

# Basic Television

SECOND EDITION

# BASIC TELEVISION

## *Principles and Servicing*

BERNARD GROB

*Instructor, RCA Institutes, Inc.*

Second Edition

McGRAW-HILL BOOK COMPANY, INC.

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**BASIC TELEVISION**

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**BASIC TELEVISION: PRINCIPLES AND SERVICING**

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*To Ruth and Harriet*

## PREFACE

This book presents a comprehensive course in television, including color television, for radio servicemen and technicians. The practical explanations of television principles, receiver circuits, and trouble shooting are planned for the beginning student in television. For each topic the essential points are first described in direct, simple terms, and then details are added progressively to provide a thorough explanation of the subject. This method of presentation, based on the author's experience in teaching television courses for many years, has proved very successful in helping students to understand television.

The book is a suitable text for television courses that follow a course in radio fundamentals. Similarly, Chapter 24 on Color Television can be used for a short course in color television receivers after the students have learned the essentials of black-and-white television receivers. No mathematics other than arithmetic and some simple algebra is needed to study the book, and even this is limited for the most part to the chapter on Practical Design of Video Amplifiers. Many simplified drawings, complete schematic diagrams, and photographs are used to illustrate each topic, in addition to practical hints and concrete examples. Charts and tables are used throughout the book to summarize a large amount of useful information and to emphasize important comparisons. Review questions are included at the end of each chapter.

The second edition, like the first presents the basic material first to allow continuity from topic to topic throughout the book. First the principles of the television system are explained, then the general requirements of television receivers, followed by a detailed explanation of receiver circuits. A separate chapter on each receiver section allows a thorough analysis of typical circuits. The advanced circuits and techniques of modern television receivers, including provision for the u-h-f television channels, are included. There are two chapters on the video amplifier circuits and on the deflection circuits, because of their importance. Each chapter on receiver circuits is concluded with essential trouble-shooting principles used for localizing troubles to one section of the receiver. The organization and continuity make the book suitable for independent home study.

Most of the material in the first edition has been rewritten and new illustrations added to bring this revision completely up to date and provide more practical information on servicing. There is a thorough explanation of color television. Receiver requirements for the u-h-f television channels are integrated into the discussions of r-f tuners, antennas, and transmission lines. Intercarrier sound is emphasized, and the discussions of a typical receiver are based on the use of intercarrier sound. The ratio detector circuit is explained more thoroughly. Automatic gain control has been expanded into a separate chapter. Automatic-frequency-control circuits for the horizontal deflection oscillator are fully explained. Details are included on horizontal output circuits with reaction scanning, boosted B+, and flyback high voltage. The chapter on receiver servicing describes principles of localizing troubles for common service problems, with additional practical information on r- and i-f alignment. The tables in this chapter summarizing receiver troubles have been enlarged, and cross references added for the sections in each chapter giving detailed explanations of the trouble symptoms.

The McGraw-Hill text-film series *Basic Television—Principles and Servicing* (six 16mm motion-picture films) is closely correlated with this text, as follows: *The Television System* is correlated with Chaps. 1 to 5; *Antenna Installation*, with Chap. 21; *Television Receivers*, with Chap. 8; *Localizing Troubles*, with Chap. 8; *Deflection Circuits*, with Chap. 18; and *Practical TV Alignment*, with Chaps. 19 and 23. There is also a series of six 35mm follow-up filmstrips with captions in the form of questions. These films and filmstrips are available from the Text-Film Department of the McGraw-Hill Book Company.

My former fellow instructors Mark Karpeles and the late Joseph Powder checked the manuscript of the first edition, besides contributing valuable material, especially on video amplifiers, and I am deeply grateful for their wholehearted assistance. For the second edition, my good friend Milton Kaufman, engineer of the CBS-Columbia Color Engineering Laboratories, reviewed the manuscript and made many helpful suggestions on color television, deflection circuits, and trouble shooting.

The schematic diagrams and photographs in the book have been made available by many organizations, as noted in each illustration, and this courtesy is gratefully acknowledged. Many of the photographs of trouble symptoms in the picture are from the booklet *TV Servicing* and the *RCA Television Picto-O-Guide*, by John R. Meagher, published by the RCA Tube Department. Photographs have also been used from the Home Study Television Servicing Course published by RCA Institutes, Inc. Schematic diagrams and photographs provided by C. E. Welsher, manager, Technical Publications and Service Clinic Group of RCA

Service Company, Inc., have been helpful. The RCA Service Company also supplied color illustrations for the chapter on Color Television. The Technical Writing Department of Admiral Corporation has provided schematic diagrams and helpful information on Admiral television receivers.

The photographs in Figs. 8-4, 19-17, 20-5, 20-12, 20-16, 21-2, 21-4, and 23-7 are from the correlated motion pictures and filmstrips of the text-film series *Basic Television—Principles and Servicing*. Photographs have also been provided by the Philco Corporation, Sylvania Electric Products, Inc., Zenith Radio Corp., Allen B. DuMont Laboratories, Inc., and other manufacturers, as noted in each illustration.

It is a pleasure to thank my wife Ruth for her excellent work in typing the manuscript.

BERNARD GROB

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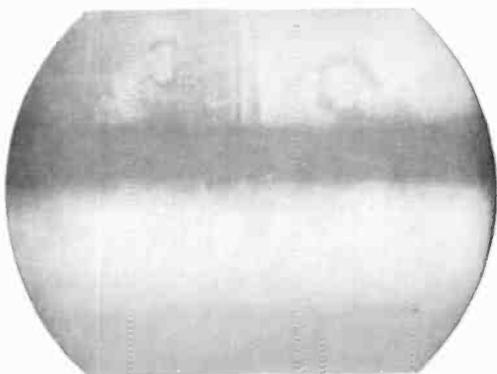


PLATE II. 60-cps hum voltage in the  $Q$  signal. (RCA.)



PLATE III. Color television picture reproduced with  $I$  signal but no  $Q$  signal. (RCA.)



PLATE IV. Color television picture reproduced with  $Q$  signal but no  $I$  signal. (RCA.)

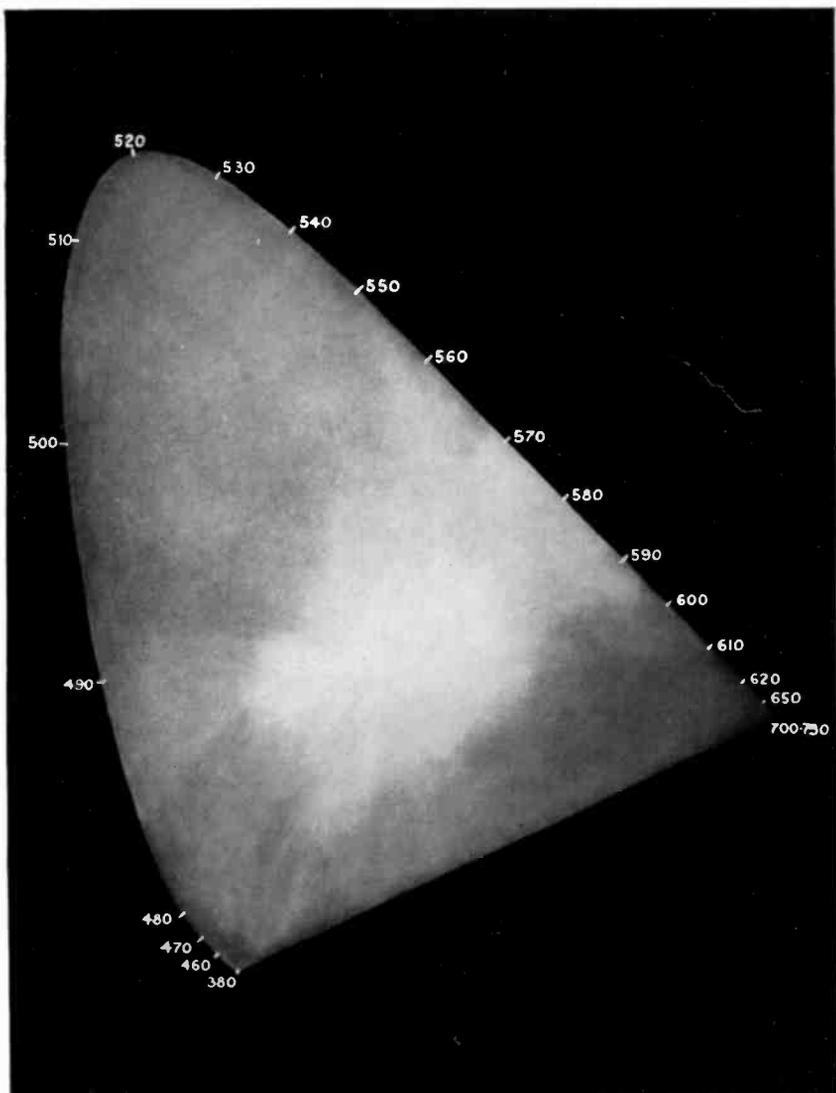


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## CHAPTER 1

### THE TELEVISION SYSTEM

Television means "to see at a distance." Our practical television system is a method of transmitting and receiving a visual scene in motion by means of radio broadcasting. The sound associated with the scene is transmitted at the same time, to provide a complete sight and sound reproduction at the receiver of the televised program. Although the end



FIG. 1-1a. A still picture.

result required is a motion picture, television is basically a system for reproducing a still picture such as a snapshot. Many of these still pictures are shown one after the other in rapid sequence during each second, to give the illusion of motion. Therefore, the first requirement of the television system is that it be capable of transmitting and receiving a simple still picture.

**1-1. Picture Elements.** A still picture is fundamentally an arrangement of many small dark and light areas. In a photographic print the fine grains of silver are distributed over the picture in a manner that provides the differences in light and shade needed to reproduce the image. When a picture is printed from a photoengraving there are many small black printed dots in the reproduction, which form the image. Looking at the still picture and its magnified view in Fig. 1-1, it can be seen that the



FIG. 1-1b. Magnified view of still picture.

printed picture is composed of small elementary areas of black and white. Note that the number 3 on the player's back is black because it contains many black printed dots of large diameter, while the uniform is white because the printed dots are small and widely separated. Shades of gray between black and white are reproduced in the same way by the proper distribution of black and white areas in the picture. This basic structure of a picture is evident in newspaper photographs. If they are examined closely, the dots will be seen because the picture elements are relatively large.

Each small area of light or shade is a basic element of the picture. These are called *picture elements*, since they contain the visual information in the scene. If all the elements are transmitted and reproduced in the same degree of light or shade as the original and in proper position, the

picture will be reproduced. Assume that it is desired to transmit an image of the black cross on a white background shown at the left in Fig. 1-2 to the right side of the figure. The picture is divided into the elementary areas of black and white shown. Those picture elements corresponding to the background are white, while the elements forming the cross are black. When each picture element is transmitted to the right side of the figure and reproduced in the original position with its shade of black or white, the original image is duplicated.



FIG. 1-2. Reproducing a picture by duplicating the picture elements.

**1-2. Transmitting and Receiving the Picture Information.** In order to transmit and reproduce the visual information corresponding to a pic-

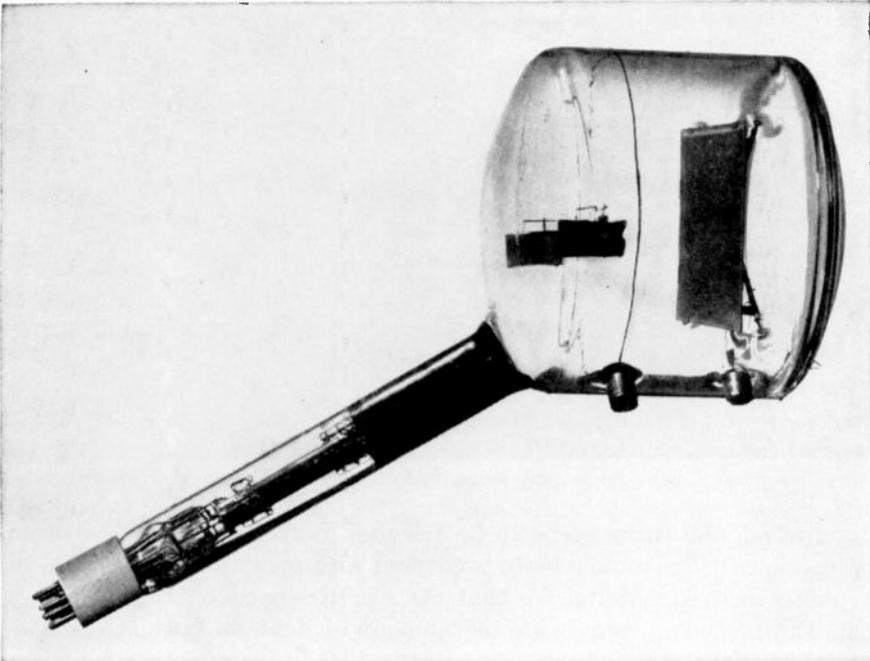


FIG. 1-3. The iconoscope camera tube. The electron-gun structure in the narrow part of the tube produces a beam of electrons aimed at the mosaic image plate. (RCA.)

ture element, the television system requires a camera tube and an image-reproducing tube. The camera tube has the function of producing an electric signal that corresponds to the visual information in a picture element, while the image-reproducing tube must be able to convert the

electric-signal voltage back into a visual image that is the duplicate of the original picture element. In comparing this with the more familiar system of transmitting sound, the camera tube corresponds to the microphone at the broadcast station and the image-reproducing tube corresponds to the loudspeaker at the receiver.

A camera tube such as the one shown in Fig. 1-3 can be used at the television broadcast station to produce an electric-signal voltage proportional to the brightness of a picture element. This particular type of camera tube, the *iconoscope*, has a mosaic plate about 3 by 4 in. in size,

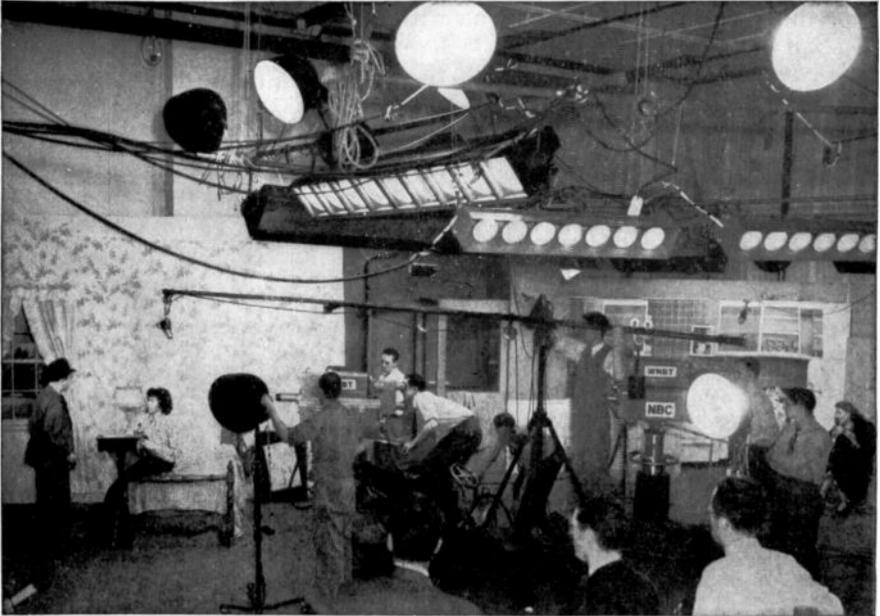


FIG. 1-4. A television studio scene. (*National Broadcasting Company.*)

upon which the entire scene to be televised is focused by means of an optical lens. The mosaic plate is covered with many minute globules of a photosensitive material, so that the electric-signal output from any part of the plate varies with the amount of light on that area. The function of the mosaic plate is similar to that of the photographic plate or film of an ordinary camera in that the optical image is focused upon it. Because it responds to variations in light intensity, the mosaic plate can translate a picture element into its corresponding electric signal.

The way that a scene is televised in the studio of the broadcast station by means of a television camera is shown in Fig. 1-4. Note the microphone on the boom overhead for the associated sound, and the television cameras in the illustration. Each contains a camera tube for providing

the required electric-signal output that corresponds to the picture elements. Several cameras are used to obtain different camera angles for the picture.

The transmitting and receiving arrangement for the television system is illustrated in Fig. 1-5, for both the picture and sound signals. The desired sound is converted by the microphone to an audio signal, which is amplified for the sound signal transmitter. For transmission of the picture, the camera tube converts the visual information into electric signals corresponding to the picture elements in the scene being televised. These electrical variations become the *video* signal, which contains the desired

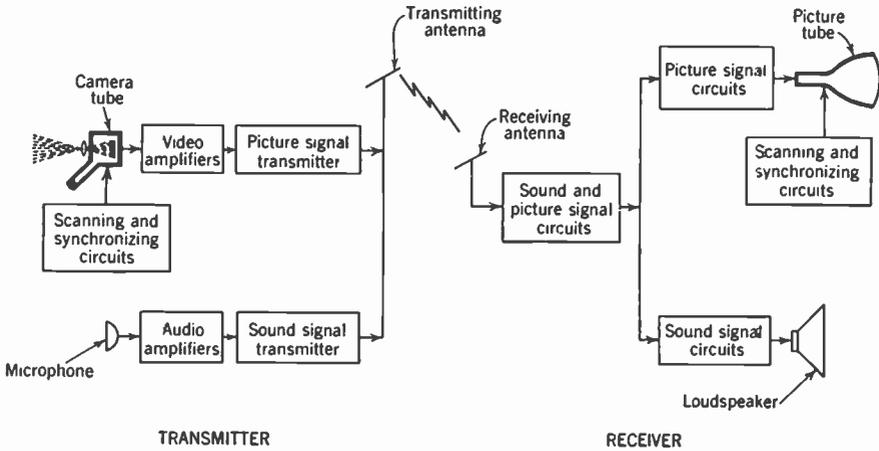


FIG. 1-5. Block diagram of the television system.

picture information. The video signal is amplified and coupled to the picture signal transmitter. At the transmitter, the video signal modulates a higher-frequency carrier wave to produce the transmitted picture signal, as in standard radio broadcasting; the audio signal modulates a separate carrier wave to produce the transmitted sound signal. The transmitter output is then coupled to the transmitting antenna so that the picture and sound signals can be radiated.

The receiving antenna intercepts the radiated picture and sound carrier signals, which are then amplified and detected in the receiver. The detector output includes the desired video signal containing the information needed to reproduce the picture. Then the recovered video signal is amplified and coupled to an image-reproducing tube that converts the electric signal back into picture elements in the same degree of black or white.

The image-reproducing tube, commonly called a *picture tube* or *kinescope*, is very similar to the cathode-ray tube used in the oscilloscope. A

typical picture tube is shown in Fig. 1-6. The glass envelope contains an electron-gun structure which provides a beam of electrons aimed at the fluorescent screen. When the electron beam strikes the screen, it emits light, and the screen becomes bright in proportion to the intensity of the electron beam.

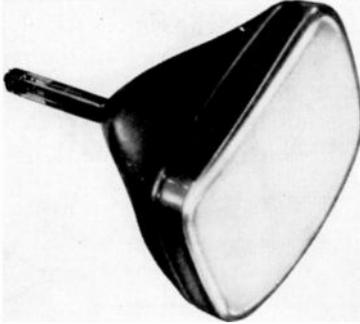


FIG. 1-6. A typical picture tube. The electron-gun structure in the narrow part of the tube produces a beam of electrons that is accelerated to the screen. (*Sylvania Electric Products, Inc.*)

Varying the intensity of this beam by varying the voltage applied to the control grid of the tube changes the intensity of the spot of light on the screen. When signal voltage makes the control grid of the kinescope less negative, the beam current is increased, making the spot of light on the screen brighter. More negative grid voltage reduces the brightness. If the grid voltage is negative enough to cut off the electron-beam current of the kinescope, there will be no light on the screen. This corresponds to black.

The video signal voltage corresponding to the desired picture information is coupled to the control grid of the picture tube so that the picture elements can be reproduced on the kinescope screen. A typical direct-view television receiver is shown in Fig. 1-7, the

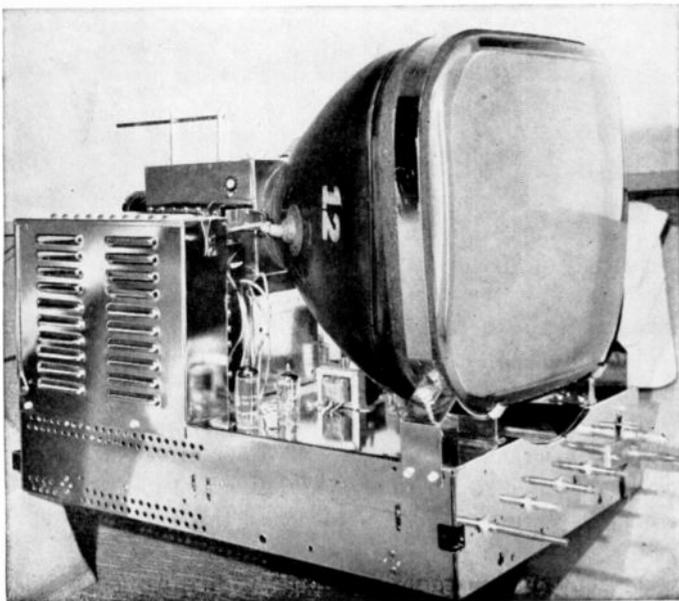


FIG. 1-7. A direct-view table-model television receiver. (*Hallcrafters Co.*)

reproduced image on the screen of the picture tube being viewed directly from the front of the table-model receiver.

**1-3. Scanning.** Although one element is not enough to reproduce a picture, the same process is carried out for all the picture elements in successive order, and each element is positioned correctly on the picture tube screen to reproduce the entire picture. This is done in the same way that a written page is read to cover all the words in one line and all the lines on the page. Starting at the top left of the frame, which is the complete picture, the scene is scanned from left to right and from top to bottom one line at a time, as illustrated in Fig. 1-8. This method of searching the picture for its information, called *horizontal linear scanning*, is the method used for scanning, both at the transmitter and at the receiver.

The scanning is done by a very narrow beam of electrons produced by the electron-gun structure contained within the camera tube and the picture tube. At the transmitter the electron beam is made to scan the image plate in the camera tube one element at a time and to cover one horizontal line after another so as to cover the entire picture. At the receiver the electron beam in the kinescope is made to scan the screen in the same way. The scanning sequence is this:

1. The beam is made to sweep across one horizontal line, covering all the picture elements in the line.

2. At the end of the line the beam is returned very quickly to the left side to begin scanning the next horizontal line. No picture information is scanned during this retrace time, as both camera tube and picture tube are blanked out for this period. The retraces must be very rapid, therefore, since they are wasted time in terms of picture information.

3. When the beam is returned to the left side, its vertical position is lowered so that the beam will scan the next lower line and not repeat over the same line. This is accomplished by the vertical scanning motion of the beam, which is provided in addition to horizontal scanning.

The number of scanning lines for one complete picture should be large in order to include the highest possible number of picture elements and, therefore, more picture details. However, other factors limit the choice of the number of scanning lines, and it has been standardized at a total of 525 for one complete picture or frame. This is the optimum number of scanning lines for the present television broadcast channels.

**1-4. Motion Pictures.** With all the picture elements in the frame tele-vised by means of the scanning process, it is also necessary to present

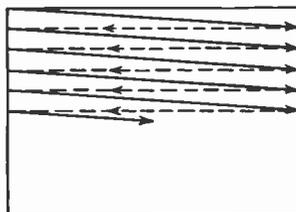


FIG. 1-8. Horizontal linear scanning procedure. The dashed lines indicate horizontal retraces.

the picture to the eye in such a way that any motion in the scene appears on the screen as a smooth and continuous change. In this respect the television system is very similar to motion-picture film practice.

Figure 1-9 shows a strip of motion-picture film. Note that it consists of a series of still pictures with each picture frame differing slightly from the preceding one. Each frame is projected individually as a still picture, but they are shown one after the other in rapid succession to produce the illusion of continuous motion in the scene. In standard commercial motion-picture practice, 24 frames are shown on the screen for every second during which the film is projected. A shutter in the projector rotates in front of the light source and allows the film to be projected on the screen when the film frame is still, but blanks out any light from the screen during the time when the next film frame is being moved into position. As a result, a continuous and rapid succession of still-film frames is seen on the screen. With all light removed during the change from one frame to the next, the eye sees a rapid sequence of still pictures that provides the illusion of continuous motion.



FIG. 1-9. A strip of motion-picture film. (Eastman-Kodak Company.)

This illusion of motion is possible because of a fortunate property of the human eye. The impression made by any light seen by the eye persists for a small fraction of a second after the light source is removed. Therefore, if many views are presented to the eye during this interval of persistence of vision, the eye will integrate them and give the impression of seeing all at the same time. It is this effect of persistence of vision

that makes possible the televising of one basic element of a picture at a time. With the elements scanned rapidly enough, they appear to the eye as a complete picture unit with none of the individual elements separately visible. To have the illusion of motion in the scene also, enough complete pictures must be shown during each second to satisfy this persistence-of-vision requirement of the eye. This can be done by having a picture repetition rate greater than 16 per second. The repetition rate of 24 pictures per second used in motion-picture practice is satisfactory and produces the illusion of motion on the screen.

However, the rate of 24 frames per second is not rapid enough to allow the brightness of one picture to blend smoothly into the next through the time when the screen is blank between frames. The result is a definite flicker of light that is very annoying to persons viewing the screen, which is made alternately bright and dark. The extent to which this flicker can be noticed depends on the brightness of the screen, the flicker effect being worse for higher illumination levels. In motion-picture practice the problem of flicker is solved by running the film through the projector at a rate of 24 frames per second but showing each frame twice so that 48 pictures are flashed on the screen during each second. A shutter is used to blank out light from the screen not only during the time when one frame is being changed for the next but once between. Thus each frame is projected twice on the screen. There are 48 views of the scene during each second and the screen is blanked out 48 times per second, although there are still the same 24 picture frames per second. As a result of the increased blanking rate, flicker is eliminated.

**1-5. Frame and Field Frequencies.** A similar process is carried out in the television system to reproduce motion in the picture. Not only is each picture broken down into its many individual picture elements, but the scene is scanned rapidly enough to provide sufficient complete pictures or frames per second to give the illusion of motion in the reproduced scene on the picture tube screen. Instead of the 24 in commercial motion-picture practice, however, the frame repetition rate is 30 per second in the television system. This repetition rate provides the required continuity of motion.

The picture repetition rate of 30 per second is still not rapid enough to overcome the problem of flicker at the light levels encountered on the picture tube screen. Again the solution is similar to motion-picture practice; each frame is divided into two parts, so that 60 views of the scene are presented to the eye during each second. However, the division of a frame into two parts cannot be accomplished by the simple method of the shutter used with motion-picture film, because the picture is reproduced one element at a time in the television system. Instead, the same effect is obtained by a method of interlaced horizontal linear scanning that divides the total number of lines in the picture frame into two groups of lines called *fields*. Each frame is divided into two fields, one field containing the odd-numbered, the other the even-numbered scanning lines. The repetition rate of the fields is 60 per second, since two fields are scanned during a single frame period and the frame frequency is 30 cps. In this way 60 views of the picture are presented to the eye during one second, providing a repetition rate great enough to eliminate flicker.

The frame repetition rate of 30 is chosen in television, rather than the 24 of commercial motion pictures, primarily because most homes in the United States are supplied with 60-cycle a-c power. Having the frame

repetition rate of 30 per second makes the field repetition rate exactly equal to the power-line frequency of 60 cycles per second. This simplifies the problem of filtering the power supplies at the receiver and transmitter.

**1-6. Synchronization.** There are 30 complete frames scanned during each second and the total number of horizontal scanning lines is 525 for each frame, which means that  $525 \times 30$ , or 15,750, horizontal lines are scanned per second. This sets the line-scanning rate at 15,750 cps. Since each frame is scanned in two fields, the standard field-scanning rate is  $30 \times 2$ , or 60, cps. Not only must the line- and field-scanning frequencies be standardized, but the scanning at the receiver and the scanning at the broadcast station must be exactly synchronized. Time in scanning corresponds to distance in the image. As the electron beam in the camera tube scans the image, the beam covers different elements of the image and provides the corresponding picture information; when the electron beam scans the screen of the picture tube the scanning must be exactly timed to assemble the picture information in the correct position. Otherwise, the electron beam in the picture tube can be scanning the part of the screen where a man's mouth should be while at that time the picture information being received corresponds to his nose. Therefore, in order to keep the transmitter and receiver scanning in step with each other, special synchronizing signals are transmitted with the picture information for the receiver. These timing signals are rectangular pulses used to control both transmitter and receiver scanning.

The synchronizing pulses are transmitted as a part of the complete picture signal for the receiver, but they occur during the blanking time when no picture information is transmitted. The picture is blanked out for this period while the electron beam retraces. A horizontal synchronizing pulse at the end of each horizontal line begins the horizontal retrace time, and a vertical synchronizing pulse at the end of each field begins the vertical retrace time, thus keeping the receiver and transmitter scanning synchronized. Without the vertical field synchronization, the reproduced picture at the receiver does not hold vertically, rolling up or down on the kinescope screen. If the scanning lines are not synchronized, the picture will not hold horizontally, as it slips to the left or right and then tears apart into diagonal segments.

In summary, then, the horizontal-line-scanning frequency is 15,750 cps, and the frequency of the horizontal synchronizing pulses is also 15,750 cps. The frame repetition rate is 30 per second, but the vertical-field-scanning frequency is 60 cps and the frequency of the vertical synchronizing pulses is also 60 cps.

**1-7. Picture Qualities.** The excellence of the picture as it appears to the observer viewing the image at the receiver depends on the following factors.

**Brightness.** This is the over-all or average intensity of illumination in the reproduced picture. The brightness of the picture must be great enough to provide a picture that can easily be seen. It is also desirable that the picture be bright enough to permit comfortable viewing in daylight or in a room with normal lighting. This presents a problem because the fluorescent screen of the kinescope is illuminated on only one small area at a time, thus making the over-all brightness of the complete picture much less than the actual spot brightness. Suitable brightness is especially difficult to obtain in a big picture because the light intensity must be distributed over a large area.

**Contrast.** By contrast is meant the relative difference in intensity between black and white parts of the reproduced picture, as differentiated

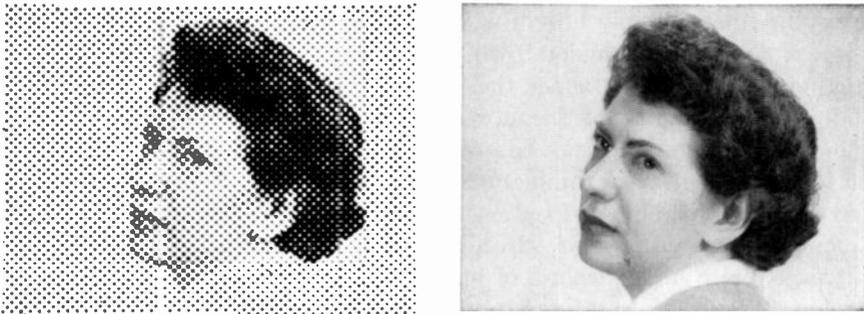


FIG. 1-10. How the picture quality is improved with more detail. (a) Coarse structure having low resolution and poor definition; (b) fine detail having high resolution and good definition.

from brightness, which is the average intensity. The contrast range should be great enough to produce a pleasing picture, with bright white and complete black for the extreme intensity values. Weak contrast, which is caused by insufficient video signal amplitude, results in a weak picture that is soft in appearance but dull and flat. Excessive contrast makes the picture appear hard, usually with distortion of the gray values.

**Detail.** The quality of detail in the reproduced picture, which is also called *resolution* or *definition*, depends on the number of basic picture elements that can be reproduced. With a great number of small picture elements, the fine detail of the image is evident. If only the larger elements can be reproduced, much of the pleasing quality of the original scene will be lost in the reproduction. Therefore, as many picture elements as possible should be reproduced on the screen of the picture tube in order to have a picture with good definition. This makes the picture clearer, as small details are evident and objects in the scene are outlined sharply. Good definition also gives apparent depth to the picture by bringing in details of the background.

The improved quality of a picture with more detail can be seen in Fig. 1-10, which shows how a larger number of picture elements can increase the definition. With 525 scanning lines per frame, and the fixed bandwidth of the transmission channels used in our commercial television system, the amount of detail is limited to a maximum of approximately 150,000 picture elements for the entire picture, regardless of its size. This normally allows a reproduced picture with the same detail as in 16-mm motion pictures.

*Viewing Distance.* With a given size for the reproduced picture and a fixed number of picture elements, which cannot be exceeded in the television system regardless of the screen size, there is a proper viewing distance for obtaining the most satisfaction in observing the picture. If the observer is too close to the screen, the picture appears coarse, with the scanning lines possibly becoming evident. At greater viewing distances some of the fine picture detail will be wasted. The best viewing distance is about four to eight times the picture height, depending on the individual visual acuity of the observer and on picture brightness. The viewing distance should not be so close that the eye must move excessively to follow the action in the picture, nor should the viewing distance be so far that the eye remains fixed on the image as a whole, since both conditions produce eye fatigue. In addition, the viewing distance should be increased with higher values of brightness.

*Aspect Ratio.* The ratio of width to height of the picture frame is called the *aspect ratio*. This has been standardized as 4:3, making the picture wider than its height by a factor of 1.33. Only the proportions of the picture frame are standardized by the aspect ratio. The actual picture size can be anything from a few square inches to 20 by 15 ft, as long as the correct aspect ratio of 4:3 is maintained. This gives the television picture practically the same proportions as a frame in motion-picture film.

**1-8. Television Channels.** The television system must reproduce a great deal of picture information within a very short period of time. It takes  $\frac{1}{30}$  sec to scan the complete picture once, and during that time 525 horizontal lines are scanned. Therefore, it takes only  $\frac{1}{15,750}$  sec to scan one line from left to right and retrace to start another line. Also, within each horizontal line there are many picture elements. Because so much picture information must be contained in an electric signal within so short a period of time, signal voltages of high frequency are produced. These video signal frequencies are as high as 4 million cycles per second. Since the frequency of the picture carrier wave that is used to transmit the signal must be above 4 Mc, television broadcast stations use transmitting frequencies much higher than the standard broadcast band (535 to 1,605 kc) for radio broadcast stations. Also, a much wider

band of frequencies is necessary for transmitting a television program. The band of frequencies assigned to a broadcast station for transmission of their signals is called a *channel*.

Each television broadcast station is assigned by the Federal Communications Commission (FCC) a channel 6 Mc wide in one of the following television broadcast bands: 54 to 88 Mc, 174 to 216 Mc, and 470 to 890 Mc. The 54- to 88-Mc and 174- to 216-Mc bands are in the very-high-frequency (v-h-f) spectrum of 30 to 300 Mc originally used for television broadcasting. The 54- to 88-Mc band includes channels 2 to 6, inclusive, which are often called the low-band v-h-f television channels; the 174- to 216-Mc band includes channels 7 to 13, inclusive, which are often called the high-band v-h-f television channels. The 470- to 890-Mc band is in the ultra-high-frequency (u-h-f) spectrum of 300 to 3,000 Mc. This band includes the u-h-f television channels 14 to 83, inclusive. These have been added to the previous v-h-f television channels to provide more channels for the expanding television broadcast service. In all the bands, each television broadcast channel is 6 Mc wide in order to accommodate the picture carrier signal, which is amplitude-modulated by the wide range of signal-voltage frequencies produced in scanning the picture, and to include the sound signal associated with the picture.

**1-9. The Associated FM Sound Signal.** The sound associated with the televised scene is transmitted simultaneously with the picture signal in a common 6-Mc channel to permit a complete visual and sound reproduction of the televised program. The sound and picture signals are transmitted on separate carrier waves, however. Although the picture signal is amplitude-modulated, the sound is transmitted as a frequency-modulated signal. With the high transmitting frequencies and the 6-Mc channel employed for transmitting the picture signal, there is easily enough space in each television channel for the FM sound signal, providing the advantages of frequency modulation for the sound. Amplitude modulation is preferable for the picture signal, however, because there is less distortion of the picture with reception of multipath signals.

**1-10. Standards of Transmission.** The problems of televising a scene bring up many factors that make the receiver dependent upon the transmitter for proper operation. This makes it necessary to set up standards for the transmitter for use by all television broadcast stations, so that a receiver will work equally well for all stations. These have been specified as a list of transmission standards<sup>2</sup> by the Federal Communications

<sup>1</sup> The specific frequencies for all the television broadcast channels are listed in the Appendix.

<sup>2</sup> The complete technical standards are in the *Federal Communications Commission Rules Governing Radio Broadcast Services*, Part 3, Subpart E, Rules Governing Television Broadcast Stations. This also gives channel assignments by states and cities.

Commission (FCC). Several points in the standards are listed here to summarize this brief outline of the television system:

1. It is standard to scan at uniform velocity in horizontal lines from left to right, progressing from top to bottom of the image, when viewing the scene from the camera position.
2. The number of scanning lines per frame period is 525.
3. The frame repetition rate is 30 per second, and the field repetition rate is 60 per second.
4. The width of the channel assigned to a television broadcast station is 6 Mc.
5. The associated sound is transmitted as an FM signal. The sound carrier signal is included in the 6-Mc television channel.
6. The picture carrier is amplitude-modulated by both picture and synchronizing signals. The two signals have different amplitudes on the AM picture carrier.
7. The aspect ratio of the picture is 4 units horizontally to 3 units vertically, or 1.33.

**1-11. Color Television.** The basic principles of color television are the same as those of the black-and-white (monochrome) television system, with the additional requirement that the colors of the visual scene are to be reproduced in the televised scene. When the image is scanned at the broadcast station, video signals corresponding to the desired picture information are obtained for the red, green, and blue colors in the scene by using optical color filters. The color video signals are then combined in order to convert the picture information into the following two signals for transmission to the receiver in the standard 6-Mc television broadcast channel:

1. The *luminance signal*, which contains the brightness variations of the picture information, including fine details. This signal is transmitted with essentially the same standards as in monochrome television.
2. The *chrominance signal*, which contains the color information only for areas large enough to require color for suitable color reproduction. Because of the restricted bandwidth necessary for the chrominance signal in the 6-Mc broadcast channel, there is no color reproduction for small details. Both the chrominance and luminance signals are transmitted to the receiver by amplitude modulation of the picture carrier wave, as in black-and-white television broadcasting, in the station's assigned 6-Mc channel in the v-h-f or u-h-f television broadcast bands.

By means of the luminance signal alone, a conventional television receiver for monochrome reproduces in black and white a picture televised in color. There is no need for an adapter, as the monochrome receiver uses the luminance signal for a black-and-white picture reproduction without interference from the chrominance signal. In a color

television receiver the chrominance signal is combined with the luminance signal to recover the original red, green, and blue color video signals. These are then used for reproducing the picture in color on the screen of a color kinescope. Color television receivers can also reproduce in monochrome a picture televised in black and white. As a result, programs televised in color are reproduced in color by color television receivers, while monochrome receivers reproduce the picture in black and white. Television programs broadcast as a monochrome picture are reproduced in black and white by both monochrome and color television receivers.

## REVIEW QUESTIONS

1. What is a picture element?
2. What is the function of the camera tube? Of the picture tube?
3. Describe briefly how a scene is televised to be reproduced on the screen of the television receiver.
4. How is the problem of flicker overcome in commercial motion-picture practice? In television? Why is this successful in eliminating flicker?
5. How is motion in the scene reproduced in the television system?
6. Describe the horizontal linear scanning procedure.
7. How many horizontal lines are scanned in  $\frac{1}{30}$  sec?
8. Why is synchronization necessary in the television system?
9. What is the picture repetition rate in television?
10. What is the vertical-field-scanning rate, in cycles per second?
11. What is the horizontal-line-scanning rate, in cycles per second?
12. What is the standard aspect ratio for the picture reproduced on the screen of the picture tube?
13. What is the bandwidth of frequencies assigned for television broadcast channel 13? What two signals are included in the one channel?
14. How is the associated sound transmitted?
15. How does the picture quality of contrast differ from brightness?
16. Why is it necessary, for good detail in the picture, to transmit and receive a great deal of picture information within a short period of time?
17. How could the speed of the scanning beam be reduced? What effect would this have on the picture?
18. How much time in microseconds is required to scan a complete horizontal line, which includes the trace and retrace?
19. In color television broadcasting, why can the luminance signal be used in monochrome receivers to reproduce the picture in black and white?

## CHAPTER 2

### CAMERA TUBES

The picture signal begins at the television camera. Here the televised scene is converted to an equivalent electric signal that can be transmitted to the receiver. In order to accomplish this the camera must first be a viewing instrument capable of "seeing" the televised scene. As illustrated in Fig. 2-1, light from the illuminated scene is focused by means of

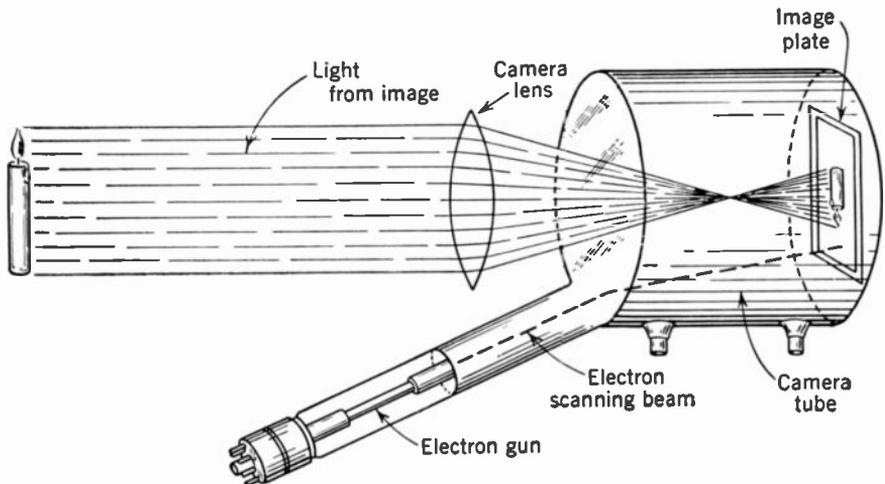


FIG. 2-1. Televising an image with a camera tube.

an optical lens to provide the desired image for the camera tube. Figure 2-2 shows a television camera with several different lenses for close-up views or long shots of the televised scene. With the scene focused on the image plate of the camera tube it is then possible to convert the variations of light intensity in the picture elements into a chain of electric impulses that correspond to the picture information, by means of the electron scanning beam and the photoelectric properties of the image plate.

**2-1. Photoelectric Effect.** Certain metals have the property of emitting electrons when light strikes their surface. The light-sensitive surface usually consists of a silver base upon which is evaporated a thin layer of a photosensitive metallic compound. Cesium oxide is often used as the photoelectric material because it is most sensitive to the light produced with incandescent lighting. When a positive collecting plate is placed close to the emitting surface, which is now a *photocathode*, the emitted

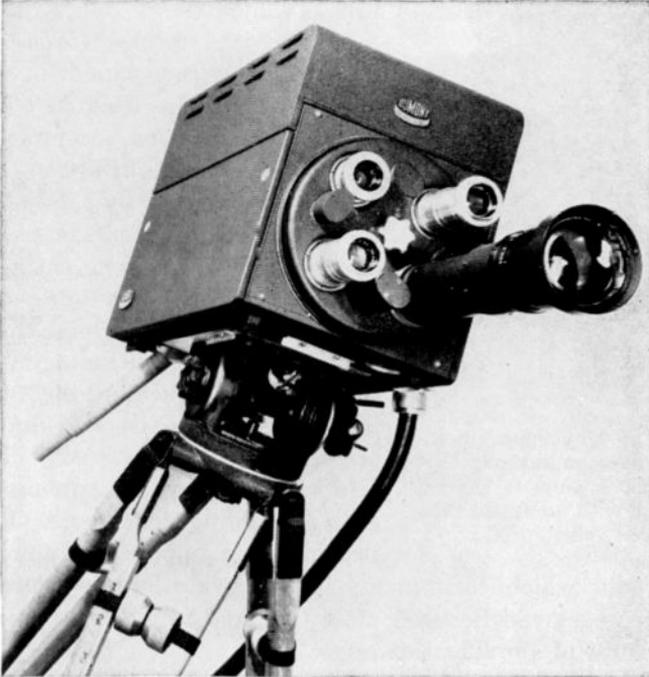


FIG. 2-2. Portable television camera, with four lenses of different focal lengths mounted on a revolving turret. The long barrel has a telephoto lens for close-ups. An image orthicon camera tube is used in this camera head. (Allen B. Du Mont Laboratories, Inc.)

electrons can be collected to produce an electric current that corresponds to the incident light. The photoelectric action is practically instantaneous, and the number of electrons emitted can be made directly proportional to the amount of incident light, thus providing an electric signal with variations that correspond exactly to the variations in light intensity. This is the fundamental action by which the optical image is converted to an electric image.

The light-sensitive surface can be enclosed in a vacuum and a positive anode provided to collect the emitted electrons, forming a phototube such

as the one illustrated in Fig. 2-3. The photoelectric tube, or *PE cell*, as it is often called, is a cold-cathode two-element tube with a large curved cathode, the inner surface of which is coated with the light-sensitive material. As large an area as possible is used for the photocathode structure in order to intercept the maximum amount of light. The narrow wire down the center of the tube is the anode.



FIG. 2-3. A typical vacuum phototube. The average cathode current for such a tube is several microamperes with an anode voltage of 100–250 volts. (RCA.)

*Phototube Amplifier Circuit.* Figure 2-4 illustrates the manner in which the phototube can be used in conjunction with a conventional amplifier circuit. When the cathode surface in the phototube is illuminated by the light source, the electron flow resulting from photoelectric emission is from cathode to the positive anode, through the anode supply, and back to cathode through the external load resistor  $R_1$ . The voltage drop across  $R_1$  produced by the photoelectric current is impressed between grid and cathode of the amplifier tube. Any changes in grid voltage vary the amount of plate current flowing in the amplifier, thus providing an appreciable signal in the

output circuit, which corresponds to the original photoelectric signal. Additional stages can be used after the first amplifier to produce the desired amount of amplification.

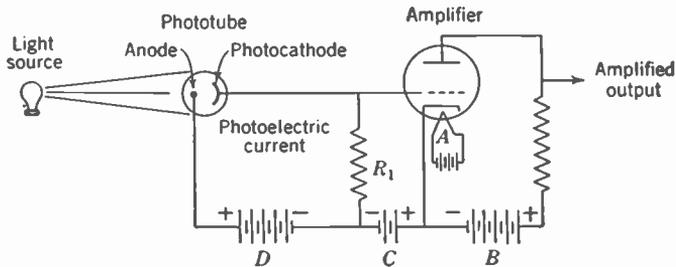


FIG. 2-4. Phototube amplifier circuit. The *A* battery supplies filament power, the *B* supply provides plate voltage and the *C* battery furnishes grid bias for the amplifier. The *D* supply makes the phototube anode positive with respect to its cathode.

*Secondary Emission and Multiplier Phototubes.* Metals have the property of emitting electrons when their surfaces are bombarded by incident

electrons having a high velocity. This is called *secondary emission*, and the electrons emitted from the metallic surface by the incident primary electrons are called *secondary electrons*. These are not the same electrons that originally strike the surface but must be new electrons ejected from the metal, because there can be more secondary than primary electrons. Aluminum, as an example, can release seven secondary electrons for each incident primary electron.

Phototubes and other low-current tubes often employ an electron-multiplier structure, making use of the secondary-emission effect to amplify the small amount of available current. In such tubes a series of cold anode-cathode electrodes called *dynodes* is mounted internally, with each at a progressively higher positive potential as illustrated in Fig. 2-5. The

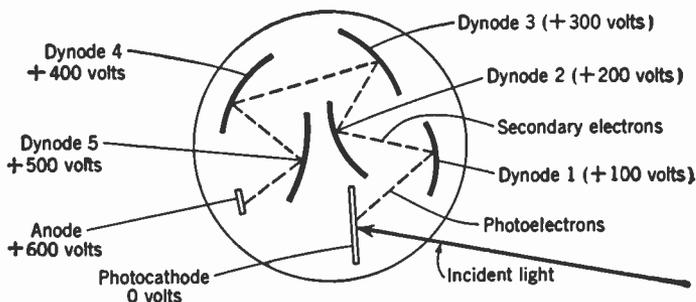


FIG. 2-5. Illustrating an electron-multiplier structure for a phototube.

few electrons emitted by the photosensitive surface are accelerated to a more positive element and in striking the dynode can force the ejection of secondary-emission electrons when the velocity of the incident electrons is great enough. The average number of electrons emitted from a dynode by each arriving electron may be only three or four, depending on the potential and the type of surface; but the number of electrons available is multiplied each time the secondary electrons strike the emitting surface of the next more positive dynode. The electron multiplier can be very useful, therefore, as a noise-free amplifier for very small photoelectric currents. Very little noise is produced while amplifying the current in an electron multiplier because it has no resistors. A relatively high voltage supply is necessary for operation, however, because each stage must be at a progressively higher potential.

**2-2. Flying-spot Camera.** The arrangement shown in Fig. 2-6 is called a *flying-spot* camera because the spot of light is made to move over the image to scan the entire picture. This system combines the phototube and amplifier circuits with an electron scanning beam to convert an optical image to the desired camera signal. The screen of a cathode-ray tube is the source of light. Within the tube, the electron-gun structure

produces a beam of electrons aimed at the luminescent screen, resulting in a small spot of light in the area where the electrons strike the screen. Deflection coils apply horizontal and vertical deflecting forces to move the electron beam and the light spot over the entire face of the tube, producing the required scanning motion. The brilliance of the light spot is not changed during the scanning but is kept at as high a value as can be obtained without burning the screen. When the scene to be transmitted is in the form of a transparent film strip or slide, as shown in the figure, the light from the scanning spot passes through the film and is collected by a phototube behind the image. The amount of light that passes through

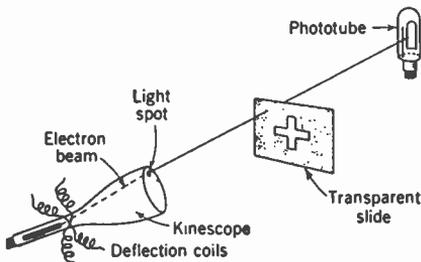


FIG. 2-6. Flying-spot scanning arrangement.

the film to the phototube varies with the density of the film. As the light spot scans the entire image to be transmitted, the output of the phototube contains signal variations for all the picture elements, thus providing the desired camera output signal.

Use of the flying-spot camera is generally limited to a small image, such as a film slide, placed close to the light source. The scanning

speed required for high-definition pictures is easily obtained, however, and special flying-spot cathode-ray tubes have been developed to produce sufficient light for the image. Sensitive phototubes containing an electron multiplier are used with the flying-spot scanner to obtain enough signal for production of a high-quality picture. Figure 2-7 shows a flying-spot camera pickup.

**2-3. Camera Pickups.** A device that can convert the light image into a corresponding electrical signal is called a *camera pickup*. This combines photoelectric conversion of light into electricity with scanning to provide signal for all the picture elements in the image, as illustrated by the flying-spot camera pickup. A pickup that includes in one unit facilities for scanning and photoelectric conversion to produce the desired camera signal output corresponding to the complete picture is called a *camera tube*. The most common types of camera tubes are the iconoscope, the image orthicon, and the vidicon.

**Camera Sensitivity.** The sensitivity of the camera pickup is the ratio of electric-signal output to the amount of light needed for producing the signal. In order to provide enough camera signal with moderate illumination, the sensitivity should be as high as possible. With increased sensitivity for the camera pickup, the associated optical system can use smaller lens apertures, improving the depth of focus and reducing the

size and cost of the optical equipment necessary to obtain enough camera signal.

*Instantaneous Pickups.* An instantaneous camera pickup is one that produces output signal for the light in a picture element only at the instant it is scanned. The flying-spot camera pickup is an example. The sensi-



FIG. 2-7. Flying-spot camera pickup using flying-spot cathode-ray tube 5WP15. (RCA.)

tivity of an instantaneous camera pickup is inherently limited to relatively low values because the illumination used to produce signal cannot be stored.

*The Storage Principle.* In order to increase the efficiency of camera tubes, several storage tubes have been developed to provide useful camera output at low light levels. In these storage tubes the effect of illumination on a picture element is allowed to accumulate between the times it is scanned in successive frames, so that the amount of camera signal avail-

able in the output is enormously increased over the amount of output available in an instantaneous system. Taking the number of picture elements in a frame as 150,000, the increased light sensitivity of a storage tube can be 150,000 times as great as in an instantaneous system if the storage tube fulfills the theoretical possibilities by storing light through the entire frame period.

*Masking Voltages.* Camera sensitivity is needed because the camera signal produced must be greater than some minimum level if it is to be useful. In practice it is impossible to amplify exceedingly small currents or voltages and obtain a useful output signal, because small random

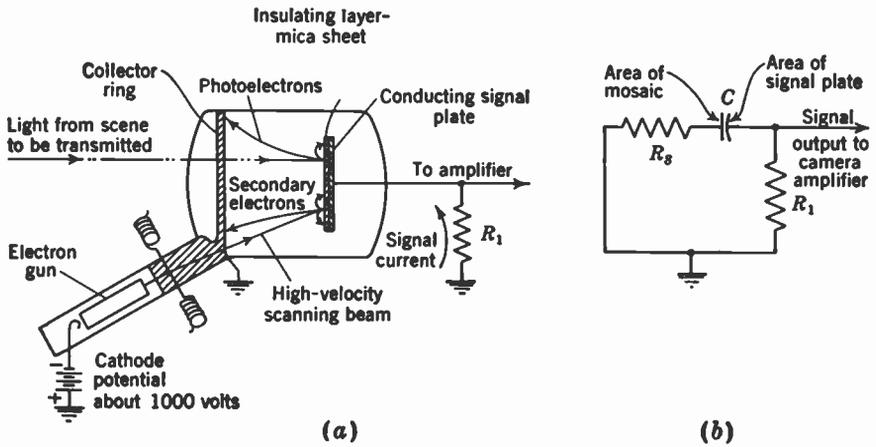


FIG. 2-8. The iconoscope camera tube.  $R_s$  is the resistance of the path for secondary electrons from the mosaic to the collector anode, which is about 5 megohms. A typical value of the load resistor  $R_1$  is 10,000 ohms. The capacitance of  $C$  is about 100  $\mu\text{mf}$  per sq cm. (a) Tube construction; (b) signal circuit.

voltages always present in the amplifier produce a spurious signal. These random currents or voltages result from heating in the circuit components and from tube noise and are amplified along with the desired signal. If the signal voltage is not large enough initially, the amplified output will not be useful because of the dominance of the spurious signal. The spurious signal has no relation to the desired signal and may be called masking voltage because it tends to obliterate the useful signal. It is called noise in a sound system. In a visual system the masking voltages produce random changes in light values that give the picture a speckled appearance which is generally described as "snow." The ratio of signal to masking voltage should be at least 20:1 and preferably more than 50:1.

**2-4. Iconoscope.** Invented by V. K. Zworykin, the iconoscope is the first of several types of camera tubes utilizing the light-storage principle to increase camera sensitivity. Excellent results are obtained with the

iconoscope where the illumination of the televised scene is 800 ft-candles,<sup>1</sup> while usable pictures are possible at lower light levels. A photograph of a typical iconoscope is shown in Fig. 1-3, and the diagrams in Fig. 2-8 illustrate construction and operation of the tube. Figure 2-9 shows how the iconoscope is mounted in a camera head for televising motion-picture film.

*The Image Plate.* The scene to be televised is focused through the glass window of the vacuum tube onto the flat image plate, called the *mosaic*

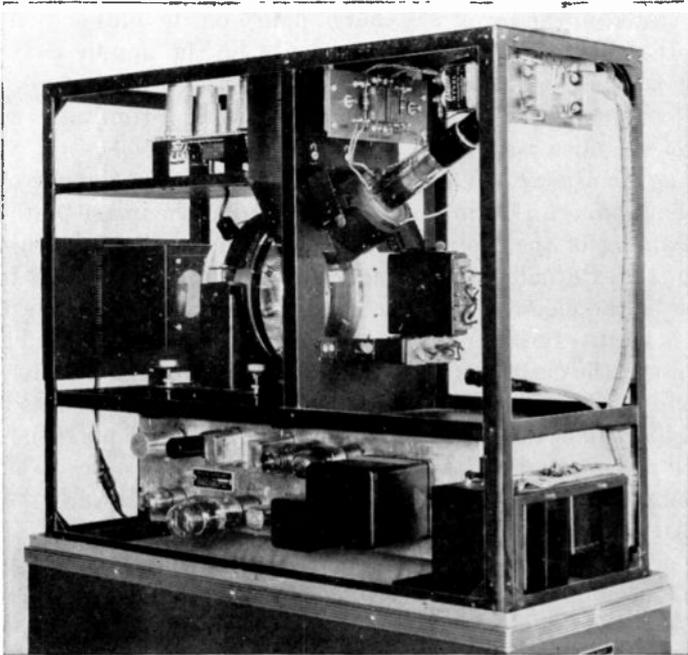


FIG. 2-9. Inside view of an iconoscope camera for film. Note back-light housing at left; rim light not shown. (Allen B. Du Mont Laboratories, Inc.)

plate because it is coated with many globules of the light-sensitive material. The mosaic plate is a very thin sheet of mica covered on one side with a large number of minute globules, photosensitized by means of an activating process that produces silver globules coated with cesium oxide and pure cesium. This side of the image plate faces the optical window and receives light from the televised scene. The opposite side of the mica sheet is coated with a conducting film of colloidal graphite, which is the signal plate. The size of the rectangular image on the mosaic plate

<sup>1</sup> The foot-candle is a unit of illumination. As an example, the desired illumination on a library reading table is about 10 ft-candles.

of a typical iconoscope is  $4\frac{3}{4}$  by  $3\frac{9}{16}$  in. Two caps sealed into the glass bulb connect internally to the signal plate and the collector ring.

An individual silver globule on the mosaic surface has a diameter of less than 0.001 in. and is insulated from neighboring globules by the mica. Each cesium-silver globule, being photoelectric, emits electrons proportional to the amount of light incident at that point and accumulates a positive charge because of the loss of electrons. Insulation by the mica between globules preserves the distribution of the charge on different parts of the mosaic. Therefore, the picture information in the optical image is stored in the form of a charge image on the mosaic plate. The value of the charge deficiency at any point on the mosaic continuously increases as light continues to strike the photoelectric surface, and the charge distribution can be maintained for as long a time as is necessary because of the mica insulation.

*Scanning the Mosaic.* The electron-gun structure in the narrow elbow of the tube produces a beam of electrons aimed at the image plate for electronic scanning of the mosaic. Horizontal and vertical deflecting coils are mounted externally on the neck of the tube to provide magnetic deflection of the electron beam to scan the entire image. Electrostatic focusing is used. In order to eliminate any obstruction in the light path to the mosaic, the electron gun is mounted at an angle to the image plate. The second anode of the electron gun is in the form of a metalized coating on the inside wall of the glass tube, which is at a positive potential of 1,000 volts with respect to the cathode of the electron gun in order to accelerate the scanning-beam electrons away from the cathode toward the mosaic plate. An inverted power supply is used, with the cathode at a high negative potential so that both the collecting ring and the image plate can be at ground potential.

*Secondary Emission.* Figure 2-8a shows that secondary electrons are emitted when the high-velocity scanning beam strikes the mosaic surface. Enough secondary electrons are ejected from the mosaic to make the potential of the point scanned approximately +3 volts. Note that this is the potential of any point on the mosaic only at the instant it is being scanned by the electron beam. The emitted secondary electrons can be collected by the anode collector ring to provide the output signal current. However, not all the secondary electrons ejected from the point on the mosaic being scanned are collected by the anode ring. Those secondary electrons not collected return to other parts of the mosaic plate in a random distribution. This redistribution current, consisting of a rain of secondary electrons, makes the mosaic plate negative, since more secondary electrons are added than are lost by photoemission. Except for the one point being scanned, therefore, the entire mosaic area becomes negatively charged because of the redistributed secondary electrons pro-

duced by the scanning beam. With a light image on the mosaic plate, photoemission makes the white parts of the picture less negative than the dark parts.

*Signal Action.* As the electron beam scans the mosaic plate, its charge distribution provides a varying secondary-emission current corresponding to the picture information. Each point ejects enough secondary electrons to change its potential to +3 volts, from some negative voltage that depends on the light value. Dark picture elements cause increases in current because more secondary electrons can be emitted from the more negative areas of the mosaic. Since white areas are less negative because of greater photoemission, these points provide less secondary-emission current. As each point on the mosaic is scanned successively, therefore, the instantaneous changes in secondary-emission current correspond to the variations of light in the image. The part of the secondary-emission current collected by the anode ring is the output signal current. The varying signal current can flow in the capacitive output circuit illustrated in Fig. 2-8b, from the mosaic to the collector ring in the tube, returning to the signal plate through the output load resistor  $R_1$ , and back to the mosaic by means of its capacitance to the signal plate through the mica dielectric. With  $R_1$  equal to 10,000 ohms and a signal current of approximately  $0.1 \mu\text{a}$  for the highlights in the scene, as typical values, the output signal voltage across  $R_1$  is  $1,000 \mu\text{v}$ .

*Shading Correction.* The secondary electrons falling back on the mosaic produce an irregular charge distribution that does not correspond to the picture information. This causes undesired variations in the output signal, therefore, which result in uneven shading of the reproduced picture. To correct for the spurious shading, it is necessary to add shading-correction signals in the output of the iconoscope, at the vertical scanning frequency of 60 cps to correct the shading from top to bottom in the picture and at the horizontal scanning frequency of 15,750 cps to correct the shading from left to right.

*Bias or Back Light.* This is a light mounted in the camera housing to illuminate the back of the iconoscope in order to increase its sensitivity.

*Rim or Edge Light.* This is a thin rectangle of light projected around the front rim of the iconoscope's image plate. The intensity of the edge lighting can be varied to minimize white flare that often appears at the edges of the picture produced by an iconoscope.

*Keystone Correction.* Because the electron gun in the iconoscope is inclined with respect to the flat image plate, the beam would scan an area shaped like a keystone, instead of the rectangular area required. The keystone effect is illustrated in Fig. 4-6. To eliminate this distortion of the scanning pattern, a keystone correction circuit is necessary with an iconoscope camera.

**2-5. Image Orthicon.** The image orthicon is the camera tube most commonly used in television broadcasting, because of its high sensitivity. Good pictures are obtained with an illumination of about 8 to 40 ft-candles in the televised scene. With the image orthicon, any scene visible to the eye can be televised.



Fig. 2-10. The image orthicon camera tube. It is approximately  $15\frac{1}{2}$  in. long, 3 in. wide at the front face, and 2 in. wide along the narrower section. (RCA.)

*Construction and Operation.* A photograph of the image orthicon camera tube is shown in Fig. 2-10, while its construction and operation are illustrated in Fig. 2-11. The tube has three main sections within the vacuumed glass envelope: the image section, scanning section, and electron multiplier. As illustrated in Fig. 2-11, light from the scene to be televised is focused onto the photocathode in the image section. Here the optical image is converted to an electrical charge image on the target plate. One side of the target plate receives the electrons emitted from the photocathode, while the opposite side of the target is scanned by the electron beam from the scanning section. As a result, signal current for the entire image is produced by the scanning beam. The signal current is then amplified in the electron-multiplier section, which provides the desired camera output signal.

*The Image Section.* The wide part of the tube is the image section. The inside of the glass faceplate at the front is coated with a photoelectric material, consisting of a silver-antimony surface sensitized with cesium, to serve as the photocathode. This semitransparent photocathode receives the light image on the front side, while electrons are released from the side facing the target.

The number of electrons emitted at any point in the photocathode is directly proportional to its illumination. Therefore, the electrons have a distribution corresponding to the light variations in the optical image, forming an electron image of the picture. This electron image extends outward from the photocathode surface like bristles of a brush. Since the target plate is about 400 volts more positive than the photocathode, the entire electron image is attracted to the target. Although the electrons tend to repel each other and break up the charge image, this scattering effect is remedied by using a long-focus coil that extends over the image section. As a result, the electron image is focused

at the target in order to produce a charge distribution on the target plate corresponding to the image. Both sides of the target plate have the required charge image.

The two-sided target is an exceedingly thin sheet of low-resistivity glass, about  $1\frac{1}{2}$  in. wide. Mounted on the side of the target toward the photocathode is a very fine and uniform wire-mesh screen, spaced approximately 0.002 in. from the glass plate. The wire screen has 500 to 1,000 meshes per inch, with an open area of 50 to 70 per cent so that the screen wires do not interfere too much with the image. The target assembly, including the glass plate and mesh screen, is connected to a

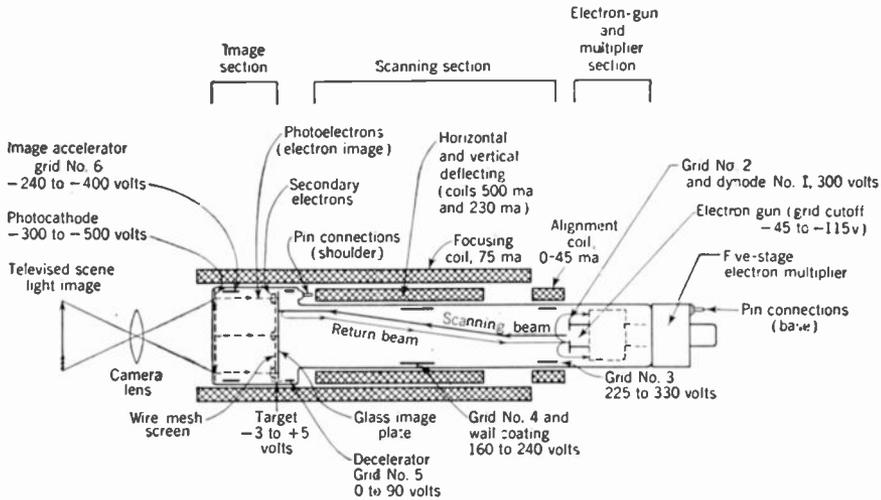


FIG. 2-11. Construction of the image orthicon. All potentials are with respect to zero volts at the cathode of the electron gun.

voltage source that can be adjusted between  $-3$  and  $+5$  volts for the best picture. Connections for all the image-section supply voltages are made through the pins mounted on the wide shoulder of the tube envelope, as can be seen in Fig. 2-10.

*The Scanning Section.* The electron-gun structure produces a beam of electrons that is accelerated toward the target. As indicated in Fig. 2-11, positive accelerating potentials of 160 to 300 volts are applied to grid 2, grid 3, and grid 4, which is connected internally to the metalized conductive coating on the inside wall of the tube. The electron beam is focused at the target by the magnetic field of the external focus coil and by the voltage supplied to grid 4. The alignment coil provides a magnetic field that can be varied to adjust the scanning beam's position exactly to correct for any mechanical misalignment of the electron gun.

Deflection of the electron beam to scan the entire target plate is accomplished by the magnetic field of the vertical and horizontal deflecting coils mounted externally on the tube.

Since the target plate is close to zero potential, the electrons in the scanning beam can be made to stop their forward motion at the surface of the glass, and then return toward the gun structure. The grid 4 voltage is adjusted to produce uniform deceleration of electrons for the entire target area. As a result, electrons in the scanning beam are slowed down near the target. Depending upon the potential of individual areas in the target, some scanning-beam electrons land on the

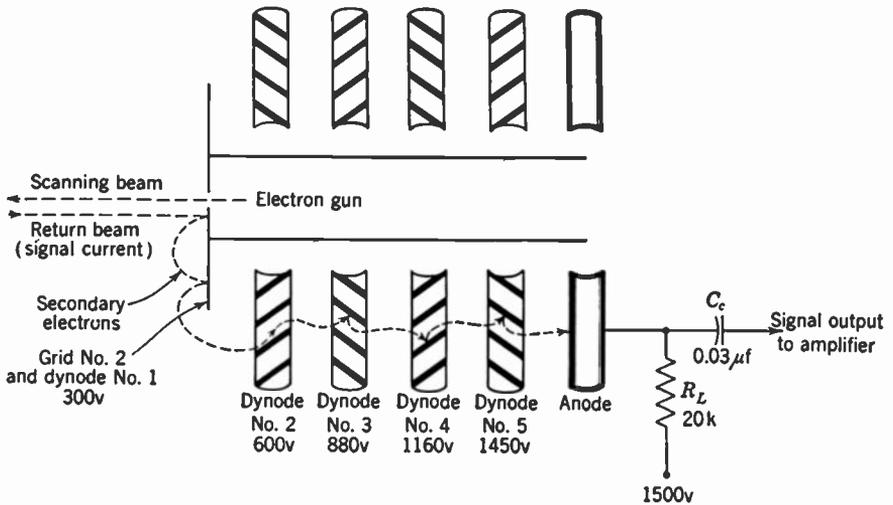


FIG. 2-12. Illustrating the electron-multiplier section of the image orthicon.

target while others stop at the glass surface and turn back to go toward the electron-multiplier structure. The electrons that return from the target provide the desired signal current. These return electrons are the primary scanning electrons, since the low-velocity scanning beam cannot produce secondary emission.

*The Electron-multiplier Section.* The return beam from the target goes to the electron multiplier for amplification of the signal current. As illustrated in Fig. 2-12, the multiplier consists of several dynodes constructed as metal disks with cutouts, like a pinwheel. Each dynode is at a positive potential 200 to 300 volts greater than the preceding dynode.

Electrons returning from the target strike a disk on grid 2, which also serves as the first dynode of the multiplier section. The second dynode is a 32-blade pinwheel mounted behind dynode 1. Primary electrons from dynode 1 strike the blades of dynode 2 to produce more secondary

electrons, which are attracted through the slots to the next stage. The same action occurs for each succeeding dynode. Five multiplier stages are used, each with a gain of approximately 4, providing a total gain of  $4 \times 4 \times 4 \times 4 \times 4$ , or about 1,000. The secondary electrons are finally collected by the anode, which is connected to the highest supply voltage of +1,500 volts, in series with load resistor  $R_L$ . The anode current through  $R_L$  has the same variations that are present in the return beam from the target, amplified by the gain of the electron multiplier. Therefore the voltage across  $R_L$  is the desired output, which is capacitively coupled to the camera signal amplifier.

*Camera Signal.* The scene to be televised is focused by means of a suitable optical lens through the glass window of the tube directly onto the photocathode. Photoelectrons are emitted from the cathode surface in direct proportion to the light and shade in the scene, converting the optical image into an electron image. The electron image is accelerated toward the target, which is at a potential about 400 volts more positive than the negative photocathode.

When the photoelectrons in the electron image strike the target, secondary electrons are emitted from the screen side of the glass plate to produce a positive charge pattern on the plate. The charge is positive because the number of secondary electrons emitted is greater than the number of primary electrons. Brighter parts of the picture produce more photoelectrons and, therefore, more secondary emission. This makes the target more positive for the bright parts of the optical image, compared with dark areas in the picture. Secondary electrons ejected from the target are collected by the mesh screen close by so that they do not accumulate and cannot retard the secondary-emission process. As a result, a charge image is produced on the target plate corresponding to the picture elements in the optical image.

The light-storage principle is utilized effectively here as light in the image continuously provides photoelectrons that produce secondary emission, and the secondary electrons are removed by the wire screen to allow the charge to accumulate on the target plate. The charge distribution is preserved because the glass target plate has high electrical resistance along the surface. In this direction the glass plate is very narrow and has enough resistance to prevent the charge distribution from equalizing itself during the frame time of  $\frac{1}{30}$  sec. In the front-to-back direction, however, the glass target plate is a conductor of very large diameter and has a low resistance. Therefore, the charge pattern appears on both sides of the target, with the brighter elements of the light image more positive than darker areas.

At the same time that the charge pattern is being formed on the target plate, it is scanned by the beam from the electron gun. The scanning-

beam electrons have very little forward velocity at the target, but where the glass plate has a positive charge some of the scanning-beam electrons are attracted to land on the target. Enough electrons must be deposited on the glass plate from the electron stream to neutralize the positive charge. Therefore, more positive parts of the target require a greater number of electrons from the scanning beam than the less positive areas, which do not need so many electrons to neutralize their charge. The electrons in the beam in excess of the amount required to neutralize the charge on the spot being scanned turn back from the target and return to the electron gun.

As the electron beam scans the target, therefore, the charge distribution corresponding to the picture elements in the light image determines how many scanning electrons are returned toward the electron gun. Darker picture elements produce less positive areas on the target, and need fewer deposited scanning-beam electrons to neutralize the charge. A larger number of electrons from the scanning beam are returned to the electron gun for these elements. Bright elements of the picture produce more positive areas on the target, and fewer electrons turn back to the electron gun from these areas. In this way, an electron stream is started on its way back to the electron gun from the target plate, with variations in magnitude that correspond to the charge distribution on the target plate and the picture elements in the optical image.

The returning stream of electrons arrives at the gun close to the aperture from which the electron beam emerged. The aperture is part of a metal disk covering the gun element. When the returning electrons strike the disk, which is at a positive potential of about 300 volts with respect to the target, they produce secondary emission. The disk serves as the first stage of an electron multiplier, therefore, since the number of secondary electrons emitted is greater than the number of incident electrons. Succeeding stages of the electron multiplier are arranged symmetrically around and back of the first stage, and secondary electrons are attracted to the dynodes at progressively higher positive potentials. Five stages of multiplication are used. The amplified output current from the final stage of the multiplier varies in magnitude with the picture information in the televised scene and is the desired camera signal.

With a signal current of  $10 \mu\text{a}$  from the highlights in the scene and a 20,000-ohm output load resistor, as typical values, the camera signal output from the image orthicon is 200,000  $\mu\text{v}$ , or 0.2 volt. This signal output from the image orthicon is more than 100 times greater than the iconoscope output, and is obtained with a much lower light level about one-fiftieth of the illumination needed for the iconoscope.

The signal action in the image orthicon can be summarized briefly as follows:

1. Light from the televised scene is focused onto the photocathode, where the light image produces an electron image corresponding to the picture elements.

2. The electron image is accelerated to the target to produce secondary emission from the glass plate.

3. The secondary emission produces on the target a pattern of positive charges corresponding to the picture elements in the scene.

4. The low-velocity scanning beam from the electron gun provides electrons that land on the target to neutralize the positive charges. Scanning-beam electrons in excess of the amount needed to neutralize the positive charges turn back from the target and go toward the electron gun.

5. As the beam scans the target, therefore, the electrons turned back from the glass plate provide a signal current that varies in amplitude in accordance with the charge pattern and the picture information.

6. The returning signal current enters the electron multiplier, where the current is amplified. The amplified current flowing through the load resistor in the multiplier's anode circuit produces the camera signal output voltage.

*Sticking Picture.* This is an image of the televised scene, with reversed black-and-white values, which tends to remain after the camera has been focused on a stationary bright image for a considerable length of time, especially if the image orthicon is operated before sufficient warm-up. The sticking picture can usually be erased, however, by operating the camera awhile with a flat image such as a blank gray wall.

**2-6. Vidicon.** As illustrated in Figs. 2-13 and 2-14, the vidicon is a small camera tube of simple construction, compared with the image orthicon. The resolution of the vidicon is less, though, and it requires 100 to 200 ft-candles of incident illumination on the televised scene. The structural arrangement is shown in Fig. 2-14. The signal electrode is a transparent conducting film on the inner surface of the glass faceplate which has a photoconductive layer on the back side, toward the electron gun. Adjacent to the photoconductive layer is a fine-mesh screen, indicated as grid 4, connected internally to the long focusing-electrode grid 3. Grid 2 is the accelerating grid and grid 1 the control grid for the



FIG. 2-13. The vidicon camera tube. It is  $6\frac{1}{4}$  in. long and 1 in. wide, approximately. (RCA.)

electron gun, which provides the electron beam for scanning the photoconductive signal plate. The rectangular scanned area is  $\frac{1}{2}$  by  $\frac{3}{8}$  in.

The signal circuit for camera signal output from the vidicon is illustrated in Fig. 2-15. Each element of the photoconductive layer on the signal plate is an insulator in the dark but becomes slightly conductive with illumination. When light strikes the image side, therefore, the opposite side becomes slightly more positive, rising toward the fixed potential on the signal plate because of the lower resistance in the photoconductive layer. The more light on any part of the plate, the more positive it becomes. As a result, the signal plate has a positive charge pattern corresponding to the picture elements in the optical image. This

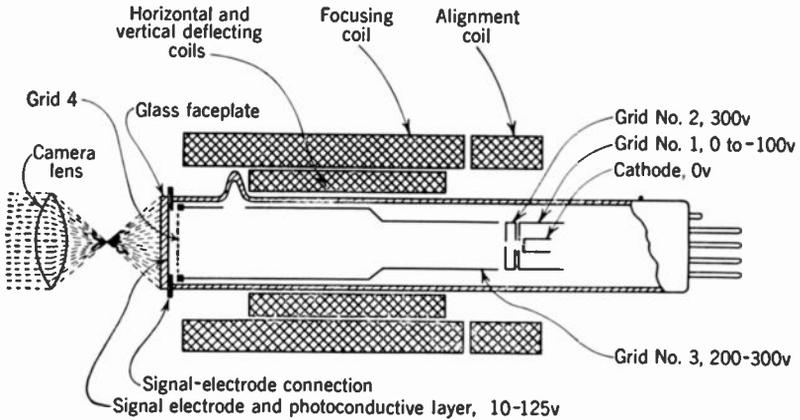


FIG. 2-14. Schematic arrangement of vidicon. (RCA.)

is scanned by the low-velocity electron beam from the electron gun. Electrons in the beam striking the signal plate are deposited until its surface potential is reduced to the cathode voltage. Excess electrons are turned back but this return beam is not utilized in the vidicon. As electrons are deposited on the photoconductive layer to neutralize its positive charge, capacitive signal current flows through the load resistor in the signal-plate circuit. With a signal current of  $0.1 \mu\text{a}$  for highlights in the scene, across a 50,000-ohm load resistor, the camera signal output voltage is  $5,000 \mu\text{v}$ .

**2-7. Camera-tube Applications.** The iconoscope and vidicon are generally used in film cameras because they provide good results with the high light level supplied by film projectors. Better resolution can be obtained with the iconoscope but the vidicon is more stable in operation, has higher sensitivity, and requires simpler control equipment. The flying-spot camera pickup can also be used for film. The image orthicon has replaced the iconoscope for studio work because of the lower light level required. Also, the image orthicon is the camera tube used in field

cameras, for operations outside the broadcast studio, such as the televising of sports and news events. Any scene with enough illumination for direct viewing either outdoors or indoors can be televised satisfactorily with the image orthicon camera because of its high sensitivity and ability to operate over a wide range of light values. In addition, the optical system used with the image orthicon is about the same size as for a 35-mm motion-picture film camera, making it suitable for the portable field equipment.

*Image Orthicon Types.* Two models of the image orthicon are manufactured in order to obtain the operating characteristics desired for different applications. One type has very high sensitivity and can operate

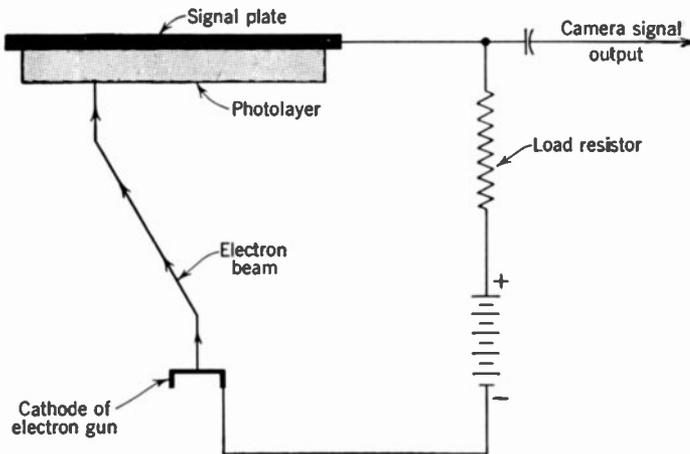


FIG. 2-15. Signal circuit for camera signal output from the vidicon.

satisfactorily over a wide range of illumination, for use in televising scenes outside the studio, where the illumination may be very low or the light level may vary considerably. The other image orthicon model, which is designed for use in the studio where the artificial lighting can be controlled, produces a picture of better quality at the expense of reduced sensitivity. The main difference in construction is the closer target-to-mesh spacing in the image orthicon for studio use.

*Monoscope.* This is a camera tube with a fixed image to provide camera signal for test purposes. Operation of the monoscope is similar to the iconoscope, but in place of the mosaic a test pattern is printed on the image plate.

#### REVIEW QUESTIONS

1. What are the two fundamental requirements of a camera pickup?
2. Show the physical arrangement of a phototube, labeling anode and cathode.
3. Describe the secondary-emission effect.
4. What is an electron multiplier?

5. Describe briefly the operation of a flying-spot camera pickup.
6. How is the light-storage principle utilized in the iconoscope?
7. Describe the signal circuit for obtaining camera signal output from the iconoscope.
8. Why is a spurious shading signal produced in the iconoscope?
9. What are the three main sections of the image orthicon camera tube?
10. Describe briefly how the desired camera signal is obtained from the image orthicon in televising a scene.
11. What are two factors that enable the image orthicon to have its high value of sensitivity?
12. Describe briefly how the desired camera signal corresponding to the optical image is obtained from the vidicon.
13. What is the function of a monoscope?
14. Give the practical applications for two camera tubes.

## CHAPTER 3

### THE ELECTRON SCANNING BEAM

Production of the camera signal requires the use of a scanning beam so that the information in all the individual picture elements is converted into a useful electric signal representing the complete picture. Similarly, the screen of the picture tube in the television receiver is scanned to reproduce the image. A narrow beam of electrons is used for scanning in both cases because the small mass of the electrons makes the electron beam ideally suited for the rapid scanning motions that must be pro-

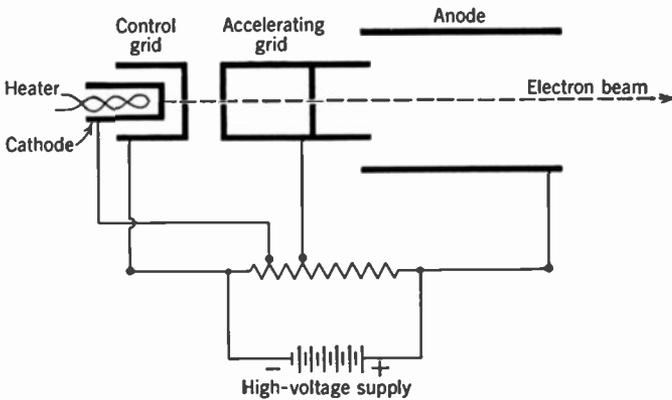


FIG. 3-1. Elements in an electron gun.

duced. The more familiar cathode-ray tube used in the oscilloscope is a common application of the electron beam and may be used as a comparison. Picture tubes used in the television system for reproducing the image are almost identical with the oscilloscope cathode-ray tube, while the camera tube uses the electron-gun structure.

**3-1. The Electron Gun.** The electron gun, which is enclosed in a vacuumed glass bulb, has the function of producing a narrow beam of high-velocity electrons. As illustrated in Fig. 3-1, the gun structure includes a heated cathode to emit electrons, a control grid to control the

flow of these electrons, and one or more additional electrodes that have a positive potential with respect to the cathode in order to accelerate the electrons away from the cathode. Instead of collecting the electrons, however, the accelerating electrodes<sup>1</sup> form a narrow electron beam that goes through to strike the scanned surface, which is the screen in a picture tube or the image plate in a camera tube. The electron gun is normally mounted in the narrow part of the tube near the base.

Since the electrons emitted from the cathode must be concentrated into a beam and pass through the grids, these have a cylindrical structure that is different from conventional vacuum-tube elements. Referring to Fig. 3-1, at the left is the cathode. This is a thin metal sleeve enclosing

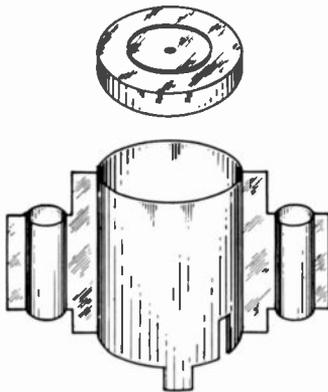


FIG. 3-2. Control-grid cylinder with aperture disk. (Sylvania Electric Products, Inc.)

the heater coil. Fitted on the cathode sleeve is a cap with a recess in the center to hold the oxide mixture that is heated to produce thermionic emission of electrons from the cathode. Directly after the cathode is the control-grid cylinder. This sleeve encloses the cathode and has a pin-hole aperture in the center to allow the electron beam to pass through. Figure 3-2 shows the construction of a typical control-grid cylinder and its disk cover with an aperture of approximately 0.040 in. diameter. The control grid is maintained at a negative potential with respect to the cathode, by means of a suitable bias voltage. The next element in the electron gun is an accelerating grid, which is also a cylinder with an aperture for the electron beam. A d-c potential of positive polarity with respect to the cathode is applied to the accelerating grid so that electrons can be attracted from the cathode. More than one accelerating grid can be used. The accelerating voltages become progressively greater and the final accelerating electrode, which is the anode, is maintained at the highest positive potential with respect to the cathode in order to accelerate the electron flow from the cathode to the scanned surface. The construction of a typical electron gun is shown in Fig. 3-3. The metal elements of the gun are generally made of nickel or a nickel alloy, mounted on ceramic insulator supports along the length of the gun structure.

The anode is generally in the form of a conductive wall on the inside of the wide flared part of the tube's envelope, as indicated by the wider

<sup>1</sup> The accelerating electrodes are sometimes called *anodes* because they have a positive potential, but the IRE standard term is *grid* for an electrode having one or more openings for the passage of electrons or ions, while an anode collects electrons.

diameter shown for the anode in Fig. 3-1. Picture tubes with a glass envelope have a black conductive coating on the inside wall, as illustrated in Fig. 3-7, covering the entire flared portion of the bulb almost up to the screen and extending into the narrow neck. The coating is usually colloidal graphite (commonly called Aquadag). For a picture tube having a metal shell instead of glass for the flared part of the envelope, the metal shell is the anode. The anode has the highest positive potential. In an electron gun with several accelerating grids, one or two of these cylindrical electrodes may be connected internally to the anode wall coating and have the same potential. A separate connector on the side of the tube is generally used for the external anode connection, instead of connecting to a pin on the tube base, when the high voltage is more than about 5 kv. In addition to serving as the final accelerating electrode, the anode wall shields the electron beam from stray electric fields.

Electrons emitted from the cathode of the electron gun are formed into a beam because of the restricted path imposed by the apertures and the acceleration produced by the positive electrodes. The electron beam has a complete circuit for current flow because of the secondary emission produced when the beam strikes the scanned surface. These secondary electrons are attracted to the positive anode, forming a complete path for current flow from cathode to the scanned surface, to the collecting anode, and back to cathode through the power supply that provides the anode voltage. The amount of beam current is very small, being in the order of microamperes. The d-c potential applied to the anode is very high, ranging from 9,000 to 80,000 volts for picture tubes in order to obtain values of beam current of several hundred microamperes to produce a bright picture on the screen. For camera tubes, however, the principal requirement of the electron gun is a very narrow, sharply defined beam, and the required beam current of approximately  $0.1 \mu\text{a}$  can be obtained with anode voltages of about 1,000 volts or less.

**3-2. First Electron Lens.** The electrons emitted from the cathode tend to form a diverging broad-angle stream because all have a negative charge and repel each other. Consequently, special attention must be given to the problem of forming the electron stream into a narrow beam. This is analogous to focusing a beam of light by means of optical lenses. Therefore, the term *focusing* is used for the action of obtaining a narrow electron

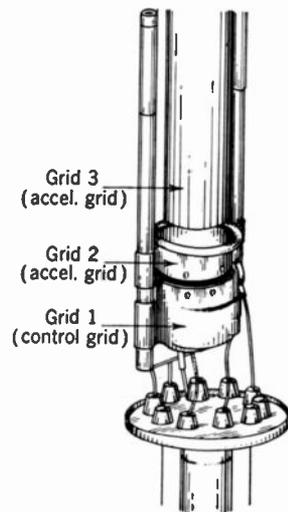


FIG. 3-3. Construction of an electron gun. (RCA.)

beam, while a focusing system in the electron gun is called an *electron lens*. Two electron lenses are generally used.

*Crossover Point.* Figure 3-4 illustrates the operation of the control grid as the first electron lens, concentrating the electrons into a converging beam. The arrows indicate the electrostatic force acting on electrons leaving the cathode. Since the control grid is negative with respect to the cathode, the electrostatic force is in a direction that tends to repel the electrons back to the cathode, with the repelling force in the direction of the arrows on the electrostatic lines of force. The anode voltages provide a forward accelerating force, however, and the net result is to bend the diverging electron paths toward the center line of the gun structure so that the electrons can travel through the grid aperture. Electrons leaving the cathode in the direction shown by the line  $KA$ , for example, are

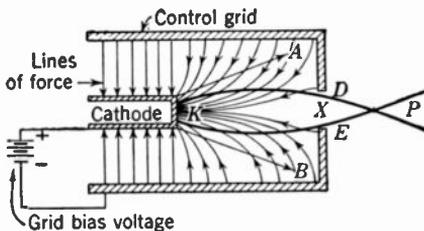


FIG. 3-4. Action of the control grid as the first electron lens.

repelled almost directly toward the center axis of the gun while they are in the region of the cathode, but the direction of the repelling force changes gradually as the control grid is approached. As a result, an electron emitted in the direction  $KA$  is made to follow a path such as  $KDP$ . Likewise, an electron emitted in the direction  $KB$  follows the curved path  $KEP$ .

Both electron paths pass through the control-grid aperture at  $X$  and cross at the point  $P$  just beyond the aperture. The same is true for any electron if its path allows passage through the grid aperture, and all useful electron paths converge at the focal point  $P$ . This is the crossover point, serving as the point source of electrons for a narrow electron beam.

*Intensity Control.* The magnitude of the control-grid voltage determines the intensity of the electron beam and, therefore, the amount of beam current that strikes the scanned surface. When the control-grid voltage is made more negative with respect to the cathode, fewer electrons from the cathode can follow paths that allow passage through the grid aperture to form the beam. Also, the attracting force on the electrons resulting from the accelerating anodes is decreased with a more negative control grid and fewer electrons can leave the cathode area. Adjustments of the control-grid voltage in the camera tube are relatively simple, the bias value being fixed for a value of beam current that produces the best camera signal. For the picture tube, the grid bias voltage is varied to adjust screen brightness, more negative voltage decreasing the illumination. The amount of negative grid voltage required to reduce the beam current to the point of visual cutoff, where no light is produced on the

screen, varies with the gun structure and anode voltage but is in the order of 50 volts. In addition to the bias, signal variations in voltage are impressed on the control grid of the picture tube to vary the beam intensity according to the desired picture information for reproduction of the image.

**3-3. Electrostatic Focusing.** Although the first electron lens converges the electron stream to provide a point source of electrons for a narrow beam, the electron paths diverge again after the crossover point and an additional lens system is needed to focus the electron beam at a point on the scanned surface. The first electron lens is fixed by the cathode and control-grid construction of the tube, but the second lens system is made variable to allow adjustment of the focus. Either of two systems is possible for the second lens, one using an electrostatic focusing arrangement similar to the first electron lens and the other employing a magnetic field for focusing.

The second electron lens for focusing the beam electrostatically can be provided by the accelerating grid and anode, as illustrated in Fig. 3-5. The focusing grid voltage is about one-fifth the anode voltage but both

are positive with respect to the cathode in order to attract electrons from the cathode. An electron in the field between the grid and the anode tends to follow along the electrostatic lines of force shown, until it strikes the more positive anode. However, the electrons in the field have a strong forward acceleration because of the anode voltage. The resultant electrostatic field near the focusing grid provides a force in a direction that moves the electrons toward the axis of the gun. As a result, the electron beam converges going from the focusing field to the more positive potential of the next accelerating field.

As illustrated in Fig. 3-5, when an electron from the crossover point at *P* enters the focusing field in the direction shown by the line *PA*, the lines of force the electron crosses force it toward the center line of the gun. Likewise, an electron tending to travel in the direction *PD* is forced to move toward the center. As the electrons approach the center, the lines of force that are crossed have a direction almost parallel to the axis of the gun, and the electrons are moved from their course by a smaller degree. Therefore, the electrons will follow curved paths such as *PAS* or *PDS* to be focused at point *S* beyond the grid. By varying the amount of voltage applied to the focusing grid, the electrostatic focusing field can be varied to produce electron paths with the curvature required to image

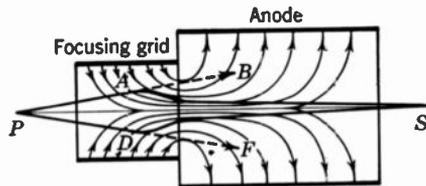


FIG. 3-5. Electrostatic field between accelerating electrodes. This serves as the second electron lens for electrostatic focusing.

the crossover point onto the scanned surface. Therefore, focusing the electron beam to obtain the smallest possible spot on the scanned surface can be accomplished by providing an adjustable focusing grid voltage in the electrostatic focusing method.

**3-4. Magnetic Focusing.** A magnetic field with lines of force parallel to the electron beam can be used for magnetic focusing, serving as the second electron lens to image the crossover point onto the scanned surface. Either a permanent magnet or an electromagnet mounted externally around the neck of the tube produces the focusing field. Figure 3-6 illustrates the magnetic field within the tube produced by direct current flowing through an external focus coil. The coil winding is concentric with the beam, circling around the tube's neck so that the magnetic field perpendicular to the wire is parallel to the beam. The direction of the magnetic field is the same as would be produced if the neck of the tube were encircled by many bar magnets, each placed lengthwise to make the

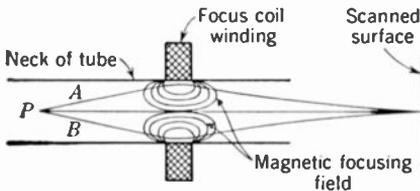


FIG. 3-6. Action of a short focusing coil.

field lines between north and south poles parallel to the beam axis. The form of a typical focus coil and its mounting on a picture tube are shown in Fig. 9-2. The focus coil is generally wound on a soft iron ring, concentrating the magnetic field and making it possible to produce the required field strength with less current. A steady direct current of about 100 ma can be used to provide the required magnetic field. Provision is made for moving the focus coil along the axis of the tube to allow rough focusing, and the amount of current flowing through the coil is adjusted by means of a variable resistance for fine focusing control. Extended focus coils that cover the entire length of the electron beam are employed with camera tubes, but the short focus coil illustrated is used for picture tubes.

Figure 3-6 shows that the magnetic lines of force in the focusing field are essentially straight and parallel to the electron beam in the region of the focus coil. Electrons in the beam traveling along the horizontal axis from the crossover point of the first electron lens have an associated magnetic field with lines of force that are circular around the beam axis in a plane perpendicular to both the electron beam and the focusing field. Because the magnetic field of the electron beam is at right angles to the focusing field, these two fields do not react with each other since the field strength is not changed with two perpendicular fields. These electrons, as a result, can proceed along the center axis toward the scanned surface, accelerated by the anode voltage. However, an electron that travels along the line  $PA$  from the crossover point  $P$  is moving at an angle

and has a component of motion perpendicular to the focusing field. Therefore, a component of its associated magnetic field can be considered in parallel with the field produced by the focus coil. Where the lines of force in the two magnetic fields are in the same direction they aid to produce a stronger field, while opposing lines of force produce a weaker field. The reaction of the two fields produces a force that moves the electrons toward the weaker field. As a result, a force is applied to those electrons that travel in such paths as  $PA$  and  $PB$  at an angle with respect to the beam axis, moving the electrons toward the center axis of the electron gun. Electrons in the outer edges of the diverging beam have paths that make the largest angle with the horizontal axis and are subject to a greater force toward the center than those electrons nearer the axis, producing a converging electron beam.

The motion of the electron resulting from the action of the magnetic field must be perpendicular both to the direction of beam current and to the focusing field. Therefore, the electrons follow a circular spiral motion toward the center axis as they are accelerated to the scanned surface by the anode voltage. By adjusting the position of the focus coil and the amount of focusing current, the electrons can be given a component of motion toward the axis of the gun which persists after the focusing field is passed, so that the electron beam converges to a point on the scanned surface. The magnetic field of the focus coil can function as the second electron lens, therefore, to produce an image of the crossover point on the scanned surface.

The direction of the focus coil current is immaterial because it is necessary only that the lines of force in the focusing field be parallel to the electron beam. A reversed field produced with opposite polarity of coil current merely reverses the direction of rotation in the spiral motion of the electrons.

**3-5. Electrostatic Deflection.** With the electron beam formed and focused by the electron gun, the next requirement is to provide means for moving the beam in the horizontal and vertical directions. The scanning procedure in television is a system for deflecting the electron beam horizontally and vertically in a standard sequence in the camera tube and picture tube. Horizontal and vertical deflection proceed simultaneously but with different velocities. Either of two deflection methods can be used, just as in the focusing arrangement. Electrostatic deflection can be accomplished by means of deflecting plates mounted internally at the neck of the tube just before it begins to flare out, while magnetic deflection makes use of deflecting coils mounted externally on the neck of the tube. This is shown in Fig. 3-7, which illustrates the electrostatic and electromagnetic types of cathode-ray tube.

*Deflecting Voltage.* Electrostatic deflection is accomplished by mount-

ing parallel plates at the end of the electron gun in a position that allows the electron beam to pass between them. As illustrated in Fig. 3-8, two pairs of deflection plates are used, with one pair mounted ahead of the other. One pair of plates is mounted in a vertical plane left and right

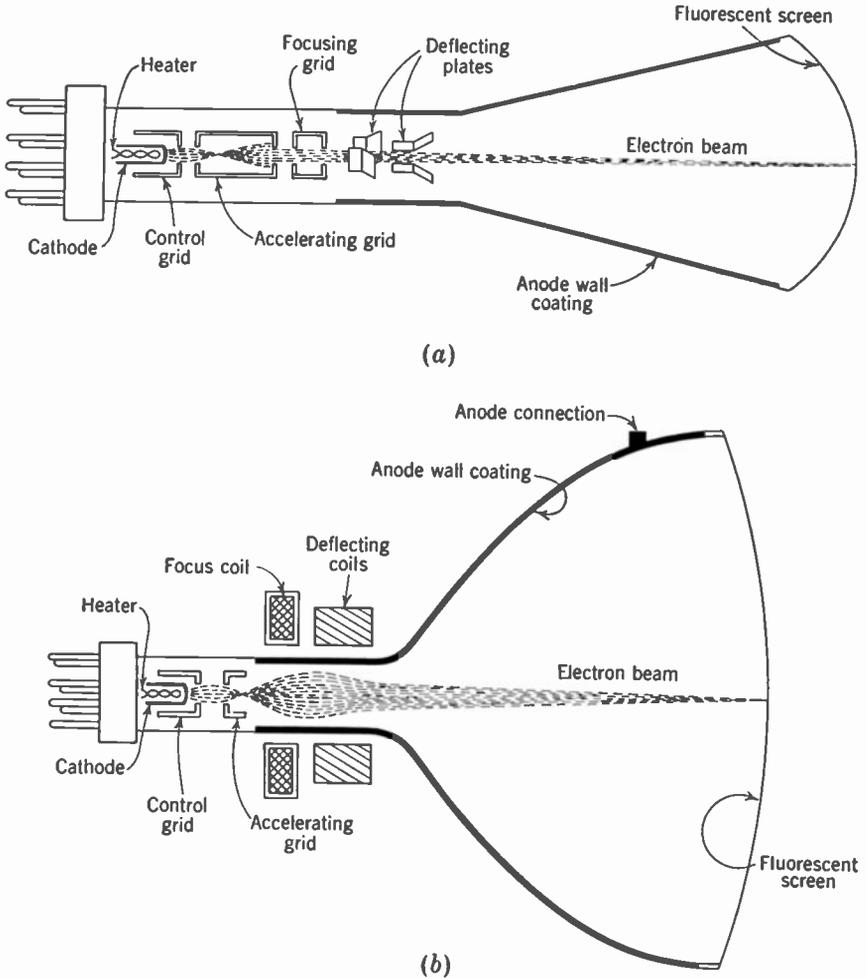


FIG. 3-7. Cathode-ray tubes. Although not shown, electrostatic focus can be used with magnetic deflection or vice versa. (a) Electrostatic cathode-ray tube; (b) electromagnetic cathode-ray tube.

of the electron beam to deflect the beam horizontally, and these are the horizontal deflection plates. The other pair of plates is placed in a horizontal plane to deflect the beam vertically; these are the vertical deflection plates.

When a positive potential is applied to a deflection plate the negatively charged electrons in the beam are attracted toward the positive plate. If one plate is made negative with respect to the other, the electron beam will be repelled away from the negative plate. Therefore, a deflecting voltage can be applied to the horizontal pair of deflection plates to produce horizontal deflection of the electron beam, either to left or right. Deflecting voltage can be applied to the vertical plates in a similar manner to deflect the electron beam up or down toward the more positive plate. As a result, deflection voltages can be applied to the horizontal and vertical plates simultaneously to deflect the beam both horizontally and vertically. While plate *A* in Fig. 3-8 is made more positive than *B* to deflect the beam upward, as an example, plate *C* can be made more posi-

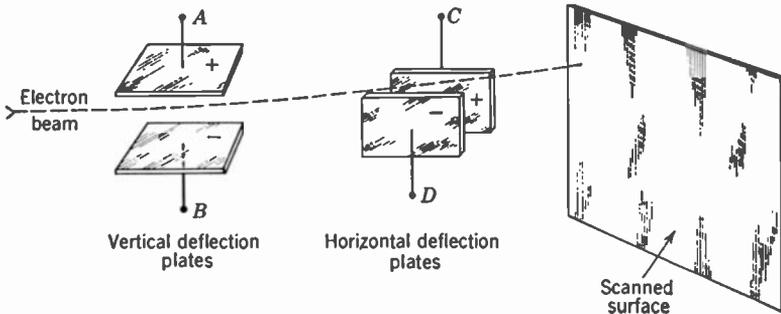


FIG. 3-8. Electrostatic deflection.

tive than *D* for deflecting the beam to the left, moving the electron beam to the upper left corner of the scanned surface in the illustration.

The deflection plates must be positive with respect to the cathode. Their potential, therefore, is approximately equal to the second-anode voltage so that the electrons can be accelerated toward the scanned surface. In addition, the potential of one plate must vary with respect to the other in order to deflect the electron beam. Establishing the potential difference necessary between a pair of plates for deflection requires a d-c deflection voltage for positioning the electron beam and a varying deflection voltage to deflect the beam in a continuous motion. The d-c positioning voltage for the deflection plates is taken from the high-voltage supply for the tube and so applied that the potential of one plate can be adjusted with respect to the other plate in order to provide a steady deflecting force for centering the beam. Provision is made for both vertical and horizontal centering. In addition to the d-c potential for centering control, a varying voltage is applied to the deflection plates so that the electron beam can be deflected continuously to produce the desired scanning pattern. An alternating voltage is applied to the hori-

zontal deflection plates to produce the horizontal scanning lines and, simultaneously, an alternating deflection voltage is coupled to the vertical deflection plates to move the beam downward slowly as it traces out the horizontal lines.

**Deflection Factor.** The amount that the electron beam is deflected varies directly with the potential difference between the two parallel plates and the distance to the scanned surface but is inversely proportional to the anode potential. With higher anode voltage, therefore, a greater deflection voltage is needed to produce a given displacement of the beam. If the anode voltage is doubled, as an example, twice as much deflection voltage will be required for the same amount of deflection.

The characteristics of the electrostatic deflection system are specified by means of a deflection factor for the tube. This is given either as the amount of d-c deflection volts required to produce 1 in. of deflection with the anode voltage equal to the maximum tube rating, or in d-c deflection

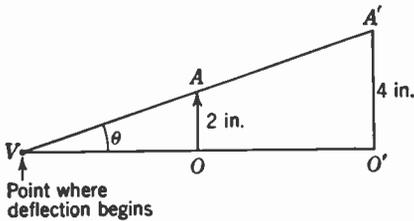


FIG. 3-9. The amount of deflection increases with the distance to the scanned surface.

volts per thousand volts of anode voltage  $E_b$  required for a deflection of 1 in. The deflection factor for electrostatic tubes is 25 to 40 d-c volts per inch per kilovolt of  $E_b$ . A tube with a maximum anode voltage of 5,000 volts may have a deflection-factor rating of 30 d-c volts per inch per kilovolt of  $E_b$ , or 150 volts per inch. In either case, a potential difference

of 150 volts between a pair of deflection plates deflects the electron beam 1 in. on the screen when the anode potential is 5,000 volts. The scanning voltage applied to the plates in this tube must have a value of 600 volts peak to peak to deflect the beam a total of 4 in., with a displacement of 2 in. from both sides of center.

The deflection factors for both pairs of plates are not exactly the same because the amount that the electron beam is displaced on the scanned surfaces varies with the distance to the point where deflection begins. In Fig. 3-9 the beam is deflected vertically from the axis  $VOO'$  through the deflection angle  $\theta$  from the point  $V$ . When the scanned surface is separated from point  $V$  by the distance  $VO$  the vertical displacement on the scanned surface is equal to  $AO$  or 2 in. However, if the distance  $VO$  is doubled the amount of deflection of the scanned surface  $A'O'$  will also be doubled to equal 4 in., although the electron beam has the same deflection angle. A larger deflection voltage is needed, therefore, for the deflection plates nearer the screen to produce the same displacement of the electron beam on the scanned surface. The deflection plates are

usually flared out in order to avoid obstructing the deflected electron beam.

**3-6. Magnetic Deflection.** Magnetic deflection of the electron beam can be used instead of electrostatic deflection. A magnetic deflecting field is established in the tube just past the electron gun to react with the magnetic field associated with the electron-beam current, producing a force that deflects the electrons at right angles to both the beam-current axis and the deflection field. The deflection field is produced by a pair of electromagnets placed around the neck of the tube. As illustrated in

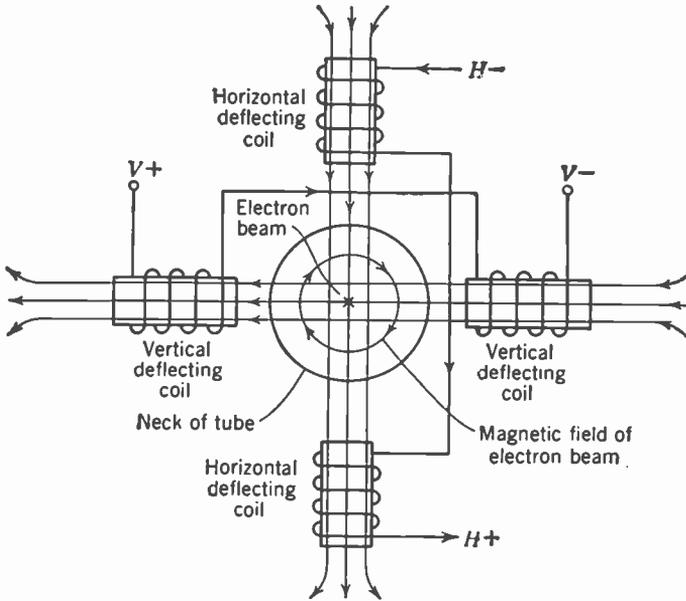


FIG. 3-10. Two pairs of coils around the neck of the cathode-ray tube for horizontal and vertical electromagnetic deflection. The electron beam will be deflected upward and to the left for the directions of current shown.

Fig. 3-10, two pairs of deflection coils are used. The two coils of one set are placed above and below the beam axis and are connected in series with each other to deflect the beam in the horizontal direction, serving as the horizontal deflection coils. The coils of the other pair are mounted left and right of the beam to deflect the beam vertically; these are the vertical deflection coils.

*Motor Action on a Conductor in a Magnetic Field.* Since the electron beam consists of electrons in motion it is equivalent to an electron current. The effect of the magnetic deflecting field on the electrons can be seen by reviewing the motor action on a wire carrying current in a magnetic field. Consider the conductor *AB* in Fig. 3-11a carrying current in a

uniform magnetic field of intensity  $H$  with lines of force perpendicular to  $AB$ . The current has its own associated magnetic field which reacts with the external field to produce a resultant force tending to move the conductor in a direction perpendicular to both the field  $H$  and the conductor. There are three factors to keep in mind regarding the resultant force on the conductor and its direction:

1. The direction of the external magnetic field  $H$  must be noted.
2. The direction of the conductor must be perpendicular to field  $H$  so the lines of force in the two fields will be parallel. The electromagnetic field associated with the current in the conductor has lines of force that are at right angles to the conductor and in a clockwise direction for the direction of current shown in the figure, in accordance with the left-hand

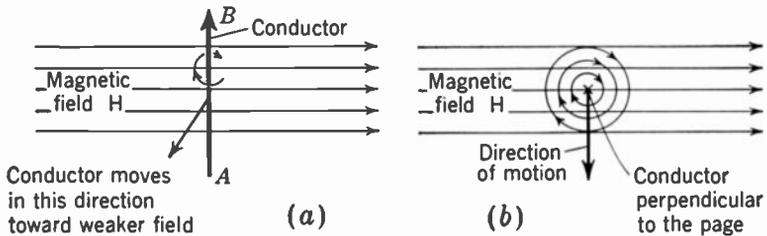


FIG. 3-11. Direction of the force on a wire carrying current in a magnetic field at right angles to the conductor.

rule. This rule states that when the conductor is grasped in the left hand with the thumb pointing in the direction of electron flow through the conductor the fingers curve in the direction of the associated magnetic field.

3. The force on the conductor results from the fact that some of the lines of force in its electromagnetic field are in the same direction as the external field to increase the field strength in some parts of the field, while other lines of force are oppositely directed to reduce the field strength. As a result, there is a force on the conductor directed from the stronger to the weaker field, producing motion in a direction perpendicular to both the conductor and the field.

In Fig. 3-11a the electron flow in the conductor is from  $A$  to  $B$ . Using the left-hand rule, the associated electromagnetic field is in a plane perpendicular to the page and in the clockwise direction. Lines of force in both magnetic fields are opposing in the direction out of the page toward the reader and aiding on the opposite side of the conductor. Therefore, the resultant force tends to move the conductor  $AB$  out of the page toward the weaker field. In Fig. 3-11b the conductor is perpendicular to the page with electron flow out of the paper. Above the conductor in the figure the lines of force are in the same direction and are aiding to produce

a stronger field. Below the conductor the lines of force are opposing and the resultant force on the conductor is directed downward.

The deflecting action and the direction in which the beam is deflected by the scanning coils in Fig. 3-10 can be seen by noting the direction of the magnetic lines of force in the deflection fields and the field associated with the beam current, remembering that the reaction of two parallel fields always exerts a force toward the weaker field. Consider the horizontal deflection coils first. The windings are in a horizontal plane above and below the beam axis. Using the left-hand rule, the thumb pointing in the direction of the field inside a coil when the fingers curve in the direction of the electron flow around the coil, the deflection field for the horizontal windings is in a vertical plane and downward for the directions shown in Fig. 3-10. When the direction of the electron beam is out of the paper, as signified by the cross in the center of the tube, its magnetic field has lines of force clockwise around the beam in a plane perpendicular to the beam axis. To the right of the beam axis, the magnetic field of the electron beam is in the same direction as the deflecting field, while the fields are opposing on the left. The electron beam, therefore, is deflected to the left, as the resultant force moves the beam toward the weaker field. In a similar manner, the vertical deflection coils deflect the electron beam upward. Deflecting current for both sets of coils can be provided simultaneously, deflecting the beam to the upper left corner of the scanned surface.

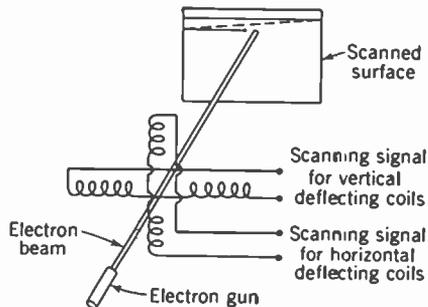


FIG. 3-12. Action of the deflection coils on the electron scanning beam.

*Deflection Current.* In order to obtain the desired deflection fields an alternating deflection current is passed through the scanning coils to produce the continuous deflecting motion required. Direct current for centering control can be obtained by means of a low-voltage tap on the d-c power supply, with provision for both vertical and horizontal positioning. In addition to the direct current, an alternating current through the horizontal deflection coils deflects the electron scanning beam to produce the horizontal scanning lines and, simultaneously, vertical deflection current moves the beam downward slowly as it traces out the horizontal lines, thus producing the desired scanning pattern. This is illustrated in Fig. 3-12.

The amount of beam deflection is directly proportional to the magnitude of deflection current and the distance to the scanned surface. How-

ever, the deflection is inversely proportional to the square root of the anode voltage in magnetic deflection. Therefore, it is easier to provide large amounts of deflection with high values of anode voltage in a magnetic tube, compared to the electrostatic method where the deflection is inversely proportional to the anode voltage itself. If in magnetic deflection the anode voltage is quadrupled, the amount of deflection current need be only twice as great for the same deflection. This is an important factor when high values of anode voltage are used, and magnetic deflection is generally employed for picture tubes with an anode voltage of about 10 kv or more.

**Deflection Yoke.** The two pairs of deflection coils are placed in one form called a deflection *yoke*, which is slipped over the narrow end of the

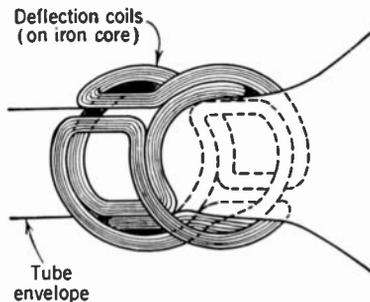


FIG. 3-13. Construction of a deflection yoke.

tube and clamped at a point beyond the electron gun just before the glass envelope flares out. A complete deflection yoke and the manner in which the yoke is mounted on a picture tube can be seen in Fig. 9-2, while Fig. 3-13 illustrates the construction of the yoke. The physical form for each of the deflecting coils is generally a flat rectangular winding, which is then curved to fit the yoke form. The entire yoke is in a housing that allows the magnetic field to penetrate the tube.

The amount of deflection current needed for full deflection depends upon the inductance of the yoke coils and the anode voltage. In a typical yoke for a picture tube using 12,000 volts on the anode, the vertical deflection coils have a total inductance of approximately 40 mh and require 300 ma peak-to-peak deflection current, while the horizontal deflection coils have a total inductance of approximately 13 mh and require about 1,000 ma peak-to-peak deflection current. The horizontal and vertical scanning coils in the yoke are not interchangeable because they have different values of inductance. Although the two sets of coils are essentially perpendicular their impedances must be different to minimize *crossstalk* interference, which is produced when the signal from one set of coils is coupled into the other.

## REVIEW QUESTIONS

1. Show the electron-gun structure for a cathode-ray tube using electrostatic focusing and magnetic deflection, labeling all the elements in the tube.
2. What is the function of the control grid in the electron gun?
3. What is meant by the crossover point in the electron gun?
4. Give one advantage of magnetic deflection over electrostatic deflection.
5. What is the effect on the amount of deflection produced when the second-anode voltage is decreased and all other factors remain the same?
6. If the distance from the point of deflection to the screen is doubled in a picture tube, with all other factors remaining the same, what will be the effect on the size of the scanning pattern?
7. What is the effect on the direction of deflection when the deflecting coil current is reversed?
8. How is the focus of the electron beam varied in magnetic focusing and electrostatic focusing?
9. How is the electron beam moved vertically and horizontally in a cathode-ray tube using magnetic deflection? Electrostatic deflection?
10. What is the function of the deflection yoke?

## CHAPTER 4

### SCANNING AND SYNCHRONIZING

The electron scanning beam is continuously deflected in a standard pattern to scan all the picture elements in the image. This must be the same at the transmitter and receiver in order to reproduce the picture elements in their correct position. The standard scanning pattern for both the camera tube and picture tube is based upon the use of saw-tooth deflecting signals for linear scanning, with synchronization of the scanning motions provided by the transmitted synchronizing signals.

**4-1. The Saw-tooth Scanning Waveform.** In linear scanning, the deflecting force provided by the deflection plates or coils must increase

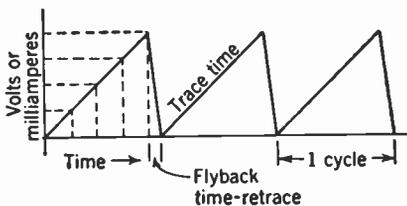


FIG. 4-1. The saw-tooth scanning waveform. During trace time, the linear rise in amplitude provides equal increments of voltage or current for equal intervals of time.

linearly with time to produce a uniform scanning motion during the trace period. When the end of the trace is reached, the deflecting force must then reverse itself and fall rapidly to its initial value to begin another cycle. Since the deflecting force is directly proportional to the magnitude of deflection voltage in the electrostatic method or directly proportional to the deflecting current in magnetic deflection, the deflection signal required to produce the desired linear scanning pattern has the saw-tooth wave shape shown in Fig. 4-1.

In electrostatic deflection a saw-tooth voltage is applied to the deflection plates to produce the desired linear sweep of the electron beam. In electromagnetic deflection the current through the deflecting coils must have the saw-tooth wave shape because deflection of the beam is produced by the electromagnetic field associated with the deflecting current.

Consider the saw-tooth wave shape of Fig. 4-1 as the scanning voltage for electrostatic deflection with a peak value of 400 volts. If the electrostatic tube requires a deflection voltage of 100 volts to produce a deflection of 1 in. on the scanned surface, then the beam will be deflected 4 in. by

400 volts. The linear rise on the saw-tooth voltage wave provides equal increments of 100 volts for the four equal intervals of time shown, deflecting the beam one additional inch for each increase of 100 volts. For horizontal scanning the uniform voltage rise thus provides a deflecting force that increases continuously during the trace time to produce a constant-velocity sweep of the electron beam from left to right across the scanned surface. At the peak of the voltage rise the saw-tooth voltage wave reverses direction and decreases rapidly to its initial value, to produce the rapid retrace or flyback. No picture information is scanned during the retrace, and the flyback time is made as short as possible to produce a rapid retrace to the left for the start of another horizontal line.

Saw-tooth deflection signal can be provided for the vertical scanning motion in a similar manner. As the electron beam is deflected horizontally the vertical saw-tooth sweep signal moves the beam downward with uniform speed to scan all the horizontal lines from top to bottom of the frame. Then the rapid retrace on the vertical saw-tooth deflection signal returns the electron scanning beam to the top of the frame for the beginning of the next vertical scanning cycle.

Both the trace period and the flyback time are included in one complete cycle of the saw-tooth wave. Since the number of horizontal lines per frame is 525 and the number of frames per second is 30, the frequency of the saw-tooth deflection signal required for horizontal scanning is  $525 \times 30$ , or 15,750, cps. The frequency of the vertical saw-tooth scanning signal is 60 cps because each frame is scanned twice from top to bottom during one frame period in an interlaced scanning pattern and the frame frequency is 30 cps. The vertical scanning motion is relatively slow, compared to the horizontal sweep, because many horizontal lines must be scanned during one cycle of vertical scanning, and even the vertical flyback time is long enough to include several horizontal lines. Although illustrated for electrostatic deflection, the saw-tooth wave shape requirements are the same for the saw-tooth scanning current in magnetic deflection.

The method of obtaining the required saw-tooth current or voltage wave shape makes use of the linear rise in current through an inductance or the linear rise in voltage across a condenser when charging voltage is applied. Various types of saw-tooth generator circuits for producing the required deflection signal are possible, but all are fundamentally the same in that provision is made for a relatively long charging time to produce the linear rise on the saw-tooth wave and a short discharge time for the flyback. Flyback or retrace time on the saw-tooth wave for horizontal deflection is approximately 10 per cent of the total line period. Practical limitations make it difficult to obtain a more rapid flyback. The lower

frequency vertical saw-tooth wave usually has a flyback time that is less than 5 per cent of the complete cycle.

**4-2. Standard Scanning Pattern.** The scanning procedure that has been universally adopted employs horizontal linear scanning in an odd-line interlaced pattern. The FCC scanning specifications for television broadcasting in the United States provide a standard scanning pattern that includes a total of 525 horizontal scanning lines in a rectangular frame having a 4:3 aspect ratio. The frames are repeated at a rate of 30 per second with two fields interlaced in each frame.

*Interlacing Procedure.* Interlaced scanning can be compared to reading a written page where the information on the page is continuous when reading all the odd lines from top to bottom and then going back to the

The horizontal scanning lines are interlaced in the odd lines are scanned, omitting the even lines. the television system in order to provide two Then the even lines are scanned to complete the views of the image for each picture frame. All whole frame without losing any picture information.

FIG. 4-2. The interlacing of lines. Read all the odd lines first from top to bottom, then the even lines.

top of the page to read all the even lines down to the bottom of the page. This is illustrated in Fig. 4-2. If a page were written and read in this interlaced pattern the same amount of information would be available as though it were written in the usual way with all the lines in progressive order. For interlaced scanning, then, all the odd lines from top to bottom of the frame are scanned first, skipping over the even lines. After this vertical scanning run and a rapid vertical retrace to bring the electron scanning beam back to the top of the frame, all the even lines which were omitted in the previous scanning run are then scanned from top to bottom. Thus each frame is divided into two fields, the first and all odd fields containing the odd lines in the frame, while the second and all even fields include the even scanning lines. With two fields per frame and 30 complete frames scanned per second the field repetition rate is 60 per second and the vertical scanning frequency is 60 cps.

*Odd-line Interlacing.* The geometry of the standard odd-line interlaced scanning pattern is illustrated in Fig. 4-3. The electron scanning beam

starts at the upper left corner of the frame at point *A* in the figure and sweeps across the frame with uniform velocity to cover all the picture elements in one horizontal line. At the end of this trace the beam is then rapidly returned to the left side of the frame, as shown by the dashed line in the illustration, to begin the next horizontal line. The horizontal lines slope downward in the direction of scanning because the vertical deflecting signal simultaneously produces a vertical scanning motion, which is very slow compared with the horizontal scanning speed. The slope of the horizontal line trace from left to right is greater than during the retrace from right to left because of the shorter time of the retrace, which does not allow the beam as much time to be deflected vertically. The beam is continuously and slowly deflected downward as it scans the horizontal

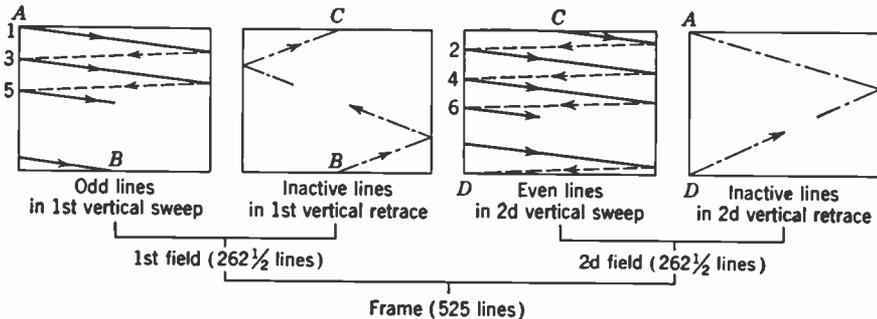


FIG. 4-3. The odd-line interlaced scanning procedure.

lines and its position is successively lower and lower as the horizontal scanning proceeds.

After the first complete horizontal line, which includes the trace and retrace, has been scanned, the electron beam is at the left side of the frame and the third line can then be scanned, omitting the second line. This is accomplished by doubling the vertical scanning frequency from the frame repetition rate of 30 to the field frequency of 60 cps. Deflecting the beam vertically at twice the speed necessary to scan 525 lines produces a complete vertical scanning period for only  $262\frac{1}{2}$  lines, with alternate lines left blank. Thus the electron beam scans all the odd lines, finally reaching a position such as *B* in the figure at the bottom of the frame.

At this time the vertical retrace begins because of the flyback on the vertical saw-tooth deflecting signal, and the beam is brought back to the top of the frame to begin the second, or even, field. The vertical flyback time is very fast compared to the trace but is slow compared to the horizontal scanning speed, and therefore some horizontal lines are produced during the vertical flyback. As shown in the figure, the beam moves from point *B* up to *C*, traversing a whole number of horizontal lines.

These lines are all inactive, though, because the beam is cut off by blanking voltage during the vertical flyback time.

Horizontal scanning during the second field begins with the beam at point *C* in the middle of a horizontal line because the first field contains 262 lines plus one-half a line, including active and inactive lines. After scanning a half line in the second field the electron beam then proceeds to scan all the even lines in the pattern lying between the odd lines and omitted during the scanning of the first field. The vertical scanning motion is exactly the same as in the previous field, giving all the horizontal lines the same slope downward in the direction of scanning, and all the even lines in the pattern are scanned down to point *D*. The vertical retrace begins at point *D* and after traversing a whole number of inactive lines the beam is brought back to *A* for the beginning of the third field.

All odd fields begin at point *A* and are the same. All even fields begin at point *C* and are the same. Since the beginning of the even-field scanning at *C* is on the same horizontal level as *A* with a separation of one-half line, and the slope of all the lines is the same, the even lines in the even fields fall exactly between the odd lines in the odd field. The essential requirement for this odd-line interlace is that the starting points at the top of the frame be separated by exactly one-half line between even and odd fields.

*A Sample Scanning Pattern.* A complete scanning pattern is shown in Fig. 4-4 with the required horizontal and vertical scanning signals to illustrate odd-line interlacing. A total of 21 lines is used for simplicity, instead of 525, and a convenient vertical retrace time is assumed to provide two inactive lines for the frame. The 21 lines are interlaced to produce two fields per frame with one-half the 21-line total, or  $10\frac{1}{2}$  lines, included in each field. Of the  $10\frac{1}{2}$  lines in each field one is inactive, occurring during the vertical retrace, which leaves  $9\frac{1}{2}$  active lines per field and 19 for the entire frame.

Starting in the upper left corner at *A* the beam scans the first line from left to right, retraces to the left for the beginning of the third line in the frame, and then scans the third and succeeding odd lines down to the bottom of the frame. The active horizontal lines slope downward to the right in the direction of scanning because of the vertical saw-tooth deflecting signal. After scanning  $9\frac{1}{2}$  active lines the beam is at point *B* at the bottom of the frame when the vertical flyback begins. One inactive horizontal line is scanned during the vertical retrace, consisting of two half lines in this illustration, sloping upward in the direction of scanning. During this vertical retrace the scanning beam is brought up to point *C*, separated from point *A* by exactly one-half line, to start scanning the second field. Because of this half-line separation between points *A* and

*C* the lines scanned in the even field fall exactly between the odd lines in the previous field. The beam then scans  $9\frac{1}{2}$  even lines from point *C* down to *D* where the vertical retrace begins for the even field. The number of inactive lines in even fields is the same as for odd fields because

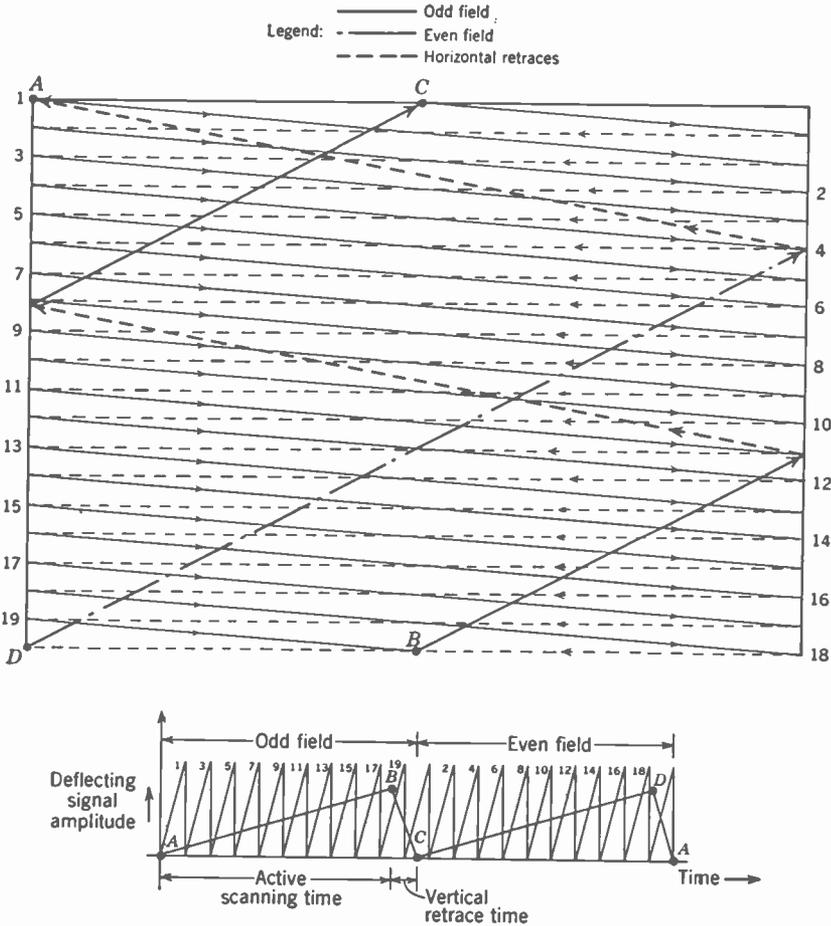


FIG. 4-4. A sample scanning pattern for 21 interlaced lines, with the required saw-tooth deflecting signals. Beginning at point *A* the scanning motion is continuous through *B, C, D*, and back to *A* again.

the flyback time on the vertical saw-tooth signal is the same. Therefore, the one inactive line in the second field can return the beam from *D* at the bottom to point *A* at the top left corner of the frame where another odd field begins.

It should be noted that the points at which vertical retrace and the active downward scan begin need not be as shown in Fig. 4-4. These

points could all be shifted by any fraction of a horizontal line without loss of interlace if the half-line difference were maintained. The half-line spacing between the starting points in alternate fields is automatically produced in the saw-tooth deflecting signals and the scanning motion because there is an odd number of lines for an even number of fields. Proper interlacing is assured, therefore, when the required frequencies of the horizontal and vertical saw-tooth scanning signals are maintained precisely and the flyback time on the vertical saw-tooth wave is constant for all fields.

With 525 lines in the scanning pattern, there must be exactly  $262\frac{1}{2}$  cycles of horizontal saw-tooth deflecting signal for a complete vertical saw-tooth wave to make a total of 525 lines for two fields or one frame,

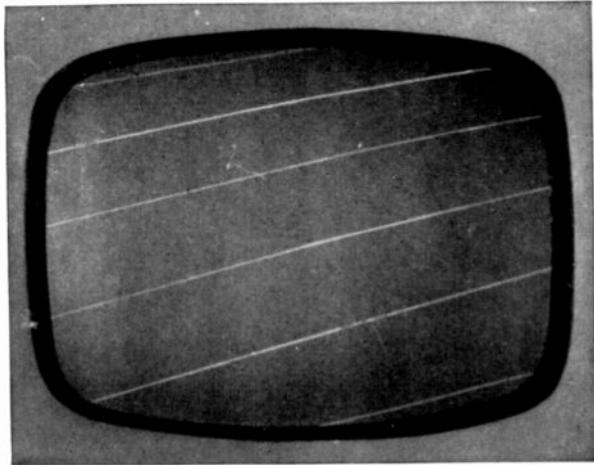


FIG. 4-5. The scanning raster on the kinescope screen.

including active and inactive lines. Since the frame repetition rate is 30 per second, the required vertical scanning frequency is 60 cps, and the horizontal frequency must be exactly  $262\frac{1}{2} \times 60$ , or  $525 \times 30$ , which is equal to 15,750 cps.

*The Scanning Raster.* As the electron beam scans the screen of the picture tube it produces a rectangular area of light that is called the scanning raster. Figure 4-5 shows the raster on the kinescope screen, without any picture information. When video signal voltage is coupled to the kinescope control grid, the picture is reproduced on the raster.

**4-3. Flicker in the Picture.** Interlaced scanning is used because the flicker effect is negligible with 60 views of the picture presented each second. Although the frame repetition rate is still 30 per second, the picture is blanked out during each vertical retrace and the change from black between pictures to the white picture occurs at the rate of 60 per second, which is too rapid to be noticeable. If progressive scanning were

used instead of interlacing, with all the lines in the frame simply scanned in progressive order from top to bottom, there would be only 30 blankouts per second and objectionable flicker would result. Scanning 60 complete frames per second in a progressive pattern would also eliminate flicker in the picture but the horizontal scanning speed would be doubled, introducing additional problems more difficult to overcome than those of interlaced scanning.

Although the increased blanking rate with interlaced scanning largely eliminates the effect of flicker in the image as a whole, the fact that individual lines are interlaced can cause flicker in small areas of the picture. Any one line in the image is illuminated 30 times per second, reducing the flicker rate of a single line to one-half the flicker rate for the interlaced image as a whole. The lower flicker rate for individual lines may cause two effects in the picture called *interline flicker* and *line crawl*. The interline flicker is sometimes evident as a blinking of thin horizontal objects in the picture, such as the roof line of a house. Line crawl is an apparent movement of the scanning lines upward or downward through the picture, due to the successive illumination of adjacent lines. These effects are more noticeable in bright parts of the picture because the eye perceives flicker more easily at high brightness levels. Usually, the flicker in small areas of the picture is not evident except on close inspection, but with particular types of picture content the flicker may be noticeable at normal viewing distances.

*Frame and Field Frequencies.* The frame and field repetition rates are chosen at exactly 30 and 60 because 60 cps is the normal base frequency for most a-c power in the United States. Ripple voltages or currents from the power line always enter the television circuits through the B supply and by stray electric and magnetic fields. The ripple voltage may affect picture reproduction in terms of horizontal or vertical deflection, and the brightness produced by the electron scanning beam. When the ripple variations are synchronous with the scanning frequencies the effect of ripple at the 60-cps fundamental of the power-line frequency or at the second harmonic at 120 cps is much less noticeable in the picture than with some other value, such as 24 and 48 for the frame and field repetition rates. The reason is that the effect of the 60- or 120-cps interference in the reproduced picture stays still when the ripple frequency is synchronous with the field-scanning frequency. Therefore, the standard frame and field frequencies are 30 and 60 per second in order to eliminate flicker, to allow a smooth reproduction of any motion in the televised scene, and to minimize the effect of interfering power-line ripple on the picture.

**4-4. Scanning Distortions.** In the formation, focusing, and deflection of the electron beam, the following distortions can be introduced in the scanning pattern.

*Defocusing at the Edges of the Scanning Pattern.* When the electron beam is deflected by either electrostatic or magnetic means the deflection of the beam may interfere with the focusing action of the gun, producing poor focus at the edges of the scanning pattern and the reproduced picture. Generally, the greater the deflection of the beam from the center axis, the more defocusing at the edges. This defocusing may be caused by several factors. If the distance from the point of deflection to the scanned surface is not always the same as the beam scans different parts of the frame, the beam will have to travel varying distances from the electron lens and the sharpness of focus will vary over the scanning pattern. Another reason for defocusing is slight variations in the speed of the electrons in the beam, because of variations in velocity as they leave the cathode. As a result, there may be small differences in the amount of deflection for the electrons in the beam and the scanning spot spreads out into an ellipse, instead of being a small, fine circle. Finally, it is essential that the deflection field be uniform and symmetrical about the beam axis. In the magnetic-deflection system the coil arrangement is inherently balanced for providing a uniform deflection field. With electrostatic deflection, however, the problem of obtaining a uniform deflection field is complicated by the fact that the deflection plates must be connected directly or indirectly to the second anode so that they can be at approximately the same potential. It is necessary to use a balanced push-pull circuit arrangement for the deflection plates in order to produce a suitable scanning pattern. In addition to poor focus, the scanning pattern will show trapezoidal or keystone distortion if unbalanced electrostatic deflection is used instead of the push-pull circuit.

*Keystone or Trapezoidal Distortion.* A trapezoidal distortion of the scanning pattern that gives it the appearance of a keystone is generally called the *keystone effect*. This is illustrated in Fig. 4-6. As an example of this type of distortion, keystoneing occurs in the iconoscope type of camera tube because the electron gun is inclined at an angle of about  $30^\circ$  with respect to the flat mosaic plate. If the horizontal deflecting signal amplitude is constant for all lines in the pattern the distance scanned at the top will be greater than at the bottom because of the greater distance from the electron gun to the image plate. The lower part of the mosaic plate is nearer to the gun, and the horizontal lines at the bottom will not be wide enough, producing the trapezoidal distortion shown instead of the required rectangular scanning pattern. This trapezoidal distortion, or keystoneing, can be corrected by using a greater amplitude of horizontal deflection signal at the base of the pattern than at the top, thus compensating electrically for the geometrical distortion. Besides this keystone effect in the iconoscope camera tube, trapezoidal distortion can also be introduced by unbalanced deflection fields for the vertical and hori-

zontal scanning, causing keystone at the top and bottom or at the sides of the scanning pattern.

*Curved Edges on the Scanning Pattern.* Two other ways in which the rectangular shape of the scanning pattern can be distorted are shown in Fig. 4-7. In *a* the reduced width of the scanning lines toward the bottom of the pattern is produced because the scanning voltage or current does not have enough amplitude to deflect the beam the full width of the scan-

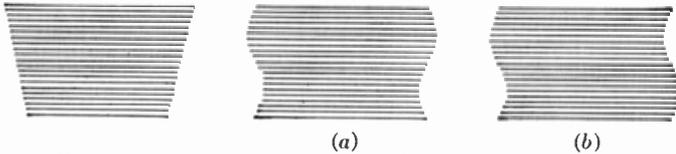


FIG. 4-6.

FIG. 4-7

FIG. 4-6. Keystone or trapezoidal distortion of the scanning pattern.

FIG. 4-7. Curved edges on the scanning pattern. (a) Sine-wave variations in horizontal scanning amplitude; (b) sine-wave displacement of the horizontal lines.

ning pattern. At the top, the scanning voltage or current has a greater amplitude and produces excessive horizontal deflection. Such nonuniform width of the horizontal scanning lines results from varying amplitudes in the horizontal scanning signal. In *b* the width of each horizontal line is the same but a superimposed ripple on the horizontal scanning voltage or current produces a regular displacement of the horizontal lines. These effects are caused by 60-cps hum voltage in the horizontal deflection circuits.

*Nonlinear Scanning.* In the linear method of scanning employed at the transmitter and receiver, the velocity of the scanning spot must be uniform as it moves horizontally to form each line as it is deflected verti-

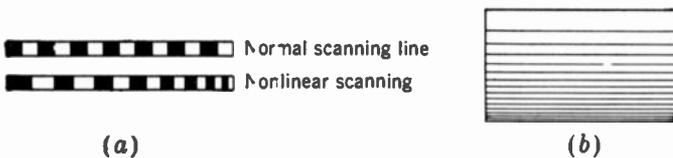


FIG. 4-8. Effects of nonlinear scanning. (a) Crowding and spreading of picture elements in a horizontal line, caused by nonlinear horizontal scanning; (b) crowding and spreading of the horizontal lines caused by nonlinear vertical scanning.

cally to scan all the lines. If the receiver scanning spot moves too slowly along a horizontal line, compared with the transmitter scanning speed, the picture elements in the received image will be crowded together. Conversely, if the receiver scanning motion is too fast the picture elements will be spread out. Usually, the effect is such that spreading of the picture elements occurs at one side of the scanning pattern while the opposite side is crowded. This is illustrated in Fig. 4-8a for a horizontal line with

picture elements spread out at the left and crowded at the right. When the same effect occurs for all the horizontal lines in the pattern, the entire picture is distorted; picture information is spread out at the left and crowded together at the right side of the picture. The vertical scanning motion must also be uniform or the horizontal lines will be bunched at the top or bottom of the pattern and spread apart at the opposite side. This is illustrated in Fig. 4-8*b* for spreading at the top and crowding at the bottom.

*Poor Interlacing.* The vertical scanning trace for each field must start exactly half a line from the beginning of the previous field in the odd-line

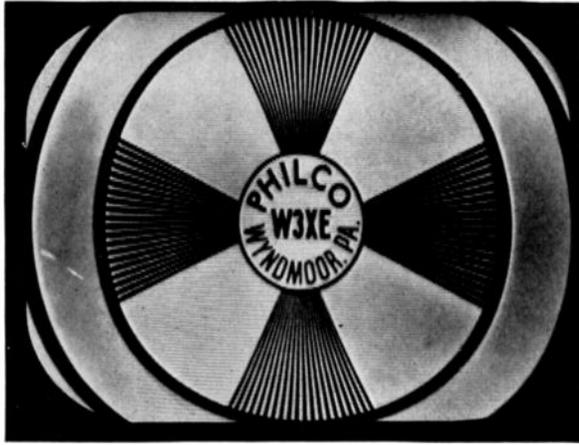


FIG. 4-9. Faulty interlacing. Note how the line divisions in the horizontal wedges appear to be interwoven, in a moiré effect. (Philco Corporation.)

interlaced scanning pattern employed. When the downward motion is displaced from the correct position by a small fraction of the line-scanning period, the spot starts scanning too close to one of the lines in the previous field instead of scanning exactly between lines. This produces a vertical displacement between the odd and even lines that is carried through the entire frame. The result is that the lines in odd and even fields overlap and to some extent leave blank spaces between lines, instead of sharing the scanned area equally. This defect in the interlaced scanning is called *pairing* of the lines. The faulty interlacing reduces the amount of picture details in the vertical direction, since there is a smaller number of effective scanning lines. In an extreme case the lines in each successive field may fall exactly on the lines in the previous field and the effective scanning pattern contains only one-half the usual number of horizontal lines. When there is a group of horizontal lines that slant downward in the image, the lines are interwoven in the moiré effect shown in Fig. 4-9 because the even and odd fields are not exactly interlaced. The inter-

woven, or moire, effect in diagonal lines is also called *fishtailing*. This is a typical effect that can be used to check for faulty interlacing. Another way of checking the interlace is to observe the spacing between horizontal lines scanned during vertical retrace time. If the retrace lines are paired, instead of being evenly distributed from top to bottom of the raster, then the interlacing of even and odd scanning lines must be poor. The lines scanned during vertical trace time are also paired when interlace is faulty, so that more dark space is produced between successive pairs of scanning lines. Poor interlacing is caused by inaccurate vertical synchronization.

**4-5. The Synchronizing Pulses.** The need for exact synchronism in the scanning at the transmitter and receiver requires synchronizing signals to keep the receiver locked in with the transmitter scanning, and to maintain

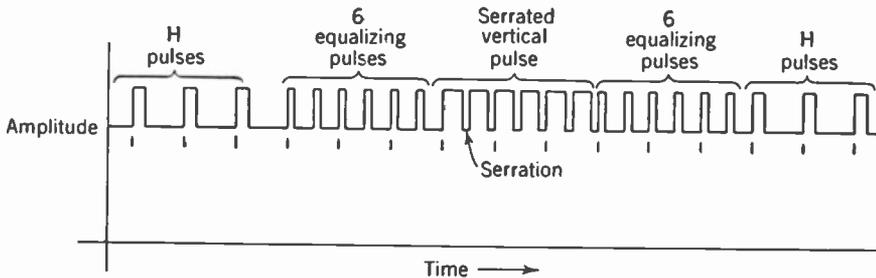


FIG. 4-10. The synchronizing pulses.

correct timing of the vertical and horizontal scanning motions. A horizontal synchronizing pulse is transmitted for each horizontal line to keep the horizontal scanning synchronized, and a vertical synchronizing pulse is transmitted for each field to synchronize the vertical scanning motion. Therefore, the horizontal synchronizing pulses have a frequency of 15,750 cps, and the frequency of the vertical synchronizing pulses is 60 cps. The synchronizing pulses are transmitted as part of the picture signal but are sent during the blanking period when no picture information is transmitted. This is possible because the synchronizing pulse begins the retrace, either horizontal or vertical, and consequently occurs during retrace time. The synchronizing signals are combined with the picture signal in such a way that part of the modulated picture signal amplitude is devoted to the synchronizing pulses and the remainder to the camera signal. The term *sync* is often used for brevity to indicate the synchronizing pulses.

The form of the synchronizing pulses is illustrated in Fig. 4-10. Note that all pulses have the same amplitude but differ in pulse width or waveform. The synchronizing pulses shown include from left to right three horizontal pulses, a group of six equalizing pulses, a serrated vertical

pulse, and six additional equalizing pulses which are followed by three more horizontal pulses. There are many additional horizontal pulses after the last one shown, following each other at the horizontal line frequency until the equalizing pulses occur again for the beginning of the next field. For every field there must be one prolonged vertical pulse, which is actually composed of six individual pulses separated by the five serrations. The horizontal pulses are repeated at the horizontal line intervals and synchronize each horizontal line, while the vertical pulses occur at the field frequency of 60 cps to synchronize each field.

Each vertical synchronizing pulse extends over a period equal to six half lines or three complete horizontal lines, making it much wider than a horizontal pulse. This is done to give the vertical pulses an entirely different form from the horizontal pulses so that they can be completely separated from each other, one furnishing horizontal synchronizing signals alone while the other provides only vertical synchronization. The five serrations are inserted in the vertical pulse at half-line intervals in order to divide the prolonged vertical pulse into six smaller intervals, each of which can serve for horizontal synchronization, thus preserving the continuity of horizontal synchronization through the vertical synchronizing time. The reason for doing this is that horizontal synchronization must be maintained at all times, including the vertical retrace time when no picture information is scanned. If the horizontal synchronizing action is lost during the vertical retrace the interlacing will suffer, and synchronization may not be reestablished in time for the start of the active horizontal scanning.

The equalizing pulses are also spaced at half-line intervals so that every other one can serve for horizontal synchronization, alternate pulses being used for even and odd fields. The reason for using equalizing pulses, however, is related to vertical synchronization. They are inserted to equalize the difference in the vertical synchronizing signals for alternate fields. Their effect is to provide identical wave shapes in the separated vertical synchronizing signal for even and odd fields so that constant timing can be obtained for good interlace.

Since the equalizing pulses are repeated at half-line intervals their repetition rate is twice 15,750, or 31,500 cps. Thus, the horizontal pulse frequency is one-half the equalizing pulse rate and the vertical pulse frequency is  $\frac{1}{525}$  of the equalizing pulse frequency. These are exact submultiples of the equalizing pulse frequency and can be obtained, therefore, by frequency division of the equalizing pulses. In this way, all the synchronizing pulses are derived from a common source at the transmitter and their frequencies are automatically interlocked in the correct ratios.

It should be noted that the synchronizing signals do not produce scanning. Saw-tooth generators are required to produce the scanning motion,

but the synchronizing signals are needed for exact timing of the horizontal and vertical scanning motions at the transmitter and receiver. Without sync the saw-tooth generators can operate to produce the scanning raster, but the picture information reproduced on the raster does not hold still. Specifically, the synchronizing pulses force the saw-tooth generators to begin the flyback on the saw-tooth scanning wave, timing each cycle of the saw-tooth deflecting signal at the synchronizing frequency.

#### REVIEW QUESTIONS

1. Explain briefly why a saw-tooth deflecting signal is the wave shape required for linear scanning.
2. Draw the interlaced scanning pattern for a total of 25 lines, showing the corresponding saw-tooth scanning signals. Assume one inactive line per field.
3. Give one advantage and one disadvantage of interlaced scanning compared with progressive scanning.
4. If progressive scanning were used with a picture repetition rate of 60 per second what would be the frequency of the vertical saw-tooth deflection signal? What is the disadvantage of this scanning procedure?
5. How much time in microseconds is required for scanning one complete horizontal line, including the trace and retrace? How long is the trace time, assuming 10 per cent for the retrace time?
6. In a 525-line interlaced pattern, how many inactive lines are produced during the vertical retrace for each field when the vertical flyback time is 3 per cent of a complete cycle? For both even and odd fields?
7. Why are the active horizontal lines closer together than the inactive lines produced during vertical retrace?
8. What is meant by keystoneing?
9. Describe the effect on picture reproduction when the saw-tooth scanning signal for horizontal deflection in the picture tube is bowed out, rising rapidly and then flattening out at the top.
10. What is the function of the horizontal and vertical synchronizing pulses? What are their frequencies? What is the frequency of the equalizing pulses?
11. What is the scanning raster?
12. Describe briefly one way to check the interlacing of the scanning lines.
13. Where is the electron scanning beam on the kinescope screen at the start of the linear rise on the saw-tooth wave for horizontal scanning? At the start of horizontal retrace? At the start of the linear rise on the saw-tooth wave for vertical scanning? At the start of vertical retrace?
14. What is interline flicker?

## CHAPTER 5

### THE COMPOSITE VIDEO SIGNAL

The composite video signal contains all the information needed to reproduce the picture. This includes (1) camera signal corresponding to the desired picture information, (2) synchronizing pulses to synchronize the transmitter and receiver scanning, and (3) blanking pulses to obliterate the retraces produced in scanning and to ensure that there is no camera signal to interfere with the synchronization. How these three components are added to produce the composite video signal is illustrated

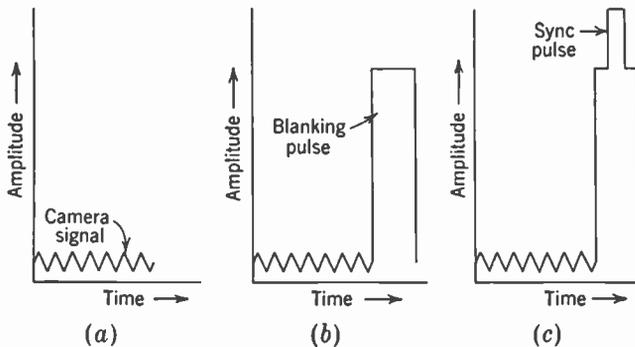


FIG. 5-1. Components of the composite video signal. (a) Camera signal; (b) camera signal plus horizontal blanking pulse; (c) composite video signal with camera signal, blanking pulse, and sync pulse.

in Fig. 5-1. The camera signal in *a* is combined with the blanking pulse in *b* and then the sync pulse is superimposed on the pedestal atop the blanking pulse to produce the composite video signal in *c*. The result is composite video signal for one horizontal scanning line.

**5-1. Construction of the Composite Video Signal.** In the composite video signal in Fig. 5-2, successive values of voltage or current amplitude are shown for corresponding values of time during the scanning of three horizontal lines in the image, together with the inactive periods, which include the synchronizing and blanking signals. The amplitude of the video signal is divided into two sections, the lower 75 per cent being

devoted to the active camera signal while the upper 25 per cent is used for the synchronizing pulses. Standardization is necessary so that the signal will be suitable for all television receivers, and the form shown is the standard construction of the composite video signal. The lowest amplitudes correspond to the whitest parts of the picture while the darker parts of the picture have higher amplitudes in Fig. 5-2 because this is the way the signal is transmitted, using a standard negative polarity of transmission. Negative transmission means that white parts of the picture are represented by low amplitudes in the transmitted picture carrier signal. Higher amplitudes correspond to progressively darker picture information until the black level is reached, which is represented by the fixed level of 75 per cent of maximum signal amplitude.

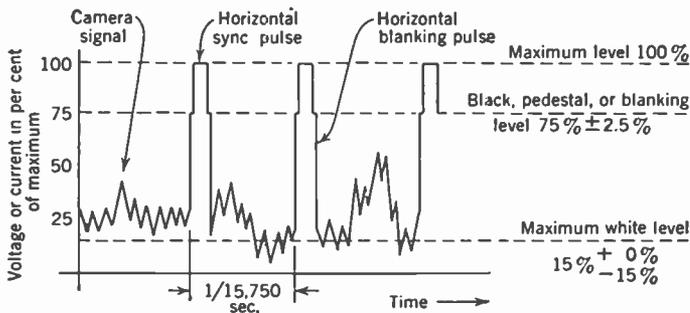


FIG. 5-2. Composite video signal for three consecutive horizontal lines.

**Black Reference Level.** The black level is constant at 75 per cent amplitude and independent of picture information, in order to maintain a brightness reference in the television system. When the image is reproduced, the 75 per cent level of the video signal corresponds to the grid cutoff voltage of the picture tube and the absence of light, thus establishing a black level. The brightness values of various shades of white and gray are then defined in terms of their amplitude relative to the black level. The 75 per cent amplitude is also the pedestal level, or blanking level, because this represents the tops of the blanking pulses, providing pedestals on which the synchronizing pulses are placed. Blanking is accomplished by making the blanking level black.

Any signal amplitude greater than the black level is called *blacker than black*, or *infrablack*, because this voltage drives the picture-tube grid voltage more negative than cutoff. The synchronizing pulses, which initiate the flyback in the scanning circuits, are blacker than black.

**The Composite Video Signal and Scanning.** Referring again to Fig. 5-2, consider the amplitude variations shown as the desired video signal obtained in scanning three horizontal lines at the top of the image. Starting at the extreme left in the figure at zero time, the signal is at a

white level and the scanning beam is at the left side of the image. As the first line is scanned from left to right, camera signal variations are obtained with various amplitudes that correspond to the required picture information. After this active trace produces the desired camera signal for one line, the scanning beam is at the right side of the image. The blanking pulse is then inserted to bring the video signal amplitude up to the black level so that the retrace can be blanked out. After a blanking interval long enough to include the retrace time, the blanking voltage is removed, since the scanning beam is then at the left side of the image ready for the active scanning of the next horizontal line. Each active horizontal line is scanned successively in this way.

*The Blanking Pulses.* The composite video signal contains blanking pulses to obliterate the retrace lines by raising the signal amplitude to the

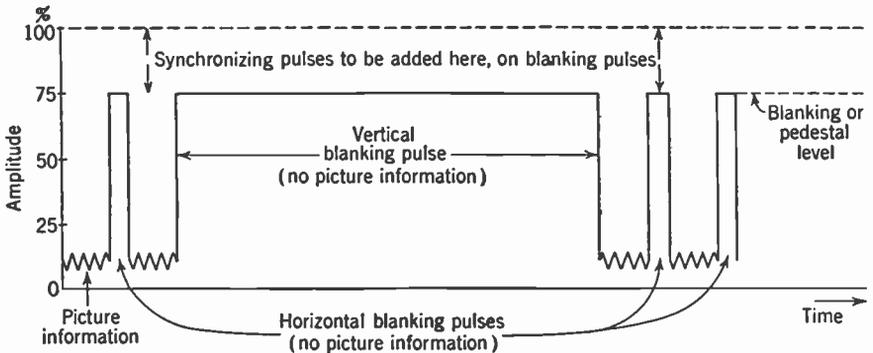


FIG. 5-3. Horizontal blanking pulses and vertical blanking pulse in the video signal. Sync pulses not shown.

black level during the time the scanning circuits produce the retraces. Retrace normally is produced during blanking time because the synchronizing pulses, which start the retrace, occur within the blanking period.

As illustrated in Fig. 5-3, there are horizontal and vertical blanking pulses in the composite video signal. The horizontal blanking pulses are included to blank out the retrace from right to left in each horizontal scanning line. The repetition rate of the horizontal blanking pulses, therefore, is the line-scanning frequency of 15,750 cps. The vertical blanking pulses have the function of blanking out the scanning lines produced when the electron beam retraces vertically from bottom to top in each field. Therefore, the frequency of the vertical blanking pulses is 60 cps.

*Horizontal Blanking Time.* Details in the horizontal blanking period are illustrated in Fig. 5-4. The interval between successive scanning lines is indicated by  $H$ . This time for one complete horizontal scanning

line is  $1/15,750$  sec, or  $63.5 \mu\text{sec}$ , since the horizontal line-scanning frequency is  $15,750$  cps. The horizontal blanking pulse has a width of  $0.14H$  to  $0.18H$ , so that the blanking time is about 16 per cent of the line period. Superimposed on the pedestals provided by the tops of the blanking pulses at the black level are the narrower sync pulses. As noted in Fig. 5-4, each horizontal sync pulse is  $0.08H$ , occupying about one-half the width of the horizontal blanking pulse. For the remaining 8 per cent of  $H$ , during the blanking time of  $0.16H$ , the signal is at the pedestal level. The part of the pedestal just before the sync pulse is called the *front porch*, as indicated in Fig. 5-4, and the *back porch* follows the sync pulse. The front porch is  $0.02H$  and the back porch  $0.06H$ , approximately.

In summary, then, 84 per cent of the horizontal line-scanning period is used for active scanning of the image, taking  $0.16H$  as a typical value of blanking time. The remaining 16 per cent is devoted to horizontal blanking, with 8 per cent utilized for the horizontal synchronizing pulses. Synchronizing does not begin until a short time after the leading edge of the blanking pulse, so that the picture is blanked out for  $0.02H$  before retrace starts and the blanking continues for  $0.06H$  after completion of the synchronizing pulse. It should be noted that the time required for the

retrace of the scanning beam depends on the scanning circuits and is not limited to the width of the synchronizing pulse. The retrace must be accomplished within the blanking time, however.

The amount of time devoted to horizontal blanking and retrace may seem relatively long, since it leaves only 84 per cent of the horizontal line period for the active scanning of picture information, but these values are necessary. It is difficult to produce scanning circuits that can supply the saw-tooth voltage or current scanning waves required for horizontal deflection with a flyback time very much less than the horizontal blanking time used. The long blanking time allows a margin of safety, therefore, ensuring that the horizontal retrace is blanked out.

In addition to the blanked-out retrace, a small portion of the forward scanning motion usually is blanked out at the beginning and end of the

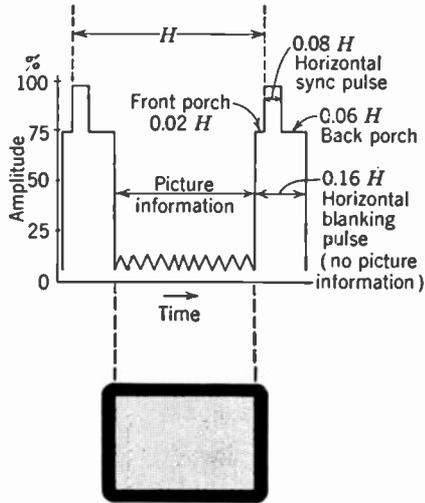


FIG. 5-4. Details of the horizontal blanking and sync pulses.  $H$  equals  $1/15,750$  sec.

active trace, as illustrated by the black bars at the left and right sides of the raster in Fig. 5-4. The black bar on the right side of the image corresponds to the front porch on the horizontal blanking pulse, before retrace starts. When the retrace is completed in less than  $0.14H$ , the trace on the left side begins during the blanking time, producing the black bar at the left side of the image corresponding to part of the back porch on the horizontal blanking pulse. The blanking bars at the sides have no effect on the picture other than decreasing the effective width slightly. Compensation is easily made, increasing the amplitude of the horizontal scanning voltage or current. By allowing some tolerance in the amount of time required for the horizontal retrace, much more reliable performance can be obtained in the scanning circuits.

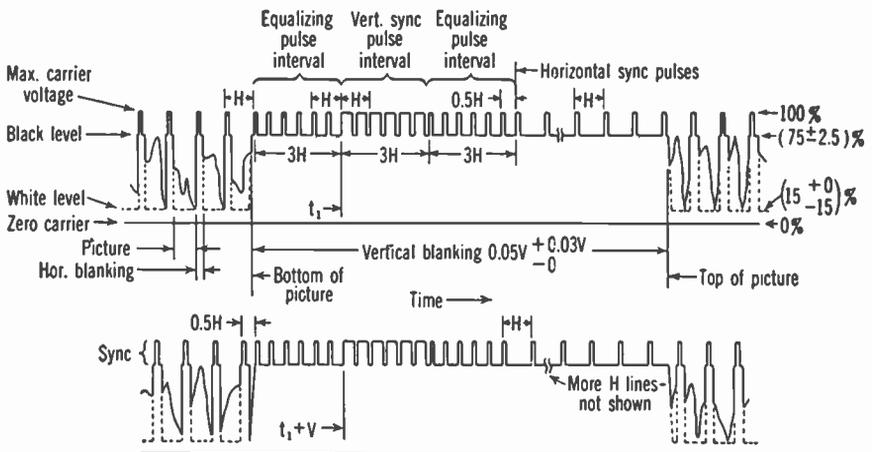


FIG. 5-5. Sync and blanking pulses for successive fields.  $V$  equals  $\frac{1}{60}$  sec.

**Vertical Blanking Time.** The vertical blanking pulses raise the video signal amplitude to the black level for the pulse duration, so that the scanning beam can be blanked out while the vertical retraces are completed. Figure 5-5 shows the composite video signal with vertical blanking pulses, during the intervals between the end of one field and the beginning of the next. The signals illustrated are identical except for the half-line displacement between successive fields necessary in odd-line interlacing. Starting at the left in the figure, the last four lines at the bottom of the image that are active in scanning information are shown with the required horizontal blanking and synchronizing pulses. Immediately following the last active line the video signal is brought up to the black level by the vertical blanking pulse in preparation for the vertical retrace. The vertical blanking period begins with six equalizing pulses spaced at half-line intervals. The serrated vertical synchronizing

pulse is next, followed by six additional equalizing pulses and a train of horizontal synchronizing pulses. No picture information is scanned during this time and the video signal is maintained at the black level so that the vertical retrace can be blanked out. The vertical blanking interval for each field equals 5 to 8 per cent of the field-scanning period of  $\frac{1}{60}$  sec and, therefore, includes 13 to 21 inactive horizontal lines, which are blanked out. After the vertical blanking pulse, active scanning is resumed and the composite video signal, including the camera signal, blanking, and synchronizing pulses for each active horizontal line, continues up to the next field.

The field retrace does not begin until about the middle of the serrated vertical synchronizing pulse, when the vertical scanning circuit is forced to produce the flyback on the saw-tooth scanning wave. About four lines are blanked out at the bottom of the picture, therefore, before the vertical retrace starts. Most of this time is devoted to the equalizing pulses, which are needed to provide exact timing of the vertical retrace motion in successive fields. The vertical flyback time depends on the scanning circuit used but must be completed within the blanking time so that the electron scanning beam will be started on the active trace down when the blanking pulse is removed for the scanning of picture information. With a vertical retrace time equal to the period of three horizontal lines, which is a typical example, some of the lines scanned at the top of the picture in the downward scan are blanked out. Thus, if the vertical flyback is assumed to start at the second serration in the vertical synchronizing pulse and if the retrace time is equal to the period of three horizontal lines, four horizontal lines at the bottom of the picture will be blanked out, three blanked-out horizontal lines will occur during the vertical flyback from bottom to top, and 6 to 14 lines at the top of the picture will not be visible. The same number of lines will be blanked out in the next successive field, with a half-line displacement from the previous field. This wide tolerance in the vertical blanking period is necessary in order to ensure that enough time is available to complete the vertical flyback and obtain the required linearity of the active trace before the scanning of picture information is resumed. The black bars at the top and bottom of the raster in Fig. 5-4 correspond to the scanning lines blanked out by the vertical blanking pulse. The only effect of the obliterated lines at the top and bottom of the frame is a slight reduction in the effective height of the picture; this is easily remedied by increasing the amplitude of the vertical scanning voltage or current.

### 5-2. Correlation between Picture Content and the Video Signal.

Two examples are shown in Fig. 5-6 to illustrate how the composite video signal is related to the picture information in the optical image. In *a* is shown the video signal that corresponds to one active horizontal line in

the scanning of an image consisting of a black vertical line down the center of a white frame. Starting at the left in the figure, the camera signal obtained in active scanning of the image is initially at the white level corresponding to the white background. The scanning beam continues its forward motion across the white background of the frame and the signal continues at the same white level until the middle of the picture is reached. When the black bar is scanned the video signal immediately rises to the black level and remains there while the entire width of the black bar is scanned. After the black vertical bar is scanned the signal amplitude drops to the white level corresponding to the white background and continues at that level while the forward scanning motion is completed to the right side of the image.

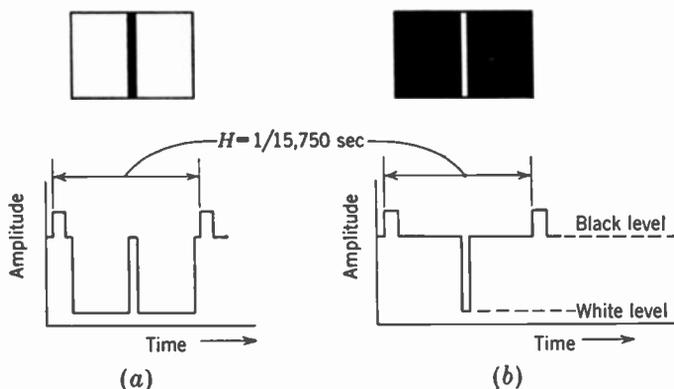


FIG. 5-6. Composite video signal obtained in scanning one line of a pattern containing a vertical line down the center of the frame. (a) Black line on white background; (b) white line on black background.

At the end of the active trace the horizontal blanking pulse is inserted to bring the video signal amplitude up to the black level in preparation for the horizontal retrace. After completion of the retrace the forward scanning motion begins again to scan the next horizontal line. Each successive horizontal line in the even and odd fields is scanned in this way, and the corresponding composite video signal for the entire picture contains a succession of signals with a waveform identical with that shown in Fig. 5-6a for each active horizontal scanning line. For the case illustrated in *b* the action is the same but the camera signal corresponds to a white vertical bar down the center of a black frame.

These are simple types of images, but the correlation can be carried over to an image with any distribution of light and shade. Thus, if the pattern contains five vertical black bars against a white background, the composite video signal for each horizontal line will include five rapid variations in amplitude from white to the black level. If the pattern consists

of a horizontal black bar across the center of a white frame, most of the horizontal lines will contain white picture information for the entire trace period, with the camera signal amplitude remaining at the white level except for the blanking intervals, while those horizontal lines that scan across the black bar will produce camera signal that remains at the black level for the complete active scanning time. A typical image consists of picture elements having various degrees of light and shade with a non-uniform distribution. In this case, then, the corresponding video signal contains a succession of signals for each horizontal line, with variations in camera signal amplitude for each picture element in a line, and waveforms for each horizontal line that vary for different lines in the frame.

**5-3. Video Frequencies and Picture Information.** Referring to the checkerboard pattern in Fig. 5-7, the square-wave signal shown represents the camera signal variations of the composite video signal obtained in scanning one horizontal line at the top of the image. It is desired to find the frequency of this square wave, because the frequency of the camera signal variations in the video signal is an extremely important factor in determining whether or not the television system can transmit and reproduce the corresponding picture information. The highest video frequency that can be transmitted is approximately 4 Mc because of the restriction of a 6-Mc transmission channel.

In determining the frequency of any signal variation, the time for one complete cycle must be known, where a cycle includes the time from one point on the signal waveform to the next succeeding point with the same magnitude and direction. The frequency can then be found as the reciprocal of the period for one cycle. Thus, the period of one horizontal scanning line is  $1/15,750$  sec and the line-scanning frequency is 15,750 cps. The camera signal variations within one horizontal line, however, necessarily have a shorter period and a higher frequency. In order to determine the frequency of the signal variations for the pattern shown in Fig. 5-7 it must first be noted that one complete cycle of camera signal includes the information in two adjacent picture elements, one white and the other black, because it is only after scanning the second square that the camera signal has the same magnitude and direction as at the start of the first square. Therefore, to find the frequency of the camera signal variations in Fig. 5-7 it is necessary to determine how long it takes to scan across two adjacent squares, which is the period for one cycle of the resultant camera signal.

The period of one complete cycle of the square-wave camera signal variations in Fig. 5-7 can be calculated as follows: The horizontal line period is  $1/15,750$  sec, or  $63.5 \mu\text{sec}$ , including the trace and retrace. With a horizontal blanking time equal to 16 per cent of the line interval, or  $10.2 \mu\text{sec}$ , this must be subtracted from the total line period to obtain

the trace time corresponding to the active scanning of picture information that is visible in the image. Therefore, it takes  $53.3 \mu\text{sec}$  for the active trace across one horizontal line. Scanning the horizontal squares in Fig. 5-7, two of the twelve squares are scanned in a time equal to  $\frac{1}{6}$  of the total active time, or  $53.3/6 \mu\text{sec}$ . This is  $8.9 \mu\text{sec}$  and is the period of one complete cycle of the square-wave signal. The frequency of the camera signal, therefore, is  $1/8.9 \mu\text{sec}$ , or  $0.11 \times 10^6 \text{ cps}$ , which is  $0.11 \text{ Mc per sec}$ . This is well within the capabilities of the television system, since the upper frequency limit for video signal frequencies that can be transmitted is approximately  $4 \text{ Mc}$ .

When a typical image with scattered areas of light and shade is scanned, the camera signal variations do not have the symmetry of the square-wave

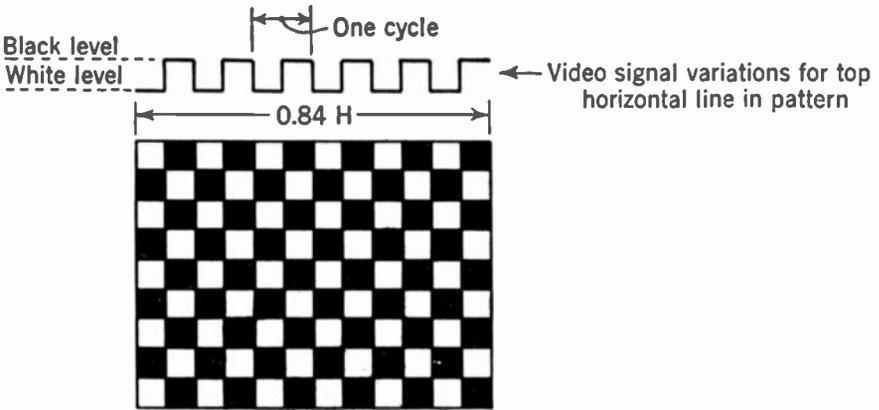


Fig. 5-7. Checkerboard pattern and corresponding camera signal variations.

signal shown in Fig. 5-7. However, the differences of light and shade in the image correspond to changes in the camera signal amplitude in the same manner, and the frequency of the resultant camera signal variations depends on the time required to scan adjacent areas that differ in light values. Thus, when large objects with a constant white or black level are scanned the resultant camera signal variations have a low frequency because of the relatively long time between changes in level. Small areas of light and shade in the image correspond to higher video frequencies. The highest signal frequencies correspond to the variations between small adjacent picture elements in a horizontal line. Therefore, the highest video frequencies correspond to the horizontal detail of the picture, and the ability of the television system to transmit and reproduce the high video frequencies determines the extent to which the fine horizontal detail in the image can be reproduced.

**5-4. Maximum Number of Picture Elements.** Referring again to the checkerboard pattern in Fig. 5-7, the maximum possible number of pic-

ture elements in the image can be determined where each square is considered as one element. The problem is to find out how many picture elements or squares can be reproduced in the vertical direction and along a horizontal line in the image, with 4 Mc as the highest video frequency that can be transmitted.

*Horizontal Detail.* The number of picture elements contained in each horizontal line is limited only by the ability of the television system to generate, transmit, and reproduce rapid changes in the camera signal amplitude. Since 4 Mc is the highest video signal frequency that can be transmitted in the television channel, this represents the best horizontal detail that can be reproduced at the receiver. Proceeding in the same manner as in the previous section, the number of elements corresponding to 4 Mc can be determined to show the maximum number of picture elements in a horizontal line and the size of the smallest possible horizontal detail. The period of one complete cycle for a 4-Mc signal variation is  $1/(4 \times 10^6)$  sec, or  $0.25 \mu\text{sec}$ . This is the time required to scan two adjacent picture elements. With two elements scanned in  $0.25 \mu\text{sec}$ , then eight elements are scanned in  $1 \mu\text{sec}$ , and  $8 \times 53.3$ , or 426, picture elements can be scanned during the entire active line period of  $53.3 \mu\text{sec}$ . Thus, 426 picture elements can be scanned along a horizontal line without exceeding the 4-Mc frequency limitations. If there were 426 squares in the horizontal direction in the checkerboard pattern in Fig. 5-7, therefore, the resultant camera signal variations would produce a 4-Mc signal, which represents the maximum capabilities of the present system in terms of horizontal detail.

In order to reproduce the squares of the checkerboard pattern as individual, discrete elements, a square-wave signal is needed. Since response up to about the fifteenth harmonic of the fundamental frequency is required to reproduce a square wave, it would be necessary to utilize a 60-Mc sine-wave signal, which is beyond the capabilities of the television system. As a result, the maximum number of 426 horizontal details in a television picture can be reproduced only as the continuous variations in shading for a 4-Mc sine-wave video signal instead of the individual, discrete elements corresponding to the 4-Mc square-wave signal.

*Utilization Ratio and Vertical Detail.* The maximum number of vertical elements in the image that can be reproduced is directly dependent on the number of scanning lines, since each scanning line can represent only one detail in the vertical direction. However, a scanning line may represent no vertical detail at all. The two opposite cases are illustrated in Fig. 5-8 where the image to be scanned is a vertical bar containing a number of alternate black and white squares. The height of each square is considered to be equal to the width of a scanning line. When a square in the image has a position such that the scanning beam passes directly over

it, as in *a*, the corresponding camera signal represents the vertical detail perfectly. This is the best possible case, and the reproduced pattern corresponds exactly to the original image. For the case illustrated in *b*, however, the details in the image are so placed that the scanning beam passes over the boundary between a black and a white square and the camera signal variation corresponds to a gray level intermediate between the black and white details, representing the average brightness of the two elements. When the scanning beam covers two picture elements in this way the vertical detail is entirely lost and the reproduced image becomes

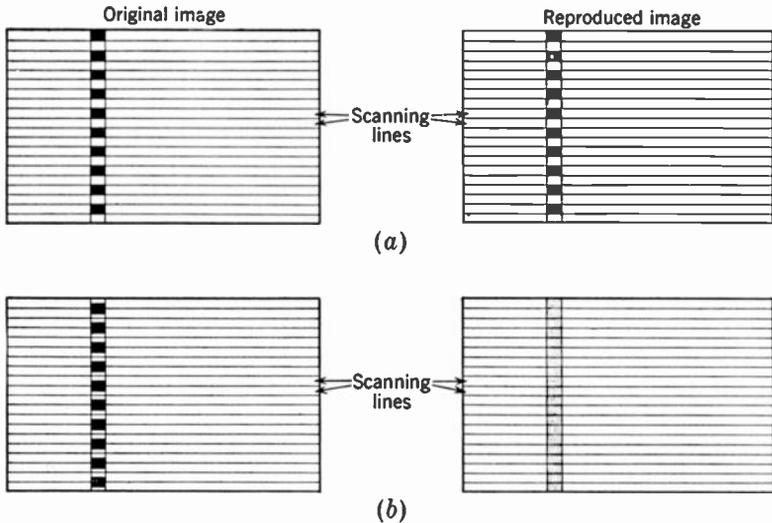


FIG. 5-8. The vertical detail depends on how the scanning lines cover the picture elements. (a) Each scanning line covers an individual vertical detail; (b) the scanning lines straddle the vertical details.

the uniform gray bar shown in *b*, with the equally spaced black and white details of the vertical bar in the original image completely lost. The number of picture elements that can be reproduced in the vertical direction, therefore, depends to a great extent on the position of the elements with respect to the scanning lines, since the vertical detail is partially or completely lost when a scanning line overlaps two adjacent vertical picture elements.

In an image with typical picture content there is a nonuniform arrangement of the picture elements and some of them fall directly on a scanning line while others straddle the scanning lines. The problem in establishing the useful vertical detail, then, is one of determining how many picture elements can be reproduced along a vertical line by a given number of scanning lines. This depends on the average number of elements that can be expected to fall directly on a scanning line when there is a random dis-

tribution of light and dark picture elements and varies with the picture content. The ratio of the number of scanning lines useful in representing the vertical detail to the total number of active scanning lines is called the *utilization ratio*. Theoretical calculations and experimental tests show that the utilization ratio ranges from 0.6 to 0.9 for different images with typical picture content, with 0.75 as an average value.

Employing the average utilization ratio of 75 per cent, the maximum possible number of vertical elements can be determined. The number of active scanning lines is equal to 525 minus those lines scanned during the blanking time. Taking a vertical blanking time equal to 8 per cent of the field-scanning period, the number of lines blanked out for the entire frame is  $0.08 \times 525$ , or 42 lines. Some of these lines occur during the vertical retrace, while others are scanned at the top and bottom of the frame during the blanking time; but all are inactive in scanning picture information. With 42 lines obliterated, 483 active lines remain. The number of active scanning lines effective in showing vertical detail is  $483 \times 0.75$  since this is the utilization ratio, providing 362 working lines. Therefore, the maximum number of vertical details that can be reproduced with 525 total scanning lines and 483 active scanning lines is about 362, the exact value depending upon the utilization ratio.

*Total Number of Picture Elements.* On the basis of the previous calculations, the maximum number of picture elements possible for the entire image is  $426 \times 362$ , or about 150,000. This number is independent of the picture size. With different images 100,000 to 200,000 picture elements may be accommodated, depending on the picture content. The total number of picture elements can be regarded as the figure of merit of the scanning pattern and may be compared with motion-picture reproduction. A single frame of 35-mm motion-picture film contains about 500,000 picture elements. The smaller 16-mm frame contains one-fourth as many, or about 125,000. The televised reproduction, therefore, can have about the same amount of details as 16-mm motion pictures. The detail in a 16-mm film reproduction is superior to a television picture, however, because the picture elements are reproduced as discrete units.

**5-5. The D-C Component of the Video Signal.** In addition to the fact that the video signal amplitude changes continuously with the variations of light for individual picture elements, the signal must have an average value that corresponds to the average brightness of the scene. The average is for complete frames and not individual lines in the frame. As an example of the importance of the background level, the camera signal corresponding to a gray line on a black background will be identical with the signal that corresponds to a white line on a gray background if there is no average brightness or background information. The detailed picture information corresponds to the variations of the signal within any hori-

zontal line, which may be considered as the a-c component of the video signal. The average value of the signal over the entire frame or several frames is the d-c level of the signal, which corresponds to the average light level because it is the axis from which the individual variations in the signal differ in accordance with the relative intensity of each picture element. Therefore, the d-c component of the video signal provides the average brightness or background information for the reproduced picture. The same detailed picture information may have an entirely different appearance when the average brightness or d-c level is changed. The d-c component is included in the video signal transmitted to the receiver, for the purpose of providing the correct average brightness in the picture reproduced on the kinescope screen.

*D-C Component for Dark and Light Scenes.* The d-c component of the composite video signal is its average value. As illustrated in Fig. 5-9,

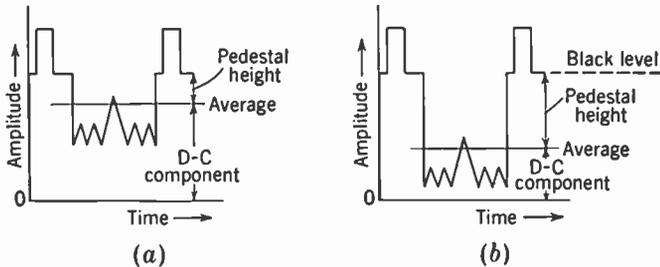


FIG. 5-9. Video signals with the same a-c variations but different average brightness values. (a) Dark scene with average value close to black level; (b) light scene with average value further from black level.

when the average value or d-c component is close to the pedestal level, as in *a*, the average brightness is dark. The same a-c signal variation in *b* has a lighter background because the average axis is farther from the black level. The average brightness of the reproduced picture, therefore, depends upon how far the average-value axis of the video signal is from the black level.

*Pedestal Height.* As noted in Fig. 5-9, the pedestal height is the distance between the pedestal level and the average-value axis of the video signal. This indicates the average brightness of the signal, since it indicates by how much the average value differs from the black level. Although it is not the d-c component of the video signal measured from the zero axis, the pedestal height is a convenient measure of the average brightness because the distance between the pedestal and average-value levels stays the same if the signal loses its d-c component. With the pedestal level as the fixed black reference level, therefore, the pedestal height can always indicate the correct brightness of the video signal.

*Setting the Pedestal Level.* The method of utilizing the pedestal level as the black reference level for the entire system can be followed by starting with the camera. The camera signal output is amplified in several stages before being coupled to a *control amplifier* at the studio of the transmitting station. At this stage the camera signal can be considered as being without the d-c level needed to indicate the brightness of the scene because of capacitive coupling in either the camera tube or the camera signal amplifiers. The synchronizing and blanking pulses are added to the camera signal in the control amplifier to produce the composite video signal, with the synchronizing pulses superimposed on the pedestals provided by the blanking pulses. Before the synchronizing pulses are added, however, the tops of the blanking pulses are cut off by a clipper amplifier. This is illustrated in Fig. 5-10. The level at which the blanking pulses

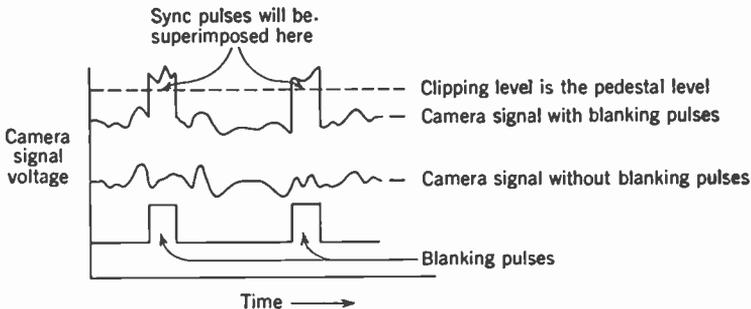


FIG. 5-10. Formation of the pedestals.

are clipped is the pedestal level and will be the black level that determines the brightness reference for the entire television system.

The pedestal level is determined in the control amplifier in terms of the average brightness of the scene being televised by varying the bias on the clipping amplifier, in order to change the level at which the blanking pulses are clipped to provide the pedestals. Manual adjustment of the clipping level is accomplished by an operator who observes the scene being televised and can set the clipper amplifier bias and the pedestal level to provide the desired brightness of the reproduced image, as observed on a monitor picture tube viewed simultaneously. Adjusting the clipping or pedestal level changes the average-value axis of the video signal and, therefore, is called *d-c insertion*.

Once the d-c insertion has been accomplished, the pedestal level becomes the desired black reference level and the pedestal height indicates the correct average brightness. The d-c level inserted in the control amplifier is usually lost in succeeding stages, though, because of capacitively coupled amplifiers that block the d-c component of the signal. However, once the pedestal level has been determined in the control

amplifier for the scene, it is not difficult to reinsert the d-c level lost in a capacitive coupling because the pedestal height is not altered. The d-c reinsertion is accomplished by a circuit that rectifies the a-c video signal to provide a d-c component proportional to the pedestal height, automatically producing the correct d-c component. At the transmitter, d-c reinsertion is employed in the modulation system to give the pedestal voltage a constant level 75 per cent of the peak amplitude of the picture carrier signal. In addition, d-c reinsertion is needed in the control-grid circuit of the picture tube when the d-c component of the video signal is lost in previous amplifiers in the receiver. The d-c reinsertion in the receiver ensures that the black-level amplitude of the video signal drives the kinescope control-grid voltage to cutoff and extinguishes the scanning



FIG. 5-11. NBC test pattern. (Copyright National Broadcasting Company.)

spot of light, thus reproducing the black reference level and all brightness values correctly.

**5-6. Test Patterns.** In order to adjust a television system conveniently and compare performances, a standard picture is desirable. This is usually in the form of a *test pattern*. Figure 5-11 shows a typical test pattern. Such test patterns, with circles, wedges, and a gray scale, are transmitted by many broadcast stations. A test pattern is helpful in making receiver adjustments and judging the performance of the receiver in terms of the reproduced picture.

The test pattern is composed of black and white lines and areas on a gray background. The pattern is constructed with approximately equal black and white areas, and the color value of the gray areas is halfway between black and white. This simulates an average scene so that no receiver adjustment should be necessary when changing from test pattern to program transmission. The black and white areas at the outer ends

of the horizontal wedges produce a signal that permits the voltage amplitude swing between white and black to be readily determined.

*Aspect Ratio, Linearity, and Brightness.* The aspect ratio is 4:3. Proper aspect ratio is obtained when the height of the picture is equal to the diameter of the large black inner circle and the diameter of the large white outer circle equals the width of the picture frame. Linearity of the scanning motions may be judged by the circles after the proper width and height have been set. If the circles appear round, the scanning linearity is properly adjusted. Linearity can also be checked by means of the horizontal and vertical wedges in the test pattern. The two vertical wedges are of equal length, as are the two horizontal wedges, and the ability to reproduce these equal lengths depends on the linearity of the vertical and horizontal scanning motions. Brightness and contrast are set by the use of the target in the center of the pattern. The target ranges from black in the center to white on the outside in five equal steps or shades.

*Focusing Adjustment.* Focus is set by varying the focus control until the black and white lines in the wedges can be distinguished as close to the center of the image as possible or when scanning lines are clearly definable in the center of the screen. Most picture tubes cannot be adjusted to perfect focus over the entire screen.

*Resolution.* The picture detail or resolution is measured on the test pattern in number of lines. If the vertical resolution is 150 lines in the reproduced picture, this means that it is possible to see 150 individual horizontal lines consisting of 75 black lines separated by 75 white lines. For equal resolution in the horizontal direction and an aspect ratio of 4:3,  $150 \times \frac{4}{3}$ , or 200, vertical lines can be resolved in the picture, consisting of 100 black lines separated by 100 white lines. However, this is still considered 150-line resolution because the resolution is measured in terms of the picture height when indicating either horizontal or vertical detail, in order to provide a common basis for comparison.

*Measuring Vertical Resolution.* The line divisions in the horizontal wedges of the test pattern measure the vertical resolution in the reproduced image and indicate the adequacy of interlacing. Both horizontal wedges are identical. At the outer edge of the horizontal wedge the vertical resolution indicated is 150 lines, and the resolution increases with the closer spacing of the lines in the wedge toward the bull's-eye circles. Where the wedge meets the outer edge of the white bull's-eye circle the resolution is 300 lines. Maximum vertical resolution is 372 lines for the test pattern. Resolution in the picture is interpreted in terms of the markers, which indicate the degree to which the individual lines in the wedge can be resolved. As an example, when the divisions in the wedge can be resolved to the point just touching the outer bull's-eye circle in

the reproduced picture, the vertical resolution is 300 lines. When the individual lines in the wedge cannot be distinguished from each other in the reproduced picture any further than the 250 marker, the resolution is 250 lines.

The vertical resolution of the reproduced image can be calculated in a very direct manner from the test pattern. There are 31 lines in the wedge, consisting of 16 black and 15 white lines. When the lines in the wedge can be resolved only to the point where the wedge meets the large outer black circle, 31 lines can be resolved in a distance equal to the height of the wedge at that point. Measuring this distance, it is slightly more than  $\frac{1}{5}$  of the total picture height, or exactly  $1/4.84$  of the height. Therefore,  $31 \times 4.84$  lines can be included in the entire picture height, with each individual line distinct from the others, and the vertical resolution is equal to 150 lines. The vertical resolution corresponding to any other part of the horizontal wedge can be measured in the same way.

*Measuring Horizontal Resolution.* The vertical wedges in the pattern measure horizontal resolution and the video-frequency response necessary to obtain the corresponding resolution. The top vertical wedge gives the horizontal resolution in number of lines referred to the picture height. For a horizontal resolution marked 300 lines, therefore, it is possible to resolve 400 black and white vertical elements in the horizontal direction. The white dots along the top wedge indicate the horizontal resolution as noted in the figure. The maximum horizontal resolution is 325 lines, which corresponds to 433 individual elements along the horizontal line. The horizontal detail corresponding to any point on the vertical wedge can be calculated by measuring the width of the wedge at the point where the individual lines can just be resolved in the reproduced picture, just as in the method used for measuring the vertical resolution. First, find the proportion of the picture width to the width of the vertical wedge at the point where the black and white divisions in the wedge can just be resolved. Multiplying this ratio by the 31 lines in the wedge, the total number of details in the horizontal direction is found. This number is then multiplied by  $\frac{3}{4}$  in order to have the resolution value in number of lines referred to the picture height. Thus, at the point where the vertical wedge meets the outer edge of the white bull's-eye circle the total picture width is 12.9 times the width of the wedge. Therefore the total number of lines that can be accommodated in the entire picture width with this resolution is  $12.9 \times 31$ , or 400. Multiplying by  $\frac{3}{4}$  to translate this value into terms of picture height, the horizontal resolution obtained is equal to 300 lines.

*Measuring Frequency Response.* The lower vertical wedge and the top one are identical; both indicate horizontal resolution, but the lower one is calibrated in frequency response. The lowest white dot near the black

circle is 2.0 Mc. The next dot toward the center of the pattern is 2.5 Mc, followed by a dot at 3 Mc; the dot nearest the center corresponds to a camera signal frequency of 3.5 Mc. The inner end of the wedge is 4.06 Mc. It is possible to mark the wedge in this manner because the extent to which the individual lines of the vertical wedge can be resolved in the image indicates the video-frequency limitations of the television system in terms of the horizontal detail reproduced.

The signal frequency corresponding to any degree of resolution in the vertical wedge can be calculated by means of a procedure similar to that used in calculating the maximum number of horizontal picture elements, as described in Sec. 5-4. As an example, if the individual lines in a vertical wedge can be resolved down to the point marked 3.5 Mc in the reproduced picture, the frequency response can be calculated in the following manner: There are 31 lines in the wedge, which correspond to  $15\frac{1}{2}$  complete cycles of the video signal obtained in scanning this part of the image. Measuring the width of the wedge at this point, it is found to be 8.2 per cent of the total picture width. The time required to scan across the wedge at this point, then, is 8.2 per cent of the total active scanning time of  $53.3 \mu\text{sec}$  for one horizontal line, which is equal to  $0.082 \times 53.3$ , or approximately  $4.4 \mu\text{sec}$ . This is the period for  $15\frac{1}{2}$  cycles of the corresponding square-wave camera signal, and the period for one complete cycle is  $4.4/15.5$ , or  $0.285 \mu\text{sec}$ . The frequency is the reciprocal of  $0.285 \mu\text{sec}$ , or 3.5 Mc. The corresponding frequency for any degree of horizontal detail in a vertical wedge can be found in this manner. It should be noted that the frequency calculations are not referred to the picture height and the factor  $\frac{3}{4}$  is not used.

*Frequency Response and Horizontal Resolution.* Both vertical wedges indicate the horizontal detail, although one is marked in frequency response while the other measures horizontal resolution in terms of the number of lines referred to the picture height. The equivalence between video-frequency response and number of lines for the corresponding horizontal resolution can be established by making the measurements just described. On this basis, the relation between the horizontal resolution in number of lines and the required video frequency for this resolution is

$$n = 80 \times f$$

$$f = \frac{n}{80}$$

where  $f_n$  is the frequency in megacycles, for a resolution of  $n$  lines (referred to the picture height). For example, a video-frequency response of 4 Mc is equal to  $4 \times 80$ , or 320-line horizontal resolution; 240-line resolution equals  $240/80$ , or 3-Mc video-frequency response.

**5-7. RETMA Resolution Chart.** The chart shown in Fig. 5-12 has been prepared by the Radio Electronics Television Manufacturers Association<sup>1</sup> for the purpose of standardizing resolution measurements. Resolution must be read only after the equipment is adjusted for minimum distortion. The RETMA chart is not generally broadcast because it is much more detailed than a typical test pattern. The chart has circles, wedges, and a gray scale, however, which can be interpreted in the same way as described for the NBC pattern.

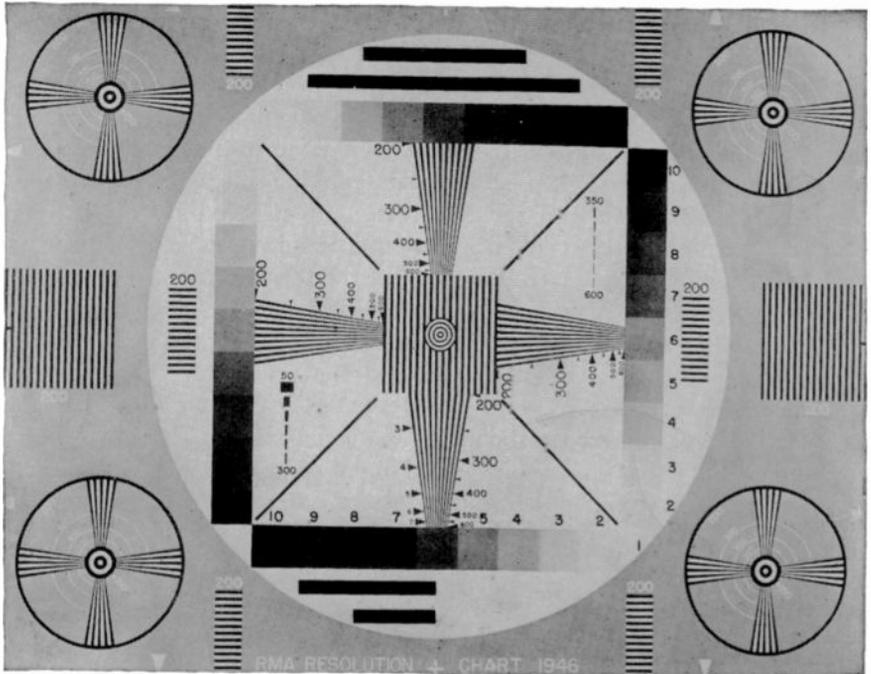


FIG. 5-12. RETMA resolution chart.

The wide horizontal and vertical bars form a large square within the large white circle. Each bar is numbered from 1 to 10 because it is constructed as a gray scale composed of 10 logarithmic steps from maximum white brightness to approximately one-tenth of this value. Aspect ratio is checked by noting that the four gray scale bars form a perfect square. The wedges in the corner circles permit linearity and resolution to be measured in the four corners of the picture. Larger wedges are also included in the large white circle in the center, and one of these is calibrated in both number of lines and frequency response. The resolution calibration extends to 600 lines and past 7 Mc so that it can be used for

<sup>1</sup> Formerly Radio Manufacturers Association (RMA).

equipment not limited to 4-Mc response. It is important to note that the 4-Mc video-frequency limitation results from the restriction of a 6-Mc transmission channel and is not the limit of the television equipment. Video frequencies as high as 8 Mc can be used at the television studio, but those above 4 Mc cannot be transmitted to the receiver. In addition to the wedges, vertical and horizontal groups of parallel bars are placed in the picture to check horizontal and vertical linearity. All these bars are spaced for 200-line resolution.

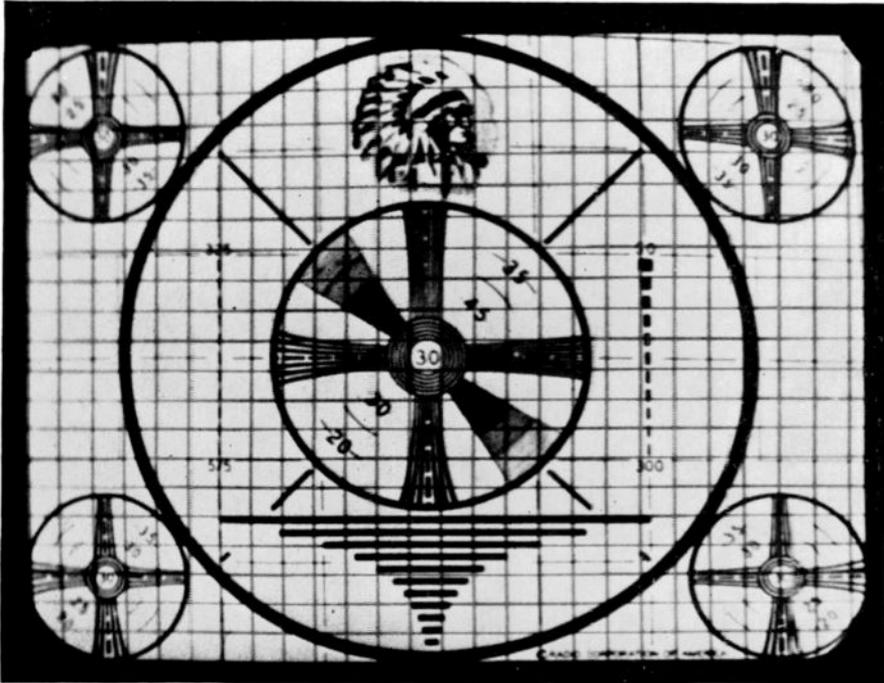


FIG. 5-13. Indian-head test pattern. (RCA.)

The small resolution circles in the center of the large white circle and in the centers of the four corner circles are for checking spot ellipticity on picture tubes. Resolution for the circles in the corners (150) is made less than the resolution in the center (300) because of the added defocusing in the corners caused by deflection of the scanning beam. The two sections of single-line widths, 50 to 300 and 350 to 600, read resolution just as do the wedges, but also provide an accurate means of checking the frequency of transient oscillations. The four diagonal lines in the large square at the center may be used to check the quality of interlacing. A jagged line indicates partial pairing of the interlaced lines. This is not effective for complete pairing of the lines, when even and odd fields exactly overlap.

However, the reduced vertical resolution in such a case will be plainly evident. The gray background of the chart provides a satisfactory balance with the white areas; a system correctly set up by the use of this chart will operate satisfactorily on an average scene without additional adjustments.

**5-8. Indian-head Pattern.** The test chart generally printed on the image plate of monoscopes is the one shown in Fig. 5-13, which is commonly called the Indian-head pattern. A feature of this pattern is the reproduction of the Indian head, which enables checking the quality of gray tones in an actual picture. The Indian-head chart also has uniformly spaced square boxes throughout the image. This *crosshatch* of lines is convenient for checking the linearity of vertical or horizontal scanning in any part of the picture. When the scanning is nonlinear the boxes are not square, showing crowding and spreading in the vertical or horizontal direction.

**5-9. Gamma.** This is a numerical factor used in television for indicating to what extent light values are expanded or compressed. Referring to Fig. 5-14, the exponent of the equations for the curves shown is called *gamma* ( $\gamma$ ), corresponding to a similar term used for indicating contrast characteristics in photographic reproduction. The numerical value of gamma is equal to the slope or gradient of the straight-line part of the curve where it rises most sharply. A curve with a gamma of less than one is bowed downward as in *a* of Fig. 5-14, with the greatest slope at the start of the curve and the relatively flat part at the end. When the gamma is more than one the curve's shape is bowed upward as in *b*, making the start of the curve comparatively flat while the sharp slope is at the end. With a gamma of one the result is a straight line as in *c*, where the slope is the same throughout.

Since the shape of the curve is determined by the value of gamma, it can be used to indicate how the over-all television system or any of its components expands or compresses the contrast of the reproduced picture. A gamma value of one means a linear characteristic that does not exaggerate any light values. When the gamma characteristic is greater than one for the white parts of the image, the reproduced picture looks "contrasty" because the increases in white level are expanded by the sharp slope, to emphasize the white parts of the picture. Commercial motion pictures shown in a darkened theater have this high-contrast appearance. Gamma values of less than one for the white parts of the image compress the changes in white levels to make the picture appear softer, with the gradations in gray level more evident.

Any component in the television system can be assigned a value of gamma to describe the shape of its response curve and contrast characteristics. As a typical example, picture tubes have the control-characteristic

curve illustrated in *b* of Fig. 5-14. The video signal voltage is always impressed on the control grid of the picture tube with the polarity required to make the signal variations for the white parts of the picture fall on that part of the response curve with the steep slope. As a result, a variation in video signal amplitude at the white level produces a greater change in beam current and screen brightness than it would at a darker level. Picture tubes have a response characteristic with a gamma greater than one, therefore, emphasizing the white parts of the picture.

Amplifiers have a gamma characteristic that is very nearly unity, using linear operation. This is illustrated in Fig. 5-14*c*, the straight-line response showing that the output signal voltage is proportional to the input voltage without emphasizing any signal level. If desired, however,

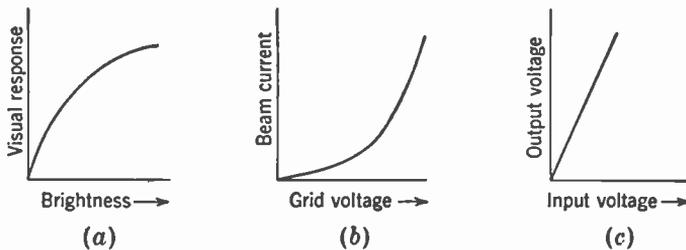


FIG. 5-14. Gamma characteristics. (a) Visual response of the eye. Gamma is less than one; (b) control-grid characteristic curve of picture tube (here gamma is more than one—equal to 2.2 in this case); (c) linear characteristic of amplifier: gamma equals one.

an amplifier can be made to operate over the curved portion of its transfer-characteristic curve by shifting the operating bias. The nonlinear amplifier can be used as a gamma-control stage, therefore, to expand or compress the white video signal amplitudes relative to the black level.

#### REVIEW QUESTIONS

1. Show the composite video signal corresponding to three horizontal lines in scanning a black frame with two equally spaced vertical white bars.
2. Show the composite video signal for three horizontal lines in scanning across a horizontal white bar.
3. What is the pedestal level?
4. How long is the horizontal blanking time for one line? Where is the scanning beam during the time corresponding to the front porch and the back porch of the synchronizing pulse?
5. Why are the synchronizing pulses inserted during the blanking time?
6. How long is the vertical blanking time for one field? Trace the action of the scanning beam from the beginning of vertical blanking until removal of the blanking pulse for active scanning of picture information.
7. Show the composite video signal for five horizontal lines in scanning a pattern with alternate black and white horizontal lines of a width equal to the diameter of the scanning beam. Assume that the scanning beam just covers every line without any straddling of lines.

8. What is the limiting factor in restricting the amount of horizontal detail that can be transmitted in the television system?

9. In the checkerboard pattern of Fig. 5-7, if the pattern is assumed to have 400 black and white squares in a horizontal line, what is the frequency of the square-wave signal variations obtained in scanning a line? The horizontal blanking time is 16 per cent of the line period.

10. What is the utilization ratio? What is a typical value for this ratio? Why does it vary with picture content?

11. Assuming a value of 0.8 for the utilization factor, what would be the maximum number of vertical details possible in the reproduced picture with a vertical blanking time of 8 per cent?

12. For the case of Question 11, what will be the maximum value of the total number of picture elements in the frame? Compare the maximum possible vertical and horizontal resolution.

13. What is the disadvantage of increasing the total number of scanning lines to a value greater than 525, with the highest possible video frequency remaining at 4 Mc?

14. Explain what is meant by the statement that picture tubes have a gamma characteristic greater than one.

15. From the reproduction of the test pattern in Fig. 5-11 calculate the vertical resolution at a point on the horizontal wedge exactly midway between the 150 and 200 markers.

16. From the reproduction of the test pattern in Fig. 5-11 calculate the horizontal resolution at a point on the vertical wedge exactly midway between the 2.5- and 3-Mc markers in number of lines referred to the picture height. With this resolution, how many horizontal details can be reproduced along a horizontal line?

17. From the reproduction of the RETMA resolution chart calculate the horizontal resolution in number of lines referred to the picture height for the vertical group of parallel lines in the center of the picture.

18. What is the purpose of the d-c insertion at the transmitter?

19. Why are the horizontal blanking pulses wider than the horizontal synchronizing pulses?

20. Why do the vertical retraces automatically occur within vertical blanking time, with synchronized vertical scanning?

## CHAPTER 6

### PICTURE CARRIER SIGNAL

The method of transmitting the AM picture signal is similar to the more familiar system of sound transmission in the standard broadcast band, where the amplitude of an r-f carrier wave is made to vary at the audio rate. In television broadcasting the composite video signal modulates a high-frequency carrier wave to produce the AM picture signal<sup>1</sup> illustrated in Fig. 6-1. The amplitude of the radiated carrier wave varies in accordance with the video modulating signal, thus producing the envelope of

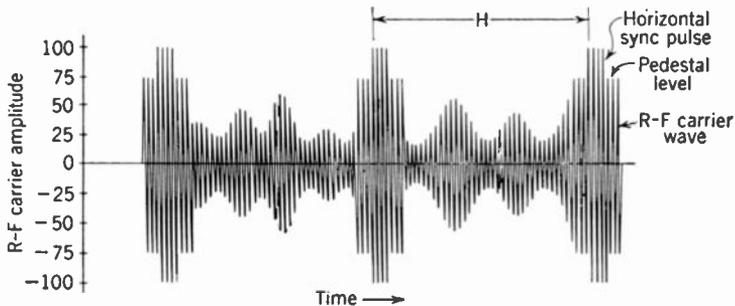


FIG. 6-1. The transmitted picture carrier wave, amplitude-modulated by the composite video signal.

amplitude variations in the r-f carrier that corresponds to the video modulating voltage. The envelope of the modulated picture carrier, then, is the composite video signal, containing all the information needed for picture reproduction. In this way the desired camera signal, blanking pulses, and synchronizing pulses are transmitted to the receiver as the envelope of the modulated picture carrier signal. At the receiver the picture signal is detected to recover the composite video signal, which is then used to reproduce the picture.

<sup>1</sup> The term *picture signal* is used here to denote the modulated r-f carrier wave, while *video* represents the signal that can be used directly to reproduce the desired visual information when applied to a picture-reproducing tube, corresponding to *audio* in a sound system.

**6-1. Negative Transmission.** Referring again to Fig. 6-1, the modulated picture carrier signal is shown with the negative polarity of transmission that is standard practice for all commercial television broadcast stations in the United States. The polarity of the video modulating signal is purposely chosen to give the modulated r-f carrier the amplitude characteristics shown, with decreasing carrier amplitudes for increasing light intensities in the televised image. The tips of the synchronizing pulses produce the peak values of r-f amplitude in the modulated carrier wave. The pedestal level is transmitted at a constant amplitude equal to 75 per cent of the peak carrier level. Smaller amplitudes in the modulated r-f carrier signal correspond to picture information that varies between black and maximum white, and the brightest parts of the picture must produce a carrier amplitude that is 15 per cent or less of the peak value.

The negative transmission is merely an arbitrarily assigned method of transmitting the video signal that depends on the polarity of the video modulating signal at the transmitter. The polarity must be standardized, however, because it determines the sense of color values in the reproduced picture at the receiver. If a receiver designed for a picture signal transmitted with negative polarity is used in a positive transmission system all the colors in the reproduced picture will be reversed, the white parts of the original image appearing as black in the reproduced picture while black areas become white.

Positive transmission could also serve as the basis for transmitting the picture signal and is used instead of negative transmission in other countries. In positive transmission, the synchronizing pulses reduce the carrier amplitude toward zero and the maximum white level corresponds to the peak carrier amplitude. The choice between the two polarities of transmission is not very definite because each has its own merit. In a negative system any temporary increase in the signal level caused by such interference as automobile ignition noise moves the signal amplitude toward the black and infrablack region. In such a case the effect of the interference in the reproduced picture is to reduce the brilliance and produce in the picture darker areas varying with the duration and amplitude of the interference. This is not so noticeable as interference effects which increase the brightness and produce white flashes in the picture. With positive transmission, any increase in the picture signal amplitude caused by interfering signals would move the carrier amplitude toward the white level, making the interference effects very obvious in the reproduced picture.

When negative transmission is used, the synchronization is more vulnerable to interference because noise pulses increase the carrier amplitude, in the same direction as the synchronizing pulses. As a result, inter-

fering noise pulses can be mistaken for synchronizing pulses in the receiver circuits. However, the effect of noise on synchronization in the receiver has been minimized by the development of stabilizing circuits that can be controlled by the synchronizing pulses but are relatively immune to interfering noise pulses.

The improvement in synchronization obtainable with positive transmission is not considered great enough to offset the advantage of reduced interference effects in the reproduced picture and several other benefits provided by the negative polarity of transmission. With negative trans-

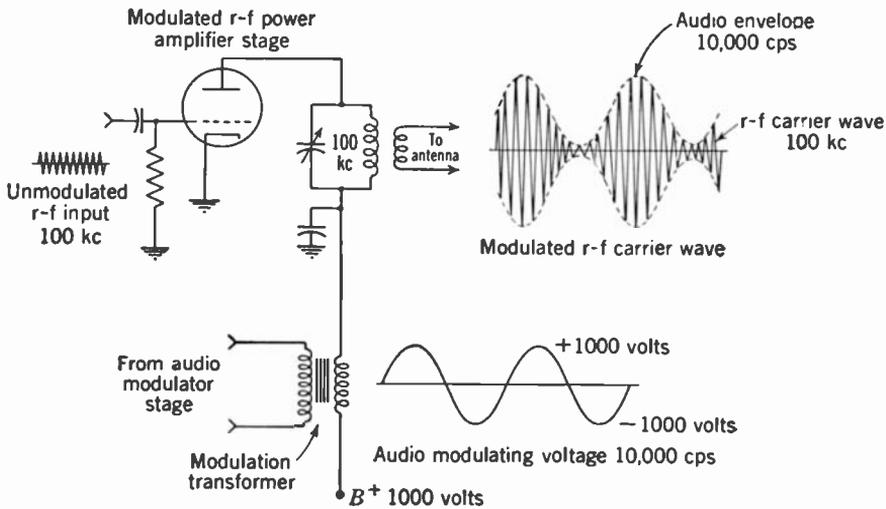


FIG. 6-2. A plate modulation arrangement for amplitude modulation of a 100-kc carrier wave by 1000-cps audio modulating voltage.

mission, the constant black and infrablack levels have a higher amplitude than the varying camera signal and, therefore, can be used as the control voltage for an automatic gain control system in the receiver with a more simple circuit arrangement than would be possible when positive transmission is used for the picture signal. Finally, there is an appreciable advantage in power efficiency at the transmitter when negative transmission is used.

**6-2. Vestigial-side-band Transmission.** The AM picture signal is not transmitted as a normal double-side-band signal. Using vestigial-side-band transmission, some of the side-band frequencies are filtered out before transmission in order to reduce the bandwidth of the channel needed for the modulated picture signal.

*Amplitude Modulation.* The concept of side-band frequencies can be illustrated in terms of the simple AM system shown in Fig. 6-2, where an

r-f carrier wave is amplitude-modulated by a sine-wave audio signal in a plate modulation arrangement. For simplicity the r-f carrier frequency is taken as 100 kc and the audio as 10,000 cps. The B supply voltage for the r-f power amplifier is assumed to be 1,000 volts and the peak value of the audio sine-wave modulating voltage is also 1,000 volts, allowing 100 per cent modulation.

The effect of the audio voltage across the secondary of the modulation transformer, which is in series with the B supply and the power-amplifier plate-to-cathode circuit, is to vary the power-amplifier plate voltage at the audio rate. When the audio voltage is zero during the modulation cycle, the amplifier plate voltage is equal to the B supply voltage of 1,000 volts, providing a circulating tank current in the tuned circuit and a current induced into the antenna circuit at the unmodulated carrier level. When the audio modulating voltage has a value of +1,000 volts the effective plate voltage for the r-f power amplifier is increased to 2,000 volts to double the circulating tank current and the antenna current. As a result, the carrier amplitude increases to twice the unmodulated level at the peak of the positive half cycle of the modulating voltage. Intermediate values of audio modulating voltage between zero and +1,000 volts increase the amplitude of the r-f carrier wave proportionately, with the carrier amplitude varying between the unmodulated level and the peak value, which is twice the unmodulated amplitude. While the audio modulating value decreases from its peak value of +1,000 volts to zero, the r-f amplifier plate voltage varies from 2,000 volts back to 1,000 volts and the carrier amplitude is reduced from its peak value down to the unmodulated level. The audio modulating voltage then swings through its negative half cycle, reducing the effective value of r-f amplifier plate voltage below the B supply voltage of 1,000 volts. At the peak of the negative half cycle of the audio modulating voltage when the audio voltage is -1,000 volts, the r-f plate voltage is reduced to zero and the carrier amplitude is also reduced to zero. As the value of the audio modulating voltage varies from its negative peak back to zero the r-f amplifier plate voltage is brought back to the B supply voltage of 1,000 volts, when the carrier amplitude is again at the unmodulated carrier level. Thus, a complete cycle of audio modulating voltage varies the carrier amplitude from the unmodulated level up to a peak value equal to twice this level for a 100 per cent modulation, down to the unmodulated carrier level, and then down to zero before returning to the unmodulated carrier level again; while the r-f amplifier plate voltage is varied from the B supply value up to twice the B voltage, back to the unmodulated value, and then down to zero voltage before returning to the unmodulated value of plate voltage equal to the B supply voltage. With continuous audio modulating voltage the same cycle is repeated over and over again.

The modulated carrier wave varies in amplitude at the audio rate, and the varying amplitudes of the r-f carrier wave provide an envelope, which corresponds to the audio modulating voltage. Both the positive and negative peaks of the r-f carrier wave are symmetrical about the center axis and have exactly the same amplitude variations, since the changes in amplitude of the negative and positive half cycles of the r-f signal are equal when the amplitude of the carrier is varied at the much slower audio rate. The result of the modulation in this case, then, is to produce an r-f carrier wave at a frequency of 100 kc with an amplitude that varies at the audio rate of 10,000 cps, producing on the carrier a symmetrical envelope that corresponds to the 10,000-cps audio modulating signal.

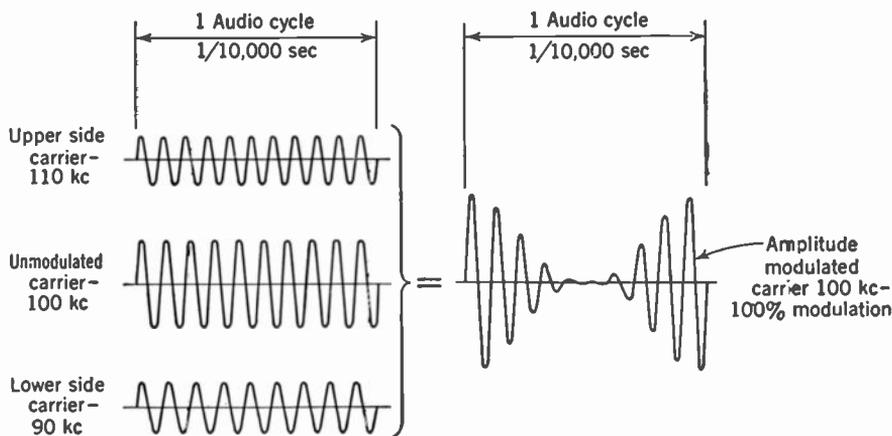


FIG. 6-3. Equivalence of the amplitude-modulated carrier to the modulated carrier plus the two side carriers produced by modulation.

*Side-carrier Frequencies.* Referring now to Fig. 6-3, it is shown that the AM wave is equal to the sum of the unmodulated r-f carrier and two side-carrier frequencies. The carrier and the side frequencies have a constant level with the amplitude of the side carriers equal to one-half the unmodulated carrier level, for 100 per cent modulation. Each side frequency differs from the carrier by the audio modulating frequency; the upper side frequency is 110 kc and the lower side frequency 90 kc in this illustration. The question as to whether the transmitted signal is a carrier with varying amplitudes or a constant-amplitude carrier with its two side carriers is without meaning, because the two concepts are the same. The constant-level side carriers plus the unmodulated carrier wave are equal to the AM carrier signal. Conversely, the AM carrier wave is equal to the unmodulated carrier plus two side carriers of proper amplitude, phase, and frequency. The equivalence of the two signals is due to the fact that the modulated r-f carrier wave is distorted slightly from true

sine-wave form by the amplitude variations, producing new frequency components. The side-carrier frequencies should not be confused with the envelope of the modulated carrier signal. For the case illustrated here the envelope is the audio modulating signal of 10,000 cps, while the r-f side-band frequencies are 110 kc and 90 kc for the upper and lower side frequencies, respectively.

*Single-side-band Transmission.* Since the AM signal is equal to the unmodulated carrier plus two side carriers, the modulation energy and information must be in the side carriers. In addition, each side carrier contains identical modulation energy and information because they are both of equal amplitude and differ in frequency from the carrier by an equal amount. Therefore, it is possible to transmit the modulating signal by means of the carrier and only one side frequency, and it does not matter

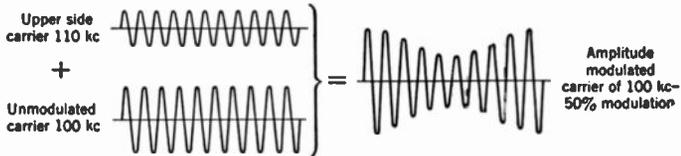


FIG. 6-4. Equivalence of the unmodulated carrier plus one side carrier to the amplitude-modulated carrier.

whether the upper or lower side frequency is used. This is called *single-side-band* transmission. The resultant modulated carrier wave is illustrated in Fig. 6-4 with amplitude variations that correspond to 50 per cent modulation instead of the 100 per cent modulation produced with both side bands. Except for the amount of amplitude swing, the envelope of the AM signal transmitted with only one side band can be considered the same as with double-side-band transmission, although there is some distortion of the envelope for high degrees of modulation. The envelope is still symmetrical about the center axis of the carrier, corresponding to the audio modulating voltage, and is not cut off at either the top or the bottom of the modulated carrier wave. In order to remove one part of the envelope it would be necessary to rectify the modulated carrier signal.

When the carrier is modulated by a voltage that includes many frequency components a band of side frequencies is produced, with a pair of side carriers for each modulating frequency. The transmitting arrangement may then allow both the upper and lower side carriers to be transmitted with normal double-side-band transmission for some modulating frequencies while transmitting only one side carrier for other modulating frequencies, in addition to the transmitted carrier frequency. Such an arrangement is called *vestigial-side-band* transmission because only part

of one side band is transmitted, with all of the other side band. The usual AM process produces the normal double-side-band frequencies plus the carrier, but any undesired side-carrier frequencies can be filtered out so that they will not be radiated from the transmitter.

**6-3. The Television Channel.** Each television broadcast station is assigned a 6-Mc channel for transmission of the AM picture signal and the FM sound signal. In order to accommodate the side-band frequencies produced with video modulating frequencies as high as 4 Mc, vestigial-side-band transmission is used for the picture signal.

*Assigned Channels.* Since the picture carrier frequency must be much higher than the highest video modulating frequency of 4 Mc, the television channels are assigned in the v-h-f 30- to 300-Mc band and the u-h-f 300- to 3,000-Mc band. Table 6-1 lists the channels and frequencies assigned by the FCC for commercial television broadcast stations in the United States. The television channel frequencies can be considered in three groups: the five low-band channels 2 to 6 in the v-h-f range; seven high-band channels 7 to 13 in the v-h-f range; and the seventy u-h-f channels 14 to 83. Frequencies between these television broadcasting bands are used by other radio services.

The number of channels available for television broadcast stations in any one locality depends upon its population, varying from one channel in a smaller city to nine for New York City, including v-h-f and u-h-f channels. One channel can be used by many broadcast stations, provided that they are separated by 155 to 220 miles in order to minimize interference between them. Stations using the same channel are *co-channel stations*. Stations that use channels adjacent in frequency, like channels 3 and 4, are *adjacent channel stations*. To minimize interference between them, adjacent-channel stations are not assigned in the same city but are separated by 55 to 60 miles or more. However, channels consecutive in number but not adjacent in frequencies, such as channels 4 and 5, channels 6 and 7, or channels 13 and 14, can be assigned in one area.

*The Standard Channel.* The structure of a standard television channel is illustrated in Fig. 6-5. The width of the channel is 6 Mc, including the picture and sound carriers with their side-band frequencies. The picture carrier is spaced 1.25 Mc from the lower edge of the channel, and the sound carrier is 0.25 Mc below the upper edge of the channel. As a result there is always a fixed spacing of 4.5 Mc between the picture and sound carrier frequencies.<sup>1</sup>

The standard channel characteristics shown in Fig. 6-5 should not be interpreted as an illustration of the picture signal. The graph merely

<sup>1</sup> The picture and sound carrier frequencies for all television broadcast channels are listed in the Appendix.

## BASIC TELEVISION

TABLE 6-1. TELEVISION CHANNEL ALLOCATIONS

Channel number	Frequency band, Mc	Channel number	Frequency band, Mc
1*	.....	42	638-644
2	54-60	43	644-650
3	60-66	44	650-656
4	66-72	45	656-662
5	76-82	46	662-668
6	82-88	47	668-674
7	174-180	48	674-680
8	180-186	49	680-686
9	186-192	50	686-692
10	192-198	51	692-698
11	198-204	52	698-704
12	204-210	53	704-710
13	210-216	54	710-716
14	470-476	55	716-722
15	476-482	56	722-728
16	482-488	57	728-734
17	488-494	58	734-740
18	494-500	59	740-746
19	500-506	60	746-752
20	506-512	61	752-758
21	512-518	62	758-764
22	518-524	63	764-770
23	524-530	64	770-776
24	530-536	65	776-782
25	536-542	66	782-788
26	542-548	67	788-794
27	548-554	68	794-800
28	554-560	69	800-806
29	560-566	70	806-812
30	566-572	71	812-818
31	572-578	72	818-824
32	578-584	73	824-830
33	584-590	74	830-836
34	590-596	75	836-842
35	596-602	76	842-848
36	602-608	77	848-854
37	608-614	78	854-860
38	614-620	79	860-866
39	620-626	80	866-872
40	626-632	81	872-878
41	632-638	82	878-884
		83	884-890

\* The 44- to 50-Mc band was television channel 1 but is now assigned to other services.

defines the signal frequencies that can be transmitted in the television channel, with their relative amplitudes. The picture carrier is shown with twice the amplitude of the side-band frequencies for 100 per cent modulation. Since the sound signal is frequency-modulated its side-band frequencies do not have the same type of amplitude characteristic as in the picture signal, and these are not shown. The sound carrier signal is a conventional double-side-band FM signal, with a bandwidth of approximately 50 kc for a frequency swing of  $\pm 25$  kc.

Vestigial-side-band transmission is used for the picture signal. All the upper side-band frequencies up to approximately 4-Mc video modulation are transmitted with their normal amplitude, as are all lower side-band frequencies that differ from the carrier frequency by 0.75 Mc or less. Lower side-carrier frequencies that differ from the picture carrier by more than 0.75 Mc but less than 1.25 Mc are gradually attenuated. The lower side-carrier frequencies that are 1.25 Mc or more below the picture carrier fall outside the channel and must be completely filtered out at the transmitter so that they will not be radiated to interfere with the lower adjacent channel. Upper side-carrier frequencies more than 4 Mc higher than the picture carrier are also attenuated in order to eliminate interference with the associated sound signal.

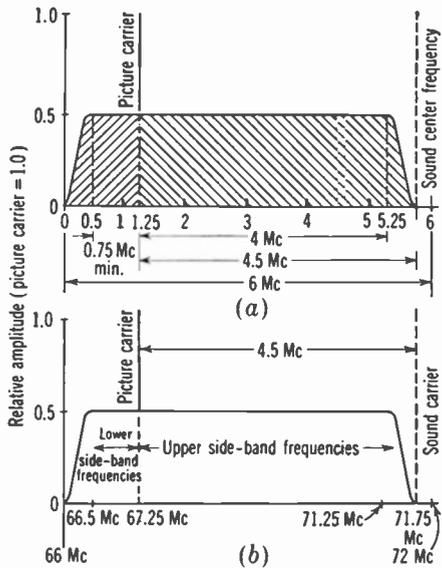


FIG. 6-5. The standard television channel. (a) The generalized structure for any channel; (b) channel 4, 66 to 72 Mc.

Numerical values for channel 4 as a typical television channel are shown in Fig. 6-5b. The channel has a bandwidth of 6 Mc from 66 to 72 Mc. The picture carrier is 1.25 Mc above the lower edge of the channel, which is 67.25 Mc for this channel, while the sound carrier is 71.75 Mc, 4.5 Mc above the picture carrier frequency. With vestigial-side-band transmission, the upper side-band frequencies to 71.25 Mc and lower side-band frequencies to 66.5 Mc, approximately, are transmitted without attenuation. As an example, when the video modulating voltage has a frequency of 0.75 Mc, both the upper and lower side frequencies of 68 Mc and 66.5 Mc, respectively, are transmitted without attenuation and the picture signal is transmitted as a normal double-side-band signal. The

same is true for any video modulating signal having a frequency that is less than 0.75 Mc. For the components of the video modulating signal with a frequency higher than 0.75 Mc, however, only the upper side carrier is transmitted with normal amplitude. With 2-Mc video modulation, as an example, the upper side frequency is 69.25 Mc, while the lower side frequency is 65.25 Mc, which is outside channel 4 and must be filtered out at the transmitter. In this case, then, only the upper side frequency is transmitted with the picture carrier, resulting in single-side-band transmission. The result is a vestigial-side-band transmission system where double-side-band transmission is used for the video modulating frequencies lower than 0.75 Mc and single-side-band transmission is used for the higher video modulating frequencies up to 4 Mc, approximately.

The advantage of using vestigial-side-band transmission for the picture signal can be seen from the fact that video modulating frequencies up to 4 Mc can be transmitted in the 6-Mc channel. A video-frequency limit of about 2.5 Mc would be necessary if double-side-band transmission were used with the picture carrier at the center of the channel. This would represent a serious loss in horizontal detail that could otherwise be utilized in the picture reproduction at the receiver, since the high-frequency response of the television system limits the amount of horizontal detail that can be obtained. It might seem desirable to place the picture carrier at the lower edge of the channel and use single-side-band transmission completely, allowing the use of video modulating frequencies higher than 5 Mc and increased horizontal detail, but this is not practicable. The elimination of undesired side-carrier frequencies is accomplished by a filter circuit at the transmitter, which cannot have ideal cutoff characteristics. Therefore, it would not be possible to remove side carriers that are too close to the carrier frequency without introducing objectionable phase distortion for the lower video signal frequencies, which causes smear in the picture. The practical compromise of vestigial-side-band transmission that is used provides for complete removal of the lower side band only where the side-carrier frequencies are sufficiently removed from the picture carrier to avoid phase distortion. The picture carrier itself and all side frequencies close to the carrier are not attenuated.

It should be noted that the required response characteristics of the transmitter channel distort the picture signal in terms of the relative amplitude for different frequencies. Since a signal transmitted with only a single side band and the carrier represents 50 per cent modulation in comparison to a normal double-side-band signal with 100 per cent modulation, the higher video frequencies provide signals with one-half the effective carrier modulation obtained for the lower video frequencies that are transmitted with both side bands. This is in effect a low-frequency boost in the video signal. However, it is corrected by deemphasizing

the low video frequencies to the same extent in the i-f amplifier of the television receiver.

**6-4. Line-of-sight Transmission.** Radio waves having frequencies higher than 30 Mc are not ordinarily returned from the ionized layers of the upper atmosphere surrounding the earth. As a result, propagation of radio waves in the v-h-f and u-h-f bands is produced mainly by ground-wave effects, rather than sky waves from the ionized atmosphere. The ground wave is that part of the radiated signal affected by the presence of the earth and can be considered as being propagated along the surface of the earth from the transmitting antenna. Since the television broadcast channels are in the v-h-f and u-h-f bands, transmission of the picture and sound carrier signals is determined primarily by ground-wave propagation.

*Horizon Distance.* The transmission distance that can be obtained for the ground-wave signal is limited by the distance along the earth's

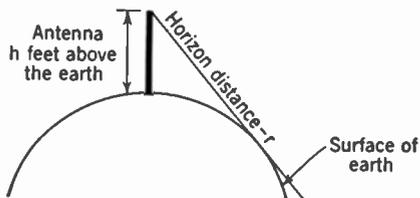


FIG. 6-6. The horizon distance  $r$  depends on the antenna height  $h$ .

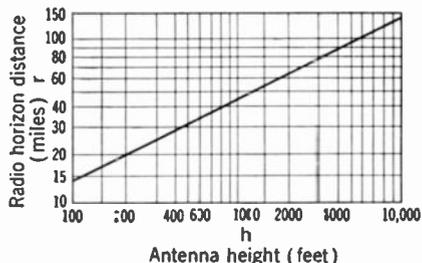


FIG. 6-7. Plot of Eq. (6-1) showing relation between horizon distance and antenna height.

surface to the horizon, as viewed from the transmitting antenna. This is called *line-of-sight transmission*, and the line-of-sight distance to the horizon is the *horizon distance*. As illustrated in Fig. 6-6, transmission from an antenna with a height  $h$  is limited to the horizon distance  $r$  because of the curvature of the earth's surface. The horizon distance for the transmitted radio wave, however, is about 15 per cent longer than the optical horizon distance because the path of the ground wave curves slightly in the same direction as the earth's curvature. This bending of the radio waves by the earth's atmosphere is called *refraction*. Making allowance for refraction, the formula for the radio horizon distance is

$$r = 1.41 \sqrt{h} \tag{6-1}$$

when  $h$  is antenna height in feet and  $r$  is the horizon distance in miles. The graph in Fig. 6-7 is a plot of this equation showing the radio horizon distance directly for any antenna height up to 10,000 ft.

Figure 6-8 shows several television transmitting antennas mounted at the top of the Empire State Building in New York City, in order to increase the antenna height and horizon distance. This antenna height is about 1,500 ft, providing a radio horizon distance of approximately 50 miles.

When considering the line-of-sight distance from the transmitting antenna to the receiving antenna, the horizon distance of each must be

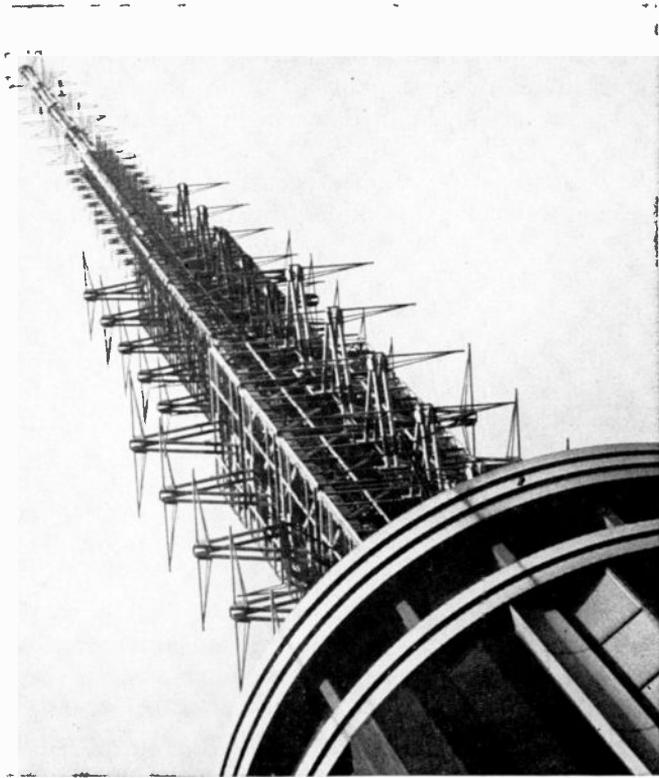


FIG. 6-8. V-h-f transmitting antennas atop the Empire State Building in New York City.

considered. For an antenna height of 150 ft at the receiver, as an example, the radio horizon distance is approximately 17 miles, and line-of-sight communications could be obtained with a transmitting antenna atop the Empire State Building for a distance of 17 miles plus 50 miles, or 67 miles. The transmitting and receiving antennas should be mounted as high as possible, therefore, for line-of-sight transmission over appreciable distances. The best signal is obtained when the receiving antenna has a line-of-sight path to the transmitting antenna.

*Service Area.* The practical service area of a television broadcast transmitter is within the radio horizon distance. The strength of the ground-wave signal decreases rapidly, however, as the distance from the transmitting antenna increases. The service area of a television broadcast station is determined by computing or measuring the boundary within which the field strength is  $500 \mu\text{v}$  or more per meter with a height of 30 ft for the receiving antenna. The  $500 \mu\text{v}$  per meter contour may extend 25 to 75 miles from the station, approximately, depending upon the height of the transmitting antenna and the amount of radiated power.

*Reflections.* As the ground wave travels along the surface of the earth, the radio signal encounters buildings, towers, bridges, hills, and other obstructions. When the intervening object is a good conductor and its size is an appreciable part of the radio signal's wavelength, the obstruction will reflect the radio wave, similar to the reflection of light from a mirror or other reflecting surface. Reflection of radio waves can occur at any frequency but is produced more easily at higher frequencies, because of the shorter wavelengths. For the television channel frequencies between 54 and 890 Mc the wavelengths are between 17 ft and 1 ft, depending on the frequency. Objects of comparable size, or bigger, can reflect the television carrier waves. When the reflected picture carrier signal arrives at the receiving antenna in addition to the direct wave or along with other reflections, the multipath signals produce multiple images called *ghosts* in the reproduced picture.

*Shadow Areas.* Where an object in the path of the ground wave reflects the radio signal, the area behind the obstruction is shadowed and therefore has reduced signal strength. The shadowing effect is more definite at higher frequencies because of the shorter wavelengths, just like reflection of the radio waves. Reception of television signal in shadow areas behind an obstruction like a tall building is often accomplished by utilizing waves reflected from other buildings nearby.

*Diffraction.* The radio horizon distance sets the limit for direct reception of the ground wave, considering both the transmitting and receiving antenna heights. However, the signal can be propagated a little further than the horizon by *diffraction* around the edge of the earth's surface at the horizon. This effect depends on the terrain, though, and the signal beyond the horizon due to diffraction decreases in strength much more rapidly than the direct wave. In unusual cases that depend on atmospheric conditions, the television carrier frequencies may be returned from the ionized layers of the earth's atmosphere to provide reception over very long distances beyond the horizon, but this is not dependable.

*Satellite Stations.* In an attempt to provide satisfactory television broadcast service to isolated communities in mountainous areas, *satellite*

or *booster* stations have been operated on an experimental basis. The satellite station can be in a suitable location for receiving television signals from a distant transmitter and rebroadcasts the program within the assigned television channels to receivers in the local area.

**6-5. Television Broadcasting.** A commercial television broadcast station includes equipment for production of the camera signal, formation of the composite video signal, and transmission of the picture signal for reproduction of the televised scene at the receiver. Since the associated sound must also be broadcast, audio facilities and an FM transmitter for the sound signal are included.

The equipment used in broadcasting a televised scene can be considered in two parts: the studio and the transmitter. At the studio the camera pickup generates the camera signal to which blanking and synchronizing pulses are added to produce the composite video signal. At the transmitter the composite video signal is amplified to the extent necessary for modulation of the picture carrier, and the modulated picture signal is radiated from the transmitter antenna. The studio normally includes facilities for direct pickup of live-talent shows and motion-picture film. In addition, field equipment may be used for remote pickups in televising sport features and other special events outside the studio. In these cases the video signal from the field equipment is relayed to the studio before being broadcast. The studio location should be convenient for the talent used in studio productions, while the transmitter must be placed where sufficient height can be obtained for the transmitting antenna. Therefore, the studio and transmitter are often at separate locations, connected by means of a cable or radio link. As an illustration, studios of television station WNBT of the National Broadcasting Company are in the RCA Building at Radio City, New York, while the transmitter is at the Empire State Building about 1 mile away.

*Live-talent Studio.* Live-talent programs are staged in a television studio, as shown in Fig. 1-4, or in a theater for big musical productions. Two or three cameras are generally employed to provide close-ups or wide-angle views for long shots and allow switching from one scene to another. Each camera has a turret with lenses of different focal length, or a *Zoomar lens* is used. The *Zoomar* is a telephoto lens of variable focal length which can easily be adjusted for quick close-up shots, while maintaining correct focus. Image orthicon cameras are generally used. The camera is mounted on a movable pedestal or dolly for flexible movement over the studio floor. The camera head is set on a universal mounting, which permits tilting up and down or movement horizontally for *panning* across the scene to obtain a panoramic view. Lamps mounted on the front of the camera are turned on from the studio control room, serving as *tally lights* or *cue lights* to indicate which camera is in use.

*Televising Motion-picture Films.* A separate small studio is employed for televising film. Projectors are available for 35- and 16-mm motion-picture film and for slides, which can be used for titles, station identification, and commercial advertisements. The projector throws the light image directly onto the image plate of the television film camera. A mirror *triplexer* can be used to enable the image from any one of three projectors to be televised with one film camera. The film camera usually employs the iconoscope or the vidicon camera tube.

A problem in televising standard commercial motion-picture film is the fact that the frame repetition rate for the film is 24 per second, while the scanning procedure in television requires 30 frames and 60 fields per second. The film cannot simply be run through the projector at the rate of 30 frames per second instead of 24, since the increased film speed would make motion in the scene appear unnatural and the sound reproduction from the sound track would be distorted. When motion-picture film is televised, therefore, a special projector is used which allows the film to travel at the speed of 24 frames per second but projects 60 light images of the scene per second, instead of the normal 48. As a result, the time for 60 scanning fields or 30 television frames matches 24 film frames.

*Television Transcriptions.* In order to distribute the same program material to many broadcast stations, or to transcribe a live-talent program to be shown some other time, the picture and sound can be recorded. The picture reproduced on the screen of a monitor kinescope at the studio is photographed on special high-quality 16- or 35-mm motion-picture film, and the audio signal is recorded on the sound track, to transcribe the program on film. This is called a *kinescope recording* or *teletranscription*. The film is run at 24 frames per second, so that the recording can be projected like commercial motion-picture film. As another transcription method, the use of tape recording is being developed to make it possible to record the composite video signal corresponding to the picture, so that the kinescope film recording will not be necessary.

*Camera Chain.* A single chain includes one camera with its control equipment. Figure 6-9 shows two image orthicon cameras, the two camera control units, and associated equipment needed for both camera chains. The portable suitcase-type units are convenient for remote pickups in the field but they can also be used for studio work. Within the camera head are the camera tube, deflection and blanking circuits for the camera tube, and a preamplifier to supply enough camera signal output for the control unit. In addition, an electronic view finder on the head displays the televised image on a small kinescope. The control unit provides remote control of gain, black level, beam current, target voltage, and electrical focus in the camera tube. Optical focus is varied by the cameraman.

*Camera Control, Monitoring, Switching, and Mixing.* The general arrangement for studio operations requires several live-camera chains to provide video signal for different views of the scene; at least one film-camera chain for televising motion-picture films; a central switching system to select the desired signal; the master control with its monitor to check the *on-the-air* video signal and the *video line amplifier* at the studio, which provides video signal with an amplitude of about 5 volts peak to peak for the line from the studio to the transmitter. In addition to

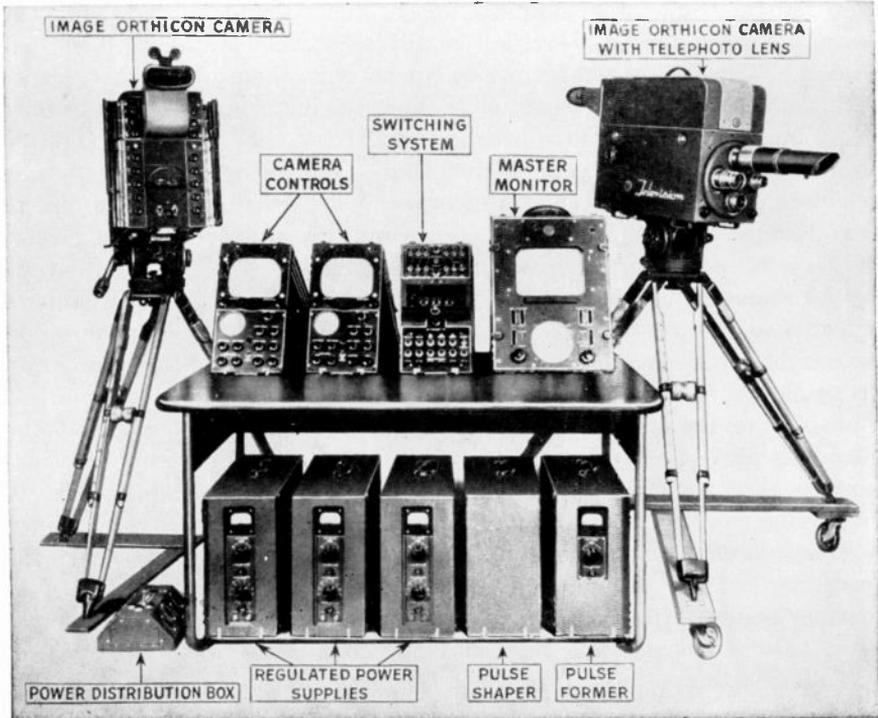


FIG. 6-9. Portable field equipment for two camera chains. (RCA.)

selecting the desired signal, the central equipment includes facilities for distributing sync signals and operating power for all the camera chains. Each camera chain has a control unit, which consists mainly of a monitor and remote-control circuits for the camera. The monitor includes a kinescope that reproduces a high-quality picture from the video signal produced by the camera chain and an oscilloscope to observe signal wave shapes. As the operator views the televised scene and the monitor, the picture quality is monitored by setting the gain, maximum black level, pedestal level, shading if necessary, and other adjustments. Black peaks in the camera signal are generally set 5 to 10 per cent from blanking,

which is called the black *setup interval*, to facilitate clamping at the pedestal level for d-c restoration.

Mixing and switching facilities are provided to switch quickly between cameras, *fade out* smoothly to black just before another camera is switched in, or to mix the pictures from two cameras for a brief period in a *lap-dissolve* transition just before one camera is cut off. Many other optical effects in the reproduced picture can be obtained by camera mixing. Also, brightness effects can be produced by adjusting the pedestal level. Interphone communication is provided for the personnel producing the program, including the camera operator, video control operator, audio console operator, technical director, and program director.

*Synchronizing Signal Generator.* The sync generator consists of two main sections: the *pulse former* that produces the pulses at the correct frequency and the *pulse shaper* to provide the standard waveforms required for the horizontal, vertical, and equalizing pulses. All the pulses are derived from a master oscillator in the pulse-forming unit, which operates at the equalizing pulse frequency of 31,500 cps. Frequency dividers are used to produce 15,750-cps pulses, which is  $\frac{1}{2} \times 31,500$ , and 60-cps pulses in steps of  $\frac{1}{7} \times \frac{1}{5} \times \frac{1}{5} \times \frac{1}{3}$  of 31,500. The master oscillator can be locked in phase with the 60-cps a-c power line or operate independently as a crystal-controlled oscillator. A typical sync generator produces the following output signals:

1. *Kinescope blanking signal.* This consists of the horizontal and vertical blanking pulses used to black out the retraces in the picture tube at the receiver. At the studio, the kinescope blanking signal is coupled to a control amplifier, where the camera signal and blanking pulses are mixed to form the semicomposite video signal.

2. *Synchronizing signal.* This is the sync for the receiver, including equalizing, horizontal, and vertical pulses with standardized timing and wave shapes. At the studio, this synchronizing signal is added to the semicomposite video signal in a control amplifier after the kinescope blanking has been added and the pedestal level set.

3. *Vertical driving signal.* This consists of pulses for synchronizing the vertical deflection generator in each camera at the vertical scanning frequency of 60 cps. These pulses are also used for vertical blanking in the camera.

4. *Horizontal driving signal.* This consists of pulses for synchronizing the horizontal deflection generator in each camera at the horizontal scanning frequency of 15,750 cps. These pulses are also used for horizontal blanking in the camera. The driving pulses are slightly narrower than the kinescope blanking pulses.

5. *Oscilloscope driving signal.* This consists of pulses at half horizontal (7,875 cps) and half vertical (30 cps) frequencies to synchronize the internal saw-tooth generators in the monitor oscilloscopes at the studio.

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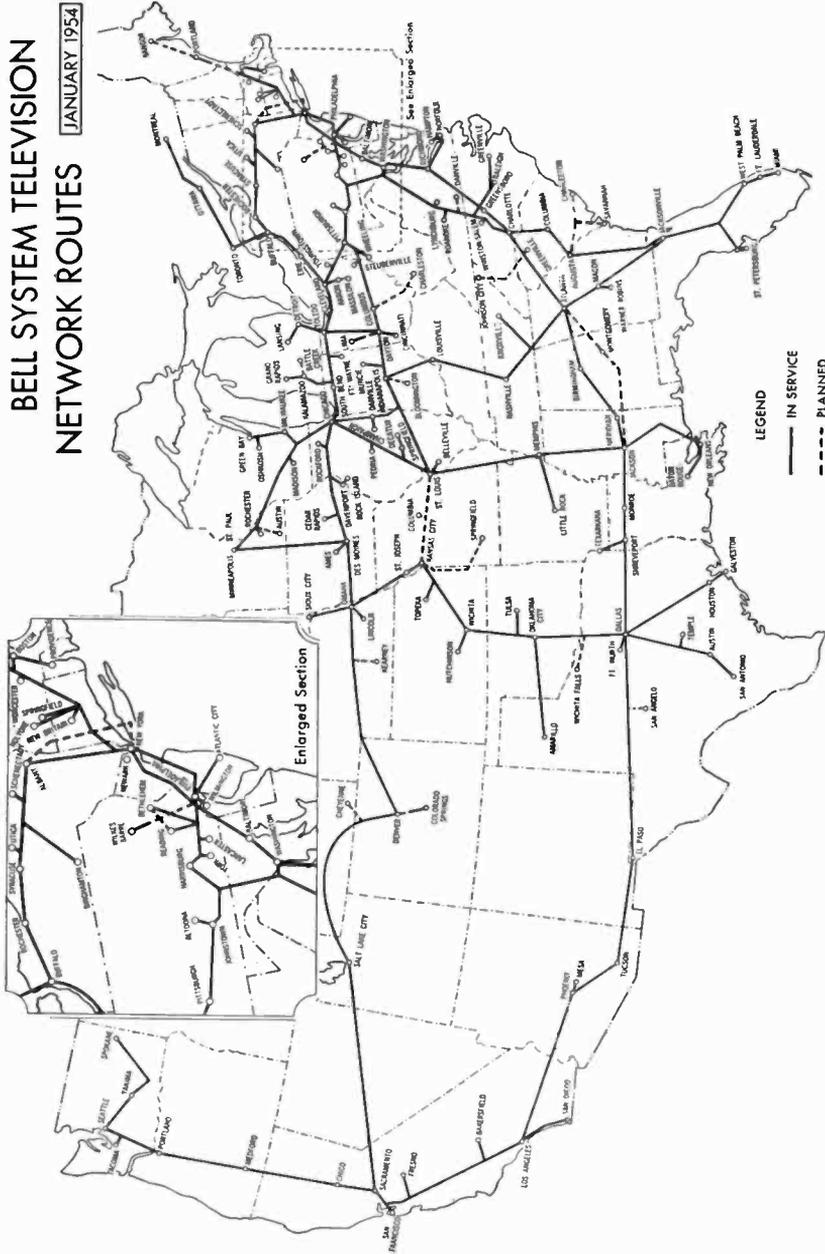


Fig. 6-10. Intercity cable and radio relay facilities for television networks. (American Telephone and Telegraph Company.)

Thus, oscilloscope wave shapes are available that include signals for two horizontal lines, or two fields, locked in to provide a stationary pattern.

*Television Relaying.* To convey the signal for a television program from one location to another, relaying is used for studio-to-transmitter links, remote pickup-to-studio links, and intercity networks. When a program is broadcast over a network, each station in the network receives the program signal by means of intercity relay links and then uses the relayed video signal as a program source to produce the standard AM picture signal broadcast in its assigned channel for the receivers in the area. The map in Fig. 6-10 indicates relay facilities throughout the country for intercity networks. Television relaying is done by cable and radio relays. Cable relaying over long distances is generally accomplished by means of special coaxial conductors in a cable, although ordinary telephone-wire pairs can be used with special terminal equipment for distances of 1 to 2 miles. With cables or telephone line, the composite video signal is relayed. The very wide range of video frequencies that must be conveyed without distortion makes cable relaying of television programs difficult, compared with audio signals, resulting in the common use of radio relaying for television. Radio relays use microwave transmitters and receivers, operating in the range of 7,000 Mc. FM is generally used to transmit the picture signal by radio relay but this is converted to the standard AM picture signal broadcast by the station in its assigned channel. For intercity radio relays, repeater stations including a microwave receiver and transmitter are mounted on high points of the terrain for maximum line-of-sight transmission distance, spaced about 20 to 35 miles apart.

*Television Transmitters.* The block diagram in Fig. 6-11 illustrates a typical transmitter for the picture signal. For frequency stability, a crystal oscillator at about 4 to 8 Mc is used to generate the carrier signal. Frequency-multiplier stages are necessary, therefore, to produce the assigned carrier frequency. The oscillator and multiplier stages form the *carrier generator* section of the transmitter. As noted previously, the assigned picture carrier frequency is 1.25 Mc above the low end of the television station's broadcast channel. However, the exact picture carrier frequencies for different stations on the same channel are offset from each other by  $\pm 10$  kc, in order to reduce the effect of interference between co-channel stations. This system is called *offset carrier operation*. After the desired carrier frequency has been obtained, the signal is coupled to an intermediate power amplifier that drives the final stage, both usually operating straight through without any frequency multiplication.

The modulation follows along the lines of conventional AM transmitters, with the video signal coupled to an r-f amplifier stage to produce amplitude modulation of the transmitted carrier signal. High-level

modulation can be used, the video signal modulating one of the r-f power stages, or low-level modulation can be employed by coupling the video modulating voltage to an r-f amplifier stage operating at a low power level. The d-c component indicating the average brightness of the scene is inserted in the video signal to line up the pedestals in the modulated carrier wave because increased transmitter efficiency is obtained with a constant peak carrier level. The modulation produces double side bands but the modulated picture signal is then coupled to a vestigial-side-band filter, which removes the lower side-band frequencies that are outside the assigned channel. As an alternate method, if the r-f amplifiers are tuned to filter out the undesired side-band frequencies, the vestigial-side-band filter is not necessary.

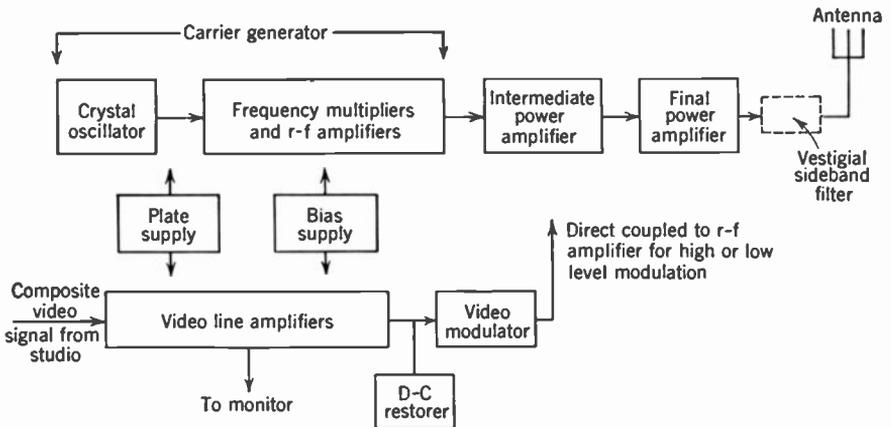


FIG. 6-11. Block diagram of television transmitter for picture signal.

Finally, the modulated picture signal, without the undesired lower side-band frequencies outside the channel, is radiated by the transmitting antenna. The same antenna is generally used for transmitting the broadcast station's AM picture signal and FM sound signal. A *diplexer* unit couples both signals to the common antenna, while isolating the picture signal transmitter from the sound signal transmitter. The FM sound transmitter includes a carrier generator similar to the picture signal transmitter, to produce the assigned sound carrier frequency 4.5 Mc higher than the picture carrier, and audio facilities for frequency modulation of the sound carrier with a maximum frequency swing of  $\pm 25$  kc. The tolerance for the picture or sound carrier frequency is  $\pm 0.002$  per cent. The peak r-f power output of a typical picture or sound signal v-h-f transmitter is 1 to 50 kw. However, the effective radiated power can be higher because it includes the gain of the transmitting antenna. The minimum effective radiated power specified by the FCC for a population of one million or more is 50 kw, with a transmitting-antenna height

of 50 ft above average terrain. For areas with a population under 50,000 the minimum effective radiated power is approximately 1 kw. The radiated power of the sound carrier signal is not less than 50 per cent or more than 150 per cent of the radiated power for the picture carrier signal.

## REVIEW QUESTIONS

1. How is the composite video signal transmitted to the receiver?
2. What is meant by negative transmission of the picture signal? Give one advantage and one disadvantage of negative transmission.
3. In negative transmission what is the amplitude of the modulated picture carrier when the maximum white parts of the image are being scanned? What is the amplitude of the black reference level?
4. Define briefly the method of vestigial-side-band transmission. Why is it used in television broadcasting?
5. A video signal frequency of 3 Mc modulates the picture carrier for channel 2, 54 to 60 Mc. What carrier and side-carrier frequencies are transmitted?
6. Draw a graph showing the frequencies transmitted in channel 14, 470 to 476 Mc, indicating the picture and sound carriers and their frequency separation.
7. Give an example of adjacent channel television broadcast stations and an example of co-channel stations.
8. Why is reflection of the transmitted carrier wave a common problem with the picture signal in television but not in the radio broadcast band of 535 to 1,605 kc? What is the effect in the reproduced picture of multipath signals caused by reflections of the picture carrier wave?
9. List the main equipment units for three camera chains, including two image orthicon cameras for live-talent studio programs and one iconoscope film camera.
10. List at least three technical operations in producing television studio programs.

## CHAPTER 7

### FREQUENCY MODULATION

The sound associated with the picture is transmitted on a separate carrier as a frequency-modulated signal. Frequency modulation has many advantages over the conventional amplitude-modulation system, and there is easily enough space in the 6-Mc television channel for the FM sound signal. For broadcasting the picture signal in the standard television channel, however, AM is used in preference to FM mainly because multipath reception of FM picture signals produces severe distortions in the picture.

**7-1. Modulation.** The process of modulation means that some characteristic of an r-f carrier current, or voltage, is made to vary in step with a modulating voltage that contains the desired information. Taking the audio voltage for the associated sound as an example of the desired signal, the audio voltage itself cannot be transmitted efficiently because its frequency is too low. By the process of modulation the audio voltage is made to vary some characteristic of a higher frequency wave, which can then act as a carrier for the lower frequency signal containing the audio intelligence needed at the receiver for reproduction. The desired information is recovered at the receiver by the detector, which demodulates the carrier signal.

The basic characteristics of any high-frequency voltage or current to be used for the carrier are its amplitude, frequency, and phase. Therefore, the carrier can be modulated by having the modulating voltage vary the amplitude of the carrier, its frequency, or its phase. With the carrier amplitude varying in step with the amplitude of the modulating voltage, the result is amplitude modulation, or AM. When the instantaneous frequency of the transmitted carrier varies with the amplitude of the modulating voltage the result is frequency modulation, or FM. If the instantaneous phase angle of the transmitted carrier is made to vary with the amplitude of the modulating voltage the result will be phase modulation, or PM.

**7-2. An FM Circuit.** Frequency modulation can be accomplished by the *wobbulator* circuit illustrated in Fig. 7-1. The inductance  $L$  and

capacitance  $C$  form the tuned circuit for the Hartley oscillator. The frequency of the oscillator is determined by the resonant frequency of the tuned circuit, being equal to  $1/(2\pi\sqrt{LC})$ . In parallel with the tuned circuit is a variable air condenser, the capacity of which can be varied

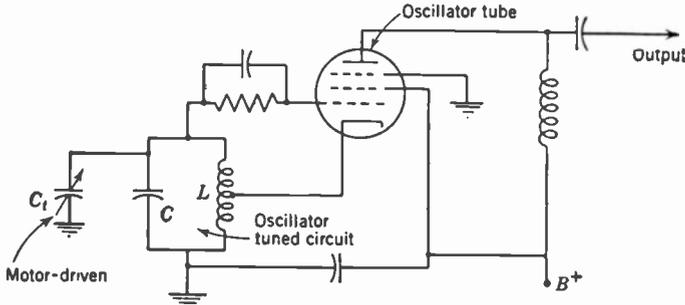


FIG. 7-1. The wobulated oscillator. The shaft of the variable air condenser  $C_1$  is mechanically coupled to a motor to make the condenser rotate in and out of mesh to change the frequency of the oscillator.

by moving the rotor plates in and out with respect to the stator plates. The shaft of this trimmer condenser is driven by a motor to rotate the plates in and out of mesh. Assuming that the frequency of the oscillator is 100 kc with the trimmer condenser halfway in mesh, the oscillator frequency varies above and below this center frequency of 100 kc as the capacitor is driven in and out of mesh. With the trimmer condenser completely in mesh, the added capacity in the tuned circuit is maximum and the output from the oscillator is at its lowest frequency. With the trimmer all the way out of mesh, the frequency of the oscillator is at its highest value. For values of capacity between the two extremes, the oscillator frequency varies continuously between its highest and lowest values around the center value of 100 kc.

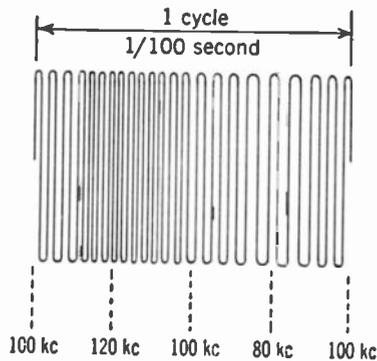


FIG. 7-2. An FM signal. The amplitude is constant, but the instantaneous frequency changes continuously.

If the time for one complete revolution of the trimmer condenser is taken as  $1/100$  sec, the r-f output of the oscillator will appear as in Fig. 7-2. The amplitude remains the same at all times but the frequency is changing continuously, with the instantaneous frequency being *wobbed* around the center carrier frequency. Starting from rest position when the trimmer condenser is half in mesh, the output frequency is at its

center value of 100 kc. As the capacity of the trimmer is reduced by moving the rotor plates more and more out of mesh, the frequency is increased until it reaches its highest value of 120 kc, at the time when the condenser is all the way out of mesh. Now the condenser is brought back to its middle position with the plates halfway in mesh, as at the start, and the oscillator frequency is brought back to its center frequency of 100 kc. For the next half cycle the capacity of the trimmer condenser is greater than its middle value, and the frequency of the oscillator is continuously decreased to its lowest value, 80 kc, before returning to the center frequency of 100 kc. This completes one cycle, which takes  $\frac{1}{100}$  sec. During this time the oscillator frequency was continuously changing from the center frequency of 100 kc, up to its maximum value of 120 kc, then down to 100 kc again, then decreasing further to its minimum value of 80 kc, to return finally to center frequency. The rate at which this cycle is repeated is the frequency of the shaft rotation, which is 100 revolutions or cycles per second in this illustration. The amount that the frequency changes from the center value depends on the amount of capacitance change in the trimmer and is  $\pm 20$  kc in this example. The amount of frequency swing can be made any amount and has no relation to the repetition rate.

This mechanical wobulator method of producing FM has applications in some FM signal generators and automatic frequency-control circuits. However, it is not readily adaptable for modulating the carrier with a voltage of varying amplitude and frequency, such as the audio signal. The same effect can be accomplished with an electronic arrangement, such as a reactance-tube modulator across the oscillator-tuned circuit, instead of the rotating condenser. The audio modulating voltage is applied to the reactance tube to change its reactance at the audio rate which, in turn, varies the oscillator frequency. The basic idea is the same as in the wobulator, with the instantaneous frequency wobbled around the center frequency at the frequency of the audio modulating voltage. The amount that the frequency changes from its center value varies directly with the amplitude of the audio modulation voltage because this decides the magnitude of the frequency change.

**7-3. The FM Signal.** The FM signal is illustrated in Fig. 7-3 for four cases of audio modulating voltage. Assume that 10 peak volts of audio modulation produce a frequency change of 20 kc with a 100-kc carrier. For the 1,000-cycle audio voltage of  $a$  it is shown that the output frequency swings between 80 and 120 kc at a rate of 1,000 cps. The frequency of the transmitted signal is 100 kc when the audio modulating voltage is at its zero value because this is the carrier frequency with no modulation. With modulation, the transmitted frequency continuously varies in value between the values of  $100 \pm 20$  kc. If the frequency

increases for positive values of modulating voltage, it will decrease for negative values. Thus, for the positive half cycles of audio the instantaneous frequency increases from 100 kc to the maximum value of 120 kc, with intermediate frequencies between 100 and 120 kc for values of audio voltage between 0 and +10 volts. For the negative half cycles the output frequency varies between 100 and 80 kc as the audio voltage varies between 0 and -10 volts.

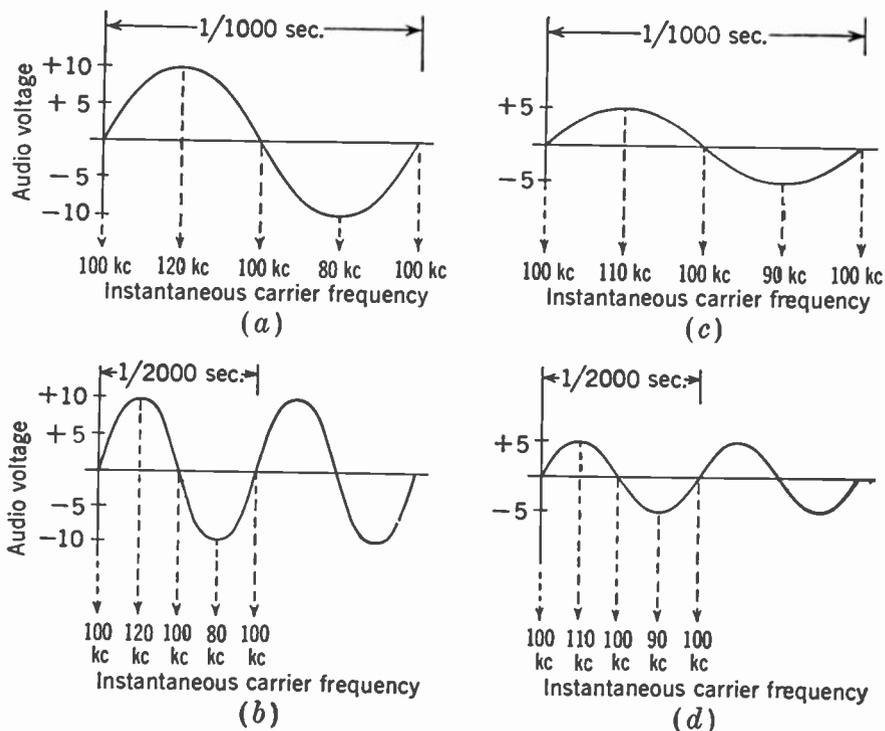


FIG. 7-3. The effect of the audio modulating voltage on the instantaneous frequency of the FM signal.

For the audio voltage shown in *b* the amount of frequency change is the same because the amplitude of audio is the same 10 volts. However, the rate at which the transmitted signal goes through its complete frequency swing is now 2,000 cps because of the 2,000 cps audio modulating voltage. Note that the carrier modulated by 2,000 cps audio voltage goes through two complete cycles of frequency swing while the carrier modulated by 1,000 cycles goes through one complete cycle. A complete cycle of frequency swing is from the center frequency up to maximum, down to the middle frequency, decreasing to the lowest frequency value, and returning to center frequency.

For the audio modulating voltage shown in *c* the maximum frequency change is now only 10 kc instead of 20 kc, because the peak value of the audio is 5 volts instead of 10. The rate at which the output signal swings about 100 kc is 1,000 cps. For the modulating voltage of *d* the maximum frequency change is still 10 kc for 5 volts audio. However, the repetition rate is 2,000 complete swings per second, which is the audio frequency.

The amount of frequency change in the transmitted carrier varies with the amplitude of the audio and should not be confused with the audio frequency. The frequency of the audio modulating voltage is the rate at which the carrier goes through its frequency swings. This determines the pitch of the sound as it is reproduced at the receiver. The amount of audio voltage determines the amount of frequency swing, and this determines the intensity or loudness of the sound reproduced at the receiver. These characteristics of FM are summarized in Table 7-1.

TABLE 7-1. COMPARISON OF FM AND AM SIGNALS

FM	AM
Carrier amplitude constant	Carrier amplitude varies with modulation
Carrier frequency varies with modulation	Carrier frequency constant
Modulating-voltage <i>amplitude change</i> determines amount of <i>frequency change</i> of carrier	Modulating-voltage <i>amplitude change</i> determines amount of <i>amplitude change</i> of carrier
Modulating frequency is rate of frequency change of carrier	Modulating frequency is rate of amplitude change of carrier

**7-4. Definition of FM Terms.<sup>1</sup>** The following terms are used often in connection with FM:

*Center Frequency.* This is the frequency of the transmitted carrier voltage or current that is present when there is no modulation and is the output frequency at the time when the modulating signal voltage is at its zero value. It is also called *rest frequency* since this is the transmitted frequency when the modulating system is at rest.

*Frequency Departure.* This states the instantaneous change of the transmitted signal frequency from the center frequency. For instance, if a transmitted carrier wave having a center frequency of 100 kc is changed to 110 kc by the modulating voltage, the frequency departure is 10 kc. The frequency departure results from modulation, and the amount of frequency change varies with the amplitude of the modulating voltage.

*Frequency Deviation.* The maximum frequency departure from center frequency, at the peak value of the modulating voltage, is the frequency deviation. As an example, an audio modulating voltage having a peak value of 5 volts might produce the frequency departure of 10 kc when the modulating voltage amplitude is 1 volt, 20 kc at 2 volts, and 50 kc for the

<sup>1</sup> Institute of Radio Engineers, Standards on Receivers: Definitions of Terms, 1952.

peak value. The frequency deviation in this case is 50 kc. Instantaneous values of the audio modulating voltage smaller than the peak values produce frequency departures less than the deviation. The amount of frequency deviation depends on the peak amplitude of the audio modulating voltage.

*Frequency Swing.* With equal amounts of frequency change above and below center, the amount of frequency swing is twice the deviation. As an example, when the audio modulating voltage has a peak amplitude on either its positive or negative half cycle of 5 volts to produce a frequency deviation of 50 kc, the frequency swing is  $\pm 50$  or 100 kc. The amount of frequency swing depends on the amplitude of the audio modulating voltage, just like frequency departure and deviation.

*Per Cent Modulation.* This term, as applied to FM, means the ratio of the actual frequency swing caused by modulation to the amount of frequency swing arbitrarily defined as 100 per cent modulation, expressed in per cent. For commercial FM broadcast stations,  $\pm 75$  kc is defined by the FCC as 100 per cent modulation. For the aural or sound transmitter of commercial television broadcast stations, 100 per cent modulation is defined as  $\pm 25$  kc. If, for example, the audio modulating voltage for the associated sound signal in television produces a frequency swing of  $\pm 15$  kc, the per cent modulation is  $1\frac{5}{25}$ , or 60 per cent. Less swing is used for the FM sound in television, compared with FM broadcasting, so that the response of the sound i-f circuits in the receiver can be made broader than the bandwidth of the signal, in order to minimize the problem of tuning in the sound with the picture.

The percentage of modulation varies with the intensity of the audio. For weak audio signals the audio voltage is small and there is little frequency swing with a small per cent modulation. Audio voltage is greater for the louder signals and there is more frequency swing, producing a higher percentage of modulation. The frequency swing produced for the loudest audio signal should be that amount defined as 100 per cent modulation. In FM, 100 per cent modulation is only an arbitrarily defined point, rather than an absolute maximum that cannot be exceeded without distortion, as in an AM system. In addition, the per cent modulation is not an absolute value in FM, but depends on the amount of frequency swing defined as 100 per cent.

*Modulation Index.* The ratio of the amount of frequency deviation to the frequency of the modulating voltage is defined as the modulation index. This is stated as a fraction or a whole number, rather than in per cent. For example, if a 25-kc frequency deviation is produced by a 5,000-cycle audio modulating voltage, the modulation index is  $25,000/5,000$ , or 5. The modulation index is useful in determining the distribution of side bands in the FM signal.

**Deviation Ratio.** This is the ratio of the maximum amount of frequency deviation to the highest audio modulating frequency. As examples, the deviation ratio in the FM broadcast band is 75,000/15,000, or 5; for the FM sound in television it is 25,000/15,000, or 1.67. The deviation ratio is important because it determines the bandwidth requirements of the FM system.

**7-5. Reactance-tube Modulator.** A common method of producing FM directly is the system of varying the frequency of an oscillator by means of a reactance tube, as illustrated in Fig. 7-4. The oscillator can be of any type except crystal-controlled. Similar to the variable trimmer condenser in the mechanical wobulator arrangement of Fig. 7-1, the plate-to-cathode circuit of the reactance tube is in parallel with the

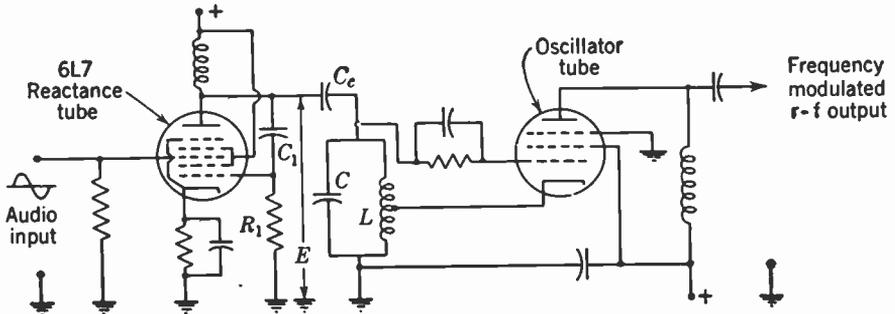


FIG. 7-4. Reactance-tube modulator for producing FM. The quadrature network  $R_1C_1$ , tuned circuit  $LC$ , and plate-to-cathode circuit of the reactance tube are effectively in parallel, with the voltage  $E$  across all three branches, since the coupling condenser  $C_c$  and cathode condenser bypass have no appreciable reactance at the operating frequency.

oscillator-tuned circuit. The basic idea of the reactance-tube modulator is the same as the motor-driven wobulator, with the reactance tube substituted for the reactance of the variable trimmer condenser to vary the oscillator frequency. Since the fundamental characteristic of any reactance is that the current flowing through such a circuit element be  $90^\circ$  out of phase with the voltage across it, this same effect can be achieved in the reactance tube to provide the required reactance. Whatever reactance the plate-to-cathode circuit of the reactance tube may have is in parallel with the oscillator-tuned circuit and affects the frequency of the oscillator.

The reactance tube can be made to appear either as a capacitance or an inductance. When the plate-to-cathode voltage lags the plate current, the tube appears as a capacitance in parallel with the tuned circuit. When the plate-to-cathode voltage leads the plate current, the tube is in effect an inductance. When the amount of inductance or capacitance is changed the frequency of the oscillator is changed. Since the amount of reactance added by the reactance tube depends upon its transconductance, the injected reactance can be made to vary at the audio rate by

applying the audio modulating voltage to the control grid of the reactance tube. In this way, the r-f output of the oscillator is made to vary above and below rest frequency by an amount proportional to the amplitude of the modulating voltage. The repetition rate of the frequency swings is the audio modulating frequency.

The schematic diagram in Fig. 7-4 illustrates the circuit for a reactance-tube modulator. The tube shown is the 6L7 pentagrid mixer, which is convenient because it has two control grids. One grid can be used for the audio modulating voltage and the other for feedback voltage from the plate. The  $R_1C_1$  branch is connected across the plate-to-cathode circuit of the reactance tube to provide the feedback voltage from plate to grid that makes the tube appear as a capacitive reactance. The voltage across  $R_1$  is the feedback voltage and must be in quadrature ( $90^\circ$  out of phase) with the plate-to-cathode voltage. The capacitance  $C_1$  is made small enough so that its reactance at the center frequency is about ten times the resistance of  $R_1$ . This makes the branch circuit of  $R_1$  and  $C_1$  in series with each other a capacitive circuit. The current that flows in this branch, therefore, leads the voltage  $E$  across it by  $90^\circ$ . The voltage across  $R_1$  is of the same phase as the current, leading  $E$  by  $90^\circ$ , because voltage and current are in phase with each other in a resistive element. This voltage across  $R_1$  is coupled to the control grid of the 6L7 tube to produce a component of plate current of the same phase also, since the plate current varies in step with the control-grid voltage. Thus, the plate current through the tube leads the plate-to-cathode voltage  $E$  by  $90^\circ$ , and the tube appears as a capacitive reactance across the oscillator-tuned circuit. To summarize the phase relations, the current and voltage for  $R_1$  lead  $E$  by  $90^\circ$  because  $R_1C_1$  is a capacitive quadrature network. The plate current through the reactance tube also leads  $E$  by  $90^\circ$ , since the plate current varies in phase with the control-grid voltage. Therefore, the plate current leads the plate-to-cathode voltage  $E$  by  $90^\circ$  and the reactance tube is capacitive.

Assume now that the tube is operating with a fixed amount of bias, and with no audio signal voltage coupled to the control grid. There will be a certain amount of plate current as determined by the transconductance of the tube,  $g_m$ , which depends on the bias. The plate current is of the same phase as the feedback voltage across  $R_1$ , and both lead the plate voltage by  $90^\circ$ . This makes the reactance tube appear as a capacitive reactance across the tuned circuit, and the oscillator operates at a center frequency—determined by the  $L$  and  $C$  of the tuned circuit and the parallel capacitive reactance of the reactance tube. When audio modulating voltage is applied to the control grid, the bias and the  $g_m$  of the tube are varied at the audio rate. For the positive half cycle of audio the bias is made less negative and the transconductance of the reactance tube increases. For the negative half cycle the bias is made more negative and

the transconductance decreases. Therefore, the effect of varying the  $g_m$  of the tube is to vary the amount of quadrature plate current through the tube.

If the amount of quadrature current through the reactance tube changes, its reactance must be different. This new value of reactance in parallel with the tuned circuit changes the output frequency of the oscillator. Physically, this apparent reactance may be regarded as the reactance tube's effect of injecting more current into either the inductive or the capacitive branch of the tuned circuit, which oscillates with maximum

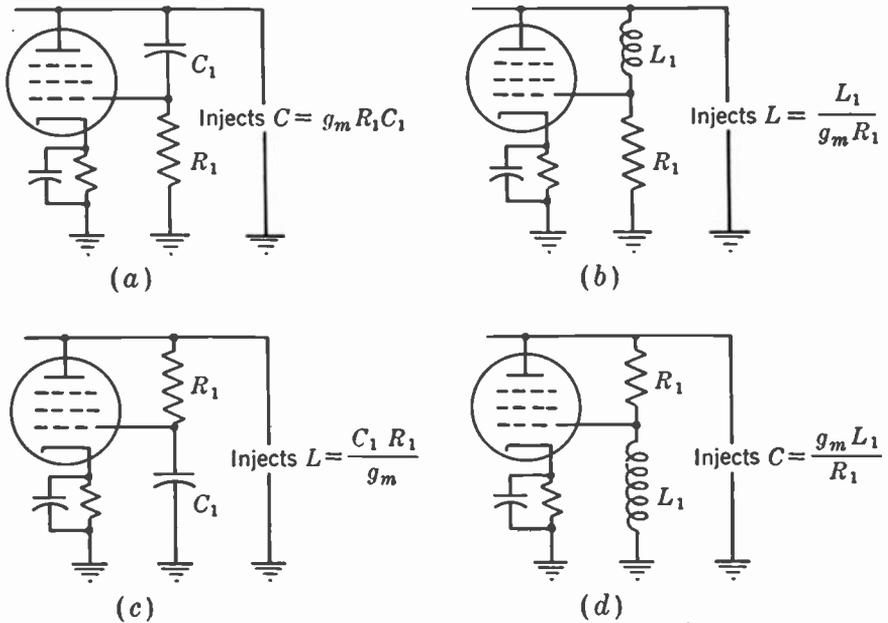


FIG. 7-5. Quadrature networks for a reactance tube.

amplitude at the frequency for which the inductive-branch current equals the capacitive-branch current. If the reactance tube's plate current leads the plate voltage, which is of the same phase as the parallel voltage across the oscillator tank circuit, more current will flow in the capacitive leg because the capacitive-branch current is also leading the tank voltage. As a result, the tuned circuit will oscillate with maximum amplitude at a new lower frequency that makes the currents in the capacitive and inductive legs equal. For a smaller amount of leading current in the parallel reactance tube the current in the capacitive leg of the tuned circuit decreases, and the tuned circuit oscillates at a higher frequency to equalize the currents in the capacitive and inductive branches.

Four possible quadrature networks are shown in Fig. 7-5. In all cases the impedance of the grid-to-plate element must be at least ten times the

impedance of the grid-to-cathode element, to make the grid feedback voltage approximately  $90^\circ$  out of phase with the plate-to-cathode voltage and thus obtain a quadrature component of plate current. Because such a quadrature network is always associated with the reactance tube, it is sometimes called a *quadrature tube*. The amount of reactance injected is in parallel with the tuned circuit of the oscillator, thus changing its frequency. As the audio modulating voltage varies the transconductance of the reactance tube, the amount of injected reactance changes and the frequency of the oscillator departs from center frequency in step with the modulating voltage. The amount of frequency departure depends on the amount of change in  $g_m$ , which varies with the amplitude of the audio modulating voltage, and the repetition rate of the frequency swings is the same as the audio modulating frequency. It should be noted that the

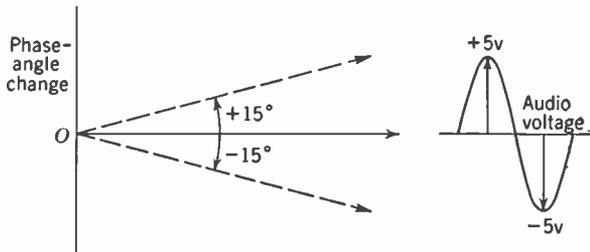


Fig. 7-6. In phase modulation the amount of phase-angle swing varies with the audio amplitude.

reactance tube need not have two control grids, since the quadrature feedback voltage and the modulating voltage can be coupled to the control grid and cathode.

The reactance-tube modulator system provides a method of obtaining FM directly with relatively wide frequency swings. A basic characteristic of this method, however, is that the average frequency of the oscillator tends to drift. A crystal-controlled oscillator for stabilizing the center frequency cannot be used in parallel with the reactance tube because the frequency stability of the crystal is so great that it would keep the oscillator frequency from changing enough to produce any appreciable amount of direct FM.

**7-6. Phase Modulation.** The characteristic of phase modulation compared to a direct FM system such as the reactance tube is that a crystal-controlled oscillator can be used for excellent stability of the center frequency. Although the oscillator is crystal-controlled, its output can be varied in phase. The basic principle is the same as in FM or AM, in that some characteristic of the carrier is made to vary in step with the audio modulation. In PM the phase angle of the transmitted carrier is shifted with respect to its phase at rest by an amount proportional to the amplitude of the audio modulating voltage. This is illustrated in Fig. 7-6.

At zero audio voltage there is no change in phase. The zero angle represents the phase of the carrier at rest, without modulation, which is the phase-angle reference. At the positive peak amplitude of the audio voltage, there is the maximum change of phase angle in one direction. The negative audio voltage peak produces maximum change of phase angle in the opposite direction. As a result, the amount that the phase angle changes from rest varies with the amplitude of the audio modulating voltage, and the repetition rate of the phase-angle swings is the audio frequency.

*Equivalent FM.* The continuous change in phase angle is equivalent to a change in frequency because there can be no change in one without a change in the other. A wave with a given phase angle is fixed so long as the phase angle is constant. Anytime there is a change in phase, though, it is equivalent to a change in frequency. This is illustrated in Fig. 7-7, in which the first

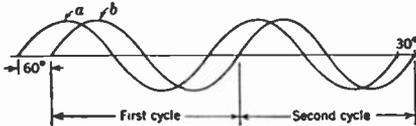


FIG. 7-7. A change in phase is equivalent to a change in frequency.

cycle is drawn for a phase angle of  $60^\circ$  between the waves *a* and *b* and the second cycle for an angle of  $30^\circ$ . The change in phase angle of *b* with respect to *a* corresponds to an equivalent change in frequency for wave *b*, since the time for a complete cycle has changed. Therefore, the result of PM is an equivalent FM signal.

The equivalent FM produced as a result of PM is very similar to direct FM, such as the output of a reactance-tube modulator, in that the frequency of the transmitted signal departs from the center frequency in step with the audio modulating voltage. The amount of frequency departure due to PM is  $f \times \Delta\theta$  where  $f$  is the audio modulating frequency and is the frequency of the phase swings. The change in phase angle  $\Delta\theta$  is the amount of phase departure, in radians. One radian is equal to  $57.3^\circ$ . As a numerical example of the amount of equivalent FM produced by a change in phase, assume that the phase departure of the r-f carrier wave is  $30^\circ$ , as determined by the amplitude of the audio modulating voltage. This phase-angle departure is  $30/57.3$ , or 0.52, radian. The repetition rate of the phase swings is assumed as 1,000 cps, which is the audio modulating frequency. The amount of frequency departure due to PM in this example, therefore, is  $1,000 \text{ cps} \times 0.52$ , or 520 cps.

An important distinction between PM and direct FM, however, is that in FM the amount of frequency departure is proportional only to the amplitude of the modulating voltage, whereas in PM the equivalent FM departure is proportional not only to the amplitude but also to the fre-

quency of the audio modulating voltage. For instance, if the frequency departure in FM is 10 kc for a given amplitude of the audio signal, the departure is 10 kc for every other audio frequency having the same amplitude. In PM, though, the equivalent frequency departure with 10,000-cycle audio modulation would be 10 times greater than at 1,000 cycles if the audio amplitude and the amount of phase-angle departure were the same for the two signals.

*Predistortion.* In order to compensate for the great inequality in per cent modulation for audio signals having the same intensity in PM it is common to insert a predistortion network between the phase modulator and the source of audio modulating voltage. An *RC* or *RL* circuit can be used as an *audio corrector* to provide for the phase modulator an audio voltage that is inversely proportional to frequency. This inverse frequency effect cancels out the frequency factor that makes the amount of equivalent frequency departure proportional to the modulating frequency in PM. The result is an FM signal whose frequency departure depends only on the amount of phase departure which, in turn, is proportional only to the amplitude of the audio modulating voltage. For example, if 1,000- and 10,000-cps signals of equal amplitude are to be used for modulating the carrier, the predistortion network will provide for the modulating grid of the phase modulator an audio voltage of 0.1 volt for the 10,000-cycle signal if its output is 1 volt for the 1,000-cycle audio. Thereafter the two signals are on equal terms so far as their effectiveness in producing frequency departure is concerned.

The amount of frequency departure still varies with the intensity of the audio modulating signal but, with audio correction, signals of all audio frequencies are equally effective in determining the frequency departure of the transmitted signal. Therefore, the PM refers only to the modulation method, since with predistortion of the audio modulating voltage the transmitted signal is FM. Many FM transmitters use a PM modulator, instead of a direct FM system such as the reactance-tube modulator. Phase modulation of a crystal-controlled oscillator allows greater stability of the center frequency, but the amount of equivalent FM produced is relatively small, requiring the use of frequency-multiplier stages to increase the frequency swing.

**7-7. Side Bands in FM.** The FM signal, like an AM signal, can be considered in terms of its side-band frequencies. The side-carrier frequencies are symmetrical above and below the carrier, and differ from the carrier by the modulating frequency, just as in the AM signal. In AM the upper and lower side-carrier frequencies will be 110 and 90 kc, respectively, if a 10,000-cps audio voltage modulates a 100-kc carrier as described previously in Sec. 6-2. However, while there is only one pair of side bands possible in AM, there may be higher orders of side-band frequencies

in FM. For the same modulation example, the side-band frequency distribution around the carrier contains lower side frequencies of 70, 80, and 90 kc, with upper frequencies of 110, 120, and 130 kc if the second- and third-order side bands have appreciable amplitude also, as they can in FM. The side-band frequencies, whether one pair or more, and whether AM, FM, or PM, are often called the *modulation product*, since they are due to modulation. Without modulation the output consists only of the unmodulated carrier. With modulation the transmitted signal includes the carrier plus the side bands or modulation product.

The number of symmetrical pairs of side-band frequencies distributed around the center carrier frequency in the FM signal depends on the modulation index and can be any amount from one pair as in AM to a few hundred. The greater the modulation index, the greater the number of side-band frequencies having appreciable amplitude. Only the first pair of side-band frequencies has appreciable amplitude with a modulation index equal to 0.5 or less. In this case the transmitted signal can be considered as the resultant of the carrier plus one upper side band and one lower side band, each as in AM differing from the carrier by the audio modulating frequency. In AM, however, the modulation product and carrier are of the same phase, while the modulation product in FM is out of phase with respect to the carrier by  $90^\circ$ . This is the basis of the Armstrong FM transmitter.<sup>1</sup>

**7-8. Requirements of the FM Receiver.** The FM receiver is a superheterodyne like a typical AM receiver, with two additional requirements:

1. The second detector in the superheterodyne FM receiver must be an FM detector, which can convert the frequency variations of the FM carrier into the corresponding voltage variations of the audio signal. Two typical circuits often used for the FM detector are the *ratio detector* and *discriminator*. Both circuits can detect variations in frequency. In general, detection of an FM signal depends upon the fact that a tuned LC circuit has different amounts of output for different input frequencies.

2. A circuit that can reject amplitude variations of the FM carrier signal is needed in the FM receiver. Since only the frequency changes in the transmitted carrier correspond to the modulating voltage, any variations of the carrier amplitude can be eliminated in an FM system without losing the desired signal. Therefore, FM provides an important method of distinguishing between the desired signal which is frequency-modulated, and undesired interfering signals that mainly cause amplitude variations of the carrier. This is probably the most important advantage inherent in an FM system for transmitting intelligence. In order to utilize this advantage of FM, the receiver must prevent amplitude variations of the

<sup>1</sup>Armstrong, E. H., Method of Reducing Disturbances in Radio Signaling by a System of Frequency Modulation, *Proc. IRE*, May, 1936.

FM signal from having an effect in the audio output. This can be accomplished by a limiter stage, or by using an FM detector that is insensitive to AM. The ratio detector is an FM detector circuit relatively immune to amplitude modulation in the FM signal.

A limiter is a class C amplifier with grid-leak bias and reduced plate and screen voltages, so that the amplifier saturates at a relatively low level of input voltage. For a wide range of input voltages that cause saturation, therefore, the output voltage from the limiter has the same amplitude. As a result, the limiter eliminates the AM because the amplitude variations of the input voltage are missing from the output signal.

**7-9. Reduction of Interference in FM.** The greatest advantage of FM is its ability to eliminate the effects of interference from the desired signal. The interference can be a modulated carrier from another FM or AM station, atmospheric or man-made static, or receiver noise. In any case, the effects of the interference on the desired signal can be made negligible in an FM system. This important advantage is an inherent part of the FM system because the instantaneous frequency variations of the modulated carrier correspond to the desired signal, while the dominant effect of an interfering signal is to change the amplitude of the carrier. Thus, an FM system immediately offers increased possibilities in separating interference from the desired signal by AM rejection in the FM receiver. In an AM system not much can be done with interference that changes the amplitude of the carrier because any attempt to limit the carrier amplitude while trying to eliminate the interference has a harmful effect on the desired signal, which is an amplitude variation of the carrier.

**AM Interference.** Whenever there are two signal frequencies in the receiver, their resultant can be amplitude-modulated because of addition of the in-phase components of the two signals. If one is the desired signal, either amplitude- or frequency-modulated, its carrier amplitude can be changed in many ways that produce interference. For two carriers of the same frequency and phase, the resultant envelope is the sum of the individual envelopes. Amplitude variations corresponding to audio beat frequencies can be introduced when the interference is of slightly different frequency; or abrupt amplitude changes corresponding to static can be added to the desired signal. In any case, the interference produces undesired AM at an audio rate that can be heard in the receiver output. The interfering effect is noticeable when the interference has an amplitude that is only 1 per cent of the desired signal.

This interfering effect is minimized in the AM system by making the amount of amplitude modulation as high as possible for the desired signal and using as much transmitter output power as possible. The extent to which this can be carried out has definite limitations, however. The per cent modulation in AM cannot exceed 100 per cent without distortion,

and this can be used only for the loudest signals. In addition, the average output power of the transmitter cannot be increased indefinitely because there is a limit above which the power cannot be increased economically.

In the FM system the transmitted signal can have its amplitude changed in the same way because of interference. Here, however, any change in amplitude is eliminated by AM rejection circuits in the receiver without losing the desired signal, which is the instantaneous frequency departure of the modulated carrier. Therefore, the use of circuits that make the receiver insensitive to AM is a most important factor in obtaining the great improvement in noise reduction for FM over AM.

*FM Interference.* The interfering signal can be varying in phase as well as amplitude, adding a varying phase shift and equivalent FM interference to the desired carrier. Thus, an FM component is added to the desired signal to produce interference that cannot be eliminated, because it is the

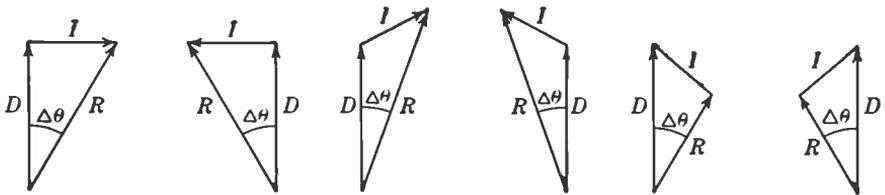


Fig. 7-8. Vector diagram illustrating the effect of PM interference.

same type of variation of the transmitted carrier as the desired signal. This parallels the case of AM interference which could not be eliminated in the AM system without disturbing the desired signal. In the FM system, however, the amount of FM interference can more easily be made negligible in comparison to the desired signal.

When two signals of varying phase or frequency are added, the total phase modulation is not the sum of the individual modulations. This is illustrated in Fig. 7-8, where the phase of an interfering PM signal is represented by  $I$ , the desired carrier by  $D$ , and the resultant sum of the two vectors is  $R$ . If the amplitude of  $I$  is less than  $D$  the change in phase angle  $\Delta\theta$  from  $D$  to  $R$  cannot exceed the angle whose tangent is  $I/D$ . For the case of a desired carrier  $D$  whose amplitude is twice the interfering signal amplitude  $I$  the amount of phase modulation added to the desired signal cannot exceed the angle whose tangent is  $1/2$ . This angle is  $27^\circ$ , or approximately 0.47 radian. The resultant PM is equivalent to an interfering FM signal, with the amount of swing greater for higher audio modulating frequencies, as in any PM system without predistortion.

The amount of FM added by the interfering signal can be made negligible in the audio output by having the desired signal produce much greater frequency swings in an FM system. In FM the desired percentage of modulation can be increased to swamp out the interference without any

increase in the transmitter power and without any distortion. If the per cent modulation of the desired signal is made large in comparison to the limited amount of FM interference the effects of the interfering signal become negligible, even though the magnitude of the interfering wave is by no means negligible in comparison to the desired carrier amplitude.

**7-10. Preemphasis and Deemphasis.** Any PM interference produces more equivalent FM for higher audio frequencies. Also, the desired signals in the higher audio-frequency range are usually of relatively low amplitude because they are harmonics of the fundamental tones and produce little frequency swing. Therefore, it is desirable in an FM system to preemphasize the higher audio modulation frequencies at the transmitter. This is done by increasing the relative amplitudes of the higher audio frequencies before modulation, so that they can produce more frequency swing in the transmitted signal and have a higher signal-to-noise ratio. In order to restore the original relative amplitudes, the audio signal should be deemphasized at the receiver to the same extent that it is preemphasized at the transmitter. At the receiver, the deemphasis network can be an RC filter in the output circuit of the FM detector, or the audio amplifier response can deemphasize the higher audio frequencies. The preemphasis and deemphasis networks are RC or LR filter circuits for the desired frequency range. For the transmitter, the FCC standards specify that preemphasis shall be employed in accordance with the impedance-frequency characteristic of an inductance-resistance network having a time constant of  $75 \mu\text{sec}$ . At the receiver the deemphasis should have the same time constant of  $75 \mu\text{sec}$ .

While it would seem that no progress is made if the audio voltage is deemphasized to the same extent that it is preemphasized, a great improvement in signal-to-noise ratio actually is accomplished. The reason is that the preemphasis precedes the effect of the interference, so that when the signal and noise are both reduced by the deemphasis the signal returns to normal while the noise is reduced below normal. This is more effective in FM than in AM because the noise level in FM increases for higher audio frequencies and the deemphasis attenuates the highest audio frequencies the most.

**7-11. Transmitter Efficiency.** Another fundamental difference between FM and AM affects the transmitting equipment to a great extent. In AM the peak carrier power varies up to four times the unmodulated carrier level when the peak antenna current doubles its unmodulated value. This requires that the transmitter be able to handle during modulation peaks a great deal more than its rated power output without introducing any distortion. Also, high-level modulation is often used in the AM transmitter with the audio voltage modulating the final r-f amplifier, and this requires a great deal of audio power. Low-level modulation can be

used in AM, with the audio voltage modulating a low-power r-f stage, but this requires that all the succeeding stages be linear amplifiers in order to amplify the modulated carrier without distortion, thus resulting in a great loss in efficiency as compared to class C amplification.

In the FM transmitter the carrier power remains the same during modulation because the amplitude of the carrier remains constant. This requires no reserve power in the transmitter for modulation peaks. Low-level modulation is used, requiring very little audio power. The power r-f stages, including the final, can still operate class C because the carrier amplitude remains constant during modulation. As a result, an FM station can transmit the same power as an AM station with about one-half the equipment and power requirements. Another advantage is that additions can be made to an FM transmitter just by adding the r-f power stages required for the desired amount of output power.

**7-12. FM vs. AM.** FM has many advantages inherent in the method of modulation itself. In actual practice, however, other factors may make the use of FM more or less desirable than AM for a given service. In evaluating the relative merits of the two methods, two comparisons can be made. One is a comparison of FM to AM with both in the same high-frequency band; the other is between the FM broadcast service in the v-h-f band and AM in the relatively low-frequency standard broadcast band of 535 to 1,605 kc.

In comparing the standard broadcast band and the v-h-f band the lower frequencies offer the advantage of a greater transmitting distance. A clear-channel AM standard broadcast station having no other interfering station on its channel has the advantage of being able to provide service to much larger areas than could be served by an FM station operating in the v-h-f band. Any FM service with wide swings must be in the v-h-f spectrum or above in order to allow room for the wider channel required, which is 200 kc for FM broadcast stations in the 88- to 108-Mc band. For the FM sound in television, 50 kc of the 6-Mc television channel is used. The v-h-f band offers the advantage of reduced atmospheric static, whether FM or AM is used. Use of the v-h-f band, though, introduces the disadvantages of reduced transmitting distance and increased interference from man-made noise, especially car-ignition noise. FM can provide a greater degree of improvement with respect to noise at these higher frequencies.

*Interference.* FM has several advantages over AM in regard to atmospheric static, interfering channels, and man-made noise such as that from diathermy machines, electrical machinery, and automobile ignition systems. The effects of atmospheric static are very much less on the very high frequencies and the use of FM makes such interference negligible. The most serious interference in the v-h-f band is car-ignition noise.

Although ignition noise is not completely eliminated by the use of FM it can be reduced to a great extent, the exact amount depending on the design of the FM receiver—especially the AM rejection circuits. In field-test comparisons, FM usually provides much more freedom from ignition-type noise than AM.

*Receiver Noise.* Receiver noise, which takes the form of a continuous hissing sound, is not usually noticeable in receivers for the AM broadcast stations in the standard broadcast band; with the high field strengths usually encountered, the gain of the receiver need not be great enough to bring the receiver noise up to the point where it can be heard. On the very high frequencies, though, receiver noise is a limiting factor in the reception of intelligible signals because of the reduced signal input to the receiver at these frequencies. With FM, receiver noise is suppressed to a great extent by the use of AM rejection circuits, balanced discriminator, and deemphasis. FM has the advantage over AM in being able to provide useful signal output from the receiver with very much less input signal than what is required for an AM receiver to overcome receiver noise.

*Service Area.* Although AM in the standard broadcast band can provide much greater transmission distances on a clear channel, FM in the v-h-f band can supply a greater service area for broadcasting because of its freedom from interference between stations in the same channel or adjacent channels. In FM the receiver has the ability to suppress completely an interfering station on the same channel if the desired signal is stronger than the other by a minimum ratio. This is called the *capture effect*. It results from the interference-reducing properties inherent in FM. While an amplitude ratio of 2:1 for the desired signal over the undesired signal will prevent the interfering station from coming through, the ratio required for complete suppression with no interfering noises may be larger, depending on the frequency swing, the type of modulating intelligence on the two signals, and their channel separation. For two stations with  $\pm 75$ -kc swings operating on the same channel there is complete suppression of the undesired signal for the worst possible case when the ratio of desired to undesired signal is 10:1 or greater. For adjacent channels a 2:1 ratio is sufficient to suppress the interfering station. In AM the desired signal would have to be very much larger than the undesired signal for similar results.

Because of this advantage of suppressing unwanted signals in an FM system, FM broadcast stations can be located much closer together geographically than AM stations and many more station assignments can be made for each channel without danger of interference. The FM broadcast station may have a larger useful service area than an AM station on the standard broadcast band suffering from shared-channel interference, even though the AM station can transmit over greater distances.

*Transmitter Efficiency.* In FM the carrier level is constant, and low-level modulation can be used with the succeeding r-f stages operating class C. This results in a smaller, more economical, more efficient, and more adaptable transmitter for FM.

*Audio-frequency Range.* In general, the use of the v-h-f band allows a wider audio modulating frequency range, because at the higher frequencies a wider channel is feasible for accommodating the resultant side-band frequencies. This is true of either FM or AM. An AM station in the standard broadcast band is limited to approximately 5 to 10 kc as the highest audio modulating frequency because of the restricted width of the transmission channel. Given a wide enough channel, though, an AM

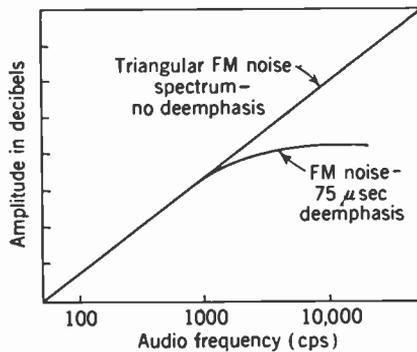


Fig. 7-9. Effect of deemphasis on FM noise.

system can use as wide a modulating frequency range as FM. The picture carrier in television, for example, is amplitude-modulated with video frequencies as high as 4 Mc.

The audio modulating frequency range is 50 to 15,000 cps for commercial FM broadcast stations and the FM sound in television. However, the extent to which this greater audio-frequency range is made useful depends on the quality of the audio system in the receiver.

*Preemphasis.* This is more advantageous in an FM system because the deemphasis in the FM receiver has an effect on noise suppression that is greater than in AM. Random noise can be considered as a continuous spectrum of signals of random phase and frequency, all of the same amplitude. In AM the detected signal contains audio noise voltages of varying frequencies but constant amplitude. In FM, however, audio output is proportional to frequency swing rather than amplitude, and the noise signals with the most swing produce the most audio output. The amplitude of the detected noise increases with audio frequency, therefore, producing the *triangular noise spectrum* of FM illustrated in Fig. 7-9. With deemphasis in the FM receiver the amplitude of the highest audio fre-

quencies is reduced the most, thus reducing the most objectionable components of the noise by the greatest amount.

*Multipath Reception.* Because v-h-f and u-h-f carrier waves are reflected from buildings, bridges, and similar obstructions, the radiated signal can arrive at the receiving antenna over multiple paths from the transmitter. The different paths are usually not the same length, with the result that the multipath signals take different amounts of time to travel the separate transmission paths to the receiver. With multipath reception of signals slightly displaced in time with respect to one another, the distortion that can result due to interference between the multipath signals is more objectionable in FM than in AM because the frequency of the FM signal is continuously changing. Since the instantaneous frequency of an FM signal varies with time, the multipath signals at the receiver generally will have different frequencies at any instant. As a result, heterodyning action between the FM multipath signals at the receiver produces interfering beats that continuously change in frequency. The changing beat frequency can produce garbled sound, similar to the effect produced by nonlinear amplitude distortion in an audio amplifier. In picture reproduction, the interfering FM beat would produce a bar interference pattern in the image with a shimmering effect, as the bars continuously change with the beat frequency. This is why AM is preferable to FM for broadcasting the picture signal, as multipath AM signals simply produce multiple ghost images.

FM is generally used for transmitting the picture and sound signals between radio relay stations, which receive and retransmit programs between stations in a network. In this service, however, multipath reception is not a problem because the relay stations operate in high-frequency microwave bands and can use very directive antennas to beam the signal from one relay transmitter directly to the receiver at the next station.

#### REVIEW QUESTIONS

1. State the manner in which the carrier varies with the amplitude and frequency of the audio modulating voltage in AM, FM, and PM.
2. What property of the FM signal determines the loudness of the reproduced audio signal? What determines the frequency of the reproduced audio?
3. A 10-volt, 100-cycle audio modulating voltage produces a frequency departure of 50 kc. What is the departure for a 10-volt, 500-cycle audio modulating voltage? for a 2-volt, 500-cycle modulating voltage? Assume that the modulation is linear.
4. A 3-volt, 1,000-cycle audio modulating voltage produces a frequency departure of 5 kc. What is the modulation index?
5. A 60-Mc carrier is modulated by 15,000-cycle audio voltage, producing first- and second-order side bands with appreciable amplitude. What are the side-band frequencies produced, in addition to the carrier?
6. Describe briefly the principle of a reactance-tube modulator using a quadrature network that makes the reactance tube appear inductive. What is the effect on

oscillator frequency when the control grid of the inductive reactance tube is made more positive by the modulating voltage?

7. What is one advantage and one disadvantage of a phase modulator?
8. Why does PM produce equivalent FM?
9. How does the equivalent FM of PM differ from FM produced directly by a reactance-tube modulator? What is done to correct this?
10. In an FM receiver, what are the functions of the limiter and discriminator stages? The ratio detector?
11. Give two reasons for the improved noise reduction in an FM system.
12. Give two advantages of FM over AM and one disadvantage of FM, with both in the v-h-f band.
13. Why is preemphasis used in FM?
14. If a 4-Mc modulating voltage produces a frequency swing of  $\pm 100$  kc, what is the modulation index? What will be the bandwidth required for the FM signal, using double-side-band transmission?

## CHAPTER 8

### TELEVISION RECEIVERS

The television receiver has the function of receiving the transmitted picture carrier signal and associated sound signal, amplifying the signals and providing detection to permit reproduction of the picture information and the sound. In effect, two receivers are included: an AM receiver for the picture signal and an FM receiver for the associated sound. The

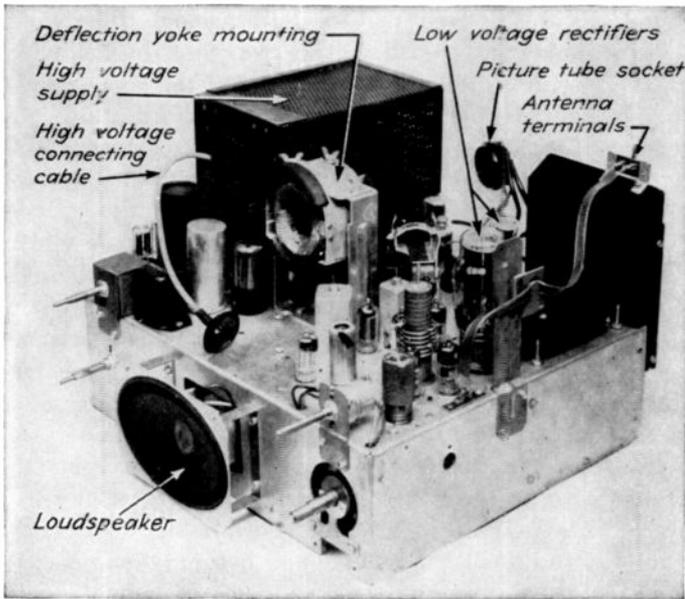


FIG. 8-1. Television receiver chassis. (RCA.)

receiver must also provide scanning and synchronization for reproduction of the image on the screen of the picture tube. Usually 15 to 30 stages are needed in the television receiver, therefore, for the picture and sound. Figure 8-1 shows a television receiver chassis.

**8-1. Forming the Image.** In order to reproduce the desired picture information at the receiver a spot of light corresponding to a basic picture

element must be formed. This is accomplished by the electron gun of the picture tube, which provides a narrow beam of electrons to strike the fluorescent screen of the tube and produce a spot of light. In addition, the spot of light must be displaced along a series of horizontal lines to form the scanning pattern. Therefore, the electron beam is deflected by providing suitable scanning current for the horizontal and vertical scanning coils in the deflection yoke mounted on the neck of the picture tube. The required scanning current for the deflection yoke is produced by the deflection circuits in the receiver. Horizontal deflection circuits are necessary for the horizontal scanning motion, while the vertical deflection circuits are needed for vertical scanning.

With a complete scanning pattern produced, it is finally necessary to vary the intensity of the light spot on the screen of the picture tube so that its brightness at any point in the scanning pattern corresponds to the brightness of the corresponding point of the scanned image in the camera tube at the transmitter. This intensity variation of the light spot is the function of the video signal, which contains the amplitude variations corresponding to the desired picture information. The video signal varies the kinescope control-grid voltage, varying the amount of beam current in the picture tube and the intensity of the spot of light on the screen, thereby reproducing the desired picture information.

**8-2. Receiver Circuits.** Figure 8-2 shows a block diagram of the circuit arrangement in a television receiver. The superheterodyne circuit is used.

*The R-F Section.* Starting at the antenna, the picture and sound r-f carrier signals are intercepted by a common receiving antenna for both signals. A transmission line connects the antenna to the input terminals of the receiver, coupling the r-f picture and sound signals to the r-f amplifier stage. The r-f circuits are tuned to the channel frequencies of the desired station and are broad enough to pass the picture and sound carrier signals, with their side-band frequencies. Although the r-f amplifier is optional, most television receivers use one r-f stage to amplify the r-f signals before coupling to the mixer or converter stage. The r-f output of the receiver's local oscillator is also coupled into the mixer, as shown in Fig. 8-2, so that the oscillator voltage can heterodyne or beat with the incoming r-f picture and sound carrier signals. When the oscillator frequency is set for the channel to be tuned in, the carrier signals of the selected station are heterodyned to the lower intermediate frequencies of the receiver. Two i-f carrier signals are produced by the mixer stage. One is the picture i-f signal corresponding to the r-f picture signal and the other is the sound i-f signal corresponding to the r-f sound signal. The original modulating information of the r-f carrier signals is present in the picture and sound i-f signals in the converter output, and the 4.5-Mc

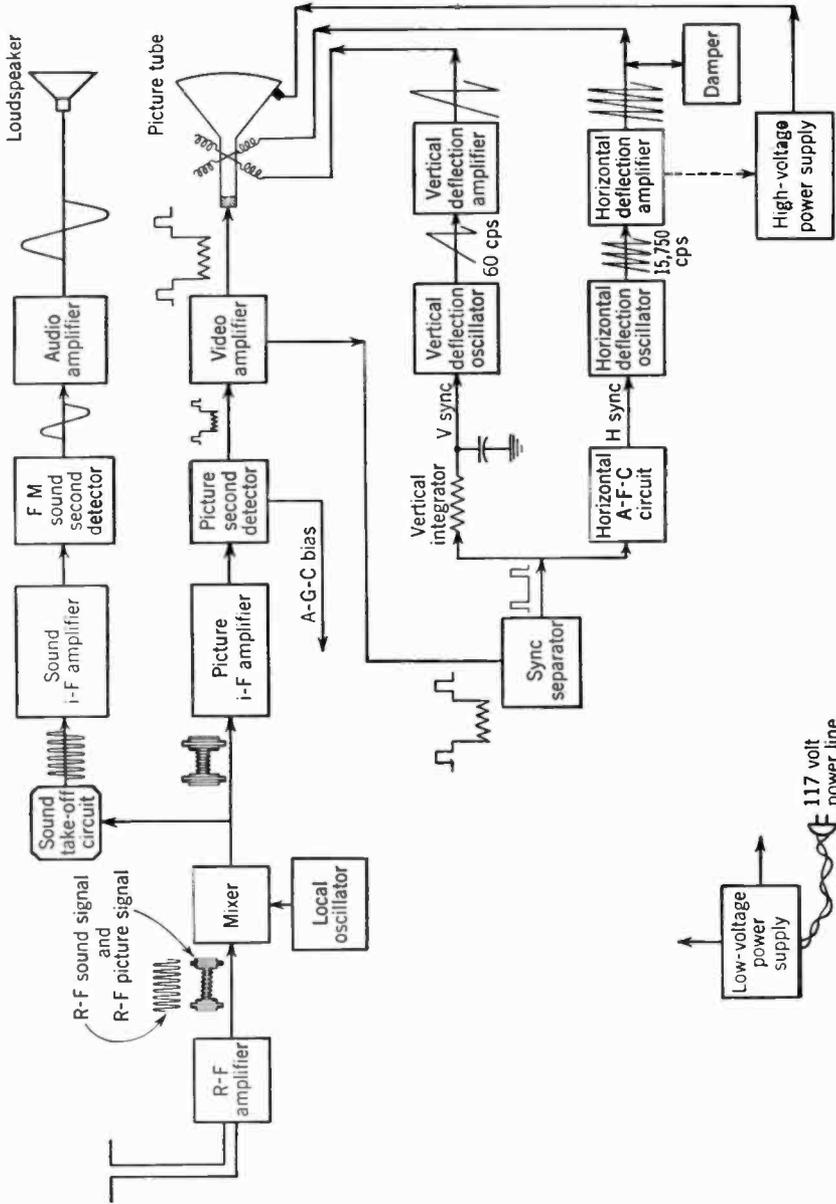


Fig. 8-2. Block diagram of a television receiver.

separation between the carrier frequencies is maintained. For most television receivers the picture and sound i-f carrier frequencies are approximately 26 and 21.5 Mc, respectively, or 45.75 and 41.25 Mc.

The r-f amplifier, converter, and local oscillator stages are usually on an individual subchassis, which is called the *front end*, *head end*, *r-f unit*, or *tuner*. The r-f unit selects the channel to be received by converting its picture and sound r-f carrier frequencies to the intermediate frequencies of the receiver, so that the selected signals can pass through the i-f stages.

*Sound Take-off.* The sound signal is separated from the picture signal after they have been converted to the lower intermediate frequencies. Since the bandwidth of the sound signal is relatively narrow, a resonant circuit tuned to this intermediate frequency can be used to filter out the sound signal and couple it to a separate sound i-f amplifier. The sound take-off block in Fig. 8-2, which is an *LC* tuned circuit rather than an amplifier stage, is shown connected to the converter output. However, the point where the sound signal is separated from the picture signal varies in different receivers. Many circuits have the sound take-off in the output of the first or second i-f amplifier stage. This method has the advantage of providing additional gain for the sound i-f signal in the same stages that amplify the picture i-f signal. In many receivers, the sound take-off is after the second detector stage. An i-f stage that amplifies both the picture and sound i-f signals is often called a *common i-f amplifier*. Regardless of where the sound take-off tuned circuit is in the receiver, this point marks the separation of the sound and picture signals. All stages before the sound take-off circuit amplify both the sound and picture signals. After the sound take-off point, the signal amplifiers are in two groups, one for the sound and the other for the picture.

*The Sound Amplifier Chain.* The FM sound i-f signal is amplified in one to three sound i-f stages to provide enough signal for the sound second detector stage. As illustrated in Fig. 8-2, the output of the FM detector is the desired audio-frequency signal. This is then amplified and coupled to a loudspeaker to reproduce the sound associated with the picture.

*The Picture Amplifier Chain.* Going back to the mixer stage in Fig. 8-2, the modulated picture i-f signal produced here is coupled to the picture i-f amplifier. Because of the broad bandwidth required to pass the picture signal, each picture i-f stage has a relatively low gain, typical values being 15 to 20. Three or four picture i-f amplifier stages generally are necessary, therefore, to provide the amount of picture i-f signal required for the picture second detector. This stage is commonly called the *video detector*. The modulated i-f picture signal is rectified and filtered in the video detector to produce the composite video signal output voltage that contains all the information needed for reproduction of the picture. The video detector output is then coupled to the video amplifier, consist-

ing of one or two stages, to provide enough video signal for the control grid-cathode circuit of the picture tube. With a peak-to-peak video signal voltage of approximately 70 volts at the kinescope control grid, the intensity of the beam current and the spot of light on the screen can be varied to allow reproduction of the picture.

*Automatic Gain Control.* The picture second detector in Fig. 8-2 is shown providing a bias voltage for automatic control of the gain of the preceding i-f and r-f stages, similar to the automatic volume-control system in conventional sound receivers. The stronger the picture carrier signal, the greater the negative a-g-c bias voltage produced and the less the gain of the receiver, resulting in relatively constant video signal amplitude for different carrier-signal strengths. Therefore, the automatic gain control in the picture amplifier chain is useful as an automatic control of contrast in the reproduced picture. No a-g-c bias is shown for the sound second detector in Fig. 8-2 because automatic volume control is not generally used in the FM sound amplifier chain. However, the a-g-c circuit affects both the picture and sound when it controls the gain of r-f and common i-f stages, which amplify the picture and sound signals.

*Synchronizing Circuits.* The video detector output includes the synchronizing pulses, which are part of the composite video signal for the picture tube. Therefore, the composite video is also coupled to the synchronizing circuits to provide the synchronizing pulses needed for controlling the frequency of the vertical and horizontal deflection oscillators in the receiver. The synchronizing circuits include one or more amplifier and separator stages. A synchronizing signal separator is a clipper amplifier stage that can separate the synchronizing pulse amplitude from the camera signal in the composite video, to provide an output consisting only of synchronizing pulses. Since there are synchronizing pulses for both horizontal and vertical scanning, Fig. 8-2 shows the output of the synchronizing separator divided into two parts. The integrator is a low-pass *RC* filter circuit that filters out all but the vertical pulses from the total separated sync voltage, to provide vertical synchronizing signals to lock in the vertical deflection oscillator at 60 cps. For horizontal synchronization, an automatic frequency-control circuit is generally used to lock in the horizontal deflection oscillator at 15,750 cps.

*Deflection Circuits.* As shown in Fig. 8-2, the deflection circuits include a deflection oscillator stage to produce the scanning voltage required for deflecting the electron beam, and a deflection amplifier stage to provide enough scanning-current amplitude to cover the screen of the picture tube. A horizontal oscillator and amplifier are needed for horizontal deflection. The damper stage minimizes oscillations produced by the deflection current in the horizontal output current, in addition to providing part of the horizontal scanning. The vertical oscillator and amplifier

produce vertical scanning. The horizontal and vertical deflection circuits produce the illuminated scanning pattern forming the raster on the kinescope screen. The raster can then be varied in intensity by the video signal coupled to the kinescope control grid, to reproduce the picture on the kinescope screen.

The deflection circuits produce the required deflection current and the resultant scanning raster with or without synchronizing pulses, because the deflection generators are free-running oscillators requiring no external signal for operation. However, the synchronizing signals are needed to hold the receiver scanning circuits exactly synchronized so that the picture information is reproduced on the raster in the correct position.

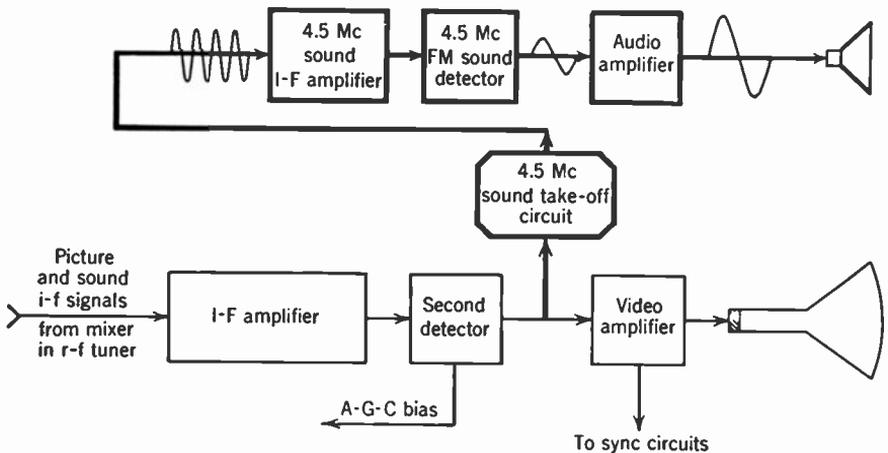


FIG. 8-3. Signal circuits in an intercarrier-sound receiver, showing sound take-off in the second detector output.

**Power Supplies.** Two power supplies are needed in the television receiver. One is the usual B supply with an output of 300 to 400 volts to provide d-c operating potentials for all amplifier stages. This is called the *low-voltage supply* in the television receiver because its output voltage is low compared with the *high-voltage supply*, which provides anode voltage for the picture tube. The anode voltage for direct-view picture tubes is generally 9 to 18 kv, while projection kinescopes require 20 to 80 kv. The dotted line to the high-voltage supply in Fig. 8-2 indicates that in many receivers the high-voltage rectifier obtains its high-voltage a-c input from the output of the horizontal deflection amplifier. With this type of high-voltage supply, the anode voltage for the picture tube depends upon operation of the horizontal deflection circuits.

**8-3. Intercarrier-sound Receivers.** Most television receivers use the circuit arrangement illustrated in Fig. 8-3 for the associated sound signal. The sound i-f signal passes through all the i-f amplifiers that amplify

the picture signal, and both signals are coupled into the second detector. Here the sound i-f signal can beat with the i-f picture carrier to provide in the detector output the desired FM sound signal converted to a lower frequency that is the difference between the two carrier frequencies. This second sound intermediate frequency is always 4.5 Mc, since this is the standard difference between the picture and sound carrier frequencies. Receivers using the second sound i-f signal of 4.5 Mc are called *intercarrier-sound receivers*. Because the 4.5-Mc sound i-f signal is produced in the second detector, the sound take-off point must be after this stage in intercarrier-sound receivers. Although shown in the output of the second detector stage in Fig. 8-3, the 4.5-Mc sound take-off can be in any part of the video amplifier circuits after the second detector. Figure 8-3 does not indicate the sync and deflection circuits and power supplies because these circuits are not affected by the intercarrier method of receiving the sound signal.

The sound take-off circuit in an intercarrier receiver is an *LC* tuned circuit resonant at 4.5 Mc. This circuit filters out the 4.5-Mc sound signal from the video frequencies and couples it to the sound i-f amplifier of the receiver. It is important to note that the sound i-f amplifier and FM detector in an intercarrier receiver are always tuned to 4.5 Mc. The 4.5-Mc sound i-f amplifier and FM detector are necessary because the 4.5-Mc sound signal is not an audio signal but is the frequency-modulated i-f signal at the lower center frequency of 4.5 Mc, which must be detected to obtain the desired audio signal. The main advantage of intercarrier-sound receivers is the fact that the 4.5-Mc sound i-f signal is automatically present when the picture is tuned in.

**8-4. Functions of the Receiver Circuits.** The television receiver can be considered in sections, each with a specific function, as follows:

*Illumination.* The picture tube with its auxiliary components and the high-voltage supply, which provides the kinescope anode voltage, have the function of producing brightness on the kinescope screen. Just the spot of light on the screen illustrated in Fig. 8-4a shows that the kinescope and high-voltage supply are operating.

*Horizontal Scanning.* The horizontal deflection circuits produce horizontal scanning lines. The single horizontal line on the kinescope screen in Fig. 8-4b shows illumination and horizontal scanning.

*Vertical Scanning.* The vertical deflection circuits produce vertical scanning to spread the horizontal scanning lines over the entire screen area, to form the scanning raster. The illuminated raster on the kinescope screen in Fig. 8-4c shows that the vertical and horizontal deflection circuits, kinescope, and high-voltage supply are operating.

*Picture.* Figure 8-4d shows a picture reproduced on the raster. The circuits for the picture from the antenna input to the kinescope grid pro-

vide video signal corresponding to the desired picture information. The video signal-voltage variations on the kinescope grid vary the intensity of the electron beam, while the deflection circuits produce scanning, to reproduce the picture as shades of white, gray, and black on the raster.

*Synchronization.* The sync circuits in the receiver hold the line structure of the picture together and make it stay still by timing the horizontal and vertical scanning correctly with respect to the reproduced picture

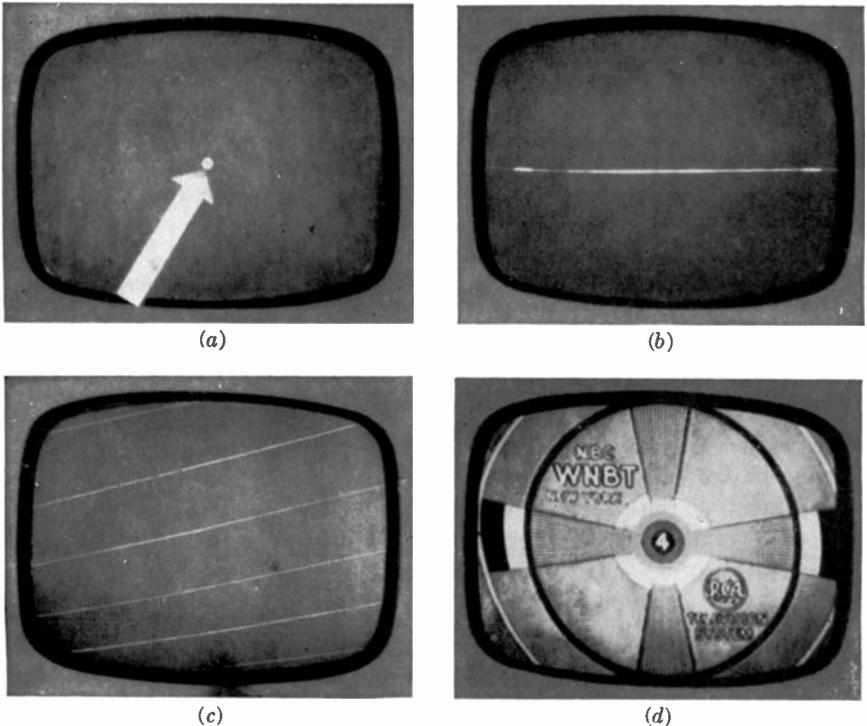


FIG. 8-4. Illustrating how the receiver puts the picture on the raster. (a) Illumination on the kinescope screen. (b) Illumination plus horizontal scanning. (c) Horizontal and vertical scanning to produce the illuminated raster. (d) The picture on the raster. (From McGraw-Hill motion-picture series, *Basic Television*.)

information. Horizontal synchronization prevents the line structure of the picture from tearing apart into diagonal segments. Vertical synchronization allows successive frames to be superimposed over each other so that the picture will not roll up or down on the screen.

*Sound.* The signal circuits for the sound provide audio signal for the loudspeaker to reproduce the sound.

Table 8-1 illustrates how the receiver can be divided into circuits that produce the illuminated raster and circuits for the picture and sound signals. The signal circuits are subdivided between picture and sound for

the typical case of an intercarrier-sound receiver that has the 4.5-Mc sound take-off circuit in the video detector output. Note that the low-voltage power supply is common to the raster and signal circuits, since all the amplifiers need the B supply voltage for operation.

TABLE 8-1. RECEIVER CIRCUIT FUNCTIONS

Circuits for the raster		Circuits for the signal		
Illumination	Scanning	Picture and sound	Picture only	Sound only
Kinescope	Horizontal deflection circuits	R-f tuner	Video amplifier	4.5-Mc i-f stages
High-voltage supply	Vertical deflection circuits	Common i-f stages Second detector A-g-c circuit		4.5-Mc FM detector Audio amplifier
Low-voltage power supply				

**8-5. Receiver Operating Controls.** The controls for adjusting the operation of the receiver circuits can be considered in two groups: the setup adjustments for the raster circuits, and the operating controls in the signal circuits. The setup adjustments for scanning are generally mounted on the rear apron of the chassis, to be set as installation or servicing adjustments to provide a suitable raster, while the operating

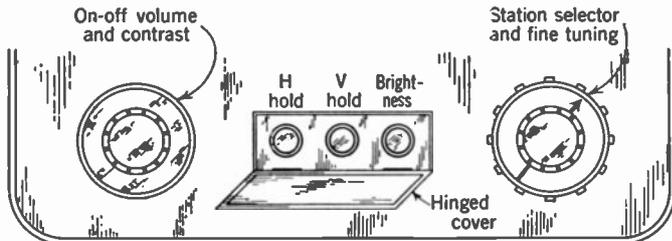


FIG. 8-5. Receiver operating controls. The hinged lid on the front panel normally covers the three controls at the center.

controls are on the front panel of the receiver, where they can be varied for different stations. Figure 8-5 illustrates a common arrangement for the operating controls. They have the following functions:

**Station Selector.** This adjusts the resonant frequency of the r-f circuits in the front end to the desired channel frequencies and sets the local oscillator at the frequency necessary to tune in the station.

**Fine Tuning.** This provides more exact setting of the local oscillator frequency. In split-sound receivers, the fine tuning control is adjusted

for the best sound, which normally results in the best picture also. With intercarrier receivers, however, the fine tuning control can be adjusted for the best picture, independently of the sound.

*Volume.* This is a typical audio level control, usually a potentiometer to vary the audio voltage input to the grid circuit of the first audio amplifier. Some receivers may also have a tone control.

*Contrast.* Since most receivers have automatic gain control to vary the gain of the r-f and i-f amplifiers, the contrast control usually adjusts the gain of the video amplifier to control the amplitude of the video signal voltage for the kinescope grid-cathode circuit.

*Brightness.* This varies the d-c bias for the kinescope grid, adjusting the over-all illumination on the screen.

*Horizontal Hold.* This adjusts the frequency of the horizontal deflection oscillator close enough to the synchronizing frequency to allow the sync to lock in the horizontal scanning at 15,750 cps. When the picture tears apart into diagonal segments, the horizontal hold control is varied to reestablish horizontal synchronization and provide a complete picture.

*Vertical Hold.* This adjusts the frequency of the vertical deflection oscillator close enough to the vertical synchronizing frequency to allow the sync to lock in the vertical scanning at 60 cps. When the picture rolls up or down, the vertical hold control is varied to reestablish vertical synchronization and make the picture stay still.

It should be noted that the physical location for the controls can vary in different receivers. Many receivers have the hold controls and brightness adjustment on the rear apron of the chassis, while others may mount the setup adjustments at the front, behind a hinged cover on the front panel.

**8-6. Projection Television Receivers.** The practical limit for the diameter of a conventional direct-view picture tube is about 30 in. To produce larger pictures for home entertainment or theater television it is preferable to employ a projection system for the television image. There are two main types of projection systems. One, called a *refractive projection system*, uses a convergent projection lens to throw the kinescope screen image onto a larger viewing screen in a manner similar to the projection of film slides or motion-picture film, as illustrated in Fig. 8-6. The other projection arrangement uses a spherical reflecting mirror instead of the lens to provide the required enlargement of the image and, therefore, is a *reflective system*. This arrangement is illustrated in Fig. 8-7.

The principal problem in television projection is sufficient brightness. The brightness of the enlarged image is very much less than the intensity of the picture on the kinescope screen because the available light must be distributed over a much larger picture area in the projected image, reducing the brightness proportionately. In addition, only 5 to 30 per cent

of the light from the kinescope is collected by the optical projection system and delivered to the viewing screen. Very high anode potentials are needed to obtain the required screen brightness in projection picture tubes, therefore, and voltages of 20 to 80 kv are used. To improve the brightness in the magnified picture the viewing screen for projection receivers usually is made directional to concentrate the light from the screen to the viewer. The brightness gain is 6 to 9. However, the directional screen limits the viewing area to  $\pm 25^\circ$  from center in the horizontal direction, and  $\pm 10^\circ$  in the vertical direction.

*Refractive Optical Projection System.* Figure 8-6 illustrates a projection-lens arrangement for television receivers. The reproduced image is

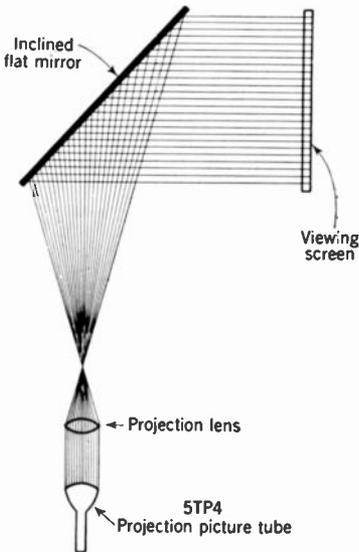


FIG. 8-6. Projection-lens system for television receivers.

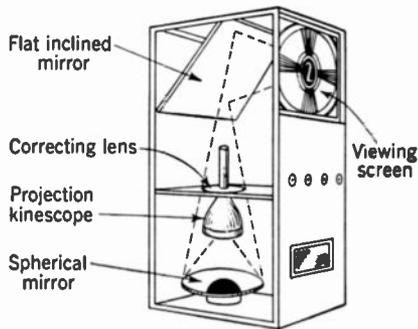


FIG. 8-7. Folded Schmidt-type reflective optical system for television projection receivers.

formed on the screen of the picture tube in the usual manner, employing the 5TP4 kinescope, a commonly used projection tube. The screen of the picture tube is small, producing a 4- by 3-in. picture, so that a projection lens of practical diameter can be used. Light from the picture-tube screen is collected by the lens and is thrown on the viewing screen as in film projection, to provide an enlarged 20- by 15-in. picture when a magnification of 5 is employed. This is a transmission type of viewing screen, as light striking it from the back produces an image that can be viewed from the front of the screen.

By adjusting the distance between lens and kinescope and the throw distance to the viewing screen, the refractive optical arrangement can pro-

vide enlarged pictures of almost any desired size. However, it is very inefficient, utilizing only 5 to 10 per cent of the available light from the kinescope screen.

*Reflective Optical Projection System.* Figure 8-7 illustrates a reflective optical system for the projection of television pictures, using the spherical reflecting mirror instead of a projection lens. The arrangement is an adaptation of the Schmidt astronomical camera and, therefore, is called a *Schmidt-type* optical projection system. Of the available light 20 to 30 per cent can be transmitted to the viewing screen. Since a greater light efficiency can be obtained, the reflective optical system is employed more often for television receivers than the projection-lens method.

In Fig. 8-7 it is seen that the projection picture tube is mounted face downward. Light from the picture tube strikes the spherical mirror and is reflected upward through the correcting lens mounted on the neck of the picture tube. The reflected light then strikes an inclined flat mirror and is again reflected to the viewing screen. With a suitable distance from picture tube to spherical mirror and from spherical mirror to viewing screen the desired enlargement of the image can be obtained, because the spherical mirror magnifies the image in the same manner as the converging projection lens. The center of the spherical reflected mirror is painted black because light rays from this part of the mirror would be blocked by the picture tube. This reduces the light efficiency, but does not block any part of the image as each picture element is a point source of light rays that have many paths to the viewing screen. The correcting lens is necessary to eliminate a form of optical distortion called *spherical aberration*, which is produced by the spherical mirror.

The reflective optical assembly, including the picture tube, spherical mirror, and plastic correcting lens, are mounted in a housing unit called the *optical barrel* that keeps out any undesired light and prevents dust from collecting on the spherical mirror. All the optical components must be spotlessly clean and free of dust for the best light efficiency. Adjustments on the reflective optical system are critical because the shape and size of the correcting lens depend on the projection distance for which the system is designed.

**8-7. Theater Television.** Theaters can provide television programs by means of large-screen projection, presenting news and sports events as they occur. The size of the projected picture can be increased to the desired value, usually 6 by 8 ft or 15 by 20 ft, by increasing the throw distance to the viewing screen. It is necessary to provide sufficient brightness in the original picture tube image to compensate for the tremendously increased size of the picture, and very high anode voltages must be used for the picture tube. The Schmidt-type reflective optical system is used to obtain the greatest light efficiency. With 30 kv on

the anode of the 5TP4 projection kinescope, suitable brightness can be obtained for a picture 6 by 8 ft; using 80 kv on the anode of the 7WP4 theater projection kinescope provides enough brightness for a picture 15 by 20 ft, which is a typical television-screen size for theaters. Figure 8-8 shows a television projection unit mounted on a theater balcony, to project the picture on the theater's screen. The base of the projection kinescope is at the front of the optical barrel.

Additional methods for theater television projection employ a mechanical form of light valve, which can modulate the intensity of light from a

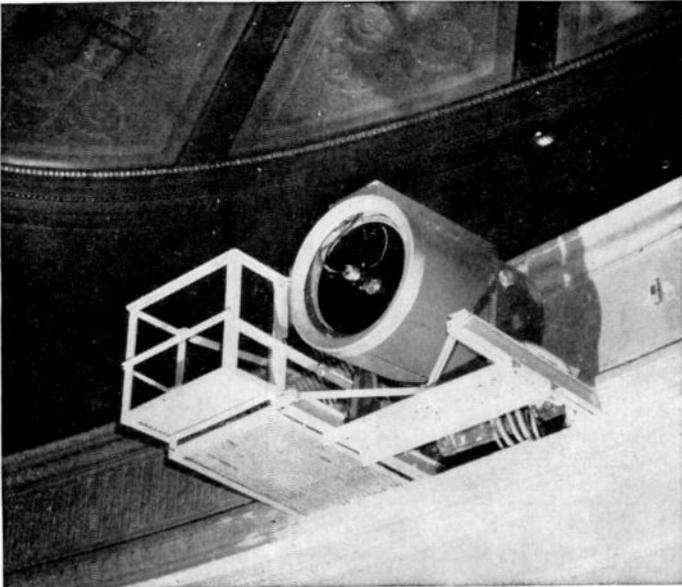


FIG. 8-8. Large-screen television projector mounted on front of balcony in theater. (RCA.)

high-intensity light source such as the carbon-arc lamp. No kinescope is used in this arrangement, since the picture information is reproduced by the light valve. Therefore, the intensity of the projected picture is not limited by the kinescope screen brightness. However, a mechanical scanning system must be used with the light valve. The picture can be reproduced in either color or black and white. Two commercial forms of theater television projection using a light valve with mechanical scanning are known as the *Eidophor* and the *Scophony* systems. Another method of providing theater television utilizes commercial motion-picture film with the theater's standard film projectors. The picture signal is transmitted to a television receiver at the theater. In the television room, the picture reproduction on the screen of the kinescope is photographed,

developed, and dried in less than 1 min to provide motion-picture film for the regular projectors.

The picture and sound signals from the source of the program can be transmitted to the theaters by special telephone lines in a closed-circuit link, or by microwave radio-relay links. In either case, the transmission for theater television is outside the commercial television broadcast channels. The scanning standards and bandwidth requirements are not limited by the 6-Mc broadcast channel, therefore, and the specifications for theater television can provide picture quality equal to 35-mm commercial motion-picture film.

**8-8. Subscriber Television.** Systems of coded or secret television transmission have been proposed where the modified signal from a con-

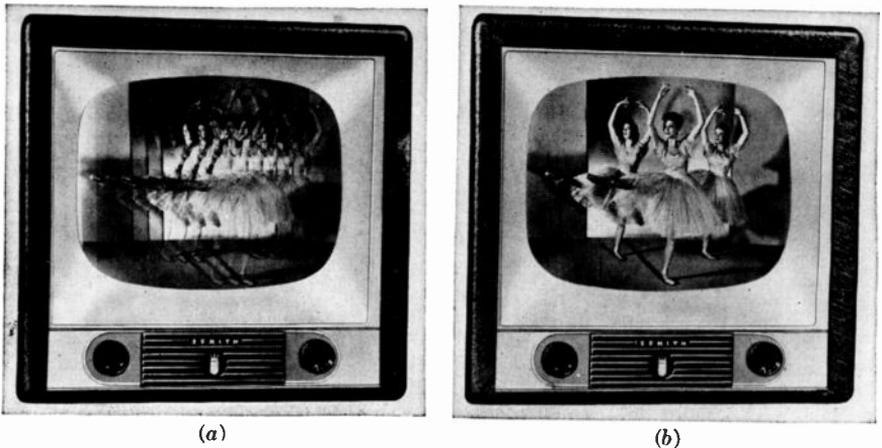


FIG. 8-9. Picture reproduction in Phonevision system. (a) Without coding signal at receiver. (b) With coding signal. (Zenith Radio Corp.)

ventional transmitter can be received as a clear intelligible picture only on a receiver supplied with a decoding signal to correct the picture reproduction. The purpose of such a coded transmission is to provide a means of charging and collecting a fee for special program material, such as first-run motion pictures, which would not normally be available to television stations and advertisers because of excessive costs. In effect, the system provides a box office for television broadcasting. In one arrangement called *Phonevision* the ordinary telephone line of the subscriber is used to supply the coding signal needed to correct the scrambled picture broadcast by the television transmitter; in another system known as *Subscriber-Vision* a punched card with coded holes in it must be used at the receiver to correct the scrambled picture. Subscriber television systems have been tried on an experimental basis as an auxiliary service to regular television broadcasting to provide special programs.

The method of making the video signal private consists of changing the phase relation between the synchronizing signals and the picture information at irregular intervals. Sometimes the camera signal variations of the video signal are phased normally with respect to sync while at other times the sync is phased incorrectly. Changes between the two types of signal are entirely random so as to give the system secrecy. Because of the

TABLE 8-2. R-F AND I-F AMPLIFIER TUBES

Type	Typical operation and characteristics— class A <sub>1</sub> amplifier				Remarks*
	Plate volts	Screen volts	Grid-bias volts (approx)	$G_m$ , $\mu\text{mhos}$	
6AG5	250	150	-1.5	5,000	7-pin miniature glass; sharp-cutoff pentode; -8-volts approx cutoff; r-f or i-f amplifier
6AN4	200	...	-1.3	10,000	7-pin miniature glass triode; -7 volts approx cutoff; u-h-f amplifier-mixer
6AU6	250	150	-1	5,200	7-pin miniature glass; sharp-cutoff pentode; -6.5 volts approx cutoff; limiter in FM circuits; i-f amplifier
6BA6	250	100	-1 to -20	4,400	7-pin miniature glass; remote-cutoff pentode; i-f amplifier
6BQ7-A	150	...	-1.2	6,400	9-pin miniature glass; medium-mu twin-triode; values for each unit; -10 volts approx cutoff; r-f amplifier for u-h-f and v-h-f
6CB6	200	150	-2.2	6,200	7-pin miniature glass; sharp-cutoff pentode; -8 volts approx cutoff; r-f or i-f amplifier
6J6	100	...	1.0	5,300	7-pin miniature glass; medium-mu twin-triode; values for each unit; -4 volts approx cutoff; local oscillator, mixer, r-f amplifier

\* The applications noted are typical but not all-inclusive.

scrambled signal the resulting picture reproduction is blurred, with the image moving back and forth horizontally at a slow irregular rate. Figure 8-9a shows how the blurred picture looks on a receiver not supplied with the decoding signal, while *b* is the same picture after being cleared up by the key signal at the receiver.

**8-9. Receiver Tubes.** Tables 8-2 to 8-5 list examples of tube types often used in television receivers. The tabulations do not include all

possible tubes since there are many types available and additions are constantly being made. However, the tubes listed are typical and they indicate some important characteristics of television receivers.

*R-F Amplifier Tubes.* The tube types in Table 8-2 are intended for wide-band or high-frequency service, featuring high values of transconductance with small interelectrode capacitances. Figure 8-10 shows the 6CB6, which is a typical miniature glass tube for high-frequency work. The height of the tube is only  $1\frac{1}{2}$  in. and the miniature construction allows smaller tube capacitances. The high-frequency twin-triode tubes

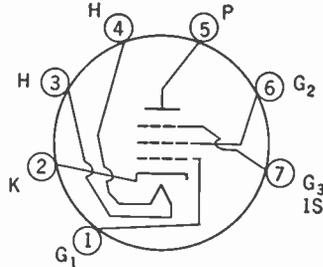


FIG. 8-10. Miniature glass tube with 7-pin base. (RCA.)

6J6 and 6BQ7-A are often used in the r-f amplifier stage because of the low noise rating of triodes.

*R-F Oscillator and Mixer Tubes.* Separate oscillator and mixer stages are generally used for frequency conversion in the superheterodyne television receiver because of the high frequencies. An r-f amplifier tube, such as the 6AG5, 6CB6, or 6J6, is suitable for the mixer stage. The local oscillator uses a miniature glass triode having medium  $\mu$ , like the 6J6, which can sustain oscillations at the high frequencies required. The 6J6 twin-triode is often used, with one triode section as the oscillator and the other triode as the mixer. The 9-pin miniature glass type 6X8 is a triode-pentode converter tube suitable for the v-h-f television channels, providing performance comparable to a 6AG5 mixer with one triode section of the 6J6 as oscillator. For the u-h-f channels, a crystal diode is commonly used for the mixer.

*Video Amplifier Tubes.* Referring to Table 8-3, the 6AC7, 6AG7, and 6CL6 feature high transconductance with low tube capacitances for wide-band amplifier operation; also, they can supply enough video signal output voltage to drive the kinescope grid-cathode circuit without introducing too much amplitude distortion. These tubes are generally used

TABLE 8-3. VIDEO AMPLIFIER TUBES

Type	Typical operation and characteristics of class A <sub>1</sub> amplifier				Remarks
	Plate volts	Screen volts	Grid-bias volts	G <sub>m</sub> , $\mu$ hos	
6AC7	300	150	-1.6	9,000	Octal, sharp-cutoff pentode
6AG7	300	150	-3	11,000	Octal, power pentode
6CL6	300	150	-3	11,000	9-pin miniature glass pentode
6K6	250	250	-18	2,300	Octal, power pentode, output tube
6V6	250	250	-12.5	4,100	Octal, power pentode, output tube
7C5	250	250	-12.5	4,100	Loctal, power pentode, output tube

when there is just one video amplifier stage. With two video amplifier stages, one of the audio output tubes 6K6, 6V6, and 7C5 is generally used in the video output stage because of the larger grid signal swing. The first video amplifier stage can use a sharp-cutoff r-f amplifier tube like the 6AU6.

*Deflection Amplifier Tubes.* Power tubes are required in the deflection circuits. Referring to Table 8-4, note that the horizontal output tubes can

TABLE 8-4. DEFLECTION POWER AMPLIFIER TUBES

Type	Maximum ratings for magnetic deflection in 525-line 30-frame system			Plate dissipation, watts	Remarks*
	Plate volts	Screen volts	Grid-bias volts		
6AU5-GT	450	200	-50	10	Octal; beam power; horizontal output
6BG6-G	700	350	-50	20	Octal; beam power; horizontal output; 19BG6-G has 19.8-volt heater for series filaments
6BQ6-GT	500	200	-50	10	Octal; beam power; horizontal output; 25BQ6-GT has 25-volt heater for series filaments
6CD6-G	700	175	-50	15	Octal; beam power; horizontal output
6CU6	550	175	-28†	11	Octal; beam power; horizontal output
6S4	500	...	-50	7.5	9-pin miniature glass; medium-mu triode; vertical output
12BH7	500	...	-50	3.5	9-pin miniature glass; medium-mu twin-triode; vertical deflection oscillator and output

\* The applications noted are typical but not all-inclusive.

† -28 volts is typical control-grid bias.

dissipate more power than the vertical output tubes. The 6BG6-G and 6CD6-G horizontal output tubes are physically larger than the GT types. Figure 8-11 shows the 6BG6-G which has the plate connection brought out to the top cap because of the high plate voltage. The 12BH7 illustrates the twin-triode tube type with medium power capabilities, commonly used in the deflection and sync circuits because of the economy of the double-tube arrangement. Additional twin-triode tubes with similar applications are the 12AU7 9-pin miniature glass tube and the 6SN7-GT octal tube. Each section of the 6SN7 has the same character-

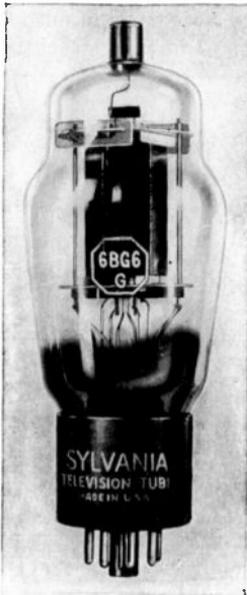


FIG. 8-11. Horizontal power output tube 6BG6. The top cap is the plate connection. (Sylvania Electric Products, Inc.)



FIG. 8-12. High-voltage rectifier tube. The top cap is the plate connection. (RCA.)

istics as the 6J5 medium-mu triode. The 12BH7 and 12AU7 have the heater center-tapped for either 6.3- or 12.6-volt operation. The damper tube in the horizontal output circuit is usually a diode power rectifier, such as the 5V4, 6W4-GT, and 6AX4-GT, featuring high-voltage insulation from cathode to heater.

*Rectifier Tubes.* Table 8-5 lists several rectifier tube types for the power supplies in a television receiver. For the low-voltage supply the maximum permissible d-c output is the most important characteristic of the rectifier. Full-wave rectifier vacuum tubes or selenium rectifiers are generally used. Two full-wave rectifiers may be used to supply the required load current of 200 to 300 ma. For the high-voltage rectifier

tubes the maximum peak inverse voltage rating is most important. This is the maximum voltage that can be applied to the tube in the inverse direction, making the plate negative, without danger of arcing in the tube. The high-voltage supply always uses a half-wave rectifier for the high-voltage output of 9 to 18 kv, with a low-current drain of less than 1 ma. Figure 8-12 shows a photograph of the 1B3-GT high-voltage rectifier, which is an octal tube. The 1X2-A is a nine-pin miniature glass tube with lower voltage and current ratings than the 1B3-GT.

TABLE 8-5. RECTIFIER TUBES

Type	Filament		Maximum ratings			Remarks
	Volts	Amperes	Peak inverse plate volts	Peak plate current, ma	Average d-c output, ma	
5AW4	5.0	4.0	1,550	750	250	Full-wave rectifier for large d-c load current
5T4	5.0	2.0	1,550	675	225	
5U4	5.0	3.0	1,550	675	225	
5Y3	5.0	2.0	1,400	375	125	Full-wave rectifier for moderate d-c load current
1B3-GT*	1.25	0.2	30,000†	17	2	Half-wave rectifier for high-voltage supply
1X2/A	1.25	0.2	18,000†	10	1	
3A3	3.15	0.2	30,000†	80	1.5	

\* Formerly tube type 8016.

† Maximum frequency of a-c supply voltage, 300 kc.

*Detectors.* The 6AL5 is a miniature glass twin-diode that is often used for the video detector, a-g-c rectifier, discriminator or ratio detector for the FM sound signal, and sync discriminator in the horizontal a-f-c circuit. Germanium and silicon crystal diodes also are commonly used in such low-power detector applications.

**8-10. Localizing Troubles to a Receiver Section.** In order to localize troubles, we can use three indicators in the television receiver: the illuminated raster, the picture, and the sound. Several examples are given here to illustrate how these indicators can help to localize a trouble. These are based on the typical receiver shown in Fig. 8-13, using inter-carrier sound with the 4.5-Mc sound take-off in the second detector output circuit and having a high-voltage supply that operates from the horizontal deflection circuits.

*No Illumination, with Normal Sound.* In this case, the kinescope screen is completely blank, without any light output or scanning lines and, there-

fore, no picture. Even with video signal at the kinescope grid, the picture cannot be reproduced unless the illuminated raster is present. The first step necessary to provide a picture, therefore, is to find out why there is no brightness. Since the sound circuits are operating, the receiver has a-c power input and the low-voltage supply is providing B+ output. In order to have brightness, the kinescope itself must be operating normally and the high-voltage supply must provide kinescope anode voltage. It is important to remember that the horizontal deflection circuits produce the a-c input for the high-voltage power supply. Therefore, the trouble causing no illumination can be in the kinescope and its associated circuits,

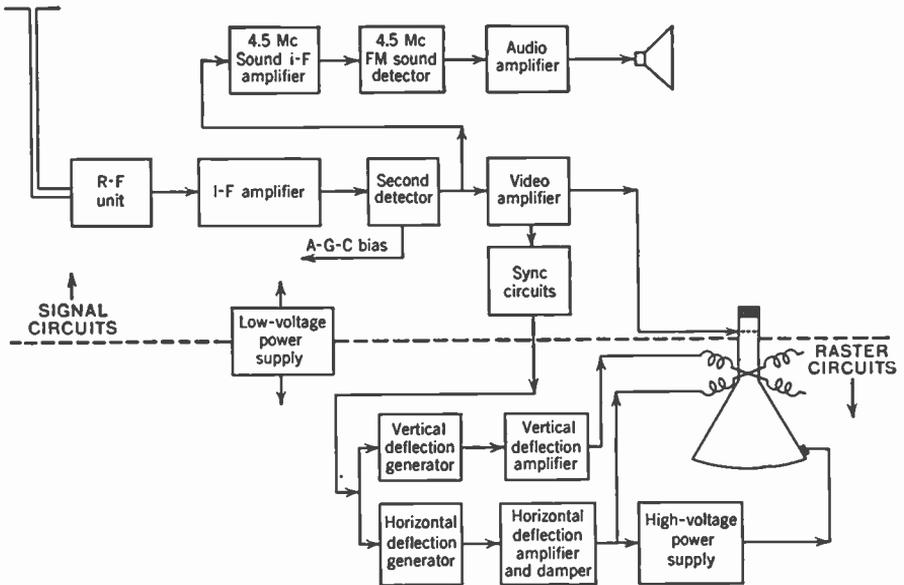


FIG. 8-13. Signal circuits and raster circuits in a typical television receiver.

the high-voltage supply or the horizontal deflection circuits. If there is d-c high voltage for the kinescope anode, the trouble is in the kinescope or its associated circuits. If there is no d-c high-voltage output, the trouble is in the high-voltage supply or the horizontal deflection circuits. The trouble can be localized further by noting whether the horizontal deflection circuits are producing the a-c high-voltage input for the plate of the high-voltage rectifier.

*No Picture and No Sound, with Normal Raster.* The normal raster indicates that the kinescope, deflection circuits, and power supplies are operating. This trouble is in the signal circuits, before the sound take-off point, because both the picture and sound are affected. Or, the trouble

can be in the a-g-c circuit when it controls stages common to both the picture and sound signals. In Fig. 8-15 the circuits common to the picture and sound signals are the r-f section, i-f amplifier, second detector, and a-g-c circuit. If the trouble occurs only on some channels but not on others the defect is probably in the r-f section, including the antenna and transmission line, since this is the only part of the receiver operating on the r-f signal frequencies for each individual channel.

*No Picture, with Normal Raster and Normal Sound.* Figure 8-4c shows the kinescope screen with just the raster. The fact that the receiver produces the raster means the kinescope and the raster circuits are functioning, which includes the vertical and horizontal deflection circuits and the power supplies. In the signal circuits all the stages operating on the sound signal must be normal. The one section in Fig. 8-15 operating only on the signal for the kinescope grid is the video amplifier; therefore, the trouble must be in this stage. It should be noted, though, that, in receivers with an a-g-c amplifier directly coupled to the video amplifier, a defect in the video amplifier can affect the sound through the a-g-c circuit.

*No Sound, with Normal Raster and Normal Picture.* The normal picture on the raster means the kinescope, deflection circuits, picture signal circuits and power supplies are operating. The trouble must be in the sound circuits, after the sound take-off point, because only the sound is affected. In Fig. 8-15 this includes the 4.5-Mc sound take-off circuit in the video detector output, the 4.5-Mc sound i-f amplifier, the 4.5-Mc FM detector, the audio amplifier, and the loudspeaker. The trouble can be localized to a specific stage in the sound circuits by signal tracing to find out which circuit does not pass signal.

*Only a Horizontal Line on the Screen.* Figure 8-4b shows how the kinescope screen looks with horizontal deflection only. The first step necessary is to find the trouble in the raster circuits. Any illumination on the screen means the kinescope and its auxiliary circuits for low-voltage and high-voltage d-c operating potentials are functioning. The horizontal deflection circuits are producing the horizontal line on the kinescope screen. Only vertical deflection is missing. Therefore, the trouble must be in the vertical deflection section of the raster circuits, which includes the vertical deflection generator and the vertical output stage.

*No Raster and No Sound.* The screen is completely blank, without illumination, and there is no sound. This trouble means the raster circuits and signal circuits are not operating. The defect must be in the low-voltage power supply, since this is the only receiver section that affects both the raster and the signal. Either there is no a-c input power, or the low-voltage supply is not producing B+ voltage output.

## REVIEW QUESTIONS

1. How is the image reproduced on the screen of the picture tube?
2. What is the function of the antenna and the transmission line?
3. What is the function of the r-f tuning section?
4. What is the function of the sound take-off circuit?
5. What is the function of the sound i-f amplifier section?
6. What is the function of the FM sound detector stage?
7. What is the function of the picture i-f amplifier section?
8. What is the function of the video detector?
9. What is the function of the video output stage?
10. What is the function of the sync separator circuits?
11. What is the function of the vertical deflection generator and amplifier stages?
12. What is the function of the horizontal deflection generator and amplifier stages?
13. What is the function of the low-voltage power supply?
14. What is the function of the high-voltage power supply?
15. Give one feature of the 1B3-GT high-voltage rectifier tube.
16. Give one advantage of miniature glass r-f amplifier tubes like the 6CB6 compared with a metal tube such as the 6SH7.
17. In projection television give two reasons for the reduced brightness of the projected picture compared with the brightness of the original image on the face of the kinescope.
18. Describe briefly two features of theater television.
19. Why is the anode voltage on a projection kinescope for large-screen theater television higher than the anode voltage for a direct-view kinescope?
20. Referring to the receiver diagram illustrated in Fig. 8-2 classify all the blocks according to raster circuits, circuits for the picture, sound signal circuits, and circuits common to both the picture and sound signals.
21. For the receiver block diagrams in Figs. 8-2 and 8-13 explain how a trouble could be localized to one section of the receiver, for the following:
  - a. No picture, with normal raster and normal sound.
  - b. No sound, with normal raster and normal picture.
  - c. No picture and no sound, with normal raster.
  - d. Only a horizontal line on screen, with normal sound.
  - e. No raster, with normal sound.
  - f. Weak picture with little contrast, on some channels but not others.
  - g. No raster and no sound. Tube filaments are lighted.
  - h. No raster and no sound. None of the tube filaments is lighted.

## CHAPTER 9

### PICTURE TUBES

As illustrated in Fig. 9-1, the picture tube, or kinescope, is a funnel-shaped cathode-ray tube consisting of a vacuumed envelope, an electron gun, and a luminescent screen. The gun serves to produce and direct a narrow beam of electrons down the length of the tube toward the screen. To form the screen, the inner surface of the wide glass face of the tube is coated with a luminescent material which produces light under impact of the bombarding electrons. A deflection system at the neck of the tube deflects the electron beam over the entire surface of the screen to produce the raster. With video signal at the control grid of the electron gun to vary the beam intensity, the picture information is reproduced on the screen.

**9-1. Deflection and Focus.** Either magnetic or electrostatic deflection may be employed for a cathode-ray tube, but practically all picture tubes use magnetic deflection. The deflection yoke containing the scanning coils is mounted externally on the neck of the kinescope, as shown in Fig.

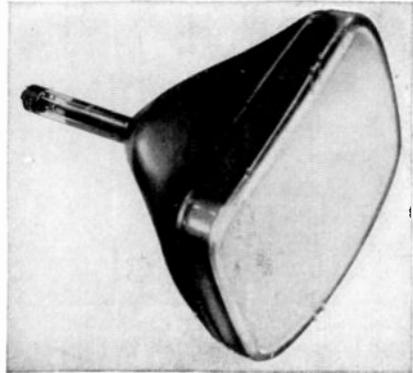


FIG. 9-1. A 20-in. rectangular glass picture tube. (*Sylvania Electric Products, Inc.*)

9-2. Focusing of the electron beam may be either electrostatic or magnetic. With electrostatic focusing, the voltage on the focusing grid of the electron gun within the tube controls the focus. When magnetic focusing is used a focusing magnet is mounted externally on the tube over the electron gun, as in Fig. 9-2. The focusing magnet may be either a permanent magnet (pm) or an electromagnetic (em) coil magnet.

As shown in Fig. 9-2, the deflection yoke is mounted where the envelope begins to flare out, with the rubber cushion at the front of the yoke's housing pressing lightly against the cone of the tube. If the yoke is placed too far back toward the tube base, the beam will be deflected off

the screen for large deflection angles and the corners of the raster will not be visible. The yoke can be turned slightly in its housing, by loosening the wing nut at the top and moving it to the left or right, to make the edges of the raster parallel to the mask of the picture tube. Figure 9-3 shows a typical kinescope deflection yoke. Yokes with nonuniform windings for the deflection coils, which become progressively thinner toward the front opening because of a *cosine* or *cosine-squared* winding distribution, are called *anistigmatic* yokes. They have the advantage of improving the focus at the edges of the raster. However, pin-cushion or barrel distortion may result, with the edges of the raster bowed in or out, requir-

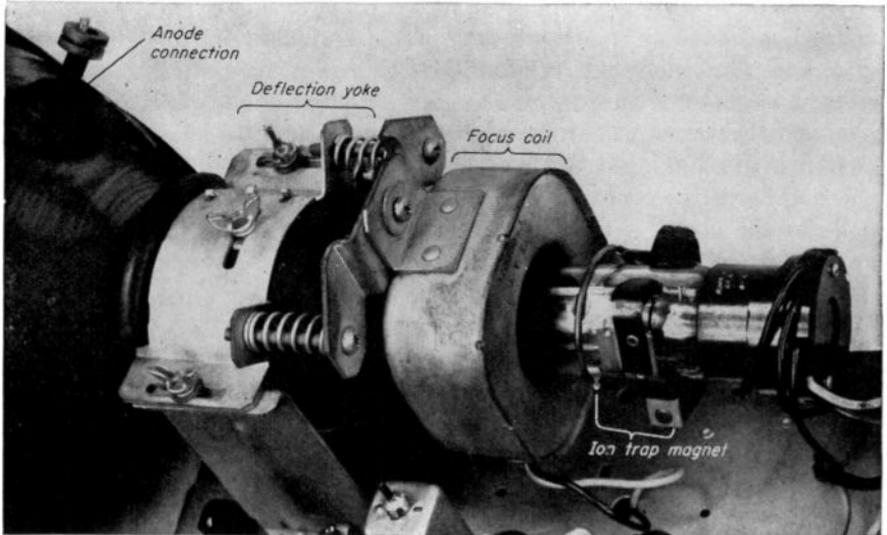


FIG. 9-2. Control units mounted on neck of picture tube using magnetic deflection and magnetic focus. (RCA.)

ing the use of correction magnets as illustrated by the pin-cushion magnets mounted on the deflection-yoke housing in Fig. 9-5.

The focus magnet is mounted just behind the yoke, with a space about  $\frac{1}{2}$  in. or less between them. In order to focus the scanning lines in the raster the physical position of the focus magnet ring can be adjusted by moving it back and forth along the length of the tube, and turning the magnet slightly to the left or right. Best focus is usually obtained with the tube neck approximately centered in the hole of the ring magnet. A focus coil with about 5,000 turns carrying approximately 100 ma direct current provides the magnetic field strength required for correct focus, with an anode voltage of 10 to 15 kv. Usually, a fine-focusing adjustment is also provided. Either a rheostat varies the current through the focus coil, or a movable iron ring functions as a magnetic shunt for a pm

focus magnet, as illustrated in Fig. 9-4. The thumbscrew focusing adjustment shown here moves the iron sleeve back and forth along the axis of the kinescope to vary the magnetic focusing field.

The pm focus magnet in Fig. 9-4 also has a centering ring, which is eccentric with the center hole in the magnet. The lever on the centering ring can be moved up or down and left or right to provide mechanical centering by changing the direction of the magnetic focusing field. Electrical centering can be provided by rheostats to vary the direct current through the horizontal and vertical scanning coils, but this is not commonly used because of the added current drain on the low-voltage power



FIG. 9-3. Kinescope deflection yoke. (RCA.)

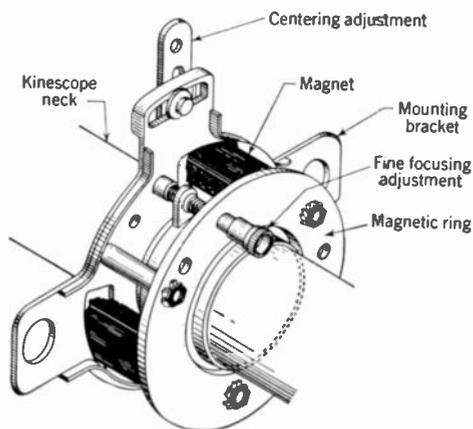


FIG. 9-4. A pm focusing ring magnet, with magnetic shunt and centering ring.

supply. In addition to these centering adjustments, moving the focus magnet off the center axis of the kinescope, or tilting the magnet, changes the centering of the electron beam. Moving the focus magnet horizontally changes the vertical positioning of the electron beam; moving the magnet vertically changes the horizontal positioning.

Kinescopes using electrostatic focus and magnetic deflection must have a separate centering magnet, when no d-c centering current through the yoke is provided, because there is no magnetic focus field that can be used for centering. The centering magnet is mounted on the neck of the kinescope about 3/4 in. behind the yoke, as shown in Fig. 9-5. Two wire rings in the magnet can be rotated to center the picture horizontally and vertically.

**9-2. Ion Spot in Magnetic Picture Tubes.** When magnetic deflection is used for the kinescope, a brown spot can form at the center of the screen because of bombardment by ions. This circular brown area is called an

*ion spot.* It has a diameter of about 1 in. on a 20-in. screen. The spot is produced by negative ions emitted from the cathode, which have the same charge as electrons but a much greater mass. The ions are part of the beam current but they are not deflected as much as the electrons when magnetic deflection is used. As a result, the center area of the screen is constantly bombarded by the ions, causing a brown burned spot that is insensitive and does not produce any light.

Since the ions have a very large mass, compared with the electrons, their velocity is low and there is little ion current in the beam. The magnetic field associated with this small value of ion current is weak.

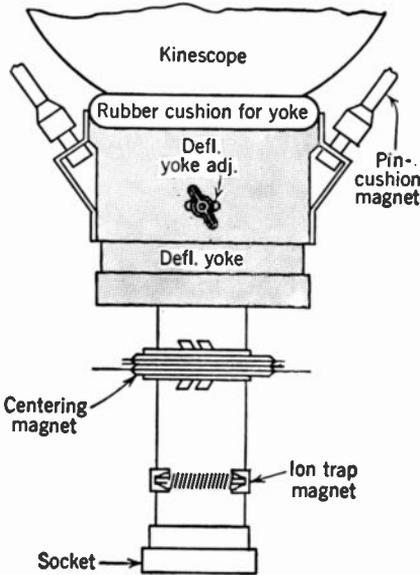


FIG. 9-5. Centering magnet mounted on kinescope.

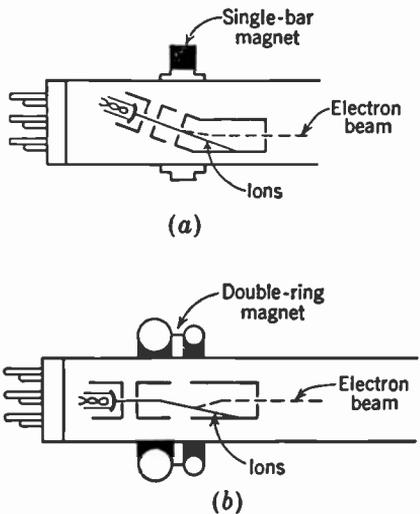


FIG. 9-6. Ion-trap arrangements. (a) Bent-gun. (b) Slashed-field type of electron gun.

The resultant force produced by the magnetic deflecting field is small, therefore, and there is very little deflection of the ions. The electrons in the beam are deflected by the required amount, however, because their velocity is high, producing a relatively large value of current and a strong magnetic field to react with the deflecting field. In electrostatic deflection, the amount of deflection is independent of the mass of the particles in the electron beam and the ions are deflected to the same extent as the electrons. No ion spot is produced with electrostatic deflection, therefore, because the energy of the bombarding ions is distributed over the entire screen.

The most common method of eliminating the ion spot in magnetic-deflection tubes employs an ion-trap arrangement to prevent the ions

from reaching the screen. The total cathode beam current, including electrons and ions, is made to leave the cathode at such an angle that the beam would strike the side of the tube instead of the screen, as illustrated in Fig. 9-6. A steady magnetic field introduced near the cathode then alters the path of the electrons only, so that the electron beam can be focused and deflected on the screen. The ions strike the side of the anode where they become part of the anode waste current that never reaches the screen. The beam current from the cathode is aimed off the screen by using either a bent gun as in *a* of Fig. 9-6 or a slashed field in *b*, which provides an electrostatic field that deflects the total beam current away from the axis of the tube.

The magnetic field required to deflect the electron beam back to the screen is provided by a small external magnet, which is called the *ion-trap*

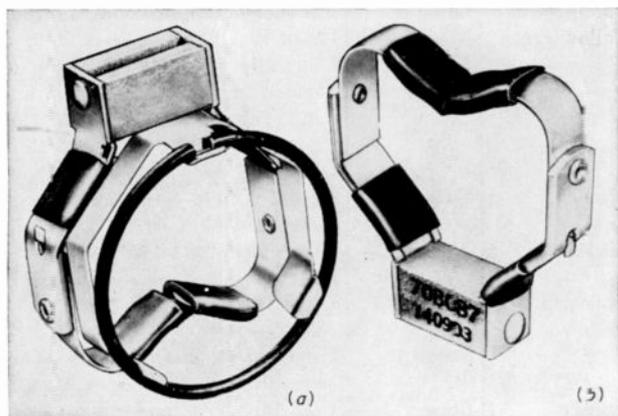


FIG. 9-7. Ion-trap magnets. (a) Double-magnet type. (b) Single-magnet type. (Clarostat Mfg. Co., Inc.)

*magnet*. This magnet provides a constant magnetic field with lines of force perpendicular to the beam axis. By rotating the ion-trap magnet, its transverse magnetic field can be positioned to provide the deflection needed to move the electron beam back toward the center. The magnetic field has no substantial effect on the ions and they do not reach the screen. The ion-trap magnet can be a coil magnet supplied with direct current for the steady magnetic field, but pm ion-trap magnets are most common. Figure 9-7 shows two types of ion-trap magnet. Such a magnet is also called a *beam bender*. The required field strength of a single-bar ion-trap magnet for a bent-gun arrangement is about 30 to 50 gaussses<sup>1</sup> with anode voltages of 10 to 16 kv. The single-bar ion-trap magnet is generally used with the bent-gun arrangement, which is most common in picture tubes.

<sup>1</sup> The gauss is an electromagnetic unit of magnetic-field strength. One gauss equals one magnetic line of force per square centimeter.

Figure 9-2 shows how the ion-trap magnet is mounted on a kinescope using magnetic deflection. Note that the smaller and weaker magnet is at the front, toward the screen. The magnet is placed approximately over grid 2 of the electron gun, near the tube base. By moving the magnet along the length of the tube about  $\frac{1}{4}$  in. forward or backward and rotating it slightly at the same time, the position can be found that bends

TABLE 9-1. SCREEN PHOSPHORS FOR CATHODE-RAY TUBES

Phosphor number	Color	Persistence	Remarks
P1	Green	Medium	Generally used for cathode-ray tube in oscilloscopes
P2	Blue-green	Long	For special oscilloscopes and radar receivers
P3	Yellow-green	Medium	Obsolete
P4	White	Medium	Generally used for television picture tubes. White is produced by combination of blue and yellow fluorescence. Has more blue than yellow to produce bluish-white like daylight
P5	Blue	Very short	Used for high-speed photography of oscilloscope traces
P6	White	Medium	Flat white, for picture tubes used with color filters in mechanical color television system
P7	Blue-white	Short	Used for radar receivers. Two-layer (cascade) screen
	Yellow	Long	
P11	Blue	Short	Used in oscilloscopes for visual or photographic observation
P12	Orange	Long	Used for radar receivers
P14	Blue	Short	Used for radar receivers. Two-layer (cascade) screen
	Red-orange	Long	
P15	Blue-green and near ultraviolet	Very short	Used as flying-spot scanner for camera pickup
P16	Near ultraviolet	Very short	Used as flying-spot scanner for camera pickup
P22	Red, green, and blue	Medium	Used for color television picture tubes

the electron beam back to the screen. A small misadjustment of the position of the ion-trap magnet, or reversed direction of the magnetic field, will throw the electron beam and the scanning raster off the screen, resulting in no brightness. The ion-trap magnet should be adjusted exactly for maximum brightness on the screen.

**9-3. Screen Phosphors.** The inside surface of the wide glass face of the picture tube is coated with a chemical phosphor, which has the luminescent property of emitting light when bombarded with electrons,

thereby converting the electrical signal into visible light. Light radiating from the screen as it is excited by the electron beam is called *luminescence*. When the radiated light is extinguished practically instantaneously after the electron-beam excitation has ceased, the screen is *fluorescent*. Continued emission of light after excitation is called *phosphorescence*.

Table 9-1 lists the screen phosphors and their characteristics. Some persistence of screen illumination after excitation by the electron scanning beam has ceased is desirable for a picture tube because this increases the brightness. However, the persistence time must be less than the frame period of  $\frac{1}{30}$  sec, so that one frame of the image cannot persist into the next frame to cause blurring of moving objects in the picture. For picture tubes, therefore, it is usually desired that the screen produce white light with a medium persistence, as typified by the P4 phosphor. In the P4 phosphor, after removal of excitation, the blue component of the emitted light decays to 1 per cent of its initial value in about 0.005 sec, and the yellow component has a decay time of about 0.06 sec.

The phosphors used for cathode-ray tube screens are compounds of the light metals such as zinc, cadmium, and calcium. Screens are generally made with several phosphors, which are combined to provide the desired color and persistence characteristics. The phosphors making up the screen material are ground until very fine particles of uniform size are obtained, and the material is applied to the glass face of the tube as a uniform layer. The phosphor materials forming the screen are good electrical insulators. The useful life of the screen depends on the phosphors used and the operating conditions of the tube, but usually is 500 to 2,000 hr. with normal beam current. Discoloration and a gradual decrease in brightness are indications of the end of the useful life of the screen.

**9-4. Metal-backed Screens.** Some picture tubes have a smooth and very thin metal coating, usually aluminum, applied to the back surface of the luminescent screen. The metallic film is thin enough to allow the electron beam to pass through to the screen when anode voltages of 10,000 volts or more are employed. Aluminum is commonly used for the metal backing because it is readily vaporized and deposited as a film. It reflects light well, allows the electron beam to pass through fairly easily, and does not react with the phosphor to spoil the screen characteristics. The aluminum film is very thin, but is sufficient to provide a light-reflecting surface and an electrical conducting plate on the back of the screen.

In ordinary cathode-ray tubes the return path for the electron-beam current that strikes the screen is obtained by secondary emission of electrons from the screen. These secondary electrons are collected by the second anode, allowing the screen to charge to approximately the second-anode potential. The amount of secondary emission from the screen is

limited, however. When very high anode voltages are used, therefore, the screen may not be able to charge to the anode potential, remaining at the highest positive potential to which it can charge. This is the *sticking potential* of the screen, and it is not possible to have a higher effective gun voltage for the tube than this value. As an example, if the second-anode voltage for a tube is 15 kv but the sticking potential of the screen is 10 kv, the effective gun voltage will be only 10 kv because this is the potential difference between the screen and cathode. Such sticking of the screen potential may be a serious disadvantage in picture tubes employing high anode voltages of about 10 to 80 kv to obtain enough screen brightness. With a metal-backed screen, however, no trouble is experienced with sticking of the screen potential. The screen potential is essentially the same as the gun voltage, even with extremely high anode potentials, making it possible to obtain very high values of screen brightness. The metalized screens are used only with tubes employing anode voltages of 10 kv or more, since there is little trouble with sticking potential for conventional tubes operated with lower anode voltages. Above 10,000 volts the energy loss due to the metal film is relatively small and the screen is at essentially the same potential as the second anode.

There are several other advantages provided by the metal-backed screen. The ion spot is eliminated without the need for an ion trap because the slowly moving ions cannot pass through the metal film to strike the screen. The conductive coating also makes the screen insensitive to stray capacitance effects. Finally, the metal-backed screen provides an important optical advantage for picture tubes. Light from the screen that would radiate toward the inside of the tube is reflected back from the smooth metal reflecting surface, increasing to a great extent the amount of light available in front of the screen. This eliminates the internal reflections, increasing the brightness and contrast in the reproduced picture. The metal back effectively doubles the brightness of the picture, compared with an unmetalized screen. The metal-backed screen is used in the projection kinescopes shown in Fig. 9-8.

**9-5. Types of Picture Tubes.**<sup>1</sup> Table 9-2 lists some of the kinescope tube types commonly used in television receivers, in order to illustrate typical kinescope characteristics.

*Type Designation.* The type designation for each tube consists of three groups of symbols. First is a number of one or two digits giving the approximate screen size by specifying the maximum dimension of the faceplate in inches, within  $\frac{1}{2}$  in. For a round tube this is the diameter; on a rectangular face the maximum dimension is the diagonal, as illustrated in Fig. 9-9. The second symbol is a letter assigned by RETMA, alphabetically in the order of registration for different tube types. Tubes

<sup>1</sup> Detailed data on kinescopes are given in manufacturers' manuals for receiving tubes.

with the same letter designation do not necessarily have the same characteristics. For instance, the 5TP4 projection kinescope has a different electron-gun structure than the 17TP4 direct-view kinescope. The third symbol in the type designation consists of the letter P and the number of the screen phosphor. As seen from the table, all the picture tubes use the P4 phosphor, which has a white fluorescence with medium persistence. Another letter may be added to the type designation, as in the 19AP4-A,

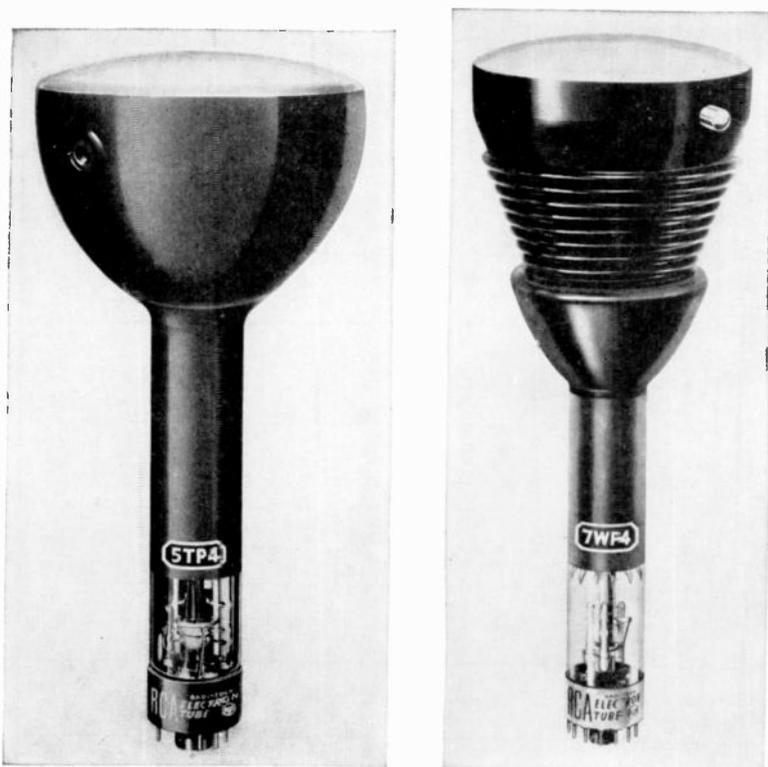


FIG. 9-8. Projection kinescopes 5TP4 and 7WP4. (RCA.)

to indicate there has been a change in the design of the tube type but not enough to justify a new type number. The new tube with the added letter will always operate in the same circuits as the original tube but the old tube may or may not work in all circuits where the new tube is used. The letter A on many oscilloscope cathode-ray tubes indicates a lower first-anode current than the original design. With picture tubes the added letter often designates only a different kind of glass faceplate.

*The Faceplate.* Picture tubes are made with a practically flat face and screen. The use of a glass faceplate approximately  $\frac{1}{2}$  in. thick provides the mechanical strength required to withstand the air pressure on the

TABLE 9-2. PICTURE TUBES

Type	Construction	Base	Length, inches	Deflection angle, degrees*	Deflection method†	Focus method†	Heater	Approx values for typical operating voltages				Remarks
								Anode volts	Focusing-grid volts	Accelerating-grid volts (grid 2)	Control-grid cutoff (grid 1)	
5TP4	Glass, round	Duodecal 7-pin	11½	50	M	E	6.3 volts, 0.6 amp	27,000 (anode 2)	5,000 (anode 1)	200	-42 to -98	Projection kinescope; has metal-backed screen
7JP4	Glass, round	Dihedral 12-pin	14½	..	E	E	6.3 volts, 0.6 amp	6,000 (anode 2)	2,000 (anode 1)	...	-75 to -168	Electrostatic tube; deflection factor for each pair of deflecting plates 36 and 30 volts per in. per kv of $E_{s1}$ ; grid 2 and grid 4 connected internally to second anode
7WP4	Glass, round	Dihedral 14-pin	19½	35	M	E	6.6 volts, 0.6 amp	75,000 (anode 2)	17,000 (grid 3)	500	-155	Projection kinescope for theater television equipment, has metal-backed screen
10BP4	Glass, round	Duodecal 5-pin	17¾	52	M	M	6.3 volts, 0.6 amp	9,000	.....	250	-27 to -63	10BP4 has clear faceplate; 10BP4A, gray
12LP4	Glass, round	Duodecal 5-pin	18¾	54	M	M	6.3 volts, 0.6 amp	11,000	.....	250	-27 to -63	12LP4 has clear faceplate; 12LP4A, gray
14EP4	Glass, rectangular	Duodecal 5-pin	16¾	65	M	M	6.3 volts, 0.6 amp	12,000	.....	300	-33 to -77	Clear faceplate
15GP22	Glass, round	Bidocal 14-pin	28¾	45	M	E	6.3 volts, 1.8 amp	19,000	4,000	200	-45 to -100	Three-gun triorion kinescope for color television receivers; convergence voltage 9kv
16AP4	Metal, round	Duodecal 5-pin	22½	53	M	M	6.3 volts, 0.6 amp	12,000	.....	300	-33 to -77	16AP4 has clear faceplate; 16AP4A, gray; direct substitute for 10BP4
16KP4	Glass, rectangular	Duodecal 5-pin	18¾	65	M	M	6.3 volts, 0.6 amp	14,000	.....	300	-33 to -77	16KP4 has gray faceplate; 16KP4A, gray, metal-backed screen
17TP4	Metal, rectangular	Duodecal 6-pin	19½	70	M	E (low-voltage)	6.3 volts, 0.6 amp	14,000	0 to 350 (grid 4)	300	-33 to -77	Gray frosted faceplate; grid 5 and grid 3 connected internally to anode
19AP4	Metal, round	Duodecal 7-pin	21½	66	M	M	6.3 volts, 0.6 amp	13,000	.....	250	-27 to -63	19AP4 has clear faceplate; 19AP4A, gray; 19AP4B, gray, frosted; 19AP4C, gray, metal-backed screen; 19AP4D, frosted
20CP4	Glass, rectangular	Duodecal 5-pin	21¾	66	M	M	6.3 volts, 0.6 amp	12,000	.....	300	-33 to -37	Gray faceplate; 20CP4A has gray faceplate also but uses typical anode voltage of 16,000 volts
20TP4	Glass, rectangular	Duodecal 5-pin	21¾	66	M	E (self-focus)	6.3 volts, 0.6 amp	14,000	0 (grid 4)	300	-33 to -77	Gray faceplate; grid 5 and grid 3 connected internally to anode
21MP4	Metal, rectangular	Duodecal 6-pin	22¾	70	M	E (low-voltage)	6.3 volts, 0.6 amp	14,000	-55 to +300 (grid 4)	300	-33 to -37	Gray frosted faceplate; grid 5 and grid 3 connected internally to anode
24AP4	Glass, round	Duodecal 5-pin	23¾	70	M	M	6.3 volts, 0.6 amp	15,000	.....	300	-33 to -37	24AP4 has gray faceplate, 24AP4A, gray with aluminumized screen
27EP4	Glass, rectangular	Duodecal 5-pin	23½	90	M	M	6.3 volts, 0.6 amp	20,000	.....	300	-28 to -72	Aluminized screen; gray faceplate; 27EP4 is same kinescope with external conductive coating

\* Diagonal deflection angle given for rectangular tubes.

† E is electrostatic; M is magnetic.

vacuumed envelope. With a flat face, the tube can accommodate a larger picture for a given diameter, as optical distortions caused by curvature of the screen are reduced. The thick glass also makes it feasible to construct the picture tube envelope with a rectangular face-plate. Most picture tubes now are made with a rectangular screen because this has the advantages of eliminating the wasted screen area of a round tube and providing more compact installation in the receiver cabinet. As shown in Fig. 9-9a, a rectangular raster with the 4:3 aspect ratio wastes part of the circular screen at the top, bottom, and both sides. Usually, the width adjustments in the receiver are set to make the raster reach the sides of the circular screen, as in *b*, although the corners of the raster are then off the screen. The rectangular screen accommodates the full raster without any wasted screen area, as shown in *c*.

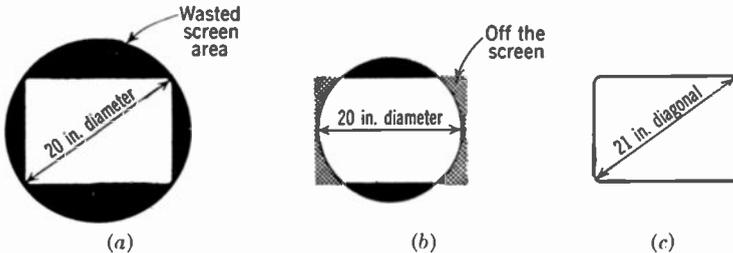


FIG. 9-9. Size of raster on round and rectangular screens. (a) Rectangular raster on round screen. (b) Full-width raster on round screen. (c) Rectangular raster filling rectangular screen.

The faceplate can be manufactured with a neutral light-absorbing material in the glass, making it gray, in order to increase the optical contrast in the picture by reducing the room lighting's reflections from the screen. The gray glass is an effective filter for reducing the intensity of the undesired reflections because these light rays must go through the dark glass at least twice before being reflected outward to the viewer, while direct light from the phosphor passes through the faceplate only once. Compared with a clear glass faceplate, approximately 65 per cent of the light from the screen is transmitted through the gray faceplate. When the screen brightness can be made high enough to allow for this attenuation, however, the contrast is then increased by about the same amount as the loss of light through the filter. In addition to the use of gray glass, the faceplate can be etched or frosted, like a frosted light bulb. The frosted faceplate diffuses reflections of bright objects in the room, which might otherwise be objectionable when reflected from dark areas of the picture. The faceplate types available for some common kinescopes are noted in the "Remarks" column in Table 9-2.

*Tube Base.* Kinescopes using magnetic deflection generally have the small-shell, *duodecal* (12-pin) base, which has a diameter of  $1\frac{7}{16}$  in.

When the spacing is for 14 pins the base is called *diheptal*. The *bilecal* base is for 20 pins; A *magnol* base is for 11 pins. Figure 9-10 shows kinescope socket connections for several tube bases. The number given with the type of base states how many pins are actually on the base.

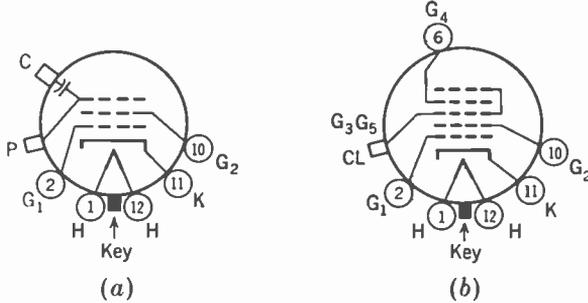


FIG. 9-10. Common duodecal base and socket connections for magnetic-deflection kinescopes, bottom view. (a) 20CP4A, 5-pin. Typical for tubes with magnetic focus or self-focus. (b) 17TP4, 6-pin. Typical for tubes with low-voltage electrostatic focus. The nomenclature is as follows:

- |                                   |                                      |
|-----------------------------------|--------------------------------------|
| H—Heater                          | P <sub>1</sub> —First anode          |
| K—Cathode                         | P <sub>2</sub> —Second anode         |
| G <sub>1</sub> —Control grid      | P—Anode                              |
| G <sub>2</sub> —Accelerating grid | CL—Metal-shell lip (anode connector) |
| G <sub>3</sub> —Ultor grid        | C—External conductive coating        |
| G <sub>4</sub> —Focusing grid     | IC—Internal connection (do not use)  |
| G <sub>5</sub> —Ultor grid        |                                      |

For instance, the duodecal 5-pin base has space for 12 pins but only pins 1, 2, 10, 11, and 12 are used.

*The Anode.* In glass kinescopes the anode is a conductive coating on the inside wall of the tube. Because of the high voltage, a separate anode



FIG. 9-11. Round glass picture tube, type 12LP4. Note the external conductive coating on the glass bulb, which is insulated from the anode connection. (Sylvania Electric Products, Inc.)

connector on the side of the cone is used, instead of connecting to the tube base. This is usually a small recessed hole of about 1/4 in. diameter into which a ball-spring connector fits to contact the anode coating on the inside wall. As shown in Fig. 9-11, glass kinescopes usually also have an external conductive coating on the cone, which is insulated by the glass

from the anode wall coating inside. The outer coating must be connected to chassis ground. The ground connection is generally made by spring contacts that brush against the coating. This provides a filter condenser of about 500 to 2,500  $\mu\text{f}$  for the anode high-voltage supply, with the external coating as one plate, the anode coating the other plate, and the glass bulb the insulator between the two plates of the condenser.

For metal tubes the metal cone is the anode. In the construction of a metal tube the faceplate, the narrow elbow containing the electron gun, and a small part of the cone are glass but these are made individually and then sealed to the metal shell, which is a chromium-iron alloy. Figure 9-12 shows a metal kinescope. The advantage of this metal-envelope construction is its lighter weight, compared with an all-glass



FIG. 9-12. Metal-cone kinescope. (RCA.)

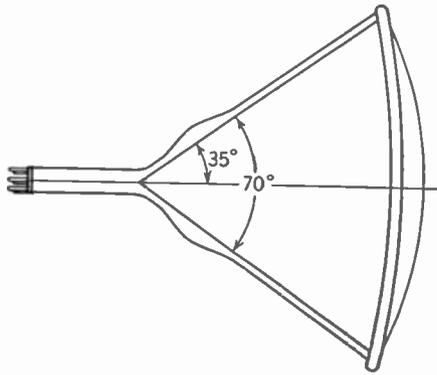


FIG. 9-13. Kinescope with deflection angle of 70°.

tube, especially for large tubes with a screen diameter of 16 in. or more. For the high-voltage connection on metal kinescopes, a metal strip contacts the metal rim around the faceplate.

**Deflection Angle.** This specifies the maximum angle through which the beam can be deflected to fill the screen without striking the sides of the tube. The deflection angle for typical kinescopes is 50 to 90°. This refers to the total angle. As shown in Fig. 9-13, a deflection angle of 70° means the electron beam can be deflected 35° from the center line of the tube. For rectangular kinescopes the maximum deflection angle is between two diagonally opposite corners of the screen. A kinescope with a wide deflection angle has a shorter bulb, allowing installation of the kinescope in a smaller receiver cabinet.

**Low-voltage Electrostatic Focusing.** In a kinescope using magnetic deflection and electrostatic focusing like the 5TP4, the focusing electrode requires about one-fifth the anode high voltage. In this type of electron gun, the beam is accelerated to one-fifth the high voltage and then

focused by further acceleration to the full anode voltage. A later type of electron gun that uses lower voltage for the focusing grid is illustrated in Fig. 9-14. Grid 1 is the control grid, grid 2 is the accelerating grid, grids 3 and 5 are connected internally to the anode, and grid 4 is the focusing electrode. With this deceleration type of electron lens the B supply can provide the focusing voltage, which may be between  $-30$  and  $300$  volts, approximately. In a self-focus kinescope, the focusing grid connects internally to the cathode through a resistor, and no focusing adjustment is needed.

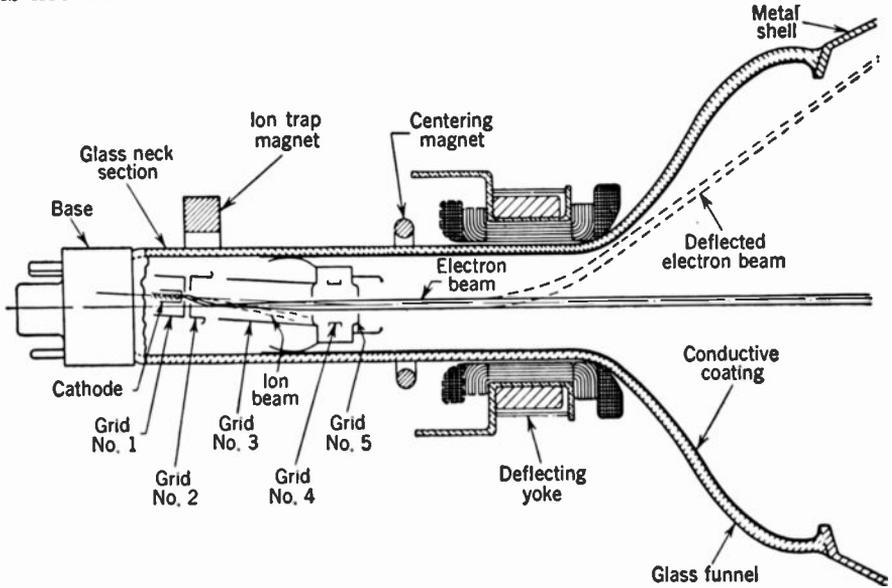


FIG. 9-14. Magnetic-deflection kinescope using low-voltage electrostatic focusing. Note the internal connection of grids 3 and 5 to the anode, and the position of the centering magnet. (RCA.)

Grid 3 and grid 5 in Fig. 9-14, which connect internally to the anode, are indicated collectively for convenience by the term *ultor*. The *ultor* is the part of the electron gun that has the highest accelerating voltage prior to deflection of the beam.

**9-6. Control-grid Characteristic Curves of Picture Tubes.** As shown in Fig. 9-15, the transfer characteristic curves of picture tubes are similar to the grid-plate characteristic curves for amplifier tubes. The kinescope control grid has a fixed negative bias and the video signal voltage varies the instantaneous grid voltage, varying the amount of beam current and the light from the screen. At cutoff, the grid voltage is negative enough to reduce the beam current to a value low enough to extinguish the beam, which corresponds to the black level. The parts of the screen without

any luminescence look black in comparison with the adjacent white areas. The cutoff bias for kinescopes is about 50 volts, ranging from -33 to -77 volts, depending upon the control grid-cathode spacing.

As the a-c video signal swings in the positive direction to make the control grid less negative, the beam current increases. The slope of the steep part of the kinescope characteristic curve is approximately 2.2, providing a gamma of 2.2. It should be noted that the values of the grid voltages shown in Fig. 9-15 are positive signal volts applied in addition to

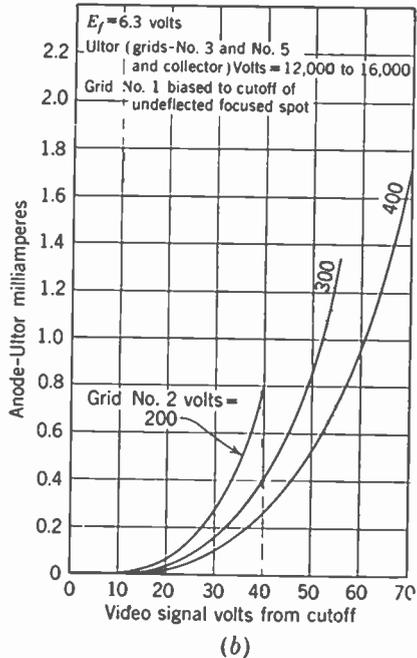
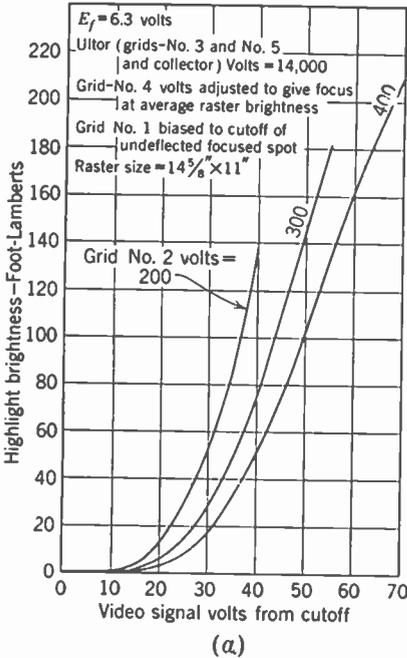


FIG. 9-15. Average grid-drive characteristics of 17TP4 kinescope. (a) Highlight brightness vs. grid drive. (b) Anode-Ultor current vs. grid drive. (RCA.)

the cutoff bias voltage. For a value of grid-driving voltage 20 volts positive with respect to a typical cutoff bias of -50 volts, the actual grid voltage becomes -30 volts. Referring to Fig. 9-15 the corresponding beam current 20 volts from cutoff is approximately 50  $\mu$ a, with a grid 2 voltage of 300 volts, and the highlight brightness is about 7 ft-lamberts.<sup>1</sup> For a peak value of video signal voltage driving 45 volts more positive than cutoff, the instantaneous grid voltage is -5 volts, the anode current is slightly less than 600  $\mu$ a, and the highlight brightness is a little more than

<sup>1</sup> The foot-lambert is a unit of brightness; 3.14 ft-lamberts equals the brightness value of one candle per square foot of light-emitting surface.

110 ft-lamberts. Projection kinescopes, using an anode voltage of 20 to 80 kv, produce highlight brightness values of 4,000 to 30,000 ft-lamberts, with anode current of 1 to 6 ma. When the video signal voltage of varying amplitudes is applied to the control grid of the kinescope, the beam current and screen illumination change with the signal variations to reproduce the desired picture information.

*Video Signal for the Kinescope.* With a cutoff voltage of  $-50$  volts, in order to use the operating range of the kinescope for full contrast the amount of composite video signal required is approximately 70 volts, peak to peak, including the sync-pulse amplitude. The video signal can be applied either to the control grid of the kinescope or to the cathode. In either case, the white parts of the video signal must make the instantaneous kinescope grid voltage less negative than the bias, with respect to the cathode voltage, producing more beam current and increased light.

*Light Distortions.* Several distortions of the signal-to-light relationship may be produced in the tube. One is a faint halo of light around the scanning spot, which is called *halation*. Some of the light radiated from the spot of light produced by the beam is reflected internally at the front and back surfaces of the glass faceplate. The internal reflections provide secondary sources of light that produce a halation ring around the scanning spot. Light from the screen can also radiate backward into the tube when the screen is not aluminized, producing *scattering* of the light from the scanning spot. Since the light is brightest on the back surface where the electrons strike the screen, the scattering of the light causes reflections from the glass walls and the electron gun in the kinescope. Both the halation and scattering effects reduce the optical contrast in the picture, as they increase the light from black parts of the picture. Another distortion of the light spot is called *blooming*. This is an increase in the size of the scanning spot produced by defocusing when the beam current is excessive for the amount of accelerating voltage used. Blooming occurs with peak positive signal voltages on the control grid, resulting in loss of detail in the highlights of the picture.

**9-7. Picture Tube Precautions.** The anode voltage for picture tubes is usually 9,000 volts or more. Therefore, the safety precautions important in working with high-voltage equipment should be observed. Before any part of the kinescope is touched, the power should be turned off and the high-voltage filter condensers discharged. Picture tubes may be operated with anode voltages up to 16 kv without producing harmful X-ray radiation and without danger of personal injury on prolonged exposure at close range. Above 16 kv, special X-ray shielding precautions may be necessary.

Picture tubes should be handled more carefully than smaller vacuum tubes. The kinescope bulb encloses a high vacuum and, because of the

large surface area, the envelope has a strong force on it produced by air pressure. For a typical 20-in. rectangular picture tube, the surface area is about 1,000 sq in., and with an air pressure of approximately 15 lb per sq in. the total force on the bulb is 15,000 lb. The envelope is strong enough to withstand this if there are no flaws and the tube is handled gently. Do not strike or scratch the tube. When installing a kinescope, if the tube sticks and does not slip into its mounting smoothly, investigate and remove any obstacles. Do not force the tube.

The envelope of the kinescope should be clean and dry to keep the anode well insulated. Dust on the glass faceplate of the kinescope can reduce the brightness appreciably. To form an airtight dust seal, usually a rubber strip is placed around the face of the kinescope where it fits into the mask on the front panel of the receiver cabinet. The mounting is illustrated in Fig. 9-16.

**9-8. Picture Tube Requirements.** In order to reproduce a picture, the kinescope must be supplied with the necessary operating potentials and video signal voltage by the receiver. These requirements include:

*Heater Power.* This is needed to heat the cathode to produce electrons for the beam current. Most kinescopes require a heater voltage of 6.3 volts at 0.6 amp. If the heater is not lighted, there will be no beam current and no screen brightness.

*Ion-trap Magnet.* Magnetic-deflection kinescopes use an ion-trap magnet. If it is not positioned correctly or if the field of the magnet is too weak, the electrons in the beam will strike the sides of the tube and there will be no screen brightness.

*High Voltage.* A separate high-voltage supply is necessary to provide the anode voltage for the picture tube. The anode voltage for 10- to 27-in. direct-view picture tubes is about 9 to 18 kv, with the higher anode voltages used to produce sufficient brightness with the larger screens. Projection kinescopes require an anode voltage of 20 to 80 kv. Without high voltage, there is no beam current and no brightness.

*Low-voltage Operating Potentials.* The low-voltage power supply provides negative d-c bias voltage, the positive accelerating grid voltage, and possibly focusing current or voltage for the kinescope. The amount of grid-bias voltage required is about -25 volts, with video signal input. Excessive negative bias can cut off the beam current, resulting in no brightness on the screen. Insufficient negative bias can cause excessive brightness, with blooming in the picture. In both cases, if the bias cannot be varied around its normal value, there is no control of brightness. The positive voltage for the accelerating grid is 200 to 400 volts. Referring to the kinescope characteristic curves in Fig. 9-15, it can be seen that higher values of accelerating grid voltage allow more screen brightness and a wider contrast range, but more video signal voltage is required. Cor-

rect operation of the picture tube, therefore, depends upon normal operation of the low-voltage power supply.

*Deflecting Current.* The deflection circuits in the receiver must produce the current required for the deflection coils in the yoke on the picture tube, in order to form the scanning raster.

*Video Signal.* With a scanning raster on the screen, it is finally necessary to couple the video signal to the control grid-cathode circuit of the kinescope, in order to vary the screen brightness in accordance with the signal variations and reproduce the desired picture. The amount of composite video signal voltage required for most picture tubes is about 70 volts, peak to peak. Without any video signal there will be no picture, although the scanning raster will be produced if all other voltages are normal.

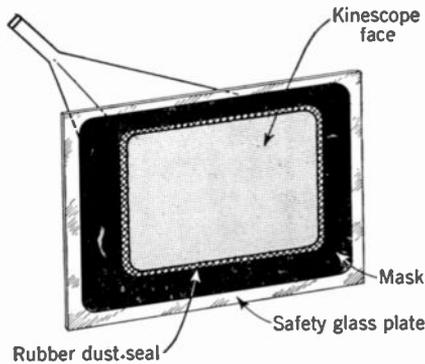


FIG. 9-16. A front-panel mounting for the picture tube. The safety-glass plate is on the outside, with the dust seal between the mask and kinescope.

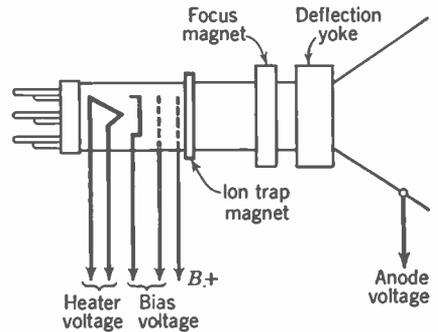


FIG. 9-17. Requirements for kinescope operation. The parts shown in dark lines are necessary to produce brightness on the screen.

These requirements are illustrated in Fig. 9-17. The heavy lines here indicate the parts that must be operating normally to have any brightness on the kinescope screen. These include all the components except the deflection yoke and the focus coil.

**9-9. Substituting Picture Tubes.** Aside from the mechanical problems of mounting the tube in the cabinet, a larger kinescope can usually be substituted for a smaller one in order to have a bigger picture. Only magnetic-deflection tubes are considered; conversion from electrostatic to electromagnetic deflection is not practicable. The conversion is simplified by using a kinescope that has approximately the same deflection angle and requires about the same anode voltage. For instance, the 16AP4 can be directly substituted for the 10BP4 because both have approximately the same deflection angle and operate with an anode voltage of 9 kv. With the same deflection angle, the deflection yoke can fill the screen on either tube. The filament power, socket, and pin

connections are the same for both tubes, as is the accelerating grid voltage. Both tubes can use the same ion-trap magnet and the same focusing coil. It should be noted that tubes of the same type designation that differ only in the kind of faceplate can be substituted without any circuit changes.

**9-10. Picture Tube Troubles.** Low emission from the cathode and an open heater are typical tube troubles. Operation of the heater can be checked by looking into the glass neck near the base of the kinescope to see if it is lighted. An open heater means no emission and no brightness. Low emission in the kinescope usually produces a distinctive weak picture, where the white parts of the picture have a silvery appearance and are as dark as the normally gray areas. It should be noted that an internal short in the kinescope from control grid to cathode, or cathode to heater, can result in no control of brightness, with a weak picture. A gassy picture tube causes blooming.

**Afterglow.** This is a luminous spot produced at the center of the kinescope screen just after the receiver is turned off. The afterglow results when the kinescope anode voltage remains on the high-voltage filter condenser while the cathode is still hot, after the deflection circuits stop operating. The remedy is to allow the high-voltage filter condenser to discharge more quickly, so that there will not be enough anode voltage to produce illumination when the receiver is off. Since the discharge path is through the cathode-anode beam current, less kinescope bias results in more beam current and a faster discharge. This is why turning up the brightness, to reduce the kinescope bias, just before or after switching the receiver off, eliminates the afterglow.

**Screen Burn.** If there is no deflection but the kinescope anode has high voltage, the scanning spot will rest at the center of the screen, producing a very bright spot of light that can burn the fluorescent coating of the screen. The result of burning the screen is a dark brown area which does not produce light when excited by the electron beam, making that part of the screen useless. Since the energy of the electron beam is concentrated on one spot without deflection, instead of being distributed over the entire screen area, the excessive excitation can destroy the fluorescent properties of the screen. If a spot or a line is seen on the screen, instead of the scanning raster, the brightness should be turned down or the power disconnected to avoid burning the screen. This is especially critical with projection kinescopes, which use very high anode voltages.

#### REVIEW QUESTIONS

1. What are the basic parts of a picture tube? Give the function of each.
2. How is a yoke for magnetic deflection mounted and adjusted?
3. How is a pm focus magnet mounted and adjusted?

4. Describe briefly three methods that may be used for centering the picture on the kinescope screen.
5. Describe briefly two methods that may be used for focusing the beam on the kinescope screen.
6. What is the ion spot? Give two methods of eliminating the ion spot.
7. Describe exactly how the ion-trap magnet is adjusted.
8. Why is the P4 phosphor generally used for picture tubes?
9. Give two advantages of an aluminized screen for a picture tube.
10. What characteristics are evident from the picture tube number 21MP4?
11. Give two types of glass faceplate available for kinescopes, with their advantages.
12. Draw the base and socket connections (bottom view) for the 20CP4 picture tube.
13. How is the high voltage generally connected to the anode in a glass picture tube? In a metal-cone picture tube?
14. What is the function of the external conductive coating on the bulb of a glass picture tube?
15. What is the advantage of a picture tube having a deflection angle of  $70^\circ$ , compared with a  $50^\circ$  tube?
16. When a larger picture tube is used in place of a kinescope with a smaller screen, explain briefly why the same receiver deflection circuits can fill the screen for either tube when both tubes have the same deflection angle.
17. What is the advantage of a kinescope using low-voltage electrostatic focusing, compared with high-voltage electrostatic focusing and magnetic focusing.
18. Define the *ultor* in a kinescope.
19. From the average characteristics of the 17TP4 kinescope in Fig. 9-15, plot the screen-brightness variations that would be produced with a sine-wave grid signal having a peak value of 40 volts and a frequency of 60 cps. The control grid is biased to cutoff. The accelerating grid potential is 300 volts.
20. Give two precautions to be observed when working with picture tubes.
21. If all operating potentials for the picture tube are normal but the heater is open, what will be seen on the kinescope screen?
22. List the kinescope requirements for producing screen brightness.
23. Can the scanning raster be produced without having brightness on the kinescope screen? Can brightness be produced on the kinescope screen without having a scanning raster?

## CHAPTER 10

### POWER SUPPLIES

Because of the different requirements in the amount of d-c power needed to operate the television receiver it generally has two power supplies. A conventional B supply provides 300 to 400 volts at about 200 to 300 ma for the plate and screen circuits of the amplifiers. The high-voltage supply produces about 9 to 18 kv anode potential for direct-view picture tubes, with a current drain of less than 1 ma. Since the high-voltage needs of the picture tube are so much different from the B supply requirements, a common power supply is not practicable.

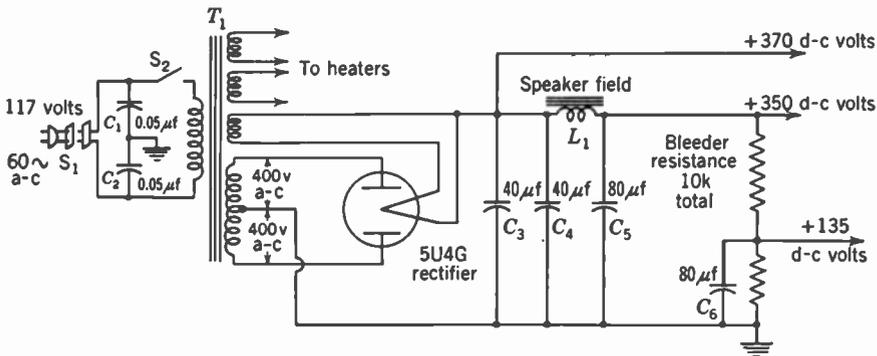


FIG. 10-1. Low-voltage power supply for a television receiver.

**10-1. Low-voltage Power Supply.** The schematic diagram of a typical low-voltage power supply is illustrated in Fig. 10-1. This is a conventional a-c operated B supply, using a step-up transformer from the 117-volt a-c power line and a full-wave rectifier. The high-voltage secondary of the power transformer furnishes about 400 volts from either side of the winding to the center tap, for the two diodes of the 5U4-G full-wave rectifier tube, providing a d-c voltage output of 370 volts at about 200 ma. In the a-c input  $C_1$  and  $C_2$  bypass both sides of the primary to ground to filter out r-f interference from the power line. The safety interlock  $S_1$  is generally part of the receiver's back panel, which must be

on to complete the a-c input circuit when the on-off switch  $S_2$  is closed. Included in the power transformer are step-down secondary windings for the filaments and heaters of all tubes in the receiver. The rectifier filament winding is separate because the B plus voltage is present here. This winding must be insulated well enough to withstand the B+ voltage to ground without arcing.

In the d-c output circuit the smoothing choke  $L_1$  is the loudspeaker field coil; therefore the speaker cable must be plugged into its socket to complete the B+ output circuit. Values for the filter choke are about 5 henrys inductance and 100 ohms d-c resistance. The large filter condensers  $C_3$ ,  $C_4$ ,  $C_5$ , and  $C_6$  are needed because the television receiver's B supply must have excellent filtering. The percentage of a-c ripple voltage in the d-c output is normally about 1 per cent. Good filtering is necessary because excessive hum voltage in the raster and signal circuits affects

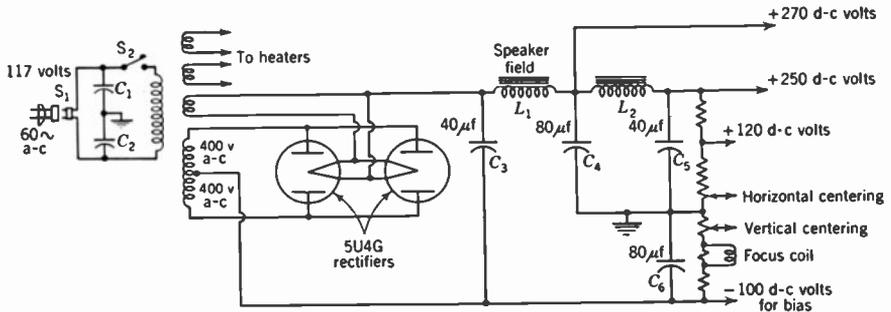


FIG. 10-2. Low-voltage power supply with two full-wave rectifiers in parallel.

the reproduced picture. In addition, low frequencies such as 30 to 60 cps must be amplified without any mutual coupling between stages through the B supply filter circuit as a common impedance.  $C_3$  and  $C_4$  in parallel provide the input filter capacitance, which is effective in maintaining a high d-c voltage output. The output filter condenser  $C_5$  bypasses the a-c ripple around the d-c load circuits. A wire-wound bleeder resistor is connected across the output terminals of the supply to stabilize the output voltage and serve as a voltage divider to tap off the different voltages required in the receiver.  $C_6$  is an additional filter condenser for the load circuits connected to the 135-volt tap.

Receivers that require about 300 ma d-c current from the B supply often use two full-wave rectifiers in parallel, as shown in Fig. 10-2, doubling the permissible load current. The two full-wave rectifiers supply a total d-c output of 370 volts at 300 ma, including +270 volts and -100 volts with respect to ground. A two-section filter is used, as a higher value of load current corresponds to a lower load resistance, which requires more filtering. The voltage-divider connections illustrate supply



d-c supply voltage. The full-wave rectifier  $V_{501}$  provides +190 volts for the remainder of the signal circuits. The focus coil  $L_{404}$  is used as a filter choke.  $L_{501}$  and  $L_{502}$  are conventional chokes, as the receiver uses a pm loudspeaker. It is important to note that the plate load circuit for each stage is omitted in Fig. 10-3, which shows just the distribution of the B supply voltages. There is no fuse for the low-voltage power supplies, but the fuse  $M_{401}$  protects the horizontal deflection circuits and

high-voltage supply. This is general practice in television receivers. Note also that the triangle and half-circle markings on  $C_{307}$ , which are at the connecting lugs, indicate individual sections. Figure 10-4 shows such a multiple-section electrolytic filter condenser.



FIG. 10-4. Multiple-section, electrolytic filter condenser can for mounting on chassis. (Cornell-Dubilier Electric Corp.)

*Voltage Polarity with Respect to Ground.* Voltages both positive and negative with respect to ground are available on the voltage divider in Fig. 10-2 because of the ground connection. The effect on the polarity of the output voltage for different ground points on a voltage divider is illustrated in Fig. 10-5. The chassis ground potential is generally used as the voltage reference, which is zero volts, but this is not necessarily the most negative point of the power supply. The center tap on the high-voltage secondary, or any other return connection on the winding, is the most negative point. When this is connected to ground, as in *a* of Fig. 10-5, all points on the voltage divider are positive with respect to ground. In *b* point *C* on the divider is grounded and point *D*, which is the negative return, becomes negative with respect to ground. This is often done to obtain negative

voltages for grid bias and other uses. In *c* point *A* is grounded and all voltages on the divider are negative with respect to ground because the most positive point in the supply is connected to ground. This can be done as long as no other point is grounded. Such an arrangement is used for an *inverted power supply*, which provides d-c output voltage negative with respect to ground.

It is important to note the arbitrary nature of assigning polarity. The power supply will operate in exactly the same way if there are no ground connections. Connecting any one point on the power supply to ground has no effect on operation of the supply but merely provides output volt-

age having the desired polarity with respect to the chassis ground, which is a good reference point because it is a convenient return connection for stages on the chassis. For the connections shown in Fig. 10-5, point *A* will always be more positive than *B*, *C*, or *D*, and *D* will always be the most negative point on the supply, regardless of where the ground connection is made.

**10-2. Filament Circuits.** The filament windings on a typical power transformer for a television receiver are illustrated in Fig. 10-6. Several

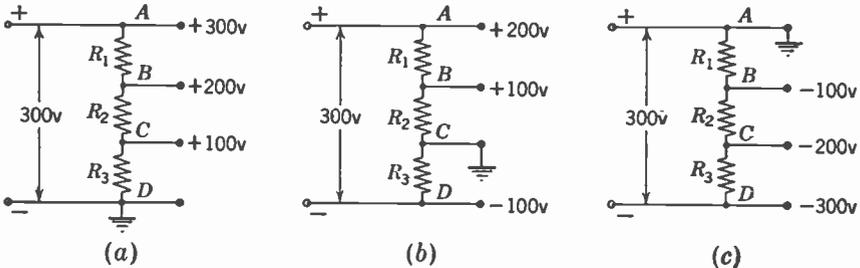


FIG. 10-5. Effect of moving the ground point on a voltage divider.

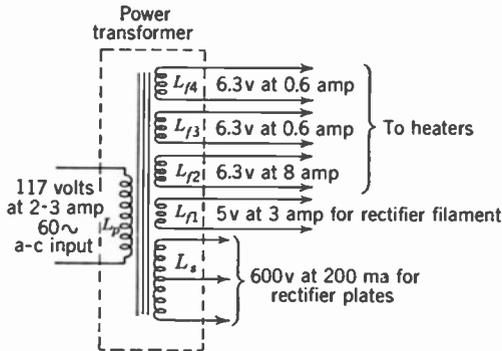


FIG. 10-6. Filament windings on low-voltage power-supply transformer.

secondary windings are usually required.  $L_{f1}$  is for only the rectifier filament, which needs a separate winding because it has the  $B+$  potential with respect to ground. The tube heaters<sup>1</sup> that have no d-c potential and can be connected to ground, which include practically all the tubes in the receiver, are in parallel across  $L_{f2}$ . However, the kinescope and damper tubes may have a high d-c cathode potential, requiring the heater and cathode to be connected together in order to eliminate arcing between these elements in the tube. In this case separate filament windings are necessary, such as  $L_{f1}$  and  $L_{f3}$ , which must have good insulation from ground.

<sup>1</sup> A filament is a directly heated cathode, which emits electrons. With an indirectly heated cathode, the heating element is generally called a heater.

*Parallel Heaters.* Figure 10-7 shows how the heaters are connected in parallel across the filament winding of the power transformer. The filament chokes  $L_1$  and  $L_2$  shown in *a* are small r-f coils for decoupling between stages to prevent feedback of r-f or i-f signal through the common filament line. They have high impedance for r-f signal but practically none for the 60-cps filament current. Therefore, the heaters are effectively in parallel with each other, as illustrated in *b*. Note that all the filaments connect

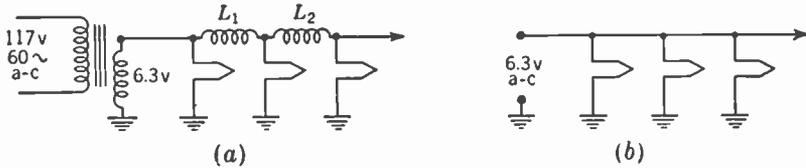


FIG. 10-7. Parallel heater connections. (a) With series r-f decoupling chokes. (b) Equivalent circuit for 60 cps.

to a common line on one side but the opposite side of each filament is returned individually to chassis ground.

*Series Heaters.* In receivers that do not have a transformer to step down the a-c power-line voltage, the heaters of all the tubes are connected in a series-parallel circuit across the a-c line. A typical circuit is shown in Fig. 10-8. The heaters in the series string at the top of the diagram all require the same filament current of 0.3 amp. The voltage

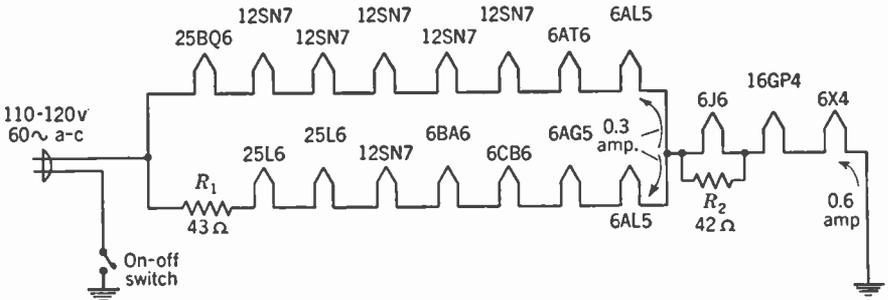


FIG. 10-8. Series heater circuit.

across the entire string of series heaters is 100.6 volts. All the tubes in the series string at the bottom of the diagram also require 0.3 amp. The voltage across the series heaters in this string, however, is 87.8 volts. Therefore, the 43-ohm series resistor  $R_1$  is added to provide an  $IR$  drop of 12.8 volts to make the total voltage 100.6 volts across this series string. The two strings are connected in parallel with each other. The parallel combination is in series with the separate string of three series heaters. Here the two individual branch currents of 0.3 amp each add to produce 0.6 amp, with 18.9 volts across the string of three series heaters. The

16GP4 kinescope filament and the 6X5 rectifier filament each require 0.6 amp at 6.3 volts but the 6J6 filament needs only 0.45 amp. Therefore, the 42-ohm filament shunt resistor  $R_2$  is added to provide a parallel path for 0.15 amp around the 6J6 heater. The sum of 100.6 volts in series with 18.9 volts equals 119.5 volts. The total filament voltage required, therefore, is equal to the a-c line voltage.

Many different arrangements of series-parallel heaters can be used. In general, the requirements are: (1) The total voltage of heaters in series across the a-c line must equal 110 to 120 volts. (2) Heaters in series must have the same current rating, or a shunt resistor like  $R_2$  is connected across the low-current filament to make the parallel combination equivalent to the higher current filament. (3) Heaters in parallel must have the same voltage rating, or a resistor like  $R_1$  is connected in series with the lower voltage heaters to provide the additional voltage drop required. The filament resistors are often in a ballast tube. (4) Heaters

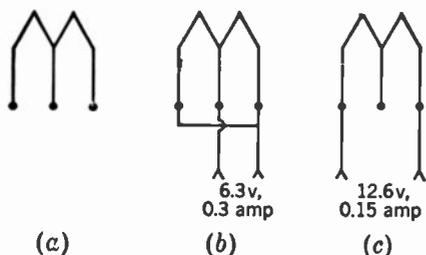


FIG. 10-9. Connections for heater with two sections. (a) Three-pin connections. (b) Parallel connection. (c) Series connection.

that have a relatively high d-c cathode potential, or are susceptible to hum pickup, are connected close to the ground end of the series string.

Some tubes have the heater in two sections that can be connected in series for 12.6 volts operation or in parallel for 6.3 volts. Figure 10-9 shows how the filament connections are wired for the two cases. The 12AU7 and 12AT7 miniature glass twin triodes are examples of tubes with three filament pins for either series or parallel connection.

**10-3. Voltage Doublers.** A voltage doubler is a rectifier circuit in which two condensers are arranged to charge on alternate half cycles and to discharge in series with each other so that the two condenser voltages are added in the output. The voltage doubler is capable of supplying a d-c output voltage approximately twice the peak voltage of the a-c input voltage. If, for example, the alternating voltage applied to the voltage doubler has an r-m-s value of 100 volts, the d-c output voltage can be  $2 \times 141$ , or 282, volts. The input must be a-c voltage for doubler operation, although often no power transformer is used.

The circuit in Fig. 10-10 is a full-wave voltage doubler. When point *A* of the a-c input is more positive than point *B*, the a-c input voltage makes the plate of the rectifier tube  $V_1$  positive with respect to its cathode.

Plate current will then flow through the a-c input voltage source and  $V_1$  to charge  $C_1$  in the polarity shown. During the next half cycle of input voltage the cathode of rectifier tube  $V_2$  is driven negative, making its plate positive with respect to cathode, and the condenser  $C_2$  is charged. The output voltage across the two condensers in series with each other is equal to twice the peak value of the a-c input because each condenser charges to the peak value of the applied voltage. The circuit is a full-wave rectifier, with the ripple in the d-c output having twice the frequency of the a-c input, because on both half cycles power is supplied to the output by each rectifier and its filter condenser.

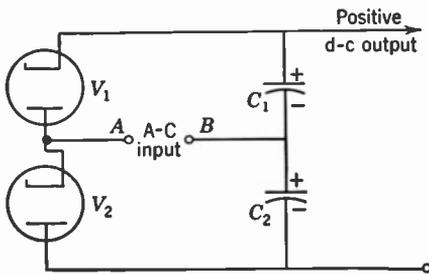


FIG. 10-10. Full-wave voltage doubler circuit.

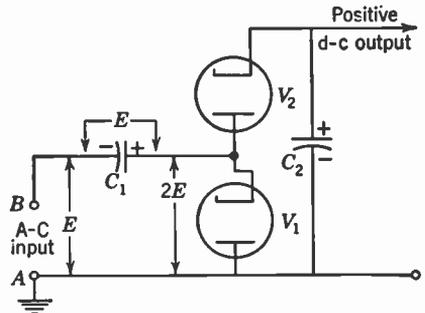


FIG. 10-11. Half-wave voltage doubler circuit.

The circuit in Fig. 10-11 is a half-wave voltage doubler. When point  $B$  of the a-c input is negative the plate of the rectifier tube  $V_1$  is driven positive and plate current flows to charge the condenser  $C_1$  to the peak value of the applied a-c voltage. This peak voltage is indicated by  $E$  in the diagram. With  $C_1$  charged to the peak voltage  $E$ , the potential with respect to ground at the  $V_2$  plate varies between zero and twice the peak voltage. Therefore, the peak voltage applied to the rectifier tube  $V_2$  is  $2E$  because it equals the sum of the d-c voltage across  $C_2$  and the a-c input voltage in series with each other. As a result, the rectified output voltage produced across  $C_2$  by  $V_2$  is twice the peak value of the a-c input voltage. This circuit is generally called a *cascade* voltage doubler because the rectified output of one diode supplies the doubled input voltage for the other diode. The cascade doubler is a half-wave rectifier, but it has the advantage of allowing a common ground connection for one side of the input and output circuits.

Energy delivered to the load from the voltage multiplier comes from discharging the output condensers. The amount of load current that can be supplied decreases with the voltage multiplication because of limitations on the peak current permissible through the rectifiers. The applications of voltage multipliers are generally limited, therefore, to doublers,

tripplers, and quadruplers. A voltage tripler may consist of a doubler in series with a half-wave rectifier, while a quadrupler can be arranged with two doublers in series with each other. A voltage tripler or quadrupler high-voltage supply is used to produce the 20 to 80 kv needed for the anode voltage of a projection kinescope. For direct-view picture tubes, the high-voltage supply may use a voltage doubler to produce 9 to 18 kv, although this can be obtained with just a half-wave rectifier. In receivers without a power transformer to step up the 117-volt a-c line voltage, the low-voltage power supply usually has a voltage doubler to produce about 250 volts for the B supply voltage.

**10-4. Transformerless Low-voltage Power Supply.** Figure 10-12 shows the diagram of a low-voltage power supply without a power transformer, including a voltage doubler to supply B+ and a series-parallel

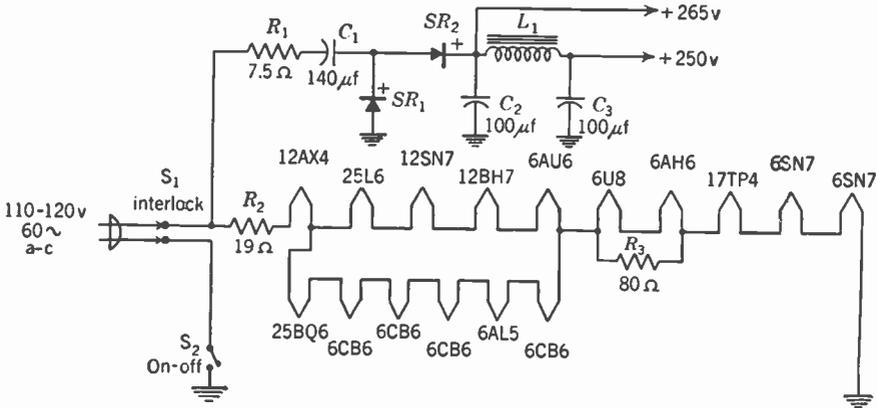


FIG. 10-12. Transformerless low-voltage power supply.

heater circuit across the a-c power line. In the B supply, the selenium rectifiers  $SR_1$  with  $C_1$  and  $SR_2$  with  $C_2$  form a cascade voltage doubler.  $R_1$  limits the peak current through the selenium rectifiers, which have a very low resistance in the forward direction. Note the large values of capacitance for the filter condensers to maintain the output voltage as the load draws current. In the heater circuit,  $R_2$  is a series resistor providing an  $IR$  drop of approximately 6 volts;  $R_3$  is a filament shunt resistor to bypass 0.15 amp around the 6U8 and 6AH6 heaters. Some receivers use a transformerless B supply but have parallel heaters with a filament transformer, in order to eliminate the relatively high heater-to-cathode potentials in series heater circuits.

**Shock Hazard.** In a transformerless B supply the B- line is one side of the a-c power line. Therefore, if the chassis is used as the B- line, the chassis is connected to one side of the power line. By touching the chassis and the other side of the power line, you are connected across 117

volts. Or, if two chassis connected to opposite sides of the power line touch, this produces a short across the line. To eliminate the danger of shock when working on this type of chassis, a power transformer with a 1:1 turns ratio can be used to isolate the receiver from the power line and supply the same line voltage.

**10-5. High-voltage Power Supplies.** The high-voltage power supply consists of a source of high-voltage a-c input, of 9 to 18 kv, the high-voltage rectifier, and the filter for the rectified d-c output voltage for the kinescope anode. Some receivers have an additional high-voltage supply to produce about 5 kv for the focusing electrode in kinescopes using high-voltage electrostatic focusing. The television receiver generates its own high-frequency a-c input for the high-voltage supply, instead of using the stepped-up 60-cps a-c line voltage directly. Either an r-f oscillator stage operating at 300 to 500 kc is used to produce the a-c input for the high-voltage supply, or the high-voltage a-c input is obtained from the horizontal deflection circuits, at the frequency of 15,750 cps. These circuits have been developed because high-voltage power supplies operating directly from the 60-cps power line have several disadvantages. Most important is the danger of serious shock, since the 60-cps line can supply large amounts of power. In addition, a 60-cps high-voltage supply is costly and bulky because of the large transformers and filter condensers needed. The higher frequency of the a-c supply voltage in the more recent types of high-voltage supply eliminates most of these disadvantages because the transformers and filter condensers can be much smaller.

*R-F High-voltage Supply.* The basic circuit of the r-f high-voltage supply is shown in Fig. 10-13. The section enclosed by the dotted lines in the figure is a conventional tuned-plate grid-feedback oscillator, which has the function of producing the alternating voltage that is rectified and filtered to provide d-c output voltage. This type of oscillator circuit is used often in r-f power supplies because it is simple and has good stability over its tuning range. Operating frequencies generally employed for the oscillator are 300 to 500 kc, with the lower frequencies used for the higher voltage supplies. An output voltage of about 10 to 15 kv at approximately 1-ma current drain can be obtained from the basic r-f power-supply circuit.

Referring to the schematic diagram in Fig. 10-13, the plate circuit is tuned by means of  $C_p$  to the desired operating frequency, which is determined by the resonant frequency of the high-voltage secondary winding  $L_1$  with the shunt capacitance  $C_t$ . Part of the oscillator output is fed back from the plate coil  $L_p$  to the grid by the tickler coil  $L_2$ . To sustain oscillations, the feedback voltage must have enough amplitude to replenish the losses in the circuit and the proper polarity to reinforce the grid

signal voltage. In some circuits the feedback voltage is capacitively coupled from the plate of the high-voltage rectifier tube back to the grid of the oscillator by means of a coil spring wrapped around the glass envelope of the rectifier tube. The oscillator does not need any external signal. The circuit operates as a typical power oscillator, and grid-leak bias is obtained by means of  $R_p C_p$  for class C operation in order to obtain maximum efficiency.  $R_1$  is used to suppress parasitic oscillations, which reduce the efficiency of the oscillator. Coil  $L_1$  is the high-voltage secondary winding, which has many more turns than the primary to obtain a high-voltage step-up. The secondary voltage is applied to the rectifier, which provides the filtered high-voltage d-c output. Varying  $C_p$  changes the amount of high voltage.

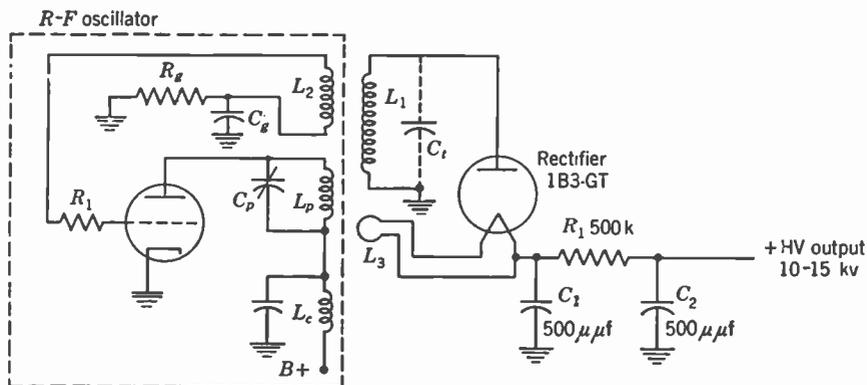


FIG. 10-13. The basic r-f high-voltage power-supply circuit.

**High-voltage Rectifier.** The rectifier tube commonly used for the high-voltage supply is the 1B3-GT, which has a maximum peak inverse voltage rating of 30,000 volts when the maximum frequency of the supply voltage is 300 kc. Energy for heating the rectifier filament is taken from the oscillator tank circuit by means of the filament winding  $L_3$ , consisting of only one or two turns. This eliminates the need for a separate filament transformer, which would require enough insulation to withstand the high d-c output voltage at the rectifier filament without arcing to ground. The coupling between  $L_3$  and the oscillator circuit can be adjusted to obtain the required filament voltage, or the coupling is fixed and a voltage-dropping resistor of 2 to 3 ohms is connected in series with the filament. In some cases the filament-dropping resistor is obtained by using resistance wire for the leads to the filament. Filtering of the rectified output is relatively easy because of the high ripple frequency and small current drain for this type of supply. A resistor-condenser filter is used, with the filter condenser usually about  $500 \mu\text{f}$ . A typical high-voltage filter condenser is shown in Fig. 10-14.

*Voltage Tripler Circuit.* Figure 10-15 shows the schematic diagram of a 30-kv r-f supply for a projection kinescope, using two 6Y6-G tubes in parallel for the r-f oscillator stage and three 1B3-GT diode rectifiers in a cascade voltage tripler circuit. The rectifiers need separate filament windings because each has a different amount of high voltage at the fila-

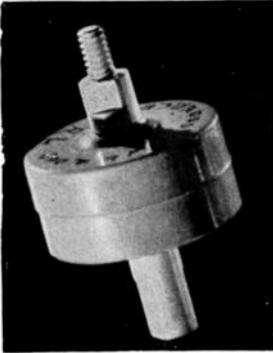


FIG. 10-14. 500  $\mu\text{f}$  high-voltage filter condenser. Voltage breakdown rating may be 10 kv or 20 kv. Diameter approximately 1 in. (Centralab.)

ment. The operation of the voltage tripler can be analyzed in three steps, as follows:

1. When the a-c input voltage across  $L_2$  makes the plate of  $V_1$  positive, it conducts to charge  $C_1$  to the peak value of the a-c input, which is approximately 10 kv.
2. On the next half cycle, the a-c voltage across  $L_2$  is in series with the d-c voltage across  $C_1$ , applying 20 kv to  $V_2$ , and the inverted diode conducts to charge  $C_2$  to 20 kv.
3. With 20 kv across  $C_2$ , when the a-c input voltage of 10 kv drives the plate of  $V_3$  positive 30 kv is applied to charge  $C_3$ .

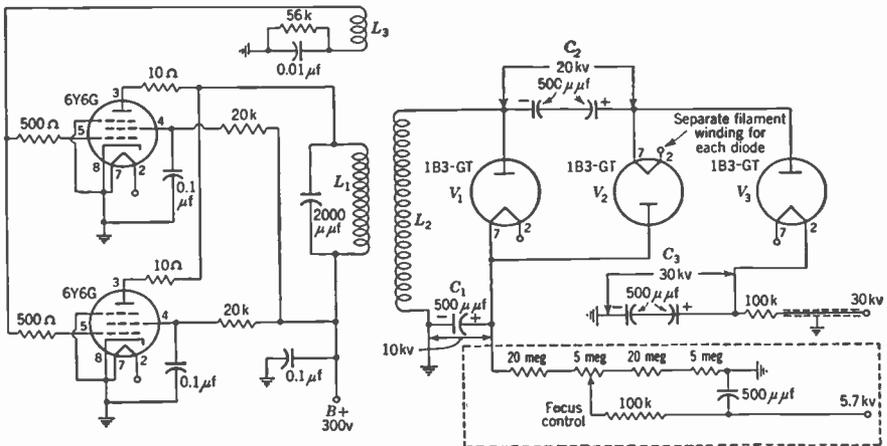


FIG. 10-15. Schematic diagram of 30 kv r-f power supply. (Essex Electronics.)

The filtered d-c output voltage of 30 kv across  $C_3$  is applied to the kinescope anode. The voltage divider across  $C_1$  provides a variable voltage of 5 to 7 kv for electrostatic focusing of the projection kinescope. Two condensers in series are used to provide the 500- $\mu\text{mf}$  capacitance for  $C_2$  and  $C_3$ , in order to increase the voltage breakdown rating from 20 kv for each condenser to 40 kv for the series combination.

*Flyback High-voltage Power Supply.* The flyback type of high-voltage supply makes use of the energy in the horizontal saw-tooth current wave required for magnetic deflection. This system can be employed only when an inductive circuit is used for horizontal scanning, because its operation depends on the sharp change in scanning current for the horizontal retrace. For this reason it is called the *flyback* or *kickback power*

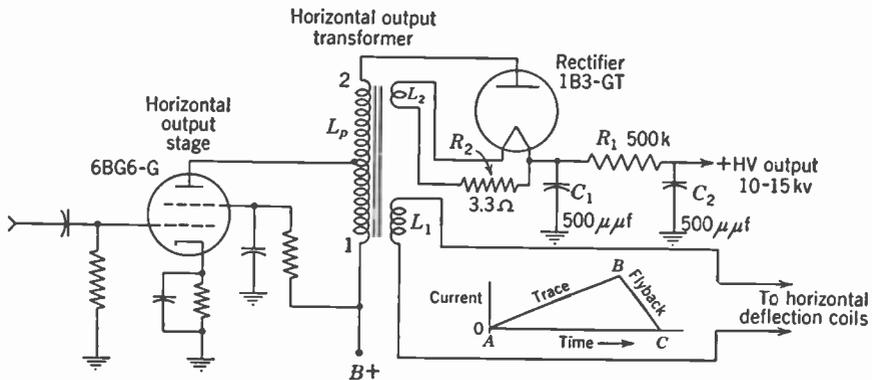


FIG. 10-16. Basic circuit of the flyback type of high-voltage supply.

*supply.* Practically all television receivers use the flyback type of high-voltage power supply.

Figure 10-16 shows the schematic diagram of a flyback high-voltage power supply and the saw-tooth current wave needed in the horizontal deflection coils for linear scanning. The scanning current through the deflection coils increases uniformly between points A and B on the saw-tooth wave to produce the relatively slow trace from left to right on the screen. At time B the current within the scanning coil decreases very rapidly for the quick flyback from right to left to complete the scanning of one horizontal line. The same procedure is repeated for each line to scan the entire frame. During the time that the scanning current is slowly increasing at a uniform rate its associated magnetic field is expanding. The flow of plate current in the output tube is suddenly cut off at time B and the magnetic field collapses very rapidly. The large induced voltage produced by the rapidly changing magnetic field shock-excites the output transformer, making it oscillate at its natural resonant frequency. This is about 100 kc, with typical values of inductance and stray capaci-

tance. Oscillation is allowed to continue for approximately a half cycle, after which it is damped out by means of a damping-tube arrangement not shown in the diagram. About 9 to 18 kv is developed across terminals 1 and 2 of the output transformer at the peak of the positive half cycle. This a-c voltage is applied between the plate of the rectifier tube and ground to be rectified and filtered, providing the high-voltage d-c output. When the rectifier plate is positive, current can flow through the

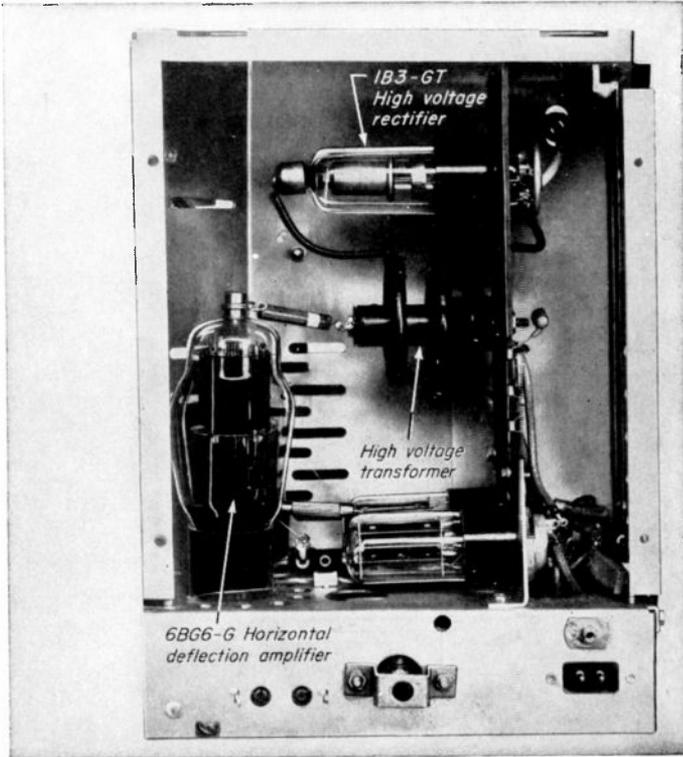


FIG. 10-17. Flyback power supply in high-voltage cage on receiver chassis. (RCA.)

rectifier tube, from terminal 2 to terminal 1 in  $L_p$ , and through the B supply to charge the  $500\text{-}\mu\mu\text{f}$  filter condenser. This size condenser is enough for adequate filtering, with the filter resistor and the  $500\text{-}\mu\mu\text{f}$  capacitance to ground that is part of the picture tube, since the a-c ripple frequency is 15,750 kc. Filament voltage is taken from the winding  $L_2$  for the 1B3-GT rectifier tube, which supplies the d-c output voltage of 10 to 15 kv, at about 1 ma. Figure 10-17 shows a flyback power supply in the high-voltage cage on the receiver chassis. The fuse for the horizontal output stage and high-voltage supply, with approximately 0.25 amp rating, is generally located in or near the high-voltage cage.

*Corona and Arcing.* At the high voltages used, with a relatively high frequency for the a-c power input, any point at a high potential can ionize the surrounding air to produce a visible corona effect, which is light blue in color. Corona results in loss of power and eventual insulation breakdown, causing arcing. It is important that there be no sharp edges in the wiring and that all soldered joints be smooth and round in order to minimize corona. The solid ring that can be seen under the high-voltage rectifier socket in Fig. 10-17 is a corona ring connected to the 1B3-GT filament in order to distribute the high potential at this point over a larger area to reduce the corona effect. Arcing and corona in the high-voltage supply can produce streaks in the reproduced picture. To recognize these effects, arcing can usually be heard as a snapping noise, while corona produces a sizzling sound, in addition to the fact that arcing and corona can be seen by looking into the high-voltage cage, with the room darkened. All high-voltage points must be well separated from the chassis to eliminate arcing.

*High-voltage Safety Precautions.* The receiver usually has a safety interlock switch that automatically opens the 117-volt a-c power input circuit when the back cover of the set, or the high-voltage-supply dust cover, is removed. If the safety interlock is disabled temporarily in order to test the receiver chassis in operation, it is a good idea to use only one hand when any part of the receiver must be touched while the high-voltage supply is functioning. The rectifier plate cap is dangerous because the amount of current possible here is limited only by the resistance of the input circuit supplying the a-c high voltage. The highest d-c voltage is at the rectifier filament. The high voltage can easily be removed if desired, though, by disabling the r-f oscillator or horizontal deflection oscillator. Disabling the high-voltage rectifier removes just the high-voltage d-c output but the a-c high voltage is still present. The high-voltage filter condensers and the anode of the kinescope should not be touched until they are discharged.

The effect of bodily contact with the high voltage depends primarily on the amount of current that passes through the body which, in turn, depends upon the amount of applied voltage and the body's resistance. For safety, then, it is desirable that the high-voltage supply have high resistance and poor regulation, so that the voltage can drop sharply with a partial short across the output. Using a high-resistance filter with a small filter condenser, instead of a choke, helps in limiting the maximum possible output current to a safe value. Another important factor is the effect of the charged filter condenser discharging through the body. Low values of filter capacitance are necessary as a safety measure. The amount of energy in the charged condenser should be limited to no more than 1 joule, which is 1 watt-second. The energy in joules is equal to

$CV^2/2$  where  $C$  is in farads and  $V$  in volts. With 15,000 volts, a filter condenser smaller than  $0.01 \mu\text{f}$  limits the stored energy to less than 1 joule.

**10-6. High-voltage Troubles.** The amount of high voltage supplied to the anode of the picture tube affects the screen brightness, focusing, and deflection sensitivity. Without any high voltage there is no brightness. Insufficient high voltage reduces the screen brightness and increases the deflection sensitivity, making the raster bigger but usually with poor focus.

*No Brightness.* This means there is no illumination at all on the screen —no raster, no line, or no spot. The reason for no brightness can be a defective kinescope, trouble in the kinescope auxiliary circuits, particularly misadjustment of the ion-trap magnet, or no anode voltage. If the trouble is no anode voltage for the kinescope, either the defect is in the high-voltage rectifier and its d-c output circuit or there is no high-voltage a-c input to the rectifier. In a flyback supply, if there is no a-c high voltage for the rectifier, the trouble can be in the horizontal deflection oscillator, deflection amplifier, and damper stages. With an r-f supply, the r-f oscillator stage must operate to supply the a-c high voltage for the rectifier. To check whether the r-f oscillator or horizontal deflection oscillator is operating, its grid-leak bias can be measured with a d-c voltmeter. Since it is produced by grid signal generated in the oscillator and fed back to the grid, the presence of the bias voltage means the oscillator is producing r-f output. If the grid-leak bias voltage is not there, the oscillator is not operating.

*Poor Regulation.* Because of its high internal resistance, the high-voltage power supply has poor regulation, as the d-c output voltage drops with an increase in load current. The output of a flyback type of high-voltage supply usually drops about 15 per cent with an increase of kinescope beam current from zero to  $50 \mu\text{a}$ . The r-f high-voltage supply has better regulation than the flyback supply and can provide correct focus over a wider range of brightness values. If the picture becomes too large and defocused as the raster blooms when the brightness is increased, this indicates the internal resistance of the high-voltage supply is too high, causing insufficient regulation. This trouble is often caused by a weak high-voltage rectifier tube.

**10-7. Troubles in the Low-voltage Supply.** Since the low-voltage supply is common to all sections of the receiver, a defect here can produce multiple trouble symptoms at the same time. Two reliable indications of trouble in the low-voltage supply are (1) no brightness and no sound at the same time; (2) a small raster, with both the height and width reduced. Table 10-1 lists the effects of low d-c output voltage caused by reduced a-c input voltage. The table shows that slightly reduced values of a-c

input voltage can lower the B+ voltage enough to decrease the size of the raster. With approximately one-third normal B+ voltage, the result is no brightness and no sound, which illustrates that the symptoms of no sound and no brightness mean trouble in the low-voltage power supply.

TABLE 10-1. EFFECTS OF CHANGES IN POWER-SUPPLY VOLTAGE

A-c input, volts	B+, volts	Effect
110	385	Normal raster, $10 \times 13\frac{1}{4}$ in.
100	340	Height reduced $\frac{1}{2}$ in., width reduced $\frac{1}{2}$ in., slight defocusing
90	300	Height reduced 1 in., width reduced 1 in., out of focus, reduced brightness
80	240	No brightness, sound volume low
65	120	No brightness, no sound volume

*Incorrect Voltage Division.* When the B supply has a voltage divider, incorrect values can cause multiple troubles in the receiver circuits using the divided voltages. It may be helpful to note that, generally, the effect of a short in a circuit is to lower its voltage, but an open in a circuit raises the voltage. As an example, in a divider that supplies positive and negative voltages with respect to ground, if the negative voltages are higher than normal while the positive voltages are too low, this indicates either an open in the negative side of the supply voltage circuits or a short in the positive side.

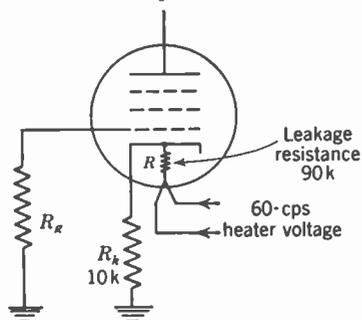


FIG. 10-18. Illustrating how heater-to-cathode leakage in a tube couples 60-cycle voltage into the cathode circuit.

**10-8. Hum.** Any of the receiver circuits can have undesired hum voltage present in the form of 60-cps a-c filament voltage, 60-cps ripple from the output of a half-wave rectifier, or 120-cps ripple in the output of a full-wave rectifier. Excessive a-c ripple in the B supply is the result of incomplete filtering. Hum from filament voltage can be caused by heater-to-cathode leakage in a tube.

Figure 10-18 illustrates how heater-to-cathode leakage in a tube can couple some of the 60-cycle heater voltage into the signal circuits. The

leakage resistance  $R$  between heater and cathode forms a voltage divider with the cathode resistor  $R_k$ . In this case  $R_k$  is 10 per cent of the total resistance in the voltage divider. Therefore, 10 per cent of the heater voltage is developed across the resistance  $R_k$  from cathode to ground. Since the control grid returns to the grounded side of  $R_k$  through the grid resistor  $R_g$ , the hum voltage in the cathode circuit varies the grid voltage and, therefore, the plate current. The higher the impedance in the cathode circuit is for 60-cycle voltage, the greater is the amount of hum voltage introduced by heater-to-cathode leakage.

Excessive 60-cps or 120-cps hum voltage can cause horizontal bars and bend in the picture.

*Modulation Hum.* Hum can be present in high-frequency signal circuits, even though they are not able to amplify the hum frequency, if the hum voltage modulates the high-frequency signal. In a picture i-f

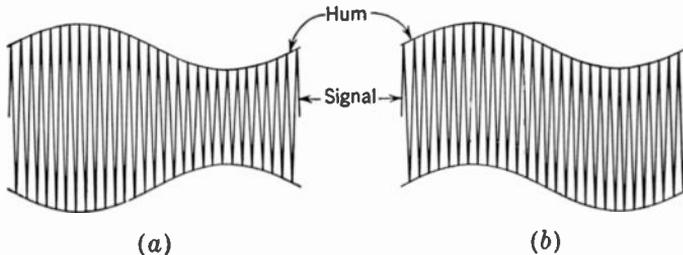


FIG. 10-19. Waveforms of hum combining with higher frequency signal. (a) Modulation hum. (b) Additive hum.

amplifier tuned to 45.75 Mc, as an example, leakage between heater and cathode can allow the filament voltage to modulate the grid signal voltage, producing i-f output that is amplitude-modulated by the 60-cps hum. Succeeding stages amplify the modulated i-f signal and when it is rectified the detected output includes the 60-cycle component in the video signal. Hum modulation usually occurs in the grid and cathode circuits of low-level stages. Modulation hum is often called *tunable hum*, since the hum is evident only when a signal is tuned in. The effects of modulation hum can be seen in the picture, therefore, but not in the raster alone. The local oscillator, r-f amplifier, converter, picture i-f, and sound i-f stages can have hum only in the form of modulation hum.

*Additive Hum.* In a circuit that has a plate load impedance for 60-cps or 120-cps a-c voltages, the hum can be present in addition to the desired signal. Modulation is not necessary, since the circuit can amplify the low-frequency hum voltage itself. The audio amplifiers, video amplifiers, sync amplifiers, vertical deflection circuits, and horizontal deflection circuits can amplify the hum voltage. The effects of additive hum are

evident in the raster and in the picture. Figure 10-19 illustrates the difference in waveforms for modulation hum and additive hum.

## REVIEW QUESTIONS

1. Draw the schematic diagram of a full-wave rectifier low-voltage supply with output voltage of +250 and -90 volts on the bleeder and a single  $\pi$ -type filter section using the loudspeaker field coil for the choke. What is the ripple frequency?
2. Give one advantage and one disadvantage of a series heater circuit compared with parallel heaters.
3. Draw the schematic diagram of a cascade voltage doubler circuit and describe briefly the principle of operation.
4. Give one advantage and one disadvantage of a voltage multiplier circuit.
5. Referring to the diagram in Fig. 10-12, how much current flows through  $R_2$ ? How much voltage is across  $R_3$ ?
6. Draw the schematic diagram of a voltage doubler B supply using two selenium rectifiers, labeling polarities of the rectifiers, the filter condensers, and the d-c output.
7. Describe briefly how to measure the grid bias on the oscillator in the r-f power supply shown in Fig. 10-13. Why does the presence of the grid-leak bias mean the oscillator is operating?
8. What are two safety measures to follow when working on high-voltage equipment?
9. Why is a bleeder resistance across the d-c high-voltage output often omitted in the r-f and flyback power supplies?
10. Describe briefly two insulation problems in a high-voltage supply.
11. What is the ripple frequency in the flyback high-voltage power supply in Fig. 10-16?
12. Referring to the diagram in Fig. 10-2, give the function of  $C_1$ ,  $L_1$ ,  $C_3$ , and  $C_6$ .
13. Referring to the diagram of Fig. 10-12, give the function of  $R_1$ ,  $R_2$ ,  $SR_1$ ,  $SR_2$ ,  $C_3$ , and  $L_1$ . What would be the effect on the picture and sound if  $R_2$  opened?
14. Referring to the voltage divider with positive and negative output voltages in Fig. 10-5b, give two troubles that could cause the voltage at point  $D$  to increase from -100 to -150 volts, while the voltage at point  $A$  decreases from +200 to +150 volts.
15. Why does no output from the high-voltage power supply result in no brightness on the kinescope screen? Give three other possible causes of no brightness, with normal sound.
16. Why does the absence of brightness and sound at the same time indicate trouble in the low-voltage supply? Give three possible troubles in the low-voltage supply that could result in no brightness and no sound.
17. What is a possible source of hum voltage causing one dark hum bar in the picture but not in the raster, and hum in the sound? The receiver is the intercarrier type, with the sound take-off circuit in the output of the second detector.
18. Describe the effect on the sound and the kinescope screen for each of the following troubles: (a) low-voltage rectifier fails; (b) low-voltage rectifier is weak; (c) high-voltage rectifier fails; (d) horizontal deflection generator fails in a receiver using the flyback high-voltage supply.

## CHAPTER 11

### VIDEO AMPLIFICATION

The video amplifier in the television receiver amplifies the output from the video detector, so that there will be enough video signal voltage coupled to the kinescope grid for proper reproduction of the picture. This corresponds to the function of an audio amplifier in a sound system, amplifying the audio voltage from the second detector to provide enough

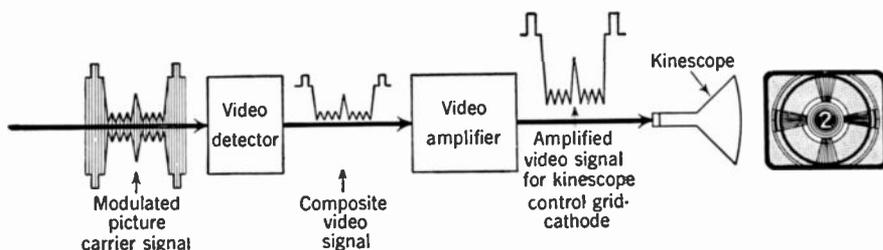


FIG. 11-1. The video signal coupled to the kinescope control-grid cathode circuit produces the picture on the raster.

signal to drive the loudspeaker. The function of the video amplifier and its place in the television receiver are illustrated in the block diagram of Fig. 11-1. Note that the entire composite video signal is amplified and used for the kinescope grid. Although the voltage above the black level does not contain any camera signal to be used for picture reproduction, the pedestal voltage is needed at the kinescope grid to establish the black level and blank out retrace lines.

**11-1. The Video Signal and Picture Reproduction.** The video signal enables the kinescope to reproduce the desired picture information. Figure 11-2 illustrates how the video signal voltage for one line, impressed between grid and cathode of the picture tube, results in reproduction of the picture elements on the kinescope screen. Assuming that the electron beam in the kinescope is scanning properly so that the elements are in their correct position, one horizontal line of the picture will be reproduced on the screen. The electron beam is made to scan the screen by the deflecting current or voltage applied to the deflecting coils or plates of the

picture tube. This scanning motion is independent of the video signal voltage coupled to the kinescope grid. The only function of the video signal is to vary the intensity of the kinescope beam current, thereby reproducing the degree of light or shade of the picture elements.

The effect of the a-c video signal on the kinescope grid is the same as any a-c signal coupled to the grid of a vacuum tube having a fixed negative d-c bias. The bias sets the operating point about which the instantaneous grid voltage varies, as the a-c signal makes the grid more or less

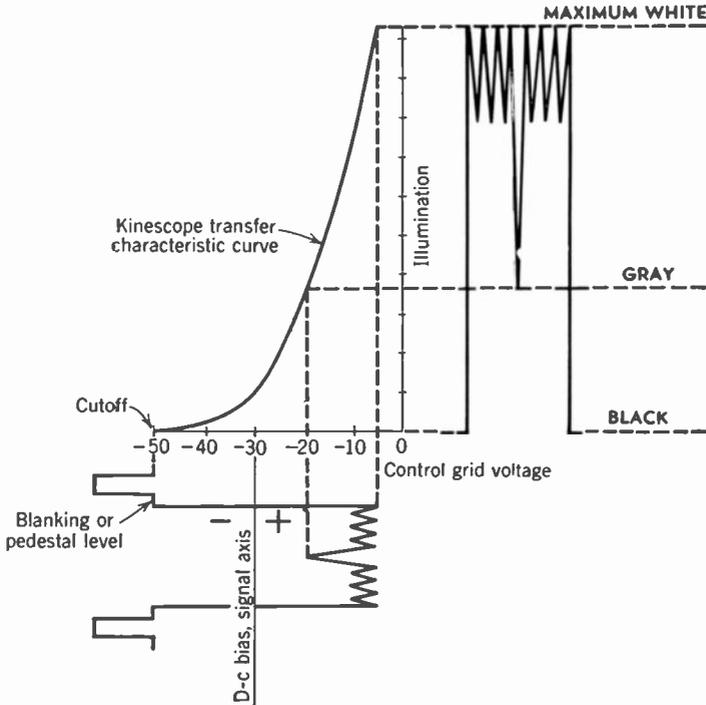


FIG. 11-2. The video signal voltage on the kinescope grid.

negative with respect to the cathode. The beam current in the tube varies with the grid voltage. As the grid goes more negative there is less beam current and the intensity of the spot on the screen is reduced. This produces a darker picture element. When the grid voltage goes as far negative as cutoff, the beam current is cut off completely and there is no spot of light on the screen. This corresponds to black. As the grid voltage is made less negative or more positive than the d-c bias value, the beam current in the tube increases to produce a brighter spot on the screen. This corresponds to the white parts of the video signal, with the maximum white driving the grid voltage the most positive. Although

the effect of the white parts of the video signal is to drive the kinescope more positive, it should be noted that the control-grid voltage is negative because of the d-c bias.

*Contrast and the Video Signal Amplitude.* The extent to which the video signal voltage swings away from its average-value axis determines the contrast of the reproduced picture. Suppose that the video voltage has only one-half the amplitude shown in Fig. 11-2. The maximum white of this weaker signal will not be so bright because the kinescope grid voltage will not be driven so far positive. This means that there will be less difference between the maximum white parts of the picture and black. The reproduced picture will not have so much contrast, resulting in a picture that appears weak and flat with no highlights. The contrast of the reproduced picture, therefore, is determined by the amplitude of the video voltage swing.

The video signal voltage coupled to the kinescope from the video amplifier must have enough amplitude to utilize the entire useful portion of the kinescope operating characteristic, so that the reproduced picture will have good contrast. The amount of composite video signal required for most picture tubes is about 70 volts peak to peak. With an output of 2 to 5 volts from the video detector and a gain of approximately 15 in a video amplifier stage, one or two stages of video amplification are enough to provide sufficient signal voltage for the control grid of the picture tube.

The amplitude of the video signal is given in peak-to-peak voltage, instead of r-m-s, average, or peak value. This is necessary because the video signal is not symmetrical, and its wave shape changes with picture content. Figure 11-3 shows that the peak-to-peak value of the composite video signal includes the total voltage swing between the tips of the synchronizing pulses and the maximum white level. Note that the peak-to-peak voltage for the white information in *a* and *b* is greater than for the gray information in *c* and *d*. The common a-c relations—that the average value is 0.636 the peak value and the effective value is 0.707 the peak value—apply only to sine waves. Since the positive and negative half cycles of the video signal are not necessarily the same, any notation other than peak-to-peak value is useless.

*Polarity.* The polarity of the signal on the picture tube grid is important because it determines the sense of the color values in the reproduced picture. If the polarity is opposite from that shown for the video signal in Fig. 11-2 a negative image in the same sense as a photographic negative will be produced, with the darker parts of the original image being white on the kinescope screen while the parts of the picture that should be white are reproduced as black. The correct polarity of the video signal makes the kinescope grid voltage less negative than the d-c bias or more

positive for white picture information, and the synchronizing pulses drive the kinescope grid voltage more negative beyond cutoff.

**Brightness and the D-C Component.** Since the a-c video signal varies the grid voltage above and below the d-c bias voltage, the picture elements reproduced on the screen vary in intensity around the brightness value corresponding to the bias voltage. Therefore, the kinescope grid bias determines the average amount of beam current and the average brightness of the reproduced picture.

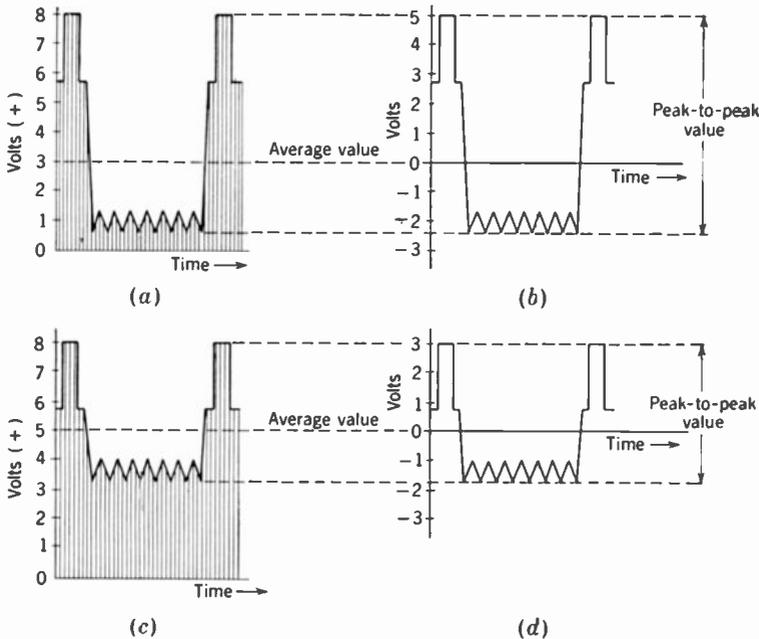


FIG. 11-3. Peak-to-peak value of the video signal. (a) D-c form for a white video signal. (b) A-c form for the same white video signal. (c) D-c form for a gray video signal. (d) A-c form for the same gray video signal.

In addition to the amount of a-c swing required for good contrast, the video signal for the kinescope grid must have the d-c component corresponding to its correct average-brightness level. This d-c component of the video signal combines with the fixed bias of the kinescope to establish the average brightness of the reproduced picture. If this d-c level of the video signal is missing, the kinescope bias will be incorrect and the picture will not be reproduced at the proper brightness level. Furthermore, the retrace lines may be visible.

**Black Level.** The grid cutoff voltage for the picture tube corresponds to black. When the negative grid voltage is equal to the cutoff value there is no beam current and no illumination of the screen. The part of

the video signal that corresponds to black, therefore, should drive the kinescope grid voltage to cutoff. Any grid voltage more negative than cutoff is blacker than black, or infrablack. The pedestal or blanking level of the composite video signal should drive the kinescope grid voltage to cutoff.

*Distortion.* The video signal at the kinescope control grid should correspond to the actual picture content. If in the process of amplification the video signal is distorted appreciably in amplitude, frequency, or phase the picture reproduction cannot be correct because of a grid signal voltage that has been distorted from the values corresponding to the correct picture information. Distortion of the relative amplitudes in the a-c video signal changes the relative light values in the reproduced picture, distorting the contrast. Frequency distortion degrades the quality of the picture information. Phase distortion can displace the reproduced picture information from its correct position, producing smear in the picture. The incorrect d-c level reproduces the wrong average brightness.

**11-2. Polarity of the Video Signal.** At the kinescope grid, the polarity of the a-c video signal voltage must be such that the synchronizing pulses occur during the negative half cycle and maximum white during the positive half cycle. This can be called either *positive picture phase* or *negative sync phase*, in referring to the polarity of the video signal. The polarities are with respect to the chassis ground in the receiver.

*Negative Transmission and Picture Phase.* It is standard practice in the United States that the polarity of the transmitted signal is negative, as described in Chap. 6. Negative transmission means that an increase of light in the original scene being scanned at the transmitter causes a decrease in amplitude of the transmitted carrier wave. At the receiver, this decrease in carrier amplitude must drive the kinescope grid voltage in the positive direction, thus causing an increase of light in the scene being reproduced on the kinescope screen.

*Producing Positive Picture Phase at the Kinescope.* The correct polarity of a-c video signal for the kinescope is the result of three factors in the receiver circuit: (1) whether the video signal is coupled to the kinescope control grid or to the cathode, (2) the number of video amplifier stages, and (3) the polarity of the video detector output. When the output from the video amplifier is coupled to the kinescope control grid, it must have positive picture phase. However, when the video signal is coupled to the kinescope cathode it must have negative picture phase because making the cathode voltage more negative drives the control grid in the positive direction. The output from an odd number of video amplifiers has opposite polarity from the input because each stage inverts the signal voltage. An even number of stages, however, has output signal of the same polarity as the input, since two phase inversions of  $180^\circ$  each produce

the original input polarity. In the video detector circuit, when the diode load resistor is in the plate circuit the video signal output has positive picture polarity, with the carrier signal modulated for negative transmission; the diode load resistor in the cathode circuit produces video signal output with negative picture polarity. Many combinations of these three factors for correct picture phase are possible, depending mainly on how many video amplifier stages are needed to produce enough video signal, but the arrangement generally used in receivers is either: (1) Positive picture phase from the video detector, inverted by one video amplifier to a video signal of negative picture phase, which is coupled to the kinescope

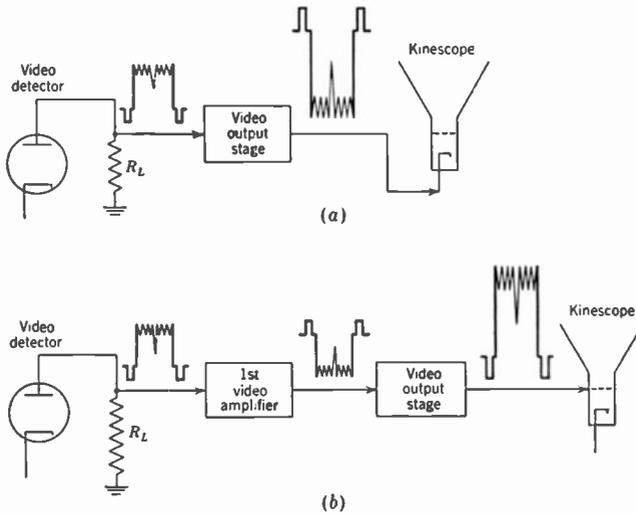


FIG. 11-4. Polarity of the video signal for the kinescope. (a) Negative picture phase at kinescope cathode. (b) Positive picture phase at kinescope grid.

cathode. (2) Positive picture phase from the video detector, inverted twice by two video amplifier stages, providing video signal of positive picture phase that is coupled to the kinescope grid. These two cases are illustrated in Fig. 11-4.

*Phase Inversion in an Amplifier.* Any amplifier with a resistive plate load inverts the phase of the input grid signal exactly  $180^\circ$ , making the output signal voltage of opposite polarity from the input voltage. When the input grid signal is the most positive, the output signal voltage is the least positive or most negative. Since the video amplifiers are resistance-coupled stages, each inverts the phase of the video signal.

The fact that the amplified signal voltage in a resistance-coupled amplifier is  $180^\circ$  out of phase with the input grid voltage can be demonstrated with the aid of the diagram in Fig. 11-5. With a B supply voltage of 300 volts and plate load resistor of 100,000 ohms, as shown, let the static plate

current of the tube be 2 ma. This is the value of plate current, set by the bias, that flows with no input signal on the grid. Assuming that the maximum positive grid voltage increases the plate current by 0.5 ma and that the maximum negative grid voltage decreases the plate current by the same amount, the plate current will increase above the 2-ma level to 2.5 ma and decrease below the static value to 1.5 ma because of the input signal voltage on the grid. As shown in the figure, this variation in plate current is in phase with the grid voltage because it follows the input signal exactly, increasing and decreasing with the grid voltage.

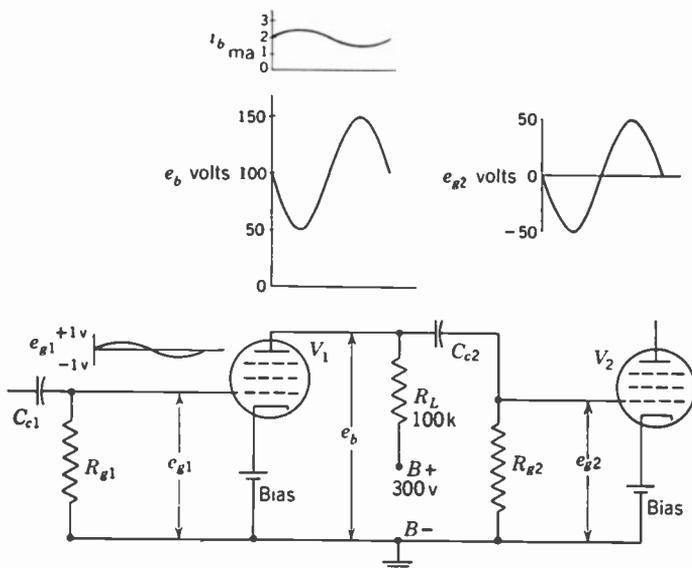


FIG. 11-5. Phase inversion in an amplifier.

However, the plate current is not the output signal. The plate-to-cathode voltage is the signal that varies the grid voltage of the next tube. This plate-to-cathode voltage is equal to the fixed B supply voltage minus the voltage across the plate load resistor  $R_L$ . The voltage across  $R_L$  is the  $IR$  drop resulting from plate current flow; this voltage changes as the plate current changes, causing the net voltage from plate to cathode of the tube to change with the varying plate current. With the 2-ma static value of plate current flowing, the  $IR$  drop across  $R_L$  is 200 volts. This makes the plate-to-cathode voltage 100 volts, since the plate voltage is equal to the difference between the 300 volts of the fixed B supply and the 200-volt  $IR$  drop across the plate load resistor. When the plate current increases to 2.5 ma, the  $IR$  drop across  $R_L$  increases to 250 volts and the net plate voltage decreases to 50 volts. When the plate current decreases to 1.5 ma, the plate voltage  $e_b$  is 150 volts. The effect of the plate current

varying between 1.5 and 2.5 ma about the 2-ma axis, therefore, is to make the plate voltage vary between 150 and 50 volts about a 100-volt axis. This varying plate-to-cathode voltage is the output signal that is coupled to the next stage.

As seen in Fig. 11-5, the output voltage has phase opposite from the input grid voltage. Its phase is opposite from the input signal even in the plate circuit, where the output voltage cannot become negative, because the plate voltage goes less positive when the grid voltage goes more positive. When the grid voltage goes more negative, the plate voltage becomes more positive. With the output signal voltage coupled to the next stage by any method that will block the steady component of the plate voltage, the variations in plate voltage are coupled to the next grid as an a-c signal, with a negative half cycle for the decreasing values

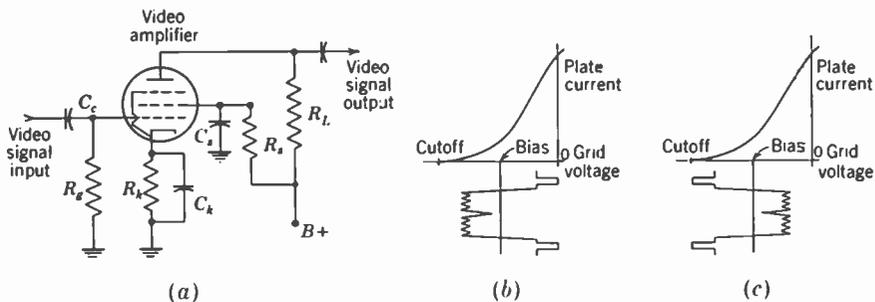


FIG. 11-6. Video amplifier circuit and its operating characteristics. (a) Basic  $RC$ -coupled amplifier. (b) Video signal voltage of negative picture phase on control grid. (c) Video signal voltage of positive picture phase on control grid.

of plate voltage and a positive half cycle for the increasing values of plate voltage. This signal voltage  $e_{o_2}$  is  $180^\circ$  out of phase with  $e_{o_1}$ .

**11-3. Operation of the Video Amplifier.** The circuit of a video amplifier stage is basically an  $RC$ -coupled amplifier, as shown in Fig. 11-6a, and its operating characteristics are essentially the same as a class A audio amplifier stage. As illustrated in b and c, the composite video signal input of the desired polarity swings the instantaneous grid voltage above and below the negative bias voltage, varying the plate current to produce the amplified video signal output voltage across the plate load resistor. Figure 11-6b illustrates the operating characteristic of the amplifier when the video signal input to the grid has negative picture phase, while c shows the conditions for a grid signal with positive picture phase. In either case, the video amplifier tube must have the correct d-c grid bias, plate, and screen voltages, in order to provide the operating characteristic required to avoid amplitude distortions due to clipping and limiting of the signal in the video amplifier.

In addition, the nonlinear operating characteristic of the amplifier affects the tones of white and gray in the picture reproduced on the kinescope screen. Notice that in Fig. 11-6b the parts of the video signal voltage corresponding to white produce instantaneous values of grid voltage near cutoff. Here the tube's operating characteristic is nonlinear, with less rapid changes in plate current for changes in grid voltage. As a result, the amplitudes corresponding to white are compressed in the video signal output. The effect of compressing the white video signal amplitudes can cause a pasty appearance in large white areas, such as the faces of people in the televised scene. In Fig. 11-6c the white parts of the video signal vary the grid voltage within the linear portion of the tube's operating characteristic. The video signal amplitudes near the black level are compressed but this is not so noticeable because the center of interest in the scene is usually in the lighter parts of the picture. Another advantage of operating the video amplifier with an input signal of positive picture phase is that noise voltages of higher amplitudes than the sync pulses can be clipped when they drive the grid voltage more negative than the cutoff voltage. For these reasons, the video amplifier usually has video input signal of positive picture phase from the video detector.

**11-4. Manual Contrast Control.** The amount of a-c video signal applied to the kinescope control-grid cathode circuit can be varied to adjust the contrast in the reproduced picture, since this varies the amount that the kinescope grid voltage swings above and below the d-c bias voltage. Corresponding to an audio volume control in a sound receiver, the *contrast control* or *picture control* is available on the front panel of the television receiver so that the operator can choose the desired amount of contrast in the picture. As the contrast control is turned clockwise, more a-c video signal is provided for the kinescope, producing a wider swing between black and the whitest parts in the reproduced picture to increase the contrast. Lowering the setting of the contrast control results in less swing of the video signal on the kinescope grid, giving the maximum white areas less intensity in the reproduced picture and reducing the contrast.

Any control that varies the amount of a-c video signal for the kinescope grid circuit will operate as a contrast adjustment. Therefore, the contrast control can be in the picture i-f section of the receiver or in the video amplifier. In a receiver that does not have automatic gain control, the contrast control is a manual bias control for the r-f and picture i-f amplifiers. Operating as a sensitivity control, it adjusts the over-all gain, varying the amount of picture signal coupled into the video detector and the amount of video signal output. In a receiver with automatic gain control to adjust the bias on the r-f and picture i-f amplifiers automati-

cally according to the signal level, the contrast control is in the video amplifier circuit.

In the video circuits, the contrast control can either vary the gain of the video amplifier or tap off the desired amount of video signal voltage. Figure 11-7 shows three typical circuits for a contrast control in the video amplifier. In *a*, the cathode resistor  $R_1$  supplies cathode bias for the video amplifier, and the control is unbypassed in order to provide degeneration. The degeneration is important in controlling the a-c signal level because with linear amplification little change in gain results from changing the bias. Moving the variable arm of the control closer to the cathode end reduces the bias and the degeneration, allowing more gain

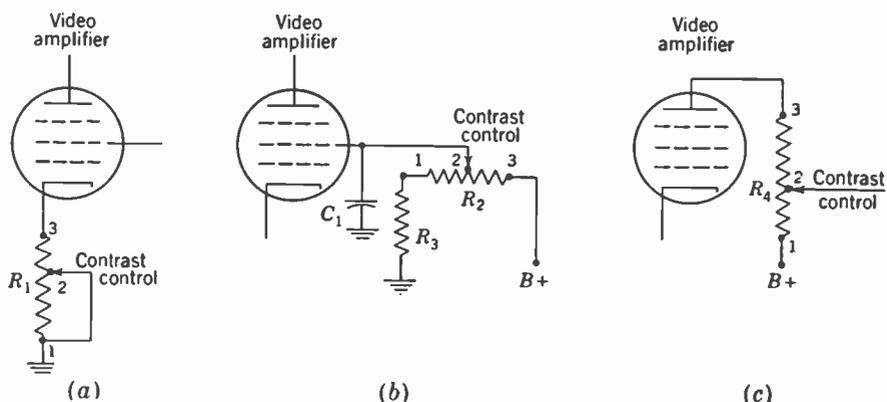


FIG. 11-7. Types of manual contrast control in a video amplifier. (a) Variable cathode bias. (b) Variable screen-grid voltage. (c) Potentiometer to tap off the desired amount of a-c video signal voltage.

in the video amplifier to increase the video signal amplitude and the contrast. In *b*, the contrast control  $R_2$  varies the gain of the amplifier by changing the screen grid voltage.  $R_3$  forms a voltage divider with  $R_2$  to provide a minimum value of screen-grid voltage, and  $C_1$  is the screen bypass condenser. Both these arrangements have the disadvantage of changing the operating characteristics of the video amplifier, when adjusting the contrast, because the d-c operating voltages change. In *c* the contrast control  $R_4$  is a potentiometer that taps off the desired amount of a-c video signal voltage. The d-c operating voltages are not changed when adjusting the contrast in this arrangement. However, the contrast control is in the video signal circuit, where the stray capacitance of the potentiometer and its connecting leads can reduce the high-frequency response of the amplifier. To minimize the shunt capacitance, the control usually is mounted near the video amplifier circuits, with the shaft mechanically linked to the front panel of the receiver.

**11-5. Video Frequencies.** It will be necessary for the video amplifier in the television receiver to amplify signal voltages that may have components ranging in frequency from 30 cycles to 4 Mc per sec. The high frequencies are produced because the video signal contains, within a line, rapid changes in voltage that occur during very much less time than the active line-scanning time of  $53.3 \mu\text{sec}$ . These rapid changes in signal voltage can correspond to frequencies infinitely high, but in present practice are limited to approximately 4 Mc by the restriction of a 6-Mc transmission channel. There are video signal voltages with frequencies lower than 4 Mc, but this sets an upper limit on the signal frequencies that the video amplifier in the receiver has to amplify.

The relationship between the frequency of a video signal variation and its associated picture information is illustrated in Fig. 11-8. As shown in

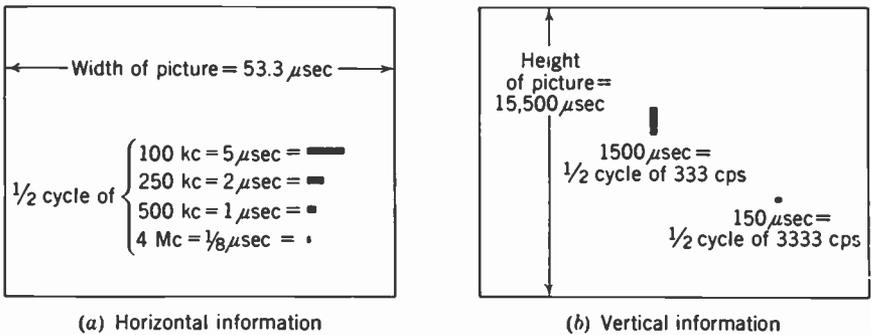


Fig. 11-8. Relation of video signal frequencies to the size of the picture information. Time for width of picture in *a* calculated as  $63.5 \mu\text{sec}$  minus 16 per cent H for horizontal blanking. Time for height of picture in *b* calculated as  $\frac{1}{60}$  sec minus 7 per cent V for vertical blanking.

the illustration, a specific length of line requires a definite amount of scanning time, and this determines how fast the corresponding variations in video signal amplitude will occur for the different light levels in the picture. To convert the size of an element of picture information into frequency, it is first necessary to calculate the time required to scan the element. This time can be considered as the period of a half cycle of the video signal required to reproduce the information. Multiplying the time by two to obtain the period of a full cycle and then taking the reciprocal of the period provides the desired frequency. Referring to *a* in Fig. 11-8, it is shown that a horizontal line with a width slightly less than one-tenth of the picture width is scanned horizontally in  $5 \mu\text{sec}$ . Therefore, the video signal for this black line preceded and followed by white information corresponds to one-half cycle of a signal variation with a period of  $10 \mu\text{sec}$  and a frequency of 100 kc. The signal frequencies needed to reproduce picture information that is given by the vertical scanning motion can be

calculated in the same manner, as shown in *b* of Fig. 11-8. Here the size of the elements in the vertical direction is converted to frequency in terms of the active vertical scanning time of 0.0155 sec. This is relatively low frequency information compared to details reproduced within a line. If the video voltage is taken from top to bottom, through all the horizontal lines in a field, this variation will correspond to a half cycle of a signal with a frequency of approximately 30 cps. When the brightness of the picture varies from frame to frame, the resultant signal frequency is lower than 30 cps, but this is considered as a change in d-c level and is reproduced by means of a d-c reinsertion circuit.

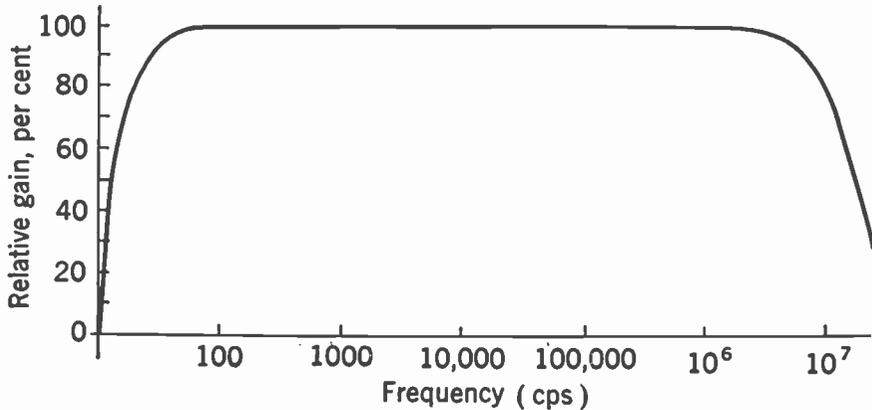


Fig. 11-9. Desired video amplifier response curve.

The a-c video signal, therefore, can be regarded as a complex waveform, not being sine wave in form, containing signal voltages that range in frequency from 30 cycles to 4 Mc per sec, approximately. In order to maintain the signal's wave shape, the video amplifier must be capable of amplifying this wide frequency range without distortion. This is not done too easily, because the amplifier inherently tends to have both phase distortion and frequency distortion when it operates over such a wide frequency range.

**11-6. Frequency Distortion.** The inability of the amplifier to amplify all frequencies of the input signal voltage equally well is called frequency distortion. This means that the amplifier has more gain for some frequencies than for others. Excessive frequency distortion cannot be tolerated because it changes the picture information. As shown in the amplifier response curve of Fig. 11-9, the amplifier response should be flat within a tolerance of about  $\pm 10$  per cent. Note that the frequency units are marked off on the horizontal axis in powers of 10, making the spacing logarithmic. This is necessary in order to accommodate the extremely

wide range of frequencies on a graph of reasonable size and still show the response at the extreme low- and high-frequency ends.

When the amplifier has a flat response curve over the video-frequency range, the relative gain of the amplifier is the same for all signal frequencies and the amplifier introduces no frequency distortion. With an actual video signal containing typical picture information coupled to the amplifier, the different frequency components do not all have the same amplitude. When the amplifier response is flat, however, all frequencies are amplified equally well and the video output signal is an amplified duplicate of the input signal with no change in wave shape.

The video amplifier response will not ordinarily be flat over the required frequency range unless precautions are taken in building the amplifier and special compensating circuits are added. Usually the response of the uncompensated video amplifier is down for the high video frequencies (about 0.5 Mc and above), making it necessary to provide high-frequency compensation for a flat frequency response up to 4 Mc. At the low-frequency end the video amplifier response is often inadequate at about 100 cps and below, so that low-frequency correction of the video amplifier may also be necessary to provide a video signal voltage that permits proper picture reproduction. The response of the video amplifier over the middle range of frequencies is normally flat and requires no compensation.

*Loss of High Video Frequencies.* Insufficient response for the high video frequencies reduces the amount of horizontal detail in the picture because the high-frequency components of the video signal correspond to the smallest picture elements in a horizontal line. If these signal frequency variations are lost, the rapid changes between black and white for small adjacent picture elements in the horizontal lines cannot be reproduced on the kinescope screen, with the resultant loss of horizontal detail. Figure 11-10 shows the effect of loss of the high video frequencies on the reproduced test pattern. Notice the lack of separations between the black and white divisions in the top and bottom wedges. The extent to which the divisions in either of these wedges can be resolved indicates the high-frequency response. Normally, the divisions can be seen all the way in to the center circle, indicating video-frequency response up to 4 Mc. The divisions in the widest part of the vertical wedges correspond to a video frequency of about 2 Mc.

In a televised scene, loss of the high-video-frequency information is evident as reduced detail, resulting in a picture that does not appear sharp and clear. Small details of picture information, such as individual hairs in a person's eyebrows and details of the eye, are not reproduced. In addition, the edges between light and dark areas, as in the outline of lettering or the outline of a person's face, are not reproduced sharply but trail off gradually instead. The effects of reduced high-frequency

response causing insufficient detail in the picture are not so noticeable when the camera at the studio presents a close-up view, since the subject then occupies a large area of the picture and the highest video signal frequencies are not necessary for adequate reproduction.

*Loss of Low Video Frequencies.* The video frequencies from 30 cps to 100 kc, approximately, represent the main parts of the picture information, such as background shading, lettering, and any other large areas of black and white. This follows from the fact that it takes a longer period of time for the scanning beam to change from black to white over large areas. Frequencies from 100 kc down to about 10 kc correspond to



FIG. 11-10. Test pattern illustrating loss of high video frequencies. Note the lack of separation of the divisions in the top and bottom wedges. Also, they are weaker in intensity than the side wedges. (*RCA Pict-O-Guide.*)

black-and-white information in the horizontal direction having a width one-tenth or more of a horizontal line. Frequencies from 10 kc down to 30 cps can represent changes of shading in the vertical direction. If a solid white frame is scanned, the signal is a 30-cps square wave. If this low-frequency square wave is not amplified with its wave shape preserved, the reproduction will show a white screen having a gradual change of intensity from top to bottom.

Figure 11-11 shows a test pattern reproduced with insufficient low-frequency response. The background is dull gray, instead of white, the lettering is not solid, and the picture in general is weak with insufficient contrast between large black and white areas. Notice that the side wedges, which represent low video frequencies, are weaker than the top and bottom wedges corresponding to the high video signal frequencies. The changes from black to white in the vertical direction between the

divisions in the side wedges represent frequencies of 4 to 8 kc, approximately, while the horizontal center line corresponds to one-half cycle of a 40-kc square-wave signal.

**11-7. Phase Distortion.** Before the harmful effects of phase distortion at low frequencies can be fully appreciated, the effect of time delay in the picture reproduction must be analyzed. The relative time delay of some parts of the signal with respect to the others is important because one element of the picture is being reproduced at a time as the scanning beam traces out the frame. As a result, a great enough time-delay distortion in



FIG. 11-11. Test pattern illustrating loss of low video signal frequencies. Background is weak, lettering is not solid, and side wedges are weaker in intensity than vertical wedges. (*RCA Pict-O-Guide.*)

the video signal can have the effect of displacing the picture information on the kinescope screen.

The time it takes the beam to complete its visible scanning run from left to right for one horizontal line is  $53.3 \mu\text{sec}$ . Consider now the scanning speed in a picture 10 in. wide reproduced on the picture tube screen. Since it takes  $53.3 \mu\text{sec}$  to cross the screen from left to right and the picture is 10 in. wide, it takes the beam  $53.3 \mu\text{sec}$  to move 10 in. on the screen or  $5.33 \mu\text{sec}$  for 1 in. If some low-frequency video signal suffers a time delay of  $5.33 \mu\text{sec}$  because of phase distortion, the variation of the spot intensity on the screen corresponding to that signal will be displaced 1 in. to the right from its proper position.

Phase delay is equivalent to time delay. If one signal voltage is  $10^\circ$  out of phase with another and lagging, as shown in Fig. 11-12, it reaches its maximum and minimum values at a later time. The time delay is the amount of time that corresponds to  $10^\circ$  of the cycle, in the illustrated

example. This varies with the frequency. For a signal voltage having a frequency of 100 cps, it takes  $\frac{1}{100}$  sec for one complete cycle of  $360^\circ$ . The amount of time equivalent to  $10^\circ$  in this cycle is  $\frac{10}{360} \times \frac{1}{100}$  sec, which is approximately 0.000278 sec, or 278  $\mu$ sec. In this time the scanning beam can be displaced in a vertical direction by more than four lines.

Phase distortion, therefore, is very important at low video frequencies because even a small phase delay is equivalent to a relatively large time delay. For the extremely high video frequencies, the effects of phase distortion are not as evident on the screen because the time delay at these high frequencies is relatively small. Normally a video amplifier stage that has flat frequency response up to the highest useful video frequency has negligible phase distortion for the high frequencies.

The effect of phase distortion in distorting the picture reproduction can also be examined in terms of the effect of time delay in changing the shape

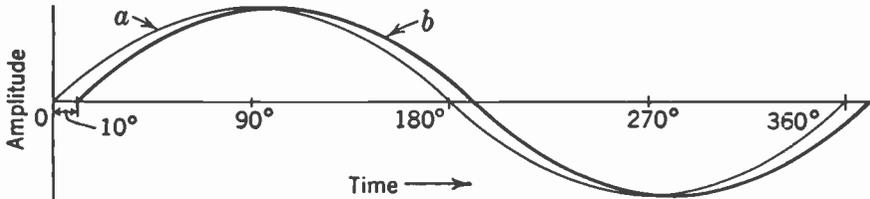


Fig. 11-12. Phase delay. Wave *b* lags behind wave *a* by the amount of time equal to  $10^\circ$  of the cycle.

of a nonsinusoidal wave. When phase distortion changes the wave shape of the nonsinusoidal video signal coupled to the kinescope grid, the reproduction cannot correspond to the correct picture information contained in the original video signal wave shape. In addition, the effect of phase distortion in changing the wave shape of a nonsinusoidal signal voltage such as a square wave gives a convenient method for observing visually the results of phase distortion in an amplifier, by means of a cathode-ray oscilloscope.

Consider the nonsinusoidal wave shape shown in Fig. 11-13*a*. This wave is actually composed of two sine waves. One is the fundamental, with the same frequency as the combined waveform. The other sine wave is the third harmonic, having a frequency three times the fundamental frequency and an amplitude one-third of the fundamental. Addition of the two sine waves, fundamental and third harmonic, produces the resultant nonsinusoidal wave, which has the same frequency as the fundamental but tends to become more of a square wave than a sine wave. If enough odd harmonics are added to the fundamental, the result will be a square wave.

The wave shape resulting from combining the fundamental and its harmonics is critically dependent on the phase relation between the waves. In Fig. 11-13b is shown the combination of the same fundamental and third harmonic as in *a*, the amplitude and frequency of each being preserved but with a different phase angle between the two waves. Assume

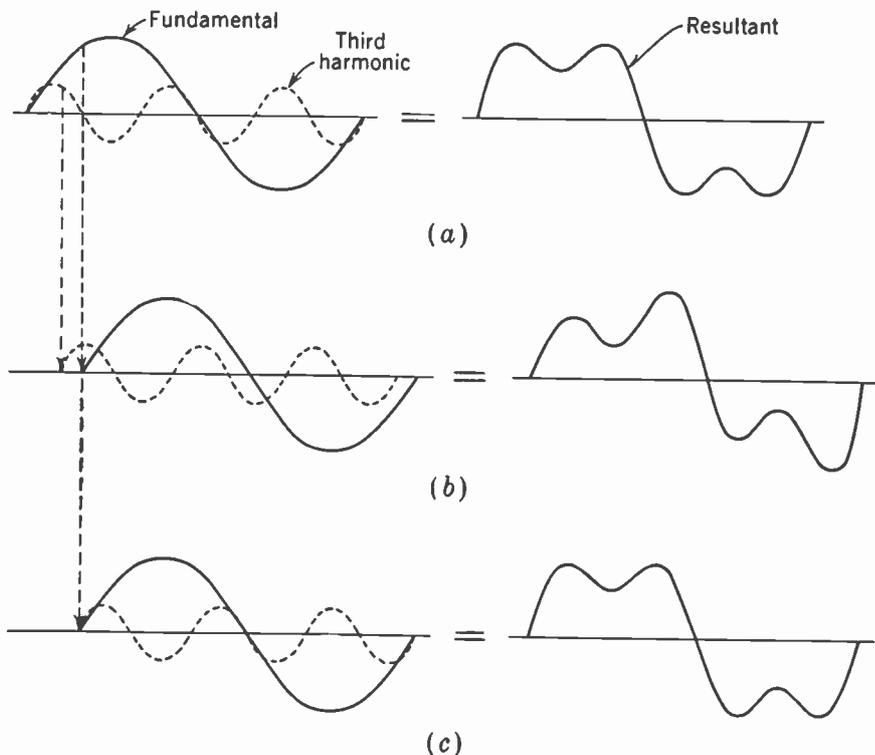


FIG. 11-13. How phase distortion changes the resultant signal wave shape. (a) Fundamental and third harmonic in phase. (b) Fundamental delayed by  $60^\circ$  and third harmonic with  $120^\circ$  delay. (c) Fundamental delayed by  $60^\circ$  and third harmonic with  $180^\circ$  delay.

that in a stage amplifying such a signal the fundamental and third harmonic are originally in phase, as shown in *a*. Then because of phase distortion in some circuit, the fundamental is made to lag by  $60^\circ$  of its cycle and the third harmonic is made to lag by  $120^\circ$  of its cycle, with the result shown in *b*. The resultant wave shape is distorted from the original wave because the original phase relations between the fundamental and harmonic are not maintained. Phase distortion always produces an unsymmetrical distortion of the wave shape, as in *b*. Notice that the peak amplitudes of the wave in *b* have different values than the original

wave and they occur at different times than the peaks for the undistorted wave. When the video signal amplitudes are distorted this way, the corresponding picture information has incorrect light values and is displaced in time, producing smear in the reproduced image.

It is important to note that the phase distortion is introduced because the amount of phase shift is not proportional to frequency; it is not caused by the phase shift in itself. When the second harmonic is delayed twice as much as the fundamental, the third harmonic three times as much, and so on, then the phase shift is proportional to frequency and there is no phase distortion. In Fig. 11-10c the third harmonic is delayed by  $180^\circ$  instead of  $120^\circ$ , making the phase shift of the fundamental and third harmonic proportional to frequency. As a result, there is no phase distortion and the wave maintains its original shape.

To see why the phase shift should be proportional to frequency, the phase delay must be translated into time delay. Consider a 1,000-cps signal, corresponding to the fundamental frequency in Fig. 11-13, delayed by  $60^\circ$ . This is a delay of  $60/360$  of the complete cycle that takes  $1/1,000$  sec. The amount of time delay is

$$\frac{60}{360} \times \frac{1}{1,000} = \frac{1}{6,000} \text{ sec}$$

When a 3,000-cps signal corresponding to the third harmonic of the fundamental is delayed by  $180^\circ$  the amount of time delay is

$$\frac{180}{360} \times \frac{1}{3,000} = \frac{1}{6,000} \text{ sec}$$

This is the same amount of time delay as for the fundamental.

With the phase shift proportional to frequency, then, the time delay is uniform. The time delay is not harmful if all frequency components have the same amount of delay. The only effect of such uniform time delay would be to shift the entire signal to a later time but with no distortion, because all components would be in their proper place in the video signal wave shape and on the kinescope screen. For uniform time delay and no phase distortion, therefore, it is required that the phase angle be proportional to frequency, as illustrated in Fig. 11-14.

Nonuniform time delay can be introduced in the video amplifier because of reactance in the plate and grid circuits. Unless the reactance of the grid coupling condenser is negligible, the signal voltage across the grid resistor will have a different phase angle for different frequencies. With a reactive plate load circuit for the video amplifier, the amplified output

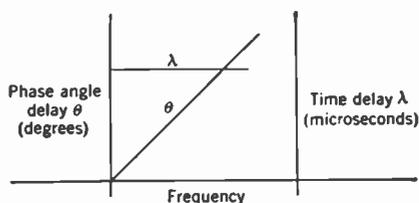


FIG. 11-14. Graph of linear phase response and constant time delay.

will be out of phase with the input by more or less than  $180^\circ$ ; and this phase angle will vary with frequency. Since the reactance in both the plate and grid circuits of the video amplifier is capacitive, this variation in phase angle is not linear; the phase angle depends on the ratio of reactance to resistance, and capacitive reactance does not vary linearly with frequency. It should be noted that an inversion of the complete signal of exactly  $180^\circ$  by the video amplifier does not contribute to phase distortion. There is no actual time delay resulting from transit time in the tube at video frequencies, and the phase inversion of  $180^\circ$  caused by the amplifier merely reverses the polarity of the video signal with no delay or advance in time.

**11-8. Direct-coupled Amplifier.** Some video amplifiers use the direct-coupled amplifier circuit, with a resistive plate load but omitting the

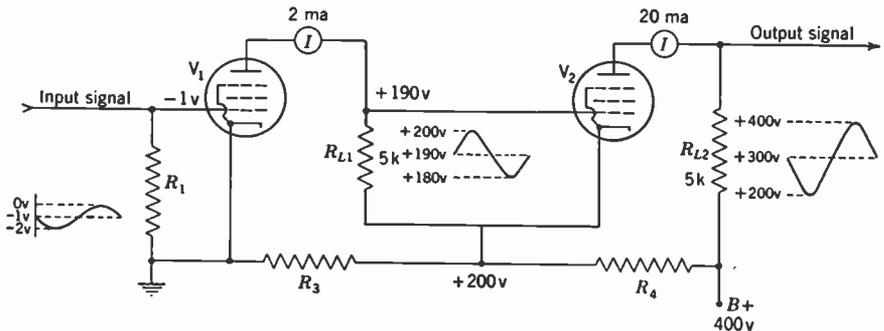


Fig. 11-15. A direct-coupled amplifier. A sine-wave signal is illustrated instead of video signal, and screen-grid voltages are omitted.

coupling condenser, as illustrated in Fig. 11-15. The advantage of this circuit is that the amplifier can have perfect low-frequency response, down to 0 cps, or direct current, because of the direct coupling without a blocking condenser. Since it can amplify direct current, the circuit is often called a *d-c amplifier*. The a-c signal variations are amplified with the d-c component of the signal input to the direct-coupled amplifier. Referring to Fig. 11-15, the signal input to the tube  $V_1$  varies  $\pm 1$  volt above and below the average value of  $-1$  volt, which can be the d-c component of the output from a previous stage. This d-c value of  $-1$  volt serves as the control-grid bias for  $V_1$ . The instantaneous grid voltage varies 1 volt above and below the bias axis of  $-1$  volt, varying the grid voltage between zero and  $-2$  volt. Assuming that a change of 1 volt in the grid varies the plate current by 2 ma, the plate-current variations in  $V_1$  are  $\pm 2$  ma, resulting in a plate-voltage change of  $\pm 10$  volts across the 5,000-ohm plate load resistor. The signal has been amplified by a factor of 10, therefore. Notice that the d-c component of the grid signal has also been increased ten times, since the average axis of the plate voltage is

10 volts from either peak of the signal, instead of 1 volt. The plate of  $V_1$  is directly coupled to the grid of  $V_2$  which amplifies the signal again by ten times, producing the output signal of  $\pm 100$  volts. Since the direct coupling makes the grid of  $V_2$  positive, with respect to ground, the cathode is returned to 200 volts, so that the control grid can still be 10 volts negative with respect to the cathode in order to provide bias. The plate of  $V_2$  is returned to the B supply voltage of 400 volts because the plate must be more positive than the cathode for plate current to flow.

In a d-c amplifier, the grid of one tube is always at the same potential as the preceding plate. This positive voltage at the control grid is canceled by putting a more positive voltage on the cathode of the second tube in order to keep the grid negative with respect to cathode. In addition, the second tube's plate voltage must be much more positive than the cathode to maintain positive plate voltage. Therefore, it is necessary to

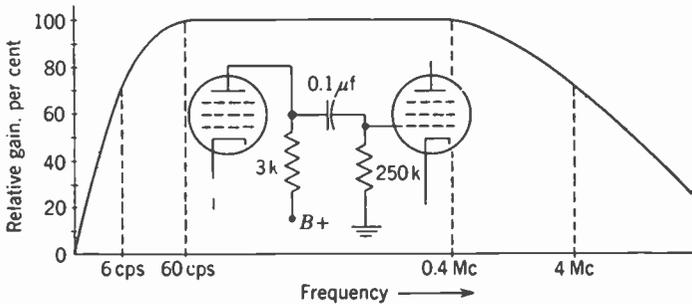


FIG. 11-16. Typical response curve for the  $RC$ -coupled amplifier.

connect each succeeding plate to a progressively higher B supply voltage. As a result, direct-coupled stages require a higher B voltage supply than a comparable resistance-capacitance coupled amplifier, where the d-c plate voltage is blocked from the next grid. The direct-coupled stage also is critical in adjustment, and sensitive to fluctuations of the screen and plate voltages because it amplifies d-c voltages. For one or two stages the requirements of a direct-coupled amplifier are not too severe, however, and it is often used for the video amplifier in television receivers.

**11-9. The Video Amplifier.** The basic circuit of the video amplifier uses a resistance for the plate load, with either direct coupling as in Fig. 11-15 or capacitive coupling as shown in Fig. 11-16. In this type of amplifier the gain is independent of frequency over a wide middle range of frequencies, making it most suitable for amplifying the band of video frequencies with the least amount of frequency and phase distortion. Audio-transformer coupling would not be suitable for the high video frequencies that correspond to radio frequencies. A video frequency as low as 2 Mc is higher than any signal in the standard radio broadcast band. This entire band from 535 to 1,605 kc would occupy only a very small por-

tion of the video band of frequencies from 30 cycles to 4 Mc. Transformer coupling using radio-frequency transformers would not be suitable for the lowest video frequencies that correspond to audio. Direct coupling can be used in order to amplify the d-c component of the video signal along with the a-c signal. The response for high frequencies, however, is not affected by the direct coupling.

*The Uncompensated Video Amplifier.* The  $RC$  amplifier is not the ideal answer to the problem of amplifying the extremely wide range of video frequencies without distortion, but it is a very practical solution. As shown in Fig. 11-16, the amplifier response is flat over a wide middle range of frequencies, which by proper choice of components can be extended to include a large part of the required video-frequency range. The response is down at the extremely low and high frequencies to an extent that would

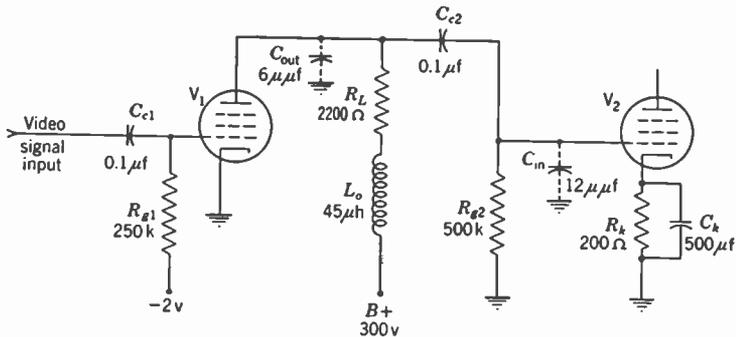


FIG. 11-17. Video amplifier with shunt peaking. Suppressor and screen-grid connections are omitted.

make the amplifier useless for video work without suitable correction. For the low frequencies, phase and frequency distortions are introduced because of insufficient bypassing in the cathode, screen, and plate circuits, and most important, by the increasing reactance of the grid coupling condenser at low frequencies. For the high frequencies, the decreasing reactance of the unavoidable shunt capacities in the plate circuit reduces the plate load of the amplifier, thus reducing the gain.

The reasons for distortion at the low-frequency end are not the same as the causes of high-frequency distortion. By using a corrective circuit for the low frequencies, therefore, and one for the high-frequency response, it is possible to obtain an over-all response for the resistance-coupled amplifier that is satisfactory for the extremely wide range of video frequencies.

*Shunt Capacitance.* The gain of the video amplifier is down at the very high video frequencies because of the shunting effect of various capacitances in the plate circuit to the chassis ground. These consist of the output capacitance of the tube itself, the input capacitance of the next stage, labeled  $C_{out}$  and  $C_{in}$ , respectively, in Fig 11-17, and stray capacitance of

the components and wiring to the chassis. The output capacitance is simply the plate-to-cathode capacitance for the amplifier tube. The input capacitance is equal to the static value of the grid-to-cathode capacitance in the following tube, plus a dynamic capacitance that is added to the input when the stage is an amplifier. This dynamic increase in the input capacitance of an amplifier stage is called the *Miller effect*. It results from the tube's grid-to-plate capacitance, as the effect of the amplified signal in the plate circuit is coupled back to the input to increase the dynamic value of the input capacitance.

The total shunt capacitance in a video amplifier stage is generally about 15 to 30  $\mu\text{f}$ . This may seem a small value, but at high video frequencies the capacitive reactance is low enough to reduce the plate load impedance and the gain of the amplifier. In order to keep the amount of shunt capacitance to the lowest possible value, tubes with small interelectrode



FIG. 11-17a. A typical video peaking coil.

capacitances are used. In addition, care must be taken in the wiring and placement of parts to keep the shunt capacitance to a minimum.

*High-frequency Compensation.* The inductance  $L_o$  in the video amplifier diagram of Fig. 11-17 is a peaking coil inserted in the plate circuit to improve the high-frequency response. Fig. 11-17a shows a typical video peaking coil. The peaking coil increases the high-frequency gain of the amplifier by resonating with the shunt capacities to form a broadly tuned circuit, thus canceling out the reactive effects of the shunt capacitance in the plate circuit. The result is an amplifier with a response that by suitable choice of values can be flat up to the required frequency. The peaking coil has no effect at the low and middle frequencies because its inductance is so low (usually 20 to 200  $\mu\text{h}$ ) that it has no appreciable reactance except for the extremely high frequencies. This method of high-frequency compensation, with the peaking coil effectively in parallel with the shunt capacities, is called *shunt peaking* and is one of several possible methods of high-frequency correction.

Two other basic methods of correcting the high-frequency response are illustrated in Figs. 11-18 and 11-19. In the series peaking circuit of Fig.

11-18 the peaking coil is effectively in series with the shunt capacities  $C_{in}$  and  $C_{out}$ . The combination peaking circuit in Fig. 11-19 combines both the shunt and the series peaking methods. Of these three methods, series-shunt combination peaking is used most often because it allows the most gain. Shunt peaking provides the least gain. Any of the three

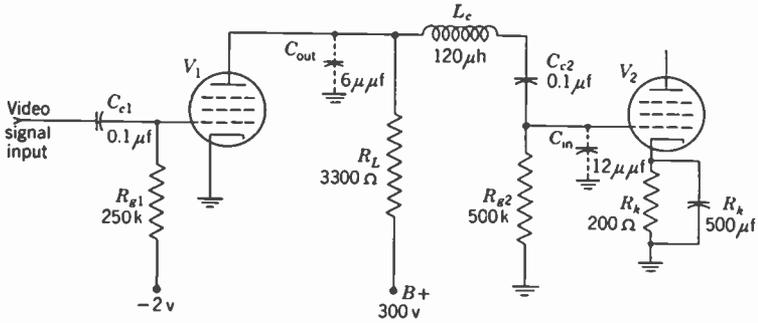


Fig. 11-18. Video amplifier with series peaking. Suppressor and screen-grid connections are omitted.

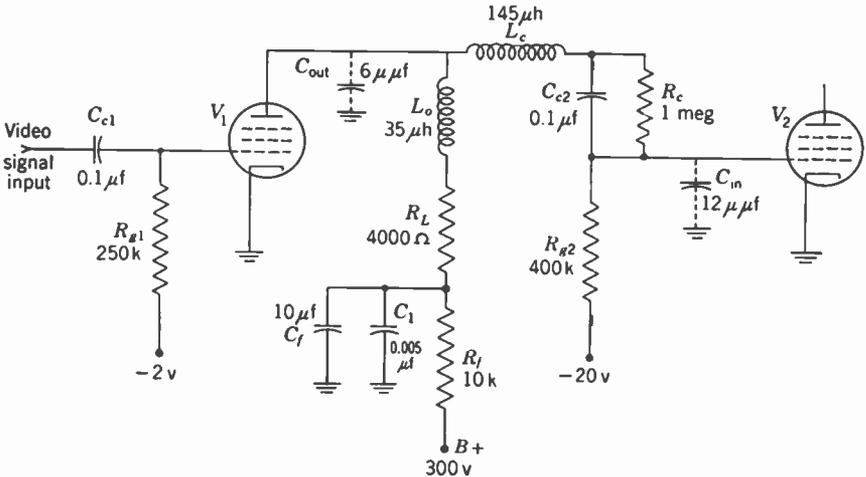


Fig. 11-19. Video amplifier with series-shunt combination peaking for high-frequency compensation, and  $R_f C_f$  and  $C_c R_c$  filters for low-frequency compensation.

methods can be made to give satisfactory response up to the highest desired frequency. The method of calculating values for these types of compensation is explained in the following chapter.

**Plate Load Resistor.** It is possible to minimize the effect of the shunt capacitance by reducing the size of the plate load resistor. The effect of the shunt capacitance in reducing the gain of the amplifier at the high-frequency end does not become noticeable until the reactance of the shunt capacitance becomes comparable to the resistance of the plate load.

Since capacitive reactance decreases for increasing frequencies, the plate load resistance is effectively in parallel with smaller reactances for the higher frequencies. The resultant impedance of the two in parallel will be less than either one. However, if one is very small compared to the other, for example, 2,000 ohms for  $R_L$  and 1 megohm for  $X_c$ , then the total impedance is practically the same as the smaller one. Therefore, until a high enough frequency is reached to make the reactance of the shunt capacitance small enough, it will have no shunting effect on the plate load resistor and no effect on the gain of the amplifier. By making the plate load resistor small in value the frequency necessary to make  $X_c$  have a shunting effect becomes higher, thus extending the high-frequency response of the amplifier. The value of  $R_L$  is 2,000 to 6,000 ohms for video amplifiers.

*Low-frequency Compensation.* Satisfactory correction of the amplifier's low-frequency response can be obtained by using large bypass condensers in the screen grid and cathode circuits, a large coupling condenser and grid resistor, in addition to special compensating filters. The decoupling filter  $R_f C_f$  in the plate circuit of  $V_1$  in Fig. 11-19 is in series with the plate load resistor  $R_L$  in order to boost the gain and reduce the phase distortion for the very low video signal frequencies. The filter has no effect on the amplifier's response in the middle range of frequencies, or at the extremely high end, because  $C_f$  is a bypass condenser for these frequencies. Notice the 0.005- $\mu\text{f}$  condenser in parallel with the large 10- $\mu\text{f}$  bypass condenser. This is done in video amplifiers to provide effective bypassing for the high frequencies where the inductance of the large tubular condenser may be enough to result in appreciable inductive reactance. The  $R_c C_c$  filter in the grid circuit of  $V_2$  in Fig. 11-19 improves the low-frequency response of the amplifier. By suitable choice of values for the plate circuit filter  $R_f C_f$  and the grid coupling circuit  $R_p C_c$  and  $R_c$ , the video amplifier's response can be corrected for frequencies down to 0 cps, or direct current. How to calculate the values for these compensating filters is explained in the following chapter.

*Gain of the Stage.* The gain is the ratio of output to input signal voltage. It varies directly with the size of the plate load because the output voltage is proportional to the plate load. This is a result of the voltage-divider effect between the internal plate resistance of the tube and the plate load resistor. The smaller the value of  $R_L$ , the smaller is the proportion of the amplified signal that is available across  $R_L$  as output signal voltage.

The gain of a pentode video amplifier over the frequency range for which the amplifier is flat is

$$\text{Gain} = g_m \times R_L$$

where  $g_m$  is the transconductance of the tube in mhos, available from the tube manual in micromhos, and  $R_L$  is the plate load resistance in ohms. For the 6AC7 tube with a  $g_m$  of 9,000  $\mu$ mhos, or 0.009 mhos, and  $R_L$  of 3,000 ohms, the gain of the stage is  $0.009 \times 3,000$ , or 27. The gain is a pure number without units because the mhos and ohms cancel, being reciprocals of each other. In general, the gain of a video amplifier stage is about 10 to 50, depending on the size of plate load resistor and the  $g_m$  of the tube.

*Tubes Used.* Pentodes and beam-power tubes are generally used for the video amplifier stage because of their high value of  $\mu$  and  $g_m$  and small grid-to-plate capacitance. The 6AC7, 6AG7, and 6CL6 are pentodes developed especially for wide-band service such as the video amplifier, where high  $g_m$  and low grid-to-plate capacitance are important. Miniature triode tubes with relatively low grid-to-plate capacitance and high transconductance can also be used. The miniature glass twin-triode tube 12AU7 is used as a two-stage video amplifier. Audio output tubes, such

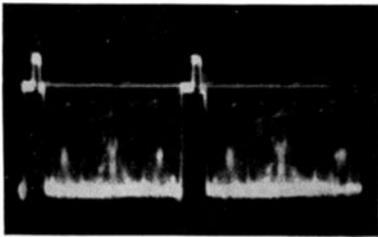


FIG. 11-20. Oscillogram of composite video signal. Oscilloscope internal sweep at 15,750/2 cps to show video signal for two horizontal scanning lines. Horizontal synchronizing pulses at top of waveform.

as the 6V6 and 6K6-GT, are used for the video amplifier output stage in the receiver because they can handle a relatively large signal swing on the control grid without excessive amplitude distortion.

**11-10. Video Amplifier Circuits.** The video amplifier in a television receiver generally has one or two stages. With input signal of positive picture phase in either case, the negative picture phase output of a single stage is coupled to the kinescope cathode, while the positive picture phase output of a two-stage video amplifier is coupled to the kinescope control grid. Figure 11-20 shows an oscillogram of the composite video signal output of the video amplifier. The amplitude of the video signal output for the kinescope is about 70 volts peak to peak. The schematic diagrams in Figs. 11-21 and 11-22 show typical video amplifier circuits.

*Single Direct-coupled Video Amplifier.* Referring to the circuit in Fig. 11-21, the composite video signal output of positive picture phase from the detector is directly coupled to the 6AC7 d-c amplifier, which amplifies the video signal with its d-c component and inverts the polarity to provide composite video signal of negative picture phase for the kinescope cathode. The picture control  $R_{316}$  in the cathode of  $V_{305}$  varies the contrast in the

reproduced picture by changing the gain of the video amplifier. Decreasing the cathode resistance to ground through  $R_{316}$  reduces the cathode bias and decreases the amount of degeneration produced by the unbypassed resistor, increasing the gain and the contrast. The 4,700-ohm resistor  $R_{322}$  is the plate load resistor of the amplifier.  $L_{303}$  and  $L_{304}$  are peaking coils to increase the response of the amplifier for high video frequencies.  $R_{320}$  and  $R_{321}$  are damping resistors across the inductances to broaden the frequency response of the coils. Each coil and its shunt resistor is a

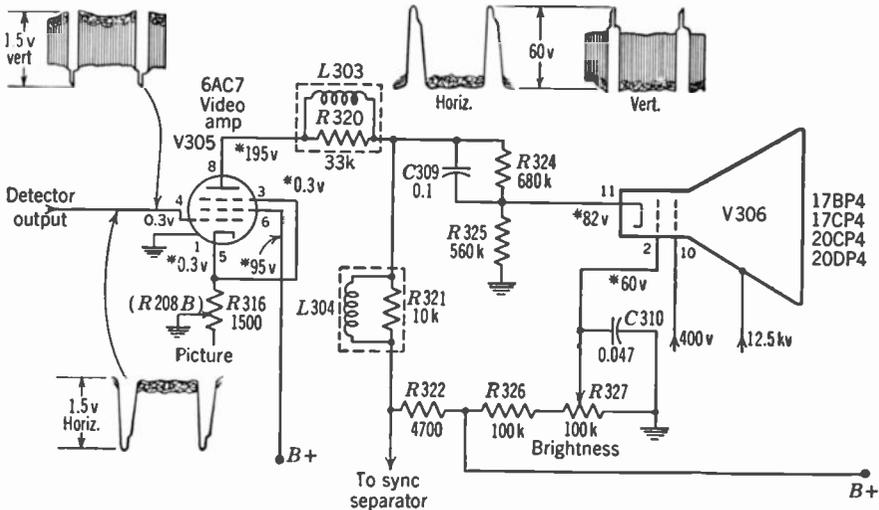


FIG. 11-21. Single-stage d-c video amplifier circuit. Capacitance values in microfarads and resistances in ohms, unless otherwise noted. Voltage values marked with asterisk vary widely with setting of controls. (Admiral Series 21 chassis.)

single unit, with the coil wound on the resistor as a coil form. The composite video signal output with its d-c component is coupled to the kinescope cathode through  $R_{324}$  and  $C_{309}$ .  $R_{324}$  decouples the kinescope cathode from the plate of the d-c amplifier for d-c voltage, so that a very slow change in the supply voltage will not vary the kinescope bias appreciably. The values of  $R_{324}$  and  $C_{309}$  can provide low-frequency compensation down to direct current, to provide the d-c component of the video signal for the kinescope cathode. In addition,  $R_{324}$  forms a voltage divider with  $R_{325}$  that reduces the d-c supply voltage at the kinescope cathode, to increase the potential difference between the cathode and accelerating grid. Note that the d-c coupling makes the kinescope cathode 82 volts positive with respect to ground, but the control-grid voltage set by the brightness control  $R_{327}$  is +60 volts, resulting in a negative grid bias of -22 volts.  $R_{326}$  forms a voltage divider with the bias control  $R_{327}$  to provide the required range of d-c voltages for the

desired brightness. The bypass condenser  $C_{310}$  across  $R_{327}$  returns the kinescope control grid to the ground, so that the video signal across  $R_{325}$  is impressed between grid and cathode of the kinescope.

*Two-stage RC-coupled Video Amplifier.* In Fig. 11-22 the composite video signal output of positive picture phase from the detector is coupled to the 6AU6 first video amplifier by the coupling condenser  $C_{138}$ , and grid resistor  $R_{138}$ . The first video stage amplifies and inverts the video signal for the 6K6-GT video output amplifier, which provides composite video signal output of positive picture phase at the kinescope grid. Since this receiver does not have automatic gain control, the contrast is varied by a manual bias control for the picture i-f stages, not shown in the figure.

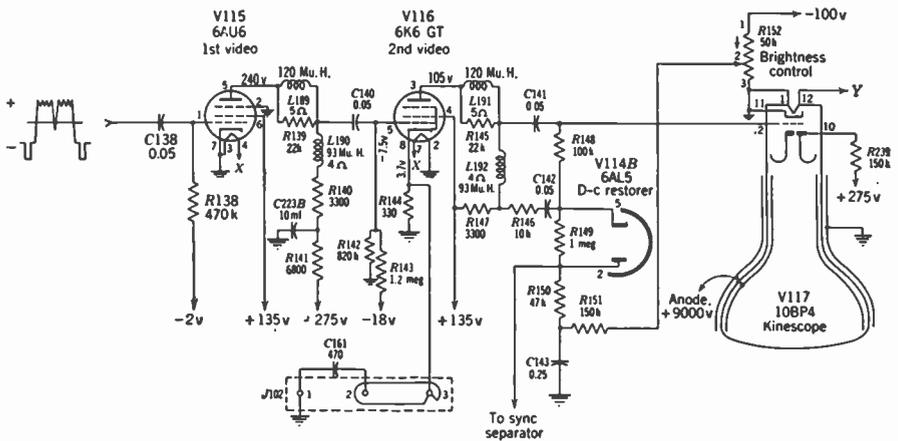


FIG. 11-22. Two-stage a-c video amplifier. Capacitance values in microfarads and resistances in ohms, unless otherwise noted. (RCA 8TS30 chassis.)

The plate load resistor for the first video amplifier stage is the 3,300-ohm resistor  $R_{140}$ . The peaking coil  $L_{189}$  has a shunt damping resistor  $R_{139}$ .  $L_{190}$  is also a peaking coil. The decoupling filter  $R_{141}$  and  $C_{223B}$  increases the low-frequency response.  $C_{140}$  couples the video signal to the grid of the 6K6 video output stage.  $R_{142}$  and  $R_{143}$  in the grid circuit form a voltage divider across the  $-18$ -volt supply to provide the required amount of grid bias for the 6K6-GT. The cathode bias produced by  $R_{144}$  is fixed, but the link on  $J_{102}$  can be changed to adjust the amount of degeneration in the cathode. With the link across positions 1 and 2,  $C_{161}$  is disconnected from the cathode resistor  $R_{144}$ , resulting in degeneration for all video frequencies. When the link is connected from 2 to 3, as in the diagram, the  $470$ - $\mu\text{f}$  capacitance of  $C_{161}$  bypasses  $R_{144}$  effectively for the high video frequencies only, reducing the amount of degeneration and increasing the response of the amplifier for high video frequencies. The plate load resistor for the video output stage is the 3,300-ohm resistor

$R_{147}$ ,  $L_{191}$  with its shunt damping resistor  $R_{145}$  and  $L_{192}$  are peaking coils.  $C_{141}$  couples the a-c video signal to the kinescope grid. Since the d-c component of the video signal is lost in the video amplifier, the a-c video signal is coupled to the 6AL5 d-c restorer, which rectifies the a-c input to provide for the kinescope grid a d-c voltage corresponding to the correct average-brightness level of the video signal.

**11-11. Hum in the Video Signal.** An excessive amount of 60 cps or 120 cps hum voltage in the video signal coupled to the kinescope produces *hum bars* on the screen, as illustrated in Fig. 11-23. Consider the case of a 60-cps sine-wave hum voltage varying the kinescope control-grid voltage in synchronism with the vertical scanning motion, as in *a* of Fig. 11-23. The positive half cycle of the hum voltage makes the grid more positive, increasing the beam current and screen illumination; the negative half cycle reduces the beam current and screen illumination. Since it takes  $\frac{1}{20}$  sec for a half cycle of the 60-cps voltage, the scanning beam moves

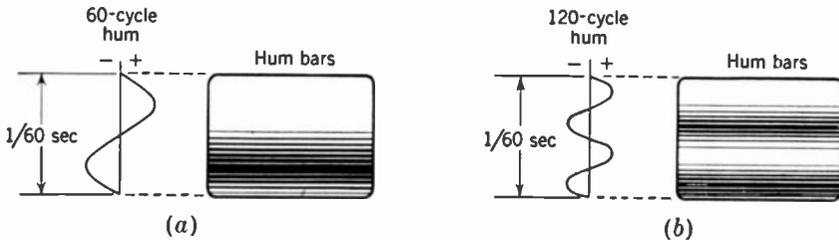


FIG. 11-23. Hum voltage and bars on kinescope screen. (a) 60 cps hum. (b) 120 cps hum.

approximately halfway down the screen during this time. Therefore, if the positive half cycle of the hum voltage begins at the same time as the vertical scan, the top half of the screen will be lighter than the bottom half. The screen then has two horizontal bars, one light and the other dark. If the frequency is 120 cps, as in *b* of Fig. 11-23, two pairs of bars are produced because one complete cycle of the hum voltage occurs during  $\frac{1}{20}$  sec to produce a pair of black-and-white bars during one-half the vertical scan. Two cycles of the 120-cps voltage produce two pairs of bars during the field interval of  $\frac{1}{60}$  sec. When the hum has the opposite polarity from the voltage shown in Fig. 11-23, the black-and-white bars will be reversed but there will be the same number. Often one of the hum bars appears in two sections, a part at the bottom of the screen and the remainder of the bar at the top, because the phase of the hum voltage does not coincide with the start of vertical scanning.

Any 60- or 120-cps hum voltage in the video section of the receiver is amplified by the video stages because these frequencies are in the video-frequency range of 30 cps to 4 Mc. The hum voltage may be present in the video amplifier as the result of modulation hum produced in either

the r-f or i-f stages and rectified in the video detector, or the hum can be introduced in the video amplifier section itself. Hum can be produced in the video section by heater-to-cathode leakage in a tube, or because of insufficient filtering of the B supply voltage. Note that the filament hum always has a frequency of 60 cps, while hum from the B supply is 120 cps in a full-wave power supply. Hum voltage that is introduced in the video section itself is the additive type of hum, which does not require any modulation and is present with or without a signal. Hum bars that are in the raster without any picture, therefore, are caused by hum introduced in the video section.

**11-12. Troubles in the Video Amplifier.** Since the video amplifier supplies the video signal to the kinescope for reproduction of the image,

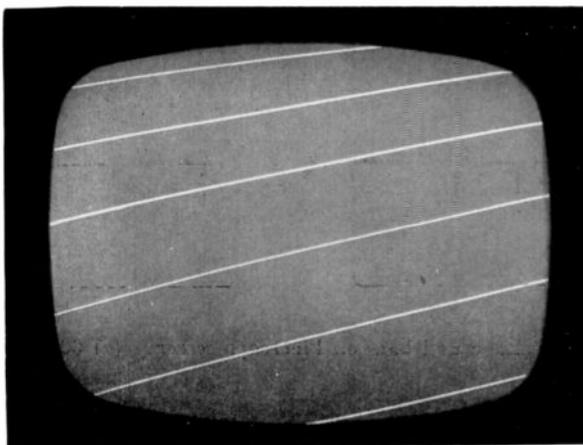


FIG. 11-24. No picture on kinescope screen. Raster is present, without snow.

troubles in the video amplifier can cause no picture, weak picture, or distorted picture quality, while the raster is normal. The grid-cathode circuit of the kinescope is part of the video signal circuits. It is important to note that these same troubles can also be produced by the i-f and r-f amplifiers because the composite video signal is the envelope of the modulated picture carrier signal amplified in the i-f and r-f stages.

*No Picture.* If the video amplifier does not provide video signal for the kinescope grid-cathode circuit, there will be no picture. The deflection circuits, however, can still produce the raster. Figure 11-24 shows a blank raster without any picture because of a defective video amplifier. In order to see the raster without a picture, it may be necessary to turn up the brightness control.

*Weak Picture.* Insufficient amplification of the video signal in the video amplifier can result in a weak picture, with insufficient contrast, as shown in Fig. 11-25.

*Smear in the Picture.* In one form of smear, the edges between black and white areas trail off to the right indistinctly within  $\frac{1}{2}$  in. on a 20-in.



FIG. 11-25. Weak picture, without snow. (RCA.)

screen, approximately, as illustrated in the test pattern in Fig. 11-10. This *short smear*, or *edge smear*, is caused by insufficient high-frequency response. However, severe smear and streaking in large areas of the picture, as in Fig. 11-28, is produced by excessive gain with phase distortion for the low video frequencies.

Figure 11-26 illustrates why insufficient high-frequency response can cause edge smear. The square wave at the top in the illustration corresponds to the picture information obtained in scanning across the black bar, which is a small part of a horizontal line. Because of loss of the high frequencies and phase distortion, the square wave is changed to the signal shown below with rounded corners, less slope, and more time delay in the leading and trailing edges. Physically, the delay is a result of the time necessary for the shunt capacitances in the video amplifier circuit to charge and discharge as the signal voltage changes rapidly at a transition between black and white. The resultant picture information does not have sharp edges, therefore, and the right-hand edge of the bar trails off slowly from black to white past the position where the black bar should end. When the right-hand edge trails off this way for every horizontal scanning line, the black bar has short smear to the right. If the bar were white on a black background, the white has short smear to the right.

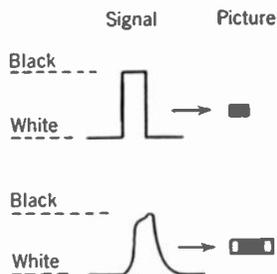


FIG. 11-26. Diagram showing how insufficient response for the high video frequencies causes edge smear.

The reason why excessive smear and streaking in large areas of the picture results from excessive low-frequency response and the associated phase distortion is illustrated in Fig. 11-27, for the case of a 15-ke square-wave signal that corresponds to a black bar having one-half the width of the picture. The excessive low-frequency amplification causes a lagging phase-angle response that produces time delay. Physically, the delay is the time necessary for the bypass condensers in the video amplifier plate decoupling filter circuits to charge and discharge when the signal amplitude changes relatively slowly. In terms of picture information, the square wave at the top of Fig. 11-27 reproduces the black bar correctly, but the delayed waveform below produces severe smear toward the right, extending the black bar across the picture. The peculiar effect where a long horizontal strip in the reproduced picture can be seen through a person standing in front of it is caused by this type of smear. A reproduced test pattern with excessive low-frequency response and time-delay

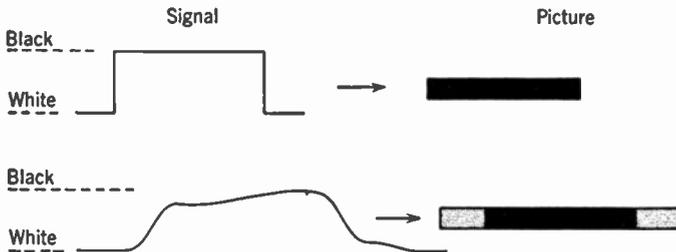


FIG. 11-27. How excessive low-frequency response with phase distortion causes severe smear.

distortion is shown in Fig. 11-28. Notice that the streaking of the black lettering causes a dark area stretching to the right, like a shadow. Note also that the smear to the right can extend over to the next line and appear at the left of the smeared picture information. A trouble that produces excessive low-frequency response in the video amplifier often causes insufficient high-frequency response also, resulting in smear over large areas and edge smear at the same time.

*Trailing Outlines in the Picture.* Excessive response for the high video signal frequencies can allow the peaking coils to produce shock-excited oscillations, or *ringing*, immediately after a rapid change in signal voltage. Since the edge between black and light areas in the horizontal direction corresponds to a sudden change in signal voltage, the ringing follows vertical outlines in the picture. The oscillations produce duplicate outlines that appear reversed from the original edge. The most noticeable effect of the ringing usually is a white outline trailing a black edge that has a light area on the right, as shown in Fig. 11-29. Generally the first outline is most obvious but it can be followed by additional evenly spaced dupli-

cate outlines. In addition, the ringing can cause a leading white outline at the left of a black edge.

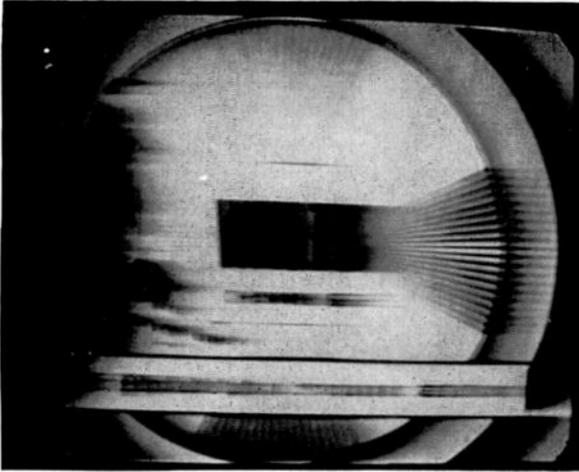


FIG. 11-28. Severe smear caused by excessive low-frequency response with phase distortion. (RCA Pict-O-Guide.)



FIG. 11-29. Test pattern with outline distortion caused by ringing in the video amplifier. (RCA Pict-O-Guide.)

The reason for the reversed outlines when the video amplifier circuit rings is illustrated in Fig. 11-30. The top waveform shows the video signal voltage corresponding to the black bar at the right. Below this are the oscillating voltage waves caused by shock excitation of the peaking coils when the signal voltage changes rapidly, with the corresponding thin black and white lines they produce in the picture. The voltage waveform

at the bottom of the figure combines the original signal with the oscillations. Notice that the oscillating voltage produces *overshoots* where the resultant wave rises too high and *undershoots* where the voltage drops too low. The undershoot at point 4 in the waveform causes a trailing white outline after the black bar. The undershoot at point 1 produces a leading white outline to the left of the black bar. The overshoots at points 2 and 3 make the signal voltage blacker than black, but this is not evident in the reproduced picture. A small amount of ringing in the video amplifier, equivalent to about 20 per cent increase in high-frequency response,

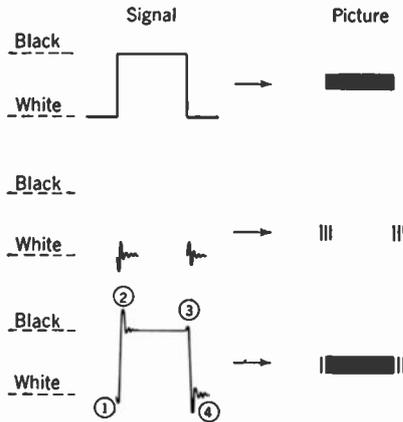


FIG. 11-30. How excessive response for the high video frequencies causes outline distortion.

may be desired in the video amplifier in order to make the picture look sharper.

*Summary of Picture-quality Distortions.* The effects in the picture of frequency distortion and phase distortion of the video signal frequencies can be summarized as follows:

1. Insufficient horizontal detail and lack of sharpness at vertical outlines between black and white edges are caused by insufficient high-frequency response. Short smear to the right of the edges also results with phase distortion for the high video frequencies.

2. Reversed outlines, primarily to the right of vertical edges between black and white, are caused by excessive response for the high video frequencies.

3. Weak picture information for large areas in the image results from insufficient response for low video frequencies.

4. Severe smear and streaking in large areas of the picture are caused by excessive gain with phase distortion for the low video frequencies.

It is important to note that these same distortions of picture quality can be caused by frequency distortion and phase distortion for the

corresponding r-f or i-f signal frequencies, especially in the i-f signal, since most of the receiver's gain is provided by the i-f amplifier stages.

## REVIEW QUESTIONS

1. What is the difference between video signal and picture signal?
2. What is the function of the video signal voltage coupled to the kinescope grid-cathode circuit?
3. What determines the average brightness of the picture reproduced on the kinescope screen?
4. What determines the contrast of the picture reproduced on the kinescope screen?
5. What is meant by *black* in the picture reproduced on the kinescope screen?
6. Describe the effects on the picture when the video signal voltage coupled to the picture tube grid is of negative picture phase.
7. Describe two methods of obtaining manual contrast control in the television receiver.
8. If a tube has the load resistance in the cathode circuit instead of the plate circuit, will there be phase inversion of the input signal? Why?
9. What kind of distortion would result from incorrect grid bias on the video amplifier?
10. What is meant by frequency distortion in an amplifier?
11. What is meant by phase distortion in an amplifier?
12. What is the effect on the reproduced picture of loss of the high video frequencies?
13. What is the effect on the reproduced picture of excessive low-frequency response?
14. What are the advantages and disadvantages of using the resistance-capacitance-coupled amplifier for video amplification?
15. Give one advantage and one disadvantage of the direct-coupled amplifier.
16. If the total shunt capacitance in the amplifier were zero, would high-frequency compensation be necessary? Why?
17. What is the video signal frequency corresponding to the information in a solid white horizontal line with a length equal to the picture width?
18. Using a 4,000-ohm plate load resistor with the 6AG7 video output tube, having a  $g_m$  of 7,000  $\mu\text{mhos}$ , what is the gain of the stage?
19. Referring to the schematic diagram in Fig. 11-21, give the function at the following components:  $R_{322}$ ,  $R_{320}$ ,  $L_{303}$ ,  $R_{208B}$ ,  $R_{326}$ ,  $C_{310}$ ,  $C_{309}$ , and  $R_{324}$ .
20. Referring to the schematic diagram in Fig. 11-22, give the function of the following components:  $R_{140}$ ,  $L_{190}$ ,  $R_{139}$ ,  $R_{141}$ ,  $C_{223B}$ ,  $C_{140}$ ,  $C_{141}$ , and  $R_{147}$ .
21. What could be the defect causing the following trouble: one pair of hum bars in the picture and in the raster, but no hum in the sound? This is an intercarrier-sound receiver with the sound take-off circuit in the video detector output circuit.
22. Why will a weak video amplifier tube cause insufficient contrast in the reproduced picture?
23. What are two troubles in the video amplifier that can cause smear in the picture?
24. Explain briefly why the trouble symptom of no picture with a normal raster, on all channels, can be due to the video amplifier.
25. Referring to the direct-coupled video amplifier circuit in Fig. 11-21, why can zero plate current in  $V_{306}$  result in no brightness on the kinescope screen?

## CHAPTER 12

### PRACTICAL DESIGN OF VIDEO AMPLIFIERS

The basic requirement of the video amplifier is that it be able to amplify the video signal and still preserve its complex waveform so that proper reproduction of the picture can be obtained. In order to do this, the amplifier must respond to a frequency range with limits of 30 cps at the low end and 4.0 Mc at the high end, approximately, without appreciable frequency or phase distortion. In general, it is not possible to maintain constant both the gain and the time delay of the amplifier over the entire video-frequency range, and a compromise must be made, with neither the gain nor the delay exactly constant but with both satisfactory in terms of the picture reproduction. For good low-frequency response, constant time delay is more important than constant gain in terms of the picture reproduction. At the high-frequency end, it is more important that the gain of the stage be uniform for all frequencies. In addition to the fact that phase distortion at these high video frequencies introduces a relatively small amount of time delay, the compensation method used for uniform gain in the amplifier usually reduces the time-delay distortion also.

**12-1. Gain.** The gain of an amplifier, which is equal to the ratio of output signal voltage divided by the input voltage, varies directly with the size of the plate load. This can be proved with the aid of Fig. 12-1, showing the plate circuit of the vacuum-tube amplifier as an equivalent a-c generator, generating a voltage equal to  $\mu e_g$ . The generator has an internal resistance equal to  $R_p$ , the internal plate-to-cathode resistance of the tube, and is connected to a load  $Z_L$  which is the external plate-to-cathode circuit. For the video amplifier,  $Z_L$  is equal to the plate load resistor  $R_L$  combined with any reactance in the plate circuit. This equivalent circuit applies for any amplifier, triode, tetrode, or pentode, as long as the tube is operating over the linear portion of its characteristic curve. That the equivalent circuit is valid can be seen from the following considerations:

1. The input signal voltage  $e_g$  is amplified by a factor equal to the  $\mu$  of the tube, but this is not the output signal voltage. The amplified signal

voltage  $\mu e_o$  is divided between the internal plate resistance of the tube and the external plate load, the two being in series as a voltage divider.

2. Any current that flows through the external plate load must flow through the cathode-to-plate resistance of the tube. Therefore,  $R_p$  and  $Z_L$  must be in series with each other across the generator.

3. The plate load  $Z_L$  can be considered as connected between plate and cathode for the a-c signal voltage because the B supply is effectively bypassed to the cathode.

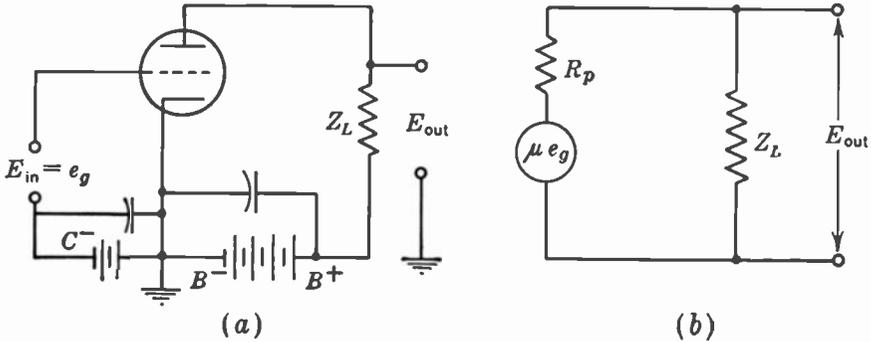


FIG. 12-1. A vacuum-tube amplifier circuit. (a) Typical operation. (b) Equivalent plate circuit.

Analyzing the equivalent circuit of Fig. 12-1, the signal plate current that flows is the generator voltage divided by the total impedance across the generator, which is equal to  $R_p$  in series with  $Z_L$ .

$$i_p = \frac{\mu e_o}{R_p + Z_L}$$

The output signal voltage available across  $Z_L$  is this plate current multiplied by  $Z_L$ .

$$\begin{aligned} E_{out} &= \frac{\mu e_o}{R_p + Z_L} \times Z_L \\ &= \frac{\mu e_o \times Z_L}{R_p + Z_L} \end{aligned}$$

The gain of the stage is  $E_{out}$  divided by  $E_{in}$ , and  $E_{in}$  is  $e_o$ . Therefore,

$$\text{Gain} = \frac{E_{out}}{E_{in}} = \frac{(\mu e_o \times Z_L) / (R_p + Z_L)}{e_o}$$

Canceling  $e_o$  and factoring out  $\mu$ ,

$$\text{Gain} = \mu \times \frac{Z_L}{R_p + Z_L} \tag{12-1}$$

The gain of the stage, therefore, is equal to the amplification factor of the tube  $\mu$  multiplied by a factor  $Z_L/(R_p + Z_L)$ , which varies with the value of  $Z_L$  compared to  $R_p$ . In any case, the gain cannot be greater than the  $\mu$  of the tube because the fraction  $Z_L/(R_p + Z_L)$  can never exceed the value of one. If the plate load is very large compared to  $R_p$ , then the fraction approaches the value of one and the gain of the stage approaches the  $\mu$  of the tube. To obtain the highest possible voltage gain with a given tube, then, the value of the plate load impedance should be made as high as possible. However, the degree of which  $Z_L$  can be increased is limited by the fact that the  $IR$  drop across the plate load resistor reduces the plate-to-cathode voltage.

**12-2. Video Amplifier Gain.** The general expression for the gain of an amplifier can be simplified for the common case of a video amplifier using a pentode tube and a small value of plate load resistor. Pentode and beam-power tubes are used because of their high  $\mu$  and  $g_m$  and low value of grid-to-plate capacitance. The low value of plate load resistor is necessary in order to minimize the effect of shunt capacitance in the plate circuit so that the gain can be extended to a frequency as high as possible. Since the pentodes have very high internal plate resistance, many of them being 1 megohm or more, and the plate load is 5,000 ohms or less,  $Z_L$  is negligible compared to  $R_p$ . Therefore, the gain of the video amplifier is given by the expression

$$\text{Gain} = \mu \frac{Z_L}{R_p} = \frac{\mu}{R_p} \times Z_L$$

because  $Z_L + R_p$  is in effect equal to  $R_p$ .

This expression for the gain of a typical video amplifier can be simplified further. The three characteristics of a vacuum tube—amplification factor ( $\mu$ ), plate resistance ( $R_p$ ), and transconductance ( $g_m$ )—are related to each other in the following ways:

$$g_m = \frac{\mu}{R_p} \quad R_p = \frac{\mu}{g_m} \quad \mu = R_p \times g_m$$

Since  $\mu/R_p$  equals  $g_m$ , the gain of the video amplifier is given directly by the expression

$$\text{Gain} = g_m \times Z_L \quad (12-2)$$

The gain of the stage depends on the  $g_m$  of the tube, rather than its  $\mu$ , because of the low value of plate load. The value of  $g_m$  for the tube is available from a tube manual, and if the value of  $Z_L$  is known the gain of the stage can be calculated.

**Example.** A video output stage uses the 6CL6 pentode tube with a plate-to-cathode voltage of 250 volts and grid bias of  $-3$  volts, providing a  $g_m$  of 11,000  $\mu$ mhos.

The plate load resistor is 4,000 ohms. What is the gain of the stage for the middle range of frequencies, where the plate load impedance is equal to the plate load resistor  $R_L$ ?

$$\begin{aligned}\text{Gain} &= g_m \times Z_L = g_m \times R_L \\ &= 0.011000 \text{ mho} \times 4,000 \text{ ohms} \\ &= 44\end{aligned}$$

In order to have uniform gain or flat response in the video amplifier, the product  $g_m \times Z_L$  must be constant throughout the desired frequency range. The value of  $g_m$  is the same for all frequencies because it is a physical constant of the tube whose value depends on the applied plate-to-cathode voltage and the grid bias. The plate load impedance  $Z_L$  is the factor that varies with frequency because it includes any reactance in the plate circuit. As the magnitude of  $Z_L$  varies with frequency because of reactance, the relative gain of the amplifier will vary in exactly the same manner. The response of the uncompensated  $RC$  amplifier decreases at the high-frequency end because  $Z_L$  decreases with the reduced reactance of the shunt capacitance in the plate circuit at high frequencies.

**12-3. Shunt Capacitance.** The total shunt capacitance  $C_t$  is the

only reason that the amplifier response is down at the high-frequency end. It is very important, therefore, to keep  $C_t$  to as low a value as possible.

*Components of  $C_t$ .* As illustrated in Fig. 12-2, the total shunt capacitance is equal to the sum of four capacitances always present in the amplifier.

$$C_t = C_{\text{out}} + C_{\text{in}} + C_m + C_{\text{stray}}$$

$C_{\text{out}}$  is the static output capacitance from plate to cathode of the tube itself. The television pentode amplifier tube 6AC7 has an output capacitance of  $5 \mu\text{mf}$ , as given in the tube manual. Similarly,  $C_{\text{in}}$  is the static input capacitance between grid and cathode. This is taken for the succeeding tube, since the input circuit of the next stage is part of the plate load for the previous tube. The 6AC7 has an input capacitance of  $11 \mu\text{mf}$ . In the case of a video output stage driving the picture tube, the input capacitance of the kinescope grid circuit is an important factor in the high-frequency response of the amplifier and must be included in its total shunt capacitance. The input capacitance of picture tubes is approximately  $6 \mu\text{mf}$ .  $C_m$  is a dynamic input capacitance, which is added to the static value of input grid-to-cathode capacitance when the tube functions as an amplifier. This increase of input capacitance in an amplifier stage is the result of the Miller effect and can increase to a great extent

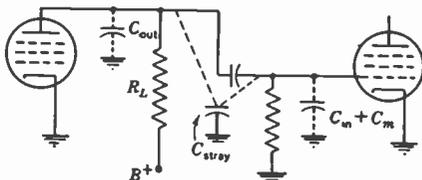


Fig. 12-2. The shunt capacitances in an amplifier.

the dynamic input capacitance of a triode, or of a pentode having a great deal of gain.

$C_{\text{stray}}$  includes capacitance to the chassis ground of the wiring, component parts such as the coupling condenser, plate load resistor, and peaking coils, in addition to capacitances between the tube elements through the tube socket. These are not included in the tube manual's listing of the capacitances but can add appreciably to the total shunt capacitance  $C_t$ .

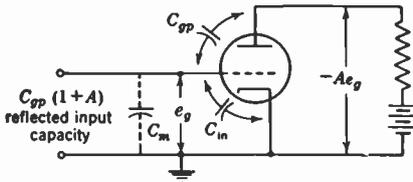


FIG. 12-3. The Miller effect reflects additional capacitance into the input circuit through the grid-to-plate capacitance.

The stray capacitances can be held to a minimum, greatly improving the amplifier's high-frequency response, by short wiring, use of low-loss sockets, and proper placement of parts. The coupling elements should be mounted away from and perpendicular to the chassis to reduce the strays. A rough estimate of the

stray capacitance in a video amplifier stage is about 5 to 15  $\mu\text{mf}$ .

**Miller Effect.** The amount of input capacitance added by the Miller effect for the case of a resistive plate load is  $C_{gp}(1 + A)$ , as shown in Fig. 12-3, where  $A$  is the gain of the stage, and  $C_{gp}$  is the grid-to-plate capacitance of the tube. This additional effective input capacitance results because the difference in potential between grid and plate charges the input grid capacitance through the grid-to-plate capacitance. In any condenser,  $Q = CV$ , where  $V$  is the potential difference between the plates,  $Q$  is the charge on either plate, and  $C$  is the capacitance. Increasing the charge on the condenser  $C_{in}$  with the same input voltage  $e_g$  effectively increases the input capacitance. The dynamic input capacitance of the tube, therefore, is

$$C_{in} + C_{gp}(1 + A)$$

where  $C_{in}$  is the static input capacitance always present in the tube, and  $C_{gp}(1 + A)$  is the capacitance added to the grid circuit because of the Miller effect when the tube is amplifying the input signal.

**Example 1.** Find the total input capacitance of the 6AC7 operating with a gain of 14. From the tube manual, the static input capacitance of the 6AC7 is 11  $\mu\text{mf}$  and  $C_{gp}$  is 0.015  $\mu\text{mf}$ .

$$\begin{aligned} C &= C_{in} + C_{gp}(1 + A) \\ &= 11 + 0.015(1 + 14) \\ &= 11 + 0.225 \\ &= 11.225 \mu\text{mf} \end{aligned}$$

**Example 2.** Find the total input capacitance of the triode amplifier 6J5 operating with a gain of 14. From the tube manual  $C_{in}$  is 3.4  $\mu\text{mf}$  and  $C_{gp}$  is 3.4  $\mu\text{mf}$ .

$$\begin{aligned}
 C &= C_{in} + C_{op} (1 + A) \\
 &= 3.4 + 3.4 (1 + 14) \\
 &= 3.4 + 51.0 \\
 &= 54.4 \mu\mu f
 \end{aligned}$$

From these examples it is seen that with a pentode tube the additional input capacitance resulting from the Miller effect is small because of the extremely low value of  $C_{op}$ . In a triode, however, the high value of grid-to-plate capacitance reflects back into the grid circuit when the tube is used as an amplifier stage, greatly increasing the dynamic input capacitance. When the gain of a stage is very high, though, even a small amount of grid-to-plate capacitance can increase the input capacitance appreciably.

*Calculating  $C_i$ .* The first step in designing the video amplifier is to find the value of the shunt capacitance, since this is a direct measure of the high-frequency response of the amplifier and how much compensation is needed at the high end.  $C_i$  can be calculated, if desired, as the sum of the individual capacitances.  $C_{out}$ ,  $C_{in}$ , and  $C_{op}$  are available from the tube manual, and the gain of the next stage can be measured or calculated from the expression  $\text{Gain} = g_m \times R_L$ .

**Example.** What is  $C_i$  for a 6AU6 video amplifier driving a 6K6-GT video output tube whose gain is 10? The stray capacitance is 5  $\mu\mu f$ .

$$\begin{aligned}
 C_i &= C_{out} + C_{stray} + C_{in} + C_{op} (1 + A) \\
 &= 5 + 5 + 5.5 + 0.5 (1 + 10) \\
 &= 5 + 5 + 5.5 + 5.5 \\
 &= 21 \mu\mu f
 \end{aligned}$$

*Measuring  $C_i$ .* Better than calculating  $C_i$  is finding its value experimentally by resonating with a known value of inductance in the plate circuit of the video amplifier. Unmodulated signal from an r-f generator is coupled to the video amplifier control-grid circuit, and the amplified output is measured with an a-c voltmeter capable of measurements at the high video frequencies used. Varying the frequency of the applied signal, the output voltage will show a marked increase at resonance when the reactance of the inductance equals the reactance of  $C_i$ . Normally, no output reading is obtained until resonance is reached because the coil is inserted in the plate circuit of the amplifier in place of the plate load resistor, which is omitted when making this measurement in order to obtain sharper resonance. The coil should be similar to the actual peaking coil that will be used for the high-frequency compensation so that  $C_i$  will not change after the measurement.

With the value of inductance and the resonant frequency known,  $C_i$  can be computed from the resonant frequency formula

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

$$f_r^2 = \frac{1}{4\pi^2 LC}$$

$$C = \frac{1}{4\pi^2 f_r^2 L} = \frac{1}{(2\pi f_r)^2 L}$$

The value of  $C_i$  is usually about 20 to 30  $\mu\mu\text{f}$ , using pentodes with the stray capacitances held down to a minimum. If the output voltmeter is not isolated from the video circuit being measured, the input capacitance of the meter must be subtracted from the measured capacitance to obtain the  $C_i$  of the amplifier itself. Another method of measuring  $C_i$  in terms

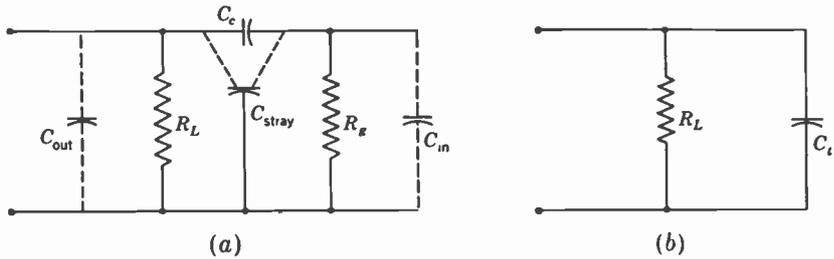


FIG. 12-4. Plate load impedance of the uncompensated video amplifier. (a) Actual circuit. (b) Equivalent circuit.

of the known high-frequency response of the amplifier is described in the next section.

**12-4. High-frequency Response of the Uncompensated Amplifier.** The manner in which the gain of the amplifier varies with frequency can be seen by examining the nature of the plate load. Since the gain is equal to  $g_m Z_L$  and  $g_m$  is independent of frequency, the gain varies with frequency in the same way that  $Z_L$  does.

*High-frequency Equivalent Circuit.* At the high-frequency end of the video band, the plate load impedance  $Z_L$  is equal to the plate resistor  $R_L$  in parallel with the total shunt capacitance  $C_i$ . That this is so can be seen from the following:

1. The reactance of the coupling condenser  $C_c$  is negligibly small for the high video frequencies.
2. This puts all the individual input and output capacitances in parallel with the plate resistor  $R_L$  and grid resistor  $R_g$ .
3. The effective plate load, then, consists of the total shunt capacitance  $C_i$ , the plate resistor  $R_L$ , and the grid resistor  $R_g$ , all in parallel. However,  $R_L$  is usually quite small, while  $R_g$  is at least 100,000 ohms. The parallel combination of  $R_L$  and  $R_g$ , therefore, is practically equal to the resistance of  $R_L$ , the smaller one.

4. Thus the effective plate load  $Z_L$  is equal to  $R_L$  in parallel with  $C_t$ , as shown in Fig. 12-4.

In order to analyze the amplifier response at high frequencies, it is only necessary to see how the equivalent plate load  $Z_L$  varies with frequency. Assuming a value for  $R_L$  of 3,000 ohms and for  $C_t$  of 30  $\mu\text{f}$ , this can be done with numeral examples. At mid-frequencies, the capacitive reactance of  $C_t$  is so large that it has practically no shunting effect and  $Z_L$  has the same value as  $R_L$ —3,000 ohms. Taking 10,000 cycles as an example, the reactance of  $C_t$  at this frequency is

$$\begin{aligned} X_c &= \frac{1}{2\pi fC} \\ &= \frac{1}{2\pi \times 10^4 \times 30 \times 10^{-12}} \\ &= 530,000 \text{ ohms (approx)} \end{aligned}$$

The resultant of 3,000 ohms resistance in parallel with 530,000 ohms of capacitive reactance is practically equal to 3,000 ohms. At a frequency of 1 Mc, however, the reactance of  $C_t$  is 5,300 ohms. This reactance in parallel with 3,000 ohms produces a resultant impedance of less than 3,000 ohms. Consequently,  $Z_L$  is less than 3,000 ohms for the high video frequencies and the gain of the amplifier is down in the same proportion as  $Z_L$ . It should be noted that the reactance of  $C_t$  and resistance of  $R_L$  must be added in parallel vectorially.

*Definition of  $F_2$ .* Since the reactance of  $C_t$  is very great compared to  $R_L$  at low and middle frequencies, and continuously decreases as the frequency becomes higher, there will be some frequency where the reactance of  $C_t$  equals  $R_L$ . This frequency at which  $X_c$  equals  $R_L$  is denoted here as  $F_2$ , and is a convenient point at which to compute the equivalent impedance of the parallel combination of  $R_L$  and  $X_c$ .

The resultant impedance  $Z_L$  of this parallel combination can be evaluated vectorially by the method illustrated in Fig. 12-5. Consider the equivalent plate load of  $R_L$  and  $X_c$  in parallel connected across an a-c generator. Since they are equal, each will draw the same amount of current  $I_1$  and  $I_2$  although they differ in phase of  $90^\circ$ . The total current drawn from the generator is the vector sum of the two currents. The actual value of the currents depends on the generator voltage and the value of  $R_L$  or  $X_c$ , but this does not matter here. Only the fact that they are equal is of consequence, and a value of 1 ma is taken for the individual branch currents. Adding the two branch currents vectorially, the total current is  $\sqrt{2}$  ma, using the parallelogram method for adding two vectors  $90^\circ$  out of phase with each other, as shown in Fig. 12-5b. With a current of 1 ma in each branch and a total of  $\sqrt{2}$  ma for both parallel legs, the total impedance of the parallel combination is  $1/\sqrt{2}$  of either branch imped-

ance because the impedance is inversely proportional to the current. The factor  $1/\sqrt{2}$  is equal to 0.707. Therefore, when  $X_{c_i}$  is equal to  $R_L$ , the total impedance  $Z_L$  of the two parallel branches is 0.707, or 70.7 per cent of  $R_L$ , and the gain of the amplifier is reduced proportionately. At mid-frequencies the plate load impedance equals  $R_L$  because  $X_{c_i}$  is too large to have any shunting effect. At the frequency  $F_2$ , where the plate load impedance is 70.7 per cent of  $R_L$ , the voltage gain is down 29.3 per cent. This is a loss of 3 db, and  $F_2$  is conveniently defined as the frequency at

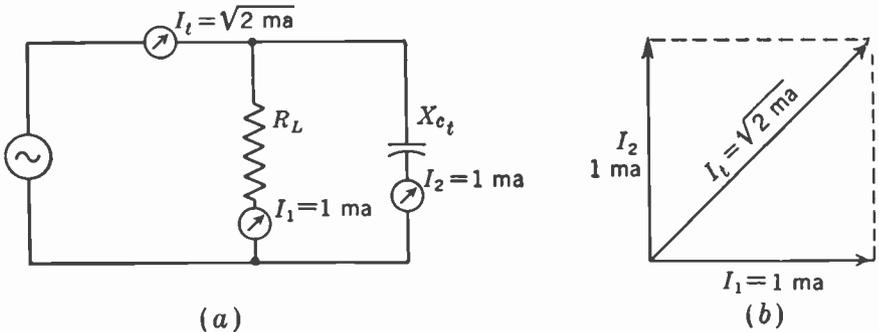


Fig. 12-5. When  $R_L$  equals  $X_{c_i}$ , the total impedance of the high-frequency equivalent plate load circuit is equal to  $(1/\sqrt{2}) \times R_L$ . (a) Plate load connected across an a-c generator. (b) Vector addition of the two branch currents.

which the high-frequency response of the uncompensated amplifier is down 3 db, compared to the gain at mid-frequencies.

The gain of the uncompensated video amplifier is down 3 db at  $F_2$  because  $X_{c_i}$  equals  $R_L$  at that frequency. The reactance of  $C_i$  at this frequency  $F_2$  must equal

$$X_{c_i} = \frac{1}{2\pi F_2 C_i}$$

Also, this value of  $X_{c_i}$  is equal to  $R_L$  by the definition of  $F_2$ . Therefore

$$R_L = \frac{1}{2\pi F_2 C_i}$$

or

$$F_2 = \frac{1}{2\pi R_L C_i} \quad (12-3)$$

where  $R$  is in ohms and  $C$  in farads to give  $F_2$  in cycles per second. The value of  $F_2$ , therefore, is inversely proportional to  $R_L$  and  $C_i$ . In order to have a high value of  $F_2$  and good high-frequency response, both  $R_L$  and  $C_i$  must be small. A few examples will demonstrate the usefulness of this formula.

**Example 1.** An uncompensated video amplifier has a plate load resistor of 4,000 ohms and a total shunt capacitance of 18  $\mu\text{f}$ . At what frequency is the gain down 3 db? What is the mid-frequency gain, using a tube with a  $g_m$  of 6,000  $\mu\text{mhos}$ ?

$$\begin{aligned}
 F_2 &= \frac{1}{2\pi R_L C_t} \\
 &= \frac{1}{2\pi \times 4,000 \times 18 \times 10^{-12}} \\
 &= 2.2 \times 10^6 \text{ cps (approx)} \\
 &= 2.2 \text{ Mc} \\
 \text{Gain} &= g_m \times R_L \\
 &= 0.006 \times 4,000 \\
 &= 24
 \end{aligned}$$

**Example 2.** What value of  $R_L$  is needed to make  $F_2$  equal 4 Mc for the amplifier of Example 1? What is the mid-frequency gain using a tube with the same  $g_m$ ?

$$F_2 = \frac{1}{2\pi R_L C_t}$$

or

$$R_L = \frac{1}{2\pi F_2 C_t}$$

$$\begin{aligned}
 R_L &= \frac{1}{2\pi \times 4 \times 10^6 \times 18 \times 10^{-12}} \\
 &= 2,200 \text{ ohms (approx)}
 \end{aligned}$$

$$\begin{aligned}
 \text{Gain} &= g_m \times R_L \\
 &= 0.006 \times 2,200 \\
 &= 13.2
 \end{aligned}$$

In general,  $C_t$  is made as small as possible with special attention to the strays, and  $R_L$  is given the value required for a desired value of  $F_2$ . While it would seem that any value of  $F_2$  can be obtained by suitable choice of the value of  $R_L$ , there are practical difficulties. Since the gain of the stage is equal to  $g_m R_L$ , the smaller the value of  $R_L$  the less the gain. It is much better to obtain a high value of  $F_2$  by keeping  $C_t$  down to a minimum. In this way,  $R_L$  can have a higher value for a given  $F_2$  and the gain of the stage will be greater. The effect of different values of  $R_L$  on the amplifier response curve, with a fixed value of  $C_t$ , is shown in Fig. 12-6.

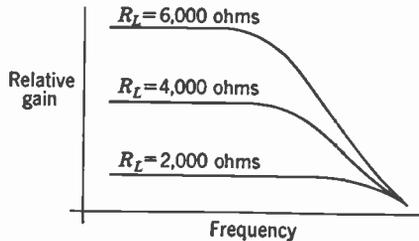


FIG. 12-6. Relative gain vs. frequency for different values of plate load resistance.

The frequency  $F_2$  is an important measure of the high-frequency response, even though the amplifier is not usable up to this frequency without compensation. The higher the value of  $F_2$  the better is the high-frequency response of the amplifier. However, it is not possible to use the

video amplifier up to  $F_2$  without compensation. At this point the gain is 70.7 per cent of the mid-frequency response, representing a loss in gain of 29.3 per cent, and a variation in gain of more than 10 per cent is not flat response in the video amplifier. The uncompensated amplifier is flat only to the frequency equal to  $0.1 F_2$ . As an example, if  $F_2$  is 4 Mc, without compensation the amplifier is very nearly flat to 0.4 Mc or 400 kc. Last and most important, when the amplifier is compensated with the design procedures given here it is made flat to the frequency  $F_2$ . When designing the amplifier by this method,  $R_L$  is chosen so that, with the experimentally determined value of  $C_t$ ,  $F_2$  will be the highest video frequency to be amplified. Then compensation is added to make the amplifier response flat to  $F_2$  within 10 per cent instead of being down 3 db at this frequency.

*Measuring  $C_t$  in Terms of  $F_2$ .* The fact that the voltage gain of the uncompensated resistance-coupled amplifier is down 29.3 per cent at  $F_2$  provides a method of calculating  $C_t$  from dynamic measurements on the video amplifier. If  $F_2$  is measured with a known value of  $R_L$  in the plate circuit,  $C_t$  can be computed from the formula

$$F_2 = \frac{1}{2\pi R_L C_t} \quad \text{or} \quad C_t = \frac{1}{2\pi R_L F_2}$$

The procedure can be as follows:

1. Insert a known value of  $R_L$  in the video amplifier plate circuit. This can be any value, say 2,000 to 5,000 ohms, as long as its value is known.
2. Measure the a-c signal voltage output of the amplifier with a known value of input voltage from a signal generator at some middle frequency, say 10 kc.
3. By varying the input frequency but keeping the input voltage constant, find the frequency at which the output voltage is down to 70.7 per cent of its mid-frequency value. This is  $F_2$ .
4. With  $F_2$  and  $R_L$  known,  $C_t$  can be computed from the formula  $C_t = 1/(2\pi R_L F_2)$ .

**12-5. Shunt Peaking.** Once  $C_t$  is known, the value of plate load resistor can be chosen for the desired value of  $F_2$ . This frequency is chosen as the limit up to which the video amplifier response is desired to be flat, because the high-frequency compensation is designed to make the amplifier response uniform up to  $F_2$ . The simplest method of doing this is with the shunt peaking circuit, as shown in Fig. 12-7.

*Plate Load Impedance.* With the peaking coil inserted in series with  $R_L$ , the equivalent plate load illustrated in Fig. 12-7b takes a form different from the case of the uncompensated amplifier. Analysis of this network to find its total impedance  $Z_L$  will show how the gain of the amplifier varies because the gain varies with frequency in exactly the same way as

$Z_L$  does. The impedance of the  $LR$  branch is the vector sum of the inductive reactance of the peaking coil  $X_{L_o}$ , in series with the plate load resistor  $R_L$ . The impedances in parallel are combined vectorially to provide the total impedance of the plate load network.

$$Z_L = X_{c_i} \sqrt{\frac{R_L^2 + X_{L_o}^2}{R_L^2 + (X_{c_i} - X_{L_o})^2}}$$

If  $R_L$ ,  $C_t$ , and  $L_o$  are known, the impedance of the shunt peaking equivalent network can be found for any frequency by substituting the values of  $R_L$ ,  $X_{c_i}$ , and  $X_{L_o}$  for that frequency.

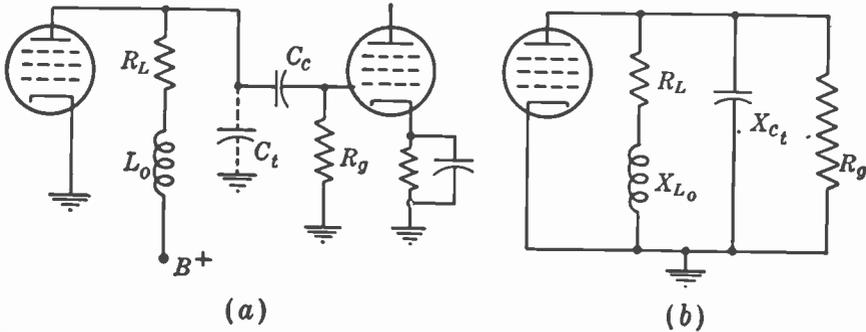


FIG. 12-7. Shunt peaking.  $R_L$  and  $L_o$  can be reversed in the circuit, but the connection shown reduces the stray capacitance. (a) Shunt-peaked stage. (b) Equivalent plate load circuit.

*Design Values.* An interesting and important case occurs when  $R_L$  equals  $X_{c_i}$ , and  $X_{L_o}$  is one-half either one. The impedance then becomes

$$Z_L = R_L \sqrt{\frac{R_L^2 + (R_L/2)^2}{R_L^2 + (R_L/2)^2}} = R_L \sqrt{1} = R_L$$

The network impedance is exactly equal to  $R_L$  for this condition. Taking a numerical case, suppose that  $X_{c_i}$  is 2,000 ohms at a given frequency. If  $R_L$  is 2,000 ohms and  $L_o$  has an inductance such that its reactance is 1,000 ohms at the same frequency, then

$$\begin{aligned} Z_L &= 2,000 \sqrt{\frac{2,000^2 + 1,000^2}{2,000^2 + 1,000^2}} \\ Z_L &= 2,000 \sqrt{1} = 2,000 \text{ ohms} \end{aligned}$$

Therefore, for the frequency at which  $X_{c_i}$  is equal to  $R_L$  and  $X_L$  is equal to  $0.5 R_L$  the impedance of the plate circuit and gain of the amplifier are exactly the same as at mid-frequencies. Since the frequency at which  $X_{c_i}$  equals  $R_L$  is  $F_2$ , if the peaking coil is made to have inductive reactance equal to  $0.5 R_L$  (or  $0.5 X_{c_i}$  at the frequency  $F_2$ ) the amplifier response will be flat to  $F_2$  instead of being down 3 db at this frequency. These values

are used for shunt peaking:

$$R_L = X_c \text{ at } F_2 = \frac{1}{2\pi F_2 C_t} \quad (12-4)$$

$$X_{L_o} = 0.5R_L = \frac{0.5}{2\pi F_2 C_t} \quad (12-5)$$

$$L_o = 0.5C_t R_L^2 \text{ (since } X_L = 2\pi fL) \quad (12-6)$$

where  $R$  and  $X$  are in ohms,  $C$  in farads,  $L$  in henrys, and  $f$  in cps.

A plot of impedance or relative gain against frequency for the amplifier shunt peaked with the above values is shown by curve 2 in Fig. 12-10. This figure shows universal response curves, applicable to any amplifier that is compensated by the method indicated. Plotting frequency in terms of the ratio  $f/F_2$  instead of actual frequency allows the curve to be used universally for amplifiers having different values of  $F_2$ . As an example, if  $F_2$  of the amplifier is 4 Mc and it is desired to know the response at 2 Mc, it can be read from the curve at  $f/F_2 = 0.5$ . Above a frequency equal to  $0.1F_2$ , where the uncompensated amplifier response would begin to fall, the gain of this shunt-peaked amplifier actually rises slightly. This rise continues, then decreases to make the gain at  $F_2$  exactly the same as at mid-frequencies. The slight rise in gain just below  $F_2$  can be tolerated because it deviates only 3 per cent from flat response.

The reason for the increase in impedance and gain over the uncompensated circuit is that the peaking coil resonates with  $C_t$  to form a parallel resonant circuit broadly tuned to the high-frequency end of the video band. The resonant circuit is broadly tuned with a very low  $Q$  because of the damping effect of  $R_L$  in series with coil  $L_o$  in the tuned circuit. The resonant frequency of the peaking circuit, where  $X_L = X_c$ , is not at  $0.6F_2$  where the maximum rise in impedance occurs, but is at some frequency higher than  $F_2$ . With such a high series resistance in the tuned circuit it has a resonance curve that is neither symmetrical nor sharply peaked, and the point of maximum impedance is not at the resonant frequency.

While these design values are not the only possibilities, they are suitable for obtaining uniform response in the shunt-peaked amplifier. If the peaking coil is too large, the  $Q$  of the compensated circuit is too high and the response rises too sharply just below  $F_2$ . If the coil is too small the plate impedance will not increase enough because of the low  $Q$ , and the response will fall down below  $F_2$ . The values given, then, represent a good compromise for maintaining the amplifier response flat up to  $F_2$ .

*Design Procedure for Shunt Peaking.* In practice, the procedure is this:

1. The highest frequency up to which the amplifier is desired flat is chosen as  $F_2$ . This may be from 2.5 to 4.5 Mc or higher, depending on the use of the video amplifier.

2.  $C_t$  is calculated or preferably measured in the chassis.
3. With  $C_t$  known,  $R_L$  is chosen as equal to the reactance of  $C_t$  at the top correction frequency  $F_2$  as decided in 1 above:  $R_L = 1/2\pi F_2 C_t$ .
4. The peaking coil  $L_o$  is chosen to have an inductance such that its inductive reactance at  $F_2$  is one-half  $R_L$

$$X_{L_o} = \frac{1}{2}R_L \text{ ohms (at } F_2\text{)}$$

or

$$L_o = 0.5C_t R_L^2 \text{ henrys}$$

5. The gain of the stage for mid-frequencies and very high frequencies up to  $F_2$  is now equal to  $g_m R_L$ , approximately, since the plate impedance is essentially equal to  $R_L$  up to the frequency  $F_2$  with the added compensation.

**Example.** An amplifier is desired to be flat to 4 Mc.  $C_t$  is 18  $\mu\text{f}$ . What are the sizes of  $R_L$  and  $L_o$  required? What is the gain of the stage when a tube with a  $g_m$  of 6,000  $\mu\text{mhos}$  is used?

1.  $F_2$  is 4 Mc, the top correction frequency.
2. The reactance of  $C_t$  at  $F_2 = X_{ct}$

$$\begin{aligned} &= \frac{1}{2\pi F_2 C_t} \\ &= \frac{1}{2\pi \times 4 \times 10^6 \times 18 \times 10^{-12}} \\ &= 2,200 \text{ ohms (approx)} \end{aligned}$$

3.  $R_L$  is 2,200 ohms, therefore, since it equals the reactance of  $C_t$  at  $F_2$ . The value for  $R_L$  is taken to the nearest 100 ohms.

4. 
$$\begin{aligned} L_o &= \frac{1}{2}C_t R_L^2 \\ &= \frac{1}{2} \times 18 \times 10^{-12} \times (2,200)^2 \\ &= 43.6 \times 10^{-6} \text{ henry (approx)} \\ &= 43.6 \mu\text{h} \end{aligned}$$
5. 
$$\begin{aligned} \text{Gain} &= g_m \times R_L \\ &= 0.006 \times 2,200 \\ &= 13.2 \end{aligned}$$

If the 2,200-ohm plate load resistor is used with the 43.6- $\mu\text{h}$  inductance in the shunt peaking circuit, the amplifier response will be essentially flat to 4 Mc with a gain of 13.

*Experimental Procedure for Determining  $L_o$ .* Further analysis of the equivalent plate circuit of the shunt-peaked amplifier will make it possible to find the correct value of  $L_o$  experimentally. If the reactance of  $L_o$  is one-half the reactance of  $C_t$  at the frequency  $F_2$ , as it should be for this type of compensation, the two reactances will be equal to each other at some frequency higher than  $F_2$  because  $X_L$  increases and  $X_c$  decreases with increasing frequency. The reactances will be equal at the resonant frequency  $f_r = 1.4F_2$ .

This is derived from the resonant frequency formula.

$$\begin{aligned} f_r &= \frac{1}{2\pi \sqrt{LC}} \\ &= \frac{1}{2\pi \sqrt{0.5C_i R_L^2 \times C_i}} \\ &= \frac{1}{2\pi \sqrt{0.5C_i^2 \times \left(\frac{1}{2\pi F_2 C_i}\right)^2}} \end{aligned}$$

since  $R_L = 1/2\pi F_2 C_i$ .

Squaring both sides of the equation,

$$\begin{aligned} f_r^2 &= \frac{1}{(2\pi)^2 \times 0.5C_i^2 \times \left(\frac{1}{2\pi F_2 C_i}\right)^2} \\ &= \frac{1}{0.5/F_2^2} = 2F_2^2 \\ f_r &= \sqrt{2}F_2 = 1.414F_2 \end{aligned}$$

Therefore, if the top correction frequency  $F_2$  is multiplied by 1.4 and the inductance of the peaking coil is adjusted to resonate with  $C_i$  at this higher frequency,  $f_r$ , the inductance will automatically be the correct size needed for the shunt peaking up to  $F_2$  because its reactance will then be one-half the reactance of  $C_i$  at the lower frequency  $F_2$ .

**12-6. Series Peaking.** The series peaking circuit has an advantage over shunt peaking because the series peaking coil, denoted  $L_e$ , separates  $C_{in}$  and  $C_{out}$ . This is illustrated in Fig. 12-8. Instead of adding in parallel as in the shunt peaking circuit,  $C_{in}$  and  $C_{out}$  are effectively in series with the peaking coil  $L_e$ . While it might seem that the series coil would reduce the signal input to the next grid because it offers increasing impedance for higher frequencies, such is not the case. For frequencies high enough to develop any voltage across  $L_e$  the coil resonates with the shunt capacitances to raise the plate impedance and gain of the stage, so that, even with the voltage-divider effect between  $L_e$  and  $C_{in}$ , the input signal voltage to the next grid is maintained uniform. With  $C_{in}$  and  $C_{out}$  in series with each other, instead of adding in parallel, the effective shunt capacity is reduced. This allows the use of a higher value of plate load resistor with a consequent over-all increase in gain, as illustrated by the response curve in Fig. 12-10 for series peaking.

In order to state definite values for the peaking coil and plate load resistor, the ratio between  $C_{in}$  and  $C_{out}$  must be known. Usually, the input capacitance is about twice the output capacitance because of the higher input capacitance of the tube, and the ratio  $C_{in}/C_{out} = 2$  is the

basis of the design values given here. Small changes in capacitance can be made to obtain the desired ratio by placement of parts. For instance, the coupling condenser  $C_c$  can be placed on either the grid or plate side of  $L_c$  to add its stray capacitance either to  $C_{in}$  or to  $C_{out}$ .  $C_{in}$  is the capacitance at the grid side of  $L_c$ ;  $C_{out}$  is the capacitance at the plate side of  $L_c$ .

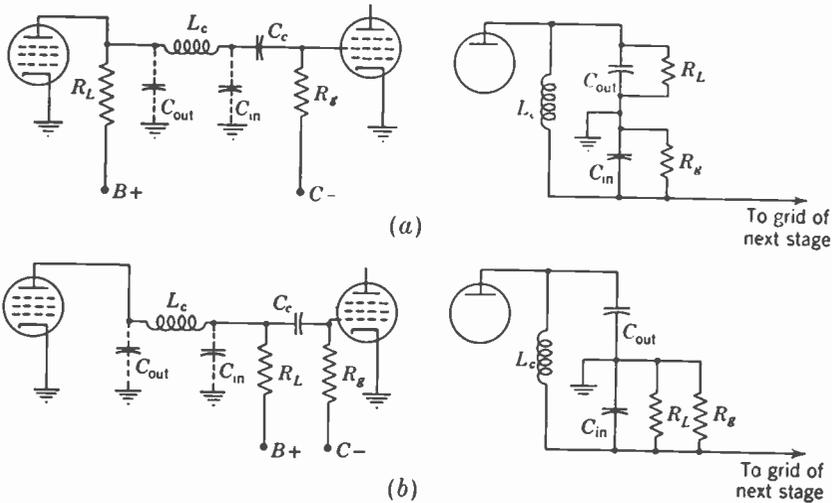


FIG. 12-8. Series peaking. (a) Series peaking circuit and equivalent plate load for  $C_{in} = 2C_{out}$ . (b) Series peaking circuit and equivalent plate load for  $C_{in} = \frac{1}{2}C_{out}$ .

For the ratio  $C_{in}/C_{out} = 2$ , the values required to make the amplifier flat to  $F_2$  for series peaking are

$$R_L = 1.5X_{c_1} \text{ at } F_2 = \frac{1.5}{2\pi F_2 C_1} \tag{12-7}$$

$$L_c = 0.67 C_1 R_L^2 \tag{12-8}$$

Since the value of plate load resistor is 50 per cent higher than in shunt peaking, the gain of the series-peaked amplifier is 50 per cent greater over the entire video-frequency range. The response is still maintained uniform up to  $F_2$ , as shown in Fig. 12-10. For the same example calculated for shunt peaking in Sec. 12-5, the plate load resistor would be 3,300 ohms,  $L_c$  120  $\mu$ h, and the gain of the stage 19.8, if series peaking were used.

Referring again to Fig. 12-8, the coil  $L_c$  and the two capacitances  $C_{in}$  and  $C_{out}$  form a  $\pi$ -type low-pass filter for coupling the signal. The filter can be turned end for end, as shown in *b*, without changing its characteristics, and this may be done if the output capacitance is larger than  $C_{in}$ . The rule is to keep the plate load resistor  $R_L$  on the low-capacitance side of the filter. In either case the ratio of input and output capacitances should be approximately 2 because this gives the best phase-angle response.

**12-7. Series-shunt Combination Peaking.** If the filter coupling of the series peaking coil is combined with a shunt peaking coil as shown in Fig. 12-9, the value of plate load resistor can be increased over either of the previous methods. The same rules apply as in the filter coupling. The

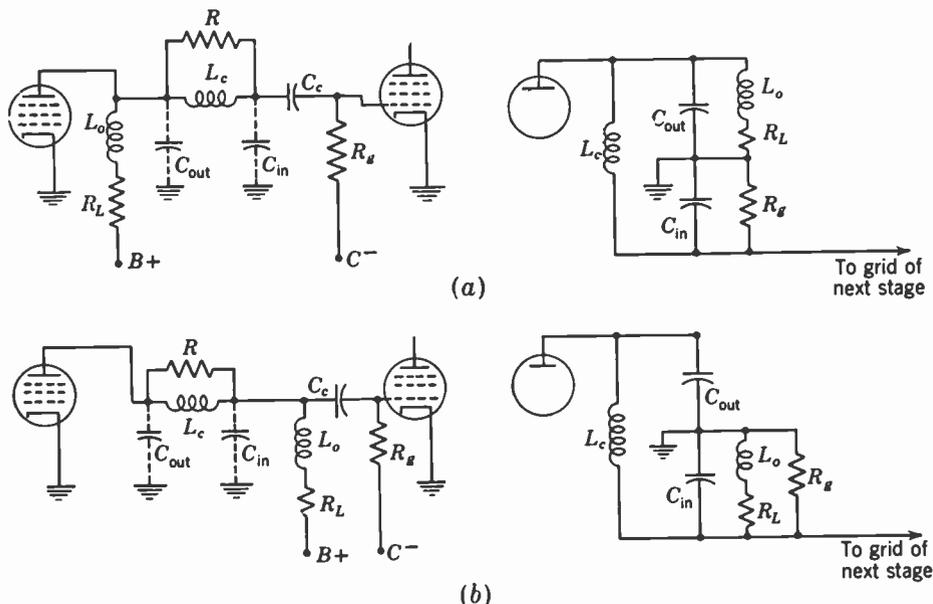


FIG. 12-9. Combination peaking, combining the shunt and series peaking methods. (a) Circuit and equivalent plate load for  $C_{in} = 2C_{out}$ . (b) Circuit and equivalent plate load for  $C_{in} = \frac{1}{2}C_{out}$ .

ratio between  $C_{in}$  and  $C_{out}$  should be approximately 2, and  $R_L$  is on the low-capacitance side of the series peaking coil  $L_c$ . The design values are

$$R_L = 1.8X_{c_i} \text{ at } F_2 = \frac{1.8}{2\pi F_2 C_i} \quad (12-9)$$

$$L_c = 0.52C_i R_L^2 \quad (12-10)$$

$$L_o = 0.12C_i R_L^2 \quad (12-11)$$

Incorrect division of the shunt capacitances in the circuit may cause an excessive rise in response, which can usually be reduced by adding the shunt damping resistor  $R$  across  $L_c$  to lower its  $Q$ . The value of  $R$  is about two to five times  $R_L$ . The coil  $L_o$  is damped by the series resistance of  $R_L$ .

With the higher value of  $R_L$ , the gain of the amplifier is 80 per cent higher than in shunt peaking, and the response is still uniform within 10 per cent up to  $F_2$ , as shown in Fig. 12-10. For the same example calculated for shunt peaking in Sec. 12-5, the plate load resistor would be

3,960 ohms,  $L_c$  145  $\mu$ h,  $L_o$  35  $\mu$ h, approximately, and the gain of the stage 23.8, if combination peaking were used.

TABLE 12-1. COMPARISON OF HIGH-FREQUENCY COMPENSATION METHODS

Type	$R_L$	$L_o$	$L_c$	Relative gain at $F_2$
Uncompensated.....	$1/2\pi F_2 C_t$	.....	.....	0.707
Shunt.....	$1/2\pi F_2 C_t$	$0.5C_t R_L^2$	.....	1.0
Series ( $C_{in}/C_{out} = 2$ ).....	$1.5/2\pi F_2 C_t$	.....	$0.67C_t R_L^2$	1.5
Combination ( $C_{in}/C_{out} = 2$ ).....	$1.8/2\pi F_2 C_t$	$0.12C_t R_L^2$	$0.52C_t R_L^2$	1.8

12-8. Summary of High-frequency Compensation. The essential design data for the shunt, series, and combination peaking methods are

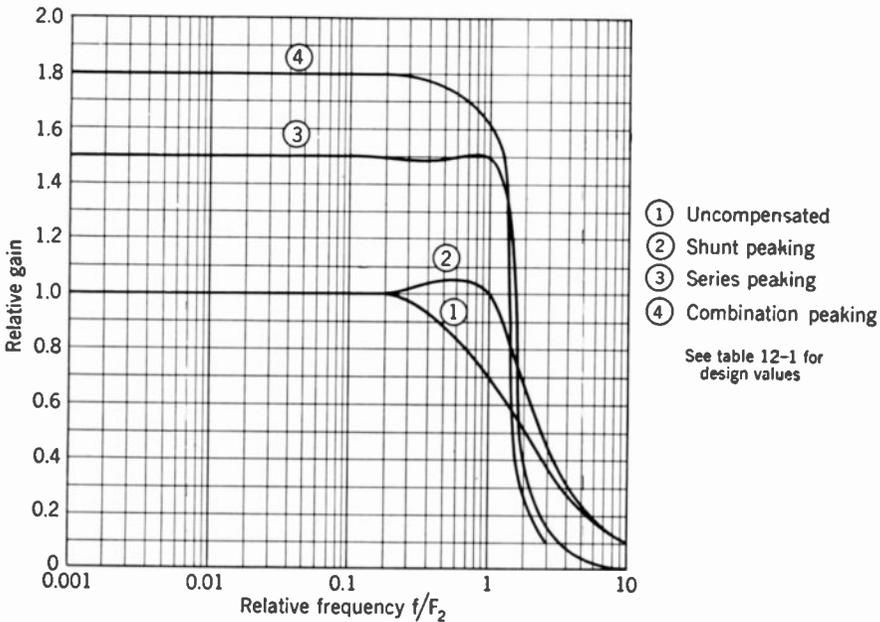


FIG. 12-10. Universal response curves for the video amplifier.

given in Table 12-1. Figure 12-10 shows the response curves. The uncompensated amplifier is also included for comparison. Combination peaking provides the most gain but the sharpest cutoff. The sharp cutoff at the high end of the frequency response makes the amplifier more susceptible to ringing, with the resultant outline distortion at edges between black-and-white picture information. However, combination peaking is used most often because of its high gain.

It should be noted that the values listed in Table 12-1 are not the only possibilities. More peaking can be obtained by using more inductance in the peaking coils, allowing a higher plate load resistance for more gain, but these advantages are at the expense of less uniform response, sharper cutoff, and more phase distortion. The design values given here provide suitable results in terms of gain, flat frequency response, and uniform time delay.

*Design Procedure.* Regardless of the top correction frequency used or the type of compensation, the design procedure is essentially the same as given previously for shunt peaking. The top correction frequency  $F_2$  is decided,  $C_i$  measured, and  $R_L$  given the correct value in terms of the reactance of  $C_i$  at the top correction frequency. With  $R_L$  and  $C_i$  known, the size of the peaking coils can then be determined. For the series and combination peaking the ratio of  $C_{in}$  to  $C_{out}$  can be made equal to 2 by proper placement of parts. This ratio can ordinarily be assumed when the tube's input capacitance is approximately twice the output capacitance.

With the compensation added, the gain-frequency characteristic of the amplifier can then be checked with a signal generator supplying constant voltage input over the required frequency range and a suitable a-c voltmeter monitoring the output. The response curve of the amplifier can be observed visually, if this is desired, in much the same way as an i-f response curve is obtained. A video sweep generator is used to supply video signal input to the amplifier the response of which is being checked, and the amplified output is coupled to the vertical input of the oscilloscope, preferably through a detector probe so that a standard oscilloscope can be used.

Every effort should be made to keep the stray and circuit capacitance down to a minimum, since the limiting factor in obtaining good high-frequency response is the total shunt capacitance. The common practice of placing a small paper or mica condenser in shunt with a large electrolytic bypass condenser in order to aid in bypassing the high frequencies can sometimes cause a nonuniform frequency response in the amplifier. The inductance of the electrolytic may resonate with the capacitance of the mica or paper condenser to cause an increase of gain in the region of the resonant frequency. Wire-wound resistors for the plate load should be avoided, unless their inductance and distributed capacitance are known and used as part of the compensating network.

*Cascaded Stages.* The amount of deviation from flat frequency response or uniform time delay that can be tolerated depends on the number of video stages of amplification used. For the one or two video stages generally used in the television receiver, compensation using the design values given will produce good results because the drop in response at the highest correction frequency is no more than 10 per cent, and the time-

delay distortion is negligible for the amount of phase distortion present. For a single stage using combination peaking, the variation in time delay from 1 ke to 4 Mc is  $0.00375 \mu\text{sec}$ .

When many cascaded stages of video amplification are used, as in studio and transmitter installations comprising 20 to 30 video amplifiers, the cumulative effects of any distortion introduced make the compensation problem much more exacting. In cascaded stages, where the plate circuit of one drives the grid of the next, the over-all frequency response is equal to the product of the individual responses because an amplifier multiplies the input signal by a factor equal to the gain. The over-all time-delay distortion is equal to the sum of the individual delays. Compensation of a multistage amplifier is more exacting, therefore, because

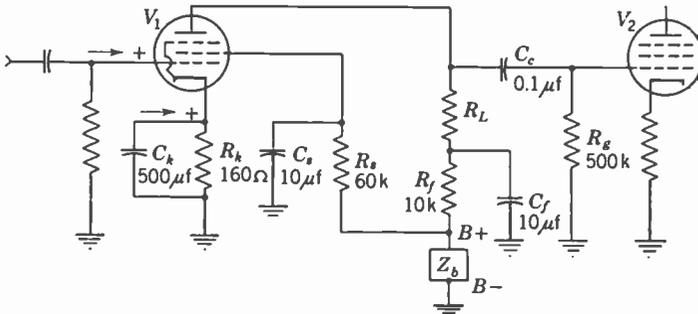


Fig. 12-11. Sources of low-frequency distortion in the conventional resistance-coupled amplifier are cathode bypass condenser  $C_k$ , screen bypass  $C_s$ , power-supply impedance  $Z_b$ , and coupling circuit  $R_g C_c$ .

secondary effects, not appreciable with one or two stages, become pronounced and may distort the signal. The problem can be solved by compensating each stage for a pass band of about 6 Mc and making up for the consequent loss in gain by using more stages.

**12-9. Low-frequency Response.** The video amplifier not only must have uniform response for the very high video frequencies but also must be capable of amplifying signal frequencies down to the frame frequency of 30 cps. There will be signal frequencies lower than 30 cps but these are reproduced by means of the d-c reinsertion circuit in the receiver.

At the low frequencies the shunt capacitance  $C_i$  has such high reactance that its shunting effect on the plate load resistor is negligible. Now, however, the increasing reactance of the coupling and bypass condensers at the low frequencies introduces distortion. Phase distortion is especially troublesome because of the large time delay at very low frequencies. Referring to Fig. 12-11, low-frequency distortion can be introduced in any of the following places in the circuit: the screen resistor and bypass net-

work  $R_s C_s$ , the power-supply impedance  $Z_b$ , the cathode resistor and bypass  $R_k C_k$ , and the coupling network  $R_o C_c$ .

*Distortion Caused by Screen-grid Impedance.* The required screen voltage, about 90 to 150 volts, is often obtained from the plate B supply by means of a series dropping resistor. This screen-grid resistor must be bypassed to the cathode or chassis ground by a condenser big enough to have little reactance at the lowest operating frequency, in order to maintain the screen at a steady d-c potential. If the screen voltage varies at the signal frequency, the gain of the stage will be reduced and phase distortion will be introduced in the amplifier.

Generally, sufficient bypassing is obtained when the resistance is at least 10 to 20 times the reactance of the bypass condenser at the lowest operating frequency. The screen dropping resistor can usually be properly bypassed with the common 8- or 10- $\mu$ f electrolytic condenser, as shown in Fig. 12-11, to eliminate the screen circuit as a source of low-frequency distortion. The value of screen dropping resistor is determined by the tube and is either obtained from the tube manual for typical operation or calculated for the desired screen voltage.

*Distortion Caused by Power-supply Impedance.* Unless an electronic type of regulated power supply is used, the impedance of the power supply  $Z_b$  is essentially equal to the reactance of the output condenser in the power-supply filter. This reactance varies with frequency and, if it is appreciable, may introduce frequency and phase distortion at low frequencies because of the corresponding variations in plate load impedance. Also, if the reactance of the output filter condenser is appreciable for the low frequencies, the video amplifier may oscillate at a low frequency of 10 to 30 cycles (motorboating) because of mutual coupling between stages through the power supply. All the amplifiers are connected to the common B supply, and unless its reactance is negligible some of the output signal voltage for one stage will be developed across the B supply and thereby be coupled into the plate circuit of the other video stages. The term *motorboating* is carried over from audio work, but the effect of the low-frequency relaxation oscillations in the video amplifier is to make the picture flicker on and off the screen at the oscillating frequency.

The problem of eliminating mutual coupling through the power supply at low frequencies can be solved by using a large output filter condenser, say 25 to 100  $\mu$ f or even larger if possible, and inserting an isolating or decoupling filter  $R_f C_f$  in series with the plate circuit as shown in Fig. 12-11. The decoupling network serves effectively as an additional filter on the power supply for the one stage alone, separating the amplifier plate load resistor from the output filter condenser of the power supply. For isolation purposes, the values for  $R_f$  and  $C_f$  are not critical, although  $R_f$  cannot be made too large because it acts as a series dropping resistor to

reduce the plate voltage applied to the tube, and  $C_f$  should be large enough to bypass  $R_f$  at the lowest operating frequency. By choosing suitable design values, however, the  $R_f C_f$  filter can be made to correct for low-frequency distortion introduced in other parts of the circuit.

Another method used to minimize mutual coupling through the power-supply impedance is an electronic voltage-regulated supply to make the power-supply impedance  $Z_b$  very low. Gas-filled voltage-regulator tubes such as the VR-150 can be placed across the power-supply output to keep the output voltage steady. Two VR-150 tubes in series across the output will hold the output voltage steady at 300 volts. Another form of regulator is shown in Fig. 12-12. The vacuum-tube regulator is so arranged that its plate current varies the  $IR$  drop across  $R_v$ . Grid bias is preset at a given value by means of the tap on the voltage divider, with the bucking battery inserted so that the grid bias can be negative with respect

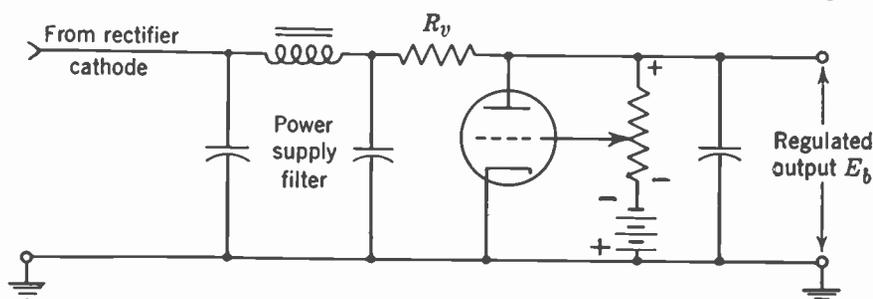


FIG. 12-12. A vacuum-tube voltage-regulator circuit for a stabilized power supply.

to cathode. As the output voltage  $E_b$  tends to change, the bias changes, providing an  $IR$  drop across  $R_v$  that opposes the change in  $E_b$  and keeps the output voltage constant. For instance, when  $E_b$  tends to increase, the bias becomes more positive, more plate current flows, and the increased  $IR_v$  drop reduces the output voltage to the steady value.

*Distortion Caused by Cathode Impedance.* Attenuation of the low frequencies can be produced by insufficient bypassing of the cathode bias resistor. If the resistance is not at least 10 to 20 times the reactance of the bypass condenser at the lowest frequency, the cathode voltage will vary at the signal frequency instead of remaining fixed as a d-c bias voltage. This results in degeneration, with reduced output, as some of the output signal is fed back to the grid circuit in phase opposite from the input to cancel out part of the input signal voltage. Referring to Fig. 12-11, it is shown that, if input signal drives the control grid in a positive direction, the cathode simultaneously will be driven more positive with respect to chassis ground because of the increased plate current, assuming that  $C_k$  is not large enough. Making the cathode more positive is the same as driving the control grid more negative, making the feedback to

the control grid a signal in opposite phase from the input to reduce the input and output signal voltage. The amount of degeneration varies with the signal frequency as the reactance of  $C_k$  varies, thus introducing low-frequency distortion.

The effects of low-frequency distortion caused by insufficient cathode bypassing can be corrected by five methods.

1. Eliminate the cathode bypass condenser, allowing the cathode resistor to degenerate the input signal the same amount for all frequencies. The gain of the stage is reduced because of the degeneration and is equal to  $A/(1 + g_m R_k)$ , where  $A$  is the normal gain without degeneration. For the case of a 160-ohm unbypassed cathode resistor with a tube having a  $g_m$  of 5,000  $\mu\text{mhos}$ , the gain of the stage will be 11.1 with degeneration if the normal gain is 20. The reduction in gain is constant for all frequencies, however, and no low-frequency distortion is produced.

2. Use a very large cathode bypass condenser, say 250 to 1,000  $\mu\text{f}$ . Such condensers are available at reasonable size and cost because they can be low-voltage electrolytics. The reactance of a 1,000- $\mu\text{f}$  condenser at 30 cycles is approximately 5 ohms and will not produce appreciable distortion in most circuits.

3. The plate decoupling filter  $R_f C_f$  can be given values that compensate for the low-frequency distortion introduced by the varying cathode impedance when the cathode bias resistor does not have a big enough bypass condenser. Design values for this are given in the next section.

4. Eliminate both the cathode bias resistor and bypass, ground the cathode, and apply fixed bias to the control grid from the power supply. This eliminates entirely the cathode bias impedance as a source of distortion, but a smaller value of grid resistance is necessary when using fixed bias rather than cathode bias. For the 6AC7, as an example, the d-c resistance in the grid circuit should not exceed 250,000 ohms when fixed bias is used, but may be as high as 0.5 megohm with cathode bias. A lower value of grid resistance aggravates the problem of the grid coupling circuit.

5. Use grid-leak bias for the video amplifier.

**12-10. Distortion Caused by the Coupling Circuit.** In Fig. 12-11, coupling condenser  $C_c$  has increasing reactance at lower frequencies, and for the extremely low video frequencies its reactance can be comparable to the resistance of the grid resistor  $R_g$ . The coupling condenser and grid resistor then act as a voltage divider for the amplified output of the preceding stage, with only the part of the signal voltage that is developed across  $R_g$  being impressed between grid and cathode for amplification. Any voltage developed across  $C_c$  is lost as far as output signal is concerned because the condenser voltage is not impressed between grid and cathode.

*Definition of  $F_1$ .* The extent to which the amplifier response is down at low frequencies because of the coupling network  $R_o C_c$  can be calculated for the case where the reactance of the coupling condenser is equal to the resistance of  $R_o$ , to provide a convenient measure of the coupling circuit's low-frequency response. For the high and middle frequencies, the reactance of  $C_c$  is negligible compared to  $R_o$ , and the entire signal voltage is applied to the grid circuit. As the frequency is decreased, the reactance of  $C_c$  increases until it is equal to  $R_o$  at some low frequency. At this frequency, the voltage across  $C_c$  equals the voltage across  $R_o$ , since they are equal arms in an effective voltage divider. The voltage across  $C_c$  is reactive, however. Adding these equal voltages vectorially, each is  $1/\sqrt{2}$  of the total applied voltage. Therefore at the frequency where  $X_{C_c} = R_o$ , denoted here as  $F_1$ , the response is down to  $1/\sqrt{2}$ , or 0.707 of the mid-frequency response. This is a voltage loss of 29.3 per cent, or 3 db.

Since the reactance of  $C_c$  is equal to  $R_o$  at the frequency  $F_1$ ,

$$X_{C_c} = R_o = \frac{1}{2\pi F_1 C_c} \quad \text{at } F_1$$

or

$$F_1 = \frac{1}{2\pi R_o C_c} \quad (12-12)$$

With a given value for  $R_o$  and  $C_c$ , the frequency at which the response is down 3 db can be found. As an example, for a 0.1- $\mu$ f coupling condenser and 0.5-megohm grid resistor,  $F_1$  is approximately 3 cycles. The amplifier is definitely not usable down to the frequency  $F_1$  because the response is down 3 db from the mid-frequency gain, as shown in Fig. 12-13. The frequency response is essentially flat down to a frequency ten times  $F_1$ .

*Phase Distortion Caused by  $R_o C_c$ .* More important than the loss in response is the phase distortion caused by the coupling circuit because it introduces a large time delay for the very low frequencies. Considering the plate output signal as an a-c generator driving the  $R_o C_c$  circuit, the current flowing in the coupling circuit leads the generator voltage by the angle with the tangent  $X_c/R$ . The signal voltage across  $R_o$  is of the same phase as the current, and leads the input voltage by the same angle. At the frequency  $F_1$ ,  $X_c$  and  $R_o$  are equal and their ratio is 1. The leading phase angle of the voltage across  $R_o$  at  $F_1$  is the angle with tangent 1, therefore, which is  $45^\circ$ . As illustrated in Fig. 12-13, the effect of the coupling circuit at the low video frequencies is to give the signal a leading phase angle that increases for lower frequencies, as the reactance of the coupling condenser  $C_c$  increases. In order to eliminate this source of

distortion it is necessary to make the phase angle zero at the lowest operating frequency.

*Values for  $R_g C_c$ .* To obtain good performance at low frequencies  $F_1$  should be a frequency as low as possible, which means that  $R_g$  and  $C_c$  should have values as large as possible. While it might seem that large enough values could be used to eliminate any distortion in the coupling circuit, there are practical difficulties.  $R_g$  cannot be made too large because of the harmful effect of reverse grid current caused by residual gas in the vacuum tube, which produces a positive grid bias. The maximum permissible value of  $R_g$  varies with different tubes and depends on whether fixed bias or self-bias is used. When the capacitance of  $C_c$  is too high, its physical size becomes too large and it adds too much shunt

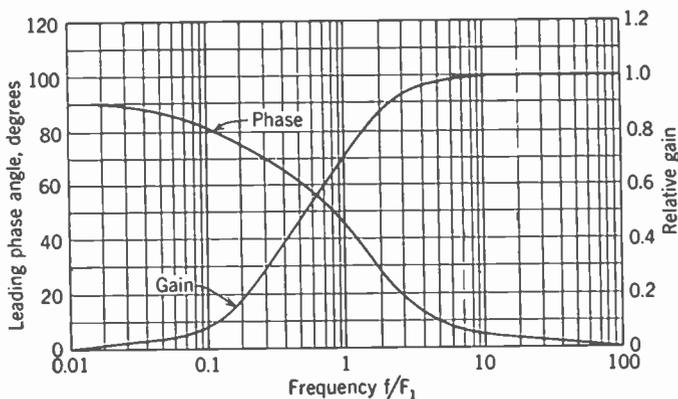


FIG. 12-13. Gain-frequency and phase-angle-frequency response of the uncompensated  $RC$  amplifier, as determined by the coupling condenser and grid resistor. At  $F_1$  the response is down 3 db and the leading phase angle is  $45^\circ$ . The leading phase angle at  $10 F_1$  is  $6^\circ$ .

capacitance to the circuit, which has an adverse effect on the high-frequency response. Also, the larger the capacitance of  $C_c$ , the greater its leakage current, which may cause trouble with a high value of grid resistance. Another disadvantage of using values of  $R_g$  and  $C_c$  that are too large is the tendency of the amplifier to motorboat. It is advisable, therefore, to use values of  $R_g C_c$  small enough to avoid these troubles and, if necessary, to use design values for the decoupling filter  $R_f C_f$  that compensate for the coupling circuit. Values used in the grid coupling circuit usually have an  $R_g C_c$  product that provides a time constant of 0.05 to 0.5 sec.

**12-11. Low-frequency Correction.** Of the four causes of low-frequency distortion, the screen-grid reactance can be made negligible, the power-supply impedance can be made very small and isolated from the amplifier by means of the plate decoupling network  $R_f C_f$ , the cathode reactance can

be eliminated or made very small and compensated by means of  $R_f C_f$ , and the coupling circuit  $RC$  product can be made as large as possible and compensated by means of  $R_f C_f$ . The compensating filter  $R_f C_f$ , then, can be very important in correcting the low-frequency response. In any case, it decouples the plate circuit from the power supply to reduce mutual coupling through the output filter condenser. In addition, this same decoupling network can be given values that allow it to compensate for distortion introduced in the cathode and the coupling circuit.

*Compensating for Cathode Impedance.* The plate current of the amplifier flows through both the cathode bias impedance and the  $R_f C_f$  impedance, so that the same kind of distortion is generated in both circuits.

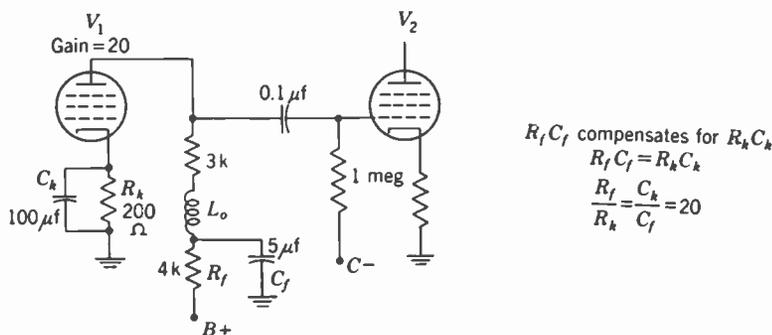


FIG. 12-14. Low-frequency compensating filter with values to correct for the cathode impedance.

However, the current flows through the networks in opposite directions, so that the  $R_f C_f$  distortion component can cancel out the distortion caused by  $C_k R_k$  when the values for  $R_f C_f$  make the two voltages equal. Perfect correction for the cathode bias impedance  $R_k C_k$  can be obtained by using the following values for  $R_f C_f$ :

$$R_f C_f = C_k R_k \tag{12-13}$$

and

$$\frac{R_f}{R_k} = \frac{C_k}{C_f} = \text{gain of the stage} \tag{12-14}$$

The calculations are easily made to find the required values for  $R_f$  and  $C_f$ .  $R_f$  is larger than  $R_k$  by the gain of the stage, and  $C_k$  is larger than  $C_f$  by the same proportion, as shown in Fig. 12-14. The value for  $R_k$  is determined by the required bias.  $C_k$  and  $C_f$  are given convenient values, say 25 to 250  $\mu\text{f}$  for  $C_k$  and 4 to 10  $\mu\text{f}$  for  $C_f$ , with  $C_k$  larger than  $C_f$  by the gain of the stage. The gain is equal to  $g_m \times R_L$ .

*Compensating for the Coupling Circuit.* The coupling circuit  $R_o C_c$  causes a loss in response and a leading phase angle that produces time-

delay distortion. The impedance of the  $R_f C_f$  network in the plate circuit,  $Z_f$ , increases the gain of the amplifier by the factor  $(R_L + Z_f)/R_L$  because the plate load impedance is increased by  $Z_f$ . Also the phase of the amplifier output is shifted in the lagging direction because  $C_f$  is a shunt capacitive reactance in the amplifier's plate load, drawing less leading plate current for the lower frequencies. Since the effect of  $Z_f$  is opposite the distortion effects produced by  $R_o C_c$ , the compensating filter  $R_f C_f$  can be given values to correct for the low-frequency distortion produced by the coupling circuit.

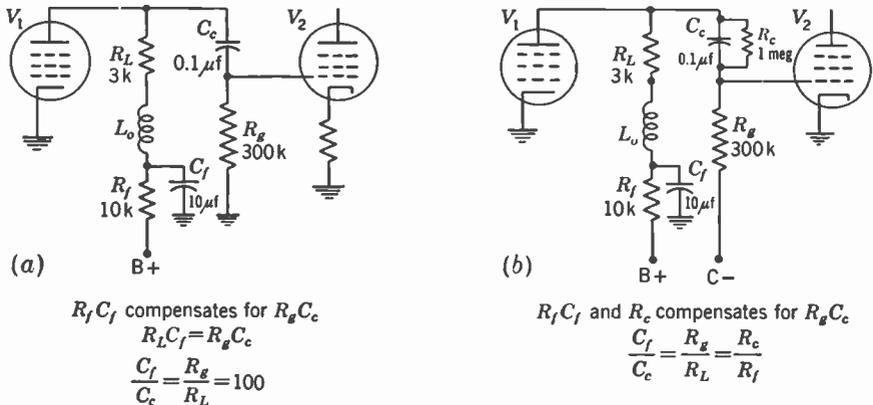


FIG. 12-15. Low-frequency compensation to correct for the reactance of the coupling condenser  $C_c$ . (a)  $R_f C_f$  filter. (b)  $R_f C_f$  and  $R_c$  across  $C_c$ .

As illustrated in Fig. 12-15a, approximate compensation for the low-frequency response of the coupling circuit can be obtained by using the following values:

$$R_L C_f = R_o C_c \quad \text{or} \quad \frac{C_f}{C_c} = \frac{R_o}{R_L} \quad (12-15)$$

Notice that  $R_L$  is the resistance of the plate load resistor, which is determined by the high-frequency response. Suitable values are 10  $\mu\text{f}$  for  $C_f$ , 0.1  $\mu\text{f}$  for  $C_c$ , and 10,000 ohms for  $R_f$ . Therefore, the required value of  $R_g$  is 300,000 ohms. The resistance of  $R_f$  must be much more than  $R_L$  for the low-frequency compensation to be effective. Also, the larger the resistance of  $R_f$  the lower the frequency that can be corrected.  $R_f$  cannot be made too large, however, because its voltage drop reduces the tube's plate voltage. With 10,000 ohms for  $R_f$  and 10-ma plate current, the voltage drop is 100 volts, which would leave 200 volts available as the available plate supply voltage with a B supply output of 300 volts.

The low-frequency compensation can be improved by adding the resistor  $R_c$  across the coupling condenser  $C_c$ , as in Fig. 12-15b. Now the coupling circuit has a d-c connection to the next stage and the low-fre-

quency response can be made perfect down to 0 cps, or direct current. The values required for this type of low-frequency compensation are

$$\frac{C_f}{C_c} = \frac{R_o}{R_L} = \frac{R_c}{R_f} \tag{12-16}$$

Using the same values as in Fig. 12-15a,  $C_f$  is 100 times as large as  $C_c$  and  $R_o$  is 100 times  $R_L$ . Therefore,  $R_c$  is 100 times  $R_f$ , or 1 megohm. The addition of  $R_c$  provides a d-c path that impresses a small positive voltage on the grid of the next stage, but this can be canceled by increasing the negative voltage applied to the grid for bias. When the previous stage is directly coupled to  $V_1$ , it can be considered as a d-c amplifier, since the d-c component of the signal will be amplified and coupled to the next stage.

**12-12. Square-wave Analysis.**

The low-frequency compensation can be determined experimentally by square-wave analysis of the amplifier's low-frequency response. A perfectly square wave is the sum of a sine wave at the fundamental frequency together with all the odd harmonics of the fundamental. The amplitude of each of the odd harmonics is inversely proportional to its harmonic number, and all the harmonics are in phase with the fundamental, as shown in Fig. 12-16a. A circuit that has flat frequency response and constant time delay for the fundamental and harmonic frequencies

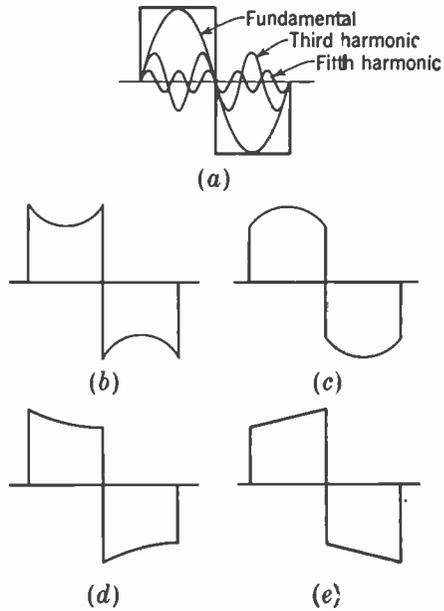


FIG. 12-16. Square-wave analysis of the video amplifier's low-frequency response. (a) Applied square wave. (b) Attenuation of low frequencies: no phase distortion. (c) Rise in gain for low frequencies: no phase distortion. (d) Phase distortion caused by excessive leading angle for low frequencies. Attenuation of low frequencies also present.  $C_f R_o$  too small. (e) Phase distortion caused by excessive lagging angle for low frequencies. No attenuation of low frequencies.  $C_f R_L$  less than  $C_f R_o$ .

transmits the square wave without any change in shape. A nonuniform frequency response or time delay must result in a change of wave shape. For a square wave of low frequency, uniform time delay is especially critical in maintaining the wave shape.

A low-frequency test of the video amplifier can be made by applying a low-frequency square-wave signal to the input, say 30 cps, and observing

the resultant output wave on an oscilloscope. The input square-wave signal should be coupled directly to the oscilloscope first to check the original wave shape and the response of the oscilloscope. Then the shape of the output wave can be interpreted in terms of the video amplifier characteristics at low frequencies. As illustrated in Fig. 12-16, curvature of the horizontal portion of the square wave is caused by a nonuniform frequency response. Note that the wave in *b* is bowed inward, indicating insufficient gain for the fundamental, which is the lowest frequency. The wave at *c* is bowed outward, showing that there is more gain for the fundamental frequency than for the higher frequency harmonics. This can be the result of a value of  $R_f C_f$  that provides too much compensation. The symmetry of a half cycle of the wave is dependent on the phase angle response of the amplifier. Lack of symmetry indicates phase distortion, with frequency distortion also present. Perfect symmetry indicates that there is no phase distortion. It is important to note that no indication of circuit performance below the fundamental frequency of the square wave is obtained. Therefore, the square-wave signal should be the lowest correction frequency desired.

The procedure for correcting the video amplifier's low-frequency response can be as follows:

1. Choose a suitable value for  $R_f$ . Larger values of  $R_f$  allow correction down to lower frequencies but its resistance cannot be too great because the  $IR$  drop reduces the plate voltage of the video amplifier, reducing the  $g_m$  and the gain of the stage for all frequencies. A value of about 10,000 ohms is practicable.
2. Use a large condenser of convenient size for  $C_f$ . This can be 4 to 10  $\mu\text{f}$ .
3. Use a coupling condenser  $C_c$  of 0.05 to 0.1  $\mu\text{f}$ .
4. Make the grid resistor  $R_g$  a variable resistance. The maximum resistance of this variable rheostat can be 0.5 to 1.0 megohm.
5. Couple the 30-cps square-wave signal to the input of the video amplifier and observe the wave shape of the amplifier's output by means of an oscilloscope. Vary the grid resistance  $R_g$  until a perfect square wave is obtained in the output. This determines the value of  $R_g C_c$  that allows the compensating filter  $R_f C_f$  to provide the best correction. Then measure the experimentally determined value of  $R_g$  and insert a fixed grid resistance of this size.

The above method provides the best over-all low-frequency compensation because maintaining the wave shape of the square-wave signal is critically dependent on minimum phase distortion. This is the condition desired in correcting the video amplifier response because minimum phase distortion at the low video frequencies is more important in picture reproduction than frequency response. The square-wave experimental proce-

ture can be used to correct distortion produced by the cathode impedance of the amplifier, or the coupling circuit, or both. With a single measurement, the optimum values are obtained for over-all low-frequency correction of a single amplifier stage or several stages. A typical square-wave signal generator is shown in Fig. 12-17.



FIG. 12-17. Square-wave signal generator. Frequency range is 5 to 100,000 cps. (Measurements Corporation.)

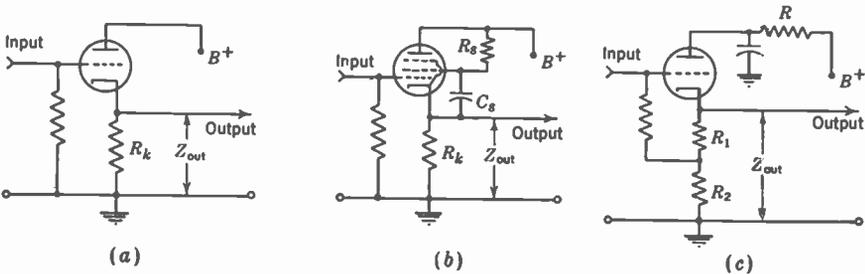


FIG. 12-18. The cathode-coupled stage. The load is in the cathode instead of the plate circuit. (a) Triode cathode follower. (b) Pentode cathode follower. (c) Triode cathode follower, with plate-dropping resistor  $R$  and grid returned to tap on cathode resistance.

**12-13. Cathode-coupled Stage.** *Cathode follower* or *cathode-coupled stage* is the name given to a stage where the load is in the cathode instead of the plate circuit, as shown in Fig. 12-18. The input is coupled to the grid circuit as usual, but the output is taken across the cathode resistor  $R_k$ , which is not bypassed in order to provide the output signal voltage. Triodes, tetrodes, or pentodes can be used. A series resistor may be used in order to drop the plate voltage from the B supply, but the resistor is bypassed so that it does not serve as the plate load for signal voltage.

An important advantage of the cathode-coupled stage is that it can operate as an impedance-matching device for coupling a high-impedance source to a low-impedance load, serving the same purpose as an impedance-matching transformer but without its limitations on frequency response. In video work, the cathode follower is used at the transmitter to couple the video signal with its extremely wide band of frequencies from the high-impedance plate circuit of a video amplifier to a transmission-line cable of much lower impedance, usually about 72 ohms. It should be noted that in the receiver the video output stage normally drives the high-impedance kinescope grid circuit directly, and no cathode follower is needed.

The effectiveness of the cathode-coupled stage for impedance matching derives from the high input and low output impedance of the stage. Both the input conductance and the grid-to-cathode capacitance are reduced by the factor  $1/(1 + g_m R_k)$ . This high input impedance can be useful in applications where it is desired to provide isolation. For instance, frequency-response checks can be made on a video amplifier by coupling the video output to a cathode follower and measuring across its cathode resistor, thus isolating the meter from the video amplifier. The lowered value of input capacitance in the cathode follower makes it possible for the preceding amplifier to provide more gain or wider frequency response because of the lower value of shunt capacitance. The cathode follower is also used very often for isolating stages in mixing amplifiers, making it possible to mix several signals in an amplifier without common coupling between circuits.

The value of output impedance is given by

$$Z_o = \frac{\frac{1}{g_m} \times R_k}{\frac{1}{g_m} + R_k} \quad (12-17)$$

This is equivalent to the reciprocal of the transconductance being in parallel with the cathode resistance. The gain of the stage is given by

$$\text{Gain} = \frac{g_m R_k}{1 + g_m R_k} \quad (12-18)$$

The gain of the stage is always less than one, as shown by the expression above, because all the output signal provides inverse feedback to the control grid to degenerate the input. However, the higher the value of  $R_k$  the closer to unity the gain of the stage becomes, and output impedances as low as 50 to 100 ohms can be obtained with a gain close to 1. When it is desired to find the value of  $R_k$  needed for a definite output impedance  $Z_o$ , Eq. (12-17) can be rearranged as follows:

$$R_k = \frac{\frac{1}{g_m} \times Z_o}{\frac{1}{g_m} - Z_o} \quad (12-19)$$

There is no phase inversion in the cathode-coupled stage, since the cathode voltage follows the grid input voltage, varying in the same direction. When the grid is driven more negative by signal voltage the decreased plate current makes the cathode less positive, and when the grid is made less negative the increased plate current makes the cathode more positive. While there is no cathode bypass condenser, the average value of cathode current produces an  $IR$  drop across  $R_k$  which can still serve effectively as the bias because it sets the operating point about which the signal variations take place. If the value of  $R_k$  needed to provide the desired amount of output impedance is too large for the required amount of bias,  $R_k$  can be divided into two parts, as in *c* of Fig. 12-18, where only the voltage across  $R_1$  serves as grid bias because of the grid return connection. When the required value of  $R_k$  is too small for the desired value of bias, an additional resistor can be added in series and bypassed for the lowest signal frequency. The output signal is then taken from the junction of the two resistors.

**12-14. Video Amplifier Tubes.** The merit of a tube as a video amplifier depends essentially on its transconductance and interelectrode capacitances. The  $g_m$  of the tube should be as high as possible in order to provide as much gain as possible with the low value of plate load that is required. The interelectrode capacitances should be as low as possible, in order to allow a higher value of plate load for a given frequency response. A figure of merit for a tube in wide-band amplifier service, then, is the ratio of  $g_m$  to the sum of the interelectrode capacitances.

$$\text{Figure of merit} = \frac{g_m}{C_{out} + C_{in} + C_{op} (1 + A)}$$

Triodes generally have a lower figure of merit than pentodes. However, triodes have a lower noise rating than the equivalent multigrid tubes and are preferred in low signal-level applications such as the first video amplifier in the camera chain, in order to maintain a good signal-to-noise ratio. In addition, the miniature triode glass tubes have small tube capacitances and can provide suitable figures of merit compared with the larger glass or metal tubes. In video amplifiers for relatively high voltage output, such as the video output stage in the receiver and the modulator section in the transmitter, it may be necessary to sacrifice figure of merit in order to use a tube capable of handling the amount of signal swing required. The beam-power tubes such as 6V6 and 6L6 are often used in

these applications. With these power tubes and triodes, the internal plate resistance of the tube may be low enough to alter the simplifications made for gain of the stage. In such a case, the gain of the stage is given by Eq. (12-1) because  $R_L$  is not negligible compared to  $R_p$ .

#### REVIEW QUESTIONS

1. Why is the response of the video amplifier down for the very high frequencies?
2. What are the causes of low-frequency distortion in the video amplifier?
3. A 3,300-ohm plate load resistor is used with the 6AC7, having a  $g_m$  of 8,000  $\mu\text{mhos}$ . What is the gain of the stage at 10 kc?
4. The total shunt capacitance  $C_t$  resonates with an inductance of 100  $\mu\text{h}$  at 3 Mc. What is the value of  $C_t$ ?
5. An uncompensated amplifier has a 2,000-ohm plate load resistor and total shunt capacitance of 40  $\mu\text{mf}$ . At what frequency is the gain down 3 db at the high end?
6. The gain of an uncompensated amplifier is down 3 db at 3 Mc, using a 3,000-ohm plate load. How much is the total shunt capacitance  $C_t$ ?
7. What two factors determine the amount of dynamic input capacitance added by the Miller effect?
8. It is desired to shunt peak an amplifier for flat response up to 3.5 Mc.  $C_t$  is 20  $\mu\text{mf}$ . A 6AC7 is used with a  $g_m$  of 9,000. What is the required value for  $R_L$  and  $L_o$ ? What is the gain of the stage?
9. Using series peaking for the amplifier of Question 8, what are the required values for  $R_L$  and  $L_c$  and what is the gain of the stage?
10. Using combination peaking for the amplifier of Question 8, what are the required values for  $R_L$ ,  $L_c$ , and  $L_o$  and what is the gain of the stage?
11. If the top correction frequency is 3.5 Mc, at what frequency will the inductive reactance of the peaking coil required for shunt peaking equal the capacitive reactance of the total shunt capacitance?
12. Calculate the resonant frequencies in terms of  $F_2$  at which the peaking coils  $L_c$  and  $L_o$  used in combination peaking will resonate with the total shunt capacity  $C_t$ . Do the same for the coil  $L_c$  in series peaking.
13. What values should the low-frequency compensating filter  $R_f C_f$  have in order to correct for the cathode impedance consisting of a 200-ohm cathode bias resistor and 100- $\mu\text{f}$  bypass condenser? The gain of the stage is 15.
14. The grid resistor  $R_g$  is 250,000 ohms and  $C_c$  is 0.05  $\mu\text{f}$ . At what frequency will the low-frequency response be down 3 db because of the coupling circuit?
15. Describe briefly how the method of square-wave analysis can be used to check low-frequency phase distortion in the video amplifier.
16. Name two tubes suitable as video amplifiers, giving their advantages.
17. What are the advantages and disadvantages of the cathode-coupled stage? Why is its gain always less than 1?
18. Using a 6AC7 tube with a  $g_m$  of 9,000  $\mu\text{mhos}$  in a cathode-coupled stage, what value of  $R_k$  is needed to provide an output impedance of 72 ohms?
19. In a resistance-coupled audio amplifier with a 500,000-ohm plate load resistor and total shunt capacitance of 30  $\mu\text{mf}$ , at what frequency is the gain down 3 db at the high end?
20. Draw the schematic diagram of a video amplifier using series-shunt combination peaking and low-frequency compensation. Calculate the values for  $R_L$ ,  $L_o$ , and  $L_c$  for flat response up to 4 Mc.  $C_t$  is 20  $\mu\text{mf}$ . Also calculate the values for the resistor  $R_c$  in parallel with  $C_c$  and  $R_f C_f$  compensating for the grid coupling circuit. Let  $R_f$  be 10,000 ohms and  $C_c$  0.05  $\mu\text{f}$ . Calculate the gain, with a  $g_m$  of 7,000  $\mu\text{mhos}$ .

## CHAPTER 13

### BRIGHTNESS CONTROL AND D-C REINSERTION

The pedestal or blanking level of the composite video signal corresponds to black. This black level is transmitted at a constant 75 per cent of the peak carrier amplitude so that the pedestal level can be a definite brightness reference, independent of the signal variations of light and shade that form the actual picture. It is necessary that this be done so that there can be a definite color reference in terms of voltage. In the receiver, the black level corresponds to the grid cutoff voltage for the picture tube because then there is no beam current and no illumination of the screen. Therefore, if the receiver is to reproduce the image with the same brightness values as the transmitted picture, the pedestal voltage of the composite video signal must drive the kinescope grid voltage to cutoff. Thus, black at the receiver corresponds to black at the transmitter, and the brightness values of the reproduced picture are the same as the transmitted picture because the variations in light and shade are from a common black level.

**13-1. Brightness Control.** The d-c bias on the kinescope grid determines the brightness of the reproduced picture because the bias sets the operating point about which the a-c video signal swings the instantaneous grid voltage, as illustrated in Fig. 13-1. The average value of the a-c video signal is zero, making the average grid voltage equal to the bias voltage. Therefore, the average beam current or kinescope illumination is the amount corresponding to the bias. While the a-c video signal contains the detailed picture information in terms of the difference in brightness of each picture element from the average, the bias determines the background information for the picture in terms of the average brightness.

The average brightness of the reproduced picture is set by adjusting the d-c bias voltage on the kinescope grid. This is the function of the manual control usually called the *brightness* or *background* control. It is generally located on the receiver's front panel or the rear apron of the chassis. In Fig. 13-2 the potentiometer *R* in both *a* and *b* is the manual brightness control. The adjustable bias voltage may be in either the control grid or cathode circuit but in both cases the bias voltage makes

the kinescope control grid negative with respect to the cathode. In *a* the negative bias voltage is connected in series with the kinescope grid load resistor. By varying the amount of negative voltage tapped off for the kinescope bias, the control *R* varies the brightness. In *b*, a positive d-c voltage is inserted in the cathode circuit of the kinescope as cathode bias.

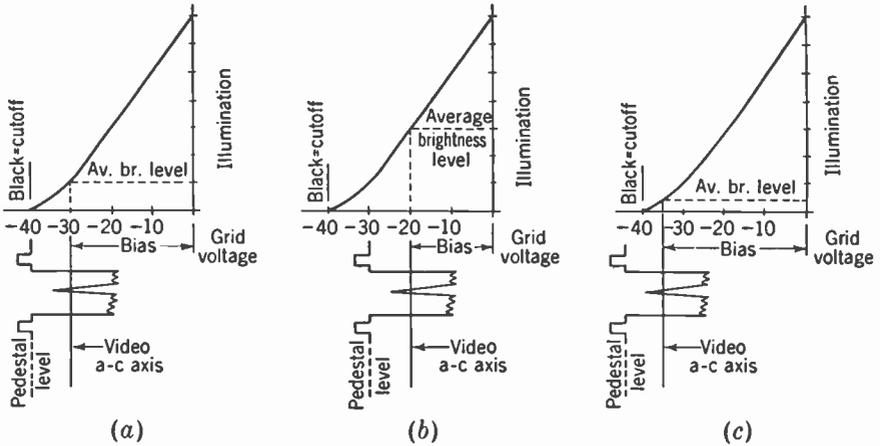


FIG. 13-1. The d-c bias on the kinescope grid determines the average brightness or background of the reproduced picture. (a) Correct bias. (b) Too little negative bias. (c) Too much negative bias.

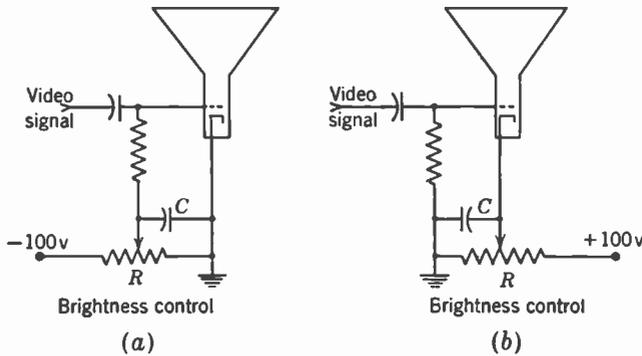


FIG. 13-2. Manual brightness-control circuits. (a) In the control-grid circuit of the kinescope. (b) In the cathode circuit of the kinescope.

Moving the variable arm of the control *R* toward the ground terminal makes the cathode less positive with respect to the control grid, which returns to ground, resulting in less negative grid bias on the kinescope and increasing the brightness. The bypass condenser *C* prevents the bias voltage from varying at the video signal frequencies and provides a signal return path around the brightness control between the cathode and grid load resistor of the kinescope.

Figure 13-3 shows a picture with the brightness control set too high. Because of insufficient negative bias, the black parts of the video signal do not come close enough to cutoff voltage on the kinescope grid and they appear too light. As a result, the picture is too bright and appears thin, without good black values. When the bias is small enough to allow the maximum white parts of the video signal to drive the kinescope grid voltage positive, the picture becomes badly defocused, with blooming. If the kinescope grid bias is too negative, some of the camera signal will be more negative than the kinescope grid cutoff voltage. These dark parts of the signal cannot be reproduced, and shadow detail in the picture is lost as the dark-gray values are cut off. The correct bias, therefore, is the

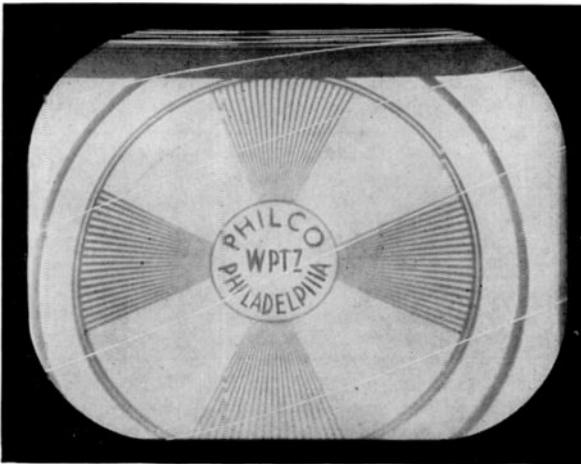


FIG. 13-3. Brightness too high. Picture appears thin with retrace lines and blanking bars visible. Retracers are not visible in receiver with internal vertical blanking circuit. (Philco Corporation.)

amount that allows the pedestal voltage of the composite video signal to drive the kinescope grid voltage to cutoff.

If the pedestal level varies for frames having different brightness levels, it will not be possible for the fixed voltage of the manual brightness control to set the bias properly. Consequently, an additional automatic biasing arrangement is necessary to shift the kinescope grid bias as needed so that the pedestal voltage of the video signal always drives the picture tube grid voltage to cutoff. This automatic brightness control is the function of the d-c component of the video signal, which makes the pedestals of the video signal line up at cutoff voltage in the kinescope grid circuit, for all frames.

**13-2. D-C Component of the Video Signal.** The composite video signal, transmitted as an amplitude variation of the picture carrier, has all the pedestals lined up at a fixed level because the signal includes the d-c

component required to bring the blanking or pedestal level to a constant 75 per cent of the peak carrier amplitude. This alignment of all the pedestals at the common black level is maintained through the r-f and i-f amplifiers in the receiver because it is only the picture carrier signal that is amplified in these stages. There is no video voltage until the modulated picture carrier is rectified in the video detector stage. Once the video signal voltage is recovered in the video detector, the d-c component inserted at the transmitter can be lost by any capacitive or transformer coupling, since with such coupling the d-c component is blocked and only the a-c variations are coupled to the next stage.

The output of the video detector contains the correct d-c component of the video signal with the pedestals lined up. If this signal were coupled directly to the kinescope grid, the manual brightness control could be adjusted to give the proper bias with all the pedestals coming at cutoff,

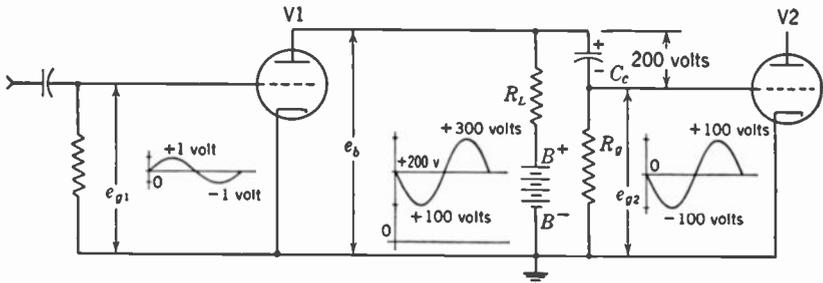


FIG. 13-4. Capacitive coupling. For simplicity, a sine-wave signal is illustrated instead of the video signal. Grid bias is omitted.

and no d-c reinsertion would be necessary. However, the output voltage of the video detector is normally not enough to drive the kinescope grid circuit and provide the desired contrast, so that one or more stages of video amplification are used. Since capacitive coupling is often employed in the video amplifier, this type of coupling will be examined in detail to see how the d-c component of the video signal is lost, causing the pedestal level to vary with the picture information.

**13-3. Capacitive Coupling.** The effect of capacitive coupling between stages is to block the steady d-c component of any plate signal voltage and allow only the variations to be developed as a-c signal in the input circuit of the next stage. How this is done is illustrated in Fig. 13-4, which shows a resistance-coupled stage amplifying a sine-wave signal. Referring to the figure, there are three parallel branches having the same applied voltage  $e_b$ . These are the internal plate-to-cathode circuit of V1, the plate load resistor  $R_L$  in series with the B supply, and the coupling circuit  $R_g C_c$ . The plate-to-cathode voltage of V1 is equal to the applied B voltage minus the  $IR$  drop across the plate load  $R_L$ . As the plate current in V1 varies

with the applied signal, the varying  $IR$  drop across  $R_L$  causes the plate-to-cathode voltage to vary.  $C_c$  and  $R_o$  are in series with each other, and this series combination is connected between plate and cathode of  $V_1$ . Effectively, then,  $V_1$  is an a-c generator with its varying plate-to-cathode voltage  $e_b$  impressed across  $R_o$  and  $C_c$ . As this voltage increases and decreases,  $C_c$  charges and discharges through  $R_o$ .

When  $e_b$  increases above its average value,  $C_c$  takes on an additional charge from the B supply. The electron flow is from B-, up through  $R_o$  to the grid side of  $C_c$ , repelling electrons from the other plate of  $C_c$ , through  $R_L$  and back to B+. This produces an  $IR$  drop across  $R_o$ , between grid and cathode of  $V_2$ , with the grid side positive. Therefore, while the plate-to-cathode voltage of  $V_1$  is increasing, a varying signal is applied between grid and cathode of  $V_2$  in a positive direction. Bias is omitted from the illustration, but in a class A<sub>1</sub> amplifier, tube 2 will have enough negative bias to keep the grid negative with respect to the cathode during the positive half of the signal swing.

When the plate voltage of  $V_1$  decreases,  $C_c$  loses some charge because the impressed voltage has decreased. The condenser neutralizes its charge by transferring the excess electrons on one plate to the other. The electron flow is from the grid side of  $C_c$  down through  $R_o$ , through the internal cathode-to-plate resistance of  $V_1$ , back to the other plate of  $C_c$ . This results in an  $IR$  drop across  $R_o$  with the grid side negative. Therefore, while the plate-to-cathode voltage of  $V_1$  is decreasing, a varying signal in a negative direction is applied between grid and cathode of  $V_2$ .

The values of  $R_o$  and  $C_c$  are always made large enough in a coupling circuit so that  $C_c$  cannot charge or discharge appreciably during the time the applied voltage  $e_b$  is increasing or decreasing. This is done in order to have any signal variations of the applied voltage developed across  $R_o$ , which is connected between grid and cathode of the next stage, rather than across  $C_c$ . As a result, the voltage across the coupling condenser  $C_c$  becomes equal to the average value of the applied voltage, since it cannot follow the variations. With the condenser voltage equal to the average value, any time that the applied voltage is at this value the coupling condenser neither charges nor discharges; there is no current through  $R_o$ , and the signal-voltage input to the next stage is zero. The zero level of voltage across  $R_o$ , then, corresponds to the applied voltage's average value, which is the voltage developed across  $C_c$ . This is blocked from the next stage because  $C_c$  is not connected between grid and cathode. Variations from the average value in the applied voltage cause  $C_c$  to charge or discharge, developing signal voltage across  $R_o$  of positive and negative polarity. The result, therefore, of coupling the fluctuating d-c plate voltage signal by means of the  $R_oC_c$  circuit is to couple to the next stage an a-c signal with variations of positive and negative polarity about a zero-volt-

age axis. This zero axis corresponds to the average value of the d-c signal, which is developed across the coupling condenser and blocked from the next stage.

**13-4. Average Value of the Video Signal.** The average value of any signal is the arithmetical mean of all the values taken over a complete cycle. For a fluctuating d-c signal the average value is some d-c level. An a-c signal has as much positive variation as negative, and the sum of all the positive and negative values is zero, making the average value zero when taken over a complete cycle. For a symmetrical sine wave the zero axis of the a-c signal is exactly in the center, as shown in Fig. 13-5*a*. For an unsymmetrical signal wave shape such as the video signal the average-value level must still be the zero axis of the a-c signal, since if the average value were not zero there would be more signal of one polarity than the

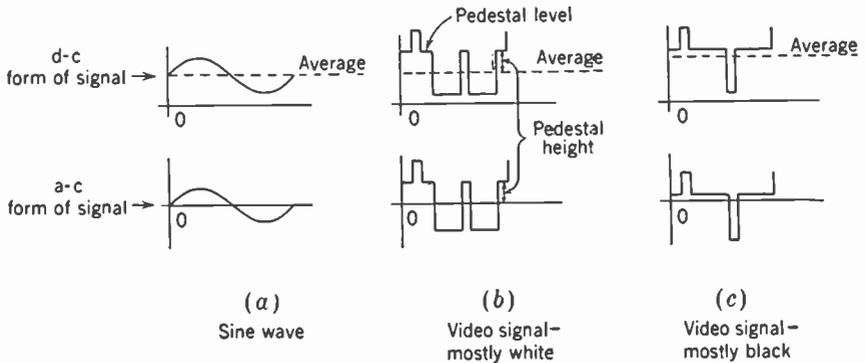


FIG. 13-5. Average value of a fluctuating signal.

other and there would be a d-c component. However, in the unsymmetrical video signal the zero axis is not in the center. The a-c axis must be a line that divides the signal so that there is just as much positive signal as negative and will be the line that makes the area above it equal to the area below. If the two areas are equal it means that, taken over a complete cycle, there will be equal amounts of positive and negative signal, even though there may be a greater instantaneous voltage in one direction.

Figure 13-5 illustrates the average value for a symmetrical sine wave and the unsymmetrical video signal corresponding to two horizontal lines. In *b* is shown the video signal resulting from one horizontal line in scanning a white frame with a vertical black line down the center. If a white line against a black frame is scanned, the video signal for one horizontal line appears as in *c*. In any case, the average-value axis divides the signal into equal areas above and below the axis and is zero. In this respect the average value is the same for all video signals, regardless of picture content. However, for video signals containing different picture infor-

mation the average-value axis is at different distances from the pedestal or blanking level, which is the color-reference voltage corresponding to black.

Putting the video signals on a common axis, the pedestals are out of line, as illustrated in Fig. 13-6. In the grid circuit of the picture tube, the zero axis of the a-c video signal corresponds to the kinescope d-c bias voltage, and the a-c signal variations swing the instantaneous grid voltage above and below the bias voltage. If the pedestals correspond to different voltage variations from the axis, it will be impossible for one setting of the manual brightness control to provide the correct bias voltage that allows the different pedestal voltages to drive the kinescope grid voltage to cutoff. When the video signal corresponds to a relatively bright scene the bias shown in Fig. 13-6 is correct. The pedestals drive the kinescope grid voltage to cutoff, making the pedestal level correspond to black. As a result, the camera signal variations produce the correct color values in screen illumination, and blanking is normal. With a dark scene, the average camera signal is closer to the pedestal level. Then the fixed bias shown in Fig. 13-6 is wrong because the pedestal voltage is not at cutoff. If the blanking or pedestal level is not reproduced as black, the color values will be incorrect and retrace lines may be visible in the picture. It should be noted that the video signals shown are not to be considered as the signals for two consecutive horizontal lines, but rather as typical lines from two different frames that do not have the same brightness.

The distance between the pedestal level and the zero axis of the a-c video signal or average-value axis of the d-c signal is the pedestal height. This is a convenient measure of the extent to which the pedestals are out of line because it measures the voltage swing of the pedestal from the zero axis in the a-c signal. In order to line up the pedestals for proper kinescope blanking and the correct brightness, it is necessary to reinsert a d-c component that is proportional to the pedestal height. If this d-c voltage is reinserted to line up the pedestals and no capacitive coupling is used after the d-c reinsertion, then the signal will appear on the kinescope grid with all the pedestals at a common level. The manual brightness control can then be adjusted to set the kinescope bias so that the pedestal voltage can provide proper blanking and the correct brightness level for all frames.

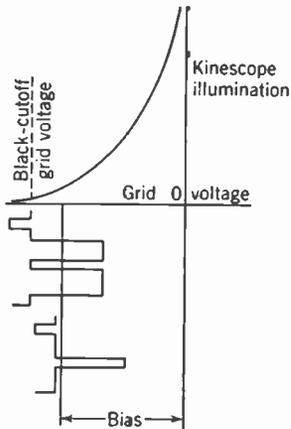


FIG. 13-6. Pedestals out of line on the kinescope grid, for the two lines of video signal in *b* and *c* of Fig. 13-5. A single value of bias cannot be correct for both signals.

**13-5. D-C Insertion.** The basic principle of d-c insertion is illustrated by the circuit in Fig. 13-7. The output voltage across  $R$  is the sum of the a-c generator output and the battery voltage because they are in series across  $R$ . With a battery voltage of 5 volts and a-c generator voltage of  $\pm 5$  volts, the effect of the battery voltage is to insert 5 d-c volts and shift the entire a-c signal to the new axis of  $+5$  volts. The variations of the a-c signal are still the same, but now they take place about a  $+5$ -volt axis instead of the original zero axis. This is positive restoration because the inserted d-c is positive with respect to the chassis ground. If the polarity of the battery is reversed, a negative 5 d-c volts will be inserted in series with the a-c input signal to provide negative restoration. The entire a-c signal will then be shifted below the zero axis, with the same variations taking place about the new axis of  $-5$  volts.

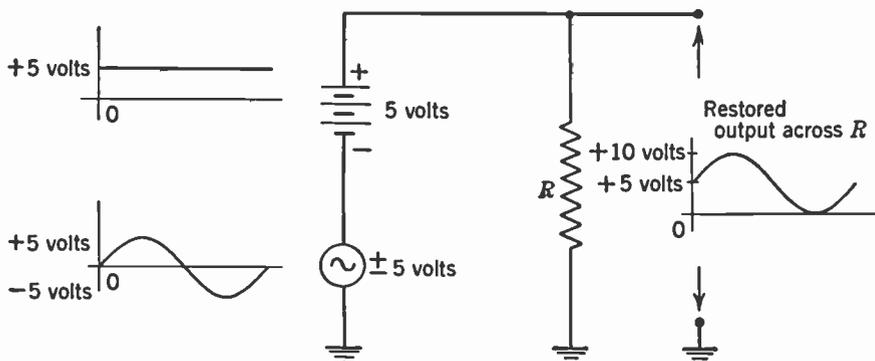


FIG. 13-7. Illustrating positive d-c insertion.

A battery cannot insert the d-c component required to line up the pedestals in the video signal because varying amounts of reinsertion are needed for video signals corresponding to picture frames that have different brightness levels. Instead, the d-c reinsertion at the transmitter or receiver is accomplished by a rectifier circuit called a *d-c restorer* or *clamping circuit*. Since the amount of d-c reinsertion must vary with the pedestal height of the video signal, the video voltage itself is rectified and used to provide the required d-c component. This type of circuit is able to insert a d-c component that is proportional to the pedestal height, thus lining up all the pedestals at a common voltage level. With the pedestals lined up, the manual brightness control can be adjusted for a value of bias on the kinescope grid that allows the pedestal voltage to drive the grid to cutoff. Then, if the d-c restorer keeps all the pedestals in line at a constant voltage level, they will all come at cutoff bias on the kinescope grid.

**13-6. Manual and Automatic Brightness-control Circuits.** To obtain the correct brightness in the reproduced picture, the d-c component of the

video signal must be present at the kinescope grid, and the manual brightness control set properly. The correct d-c component is obtained either by employing a d-c video amplifier for direct coupling of the video signal from the detector output, or by using a d-c restorer. The effect of the d-c restorer in inserting the correct d-c component for the required kinescope bias is illustrated in Fig. 13-8. The manual brightness control is adjusted so that the kinescope bias voltage is at cutoff with zero a-c video signal. Then, with a-c video signal input to provide the desired contrast, the d-c restorer automatically provides the amount of positive d-c voltage on the kinescope grid required to back off the kinescope bias from cutoff to the point that allows the negative pedestals to drive the instantaneous kinescope grid voltage to cutoff during blanking time. A light scene has more a-c signal swing than a dark scene, and a larger pedestal height, resulting in a larger positive reinserted d-c voltage from the restorer that moves the kinescope bias farther from cutoff to reproduce a brighter picture.

**Diode Restorer Circuit.** Figure 13-9 shows a diode restorer directly in the kinescope grid circuit. Here only the video output voltage across  $R_L$  is applied to the d-c restorer circuit consisting of  $R_1$ ,  $R_2$ ,  $C_1$ , and the inverted diode, which is one section of the 6AL5 twin diode. The video signal voltage for the restorer is taken across  $R_L$  so that the capacitances in the restorer circuit will not be in shunt with the video output tube's plate-to-cathode circuit. The 470,000-ohm resistor  $R_3$  couples the d-c restorer output directly to the kinescope grid and isolates the diode's shunt capacitance from the video amplifier. Additional isolation is provided by  $R_2$ , which is only 10,000 ohms because it is in series with  $C_1$  and limits the current that flows when the diode conducts.

With the positive picture phase required for the kinescope grid, the negative synchronizing pulses of the composite video signal make the diode cathode negative with respect to its grounded plate. Driving the cathode negative causes diode current to flow, producing a voltage across  $C_1$  that is proportional to the amount of negative voltage swing. Electrons can flow from cathode to plate in the diode to chassis ground, through the video output tube from cathode to plate, and through  $R_2$  to

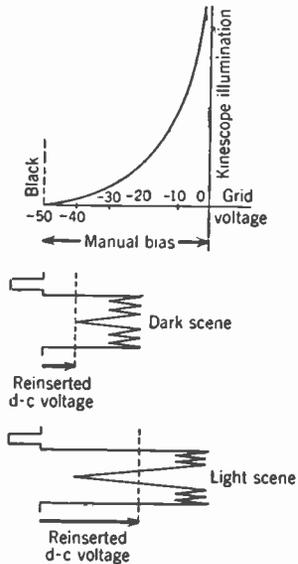


FIG. 13-8. Video signals on the kinescope grid with the pedestals in line at cutoff because of the d-c reinsertion. Manual brightness control set at cutoff.



restoration is necessary when the televised scene is one with a constant average brightness level, as is the case when the test pattern is transmitted.

Since the reinserted d-c voltage should vary from frame to frame when the average brightness changes, but not from line to line, the time constant of the d-c restorer is very long compared with the horizontal-line period but not with respect to the frame time. The time-constant value of the restorer circuit includes the coupling condenser for the diode rectifier and the diode shunt resistor through which the condenser charges or discharges when the diode is not conducting. For the case of grid-leak bias reinsertion, the grid coupling circuit  $R_g C_c$  determines the discharge time constant for the grid-cathode circuit, which operates essentially the same as a diode rectifier. A suitable value for the d-c restorer time constant is 0.03 to 0.1 sec, generally with a resistance of about 1 megohm and a coupling condenser of 0.03 to 0.1  $\mu f$ . With this time constant, which is very long compared with the horizontal-line period of 63.5  $\mu sec$ , the amount of reinserted d-c voltage across the diode coupling condenser is approximately 90 per cent of the peak amplitude of the sync pulses in the a-c composite video signal.

*D-C Component in Direct-coupled Video Amplifiers.* In a video amplifier having d-c coupling from the video detector output to the kinescope grid-cathode circuit, the d-c component is amplified with the a-c video signal and the pedestals are in line at the kinescope grid. The average value of the d-c video signal is a voltage of negative polarity at the kinescope grid, since the pedestal voltage must drive the grid negative from zero. The d-c component, therefore, makes the d-c grid voltage on the kinescope more negative than the manual bias. Furthermore, the video signal for a dark scene has a larger negative d-c component than a white scene. The characteristics of brightness control with the d-c component of the video signal can be compared to a d-c restorer system as follows:

#### Negative D-C Component

1. Manual brightness-control bias on kinescope grid near zero without video signal
2. Therefore, raster is visible without video signal at normal brightness setting
3. D-C component has negative polarity at kinescope grid
4. With a light scene, the kinescope bias shifts slightly more negative than the manual bias
5. With a dark scene, the kinescope bias shifts far negative near cutoff

#### Positive D-C Restorer

- Manual brightness-control bias on kinescope grid at cutoff without video signal
- Therefore, raster is not visible without video signal, unless brightness setting is increased
- Restorer reinserts positive d-c voltage at kinescope grid
- With a light scene, kinescope bias moves far in the positive direction, away from the manual bias at cutoff
- With a dark scene the kinescope bias shifts slightly in the positive direction, near the manual bias at cutoff

*Kinescope Grid-leak Bias.* In addition to the use of a diode restorer or a d-c coupled video amplifier, the required d-c component can be obtained by grid-leak bias in the kinescope grid-cathode circuit. If the video signal input drives the kinescope grid slightly positive, grid current flows to charge the coupling condenser. With the grid-cathode circuit operating effectively as a diode rectifier the resultant grid-leak bias provides the d-c voltage for restoration. The polarity of this reinserted d-c voltage is negative at the kinescope grid, with the white signal amplitudes clamped at zero grid voltage.

**13-7. D-C Restorer Troubles.** When the scene changes from light to dark, the background of the reproduced picture is not dark enough and retrace lines may show without the correct d-c component in the video signal for the kinescope grid. If the brightness control is reduced to eliminate the retrace lines the background will be too dark for light scenes. These troubles in obtaining the correct background level occur when the receiver has capacitive coupling in the video amplifier but no d-c restorer, or if the d-c restorer is not operating. When the receiver has internal vertical blanking the vertical retrace lines are not visible but the background level is still wrong without the correct d-c component.

Operation of the restorer circuit can be checked by measuring the d-c voltage between the kinescope grid and cathode, while turning the channel selector on and off a station. With video signal input to the restorer when a station is being received, the kinescope bias voltage changes about 20 to 30 volts.

*Effect of Noise Voltages on the D-C Restorer.* The d-c restorer must respond quickly to peak voltages in the direction of the sync pulses when the diode rectifier conducts, but the circuit recovers slowly as the coupling condenser charges or discharges through the high resistance in shunt with the diode. As a result, the restorer can change the reinserted d-c voltage more quickly in the direction of increasing brightness than for decreasing brightness. Noise peaks in the video signal, therefore, can raise the level of the reinserted d-c voltage, increasing the brightness. The effect of noise pulses on the reinserted d-c voltage is aggravated by too high a resistance in shunt with the diode.

*No Control of Brightness.* When the kinescope screen has illumination but its intensity cannot be varied with the manual brightness control, the trouble is in the d-c bias circuit of the kinescope. Measure the bias with a d-c voltmeter to see if it varies with rotation of the brightness control. To check for an internal short in the kinescope grid-cathode circuit, measure the bias at the kinescope with the kinescope socket on and with it off.

## REVIEW QUESTIONS

1. Why is the pedestal level voltage transmitted at a constant percentage of the picture carrier amplitude?
2. If the pedestal voltage drives the kinescope grid voltage to cutoff, what will be the effect of the synchronizing pulses on the kinescope grid voltage?
3. What kinescope voltage is varied by the manual brightness control? Describe how to adjust the brightness control by watching the picture.
4. If the manual brightness control is set for too much negative bias, what will be the effect on the picture? What is the effect when the control is set for too little bias?
5. A video detector is directly coupled to the video output amplifier, which is capacitively coupled to the kinescope grid. Is d-c reinsertion required? Why?
6. One video amplifier of two stages has two capacitive couplings, while another amplifier has only one capacitive coupling. Is there any difference in the amount of d-c reinsertion required for the amplifiers? Explain.
7. Is d-c reinsertion necessary in a receiver using direct-coupled video amplifiers? Why?
8. Referring to the diagram in Fig. 13-9, give the function of  $R_1$ ,  $R_2$ ,  $R_L$ ,  $C_1$ , and  $C_2$ .
9. In a receiver having a d-c restorer, with the manual brightness control at its normal setting, why is the raster normally not visible without any video signal on the kinescope grid?
10. Describe briefly how to check whether the d-c restorer is operating.

## CHAPTER 14

### VIDEO DETECTOR

The video detector is the second detector for the picture signal, following the last i-f stage in the superheterodyne television receiver. The AM picture signal is rectified and filtered in the video detector stage to provide an output signal that corresponds to the modulation envelope of the input, as illustrated in the diode detector circuit of Fig. 14-1. This envelope is the original composite video signal containing all the information required for reproduction of the picture.

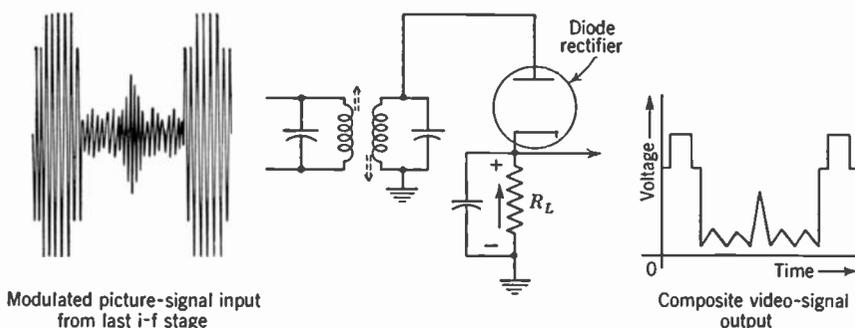


FIG. 14-1. Function of the video detector. The modulated picture signal is rectified and filtered to provide the composite video signal in the output.

**14-1. Detection.** The video detector action is the same as in any detector for an AM signal. The diode conducts plate current only when the a-c input voltage makes the plate positive with respect to its cathode, resulting in a voltage drop across the load resistor  $R_L$  in Fig. 14-1. The output voltage is of the polarity shown because the diode conducts from cathode to plate only when the plate is positive, producing the d-c voltage across  $R_L$  that is the rectified input signal. With more a-c signal input, the d-c output voltage across the load resistance is greater, but of the same polarity. When the a-c input signal voltage decreases, the rectified output voltage also decreases.

It is necessary to rectify the modulated input signal because the amplitude variations that correspond to the desired intelligence have an average

value of zero, resulting from the symmetrical envelope. After rectification, the amplitude variations of the modulated signal can be obtained. With adequate filtering of the intermediate frequencies so that the output voltage cannot follow the rapid variations in the individual cycles of the i-f input signal, but only the relatively slow variations in amplitude corresponding to the envelope, the output voltage across  $R_L$  corresponds to the modulation envelope of the input. It does not matter which polarity of the a-c input signal is used for the video detector, since both sides of the modulation envelope are the same even though vestigial-side-band transmission is used. The video detector, then, is basically the same as any ordinary AM detector. However, the questions of polarity, uniform frequency response, and adequate filtering make the problem of video detection more difficult.

**14-2. Detector Polarity.** Two polarities are possible for the output voltage of a diode detector, depending on whether the load resistor is in the

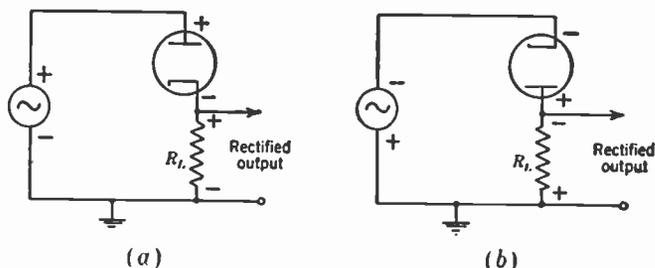


FIG. 14-2. Rectifier polarity. (a) The diode load resistance in the cathode circuit provides output voltage of positive polarity. (b) The diode load resistance in the plate circuit provides negative output voltage.

cathode or plate circuit. The polarity is not important in an audio system, because the phase of the a-c audio signal for a loudspeaker does not matter in the sound reproduction. For the picture reproduction, though, phase reversal of the video signal driving the kinescope grid produces a negative picture.

The question of polarity in a rectifier is sometimes confused because of the lack of a reference point. In Fig. 14-2 equivalent diagrams are shown for a rectifier with negative voltage output and one having positive output, both with reference to the chassis ground. The input signal voltage of the a-c generator reverses in polarity for each half cycle, driving the diode plate alternately positive and negative. Current flows through the load resistor  $R_L$  only when the generator makes the plate positive with respect to cathode. For this polarity of applied voltage, the voltage across the diode and the load resistor must be as shown in Fig. 14-2a. The plate is positive with respect to cathode, while the cathode side of the load resistor is more positive than the end connected to the negative side of

the generator and chassis ground. Output voltage taken from the cathode is positive with respect to the chassis, although still negative with respect to the diode plate. For the negative half cycle of the a-c input signal, the diode is effectively an open circuit and there is no voltage drop across the diode load resistor. This circuit arrangement, therefore, can be used to obtain from the diode cathode a rectified output voltage that is positive with respect to the chassis and varies in magnitude with the amount of a-c signal input during the positive half cycle.

With the diode inverted as in Fig. 14-2*b*, current flows through the load resistor only when the generator voltage makes the diode cathode negative with respect to its plate. When diode current flows, the voltages across the diode and the load resistor  $R_L$  must be as shown, with the plate positive with respect to its cathode, and the plate side of the load resistor negative with respect to the end connected to the positive side of the generator and chassis ground. This circuit, then, can be used to obtain from the diode plate a rectified output voltage that is negative with respect to the chassis.

The polarity of the video detector output signal voltage can now be examined, keeping in mind the fact that negative transmission is used for the picture signal. This means that the tips of the synchronizing pulses correspond to the positive and negative peaks of the modulated picture carrier and represent the greatest amplitude of the a-c input signal voltage to the video detector. Maximum white corresponds to the smallest amplitudes in the i-f input signal.

In the video detector circuit shown in Fig. 14-3*a* the diode load resistance is in the cathode circuit and the polarity of the output voltage is positive with respect to the chassis ground. Since the synchronizing pulses supply maximum signal input, the rectified output voltage has its maximum positive values for the peaks of the synchronizing pulses. As shown in the figure, the output voltage varies in magnitude with the amount of input signal to provide the desired composite video signal. The video signal output has negative picture phase, since the maximum white camera signal is represented by the least positive or most negative voltage, and positive sync phase because the synchronizing pulses are the most positive part of the video signal. Using capacitive coupling, the signal voltage coupled to the next stage is in its a-c form with the synchronizing pulse voltage producing the maximum positive swing and the maximum white camera signal equal to the maximum negative swing.

In Fig. 14-3*b*, the output of the video detector with the load resistance in the plate circuit has positive picture phase and negative sync phase. The detector's output voltage is most negative for the synchronizing pulses, which correspond to maximum i-f signal input. With capacitive coupling to the next stage, the a-c form of the video signal has the syn-

chronizing pulses as the maximum negative swing and the white camera signal voltage as the most positive swing.

The video detector circuit generally used is the one with the load resistance in the plate circuit. Its output of negative sync phase is the preferable polarity for the signal input to the video amplifier stage following the detector, because noise pulse voltages greater than the sync pulses can be cut off. With the load resistor in the plate, this detector circuit has less output capacitance because the cathode-to-heater capacitance is not across the output load resistance. Also, this circuit arrangement is less

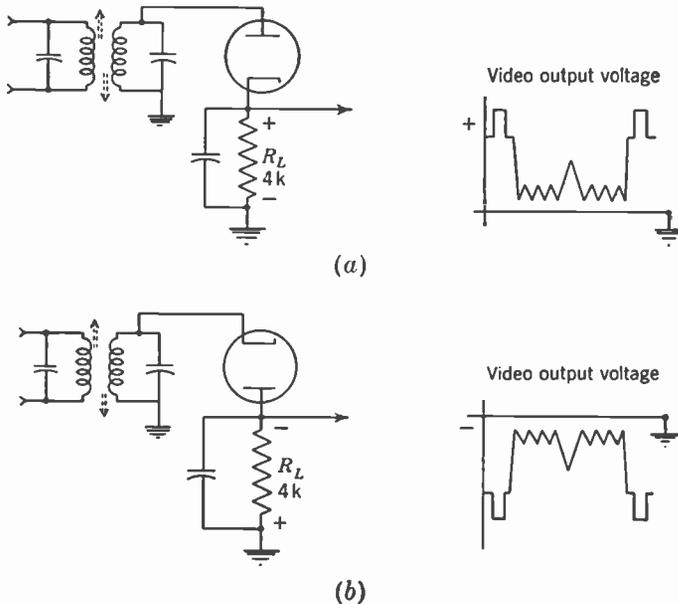


FIG. 14-3. Video detector polarity. (a) Diode load resistance in the cathode circuit. (b) Diode load resistance in the plate circuit.

susceptible to hum introduced in the detector by heater-cathode leakage, since the cathode is grounded and has no load impedance for the 60-cps hum voltage. The positive picture phase output is the polarity required for the kinescope grid signal. With two video amplifier stages following this detector, the video output is coupled to the kinescope grid; with one video amplifier stage the video output is coupled to the kinescope cathode.

**14-3. Video Detector Load Resistance.** The frequency response of the video detector is just as important as the relative gain of the video amplifiers for the wide range of video frequencies. Although not an amplifier, the video detector must have uniform output up to the highest video frequency in order to provide for the amplifiers a video signal that contains the original picture information without frequency distortion.

The problem of high-frequency response in the video detector is the same as in the video amplifier. Because of the tube and stray capacitances shunting the detector load resistance, the video detector output tends to decrease for the higher video frequencies, since the plate resistance of the tube and the load resistance act as a voltage divider for the signal. The decreasing reactance of the shunt capacitance reduces the plate load impedance for the higher video frequencies, providing less output voltage at these frequencies. In order to minimize the effect of the shunt capacitance and provide uniform output voltage up to 4 Mc, the load resistance of the video detector is usually about 4,000 ohms, which is low compared with the 0.25 to 2.0 megohms commonly used in an audio detector. Using so low a value of load resistance in the video detector, in order to obtain uniform high-frequency response, can reduce the amount of output signal because of the voltage-divider action between the load resistance and the plate resistance of the tube. The 6AL5 twin diode is

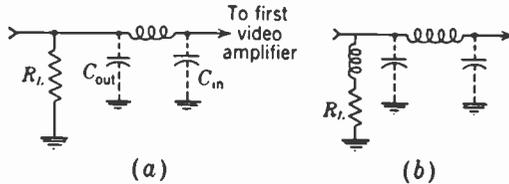


FIG. 14-4. Filter coupling circuits for the video detector output. (a)  $\pi$ -type filter coupling equivalent to series peaking. (b) Combination peaking circuit.

commonly used for the video detector because of its low plate resistance of about 300 ohms for each diode section, compared with about 4,000 ohms for the 6H6.

**14-4. Video Detector Filter.** In addition to rectifying the modulated carrier signal, the detector must provide adequate filtering of the intermediate frequencies. The filtering corresponds to the function of the usual r-f bypass condenser across the diode detector load resistance, which cannot discharge at the r-f rate but can follow only the relatively slow variations corresponding to the envelope. However, the simple bypass condenser cannot provide adequate filtering for the video detector because of the high video frequencies. With an intermediate frequency of 26 Mc for the picture carrier, the ratio of intermediate frequency to the highest video frequency of 4 Mc is only 6.5. A simple bypass condenser does not provide sufficient discrimination between the intermediate frequency and the high video frequencies. A condenser large enough to bypass the intermediate frequencies adequately would have a reactance too low for the high video frequencies; and a condenser small enough to have an appreciable reactance for the high video frequencies would not bypass the intermediate frequencies effectively.

It is usual to employ filter coupling in the video detector output in order to allow adequate bypassing of the intermediate frequencies without reducing the coupling impedance for the wide band of video frequencies. This is often a  $\pi$ -type low-pass filter, as shown in Fig. 14-4a, providing much more pronounced band-pass characteristics than a single bypass condenser. The filter coupling is the same as series peaking for high-frequency compensation of the video amplifier, and the design can be carried out in terms of the input and output capacitances by using the same procedure described in Chap. 12 for series peaking. Thus, the filter coupling has the dual functions of filtering out the intermediate frequencies and compensating the video detector for uniform high-frequency response.

In many cases, the filter coupling includes an additional peaking coil in series with the load resistance, as shown in Fig. 14-4b, compensating the high-frequency response of the video detector in the same way as combination peaking for the video amplifier.

**14-5. Video Detector Circuits.** The picture signal from the last i-f stage has a peak amplitude of about five volts or more allowing use of the

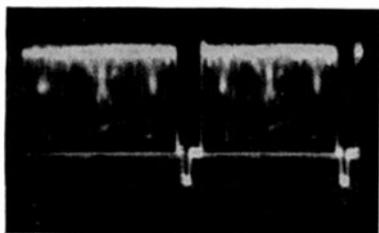
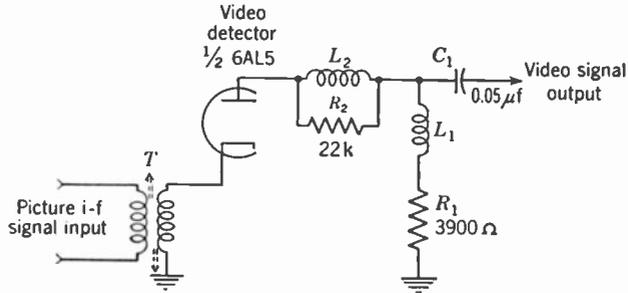


FIG. 14-5. Oscillogram of composite video signal. Oscilloscope internal sweep at 15,750/2 cps to show video signal for two horizontal scanning lines. Horizontal synchronizing pulses at bottom of waveform.

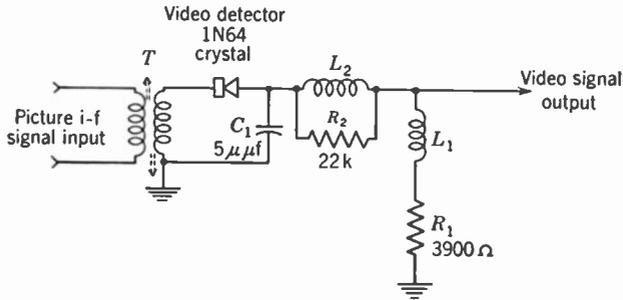
diode detector circuit. One section of the 6AL5 twin diode is commonly the half-wave rectifier for the video detector, or a crystal diode can be used. As illustrated by the typical circuits in Fig. 14-6, the video detector circuit usually has the diode load resistor in the plate circuit, to produce video output of positive picture phase with the synchronizing pulses the most negative part of the composite video signal. The video detector can be directly coupled to the video amplifier, or a coupling condenser can be used to block the d-c level of the detector's output signal. Figure 14-5 shows an oscillogram of the composite video signal output of the detector.

Referring to the video detector circuit in Fig. 14-6a, the modulated picture signal output of the last i-f stage is coupled to the cathode of the video detector, which uses one-half the 6AL5 twin diode for the half-wave rectifier. The secondary of the i-f transformer  $T$  tunes with the stray capacitance to provide at the detector cathode the i-f signal input voltage with respect to ground. The diode load resistor is  $R_1$  in the plate circuit. Uniform response for the high video frequencies up to 4 Mc is obtained by using the relatively low value of 3,900 ohms for the diode load resistance, with the peaking coils  $L_1$  and  $L_2$ , which resonate with the stray capaci-

tances in the detector's output circuit. This low-pass filter circuit also filters out the intermediate frequencies from the detector output. The 22,000-ohm resistor  $R_2$  in parallel with  $L_1$  is a damping resistor for this peaking coil. The coupling condenser  $C_1$  couples the a-c variations of the video signal output from the detector to the control grid of the first video amplifier. With the diode load resistance in the plate circuit, the detected video signal has positive picture phase and negative sync polarity. Two



(a)



(b)

FIG. 14-6. Typical video detectors. (a) Circuit using one diode section of 6AL5 tube. (b) Circuit using crystal diode.

video amplifier stages follow the detector in this circuit and the video output is coupled to the kinescope control grid.

The video detector circuit in Fig. 14-6b uses a 1N64 germanium crystal-diode rectifier. The modulated picture i-f signal is coupled to the cathode of the crystal detector, with the 3,900-ohm diode load resistor  $R_1$  in the anode circuit. The peaking coils  $L_1$  and  $L_2$  resonate with the stray capacitances in the detector's output circuit to form a low-pass filter circuit that provides uniform response for the high video frequencies up to 4 Mc, and bypasses the intermediate frequencies. The composite video signal of positive picture phase and negative sync polarity from the

detector is directly coupled to the video output amplifier in this circuit. With only one video amplifier stage, the video detector output is often directly coupled to the control grid of the video output tube and its plate circuit is directly coupled to the kinescope cathode. This circuit arrangement provides direct coupling of the video signal from the detector output circuit to the kinescope input circuit, and no d-c restorer is necessary.

The advantages of using a crystal diode for the video detector are its low shunt capacitance, compactness, and the fact that it does not have any heater current. Because of its small size, the crystal diode can be mounted in the shield can of the last i-f transformer. However, the crystal diode has the disadvantage of a relatively low back-resistance compared with a vacuum tube.

**14-6. Functions of the Composite Video Signal.** The composite video signal output of the video detector contains all the information needed for reproducing the picture.

Consequently, the video detector output is used in the various receiver circuits that require the information contained in the composite video signal. As illustrated in Fig. 14-7, the three functions of the composite video signal in the receiver circuits are:

1. The camera signal variations of the composite video contain the picture information corresponding to the basic elements of light and shade in the original scene. Therefore, the video signal is needed at the control grid-cathode circuit of the picture tube in order to vary the intensity of the electron scanning beam and reproduce the picture information. The complete composite video signal with the blanking and sync pulses is coupled to the kinescope, not just the camera signal. The pedestal level provides the brightness reference and the blanking pulses blank out retrace lines.

2. The synchronizing pulses of the composite video signal provide the timing information needed to synchronize the deflection circuits in the receiver. Therefore, the composite video signal must be coupled to a sync separator circuit. This is a clipper stage that separates the synchronizing pulses from the rest of the composite video signal. The separated synchronizing pulses can then be used by the deflection circuits for controlling the frequency of the receiver's scanning generators.

3. The level of the pedestals or any part of the synchronizing pulses in the composite video signal output of the video detector indicates the

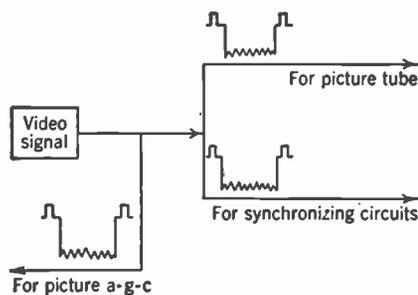


Fig. 14-7. Three functions of the composite video signal.

strength of the r-f picture carrier signal. Therefore, it can be used for automatic gain control in the receiver.

These functions of the composite video signal require that it be coupled to several different circuits in the receiver. However, one circuit can function independently of the others. For instance, clipping the synchronizing pulses in the sync separator stage does not interfere with normal operation of the video amplifier, which still provides the complete composite video signal for the picture tube.

**14-7. Video Detector Troubles.** Since the video detector is part of the signal circuits for the picture, a trouble in the detector affects the picture, while the raster is normal. The output circuit of the video detector is the input circuit of the video amplifier and can produce the same picture troubles. When the sound take-off circuit is after the detector, trouble in this stage can also affect the sound.

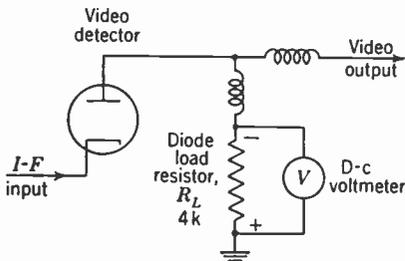


FIG. 14-8. Measuring the rectified output across the detector load resistor with a d-c voltmeter.

**Checking Detector Output.** As illustrated in Fig. 14-8, the voltage across the diode load resistor of the detector can be measured with a d-c voltmeter to see if the detector is producing rectified signal voltage in the output. The video output signal of the rectifier is a varying d-c voltage of fixed polarity because the rectifier conducts in only one direction. No B+ voltage is applied to the detector for plate voltage. The i-f signal operates the detector by supplying the a-c input voltage that drives the diode plate positive to produce rectified plate current. The d-c output voltage, therefore, is rectified signal. With i-f signal input, the d-c output voltage of the detector is about 2 to 5 volts. When the receiver is switched off channel to remove the signal input, the detected output should drop to a much lower value. The small output voltage without signal is rectified receiver noise.

**Defective Crystal Rectifiers.** A crystal rectifier can be checked by measuring its front and back resistances with an ohmmeter. The forward resistance should be 50 to 200 ohms, approximately, and the back resistance 50,000 to 200,000 ohms. When replacing a crystal diode, it is important to note the correct polarity.

**Hum in the Detector.** Without any B+ voltage applied, hum introduced in the detector would be 60-cps hum produced by heater-cathode

leakage. A crystal-diode detector cannot introduce hum because there is no heater current.

## REVIEW QUESTIONS

1. What is the function of the video detector stage?
2. What are two requirements for obtaining rectified output signal from a detector?
3. Explain briefly why a detector must rectify the modulated carrier signal.
4. Why is the load resistance in a video detector stage much lower than the load resistance in a detector stage for audio signal?
5. Draw the diagram of a video detector circuit that provides output signal having negative sync polarity. Show the composite video signal output of the detector, indicating polarity.
6. Give one advantage of a video detector that has the load resistance in the plate circuit, compared with the load resistance in the cathode.
7. Give three functions of the composite video signal output of the video detector.
8. Referring to Fig. 14-6a, state briefly the function of  $R_1$ ,  $R_2$ , and  $C_1$ .
9. Explain briefly why the output of the video detector can be checked with a d-c voltmeter. Referring to Fig. 14-6, where would the d-c voltmeter be connected in the detector circuits in *a* and *b*?

## CHAPTER 15

### AUTOMATIC GAIN CONTROL

The automatic gain control circuit controls the amount of amplification of the signal in the receiver by automatically adjusting the gain of the i-f and r-f amplifiers according to the strength of the carrier signal received. The advantage of automatic gain control is that it provides relatively constant output signal from the second detector with wide variations of signal input to the receiver. In a receiver for sound signal the automatic gain control (a-g-c) is commonly called automatic volume control (a-v-c). The television receiver usually has an a-g-c circuit for the stages amplifying the picture signal but generally there is no separate automatic gain control for the FM sound circuits.

**15-1. Requirements of the A-G-C Circuit.** The main requirement for the automatic gain control is a negative d-c voltage proportional to the carrier signal strength, to control the bias on the i-f and r-f amplifiers. In order to obtain the d-c voltage indicating carrier strength, the signal itself is rectified and filtered to supply the control bias. As illustrated in Fig. 15-1, this d-c bias voltage is applied in series with the a-c signal input to the control grid of the i-f and r-f amplifier stages controlled by automatic gain control. With more negative bias added to the grid of the amplifier, the gain of the stage is reduced. The gain is reduced by the a-g-c circuit more for strong signals than for weak signals.

*A-G-C Rectifier.* The a-g-c rectifier stage rectifies the signal to produce the negative d-c control voltage. In Fig. 15-1, the rectified output voltage is negative with respect to chassis ground because the diode load resistor  $R_L$  is in the plate circuit of the a-g-c rectifier. This is the required polarity, since the negative bias is applied to the control grid of each controlled stage, with the a-g-c line serving as a high-resistance d-c return path for the grid. Any increase in carrier signal strength produces more rectified output voltage and more negative bias, which reduces the gain of the controlled stages. With less signal input to the receiver, the a-g-c circuit develops less negative bias voltage, allowing more gain than for strong signals.



cuit. Generally, the controlled stages also have some cathode bias, which provides a minimum bias of 1 to 3 volts when there is little or no signal input.

As the a-g-c voltage makes the bias on the controlled stages more negative toward cutoff, the grid has less control on the plate current because fewer electrons pass through to the plate. The grid-to-plate transconductance ( $g_m$ ) and the amplification factor ( $\mu$ ) of the tube decrease when

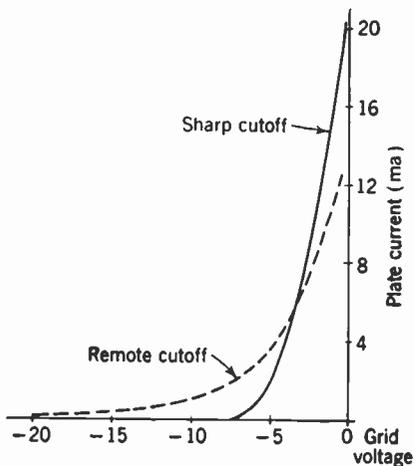


FIG. 15-2. Comparison of the grid-plate transfer characteristic curves of the 6CB6 sharp cutoff pentode and the 6BA6 remote cutoff pentode.

the a-g-c bias voltage becomes more negative, therefore, reducing the gain of the amplifiers controlled by the a-g-c line.

Stages where the bias voltage can be varied over a wide range often use remote-cutoff tubes. These are also called *variable- $\mu$*  or *super-control* tubes. As illustrated by the grid-plate transfer characteristic curves in Fig. 15-2, the remote-cutoff curve has a more gradual change in plate current for different grid voltages and a more negative grid-cutoff voltage. As a result, remote-cutoff tubes can operate with a wider range of grid-bias values and can have more grid-voltage signal swing without intro-

ducing excessive amplitude distortion in the plate circuit. Remote-cutoff tubes have less transconductance than the equivalent sharp-cutoff tubes, however, resulting in less gain. When the signal level is low and the a-g-c bias does not vary over a wide range, sharp-cutoff tubes can be used in the controlled stages.

**Delayed Automatic Gain Control.** The disadvantage of the simple a-g-c circuit in Fig. 15-1 is that it reduces the gain for very weak signals, when maximum sensitivity in the receiver is desired. Automatic gain control reduces the gain for all signals. The uniform output is obtained by cutting down the gain for strong signals more than for weak signals. Therefore, it is preferable to bias the a-g-c rectifier so that no diode current flows and no a-g-c voltage is developed for signals too weak to make the rectifier conduct. In this way, there is no a-g-c voltage to reduce the gain until the signal has enough amplitude to overcome the rectifier bias. The bias on the a-g-c rectifier is a delay bias, therefore, and the circuit is called *delayed automatic gain control*. This requires a separate a-g-c rectifier because it is not desired to bias the detector, which should be allowed

to operate for the weakest signals. The bias for the a-g-c rectifier comes from a source other than the signal itself, since the bias must be present with no signal input.

#### 15-2. Advantages of Automatic Gain Control for the Picture Signal.

Automatic gain control maintains a relatively constant signal level of video signal from the video detector for wide variations of signal input to the receiver. The signal strength will be different when tuning from one station to another. Slow changes in the plate and filament supply voltages, due to power-line variations, may change the gain of the amplifiers. In addition, the picture signal can vary in strength because of moving conductors in or near the receiver antenna system, especially airplanes flying nearby. The picture automatic gain control minimizes the effects of these changes in signal strength on the reproduced picture.

*Automatic Contrast Control.* The main advantage of automatic control of the gain of the picture i-f stages is the relatively constant contrast in the reproduced picture for a given setting of the manual contrast control. For this reason, picture automatic gain control may be called *automatic contrast control*. With the manual control set for the desired contrast, the picture a-g-c circuit maintains this contrast level by providing constant video signal amplitude from the detector. Although the strength of the received signal is usually different when switching from one station to another, the picture a-g-c circuit can automatically keep the video signal at the same level so that the manual contrast control need not be readjusted. A typical a-g-c circuit can maintain the video signal amplitude within a range of 2 to 1 while the picture signal input at the antenna terminals varies over a range of 100 to 1.

In a receiver with picture automatic gain control, the manual contrast control is in the video amplifier with the function of setting the contrast over a relatively narrow range, while the picture automatic gain control controls the gain of the i-f and r-f amplifiers automatically to adjust for the wide range of input signal levels. Adjusting the manual contrast control for the desired contrast in the picture is easier, therefore, with picture automatic gain control. Picture distortion and loss of synchronization, caused by overloading a picture i-f stage when the previous stages operate at full gain with a strong input signal, are minimized by the picture a-g-c circuit. Also, separation of the synchronizing pulses is simplified when the composite video signal has constant amplitude.

*Airplane Flutter.* This is the rise and fall in picture intensity caused by reflections of the picture signal from airplanes flying nearby. In addition to the variations in intensity, the signal reflections from the moving airplane produce a changing ghost in the picture. If the airplane is transmitting a radio signal it may produce diagonal bars in the picture just like typical r-f interference. Also, the sound may be garbled and

pulsate in volume. The effects of fading in the signal caused by airplane flutter can be reduced by the picture a-g-c circuit.

**15-3. Picture A-G-C Circuits.** The a-g-c circuit must provide an a-g-c bias voltage that indicates the strength of the picture carrier signal, independent of the picture information. In a sound system, the filtered d-c voltage from the detector is proportional to the average amplitude of the carrier, which is suitable for the a-g-c bias voltage because it indicates carrier strength. For the picture signal, though, the average value of the carrier varies with the picture information and is not a measure of the strength of the carrier signal.

In the negative transmission system, the blanking pulses bring the carrier amplitude up to the constant 75 per cent pedestal level, and the tips of the synchronizing pulses produce the peak carrier amplitude. As a result, the synchronizing pulses have constant amplitude for any level of picture carrier signal, regardless of the picture information. Therefore, the peak carrier amplitude produced by the tips of the synchronizing pulse voltage indicates carrier-signal strength.

The a-g-c rectifier for the picture a-g-c circuit produces d-c output proportional to the peak amplitude of the signal input, in order to provide an a-g-c bias voltage that indicates carrier strength. Self-bias with a filter circuit having a time constant much longer than the time between horizontal synchronizing pulses makes the circuit a peak rectifier. Either the modulated picture signal or the composite video signal can be rectified to provide the a-g-c bias voltage. When the video signal is used, however, it must have d-c coupling to the a-g-c rectifier in order to preserve the d-c component and keep the pedestals in line so that the peak of sync indicates carrier strength correctly.

The stages controlled by the picture a-g-c circuit are generally the r-f amplifier and two i-f stages. Usually, the last i-f stage has fixed bias without automatic gain control. Automatic gain control is applied to earlier stages in the receiver because the effectiveness of the a-g-c bias is proportional to the amount of gain from the grid of the controlled stage to the a-g-c rectifier.

Figure 15-3 shows a simple a-g-c circuit for the picture signal. The i-f transformer  $T_{303}$  supplies the i-f picture signal from the third i-f stage to the video detector and a-g-c rectifier in the 6AL5 twin diode.  $C_{305}$  couples the signal to the a-g-c diode plate. The i-f signal makes current flow in the a-g-c rectifier, producing negative d-c output across the load resistor in the plate circuit, which is  $R_{314}$  because it returns to the cathode through chassis ground.  $C_{306}$  and  $R_{313}$  form the a-g-c filter, with a time constant of approximately 0.1 sec. Since  $C_{306}$  cannot discharge appreciably during the time between sync pulse peaks, the voltage across the condenser is approximately equal to the peak carrier amplitude, which is a

measure of carrier-signal strength. The negative d-c voltage across  $C_{306}$  is the a-g-c control bias connected to the a-g-c line. This a-g-c bias is connected to the control grid of the first two i-f stages and the r-f amplifier.  $C_{302A}$  is a bypass condenser for the a-g-c side of the grid coil of the second i-f amplifier.  $R_{302}C_{301A}$  is a decoupling filter for the grid of the first i-f amplifier, and  $R_{304}C_{313}$  is a decoupling filter for the grid of the r-f amplifier.

*Amplified Automatic Gain Control.* By amplifying the a-g-c control voltage, an amplified a-g-c circuit develops more a-g-c bias than is available from the i-f carrier signal at the second detector. The advantage of amplified automatic gain control is that the constant video signal amplitude

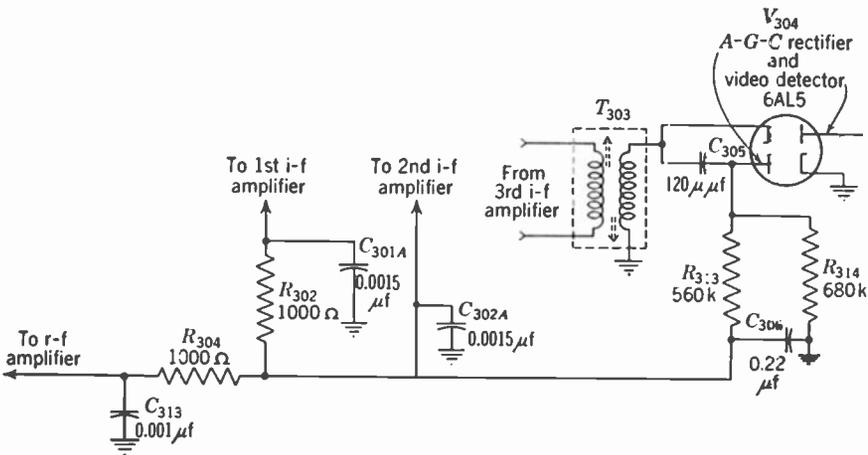


FIG. 15-3. A picture a-g-c circuit.

can be maintained over a wider range of carrier-signal levels. Amplification of the control voltage can take place before or after the a-g-c rectifier. An a-g-c amplifier for the rectified a-g-c voltage, however, must be a d-c amplifier.

*Bias Clamp.* In some television receivers the picture a-g-c circuit uses a diode to clamp the bias voltage for the r-f amplifier at zero just for low signal levels. This bias-clamp arrangement is illustrated in Fig. 15-4. The diode connects to the d-c supply voltage through  $R_1$  to make the plate several volts positive—say 5 volts. Also applied to the diode plate is the negative a-g-c bias voltage developed by the signal. As long as the signal level is so low that less than 5 volts of negative a-g-c bias voltage is developed, the diode plate is positive and it conducts. Therefore, the bias line to the grid of the r-f amplifier is effectively grounded through the low resistance of the conducting diode. When the a-g-c bias is more than -5 volts, however, the diode plate is negative and it cannot conduct.

Then the r-f bias equals the difference between the negative a-g-c voltage and the small positive voltage from the d-c supply, so that the r-f amplifier has enough bias to prevent cross modulation on strong signals. As a result, the r-f bias is clamped at zero for maximum gain for weak signals and delayed until the negative a-g-c bias voltage is large enough with a strong signal to stop the diode from conducting and provide the required negative bias. The decoupling resistor  $R_2$  isolates the i-f bias line from the bias-clamp circuit, allowing the normal a-g-c bias for the i-f amplifier. Note that the clamp circuit always keeps the r-f bias below the i-f bias for a better signal-to-noise ratio in the r-f amplifier.

**Keyed Automatic Gain Control.** A keyed, or gated, a-g-c circuit is pulsed into operation to produce rectified a-g-c voltage only during the time of the keying pulses. The a-g-c filter provides a steady d-c bias voltage, however, for the a-g-c line. The advantage of keyed automatic

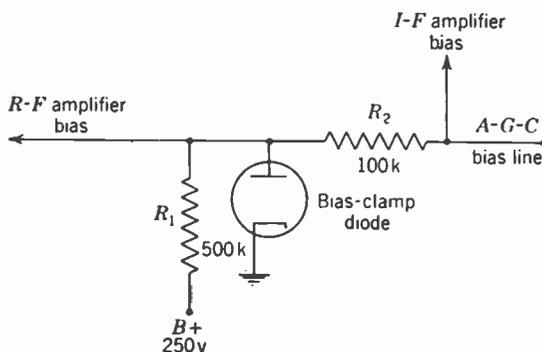


FIG. 15-4. A bias-clamp circuit.

gain control is that it permits a fast-acting a-g-c circuit which is relatively immune to noise.

**15-4. Keyed A-G-C Circuit.** The main problem in a picture a-g-c circuit is producing a bias control voltage that measures the peak carrier level in order to indicate the correct carrier strength, without making the a-g-c filter time constant too long. With a slow a-g-c circuit, having an  $RC$  time constant of about 0.3 sec or more, the a-g-c bias may not be able to come down to normal quickly enough after noise pulses have temporarily increased the peak carrier amplitude. The excessive a-g-c bias reduces the video signal level and there may be loss of synchronization. In addition, the slow a-g-c circuit cannot control the receiver gain rapidly enough to compensate for the fading signal caused by airplanes flying nearby, resulting in airplane flutter. To eliminate these difficulties a fast a-g-c circuit, with a time constant of about 0.03 sec or less, is needed. However, if the  $RC$  time constant allows the a-g-c voltage to change too quickly, the peak signal amplitude cannot be measured to indicate the

carrier strength correctly. One particular trouble is the fact that a fast a-g-c circuit can interpret the vertical synchronizing pulses as an increase in carrier level because of their greater pulse width. As a result, the a-g-c voltage rises slightly during this time, resulting in 60-cps sync voltage on the a-g-c line. Also, the increased bias during the vertical synchronizing pulse reduces the receiver gain and puts the vertical pulses *in a hole* with respect to the amplitude of the other pulses.

A solution to the problem of a fast noise-immune picture a-g-c circuit is a system for automatically keying the a-g-c rectifier or amplifier on and off. In this circuit, the a-g-c tube is driven into conduction by keying or gating pulses at the horizontal scanning frequency. The keying pulses are taken from the horizontal output circuit in the deflection circuits of the receiver, in order to pulse the a-g-c tube and allow current to flow only during the horizontal synchronizing pulse time. As a result, the output of the a-g-c tube is comparatively free from noise because the a-g-c circuit does not operate between keying pulses. In addition, fast

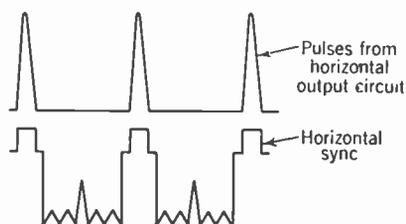


FIG. 15-5. Keying pulses occurring during horizontal sync pulse time for keyed a-g-c tube.

a-g-c action can be obtained without any error from the vertical synchronizing pulses because their increased width is not measured by the keyed a-g-c tube.

The pulses in the horizontal output circuit occur at the horizontal line-scanning frequency, which is the same as the frequency of the horizontal synchronizing pulses. With normal phasing of the horizontal scanning, the voltage peak produced in the horizontal output circuit during flyback occurs during blanking and coincides with the time of the horizontal synchronizing pulses on the composite video signal. This is illustrated in Fig. 15-5. Then the a-g-c tube is keyed into conduction during the horizontal synchronizing pulse time and produces the desired a-g-c voltage. If the horizontal scanning is not in synchronization, however, there will be no output from the keyed a-g-c rectifier because it is on at the wrong time. The polarity of the horizontal flyback pulse coupled to the a-g-c circuit is positive at the plate of the a-g-c tube, to make it conduct. This type of a-g-c circuit where the a-g-c rectifier is keyed on to conduct only during the horizontal synchronizing pulses is generally called *keyed automatic gain control* or *gated automatic gain control*. In circuits that have separate a-g-c rectifier and a-g-c amplifier tubes, the

amplifier can be pulsed on and off, while the rectifier produces the d-c control voltage independent of the keying pulses.

A typical keyed a-g-c circuit is shown in Fig. 15-6. The 6AU6 pentode  $V_{307}$  serves as both an a-g-c rectifier and amplifier. Composite video signal, with the d-c component, is directly coupled from the video output circuit to the control grid of the a-g-c tube. At the same time, pulses from the horizontal output circuit drive the plate of the 6AU6 positive during the horizontal sync pulse time. With no video signal input, the a-g-c tube is biased to cutoff by the negative grid-bias voltage between the control grid and cathode. When composite video signal is coupled to the grid of the 6AU6 the positive sync can make the a-g-c tube conduct,

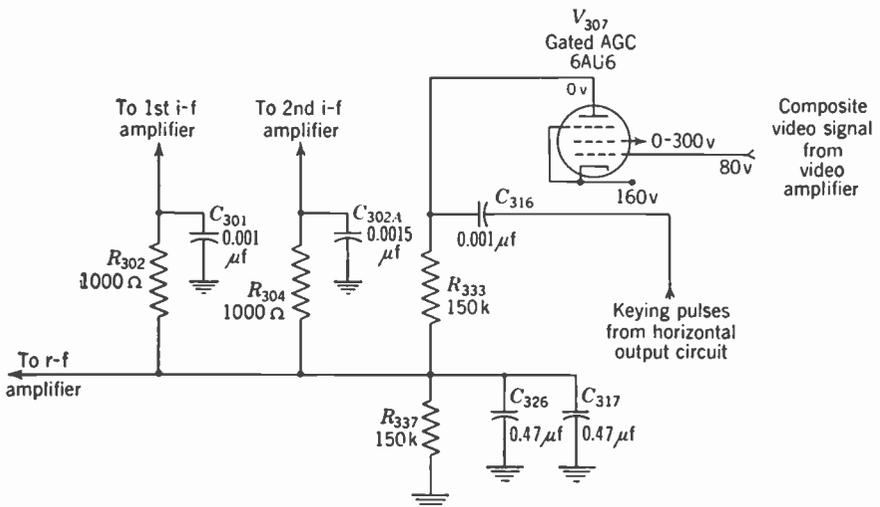


FIG. 15-6. Keyed a-g-c circuit. Voltages indicated are positive d-c volts. A-g-c voltage at plate is zero with no signal input. (Admiral Series 23 chassis.)

as the plate voltage is pulsed positive during this time. The grid bias on the a-g-c tube is negative enough to allow only the synchronizing pulses to produce plate current. Therefore, the rectified output in the plate circuit of the a-g-c tube is proportional to the peak amplitude of the composite video signal with its correct d-c component, which indicates the picture carrier-signal strength.

The plate circuit of the 6AU6 must produce negative voltage for the a-g-c bias line. Without any negative supply voltage available for the a-g-c tube, the negative voltage in the plate circuit is produced by charging  $C_{316}$ , with the plate side negative, when plate current flows during the keying pulses, and allowing it to discharge through  $R_{333}$  and  $R_{337}$  while the a-g-c tube is not conducting between horizontal sync pulses.  $R_{333}$  isolates the a-g-c line from the horizontal output circuit supplying the keying

pulses. With  $C_{326}$  and  $C_{317}$ ,  $R_{333}$  forms the a-g-c filter. The a-g-c voltage on the bias line controls the gain of the first and second i-f amplifiers and the r-f stage. The  $R_{304}C_{302A}$  combination is a decoupling filter for the grid of the second i-f amplifier, while the  $R_{302}C_{301}$  filter decouples the grid of the first i-f amplifier from the a-g-c line.

**15-5. A-G-C Level Adjustment.** The bias supplied by the a-g-c circuit can be made adjustable to provide the amount of gain control required for different signal strengths in different locations. In strong signal areas, more a-g-c bias is needed to prevent excessive signal from overdriving an amplifier stage, which results in a reversed, dark picture, out of sync. In weak signal areas, less a-g-c bias, or no automatic gain control, is desirable to allow more gain in the i-f and r-f stages. Figure 15-7 illustrates three circuit arrangements for adjusting the a-g-c level of the receiver. In *a*, only part of the filtered a-g-c voltage is coupled to the a-g-c bias line, as

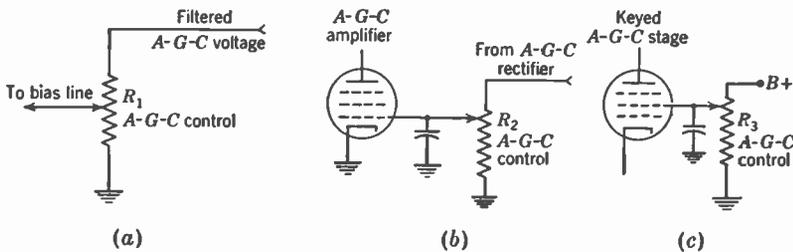


FIG. 15-7. Circuit arrangements for a-g-c level control. (a) Adjusting the bias voltage. (b) Adjusting the input voltage to the a-g-c amplifier. (c) Adjusting the screen-grid voltage of the keyed a-g-c tube.

determined by the setting of the potentiometer  $R_1$ . The a-g-c level control in *b* varies the amount of a-g-c voltage applied to the a-g-c amplifier from the a-g-c rectifier. In *c*, the screen-grid voltage of the keyed a-g-c stage is adjusted by  $R_3$ , which varies the plate current and the amount of a-g-c bias voltage available in the plate circuit.

The a-g-c control is usually mounted on the rear apron of the chassis as a setup adjustment. It may be a variable control or switch marked for local, suburban, and fringe areas. Normally, the a-g-c level can be set by adjusting the control for less a-g-c bias until the top of the picture begins to bend; then back off the a-g-c control enough to remove the bend. This should be done for the strongest station, with the manual contrast control at maximum. Otherwise the picture may bend on other stations, or when the contrast control is turned up. The a-g-c control can be set more exactly, if desired, by connecting an oscilloscope to the video detector output circuit to view the composite video signal waveform. Adjust the a-g-c level until the tips of the sync pulses start to compress, then back off the control just enough to remove the compression. When the

signal is not strong enough to produce bend in the picture, the control can be adjusted by reducing the a-g-c bias to the point where the snow in the picture becomes more obvious and then back off the control a little. In weak signal areas, it is generally necessary to reduce the a-g-c bias to minimum, or short the a-g-c bias completely, for maximum gain in the receiver.

**15-6. Troubles in the A-G-C Circuit.** A-G-C troubles affect the contrast or intensity of the picture, while the raster is normal, because the stages controlled by the a-g-c circuit amplify the picture signal. Too

TABLE 15-1. A-G-C TROUBLES

Trouble	Effect	Cause
Zero a-g-c bias . . .	Overloaded picture and buzz in sound	Short in a-g-c bias line, or no conduction in a-g-c amplifier, or no conduction in a-g-c-rectifier
Excessive negative a-g-c bias	No picture and no sound	Excessive conduction in a-g-c amplifier; can be caused by no conduction in video amplifier directly coupled to a-g-c tube

much a-g-c bias reduces the receiver gain, causing a weak picture, or no picture on a blank raster if the bias is negative enough to cut off the picture signal amplifiers. When a stage cut off by excessive bias is common to the picture and sound signals, this results in both no picture and no sound. Insufficient a-g-c bias allows excessive gain, which may result in excessive contrast that cannot be reduced with the contrast control, or an overloaded picture that has reversed black-and-white values and is out of sync. Figure 15-8 shows an overloaded picture. Also, insufficient a-g-c bias can increase the volume, with 60-cps sync buzz in the sound caused by cross modulation of the picture and sound signals in an overloaded amplifier stage common to the picture and sound. These a-g-c troubles are summarized in Table 15-1, for an intercarrier-sound receiver.

When the a-g-c circuit is directly coupled to the video amplifier, as in the keyed a-g-c arrangement in Fig. 15-6, the absence of plate current in the video amplifier results in excessive negative a-g-c bias voltage. Figure 15-9 illustrates how the a-g-c tube depends upon conduction in the video amplifier for the correct operating voltages, because of the direct coupling. When the video amplifier conducts its normal plate current of 5 ma, as an example, the plate voltage shown is 125 volts; this is 25 volts less than the plate-supply voltage of 150 volts because of the  $IR$  drop across the 5,000-ohm plate load resistor  $R_L$ . The video amplifier's plate voltage is the control-grid voltage for the a-g-c tube, as the direct coupling is used to obtain video signal with the correct d-c component. There is no  $IR$  voltage drop across the decoupling resistor  $R_1$ .

To provide correct bias for the a-g-c tube, its cathode returns to +130 volts, making the control grid 5 volts negative with respect to cathode. The 5-volt negative bias on the control grid is assumed to allow the amount of plate current that produces -2 volts across the a-g-c tube's plate load resistor  $R_2$ . However, if no plate current flows in the video

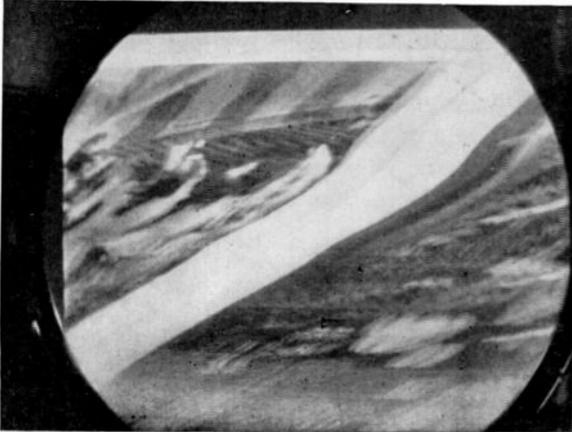


FIG. 15-8. Overloaded picture. Black and white are reversed, and picture is out of sync. (RCA Institutes, Inc.)

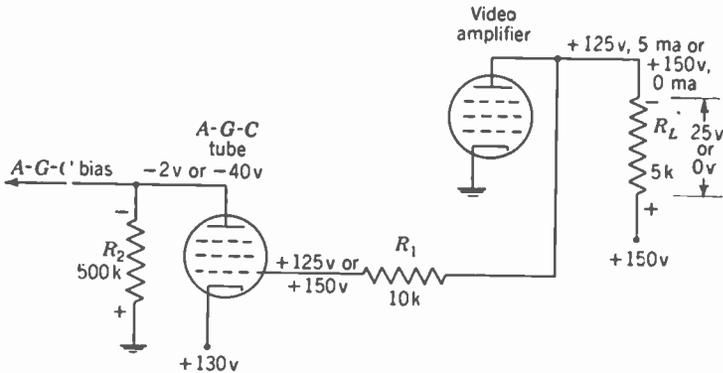


FIG. 15-9. Illustrating how the control-grid voltage of the a-g-c tube depends upon plate current in the video amplifier. Voltages above line are normal; voltages below line are with zero plate current in video amplifier.

amplifier, its plate voltage rises to the 150-volt value of the plate-supply voltage, as there is no voltage drop across  $R_L$ . Then the control-grid voltage of the a-g-c tube becomes 25 volts more positive, from +125 to +150 volts, producing more conduction. As a result, the increased plate current produces a larger negative bias voltage across the a-g-c plate load resistor, resulting in the excessive negative a-g-c bias voltage of -40 volts. Conversely, if the grid-cathode voltage on the a-g-c tube is too

negative and cuts off the plate current, it does not conduct and the a-g-c bias voltage in the plate circuit becomes zero.

*Localizing to the A-G-C Circuit.* Since the symptoms produced by a-g-c troubles are similar to troubles in the signal amplifiers, it is helpful to localize the trouble to the a-g-c circuit. Taking an example of no picture and no sound, the trouble can be localized to the picture a-g-c circuit by reducing the a-g-c bias to zero. This can be done by shorting the a-g-c line directly to chassis ground. If the sound is heard and an overloaded picture results with zero bias, but there is no picture and no sound when the a-g-c bias is on the amplifiers, the trouble must be excessive bias produced by the a-g-c circuit. This trouble can occur in amplified a-g-c circuits. With only an a-g-c rectifier operating on the i-f signal, though, it

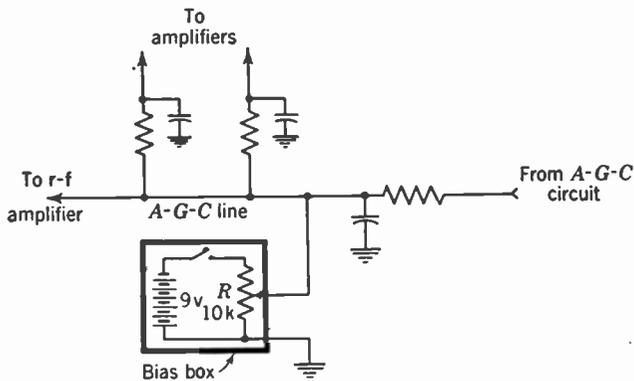


FIG. 15-10. Bias box connected to a-g-c line.

cannot produce enough a-g-c bias to cut off the stages amplifying the signal because without any signal input there will be no a-g-c bias.

A more general method of localizing troubles in the a-g-c circuit requires the use of a *battery bias box*. As shown in Fig. 15-10, the bias box provides an adjustable negative voltage that can be connected to the a-g-c line to take the place of the a-g-c bias. Connect the negative output lead of the bias box directly to the a-g-c line and the positive lead to chassis ground. If the picture and sound are normal when the bias box supplies the amplifier bias, but not when the a-g-c bias is functioning, the trouble must be in the a-g-c circuit.

#### REVIEW QUESTIONS

1. Explain briefly how an a-g-c circuit operates, giving the function of the a-g-c rectifier, the a-g-c filter, and the a-g-c line.
2. How does more negative bias for the grid of an amplifier affect the gain of the stage?
3. What is the effect of an increased carrier level on the amount of a-g-c bias and the amplification in the receiver?
4. Give two advantages of automatic gain control for the picture signal amplifiers.

5. What is the important difference between an a-g-c circuit for the picture signal and the a-g-c circuit for an AM sound signal?
6. What is the advantage of using the signal from the video amplifier to produce the a-g-c voltage, compared with the picture signal at the second detector?
7. Why is it necessary that the composite video signal for the a-g-c circuit have the correct d-c component?
8. Give one advantage and one disadvantage of a slow a-g-c circuit compared with a fast a-g-c circuit.
9. Explain how it is possible to have 60-cps sync voltage on the a-g-c line. Give two possible trouble symptoms that can result.
10. Describe briefly what a keyed a-g-c circuit is. What advantage does it have compared with picture a-g-c circuits that are not keyed?
11. Describe the construction of a bias box and how it is connected to the receiver to substitute for the a-g-c bias.
12. When the a-g-c bias line is shorted to ground, what is the effect on the amount of bias and the gain of the controlled stages?
13. Referring to the a-g-c circuit in Fig. 15-3, give the function of  $R_{314}$ ,  $R_{313}$ ,  $C_{306}$ ,  $R_{302}$ , and  $C_{301A}$ .
14. Referring to the a-g-c circuit in Fig. 15-6, give the function of  $C_{316}$ ,  $R_{322}$  with  $R_{327}$ ,  $C_{326}$  with  $C_{317}$ ,  $R_{304}$ , and  $C_{302A}$ .
15. Where would a d-c voltmeter be connected to measure the a-g-c bias produced in the circuit in Fig. 15-3? In Fig. 15-6?
16. Describe briefly how to adjust the a-g-c level control, for the typical case of average signal strength on all stations.

## CHAPTER 16

### SYNC SEPARATION

The synchronizing pulses are included in the composite video signal transmitted to the receiver in order to time the scanning with respect to the camera signal variations, so that the picture information reproduced on the kinescope screen can be in the correct position on the raster. At the broadcast station, the synchronizing signal generator produces pulses to time the scanning in the camera tube; the same generator supplies the synchronizing pulses that are added to the video signal transmitted for the receiver. These synchronizing pulses, generally called *sync*, are separated by the sync circuits in the receiver and coupled to the deflection circuits in order to control the timing of the scanning for the picture tube. As a result, the picture information reproduced on the screen of the picture tube is in the same relative position as on the image plate of the camera tube, since the scanning for both tubes is synchronized by one common source—the sync generator at the broadcast station. The amount of time for the transmitted signal to travel to the receiver has no effect on synchronization because the synchronizing pulses must be present at the same time as the camera signal variations in the composite video signal at the receiver.

The sync separation circuits in the receiver provide vertical synchronizing signals to time the vertical scanning frequency so that every field is timed correctly, and horizontal synchronizing signals to time the scanning lines. The sync is able to time the vertical and horizontal scanning by controlling the frequency of the vertical and horizontal deflection oscillators. It is important to remember that the deflection circuits in the receiver can produce vertical and horizontal scanning to form the raster with or without sync, but the position where the picture information is reproduced on the raster depends upon the vertical and horizontal synchronization.

**16-1. Vertical Synchronization of the Picture.** The vertical synchronizing signals at the rate of 60 per second for each scanning field provide vertical synchronization. When every field is produced at the correct time, the frames are superimposed on each other to provide a steady

picture locked in frame on the kinescope screen. If the vertical scanning is not locked in at the 60-cps vertical synchronizing frequency, successive frames cannot overlap but instead will be continuously displaced either above or below the first frame. Without vertical synchronization, therefore, the picture on the kinescope screen appears to roll up or down.

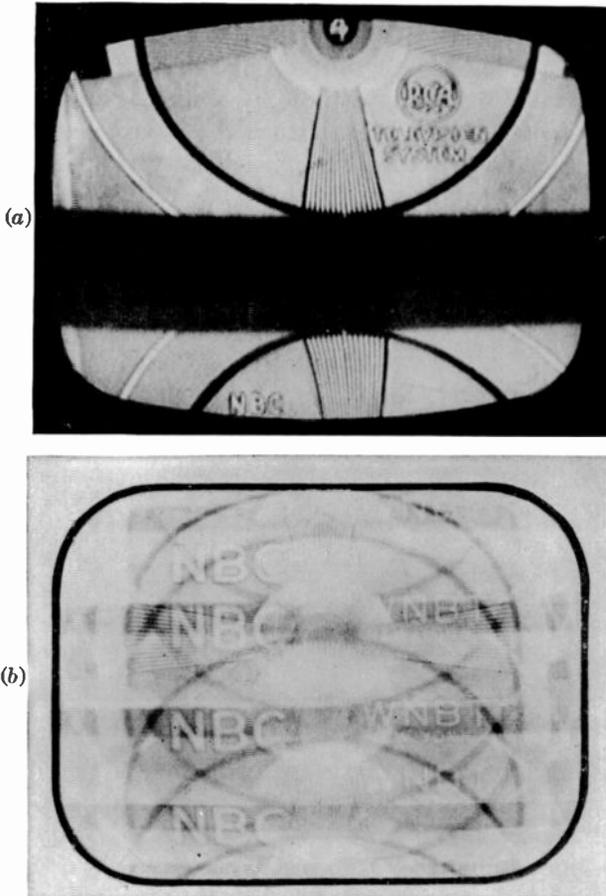


FIG. 16-1. Vertical scanning not synchronized, resulting in no vertical hold. (a) Picture slowly slips frames vertically. (b) Picture rolls fast vertically.

The faster the picture rolls the farther the vertical scanning frequency is from 60 cps. If the vertical scanning frequency is just slightly off the 60-cps synchronizing rate the picture will be recognizable, as shown in Fig. 16-1a, but it slips out of frame slowly and continuously. In b the picture is rolling fast. The wide black bar across the picture in Fig. 16-1a is produced by vertical blanking, which now occurs during vertical trace time because the scanning is out of sync. In some receivers white retrace

lines appear while the picture is rolling. These are horizontal scanning lines produced during the vertical flyback. They are visible because vertical retrace is not occurring during vertical blanking time, without synchronization. When the picture is in vertical sync, though, it stays still, locked in frame vertically, with the black vertical blanking at the top and bottom edges, and the retrace lines are blanked out.

**16-2. Horizontal Synchronization of the Picture.** The horizontal synchronizing signals at the rate of 15,750 per second for each scanning line provide horizontal synchronization. When every scanning line is produced at the correct time, the line structure of the reproduced picture holds together to provide a complete image that stays still horizontally. If the horizontal scanning is just slightly off the 15,750-cps synchronizing

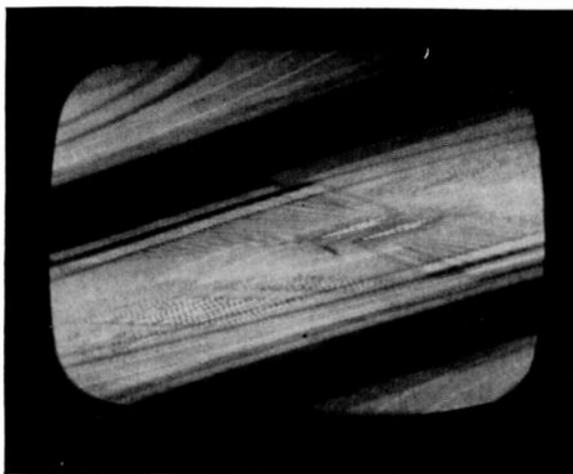


FIG. 16-2. Horizontal scanning not synchronized, resulting in no horizontal hold.

frequency rate, the line structure is complete but the picture slips horizontally, as the picture information on the lines is displaced horizontally in successive frames. The faster the picture slides horizontally the farther the scanning frequency is from 15,750 cps. When the horizontal scanning frequency departs from the 15,750-cps synchronizing frequency by 60 cycles or more, the picture tears apart into diagonal segments, as shown in Fig. 16-2. The black diagonal bars represent parts of horizontal blanking, which is normally at the sides of the picture. The more bars, the farther the horizontal oscillator is from the horizontal sync frequency of 15,750 cps.

**16-3. Separating the Sync.** In order to obtain the synchronizing signals needed for timing the scanning correctly the sync pulses must be separated from the composite video signal. Figure 16-3 shows oscillograms of the composite video signal input to a sync separator stage and

the separated sync output. This amplitude separation of the synchronizing pulses from the composite video signal can be accomplished by any one of several arrangements of a clipper stage. All are fundamentally the same in that the tube is biased beyond cutoff by an amount great enough to allow plate-current flow and output signal only for the most positive swing of the input signal. Since the synchronizing pulses have the greatest amplitude in the composite video signal, the output of the clipper stage can be made to correspond only to the synchronizing pulses of the input.

*Grid-leak Bias Sync Separator.* Figure 16-4 illustrates the operation of a typical sync separator circuit using grid-leak bias. The grid input voltage is composite video signal of positive sync polarity, so that the

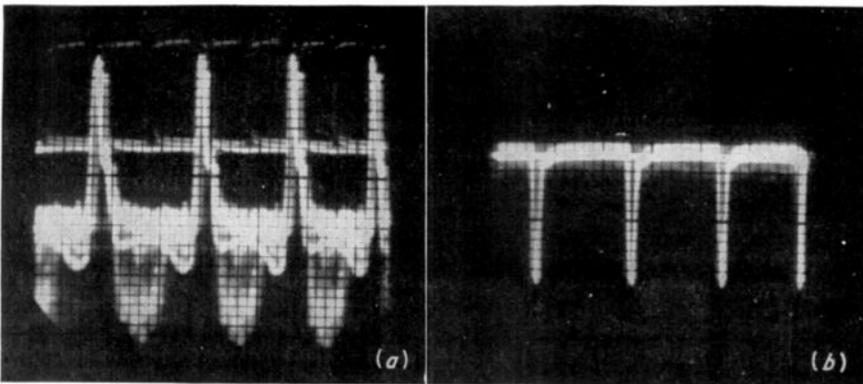


FIG. 16-3. Oscillograms of the composite video signal and separated sync. Only horizontal sync is evident with the oscilloscope internal sweep frequency at 15,750/4 cps, but vertical sync and equalizing pulses are also in the signal. (a) Video signal with positive sync polarity. (b) Separated sync with negative polarity. (RCA.)

synchronizing pulses can drive the instantaneous grid voltage positive to produce grid current. Since the negative grid-leak bias voltage produced by grid current automatically adjusts itself to the value that allows just the peaks of the input signal to drive the grid voltage slightly positive, the tips of the synchronizing pulses are clamped at approximately zero grid voltage, exactly like a grid-leak bias d-c restorer circuit. With the synchronizing pulses in line at a constant voltage level in the grid circuit, it is now necessary only to reduce the grid cutoff voltage of the amplifier to a value corresponding to the pedestal voltage on the grid, or less. By reducing the plate voltage applied to the amplifier, and the screen voltage for a pentode, the amount of negative grid voltage necessary to cut off the flow of plate current is reduced to the value required. Then plate current flows only during the time corresponding to the synchronizing pulses of the composite video signal input, producing separated sync output.

In Fig. 16-4, the amount of grid-leak bias shown is  $-4$  volts, approximately equal to the peak positive swing of the input voltage. Since the grid cutoff voltage is  $-2$  volts, the input signal must drive the instantaneous grid voltage at least  $2$  volts more positive than the negative bias to produce plate current. Only the synchronizing pulses in the input signal are positive enough to drive the instantaneous grid voltage to a value less negative than cutoff, allowing plate current to flow. Therefore, the camera signal variations of the input are missing from the output because they do not produce plate current, and the output signal voltage contains only the synchronizing pulses. It should be noted that the composite video signal input can be clipped at any amplitude from the pedestal level to the tip of sync and still provide the desired separated sync output

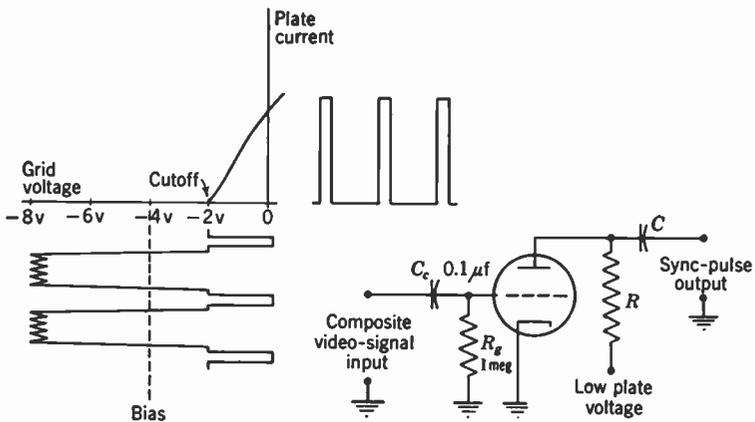


FIG. 16-4. Operation of sync separator using a grid-leak bias clipper amplifier. A pentode can be used instead of the triode.

because plate current would flow only during sync pulse time. Also, more negative values of grid cutoff voltage can be used, as long as the input signal has enough amplitude to allow just the sync pulse voltage to drive past cutoff and produce plate current.

*Types of Clipper Stages.* Several arrangements of a clipper stage to separate the sync are illustrated in Fig. 16-5. In all cases, the tube has self-bias produced by the input signal, so that the synchronizing pulses can be lined up for a constant clipping level, in order to provide sync output free of interference from the camera signal. The input for the sync separator can be modulated picture carrier signal from the i-f amplifier, or composite video signal from either the video detector or video amplifier, since they all contain the desired synchronizing pulse voltage. Usually, though, the input to the sync separator is the amplified composite video signal from the video amplifier because this arrangement has the advantage of using the video amplification in the receiver for the sync

voltage. A diode can be used for the sync separator but a triode or pentode is more common because it amplifies the clipped sync. The polarity of the composite video signal input to the sync separator must allow the synchronizing pulses to drive the grid voltage in the positive

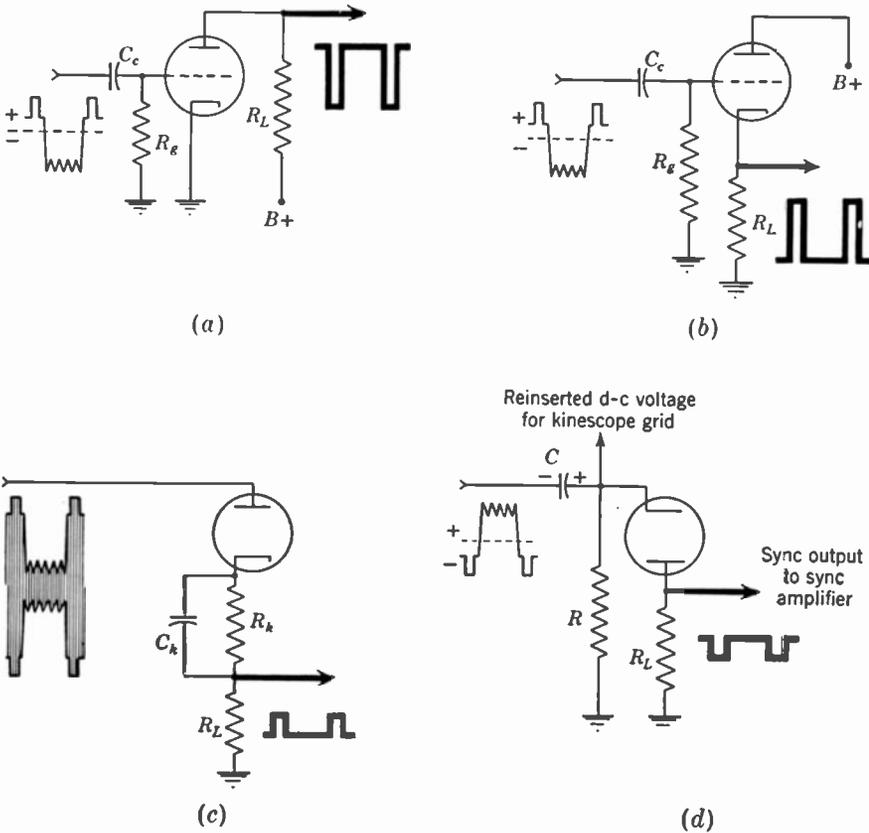


FIG. 16-5. Sync separator circuits. All use self-bias to allow plate-current flow only for the synchronizing pulses. Pentodes can be used in place of the triodes. (a) Grid-leak bias separator with negative sync output across plate load resistor. (b) Grid-leak bias separator with positive sync output across cathode load resistor. (c) Diode separator with modulated i-f signal input. (d) Inverted diode operating as combined d-c restorer and sync separator.

direction, or the cathode negative, to produce plate current. If the opposite polarity is used, the maximum white camera signal will be clipped instead of the sync.

Referring to Fig. 16-5a, this shows a triode clipper with grid-leak bias, produced by the composite video signal input of positive sync polarity. Note that the separated sync in the plate circuit is amplified and inverted, since the stage is a resistance-coupled amplifier for the sync pulses. The

circuit in *b* is also a grid-leak bias clipper but the separated sync voltage output is taken across the cathode resistor  $R_L$ . This cathode voltage is the separated sync because cathode current flows only for the sync pulses. The sync output voltage from the cathode, however, has the same polarity as the sync input at the grid. The diode clipper circuit in *c* uses cathode bias to permit plate-current flow only for the peak voltage swing of the modulated picture signal input, which is produced by the synchronizing pulses. There is no sync pulse voltage across  $R_k$  because it is bypassed by  $C_k$  to provide a steady bias, but the diode current flowing during sync pulse time produces separated sync voltage across  $R_L$ .

A circuit that combines the functions of the sync separator and a positive d-c restorer is illustrated by the inverted diode in Fig. 16-5*d*. The composite video signal input produces self-bias for the diode by means of the  $RC$  coupling circuit, as in grid-leak bias. The sync input voltage drives the cathode negative, producing diode current. The diode is inverted in order to provide reinserted d-c voltage of positive polarity. As a result, the pulses of diode current flowing through the load resistor  $R_L$  produce separated sync voltage output, while the d-c voltage across  $C$  can be directly coupled to the kinescope grid for d-c restoration. It should be noted that the grid-cathode circuit of a triode clipper can also be used as an inverted diode by coupling the composite video signal to the cathode instead of the grid. When this is done, the plate circuit can produce separated sync output, while the positive d-c voltage reinserted in the cathode is directly coupled to the kinescope grid for d-c restoration.

*Amplitude Separation.* Clipping the synchronizing pulses from the composite video signal is amplitude separation of the sync, and the stage having this function is a *sync separator*. A stage for additional clipping of the separated sync voltage is generally called a *sync clipper*.

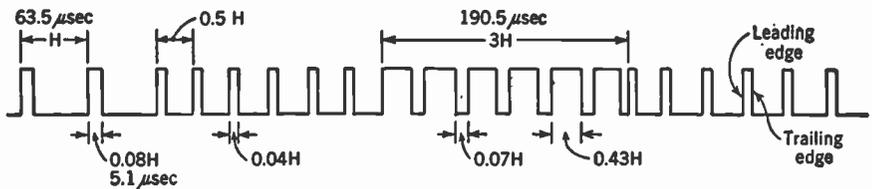


FIG. 16-6. Waveform of the synchronizing pulses.  $H$  is the horizontal line period of  $63.5 \mu\text{sec}$ .

*Waveform Separation.* In addition to the amplitude separation of the sync voltage, there still remains the problem of filtering out from the total separated sync voltage the vertical synchronizing pulses alone, for synchronization of the vertical scanning, and providing for horizontal synchronization. The total separated sync voltage includes the horizontal, vertical, and equalizing pulses, as illustrated in Fig. 16-6. All have the

same amplitude but they differ in frequency and pulse width. The repetition rate of the horizontal pulses is 15,750 per second, one for every scanning line, and their pulse width is  $5.1 \mu\text{sec}$ , providing narrow pulses with a frequency of 15,750 cps. The serrated vertical pulse is repeated at the rate of 60 per second, one for each scanning field, with a pulse width of  $190.5 \mu\text{sec}$ , providing wide pulses at the frequency of 60 cps. Since the horizontal and vertical synchronizing pulses differ in frequency and pulse width, they have different waveforms that can be separated from the total sync voltage by either  $RC$  or  $RL$  circuits.

**16-4. RC Transients.** The voltage or current in  $RC$  and  $RL$  circuits immediately following a sudden change of applied voltage is called the *transient response*. The transient response of  $RC$  circuits is analyzed here now because this has many important applications, including waveform separation of the synchronizing signals from the total sync voltage and generation of the saw-tooth voltage waveform for deflection. A few examples are given here to show a condenser charging or discharging through a series resistance. The arrangement of a battery and switch used to illustrate the charge or discharge circuit is equivalent to a square-wave pulse of applied voltage or rectangular pulses such as the synchronizing pulses. The leading edge of the pulse means that the maximum amplitude of signal voltage is instantaneously applied. The voltage is maintained for the time corresponding to the width of the pulse. This corresponds to the closing of the switch in the simplified circuit and keeping it closed for the duration of the pulse width. The trailing edge of the pulse is equivalent to removal of the applied voltage.

*Condenser Charge.* Figure 16-7a shows a circuit for charging the condenser  $C$  through a series resistance  $R$  by means of the 100-volt battery. When the switch is closed, current flows in the direction shown to charge the condenser. Since the voltage across the condenser is proportional to its charge, a small voltage  $e_c$  appears across the condenser. The charging process continues until the condenser is fully charged and the potential difference across  $C$  is equal to the battery voltage. With the voltage drop across the condenser equal to the applied battery voltage, the charging current drops to zero. While it might seem that the battery could not cause current flow in this capacitive circuit often used to block direct current, it must be remembered that the applied voltage is continuously changing during the transient response from the time the switch is closed until the condenser is completely charged. After the steady-state condition is reached, with the condenser charged to the same potential difference as the applied voltage, no current can flow.

Initially, when the switch is closed the amount of current flowing is maximum, since at that time there is no voltage drop across the condenser to oppose the applied voltage. Therefore, the condenser charges most

rapidly at the beginning of the charging period. As the condenser becomes charged, the voltage opposing the battery becomes greater and the net applied voltage for driving current through the circuit becomes smaller. Plotting the condenser voltage against time as it charges, the  $RC$  charge curve of Fig. 16-7a is obtained. The condenser voltage

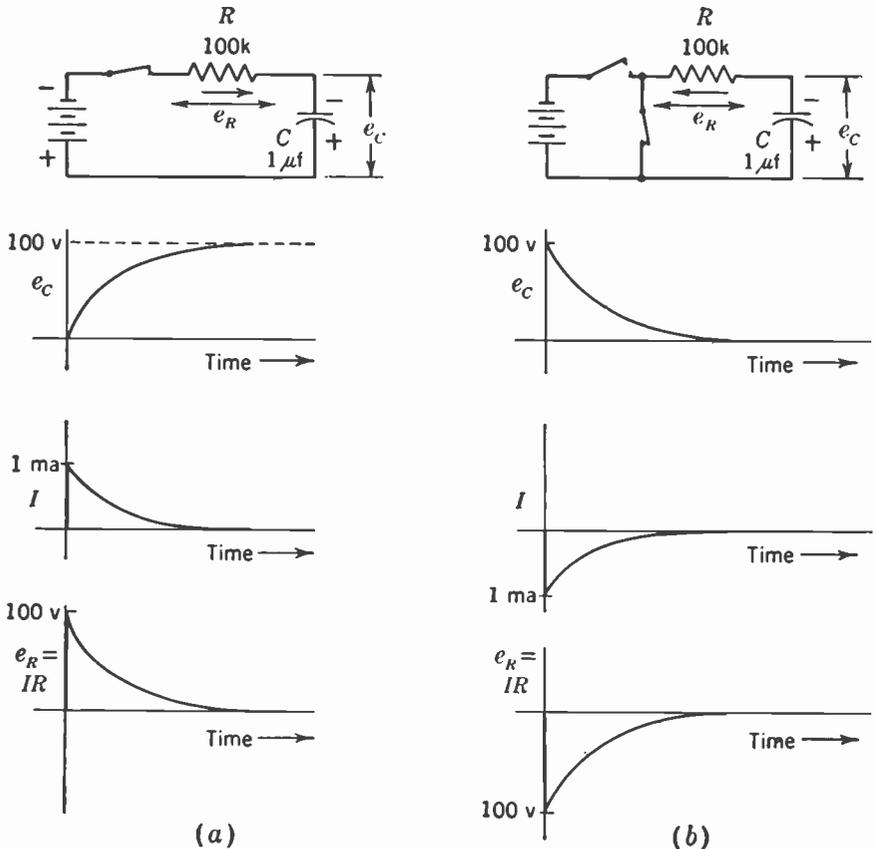


FIG. 16-7. Charge and discharge of an  $RC$  circuit. (a) Charge. (b) Discharge.

increases in the manner shown by the exponential curve because the charging current decreases as the amount of charge on the condenser increases. The equation for this curve is

$$e_c = E(1 - e^{-t/RC})$$

where  $e_c$  is the amount of voltage added to the condenser at any instant of time because of the net applied charging voltage  $E$ . Epsilon,  $\epsilon$ , is the base in the Napierian or natural system of logarithms and is a constant

equal to 2.718. The exponent  $t/RC$  includes the time in seconds, resistance in ohms, and capacitance in farads.

The current that flows in the circuit to charge the condenser has its maximum value, equal to  $E/R$ , at the first instant of charging. As the condenser takes on more charge, the net applied potential available for charging the condenser decreases because the condenser voltage opposes the battery voltage, and the current flow decreases in the manner shown in Fig. 16-7a. The current reaches its maximum value instantaneously after the application of the charging voltage and then decreases exponentially, decreasing most rapidly at first when the condenser is charging most rapidly, then declining more slowly to reach zero when the condenser is completely charged. Theoretically, the condenser never becomes completely charged and the charging current is never reduced to zero, but the condenser can become more than 99 per cent charged within a definite amount of time.

The voltage across the resistor is always equal to  $IR$  and has the same wave shape as the current because the resistance is constant, making the voltage vary directly with the current. When the condenser is completely charged the value of the current is zero and the voltage across the resistor is also zero. At any instant the voltage drops around the circuit must equal the applied voltage, and the sum of the resistor voltage and condenser voltage is equal to the battery voltage at all times.

*Condenser Discharge.* Figure 16-7b shows a circuit for discharging the condenser  $C$  through the resistor  $R$ . Let the voltage across the condenser be equal to 100 volts at the instant the switch is closed for discharging the condenser. As electrons from the negative plate of the condenser flow around the circuit to the positive plate, the condenser loses its charge and the voltage across the condenser  $e_c$  decreases exponentially as shown in Fig. 16-7b. The voltage decreases most rapidly at the beginning of the discharge because the condenser voltage, now acting as the applied voltage, then has its highest value and can drive maximum discharge current around the circuit. The magnitude of the condenser voltage  $e_c$  at any instant during the discharge is given by the equation

$$e_c = E_c(\epsilon^{-t/RC})$$

where  $e_c$  is the instantaneous value of the voltage across the capacitor on discharge, and  $E_c$  is the initial condenser voltage at the beginning of discharge. The factor  $\epsilon^{-t/RC}$  is the same as in the  $RC$  charge formula.

The discharge current has its maximum value of  $E_c/R$  at the first instant of discharge and, as shown in the diagram, has direction opposite to the current flow on charge because the condenser is now acting as the generator and is connected on the opposite side of the series resistance  $R$ .

As the condenser discharges, the discharge current decreases in magnitude because of the declining value of  $e_c$ , to produce the wave shape shown in Fig. 16-7b. After the condenser is completely discharged the current is zero. The voltage across the resistor  $R$  is equal to  $IR$  and has the same wave shape as the current.

*Time Constant.* It takes the condenser a definite amount of time to charge or discharge in the  $RC$  circuit. The series resistance limits the amount of current flow, larger values of resistance resulting in a longer time required for charge or discharge. Larger values of capacitance also require a longer time for charge or discharge. A convenient measure of the charge or discharge time of the circuit, therefore, is the  $RC$  product. The time constant of the circuit for charge or discharge is defined as

$$R \times C = t$$

where  $C$  = capacitance in farads,  $R$  = resistance in ohms, in series with the charge or discharge current, and  $t$  = time constant in seconds. Useful relations often used in calculating  $RC$  time constants are

$$\begin{aligned} R \text{ (ohms)} \times C \text{ (farads)} &= t \text{ sec} \\ R \text{ (megohms)} \times C \text{ (\mu f)} &= t \text{ sec} \\ R \text{ (ohms)} \times C \text{ (\mu f)} &= t \text{ \mu sec} \\ R \text{ (megohms)} \times C \text{ (\mu\mu f)} &= t \text{ \mu sec} \end{aligned}$$

The value of resistance used in calculating the time constant must be the resistance in series with the condenser charging current for a charge circuit or the series resistance on discharge. The charge and discharge paths for the condenser are not necessarily the same, and if the series resistance is different, the time constant will not be the same on charge as for discharge.

The time constant of the circuit states the time required for the condenser to charge to 63 per cent of the applied voltage, on charge. For discharge, the time constant states the time required for the condenser to discharge 63 per cent, or the time required for the condenser to discharge to 37 per cent of its original voltage. These values are obtained from the equations given previously for the instantaneous voltage across the condenser at any instant of time as it charges or discharges. On charge, the condenser voltage is given by

$$e_c = E(1 - e^{-t/RC})$$

The voltage after  $RC$  time can be found by substituting  $RC$  for time in the equation to obtain

$$\begin{aligned}
 e_c &= E(1 - e^{-RC/RC}) = E(1 - e^{-1}) = E\left(1 - \frac{1}{e}\right) \\
 &= E\left(1 - \frac{1}{2.718}\right), \text{ since } e \text{ is equal to } 2.718 \\
 &= E(1 - 0.37) \text{ (approx)} \\
 &= 0.63E
 \end{aligned}$$

The condenser voltage at any instant of time as it declines because of discharge is given by the equation

$$e_c = E_c(e^{-t/RC})$$

Substituting  $RC$  for time, the condenser voltage after one time constant can be found.

$$\begin{aligned}
 e_c &= E_c(e^{-RC/RC}) = E_c(e^{-1}) \\
 &= E_c\left(\frac{1}{2.718}\right) \\
 &= 0.37 E_c \text{ (approx)}
 \end{aligned}$$

The  $RC$  time constant is a convenient unit of measurement for the  $RC$  circuit, defining the amount of time required for the condenser to charge to 63 per cent of the applied voltage on charge, or the amount of time required for the condenser to discharge to 37 per cent of its initial voltage when the condenser is being discharged. These values hold for any  $RC$  circuit, since they are percentages that can be applied universally. The condenser is practically completely charged to the applied voltage after a time equal to five time constants, as shown in the universal time-constant chart in Fig. 16-12. On discharge, the condenser voltage is practically down to zero after five time constants. A few examples will demonstrate how the time constant can be applied.

**Example 1.** A 1-megohm resistor and 1- $\mu$ f condenser are connected in series across a 100-volt battery. If the condenser has no charge initially, how long will it take to charge to 63 volts?

The time constant of the  $RC$  circuit is 1 sec. Since 63 volts is 63 per cent of the applied voltage and the condenser charges to 63 per cent of the applied voltage in 1 time constant, it takes 1 sec for the condenser to charge to 63 volts.

**Example 2.** The same 1-megohm resistor and 1- $\mu$ f condenser are connected in series across a 45-volt battery. What is the condenser voltage after 1 sec if the condenser has no initial charge?

The time constant of the circuit is 1 sec, and the condenser will charge to 63 per cent of the applied 45 volts after 1 sec, or 28.4 volts.

**Example 3.** A 0.1- $\mu$ f condenser is charged to 10 volts and is then allowed to discharge through a 100-ohm resistor. What is the condenser voltage after 10  $\mu$ sec of discharge?

The time constant for discharge in the  $RC$  circuit is 10  $\mu$ sec. The condenser voltage discharges to 37 per cent of the original 10 volts, or 3.7 volts.

There are several precautions to keep in mind when figuring  $RC$  problems. The time constant gives the rate of charge but it does not determine for how long the condenser charges, nor does the time constant decide whether or not the condenser is able to charge. When the applied voltage is more than the voltage across the condenser, it will charge. Furthermore, the condenser will continue to charge as long as the applied voltage is maintained and is greater than the condenser voltage. When the condenser voltage equals the applied voltage the condenser cannot take on any more charge. Or, if the applied voltage is removed, the condenser cannot charge. Suppose that in an  $RC$  circuit with a time constant of 1 sec, 100 volts is applied for  $\frac{1}{2}$  sec. In  $RC$  time, the condenser could charge to 63 volts if the voltage were applied for 1 sec. However, the applied voltage is removed after  $\frac{1}{2}$  sec, allowing the condenser to charge only for one-half of  $RC$  time to produce approximately 40 per cent of the applied voltage, or 40 volts, across the condenser. This value is obtained from the universal time-constant chart in Fig. 16-12, which gives the per cent of charge or discharge for different amounts of time in terms of the time constant. Similarly, the time constant on discharge gives the rate of discharge but it does not determine for how long the condenser discharges nor does the time constant decide whether or not the condenser is able to discharge. The condenser will discharge as long as the voltage across it is greater than the applied voltage. When the condenser voltage discharges down to zero it cannot discharge any more. Or if the condenser is discharging and the applied voltage changes to become greater than the condenser voltage, the condenser will stop discharging and start charging. Finally, it should be noted that the 63 per cent change of voltage in  $RC$  time refers to 63 per cent of the net voltage available for producing charging current on charge, or discharging current on discharge. As an example, if 100 volts is applied to charge a condenser that already has 20 volts across it, the condenser voltage will increase by 63 per cent of 80 volts, adding 50.4 volts to the 20 volts to produce 70.4 volts across the condenser after  $RC$  time.

*Charge and Discharge in  $RC$  Time.* The voltage and current wave shapes in an  $RC$  circuit are shown in Fig. 16-8 for the case where a condenser is allowed to charge through a resistance for  $RC$  time and then discharges through the same resistance for the same amount of time. After the switch is closed for charge, the condenser charges to 63 volts in the  $RC$  time of 1 sec because this is 63 per cent of the applied 100 volts. Then the condenser is allowed to discharge and the condenser voltage declines to 37 per cent of 63 volts, or 23.3 volts, in  $RC$  time. The next charge cycle begins with the condenser voltage equal to 23.3 volts, therefore. After  $RC$  time an additional 48.3 volts is added, since this is 63 per cent of the net applied charging voltage of 76.7 volts, making the con-

denser voltage equal to 71.6 volts. The current has the wave shape shown in the illustration, with its maximum value at the beginning of the charge or discharge and then declining as the condenser becomes charged or discharged. The voltage across the resistor  $e_R$  is equal to  $IR$  and has the same wave shape as the current. The negative polarity of voltage across  $R$  is obtained on discharge because the current flow is in the direction opposite from the charging current.

*Long and Short Time Constants.* The example illustrating the action in an  $RC$  circuit when a charging voltage is applied for  $RC$  time and then removed to allow the condenser to discharge for the same period of time is equivalent to having an applied square wave of voltage with a pulse width equal to  $RC$  time. The time allowed for discharge or charge in this case is neither long nor short compared with the width of the applied pulse. By using  $RC$  circuits with a long time constant or short time constant, however, various useful types of voltage wave shapes can be obtained. A long time constant is at least five times as long as the period during which voltage is applied. This does not allow the condenser to become completely charged during the time that the voltage is applied. The time constant can be called long because it is too long to

allow the condenser to become appreciably charged. A short time constant is no more than one-fifth the period of the applied voltage. This permits the voltage to be applied for a period equal to five time constants, allowing the condenser to become completely charged, as shown by the universal  $RC$  curve in Fig. 16-12. The same conditions of long or short time constants apply on discharge as well as on charge. Usually, when a long or short time constant is desired, the time constant is made greater or less than the period of the applied voltage by a factor much greater than 5 in order to obtain the desired wave shapes.

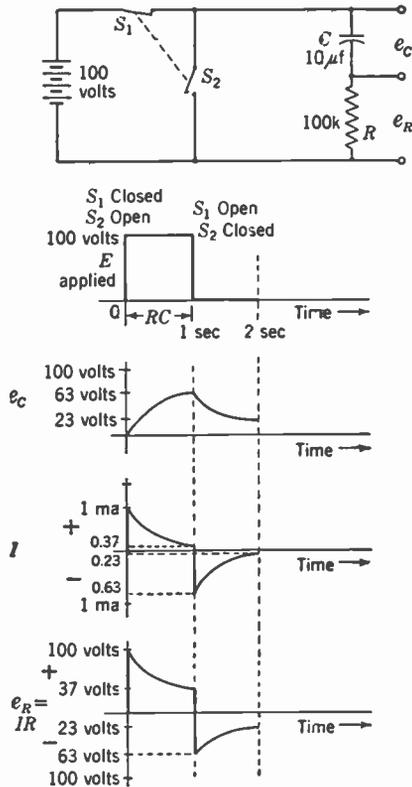


FIG. 16-8. Charge and discharge of an  $RC$  circuit in  $RC$  time.

*Charge and Discharge with a Short Time Constant.* Figure 16-9 shows the wave shapes obtained in an  $RC$  circuit when the time constant is short and is the same for charge or discharge. The voltage across the condenser is essentially the same as the applied square wave of voltage because the short time constant of the  $RC$  circuit allows the condenser to

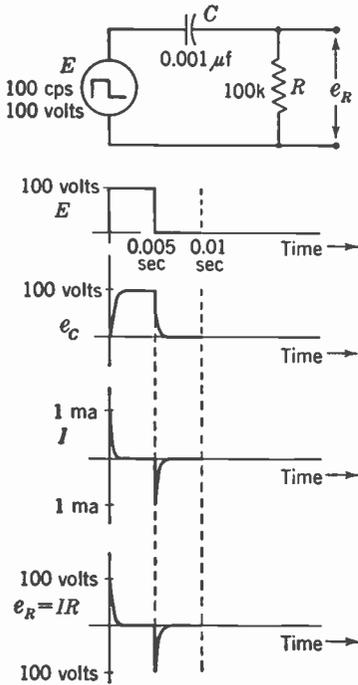


FIG. 16-9. Charge and discharge of an  $RC$  circuit with a short time constant.

charge or discharge completely within a period of time that is very much less than the width of the square-wave pulse.

Sharp peaks of current flow in the circuit because of the short time constant. The leading edge of the pulse coincides with the application of charging voltage. The current has its initial maximum value of  $E/R$  instantaneously at that time and is then rapidly reduced to zero as the condenser becomes completely charged. For most of the pulse width the condenser voltage is equal to the applied voltage and the current remains zero. The trailing edge of the applied pulse coincides with the removal of the applied voltage and the beginning of condenser discharge. The discharge current flows in the opposite direction, instantaneously reaching its maximum value at the time of discharge, and is rapidly reduced to zero as the condenser becomes completely discharged. During the remainder of the time, when the condenser voltage is zero and the applied

voltage is zero, there is no current flow. With the next pulse, peaks of current are again obtained for the leading and trailing edges of the applied square wave. The voltage across the resistor  $e_R$  is equal to  $IR$  and has the same wave shape as the current, with sharp voltage peaks for the leading and trailing edge of each input pulse.

*Long Time Constant on Charge and Short Discharge Time.* It is not necessary that the condenser charge and discharge through the same path. Figure 16-10 shows a circuit where the condenser is allowed to charge through a high resistance, providing a long time constant, and is then rapidly discharged through a low resistance. After the switch  $S_1$  is closed to apply the charging voltage the voltage across the condenser rises slowly, following the exponential charging curve for a condenser. When the applied voltage is removed by opening  $S_1$ , and the switch  $S_2$  is

closed for discharge through the low-resistance path, the condenser voltage drops rapidly to zero because of the short time constant. The result is a saw-tooth wave of voltage across the condenser. This illustrates the method generally used for producing saw-tooth deflection voltage in scanning circuits.

**16-5. RL Transients.** When voltage is applied across an inductive circuit the current flowing in the circuit cannot attain its steady-state value instantaneously but must build up gradually because of the self-induced voltage across the coil. As soon as the applied voltage produces a change in current flowing through the inductance, its magnetic field moves out to cut the turns of wire in the coil and generates a counter e.m.f. that opposes the applied voltage, thus reducing the net applied voltage available for producing current flow in the circuit. When the applied voltage is removed the coil's magnetic field collapses, again cutting the turns of

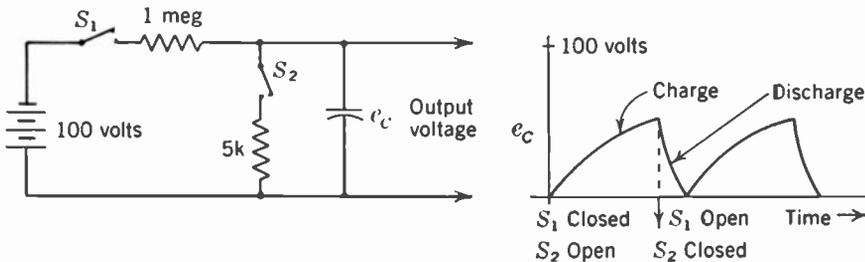


FIG. 16-10. RC circuit with a long time constant on charge and short time constant for discharge. The output across the condenser is a saw-tooth wave of voltage.

wire in the inductor and generating a self-induced e.m.f. across the coil that prolongs the flow of current in the same direction.

**RL Charge and Discharge.** Figure 16-11b shows the wave shapes obtained in an *RL* circuit with a square wave of applied voltage. When voltage is applied to the *RL* circuit the current builds up gradually toward its maximum value of  $E/R$ , following the exponential wave shape shown. The increase in current is most rapid at the beginning of the charge and then increases at a slower rate because of the induced voltage across the coil that opposes the applied voltage. The self-induced voltage across the inductance  $e_L$  reaches its maximum value instantaneously and then decreases in magnitude as the rate of change in the current decreases. When the applied voltage is removed, the collapsing magnetic field induces across the coil a voltage of polarity opposite from the voltage induced during the charge. The current flow is maintained in the same direction, however, because the reversed voltage across the inductance is now applied voltage and is connected on the opposite side of the resistance  $R$ . The voltage across the resistor  $E_r$  is equal to  $IR$  and has the same wave shape as the current.

The transient response of the  $RL$  circuit exactly parallels the  $RC$  response, but the current is now the factor in the circuit that builds up and decays exponentially, corresponding to the condenser voltage in the  $RC$  circuit, because its value cannot change instantaneously. As seen in

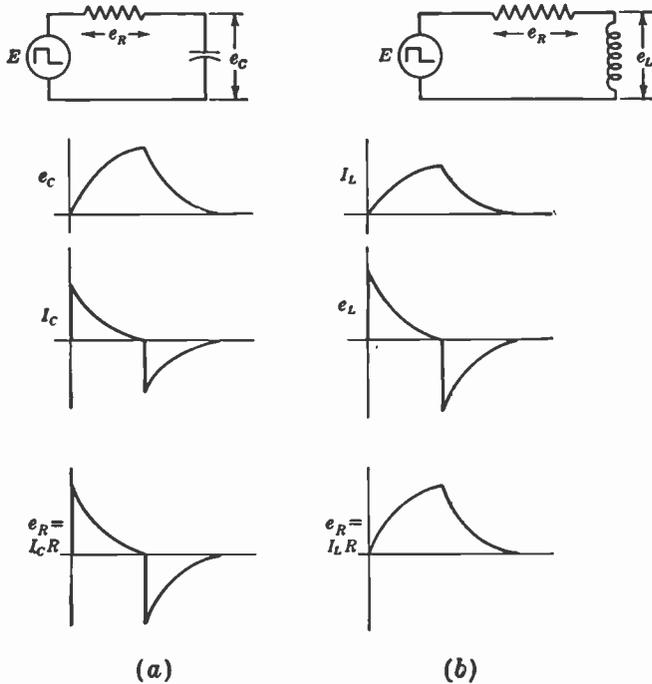


FIG. 16-11. Comparison of wave shapes in  $RC$  and  $RL$  circuits. The current in the  $RL$  circuit cannot change instantaneously but must build up or decay exponentially in the same manner as the condenser voltage in the  $RC$  circuit. (a)  $RC$  circuit. (b)  $RL$  circuit.

Fig. 16-11, the wave shapes for the  $RL$  circuit are exactly the same as in the  $RC$  circuit, with the current and voltage interchanged.

*RL Time Constant.* The time constant of an  $RL$  circuit is defined as

$$\frac{L}{R} = t$$

where  $L$  is the inductance in henrys,  $R$  the series resistance in ohms, and  $t$  the time constant in seconds. The time constant applies either to current build-up or to current decay in the circuit and has the same significance as the time constant in an  $RC$  circuit. In the inductive circuit the current rises to 63 per cent of its maximum value or decays to 37 per cent of its initial value during the period of one time constant. Since the shapes of the current and voltage wave shapes are exactly the

same in  $RC$  and  $RL$  circuits, the universal time-constant chart in Fig. 16-12 can be used for either type of circuit. Note that the voltage rise across the condenser corresponds to the current rise in the inductance.

*Integrating and Differentiating Circuits.* In an  $RC$  or  $RL$  circuit with a short time constant, the voltage that can change its value instantaneously with the applied voltage is often called the *differentiated* output. This is the voltage across the resistor in an  $RC$  circuit, or across the inductance in an  $RL$  circuit. A differentiating circuit is used when waveform separation of the horizontal sync is desired. As shown in Fig. 16-9, the output of a differentiating circuit consists of sharp peaks of voltage of opposite polarities for the leading and trailing edges of the input pulses. With the total separated sync voltage as input signal, the differentiated output contains sharp voltage peaks corresponding to the voltage changes

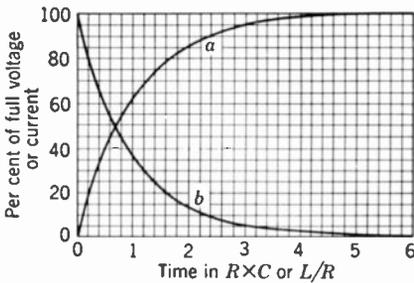


FIG. 16-12. Universal time-constant chart for  $RC$  and  $RL$  circuits. (a) Inductor current or capacitor voltage on charge. (b) Inductor current or capacitor voltage on discharge.

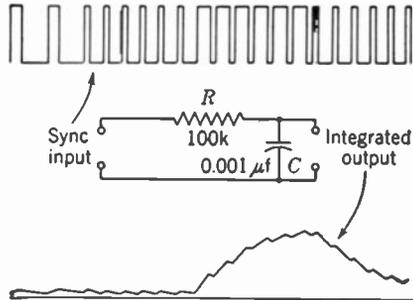


FIG. 16-13. Waveform separation of the vertical synchronizing signals from the total sync by means of an integrating circuit.  $RC$  is equal to  $100 \mu\text{sec}$ .

produced by each pulse. The voltage peaks for the leading edges separated by the horizontal line interval  $H$  have the frequency of 15,750 cps required for horizontal synchronization. In an  $RC$  or  $RL$  circuit where the time constant is not short, the voltage that cannot change instantaneously but must build up exponentially is called the *integrated* output. This is the voltage across the condenser in an  $RC$  circuit, or across the resistor in an  $RL$  circuit. The integrating circuit is used to separate the vertical sync.  $RC$  circuits are generally preferred for differentiating and integrating circuits because a wide range of time constants can be obtained more conveniently than with  $RL$  circuits.

**16-6. Separation of the Vertical Synchronizing Signals.** An integrating circuit, having a time constant long compared with the duration of the horizontal pulses but not with respect to the vertical pulse width, is used to provide the waveform separation needed for vertical synchronization. As shown in Fig. 16-13, the total sync voltage is coupled to the  $RC$  inte-

grating circuit, and the output voltage across the condenser provides vertical synchronizing voltage alone. Since the time constant of the  $RC$  circuit in the illustration is  $100\ \mu\text{sec}$  and the horizontal pulse width is  $5.1\ \mu\text{sec}$ , the condenser can charge to only a small percentage of the applied voltage for the short period during which the horizontal pulse is applied. The period between horizontal pulses, when no voltage is applied to the  $RC$  circuit, is so much longer than the horizontal pulse width that the condenser has time to discharge almost down to zero during this time. The equalizing pulses apply voltage at half-line intervals, but their duration is only one-half the horizontal pulse width, and the condenser cannot charge to any appreciable voltage.

When the vertical pulse is applied, however, the integrated voltage across the condenser can build up to the value required for triggering the vertical scanning generator. The serrated vertical pulse consists of six individual pulses, each of approximately  $27\text{-}\mu\text{sec}$  duration, with serrations having a pulse width of  $4.4\ \mu\text{sec}$ . Since the time constant of the  $RC$  circuit is  $100\ \mu\text{sec}$  and the pulse width is  $27\ \mu\text{sec}$ , charging voltage is applied for  $0.27 RC$  time and the condenser charges to 27 per cent of the applied voltage, as obtained from the universal charge curve in Fig. 16-12. During the serration the applied voltage is removed and the condenser discharges. This is only for  $4.4\ \mu\text{sec}$ , or  $0.044 RC$  time, however, and the condenser loses little of its charge before the next pulse provides sync voltage to recharge the condenser. Thus, the integrated voltage across the condenser builds up to reach its maximum amplitude at the end of the vertical pulse and then declines practically to zero for the equalizing pulses and horizontal pulses that follow, producing a pulse of the triangular wave shape shown for the complete vertical synchronizing pulse. These pulses are repeated at the field frequency of 60 per second. Therefore, the integrated output voltage across the condenser can be coupled to the vertical scanning generator to hold the vertical synchronization.

*Effect of the Equalizing Pulses.* The function of the six equalizing pulses immediately preceding and six following the vertical pulse in improving the accuracy of the vertical synchronization for better interlacing can be seen from an analysis of the  $RC$  integrating circuit. Since there is a half-line difference between the even and odd fields, without equalizing pulses the vertical pulse for the first and odd fields would have to begin a full line after the last horizontal synchronizing signal; the vertical pulse for the second and even fields would begin a half line after the preceding horizontal synchronizing signal, as illustrated in Fig. 16-14. Consequently, the voltage across the vertical integrating condenser for the even fields would not have as much time to discharge down to zero during the half-line interval between the last horizontal pulse and the beginning of the vertical pulse. The integrated vertical voltage would start building up

from a higher voltage than on the odd field, therefore, producing the difference in wave shape for even and odd fields shown in Fig. 16-14. Although this difference may appear slight, it would be sufficient to impair the interlacing because of the slightly different timing of the vertical sweep generator that would be produced for the even and odd fields. With equalizing pulses, this difference in the vertical synchronizing voltage is minimized because the half-line difference between fields precedes the beginning of the vertical pulse by six equalizing pulses. These pulses allow the integrated voltage across the condenser to adjust itself to practically equal values for even and odd fields before the vertical pulse

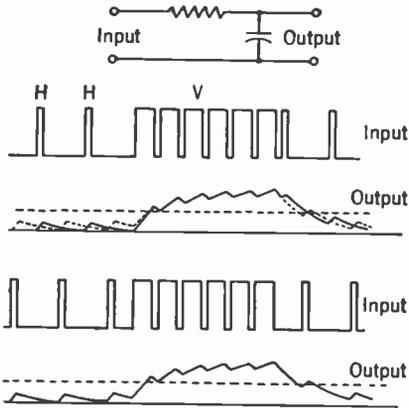


FIG. 16-14. Difference in integrated output between even and odd fields with no equalizing pulses. The dotted waveform in the upper figure is the lower waveform superimposed. The horizontal dashed lines in both figures represent the level to which the condenser voltage must rise to trigger the vertical scanning generator.

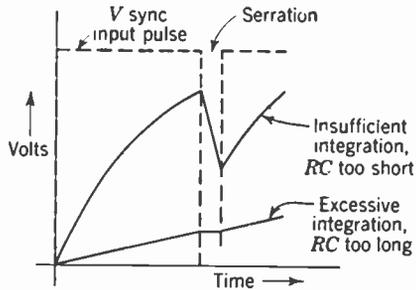


FIG. 16-15. Integrated vertical sync voltage for time constant too long and too short.

begins, since the average value of applied voltage is made more nearly the same for the two cases. The six equalizing pulses that follow the vertical pulse minimize any difference in the trailing edge of the vertical synchronizing signal for even and odd fields.

*Cascaded Integrator Sections.* A very long time constant for the integrating circuit removes the horizontal sync pulses but reduces the vertical sync amplitude across the integrating condenser and results in a slowly rising edge on the integrated vertical pulse. With a time constant that is not long enough, the horizontal sync pulses cannot be filtered out and the serrations in the vertical pulse produce notches in the integrated output. The notches should be filtered out because they give the integrated vertical synchronizing signal the same amplitude value at different times. The wave shapes in Fig. 16-15 illustrate the extreme cases of too much

integration and not enough integration. For good vertical synchronization, the integrated vertical pulse should have sufficient amplitude and rise quickly with a smooth increase up to the amount of voltage required to trigger the vertical deflection oscillator. In order to provide a compromise between good filtering of the horizontal sync with a long time

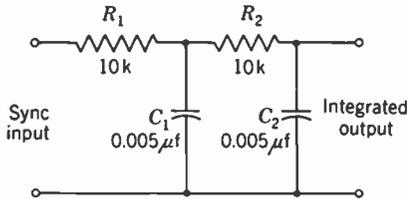


FIG. 16-16. Two-section integrating circuit.

constant and the sharper leading edge and higher amplitude produced by a shorter time constant, the integrating circuit generally consists of two or three sections in cascade, as shown in Fig. 16-16. This is a two-section integrating circuit with each  $RC$  section having a time constant of  $50 \mu\text{sec}$ . The operation of the circuit can be considered as though the  $R_1C_1$  section provided integrated voltage across  $C_1$  that is applied to the next integrating section  $R_2C_2$ . The over-all time constant for both sections is long enough to filter out the horizontal sync, while the shorter time constant of each section allows the integrated voltage to rise more sharply because each integration is performed with a time constant of  $50 \mu\text{sec}$ . The time constant for charging the output condenser in a two-section filter can be calculated as  $R_1(C_1 + C_2) + R_2C_2$ . Referring to Fig. 16-16, the over-all time constant is  $150 \mu\text{sec}$ . For a three-section filter, the over-all time constant would be  $R_1(C_1 + C_2 + C_3) + R_2(C_2C_3) + R_3C_3$ .

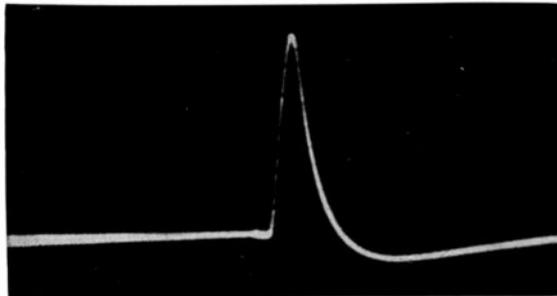


FIG. 16-17. Oscillogram of the integrated vertical synchronizing signal. Vertical deflection oscillator must be disabled to see this pulse without oscillator grid voltage. Oscilloscope internal sweep frequency at 60 cps. (RCA.)

The oscillogram in Fig. 16-17 shows the vertical sync output of a cascaded integrating circuit.

**16-7. Noise in the Sync.** Noise pulse voltages in the sync can cause poor synchronization, resulting in vertical rolling or horizontal tearing of

the picture. The noise pulses are produced by interference sources external to the receiver, such as electrical machinery and auto ignition systems, but the noise can be radiated to become part of the received signal. The noise produces voltage peaks of short duration, as illustrated in Fig. 16-18. In the integrating circuit, the noise pulses charge the integrating condenser, allowing it to reach the voltage amplitude required for synchronizing the vertical deflection oscillator too soon. In horizontal synchronization, the noise pulses can be mistaken for horizontal synchronizing signals. To reduce the effect of noise on synchronization,

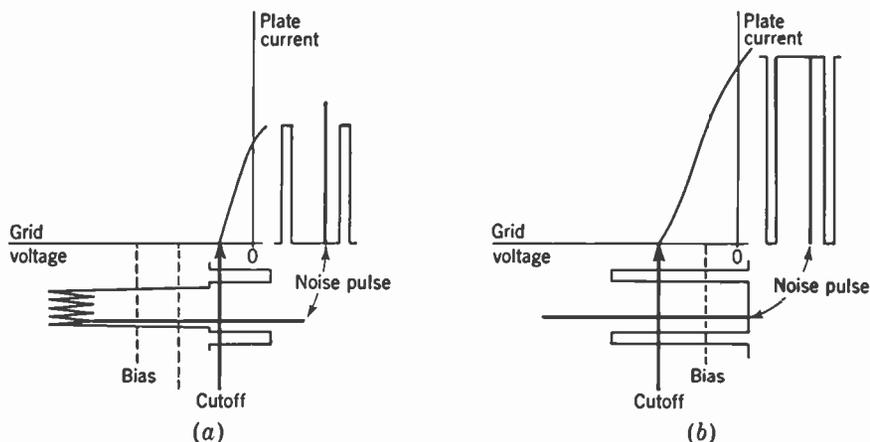


FIG. 16-18. Noise pulse and signal in sync stages. (a) Composite video with positive sync and noise pulse at grid of sync separator. (b) Separated sync of negative polarity with noise pulse at grid of succeeding sync clipper.

the sync separation circuits generally include short-time-constant  $RC$  circuits for the noise pulses in the grid of the sync separator, and a sync clipper stage for the separated sync.

*Clipping the Separated Sync.* The sync output of the separator is often clipped and amplified in the next stage, to ensure sufficient amplitude with sharp synchronizing pulses free from interfering noise and camera signal variations. Clipping in successive stages allows the top and bottom of the sync pulses to be clipped by cutoff, which is sharper than limiting by plate saturation. The operation of a sync separator followed by a sync clipper is illustrated in Fig. 16-18. The operating characteristic in *a* shows composite video signal at the grid of the sync separator, with positive sync polarity, and the separated sync output in the plate circuit. Notice that the noise pulses shown as part of the separated sync signal can have a higher amplitude than the sync voltage, if the noise is stronger than the signal. When the negative sync output in the plate of the separator is coupled to the grid of the next stage, operating as

a sync clipper, the side of the sync pulses that was positive at the grid of the separator is negative at the grid of the clipper. This is shown in *b*. As a result, both the top and bottom of the sync pulses are clipped by cutoff, one side in the sync separator and the other side in the following sync clipper. Although the noise pulse in *a* has higher amplitude than the sync voltage, the noise pulse in *b* has been reduced to the amplitude of the sync voltage because all the pulses are clipped at the cutoff level.

*Sync Separator Time Constant.* The time constant of the grid-leak bias circuit in the input to the separator must be long enough to maintain the bias from line to line and through the time of the vertical synchronizing pulse, in order to maintain a constant clipping level. Typical values are 0.1  $\mu\text{f}$  for the coupling condenser  $C_c$  and 1 megohm for the grid resistor  $R_g$ , providing a time constant of 0.1 sec. These values allow the bias to vary from frame to frame for different brightness values, keeping the tip of sync clamped at zero grid voltage. However, a time constant of 0.1 sec is too long for the bias to follow amplitude variations produced by noise pulses occurring between lines, with a frequency higher than the horizontal sync pulses.

Noise pulses in the input signal to the grid of the sync separator increase the amount of grid-leak bias produced, making the bias more negative than is required for clipping the sync from the composite video signal. This effect of noise pulses increasing the amount of bias produced by the signal is often called *noise setup*. As a result, the gain is reduced for the separated sync. In addition, the noise is amplified with the sync. If the time constant of the grid-leak bias circuit in the sync separator is made smaller for the noise pulses, it will not be long enough to maintain the bias between sync pulses, especially during the vertical sync pulse time. The result may be inadequate sync separation during and immediately after the vertical pulse. Some receivers have separate horizontal and vertical sync separator stages to provide a long time constant for vertical sync separation and a much shorter time constant for horizontal sync separation.

In order to reduce the effect of high-frequency noise pulses on the grid-leak bias for the sync separator, an  $RC$  circuit with a short time constant can be added to the grid circuit of the sync separator, as shown in Fig. 16-19. The grid coupling circuit  $R_gC_c$  provides normal grid-leak bias, with a time constant of 0.1 sec, for the sync signal. The small 150- $\mu\text{f}$  condenser  $C_1$  and the 270,000 ohm resistor  $R_1$  provide a short-time-constant circuit that allows the grid-leak bias to change quickly to reduce the effect of noise pulses in the input to the sync separator.  $C_1$  can charge quickly when noise pulses produce grid current, increasing the bias for noise. Since the  $R_1C_1$  time constant is approximately 40  $\mu\text{sec}$ ,  $C_1$  can discharge through  $R_1$  between sync pulses to keep the bias at the voltage

provided by  $R_0C_0$  for the sync. As a result, the bias stays approximately the same for the sync voltage, allowing the sync to be separated at a constant clipping level. The gain is reduced for noise, however, when  $R_1C_1$  increases the bias momentarily for noise pulses. When there is no noise,  $R_1$  keeps  $C_1$  discharged so that no additional bias is produced.

**16-8. Sync Separation Circuits.** Several circuits are shown here to illustrate typical arrangements for obtaining the synchronizing signals needed to time the deflection oscillators. The first step necessary is separation of the synchronizing pulses from the camera signal variations, which is usually done by clipping the sync from the composite video signal taken from the video amplifier. Grid-leak bias is generally used for the sync separator, which reinserts the d-c component necessary to line up the pulses for a constant clipping level. The separated sync signal can be amplified and clipped further without regard to the d-c level. The high-frequency response of an amplifier for the horizontal synchronizing pulses is important, however, to preserve the sharp leading edge. Values of 5,000 to 50,000 ohms are used for the plate load resistor to minimize

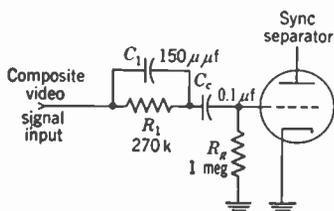


FIG. 16-19. Short time-constant network  $R_1C_1$  to reduce noise pulse amplitude at grid of sync separator.

the effect of shunt capacitance and provide suitable amplification for harmonic frequency components of the horizontal sync up to about 1 Mc. After sufficient amplification and clipping, the total separated sync voltage is coupled to the integrating circuit, which provides only the integrated vertical synchronizing signals with a frequency of 60 cps to the grid of the vertical deflection oscillator. For horizontal synchronization, however, the sync voltage generally is not coupled to the horizontal deflection oscillator. Practically all television receivers have an automatic frequency control (a-f-c) circuit for the horizontal scanning, to minimize the effect of noise on synchronization of the horizontal oscillator. In this type of circuit, the sync voltage is coupled to the oscillator control stage, which can then hold the horizontal oscillator at the horizontal line frequency of 15,750 cps.

**Sync Separator, Clipper, and Phase Inverter.** In the sync circuits shown in Fig. 16-20, composite video signal from the output of the video amplifier, with positive sync polarity, is coupled to the grid of the first sync stage, which uses one triode section of the 12AU7 twin-triode  $V_{304}$ . This stage is a grid-leak bias sync separator to provide separated sync voltage

of negative polarity in the plate circuit, without camera signal. The negative sync output from the plate is coupled through  $C_{409}$  to the grid of the next sync stage, which amplifies the separated sync and clips the negative side of the sync pulses. The amplified output in the plate of this clipper is positive sync that is coupled through  $C_{410}$  to the grid of the sync inverter, which uses one triode section of the 6SN7-GT twin-triode  $V_{401B}$ .

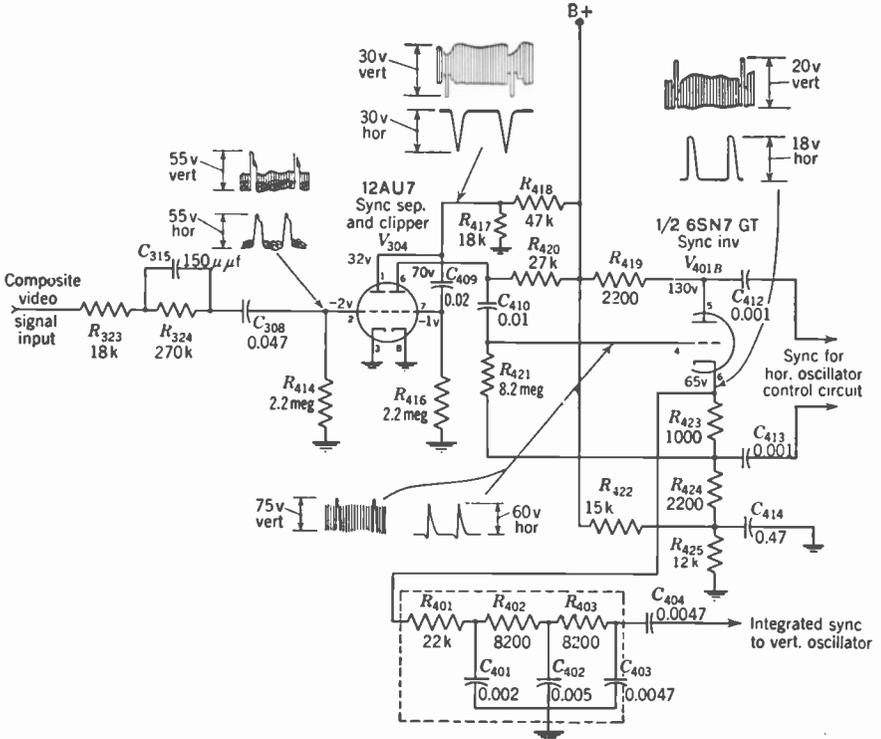


FIG. 16-20. Sync separation circuit with grid-leak bias separator, clipper, and phase inverter stages. Vertical wave shapes can be observed on oscilloscope with internal sweep frequency at 30 cps, horizontal at 15,750/2 cps. Wave-shape amplitudes are peak-to-peak voltage. Capacitance values less than 1 in  $\mu\text{f}$ , more than 1 in  $\mu\text{mf}$ . (*Admiral Series 21 chassis*.)

The inverter operates as a phase splitter to provide push-pull sync output voltage, because the horizontal a-f-c circuit in this chassis has push-pull input, requiring sync voltages of equal amplitude and opposite polarity. Negative sync voltage from the plate of the phase inverter, across the 2,200-ohm plate load resistor  $R_{419}$ , is coupled by  $C_{412}$  to one side of the a-f-c circuit, while positive sync voltage of equal amplitude across the 2,200-ohm cathode resistor  $R_{424}$  is coupled by  $C_{413}$  to the opposite side of the a-f-c circuit. The bypass condenser  $C_{414}$  returns  $R_{424}$  to chassis

ground for the horizontal sync signal, so that no horizontal sync voltage will be developed across  $R_{425}$  to unbalance the push-pull circuit. The total sync voltage of positive polarity at the cathode, across  $R_{423}$ ,  $R_{424}$ , and  $R_{425}$ , is connected to the three-section integrating circuit for vertical synchronization. The integrated voltage across  $C_{403}$  is vertical sync only, for the grid of the vertical deflection oscillator, to synchronize the vertical scanning. The dotted lines around the integrator indicate it is a printed circuit in one unit.

The sync clipper stage has grid-leak bias, provided by  $C_{409}$  and  $R_{416}$ , although the separated sync signal here has negative polarity. This just means the tip of sync drives the grid voltage in the negative direction. The separated sync voltage in the grid circuit is an a-c signal, however, with positive and negative variations from the average-value axis. Therefore, the grid can be driven positive, as illustrated by the sync signal shown in Fig. 16-18*b*, to produce grid current and the resultant grid-leak bias.

In the cathode circuit of the sync inverter,  $R_{425}$  forms a voltage divider with  $R_{422}$ , across the B supply voltage, applying positive voltage across  $R_{425}$  to the cathode. This reduces the plate-to-cathode voltage, allowing additional clipping of the sync. Notice that the grid circuit of the inverter returns to the junction of  $R_{423}$  and  $R_{424}$  in the cathode, making only the cathode voltage across  $R_{423}$  effective as bias for the grid signal.

*D-C Restorer with Sync Amplifier, Separator, and Clipper.* In Fig. 16-21 the d-c restorer stage  $V_{114B}$ , using one diode of the 6AL5, also separates the camera signal variations from the sync to provide partially separated sync for the three-stage sync circuit. The diode plate current through  $R_{150}$  during sync pulse time develops the sync output voltage, of negative polarity. This is coupled by  $C_{144}$  to the first sync amplifier  $V_{118}$ , which uses the remote-cutoff pentode 6SK7. The composite video signal input to the restorer and the partially separated sync output for the first sync amplifier are shown in the oscillograms in Fig. 16-22*a* and *b*. The polarity of the sync voltage is negative at the grid of the first sync amplifier and the stage has enough fixed bias plus grid-leak bias to cut off noise pulses of greater amplitude than the sync pulses. The sync output of the first sync amplifier has positive polarity, as shown in *c* of Fig. 16-22. This positive sync voltage is coupled by  $C_{147}$  to the grid of the sync separator  $V_{119}$ , which uses the sharp-cutoff pentode 6SH7. In this stage, the grid-cutoff voltage is low enough to produce output only for the most positive part of the sync voltage, while the more negative video and blanking signals are cut off. The sync output is shown in *d*. This sync signal, of negative polarity, is coupled by  $C_{148}$  to the second sync amplifier  $V_{120A}$ , using one triode section of the 6SN7-GT. Grid-leak bias is used here and the plate voltage is relatively low, allowing the negative side of the sync

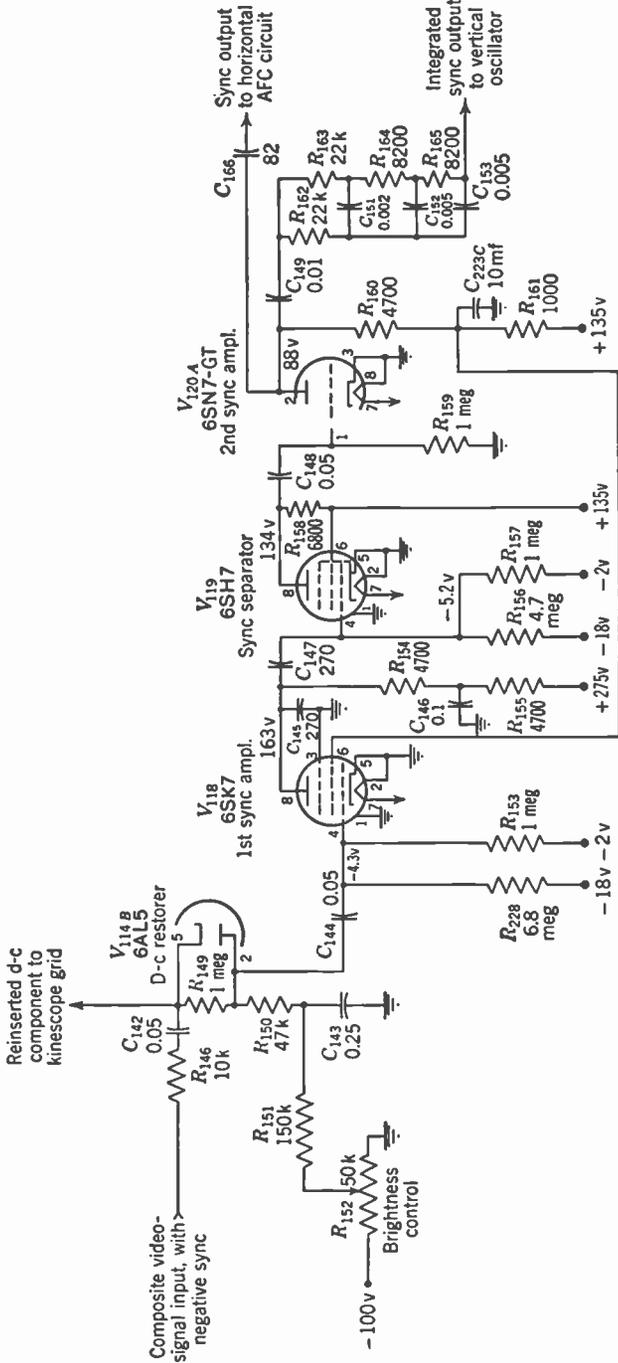
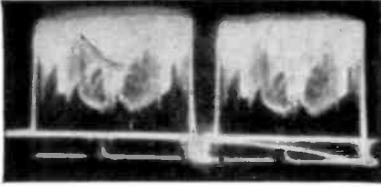
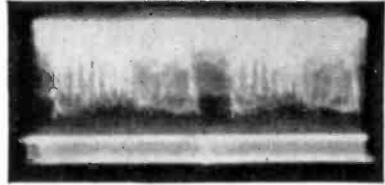


Fig. 16-21. Sync separation circuits with d-c restorer, amplifier, separator, and clipper. Capacitance values less than 1 in  $\mu f$ , more than 1 in  $\mu f$  unless otherwise noted. (RCA Model 630 TS chassis.)

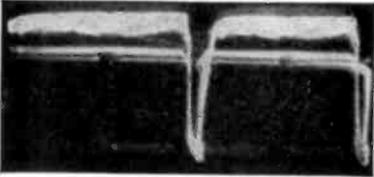


Horizontal 36 volts

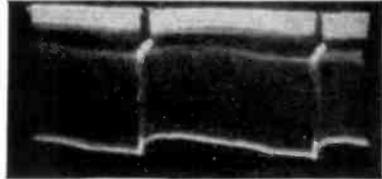


Vertical 36 volts

(a) Composite video signal input to d-c restorer-sync separator at cathode pin 5 of 6AL5.

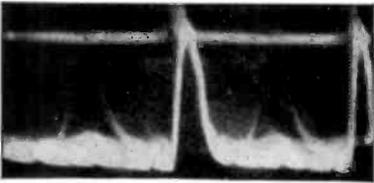


Horizontal 9 volts

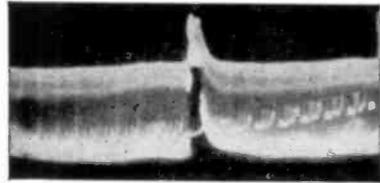


Vertical 9 volts

(b) Partially separated sync output of d-c restorer-sync separator at plate pin 2 of 6AL5.

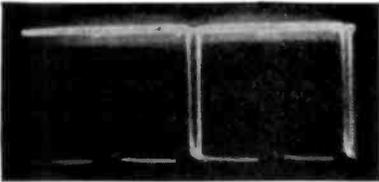


Horizontal 40 volts

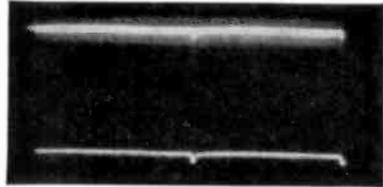


Vertical 58 volts

(c) Partially separated sync output of first sync amplifier at plate pin 8 of 6SK7.

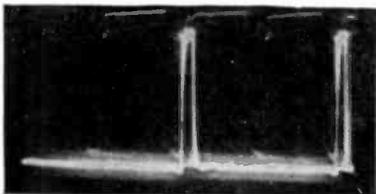


Horizontal 75 volts

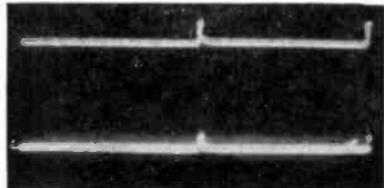


Vertical 75 volts

(d) Sync output of sync separator at plate pin 8 of 6SH7.



Horizontal 29 volts



Vertical 35 volts

(e) Sync output of second sync amplifier at plate pin 2 of 6SN7-GT.

FIG. 16-22. Oscillograms of wave shapes in sync separation circuits shown in Fig. 16-21. Oscilloscope internal sweep frequency at  $15,750/2$  cps for horizontal wave shapes and 30 cps for vertical wave shapes. Amplitudes given in peak-to-peak voltage. (RCA.)

input voltage and noise pulses to be clipped by cutoff, while the positive side is limited by plate-current saturation, to provide constant-amplitude sync output. The positive sync output of the second sync amplifier is shown in *e* of Fig. 16-22. This sync voltage is coupled by  $C_{149}$  to the integrator consisting of  $R_{163}C_{151}$ ,  $R_{164}C_{152}$ , and  $R_{165}C_{153}$ . The integrated voltage output across  $C_{153}$  is coupled to the grid of the vertical deflection oscillator to lock it in at the vertical synchronizing frequency. The output of the second sync amplifier is also coupled by  $C_{166}$  to the a-f-c circuit of the horizontal deflection oscillator to hold it synchronized at the horizontal scanning frequency.

*Gated Sync Separator and Clipper.* The sync separator and clipper circuit in Fig. 16-23 uses a tube with two control grids, such as the 6BN6 gated-beam tube or the 6BE6 and 6BY6 pentagrid tubes, as a gated

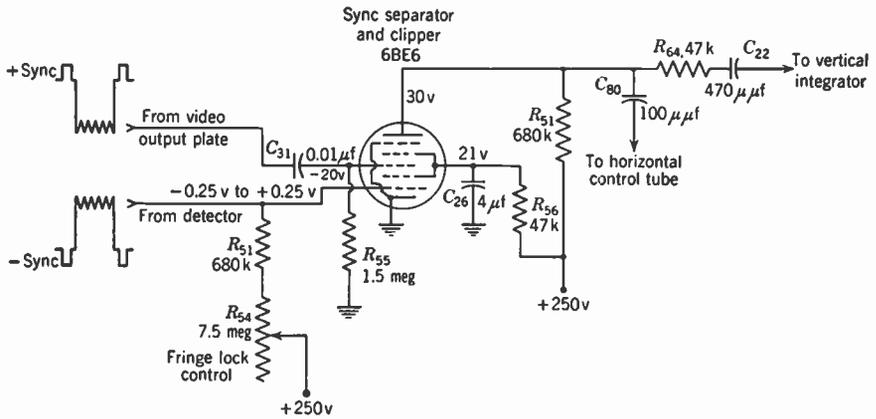


FIG. 16-23. Gated sync separator and clipper stage. (Zenith chassis 22L20.)

amplifier to combine the functions of separating the sync from the composite video signal input, while clipping noise pulses to provide noise-immune separated sync output in the plate circuit. In the 6BE6 in Fig. 16-23, the first grid next to the cathode is control grid 1 and the third grid is control grid 2. Cutoff voltage for each control grid is approximately  $-2$  volts, with plate and screen voltages of 20 to 30 volts. Either control grid can cut off the plate current. Both control grids must be less negative than cutoff at the same time, therefore, to produce the gated output.

Composite video signal input having positive sync polarity from the plate of the video amplifier is coupled to control grid 2 by  $C_{31}$  with  $R_{55}$ . The negative grid-leak bias allows just the peak positive sync voltage amplitudes of the input signal to produce plate current, resulting in separated sync output across the plate load resistor  $R_{51}$ , as grid voltages more negative than  $-2$  volts are beyond cutoff. Control grid 1 has composite video signal input of negative sync polarity, and relatively low amplitude,

from the video detector. The low signal amplitude, with a slight positive bias on control grid 1, enables the sync input voltage to remain less negative than cutoff, allowing plate current to flow for producing the separated sync output. However, noise pulse voltages in the composite video signal, which drive the grid 1 voltage more negative, cut off the plate current. As a result, noise pulses in the composite video signal input are not present in the separated sync output. In addition, the grid-leak bias voltage on control grid 2 cannot be increased by noise setup, which tends to reduce the sync output, since noise pulse voltage cannot charge  $C_{31}$  while current in the tube is cut off by control grid 1.

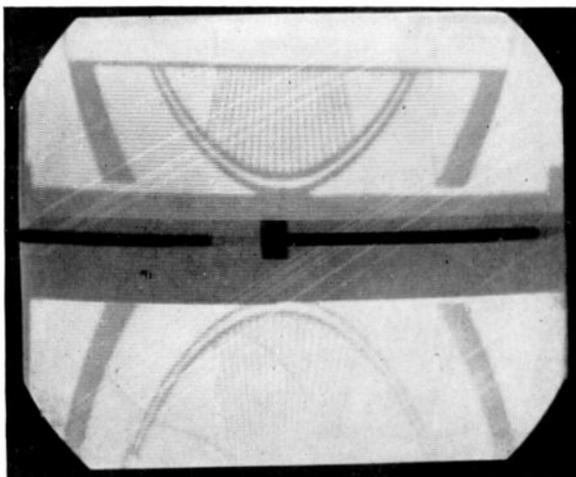


FIG. 16-24. The hammerhead pattern with vertical sync and equalizing pulses on the kinescope screen within the vertical blanking bar. (RCA.)

The bias on control grid 1 is adjusted by the fringe lock control  $R_{54}$  for normal synchronization. Less positive bias provides better noise immunity with weak signals, enabling noise pulses to drive the grid voltage negative enough to cut off plate current. Too little positive bias can reduce the separated sync output, however, causing weak vertical and horizontal hold in the picture.

**16-9. Sync and Blanking Bars on the Kinescope Screen.** Although the video signal is coupled to the sync separator to provide the desired sync pulses, the entire composite video signal with sync and blanking pulses is also coupled to the kinescope grid-cathode circuit. The amplitude variations of the sync and blanking pulses can be seen, therefore, as relative intensities on the kinescope screen. Figure 16-24 shows the details of the vertical sync and blanking bars on the kinescope screen. This can be seen by varying the vertical hold control to allow the picture to roll slowly, out of sync, so that the vertical blanking bar is in the mid-

dle of the picture instead of at the top and bottom. Brightness is turned up higher than normal to make the blanking level gray instead of black. The sync amplitude, which is 25 per cent above the blanking level, then becomes black. The black bar within the vertical blanking bar, often described as the "hammerhead" pattern, represents the equalizing and vertical sync pulses occurring during vertical blanking time. The appearance of the horizontal sync within the horizontal blanking bar on the kinescope screen is shown in Fig. 16-25. This can be seen by adjusting the phasing of the horizontal deflection oscillator to put the horizontal blanking bar in the picture, with the brightness higher than normal. Normally the sync and blanking bars are at the edges of the picture

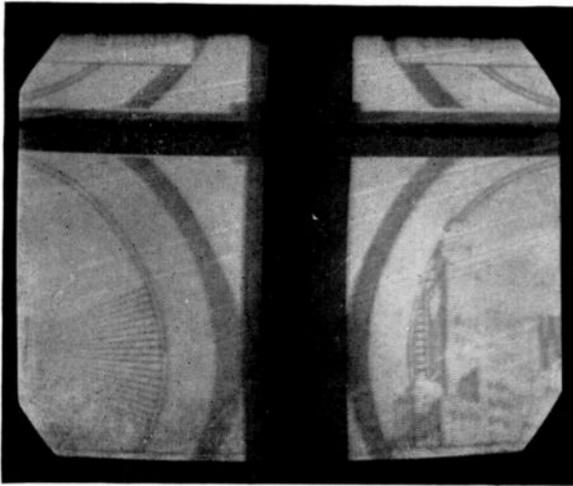


FIG. 16-25. Horizontal sync and blanking bars on the kinescope screen. (RCA.)

behind the mask of the screen and are not visible. The bars can be examined, however, to check the sync voltage at the kinescope grid. Sync that is blacker than blanking and the darkest parts of the picture on the kinescope screen shows normal sync voltage in the composite video signal input to the sync separator.

**16-10. Sync Troubles.** Vertical rolling and horizontal tearing or bending in the picture mean faulty synchronization of the scanning raster with respect to the reproduced picture information. It should be noted, though, that the separated sync is part of the received picture signal and therefore the receiver must have enough signal to provide good synchronization. In most receivers a weak picture with hardly enough contrast to be visible cannot be synchronized. Also, interfering noise pulses can easily interrupt the synchronization with a weak signal. When the picture has good quality and contrast, however, poor synchronization indicates sync trouble. All channels are affected by a trouble in the sync circuits.

A trouble in the sync circuits where only vertical sync is used such as the vertical integrating circuit, can cause loss of vertical synchronization while horizontal sync is normal. Or trouble in a horizontal sync circuit can cause loss of just the horizontal synchronization. The a-f-c circuit for the horizontal deflection oscillator provides horizontal synchronization only. Some receivers have individual sync separator stages for horizontal and vertical sync. In these circuits, a trouble in one stage can cause poor vertical synchronization, while the horizontal sync is normal, or vice versa. When there is trouble in a stage where horizontal and vertical sync are present, as in a common sync amplifier, both vertical and horizontal synchronization can be poor at the same time.

It is important to remember that the vertical and horizontal sync pulses are part of the r-f and i-f picture signal and the video signal for the sync separator. Distortion of the signal by limiting or clipping in the signal circuits can reduce the amplitude of the sync. Poor response for the low video frequencies can distort the vertical sync. Checking the sync and blanking on the kinescope screen will show whether the sync is normal in the composite video signal for the sync separator. In cases of distorted sync in the signal circuits, the picture contrast will be distorted, usually, in addition to the faulty synchronization.

*No Vertical Hold.* This is illustrated in Fig. 16-1. Vertical rolling of the picture means no vertical synchronization. The trouble may be either no vertical sync to lock in the vertical deflection oscillator, or the oscillator is so far off the correct frequency of 60 cps that the vertical sync voltage present cannot lock in the oscillator. In order to hold an oscillator synchronized not only is sync voltage necessary but the oscillator frequency must be close enough to the synchronizing frequency to enable the sync voltage to lock in the oscillator. To check whether the trouble is no vertical sync or incorrect oscillator frequency, vary the vertical hold control to see if one complete picture can be stopped, in frame, for just an instant. If varying the vertical hold control cannot stop one complete picture, the trouble is incorrect frequency of the vertical oscillator. When the hold control stops the picture but it slips vertically out of hold, the trouble is no vertical sync input to the vertical deflection oscillator.

*No Horizontal Hold.* When the picture tears apart in diagonal segments this means no horizontal synchronization. This is shown in Fig. 16-2. Again, the trouble can be in either the oscillator or the sync. To check, vary the horizontal oscillator frequency control to see if a whole picture can be produced, although it will slip horizontally. If a whole picture cannot be produced, the trouble is incorrect frequency of the horizontal oscillator. When varying the frequency control can produce a whole picture but it does not stay still horizontally, the trouble is no horizontal sync input to the a-f-c circuit for the horizontal deflection

oscillator, or the control circuit is not synchronizing the oscillator. It should be noted that many receivers have a separate setup adjustment for the horizontal frequency control, in addition to the horizontal hold control in the horizontal a-f-c circuit.

*Poor Interlace.* Inaccurate timing of the vertical scanning in even and odd fields causes poor interlacing of the scanning lines, resulting in partial pairing or complete pairing, which reduces the detail in the picture. Stray pickup of pulses generated by the receiver's horizontal deflection circuits and coupled into the vertical sync and deflection circuits can cause



FIG. 16-26. Bend just at the top of the picture, with straight edge on raster. (RCA.)

interlace troubles. The pulses generated locally by the horizontal deflection circuits for scanning are high in amplitude and do not have exact half-line difference in timing for even and odd fields. They can override the vertical sync, therefore, and produce inaccurate timing of the vertical scanning, with poor interlace.

*Horizontal Pulling in the Picture.* Weak horizontal sync allows the picture to bend or pull horizontally, as successive scanning lines are slightly displaced with respect to the picture information, but not enough to make the picture tear apart. Horizontal pulling often appears only at the top of the picture, as illustrated in Fig. 16-26. Note that the edges of the raster are straight, showing that the weak horizontal sync causes bend in the picture but not in the raster. The bend at the top of the picture indicates weak horizontal sync just after the vertical sync pulse, since the horizontal scanning at the top immediately follows the vertical flyback. Weak horizontal sync all the time generally makes the picture tear apart, but when the horizontal sync is weak for only part of the picture it pulls or bends. The bend in the picture is caused by a small con-

tinuous change in the frequency of the deflection oscillator. The scanning amplitude can remain the same to produce constant width with straight edges on the raster.

Figure 16-27 illustrates how the horizontal sync voltage can be reduced, starting with the vertical sync pulse, to cause weak horizontal sync for the top of the picture. This can happen because the average value of the sync voltage increases during the vertical pulse. Therefore, an *RC* bias circuit with too short a time constant for the vertical sync will temporarily

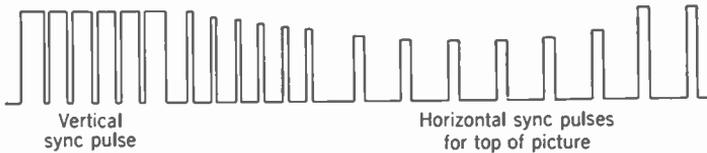


FIG. 16-27. Weak horizontal sync following the vertical pulse. Not drawn to scale, as reduced amplitude of horizontal pulses can continue for many more scanning lines at the top of the picture.

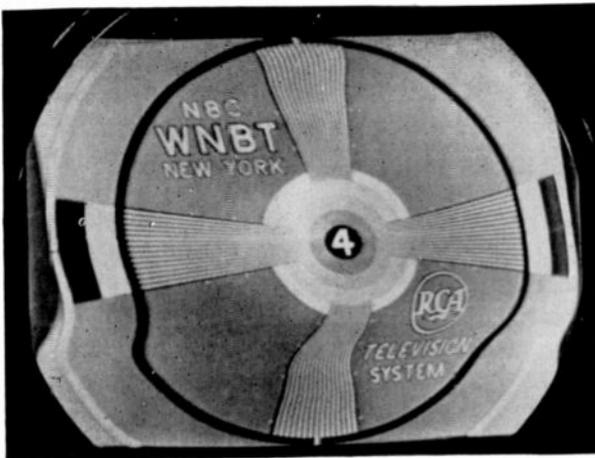


FIG. 16-28. 60-cycle hum bend in the picture, but not in the raster, without hum bars. See text for explanation of these symptoms that indicate heater-to-cathode leakage in a horizontal sync tube. (RCA.)

charge to a value that is too high for the average value of the total sync. As an example, if the vertical sync pulse makes the grid-leak bias too negative in the sync separator stage, the amplification will be reduced for the horizontal sync. Then the horizontal sync is weak following the vertical pulse, resulting in bend at the top of the picture.

*Hum in the Sync.* Excessive hum in the horizontal sync produces bend in the picture as shown in Fig. 16-28. The hum in the sync bends the picture but the edge of the raster is straight. This shows that the hum is in the horizontal sync but not in the raster circuits. With excessive hum in the vertical sync, the picture tends to lock in halfway out of phase, with

the vertical blanking bar across the middle of the picture, staying still in the position shown in Fig. 16-1*a*. Hum voltage can combine with the synchronizing pulses either in the sync circuits or in any of the picture signal circuits and video signal circuits before the sync is separated. When

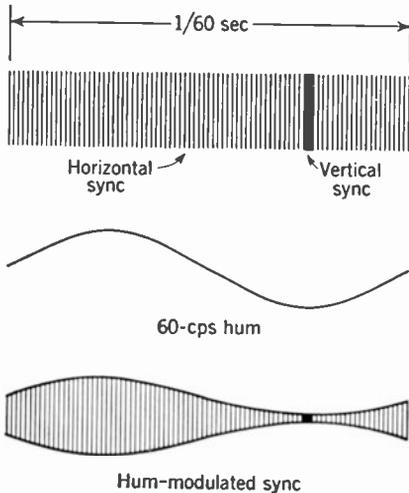


FIG. 16-29. 60-cycle hum voltage modulating sync amplitude. Notice that the sync amplitude can be reduced practically to zero by the hum voltage.

2. Hum bend in the picture with dark and light hum bars across the screen shows that the hum is introduced in the signal circuits, before sync separation.

3. Hum bend in the picture without hum bars shows that the hum is introduced in the sync circuits after sync separation.

4. Heater-to-cathode leakage in a tube causes 60-cps hum; 120-cps hum is B supply ripple from a full-wave power supply.

The horizontal sync waveform resulting from 60-cps hum voltage modulating the amplitude of the synchronizing pulses is shown in Fig. 16-29. Note that, when the hum voltage is strong enough during the negative half cycle, it can reduce the sync amplitude to zero. With or without sync, though, the raster is scanned. The temporary loss of sync causes bend in the picture, instead of tearing, as the *a-t-c* circuit keeps the horizontal deflection oscillator from changing its frequency abruptly.

The hum voltage can reduce the amplitude of the vertical sync also, as illustrated in Fig. 16-29. Excessive hum voltage may result in practically no vertical sync, allowing the hum voltage itself to trigger the vertical deflection oscillator. If the vertical deflection oscillator were triggered

introduced in the signal circuits before sync separation, though, the hum will be combined with the video signal at the kinescope grid. The resultant dark and light hum bars on the screen then show that the hum is in the video signal. The hum voltage may be 60 cps, with one cycle of bending from top to bottom of the picture, or 120 cps producing two cycles. The 60-cps hum is usually from heater-cathode leakage.

Summarizing these effects, then, the source of the hum voltage introduced into the horizontal sync can be localized by noting that:

1. Hum bend in the picture while the raster is straight shows the hum is in the horizontal sync but not in the deflection circuits.

by the positive peak of the 60-cps hum voltage in Fig. 16-29, which is displaced from the vertical sync pulse by the time of one-half a scanning field, the picture would look like Fig. 16-1a, with the black blanking bar across the middle and the center of the picture at the bottom of the raster where the vertical retrace starts.

## REVIEW QUESTIONS

1. What is the function of the synchronizing signals?
2. What is the effect on the picture if the receiver does not hold its vertical synchronization? What is the effect when the horizontal synchronization does not hold?
3. Describe the sequence used in obtaining the synchronizing signals for the scanning generators in the receiver.
4. What is meant by a clipper stage?
5. Draw the schematic diagram of a sync amplitude separator stage and explain briefly its operation. Show the input and output voltage wave shapes.
6. Why must self-bias be used in the sync separator stage?
7. What is the function of the horizontal synchronizing pulses, the vertical pulses, and the equalizing pulses?
8. A grid-leak bias circuit has a  $150\text{-}\mu\text{f}$  condenser that can discharge through a  $270,000\text{-ohm}$  resistor. How long will it take for the condenser to discharge down to one-half the voltage across the condenser at the start of the discharge?
9. 100 volts is applied to an  $RC$  circuit having a time constant of 1 sec. With 10 volts across the condenser at the start, how much voltage will there be across the condenser after 2 sec?
10. The total sync voltage is applied to an  $RC$  circuit consisting of a  $22,000\text{-ohm}$  resistor in series with a  $0.005\text{-}\mu\text{f}$  condenser. Show the wave shape of the output voltage across the condenser for the serrated vertical synchronizing pulse.
11. Referring to the sync separation circuits in Fig. 16-20, what is the sync voltage polarity in the composite video signal input to the sync separator stage? What is the polarity of the sync voltage output at the cathode of the sync inverter stage?
12. Referring to the sync separation circuits in Fig. 16-20 give the function of the following components:  $C_{315}$ ,  $R_{425}$ ,  $R_{422}$ ,  $C_{410}$ ,  $R_{417}$ ,  $R_{418}$ ,  $R_{419}$ ,  $R_{420}$ ,  $C_{403}$ ,  $C_{412}$ ,  $C_{413}$ , and  $C_{414}$ .
13. Referring to the sync separation circuits in Fig. 16-21, what is the sync voltage polarity at the grid of the 6SH7 sync separator? At the plate of the 6SN7-GT second sync amplifier?
14. Referring to the sync separation circuits in Fig. 16-21, give the function of the following components:  $C_{147}$ ,  $C_{148}$ ,  $R_{154}$ ,  $R_{155}$ ,  $C_{146}$ ,  $R_{158}$ ,  $R_{159}$ , and  $R_{160}$ .
15. Referring to the gated sync separator and clipper circuit in Fig. 16-23, what input signals are applied to the two control grids? Where is the separated sync output obtained?
16. What sync pulses are in the hammerhead pattern? Describe briefly how the hammerhead pattern can be used to localize a sync trouble in the receiver.
17. Describe briefly how to determine whether the trouble of vertical rolling of the picture is caused by no vertical sync or incorrect vertical oscillator frequency.
18. Give two causes of horizontal pulling in the picture.
19. Give two effects in the picture that can be caused by 60-cps hum produced by heater-to-cathode leakage in a picture i-f amplifier stage. How could this be distinguished from hum introduced in the sync amplifier?

## CHAPTER 17

### DEFLECTION OSCILLATORS

The scanning of the electron beam in the picture tube is made possible by the deflection oscillator stages in the receiver. As illustrated in Fig. 17-1, the deflection voltage generated by the vertical oscillator is coupled to the vertical deflection amplifier, which supplies the amount of current needed for the vertical scanning coils in the yoke mounted on the picture tube. Similarly, the horizontal oscillator produces the deflection voltage required for horizontal scanning. The deflection oscillators are able to operate without any external signal. Therefore, they can produce scanning with or without synchronizing signal input. In order to time the

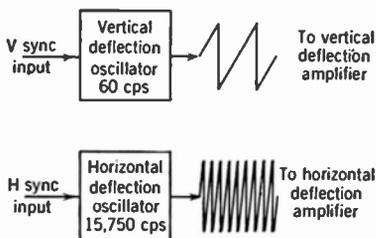


FIG. 17-1. The horizontal and vertical deflection oscillators generate saw-tooth voltage for scanning.

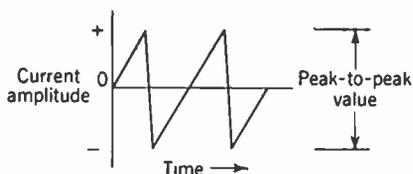


FIG. 17-2. The saw-tooth deflection waveform.

scanning correctly with respect to the transmitted picture information, however, synchronization of the deflection oscillators is necessary. The deflection oscillator stage is often called a *deflection generator*, *sweep oscillator*, *saw-tooth oscillator*, or *saw-tooth generator*.

**17-1. Saw-tooth Deflection.** The saw-tooth waveform is required for scanning because it has a linear rise in amplitude to deflect the electron beam at uniform speed for the linear trace, with a sharp drop in amplitude for the fast retrace, as illustrated in Fig. 17-2. Note that the saw-tooth scanning current is an a-c wave. The electron beam is centered by positioning controls, and the a-c saw-tooth current in the deflection coils deflects the beam away from center. Zero amplitude on the saw-tooth

deflecting wave is the time when the beam is undeflected, at the center. If positive polarity of the saw-tooth wave for horizontal deflection moves the beam toward the right, negative polarity deflects to the left. Similarly, if positive polarity of the saw-tooth wave for vertical deflection moves the beam downward from center, negative polarity deflects upward. The electron beam is at the extreme positions at the left and right sides, or top and bottom, of the raster when the saw-tooth deflection wave has its peak negative and positive amplitudes. The peak-to-peak amplitudes of the a-c saw-tooth current in the horizontal and vertical deflection coils determine the width and height of the raster.

**17-2. Neon-tube Saw-tooth Oscillator.** Figure 17-3 shows an oscillator circuit that generates saw-tooth voltage output. The circuit uses an  $RC$  network with a gas tube in parallel with the condenser. The gas tube is the neon cold-cathode glow type. Unless the voltage across the tube is great enough to ionize the gas in the tube it has almost infinite resistance, acting as an open circuit. When the tube potential becomes great

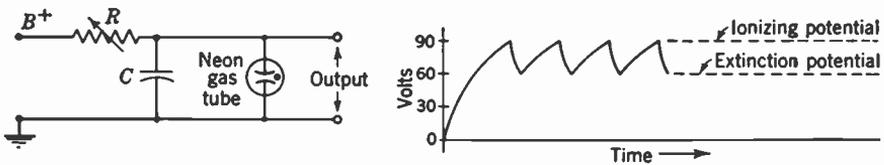


FIG. 17-3. Neon-tube saw-tooth oscillator.

enough to ionize the gas, the tube becomes a low resistance because of the ionization current. The tube remains ionized for voltages very much less than the value required to ionize the gas, but when the tube voltage becomes too low the gas deionizes, current flow ceases through the tube, and it becomes an open circuit. The potential at which the gas ionizes and conduction begins in the tube is the ionizing potential, often called the *striking* or *firing potential*. The voltage at which deionization takes place is the *deionizing* or *extinction potential*. The tube can be considered as a switch that is closed when the gas is ionized and opens with deionization.

When d-c voltage is applied to charge the condenser  $C$  through the series resistance  $R$  in Fig. 17-3, the condenser charges to the supply voltage with the exponential charging curve shown, at a rate determined by the  $RC$  time constant. If the ionizing potential is 90 volts for the neon tube used, the tube will ionize at the time when the capacitor is charged to 90 volts, since the condenser voltage is applied directly across the tube. During the time it takes for the condenser to charge to 90 volts the tube is deionized and is in effect an open circuit, allowing the condenser to charge as though the tube were not in the circuit. When the voltage across the

condenser reaches 90 volts, however, the tube ionizes to become a low-resistance path for condenser discharge. The condenser then discharges very rapidly through this relatively short time-constant circuit, and the voltage across the condenser declines toward zero. When the condenser voltage is down to the extinction potential, the voltage across the tube is not great enough to maintain ionization and the gas deionizes. Deionized, the tube is an open circuit. The condenser then begins again to charge toward the d-c supply voltage. The condenser voltage rises along the normal  $RC$  charge curve to the firing potential and falls as the condenser discharges rapidly through the low-resistance path of the ionized tube. This process continues as long as the d-c supply voltage is maintained, to produce saw-tooth voltage output across the condenser.

The frequency of the saw-tooth wave is the rate of repetition per second of a complete cycle, which includes a voltage rise and fall. The frequency

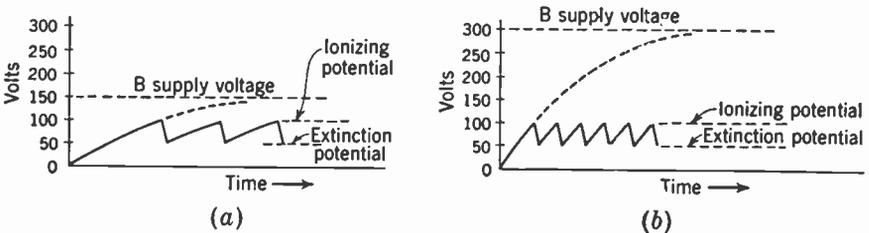


FIG. 17-4. Effect of B supply voltage on linearity of output from gas-tube saw-tooth oscillator. Increasing the supply voltage improves the linearity. (a) 150-volt supply. (b) 300-volt supply.

can be varied by changing the supply voltage, the  $RC$  time constant, or by using a tube with a different ionizing potential. Varying the time constant changes the amount of time required for the condenser to charge to the ionizing potential. A shorter time constant reduces the time required to build up to the tube's ionizing potential, increasing the frequency, while a longer time constant decreases the frequency. Normally, the frequency is varied by using the variable resistor  $R$  to vary the  $RC$  time constant of the charging circuit. Various values of condensers can also be used in a switching arrangement to provide a rough adjustment of frequency.

The linearity of the voltage rise across a condenser can be improved by using only a small part of the normal exponential  $RC$  charge curve. If a higher supply voltage is used, the condenser voltage can still charge to the required value while using only the initial linear portion of the charge curve. As illustrated in Fig. 17-4, when 300 volts is used instead of 150 volts for supply voltage, the condenser can charge up to 90 volts while still on the linear portion of the charge curve, because 90 volts is now a smaller percentage of the applied voltage. If the charging time is restricted to

the initial 40 per cent of the charge curve, the departure from linearity on the voltage rise will be no more than 1 per cent.

**17-3. Blocking Oscillator and Discharge Tube.** Figure 17-5 shows how a vacuum tube can be used, instead of a gas tube, as a discharge tube

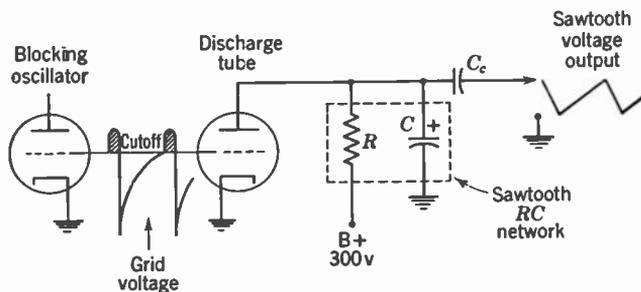


FIG. 17-5. Blocking oscillator grid voltage driving discharge tube to produce saw-tooth voltage output across saw-tooth condenser  $C$  in plate of discharge tube. Shaded area in grid-voltage waveform indicates when the discharge tube conducts.

in parallel with the condenser  $C$  to produce saw-tooth voltage output. The plate circuit of the discharge tube has the  $RC$  network for producing saw-tooth voltage output across  $C$ , which is often called the *saw-tooth condenser*. However, the discharge tube needs voltage from the blocking oscillator. The grid voltage applied by the blocking oscillator keeps the discharge tube cut off for a relatively long period of time and then makes the discharge tube conduct for a short time. During the time the grid voltage is more negative than cutoff, the discharge tube cannot conduct plate current and is effectively an open circuit. While the discharge tube is cut off, therefore, the saw-tooth condenser  $C$  in the plate circuit charges toward the  $B$  supply voltage, through the series resistance  $R$ , to produce the linear rise on the saw-tooth voltage wave. When the grid voltage drives the discharge tube into conduction, its plate-to-cathode circuit becomes a low resistance equal to the plate resistance of the tube. Then the saw-tooth condenser discharges quickly from cathode to plate through the discharge tube, producing the rapid fall in voltage for the flyback on the saw-tooth

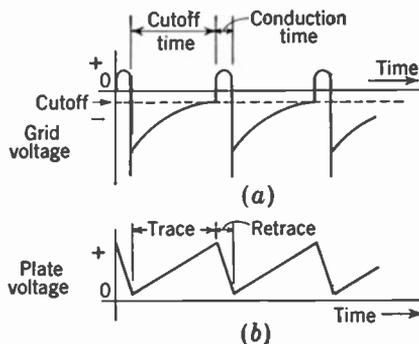


FIG. 17-6. The positive pulse of grid voltage produces flyback in the saw-tooth wave; cutoff corresponds to the linear rise on the saw-tooth wave for the trace. (a) Grid voltage of blocking oscillator and discharge tube. (b) Plate voltage of discharge tube.

voltage wave. Therefore, by applying narrow positive pulses to the grid of the discharge tube and keeping it cut off between pulses, a saw-tooth wave of voltage is produced in the output. Figure 17-6 illustrates how the linear rise of voltage in the output corresponds to the time when the grid is more negative than cutoff, while the flyback time coincides with the positive grid pulse, resulting in a saw-tooth wave of voltage output from the discharge tube with the same frequency as the grid pulses from the blocking oscillator.

*Blocking Oscillator Circuit.* Figure 17-7 shows the schematic diagram of the blocking oscillator circuit that supplies the grid voltage for the dis-

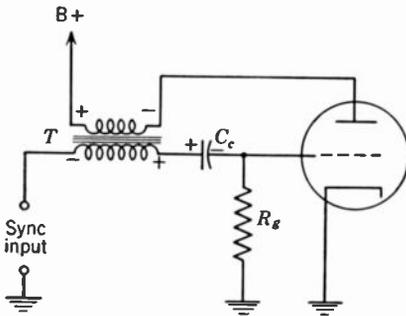


FIG. 17-7. Blocking oscillator circuit diagram.

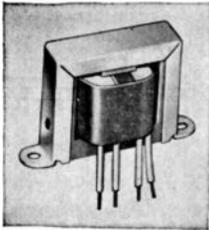


FIG. 17-8. Horizontal blocking oscillator transformer. Primary-to-secondary turns ratio 2:1. Physical dimensions are  $1\frac{1}{2}$  in.  $\times$   $2\frac{1}{2}$  in.  $\times$   $1\frac{1}{2}$  in. high. (Standard Transformer Corporation.)

charge tube. The blocking oscillator transformer  $T$  is connected to produce grid feedback voltage with the polarity required to reinforce the grid signal and start oscillations. When the grid is driven positive, grid current flows, developing grid-leak bias approximately equal to the feedback voltage. This regenerative circuit ordinarily would oscillate with continuous sine-wave output at the natural resonant frequency of the transformer, depending on its inductance and stray capacitance. However, the transformer has a low  $Q$  and the  $R_g C_c$  time constant of the grid circuit is made long enough to allow the grid-leak bias to block the tube, cutting off plate current. The tube remains cut off until the grid coupling condenser  $C_c$  can discharge through the grid resistor  $R_g$  to the point where the grid bias voltage is less than cutoff and plate current can flow again to provide feedback signal for the grid. As a result, the circuit

operates as an intermittent or blocking oscillator. A typical blocking oscillator transformer is shown in Fig. 17-8.

One sine-wave cycle of high amplitude is produced at the blocking rate, as illustrated in Fig. 17-9. The succeeding sine waves shown in dotted lines do not have enough amplitude to overcome the grid-leak bias blocking the oscillator. The number of times per second that the oscillator blocks and produces the high-amplitude pulse is the *pulse repetition frequency*. This frequency is determined by the  $R_g C_c$  time constant. For a horizontal deflection oscillator the values of  $R_g$  and  $C_c$  allow the blocking oscillator to operate with a pulse repetition frequency of 15,750 cps; a vertical blocking oscillator operates at 60 cps.

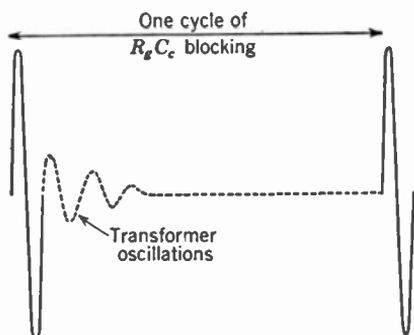


FIG. 17-9. Output of blocking oscillator. Each cycle consists of a high-amplitude pulse at the blocking rate followed by several low-amplitude oscillations.

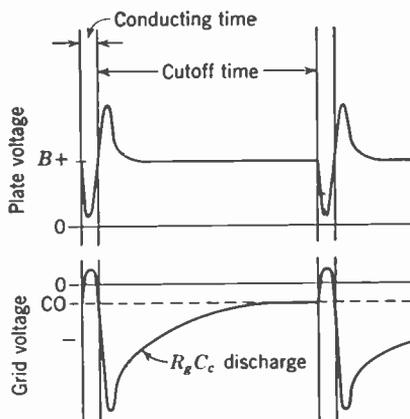


FIG. 17-10. Plate-and-grid-voltage waveforms of blocking oscillator.

The total grid voltage of the blocking oscillator, with the pulses and grid-leak bias, is shown in Fig. 17-10. Note that the blocking oscillator grid voltage is exactly the wave shape needed to operate the discharge tube. The narrow positive pulses drive the discharge tube into conduction, allowing the saw-tooth condenser voltage in the plate of the discharge tube to drop sharply for the flyback. The negative grid voltage beyond cutoff keeps the plate current of the discharge tube cut off for a relatively long time between pulses so that the saw-tooth condenser voltage can rise toward the B supply voltage for the linear rise on the saw-tooth wave. The frequency of the saw-tooth voltage output from the plate circuit of the discharge tube is the same as the blocking oscillator frequency. Also, the flyback time on the saw-tooth wave corresponds to the width of the pulse from the blocking oscillator, since the saw-tooth condenser can discharge only while the blocking oscillator pulse makes the discharge tube conduct.

*Blocking Oscillator Action.* In analyzing the operation of this circuit, the waveform of instantaneous grid-voltage variations can be considered in two parts. One is the feedback signal voltage developed across the transformer secondary by a change in primary current. The other is bias voltage developed when the feedback signal drives the grid positive to cause the flow of grid current and charging of the grid condenser. This bias voltage can increase rapidly as grid current flows through the tube to charge the condenser when the positive signal voltage increases, but must decrease relatively slowly as the condenser discharges through the grid-leak resistance when the feedback signal voltage decreases. The instantaneous grid voltage at any instant is equal to the sum of the bias and the signal drive voltage across the secondary of the feedback transformer. The grid signal voltage caused by feedback can drop to zero instantaneously when the feedback ceases, but the bias cannot.

The cycle of operations can be followed from the time power is applied. Plate current flows immediately because, initially, there is no bias in the grid-leak bias arrangement. Since the plate current flows through the primary of the coupling transformer, this increase in current from zero induces a voltage across the secondary which is coupled to the grid circuit. The windings are poled so that this feedback voltage drives the grid positive, drawing grid current and developing a bias approximately equal to the signal swing. As the plate current increases, positive signal voltage is induced across the secondary to maintain the grid voltage at approximately zero. It must be noted now that the voltage induced across the secondary of the transformer is dependent upon the rate of change of the flux produced by the primary current. As soon as the rate of change of the expanding magnetic field decreases, the result either of saturation of the transformer or of plate-current saturation, the positive feedback signal voltage decreases in magnitude. Since the bias cannot change instantaneously, the instantaneous grid voltage is now more negative, decreasing the plate current. With the decrease in plate current the transformer's magnetic field collapses, inducing a voltage across the secondary of opposite polarity from the feedback obtained when the plate current was increasing, and the grid is driven more negative. The effect of decreasing the plate current is thus amplified to cut off the flow of plate current very rapidly, and a large negative swing of signal voltage is developed across the secondary by the rapidly collapsing field, producing the extreme negative voltage beyond cutoff on the grid waveform shown in Fig. 17-10.

When plate current ceases there is no feedback signal. The grid voltage then consists only of the bias, which declines in value, following the typical exponential  $RC$  discharge curve. After a period of time that depends on the  $R_g C_g$  grid time constant, the bias is reduced to a value

equal to the grid cutoff voltage, allowing plate current to flow again, and the complete cycle is repeated. The time consumed by the rise and decay of plate current, which is equal to the width of the positive grid pulse, is determined by the inductance and capacitance of the transformer. The time between pulses is determined by the grid  $R_g C_g$  time constant.

**Blocking Oscillator and Discharge Tube Circuits.** Figure 17-11 shows a saw-tooth generator circuit for horizontal deflection, using a twin triode for the blocking oscillator and the discharge tube. The 820- $\mu\mu\text{f}$  grid coupling condenser  $C_c$  and the variable 50,000-ohm grid resistor  $R_g$  allow the oscillator to operate at the horizontal line frequency of 15,750 cps. The grid of the oscillator is connected directly to the grid of the discharge tube. Therefore, the oscillator and discharge tube grid voltage consists of short positive pulses followed by relatively long periods of cutoff.

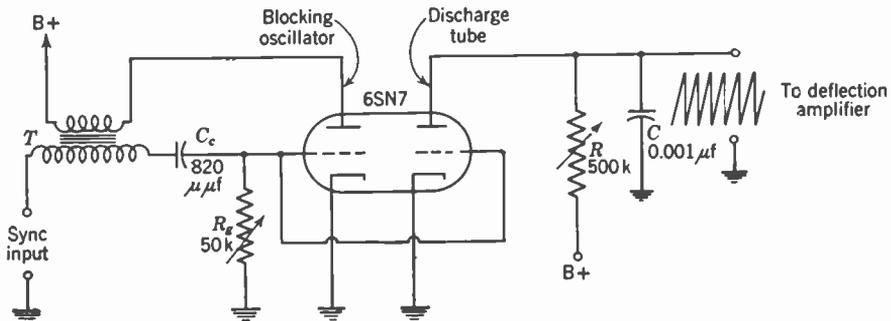


FIG. 17-11. Blocking oscillator and discharge tube circuit. Component values given are for horizontal deflection oscillator, but the same circuit can be used with different components for vertical deflection oscillator.

The 0.001- $\mu\text{f}$  condenser  $C$  between plate and cathode of the discharge tube is the saw-tooth condenser. As the blocking oscillator goes through a cycle,  $C$  charges toward the B supply voltage through the series resistor  $R$  for the linear rise on the saw-tooth wave while the discharge tube is held cut off by the oscillator grid voltage. When the oscillator conducts, developing a sharp positive pulse of grid voltage, the discharge tube also conducts. Then the saw-tooth condenser discharges rapidly through the low resistance of the discharge tube and the voltage across  $C$  drops rapidly to produce the flyback on the saw-tooth wave. The saw-tooth voltage output is then coupled to the horizontal deflection amplifier. Although this diagram illustrates a horizontal deflection generator, the same circuit can be used for the vertical deflection generator, with different values of the components to operate at the field frequency of 60 cps.

Referring to Fig. 17-11, note that the grid and cathode voltages of the discharge tube are the same as in the blocking oscillator. Therefore, the blocking oscillator itself can function as the saw-tooth generator by

connecting the saw-tooth condenser in the plate of the oscillator. Figure 17-12 shows the diagram of a blocking oscillator and discharge tube combined in a single triode. The values of the components shown are for a vertical deflection oscillator but the same circuit can be used as a horizontal deflection oscillator with different values. The stage is essentially a blocking oscillator. However, the saw-tooth condenser  $C$  is in the oscillator plate circuit instead of using a separate discharge tube. Since the oscillator grid voltage pulses the tube into conduction for short periods of time, followed by relatively long periods of cutoff, the voltage across  $C$  is a saw-tooth wave of voltage at the frequency of the oscillator. While the tube is cut off by its blocking action, the saw-tooth condenser charges through the plate load resistor  $R$  toward the B supply voltage.  $R$  is made variable to adjust the amplitude of the saw-tooth voltage output. When

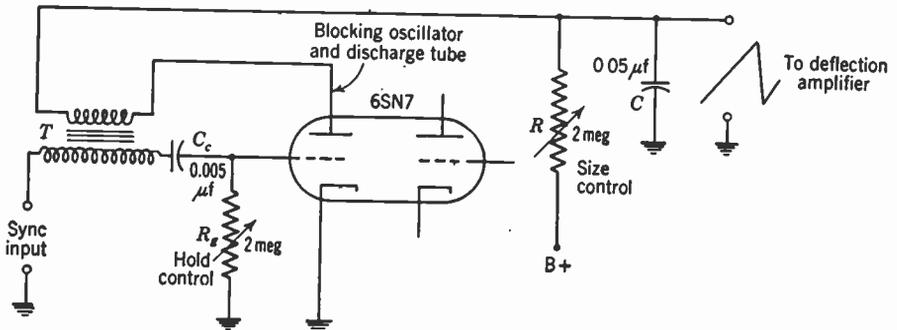


FIG. 17-12. Blocking oscillator and discharge tube circuit combined in one stage. Component values given are for vertical deflection oscillator, but same circuit can be used with different components for horizontal deflection oscillator.

plate current flows during the positive grid pulse,  $C$  discharges from cathode to plate through the tube and through the plate winding of the transformer. The discharge current of  $C$  is in the same direction as normal plate current during oscillator conduction. The frequency of the saw-tooth voltage output is the frequency of the blocking oscillator, which depends upon the grid circuit  $R_g C_c$  time constant.  $R_g$  is made variable to adjust the frequency. This type of blocking oscillator and discharge tube circuit is commonly used because it eliminates the additional stage required for a separate discharge tube. However, the frequency and amplitude controls are more independent of each other with separate stages for the blocking oscillator and discharge tube, and the amount of saw-tooth voltage output is greater.

**17-4. Synchronizing the Blocking Oscillator.** The blocking oscillator is easily synchronized by small positive pulses injected in the grid circuit to trigger the oscillator at the frequency of the synchronizing pulses. The sync is applied in series with the grid winding of the transformer so

that the positive synchronizing voltage can cancel part of the negative grid voltage produced by the oscillator. The synchronizing action is illustrated in Fig. 17-13. The positive synchronizing pulses arriving at the times marked *S* in the figure, when the declining negative grid voltage is just approaching cutoff, will be enough to drive the grid voltage momentarily above cutoff. As soon as plate current starts to flow, the oscillator goes through a complete cycle. With another positive pulse of synchronizing voltage applied at a similar point of the following cycle, the oscillator again begins a new cycle at the time of the pulse. As a result, the synchronizing pulses force the oscillator to operate at the sync frequency.

The frequency of the oscillator without any synchronizing voltage is called the *free-running frequency*; the synchronized oscillator frequency is the *forced frequency*. The free-running frequency of the oscillator

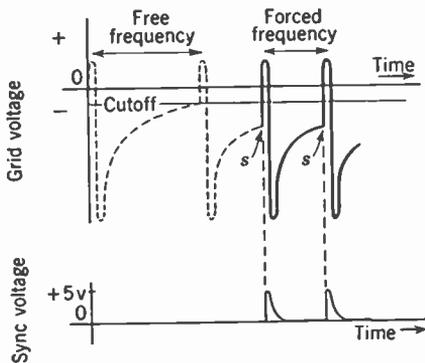


FIG. 17-13. Synchronizing the blocking oscillator.

must be set lower than the synchronizing frequency, so that the sync pulses will drive the grid voltage in the positive direction at the time when the oscillator is ready for triggering. This is the time when the grid bias has declined practically down to cutoff by itself, and needs only a slight additional positive voltage to start the flow of plate current and the beginning of a cycle. A positive synchronizing pulse that occurs in the middle of the oscillator cycle must have a much higher value to drive the grid voltage to cutoff. The peak negative swing of the grid-voltage wave may be more than 200 volts, but toward the end of the cycle a few volts of positive synchronizing voltage can be enough to trigger the oscillator and lock it in at the synchronizing frequency.

Synchronizing voltage of negative polarity in the grid cannot trigger the blocking oscillator. Also, the oscillator cannot be triggered if the free frequency is slightly higher than the synchronizing frequency, because then the synchronizing pulses will occur after the oscillator has started to conduct by itself and they will have no effect. Operating the oscillator at the

same frequency as the synchronizing pulses does not provide good triggering because the oscillator frequency can drift above the sync frequency, resulting in no synchronization for part of this time. For best synchronization, the free-running oscillator frequency is adjusted slightly lower than the forced frequency, so that the time between sync pulses is shorter than the time between pulses of the free-running oscillator. Then each synchronizing pulse occurs just before an oscillator pulse and forces the tube into conduction, thereby triggering every cycle to hold the oscillator locked in at the sync frequency.

The triggering action can be made less sensitive to noise pulses by applying a small positive voltage to the blocking oscillator grid, instead of returning it to chassis ground. With the added positive voltage on the grid, the negative bias declines to cutoff sooner and approaches the cutoff voltage with a sharper slope, as a smaller part of the  $R_g C_c$  discharge curve is used. Because of the sharper slope of the declining negative grid voltage, an interfering noise pulse occurring just before the synchronizing pulse must have much more amplitude to trigger the oscillator.

**17-5. Multivibrators.** The multivibrator is frequently used in television as a saw-tooth generator. The basic multivibrator circuit is shown in Fig. 17-14. This is a two-stage resistance-coupled amplifier with the output of the second stage coupled back to the input of the first stage. Since each amplifier produces a  $180^\circ$  phase inversion of its input signal, the feedback voltage is in the same direction as the input grid signal and oscillations can take place. The multivibrator oscillator is used in many applications because it is a compact, economical circuit that is easily synchronized and can produce various useful waveforms. Its many uses include audio oscillator, frequency divider, saw-tooth generator, electronic switch, and square-wave generators.

*Plate-coupled Multivibrator.* When supply voltages are applied to the multivibrator of Fig. 17-14, plate current begins to flow in both tubes and the circuit immediately begins oscillating. The amount of plate-current flow cannot be identical for the two tubes even if both use the same supply voltage and have plate load resistors of the same size. No matter how small this difference in plate current may be, it is immediately amplified to produce the result of one tube conducting while the other is cut off. Assume that tube 1 in Fig. 17-14 conducts slightly more than tube 2 when plate voltage is applied, driving the grid of tube 2 slightly more negative. This negative signal is amplified and inverted to provide feedback that drives the grid of tube 1 more positive, which in turn allows the first tube to drive the grid of tube 2 still more negative. The amplification of the unbalance in the stage takes place almost instantaneously to drive the grid of tube 2 to cutoff immediately. The tube remains cut off for a period of time that depends upon the  $R_g C_g$  grid time constant.

As the coupling condenser discharges through the grid resistor, the grid voltage declines towards zero. As soon as the grid voltage is reduced to less than cutoff bias, plate current begins to flow and the other tube is driven to cutoff. Thus, the slight initial unbalance sets up a regenerative switching action with first one tube conducting and then the other. One tube conducts for a period equal to the time during which the other is cut off.

In analyzing the plate and grid waveforms shown in Fig. 17-15, the following fundamental properties of vacuum-tube circuits are reviewed as an aid in following the multivibrator action:

1. When the grid is driven in the positive direction, plate current increases and the plate-to-cathode voltage decreases because of the voltage drop across the plate load resistor, which is in series with the tube

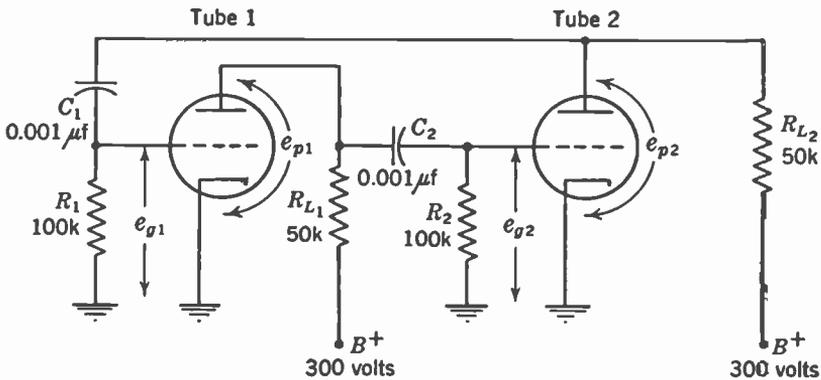


FIG. 17-14. Basic multivibrator circuit.

across the B supply. With the grid driven negative, plate current decreases and the plate-to-cathode voltage increases. When the tube is cut off, its plate-to-cathode voltage is equal to the B supply voltage, since there is no  $IR$  drop across the plate load.

2. When the plate-to-cathode voltage decreases, the plate signal voltage is in the negative direction because it is going less positive. The decrease in voltage across the  $RC$  grid coupling circuit allows the coupling condenser to discharge, producing a negative signal across the grid resistor. When the plate-to-cathode voltage increases, the coupling condenser charges from the B supply, producing a positive signal voltage across the grid resistor.

3. The coupling condenser cannot charge or discharge instantaneously, but is limited to a rate determined by the time constant of the capacitance and series resistance. When charging voltage is instantaneously applied across the  $RC$  circuit or instantaneously removed, the entire value of the

change in applied voltage must appear instantaneously across the resistor. The voltage across the resistor then declines in value as the condenser charges or discharges.

Beginning at time *A* in Fig. 17-15, tube 2 has just been cut off because of conduction in tube 1. Assume that 5 ma of plate current flows through the 500,000-ohm plate load resistor of the first tube, producing a 250-volt drop in plate voltage. Since the plate-to-cathode voltage applied across the  $R_2C_2$  coupling circuit is reduced abruptly from 300 volts to 50 volts,  $C_2$  must discharge. Instantaneously, the entire 250-volt drop in applied voltage is developed across  $R_2$ , with the grid side negative, cutting off tube 2.

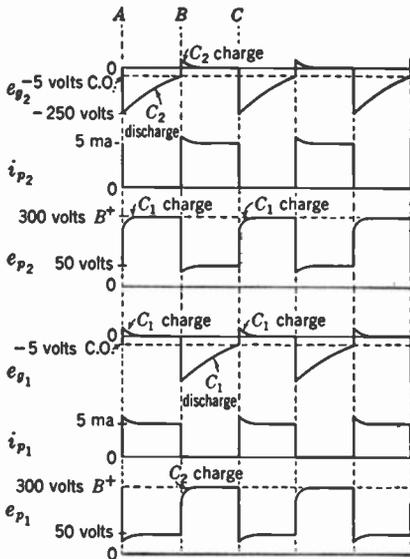


FIG. 17-15. Plate- and grid-voltage waveforms for the multivibrator in Fig. 17-14.

As the coupling condenser discharges, the voltage across  $R_2$  declines toward zero with the typical  $RC$  discharge curve shown. When the grid voltage is down to cutoff, tube 2 begins to conduct plate current. The resultant decrease in plate voltage in tube 2 drives the grid of the first tube more negative, reducing its plate current and increasing the plate voltage. This drives the grid of tube 2 more positive, further reducing the plate voltage and allowing tube 2 to drive tube 1 still more negative. This amplification of the unbalance in the stage reverses the action of the two tubes almost instantaneously, with tube 1 now cut off and tube 2 conducting, as shown at time *B* in Fig. 17-15. The plate-

to-cathode voltage of the first tube, now cut off, rises immediately to the *B* supply voltage, driving the grid of tube 2 positive. The coupling condenser  $C_2$  charges rapidly through the low resistance of the grid-to-cathode circuit of tube 2, and the grid voltage is reduced to zero very soon as the coupling condenser becomes completely charged. The grid voltage for tube 2 remains at zero and zero-bias plate current flows in tube 2 as long as tube 1 remains cut off. Meanwhile, the coupling condenser for the first tube,  $C_1$ , is discharging through its grid resistor  $R_1$  and the negative grid voltage of tube 1 declines toward zero. When cutoff voltage is reached at time *C* in the illustration, conduction begins again in the first tube, cutting off tube 2 again to repeat the cycle. The waveforms for both tubes are exactly the same but displaced in time by  $180^\circ$ , since one tube is conduct-

ing while the other is cut off. The period of conduction for either tube is equal to the cutoff time of the other tube. It is the change from cutoff to conduction that initiates the switching operation.

The output voltage from either plate is a symmetrical square wave of voltage as the plate voltage rises sharply to the B supply voltage, remains at that value for a period of time equal to the cutoff time, and then drops sharply to some low value resulting from plate-current flow. The slight departure from square corners is caused by charging of the coupling condensers. At the time *A* in Fig. 17-15, the coupling condenser  $C_1$  charges to the B supply voltage because of cutoff in the second tube. The charge path is from B- to the grid side of the condenser through the low resistance of the cathode-grid circuit as the grid is driven positive and the grid current flows and from the plate side of the coupling condenser through the plate load resistor of the previous tube to the B supply. The resistance of this path is relatively low; the time constant is short; and  $C_1$  rapidly charges to the supply voltage, reducing the grid voltage of tube 1 to zero. During the time  $C_1$  is charging, the charging current flows through the plate load resistor of tube 1 in the same direction as plate current; although the tube is cut off, the plate voltage cannot rise to the supply voltage until the coupling condenser is completely charged. Also, during the time  $C_1$  is charging, the grid voltage for tube 1 is slightly positive and its plate current is more than the zero-bias plate current, reducing the plate voltage below the value obtained with zero grid voltage after the coupling condenser is completely charged. The coupling condenser  $C_2$  charges at the time *B* in the figure, producing the same effects on the wave shapes. On discharge, the coupling condenser discharges through its grid resistor and the cathode-to-plate circuit of the previous stage, which is conducting.

The time from *A* to *C* is one complete cycle, including a complete flip-flop of operating conditions. The frequency may have an approximate range from 1 to 100,000 cps, depending primarily on the  $RC$  time constant of the grid coupling circuits. The period of one cycle is exactly equal to the sum of the cutoff periods of both tubes. While proportional to the  $RC$  time constant, the period during which one tube is cut off also depends on the amount of grid voltage required to cut off plate current in the tube, the value of any additional bias used, and the amount of negative grid voltage produced by the drop in plate voltage of the preceding tube when it conducts. The amount of negative swing produced by the drop in plate voltage is equal to the  $i_p R_L$  voltage drop. If these values are known, the cutoff period for each tube can be calculated as the time it takes the grid voltage to decline from its maximum negative value to the value required to initiate plate-current flow, and the total time of one cycle is equal to the sum of the cutoff periods for both tubes. The fre-

quency in cycles per second is equal to the reciprocal of the period in seconds.

*Cathode-coupled Multivibrator.* The multivibrator circuit in Fig. 17-16 is cathode-coupled because the coupling for feedback from tube 2 to tube 1 is obtained by means of the common cathode resistor  $R_k$ . Tube 1 drives tube 2 with the usual grid coupling circuit  $R_2C_2$ . Therefore, tube 2 can be cut off by conduction in tube 1 as  $C_2$  discharges through  $R_2$  because of the first tube's drop in plate voltage, just as in the conventional plate-coupled multivibrator. Tube 1 is cut off by the cathode bias voltage produced across  $R_k$  when tube 2 conducts, since the plate current for both tubes flows through  $R_k$ . The circuit oscillates as a free-running multivibrator with first one tube cut off and then the other.

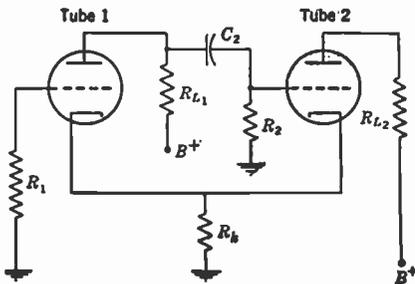


FIG. 17-16. Cathode-coupled multivibrator, using direct coupling between the cathodes. The circuit can also be arranged with separate cathode resistors and capacitive coupling between the two cathodes.

When plate voltage is applied, both tubes start to conduct. The flow of plate current in tube 1 reduces its plate voltage, driving the grid of tube 2 negative. The plate current in tube 2 is reduced because of the negative grid signal, decreasing the voltage across  $R_k$ . This allows tube 1 to conduct more plate current, driving the grid of tube 2 more negative, and the unbalance is amplified to drive tube 2 to cutoff almost instantaneously. Tube 2 is held cut off during the time that the grid coupling condenser  $C_2$  discharges through  $R_2$ ,  $R_k$ , and the plate-to-cathode resistance of tube 1. The negative grid voltage across  $R_2$  declines exponentially with the normal condenser discharge curve until the grid voltage for tube 2 has been reduced to a value that allows plate-current flow. With the plate current of tube 2 now flowing through the common cathode resistor, the cathode bias for tube 1 is increased, driving its grid negative. Plate current in tube 1 is reduced because of the additional cathode bias, allowing its plate voltage to rise toward the B supply voltage. The increase in plate voltage drives the grid of tube 2 more positive as  $C_2$  charges from the B supply through the grid-to-cathode circuit of tube 2 and  $R_{L1}$ . With tube 2 driven more positive its plate current increases, and more bias is developed across  $R_k$  as the cathode voltage follows the applied grid voltage. The action is cumulative and results in tube 1 being cut off almost instantaneously by the plate current of tube 2. Tube 1 is held at cutoff for a period of time that depends on how long it takes the coupling condenser  $C_2$  to charge.

As  $C_2$  becomes charged, the grid of tube 2 becomes less positive and the plate current of tube 2 is decreased.

It should be noted now that the grid of tube 1 is at ground potential, since there is no coupling circuit to provide grid signal for the first tube. Therefore, tube 1 remains cut off as long as the cathode bias exceeds its cutoff voltage. When the cathode voltage drops below cutoff because of decreasing plate current in tube 2 as  $C_2$  charges, tube 1 can conduct again and immediately cuts off tube 2 to repeat the cycle. Thus, the circuit oscillates as a free-running multivibrator as each tube alternately conducts to cut off the flow of plate current in the other tube. The period of

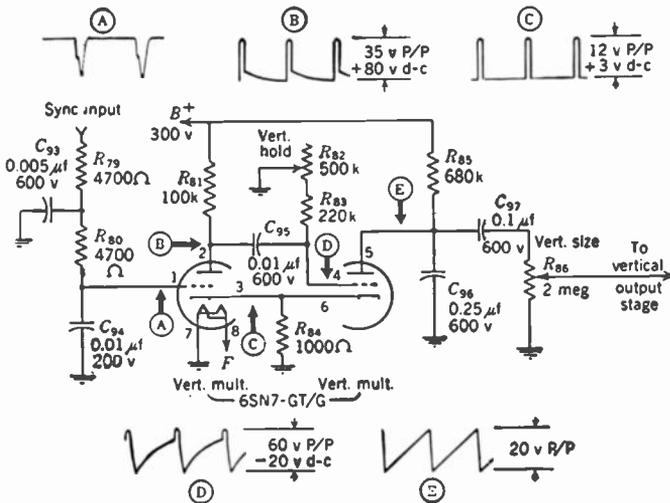


Fig. 17-17. Vertical deflection oscillator using the cathode-coupled multivibrator, with typical wave shapes. (Belmont Radio Corporation.)

time during which tube 2 is cut off depends upon the time constant on discharge for the coupling condenser  $C_2$ . The period of cutoff for tube 1 depends upon the time constant on charge for  $C_2$ , and this can be made very much shorter than the discharge time to provide unsymmetrical output from the multivibrator. The reason why tube 2 is not cut off when the increased voltage drop across  $R_k$  cuts off tube 1 is that this is the time when the grid of tube 2 is being driven positive because of decreased plate current in the first tube.

**17-6. Multivibrator Saw-tooth Generator.** An unsymmetrical multivibrator can be used as a saw-tooth generator by connecting a saw-tooth condenser across the plate-to-cathode circuit of the tube that is cut off for a relatively long period of time and conducts for a short time. Figure

17-17 shows a typical unbalanced cathode-coupled multivibrator circuit which is often used as a deflection generator in television. The voltage wave shapes at points *A*, *B*, *C*, *D*, and *E* in the circuit are also shown.

The saw-tooth generator action in this unsymmetrical multivibrator is much the same as in the blocking oscillator and discharge tube circuit, with tube 1 corresponding to the blocking oscillator and tube 2 to the discharge tube. The saw-tooth condenser is  $C_{96}$ . While negative signal from the first tube holds the second tube cut off for a relatively long time,  $C_{96}$  charges to the B supply voltage through  $R_{85}$  for the linear rise on the saw-tooth voltage output. When the grid of the discharge tube is driven

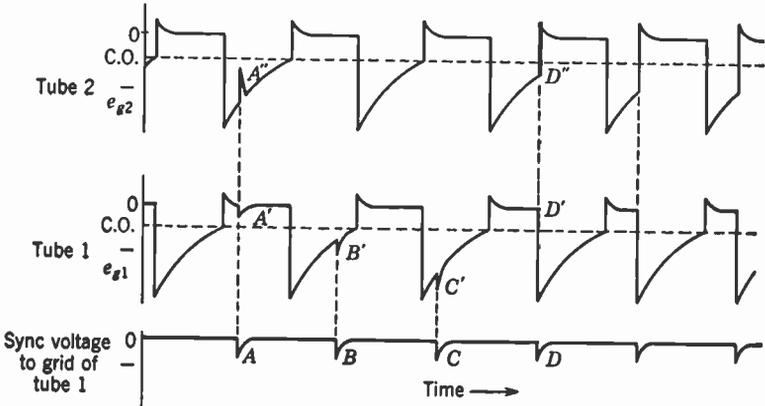


FIG. 17-18. Waveforms of grid voltage in multivibrator synchronized by negative pulses. The oscillator is pulled in to the synchronizing frequency at time *D*.

positive,  $C_{96}$  discharges through  $R_{84}$  and the low resistance of the plate-to-cathode circuit of the triode for the rapid flyback. The variable resistor  $R_{86}$  in the grid circuit of the vertical output stage is the vertical height control varying the amplitude of the signal input to the deflection amplifier. The frequency of the oscillator is varied by means of the variable grid resistor  $R_{82}$ , which is the vertical hold control. Decreasing the grid resistance reduces the period of cutoff for tube 2 to increase the free frequency of the oscillator, and increasing  $R_{82}$  lowers the frequency. The multivibrator can be accurately synchronized, and the hold control is adjusted to set the free frequency below the frequency of the synchronizing pulses so that the oscillator can be forced to lock in at the synchronizing frequency. While the deflection generator shown here is for vertical scanning, the multivibrator is also used as a horizontal scanning generator.

**17-7. Synchronizing the Multivibrator.** The multivibrator can be synchronized with trigger pulses of either positive or negative polarity. A positive triggering pulse applied to the grid of a nonconducting tube can cause switching action to take place if the pulse is large enough to

raise the grid above cutoff voltage, as in the blocking oscillator. A negative triggering pulse applied to the grid of a conducting tube can synchronize the multivibrator, provided that amplification of the pulse by the conducting tube produces a positive pulse large enough to raise the grid voltage of the nonconducting tube above cutoff. As shown in Fig. 17-18, the negative trigger pulse applied to the grid of tube 1 at time *A* reduces its plate-current flow, but the amplified pulse is not large enough to force tube 2 to conduct and no switching occurs. The negative trigger pulses *B* and *C* are completely ineffective because they are applied to tube 1 while it is cut off. At time *D*, the amplified trigger pulse drives tube 2 from cutoff into conduction, cutting off tube 1 as operating conditions are switched in the two tubes. Note that it is not necessary for the trigger pulse itself to cut off the conducting tube. The negative pulse need only reduce plate current enough to produce a resultant positive pulse having sufficient amplitude to drive the grid of the succeeding tube above cutoff.

Just as in the blocking oscillator synchronization, the natural period of the multivibrator oscillator must be greater than the interval between synchronizing pulses. Then the pulses occur just before the natural switching action would take place, when the oscillator is ripe for triggering. The trigger pulses force the multivibrator to switch earlier in the cycle than it would if free running, and the oscillator is forced to lock in at the synchronizing frequency. The cathode-coupled multivibrator sawtooth generator is usually synchronized with trigger pulses of negative polarity applied to the grid of the tube that does not have the discharge condenser, as shown in Fig. 17-17.

**17-8. Saw-tooth Deflection Current.** In magnetic scanning, deflection of the electron beam is produced by varying the magnetic field associated with the current flowing through the vertical and horizontal deflection coils in the yoke mounted externally on the neck of the cathode-ray tube. For linear scanning, the deflection current must rise linearly with time for the trace and then fall rapidly to zero for the flyback. Therefore, a sawtooth wave of current is required. The voltage applied across the scanning coil to produce this sawtooth current will not be a sawtooth wave because the inductance opposes any change in current.

The voltage required for producing sawtooth current through a pure inductance has the rectangular wave shape shown in Fig. 17-19*b*. The relatively long period of applied voltage produces the linear rise on the sawtooth wave of current. This constant voltage, applied instantaneously across the inductance, allows the current to build toward its maximum value with the same charge curve as the voltage across a condenser. During the relatively short time when the applied voltage is reduced, corresponding to the narrow pulse in the rectangular wave shape, the

current decreases to zero. The inductive voltage must be reduced to zero and reversed in polarity, producing a large pulse of opposite polarity from the charging voltage, in order to provide a rapid flyback on the saw-tooth current wave. In a purely resistive circuit, a saw-tooth applied voltage produces saw-tooth current through the resistance, since there is no lead or lag between the applied voltage and the current. Therefore, the applied voltage required for an inductive circuit containing series resistance that is appreciable when compared to the inductive reactance will be some combination of the rectangular and saw-tooth wave shapes.

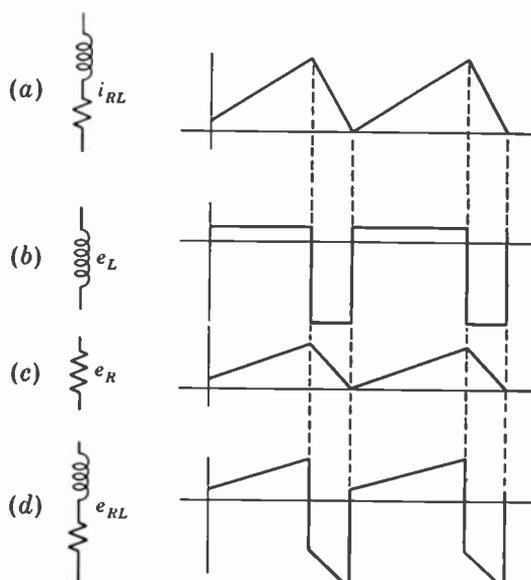


FIG. 17-19. For saw-tooth current, the voltage across an  $RL$  circuit has the trapezoidal wave shape shown in  $d$ . (a) Saw-tooth current through  $RL$  circuit. (b) Rectangular voltage across inductance. (c) Saw-tooth voltage across resistance. (d) Combined voltage across  $R$  and  $L$  in series.

This is shown in  $d$  of Fig. 17-19, which is the sum of waves  $b$  and  $c$ . The combined voltage has a *trapezoidal* waveform and the negative peak is called a *spike*. Note that the linear rise on the trapezoidal voltage wave provides the linear rise of saw-tooth current for the resistive component of the  $LR$  circuit, while the rectangular voltage with the negative peak is for the inductive component.

The circuit used to produce the trapezoidal waveform of voltage is shown in Fig. 17-20. This is the usual circuit for producing saw-tooth voltage across a saw-tooth condenser  $C$  by means of a discharge tube driven with positive pulses, with the addition of a peaking resistor  $R$  in series with the condenser on charge and discharge. The typical value of

4,000 ohms for  $R$  is small compared with the plate load resistor  $R_L$ , and the voltage across the peaking resistor is relatively small when  $C$  is charging from the B supply through  $R_L$ . The long time constant of the charging circuit allows use of only the initial part of the charging curve, providing a charging current that is constant in amplitude. This constant current develops a constant voltage of relatively small magnitude across  $R$ . On discharge, though, the peaking resistance is large enough in comparison with the low resistance of the conducting discharge tube to develop an appreciable voltage. When the discharge tube conducts,  $C$  discharges rapidly, developing a large negative pulse of voltage across  $R$ . The voltages across  $R$  and  $C$  are in series with each other across the plate-to-cathode circuit of the discharge tube, and the output voltage is the sum of the saw-tooth voltage across  $C$  and the peaked voltage across

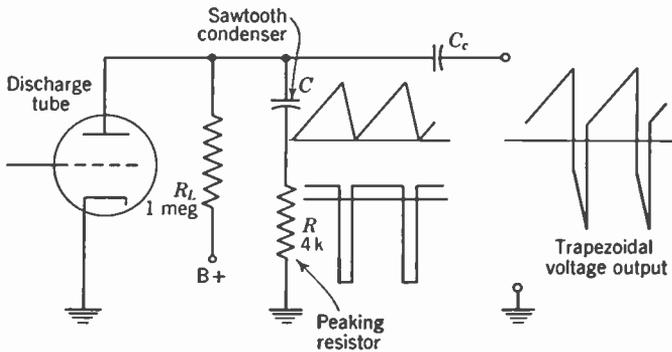


FIG. 17-20. Discharge tube with saw-tooth condenser and peaking resistor to generate trapezoidal voltage wave shape required for saw-tooth current in an  $RL$  circuit.

$R$ . As shown in the figure, the sum of these two voltages produces the trapezoidal voltage waveform required for saw-tooth current through an inductive-resistive circuit.

It should be noted that the peaking resistor provides the voltage peak required for the inductive component of the load circuit. With different values for the peaking resistor, different amounts of peaked voltage can be obtained in the output voltage waveform. Where inductance predominates in the  $RL$  load circuit, a relatively large value of peaking resistance is necessary to provide large voltage peaks. Where the load circuit has more resistance, a smaller voltage peak and smaller peaking resistor are necessary for saw-tooth current in the load. For the extreme case where the load is entirely resistive, no voltage peak is necessary, the required value of peaking resistor is zero, and the discharge tube becomes the usual saw-tooth voltage generator.

The ratio of resistance to inductance in the load circuit for the deflection oscillator depends upon the output tube and the deflection coils.

With an output tube having a high plate resistance,  $R_p$  can be large enough to make the circuit a resistive load for the generator, since the deflection amplifier's internal plate resistance is effectively in series with the inductive plate circuit. Then saw-tooth voltage from the deflection generator will produce saw-tooth current in the scanning coils. A low-resistance tube for the deflection amplifier allows the inductance to predominate and peaking is necessary in the deflection oscillator to produce the trapezoidal voltage waveform required for saw-tooth current in the resistive-inductive circuit. In more practical terms, a pentode or beam power tube for the deflection amplifier has a high plate resistance and, therefore, can produce saw-tooth current in an inductive plate circuit with saw-tooth grid voltage, as the instantaneous values of plate current are relatively independent of plate voltage. A triode deflection amplifier, which has a low  $R_p$ , needs trapezoidal grid voltage for saw-tooth plate current because the instantaneous values of plate current depend upon plate voltage.

**17-9. Deflection Controls.** Variable controls are provided to adjust the amplitude, linearity, and frequency of the saw-tooth deflection current or voltage. The horizontal and vertical amplitude controls adjust the size of the raster to provide the height and width required for the correct aspect ratio of 4:3. Incorrect setting of the width and height controls is shown in Figs. 17-21 and 17-22. The oscillator frequency controls are adjusted to set the free-running frequency of the deflection oscillator slightly below the synchronizing frequency, so that the synchronizing voltage can lock in the oscillator to hold the picture still. For this reason the deflection oscillator frequency control is often called the *hold* control. The linearity controls in the deflection amplifier circuits adjust the linear rise for the trace part of the saw-tooth deflection wave, to eliminate crowding and spreading of the picture information on the scanning raster. The linearity and size controls are usually screw-driver adjustments on the rear apron of the chassis, since they normally require no change after being set up correctly. The vertical and horizontal hold controls are often brought out to the front of the receiver as front-panel operating controls, so that they can be easily readjusted to make the oscillator pull into synchronization and hold the picture still if the sync is temporarily interrupted.

*Height Control.* The variable resistor  $R$  in the plate circuit of the vertical blocking oscillator and discharge tube circuit in Fig. 17-12 illustrates a typical vertical height control. The same arrangement of a variable resistance in series with the vertical saw-tooth condenser can be used for the height control in the multivibrator deflection oscillator. By adjusting the resistance of  $R$ , the time constant of the charging circuit for the saw-tooth condenser is varied to control the amplitude of the voltage rise

on the saw-tooth wave. The peak-to-peak amplitude of the vertical saw-tooth wave determines the amount of vertical deflection on the kinescope screen, which is the height of the raster.

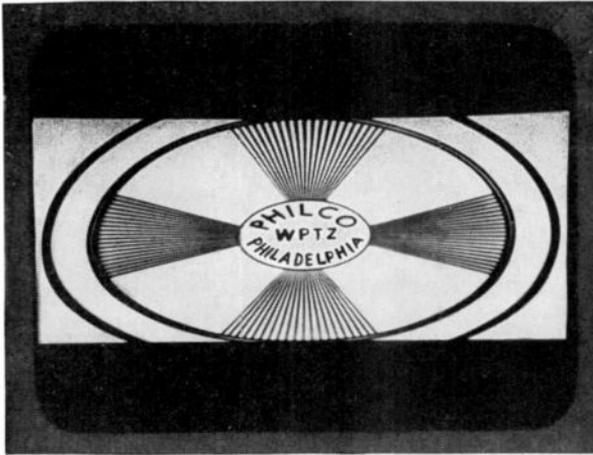


FIG. 17-21. Insufficient height. (*Philco Corporation.*)



FIG. 17-22. Insufficient width. (*Philco Corporation.*)

Changing the time constant of the charging circuit for the saw-tooth condenser changes the amplitude of the output because the rate of charging varies, resulting in a different amount of voltage across the saw-tooth condenser by the time the blocking oscillator pulse produces discharge. This is illustrated in Fig. 17-23. Increasing the resistance of the height control makes the saw-tooth condenser charge more slowly, reducing the amplitude of saw-tooth voltage available before discharge begins. With a shorter time constant, the saw-tooth condenser charges more rapidly to

produce a higher voltage. Note that the frequency of the saw-tooth voltage output is the same as the blocking oscillator frequency because the saw-tooth condenser discharges at the repetition rate of the oscillator pulses. Unless a separate discharge tube is used, however, the oscillator frequency varies slightly when the height control is adjusted, as the change in plate voltage results in a different value of grid cutoff voltage for the oscillator.

The size and linearity of the saw-tooth voltage output depend upon the  $RC$  time constant of the saw-tooth condenser and charging resistance in the plate circuit of the discharge tube. Typical values are about  $680 \mu\text{f}$  for the horizontal saw-tooth condenser, with  $0.5$  megohm of charging resistance, providing a time constant of  $340 \mu\text{sec}$ . This is approximately five times longer than the horizontal-line-scanning period of  $63.5 \mu\text{sec}$ , resulting in saw-tooth voltage amplitude about one-fifth the  $B+$  voltage, or  $60$  volts peak to peak with a  $B$  supply voltage of  $300$  volts. For a vertical saw-tooth condenser of  $0.05 \mu\text{f}$  capacitance and  $2$  megohms of

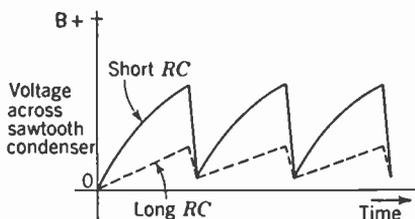


FIG. 17-23. Effect of varying the  $RC$  time constant on charge for the saw-tooth condenser. Shorter time constant allows higher amplitude but poorer linearity. Longer time constant allows better linearity but lower amplitude.

charging resistance, the time constant is  $0.1$  sec. This is six times the vertical field-scanning period of  $0.016$  sec, approximately, resulting in saw-tooth voltage amplitude one-sixth the  $B+$  voltage, or  $50$  volts peak to peak with a  $B$  supply voltage of  $300$  volts.

**Width Control.** The width control for varying the amplitude of the horizontal saw-tooth deflection wave to adjust the width of the raster is usually in the output circuit of the horizontal deflection amplifier. The amplitude of the output from the horizontal deflection oscillator is not made adjustable in receivers using the flyback type of high-voltage supply, because it would change the amount of high voltage for the kinescope anode.

**Vertical Hold Control.** This varies the free-running frequency of the vertical deflection oscillator. The frequency can be adjusted by means of a variable grid resistor, as illustrated by  $R_0$  in the blocking oscillator circuit in Fig. 17-12. The same arrangement is used in the multivibrator circuit in Fig. 17-17, where the vertical hold control  $R_{32}$  is a variable resistor in the grid of the triode with the saw-tooth condenser in the plate circuit. In order to synchronize the oscillator, the free-running frequency must be set lower than the synchronizing frequency. The greater the

difference between the two frequencies, however, the more synchronizing voltage required to lock in the oscillator. Therefore, the vertical hold control is adjusted to set the free frequency of the oscillator below 60 cps by the amount that allows the available synchronizing voltage to be just enough to trigger the oscillator. This is done by watching the picture on the kinescope screen. The vertical hold control is adjusted to make the picture stop rolling vertically and hold in frame, as the vertical oscillator is locked in at the 60-cps synchronizing frequency.

*Horizontal Hold Control.* This can be a variable grid resistor to vary the free-running frequency of the horizontal deflection oscillator, similar to the vertical hold control. However, when the horizontal deflection oscillator has automatic frequency control the horizontal hold control may be in the control-tube circuit. In either case, the horizontal hold control is adjusted to make the picture stop tearing apart in diagonal segments and hold horizontally, as the horizontal oscillator is locked in at the 15,750-cps synchronizing frequency.

**17-10. Automatic Frequency Control.** A deflection oscillator triggered by individual synchronizing pulses for each cycle is capable of providing exact synchronization, if there is no noise interference. However, interfering noise pulse voltages of approximately the same amplitude as the sync voltage can be mistaken for synchronizing pulses and trigger the oscillator, especially at the time of the oscillator cycle just before the desired sync pulse occurs, causing loss of synchronization. In order to make the synchronization more immune to noise pulse voltages, a-f-c circuits for the deflection oscillator have been developed. These are generally called *a-f-c*, *flywheel sync*, or *stabilized sync* circuits. Practically all television receivers have a frequency-stabilizing circuit for the horizontal deflection oscillator. Automatic frequency control can be applied to the vertical deflection oscillator also but is not generally used because the filter time constant required to provide a d-c control voltage proportional to the average vertical sync frequency would have to be relatively large and the circuits would take too long to pull into synchronization. In addition, the vertical sync circuit is not so susceptible to interference from noise pulse voltages because of the comparatively large integrating condenser.

The typical arrangement of an a-f-c circuit for the horizontal deflection oscillator is illustrated in Fig. 17-24. The operation can be considered in three steps:

1. Horizontal sync voltage and a fraction of the horizontal deflection voltage from the scanning circuits are coupled into the frequency-comparing circuit, which can produce a d-c output voltage proportional to the difference in frequency or phase between the two input voltages. This d-c control voltage is used to hold the deflection oscillator at the average

sync frequency, instead of triggering each cycle with individual pulses. The greater the difference in frequency between the oscillator frequency and the synchronizing frequency, the larger the control voltage.

2. The d-c control voltage is filtered by an  $RC$  circuit so that rapid noise pulses cannot change the value of the control voltage appreciably.

3. The filtered d-c control voltage changes the frequency of the deflection oscillator by the amount necessary to make the scanning frequency

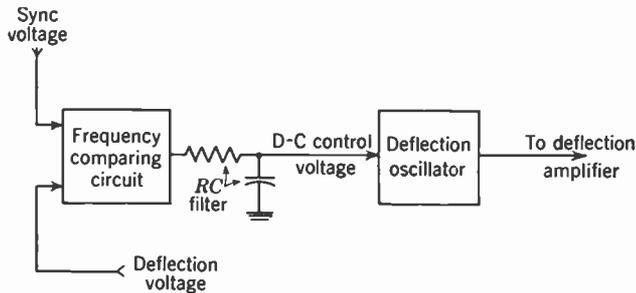


FIG. 17-24. Block diagram of a-f-c circuit for horizontal deflection oscillator.

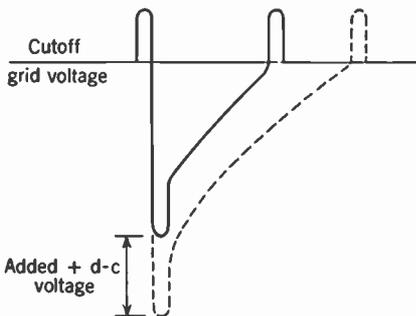


FIG. 17-25. Effect of injecting positive d-c grid voltage to increase the frequency of blocking oscillator, or multivibrator. Dotted wave is lower frequency without the added d-c voltage.

the same as the sync frequency. With a multivibrator or blocking oscillator, the d-c control voltage is coupled directly to the grid of the deflection oscillator to correct its frequency. Making the grid more positive, or less negative, increases the frequency, as illustrated in Fig. 17-25. The time between cycles is shortened when the grid has an added positive d-c voltage, as the grid cannot be driven so far negative and it takes less time for the grid voltage to decay to the cutoff value. When the d-c control voltage makes the grid more negative, the deflection oscillator frequency decreases. The d-c control voltage holds the oscillator at the sync frequency, therefore, as the difference between the oscillator frequency and the synchronizing frequency is measured by the control stage to produce the required amount of correction voltage.

*D-C Control Tube.* The a-f-c circuit in Fig. 17-26, which is generally called *Synchro-Guide*, uses a triode amplifier as a control tube to produce

the d-c control voltage for the grid of a blocking oscillator or multi-vibrator deflection generator. The d-c control voltage is taken as the voltage drop across the cathode resistor  $R_3$ , filtered by  $C_3$ . This cathode voltage is directly coupled to the grid of the deflection oscillator to correct its frequency. The amount of d-c correction voltage produced by the cathode resistor depends upon how much plate current flows. In order to regulate the amount of plate current through the control tube in accordance with the difference between the sync and scanning frequencies,

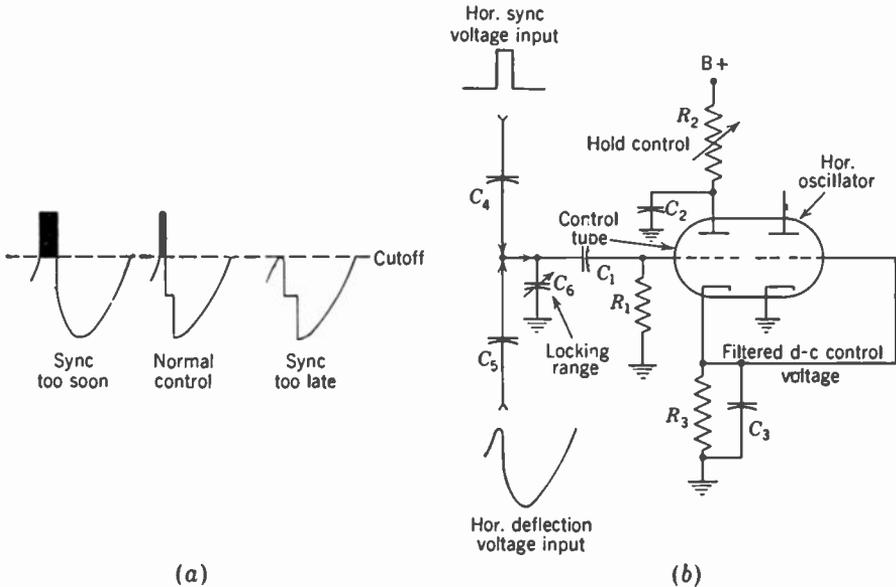


FIG. 17-26. D-C control tube for automatic control of horizontal oscillator frequency. (a) Wave shapes at grid of control tube. The shaded area is effective in producing plate current and control voltage. (b) Circuit arrangement.

the sync and scanning voltages are coupled to the control grid. The horizontal sync voltage from the sync circuits is coupled by the condenser  $C_4$ , while  $C_5$  is the coupling condenser for the scanning voltage taken from the deflection circuit. The deflection voltage at the grid of the control tube has a parabolic wave shape, like distorted half-sine waves, obtained by integrating the horizontal saw-tooth voltage. The parabolic waveform is used because it has a sharper slope than the saw-tooth wave just before the peak corresponding to the start of flyback. In some circuits, a rectangular pulse from the horizontal output circuit may also be coupled to the grid of the control tube, in order to provide a sharper slope just after the peak on the deflection voltage, but this can be omitted when the flyback is fast enough to provide the required slope.

The combined grid voltage for the control tube then consists of the horizontal sync pulses atop the peaks on the deflection voltage. The amount of plate current that flows in the control tube depends upon how much of the sync pulse is on the peak of the deflection voltage. The width of the sync pulse on the peak of the deflection voltage varies with the phase or frequency difference between the two input voltages. As shown by the grid voltage wave shapes in Fig. 17-26a, when the phase between the two input voltages changes, the grid voltage effective in producing plate current varies. As a result, the average plate current and the d-c control voltage across the cathode resistor vary with the phasing of the sync and deflection voltages. With sync and scanning voltages of the same frequency and phase, approximately one-half the width of the sync pulse is on the deflection voltage peak to produce plate current for the required amount of correction voltage. The remaining one-half of the sync pulse occurs after the peak and produces the step shown at the lower part of the deflection voltage. When the sync pulse occurs too soon, because the scanning frequency is slightly lower, more of the sync-pulse width is effective in producing plate current, resulting in a greater positive d-c control voltage from the cathode of the control tube to increase the oscillator frequency. If the scanning frequency is higher than the sync frequency, less control voltage is produced, reducing the oscillator frequency. As a result, the control tube keeps the horizontal deflection oscillator locked in phase with the horizontal synchronizing pulses, as the correction voltage continuously corrects the oscillator frequency.

The horizontal control-tube circuit usually includes the horizontal hold control, which is  $R_2$  in Fig. 17-26, and the *locking-range* control  $C_6$ . The hold control varies the plate voltage and plate current of the control tube, varying the amount of d-c control voltage produced in the cathode circuit, to change the frequency of the oscillator. The locking-range control  $C_6$  forms a capacitive voltage divider with  $C_4$  for the sync voltage input and with  $C_5$  for the deflection voltage input, to adjust the amount of grid voltage applied to the control tube. This determines how far off the deflection oscillator frequency can be from 15,750 cps and still be pulled into synchronization by the control tube.

The control-tube circuit is relatively immune to noise because interfering noise pulses that occur between synchronizing pulses cannot ordinarily cause plate current to flow in the control tube. However, the sync pulse amplitude must be constant so that the control voltage will vary only with a change in phasing between the sync and scanning frequencies. Since the amount of control voltage depends upon the portion of the sync pulse that produces plate current, the Synchro-Guide circuit is often called a *pulse-width, pulse-time, or pulse-area* control system.

*Horizontal Sync Discriminator.* Figure 17-27 illustrates an a-f-c circuit with a sync discriminator, using the twin diode as a balanced detector to produce d-c output voltage proportional to the difference in frequency between the sync and deflection voltages coupled into the discriminator. The d-c control voltage then corrects the frequency of the oscillator, usually a cathode-coupled multivibrator. Push-pull synchronizing pulse voltage from the sync circuits is coupled to the two diodes in opposite polarities.  $R_1C_1$  couples the sync voltage input for diode 1 while  $R_2C_2$  couples the sync voltage into diode 2. In addition, deflection voltage from the horizontal output circuit is applied to the two diodes as the voltage developed across  $C_4$ , in the same polarity for both diodes.

The polarity of the deflection voltage is chosen to make the slope of the flyback voltage increase in the positive direction. As a result, the sync

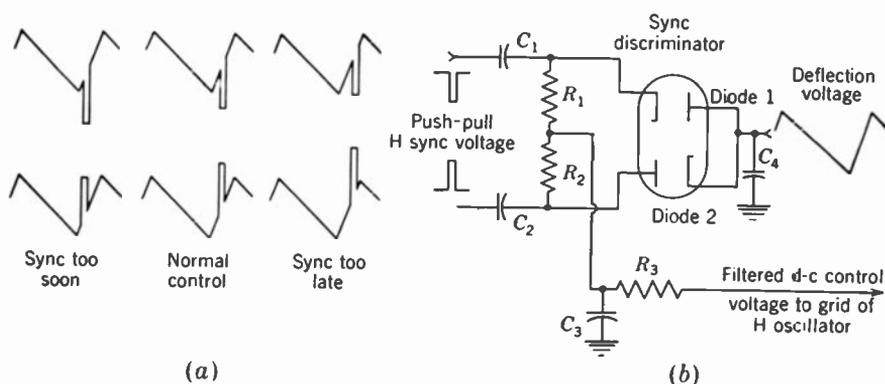


FIG. 17-27. Sync discriminator circuit for automatic control of horizontal oscillator frequency. (a) Wave shapes of the two diode voltages. (b) Circuit arrangement.

pulses combine with the flyback voltage as illustrated by the wave shapes in Fig. 17-27a. Each diode functions as a peak rectifier. When the input voltages for the two diodes have the same peak value, both diodes produce equal output. Since  $C_3$  is in the cathode return circuit to ground of diode 1, it makes the d-c control voltage more positive with respect to chassis ground, while diode 2 makes the control voltage across  $C_3$  more negative. When the sync and scanning frequencies are the same, therefore, the d-c control voltage is zero. If the scanning frequency is too high, the sync pulse will produce more peak voltage for diode 1 and less for diode 2, resulting in positive d-c control voltage to decrease the multivibrator frequency. If the oscillator frequency is too low, diode 2 will produce negative d-c control voltage to increase the frequency. The sync discriminator continuously measures the difference in peak voltage, as a result, to produce the d-c correction voltage that locks in the multivibrator oscillator at the synchronizing frequency. There are usually no controls

to adjust in the sync discriminator stage, as the horizontal hold and locking controls are in the oscillator circuit.

**Sine-wave Stabilizing Tuned Circuit.** In many blocking oscillator and multivibrator deflection generator circuits, an  $LC$  tuned circuit is added to stabilize the oscillator frequency by making the oscillator grid voltage approach cutoff with a sharp slope. Figure 17-28 illustrates the  $LC$  tuned circuit in a blocking oscillator. The multivibrator in Fig. 17-30 has a stabilizing tuned circuit, consisting of  $L_{401}$  and  $C_{418}$ . Referring to Fig. 17-28a, note that the  $LC$  tuned circuit is in the plate and grid circuits of the blocking oscillator. When the oscillator conducts, the plate current shock-excites the tuned circuit to produce sine-wave oscillations.

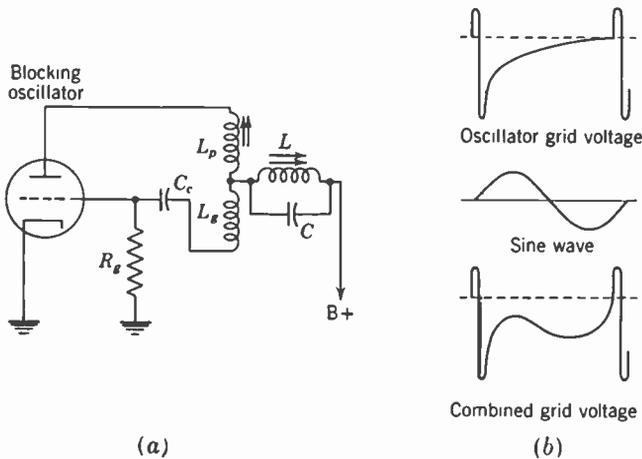


FIG. 17-28. Sine-wave stabilizing tuned circuit in blocking oscillator. (a) Circuit arrangement. (b) Grid waveforms. Dotted-line axis is cutoff voltage.

This sine-wave voltage is coupled to the grid, producing the combined grid-voltage wave shown in Fig. 17-28b. The resultant waveform has a much sharper slope as the grid voltage approaches cutoff because of the sine wave added in the correct phase. The sharper slope as the grid voltage approaches cutoff means that noise pulses or other interfering voltages must have a much higher amplitude to have any effect on the oscillator. As a result, the oscillator is stabilized to minimize frequency changes that might be produced by undesired voltages. The inductance in the stabilizing tuned circuit for the deflection oscillator is sometimes called either a *stabilizing coil* or a *ringing coil*.

**17-11. Horizontal Oscillator Circuits with Automatic Frequency Control.** The schematic diagrams in Figs. 17-29, 17-30, and 17-31 show three common types of horizontal deflection oscillator circuits with automatic frequency control.

*Blocking Oscillator with Stabilized Synchro-Guide.* The circuit in Fig. 17-29 uses a d-c control tube to correct the blocking oscillator, which has a sine-wave stabilizing tuned circuit, in an arrangement generally called *stabilized Synchro-Guide*. One twin-triode tube 6SN7-GT is used for the control tube and the oscillator stages. The triode at the right in the diagram is the blocking oscillator and discharge tube stage, with  $C_{190}$  the saw-tooth condenser in the plate circuit to produce saw-tooth voltage output for the grid of the horizontal output amplifier. Part of the saw-tooth voltage developed across  $C_{190}$  is coupled back from terminal *D* on the oscillator transformer, through  $R_{221}$ , to the grid of the control tube.

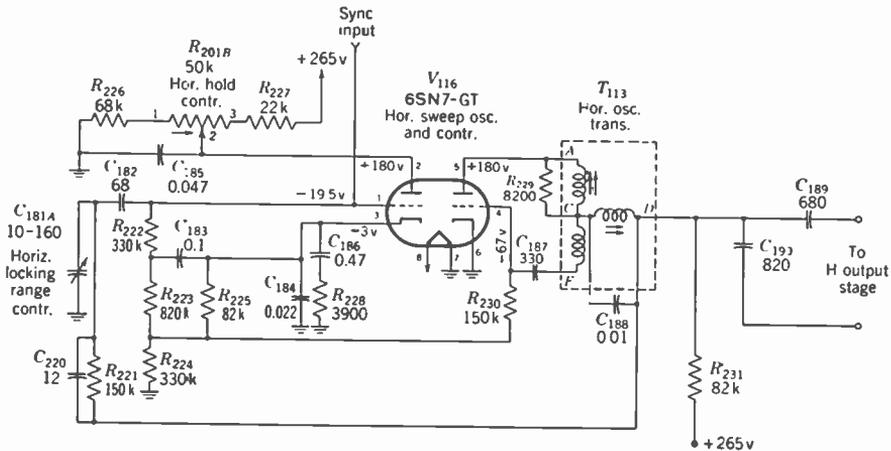


FIG. 17-29. Horizontal blocking oscillator circuit with stabilized Synchro-Guide. Capacitance values more than one in  $\mu\mu\text{f}$  and less than one in  $\mu\text{f}$ ; resistance values in ohms unless otherwise noted. Direction of arrows at controls indicates clockwise rotation. (RCA KCS66 chassis.)

Horizontal sync voltage with positive polarity is also coupled to the grid of the control tube. The phasing of the deflection voltage with respect to the horizontal sync pulses determines how much plate current flows to develop d-c output voltage across the cathode resistors  $R_{224}$  and  $R_{225}$ . The cathode voltage is filtered by  $C_{184}$ , which bypasses both  $R_{224}$  and  $R_{225}$ , and by the series filter  $R_{228}C_{186}$  also across both cathode resistors. The filtered d-c control voltage across the cathode resistor  $R_{224}$  is directly coupled to the oscillator grid circuit to correct its frequency. Negative bias for the control tube to keep it cut off except for sync pulse peaks is obtained from the oscillator grid circuit with  $R_{223}$  serving as a decoupling resistor.

Notice that with the blocking oscillator and control-tube arrangement in Fig. 17-29 the horizontal hold control  $R_{201B}$  is in the control-tube circuit. Therefore, it can vary the horizontal oscillator frequency only

when the control stage is operating. The frequency control for the oscillator is in the transformer  $T_{113}$ , which has an adjustable slug to change the inductance of coil  $A-C$  and vary the free-running frequency of the horizontal oscillator. The adjustable slug in the coil  $C-D$  for the stabilizing tuned circuit is called the oscillator *waveform* adjustment. The locking-range control  $C_{181A}$ , which varies the amount of grid-voltage input to the control tube, is usually a variable mica condenser mounted on the rear apron of the chassis as a screw-driver adjustment. In general, the locking

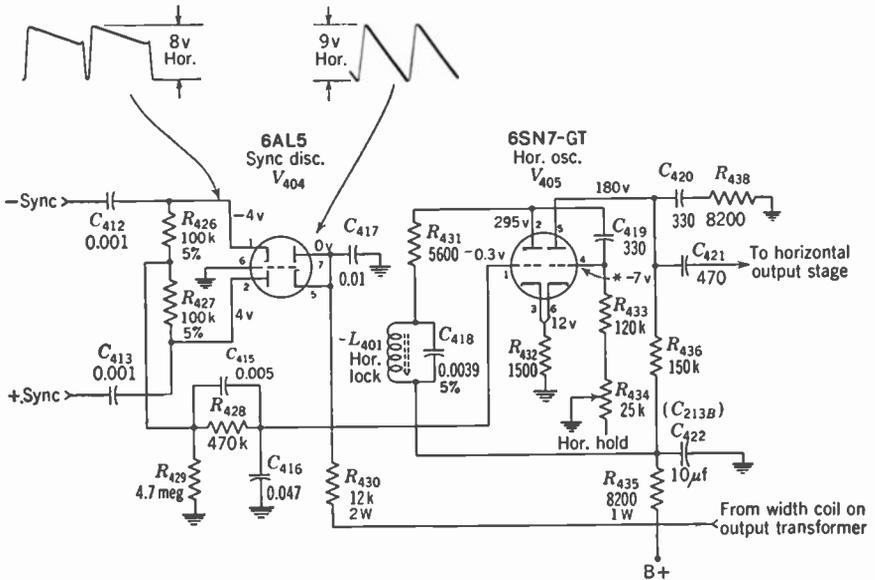


FIG. 17-30. Multivibrator with sync discriminator. Capacitance values more than one in  $\mu\text{mf}$  and less than one in  $\mu\text{f}$ ; resistance values in ohms unless otherwise noted. Voltage at pin 4 of 6SN7-GT varies widely with setting of hold control. (*Admiral Series 21 chassis.*)

range is adjusted to allow the oscillator to be pulled into synchronization within the range of frequencies approximately 180 cps lower than 15,750 cps. This is indicated by three diagonal bars sloping down to the left in the picture just before it pulls into sync.

**Multivibrator with Sync Discriminator.** The circuit in Fig. 17-30 uses the 6AL5 sync discriminator to produce d-c control voltage that corrects the frequency of the 6SN7-GT multivibrator, which is the horizontal deflection oscillator. The voltage pulses across the width coil in the secondary of the horizontal transformer are coupled to pins 7 and 5 of the 6AL5, by  $R_{430}$  and  $C_{417}$ , providing saw-tooth voltage across  $C_{417}$  at the horizontal scanning frequency. At the same time, sync voltages of

equal amplitude and opposite polarity are coupled to pins 1 and 2 of the 6AL5 from the previous sync inverter stage. The two diodes in the discriminator produce d-c voltage proportional to the phasing between the sync and deflection voltages, as a result, since this determines the peak voltages applied to the diodes. The difference between the two diode output voltages is taken from the junction of  $R_{426}$  and  $R_{427}$ , to produce the filtered d-c control voltage across  $C_{416}$ . The correction voltage is directly connected to the synchronizing grid (pin 1) of the multivibrator, to lock in the horizontal deflection oscillator at the synchronizing frequency. The deflection oscillator is a cathode-coupled multivibrator.  $C_{420}$  is the saw-tooth condenser in the plate circuit of the stage cut off for a longer period of time than it conducts, with  $R_{438}$  a peaking resistor to produce trapezoidal voltage output. This is coupled by  $C_{421}$  to the grid circuit of the horizontal output stage.  $L_{401}$  with  $C_{418}$  form the sine-wave stabilizing tuned circuit in the plate circuit of the triode at the left in the diagram, coupled by  $C_{419}$  to the next grid to make its grid-voltage wave shape approach cutoff with a sharper slope.

With the multivibrator and sync discriminator arrangement in Fig. 17-30, the horizontal hold control  $R_{434}$  is a variable resistor in the grid circuit of the oscillator to vary its free-running frequency, independently of the control circuit. The variable inductance  $L_{401}$  in the stabilizing tuned circuit is called *horizontal lock* because it can be adjusted to hold the oscillator synchronized through at least one-half the rotation of the horizontal hold control, so that the picture will not tear apart when changing channels. The horizontal lock control is usually an adjustable slug in the stabilizing coil mounted on the rear apron of the chassis.

*Sine-wave Oscillator with Reactance Tube.* The circuit in Fig. 17-31 for automatic frequency control of the horizontal scanning is generally called *Synchro-Lock*. The features of this circuit are:

1. Instead of a blocking oscillator or multivibrator, the horizontal deflection generator uses the 6K6-GT beam-power pentode in a stable sine-wave Hartley oscillator circuit tuned to 15,750 cps, which drives a separate discharge tube to produce deflection voltage at the horizontal scanning frequency.

2. The frequency of the sine-wave oscillator is controlled by the 6AC7 reactance tube in shunt with the oscillator's tuned circuit. The reactance tube is necessary because the frequency of the Hartley oscillator cannot be corrected effectively just by controlling its d-c grid voltage. Instead, the d-c control voltage is applied to the control grid of the reactance tube to vary the reactance across the tuned circuit of the oscillator.

3. The 6AL5 sync discriminator produces the d-c control voltage that indicates the difference between the sync and scanning frequencies. As a result, the d-c control voltage applied to the grid of the 6AC7 varies its

reactance, which changes the oscillator frequency, correcting the oscillator to lock in the horizontal scanning at the synchronizing frequency.

Horizontal deflection voltage is produced by the separate discharge tube  $V_{120B}$  at the frequency of the oscillator. The oscillator output voltage across the plate load resistor  $R_{200}$  is a distorted sine wave differentiated by  $R_{202}C_{176}$  to produce sharp pulses that drive the grid of the discharge tube positive. The resultant grid-leak bias produced by  $R_{203}C_{177}$  keeps the discharge tube cut off between pulses.  $C_{179}$  in the plate of the discharge tube is the saw-tooth condenser. While the discharge tube is held cut off,  $C_{179}$  charges toward +275 volts through  $R_{204}$ . When the oscillator drives the discharge tube grid positive to make it conduct and

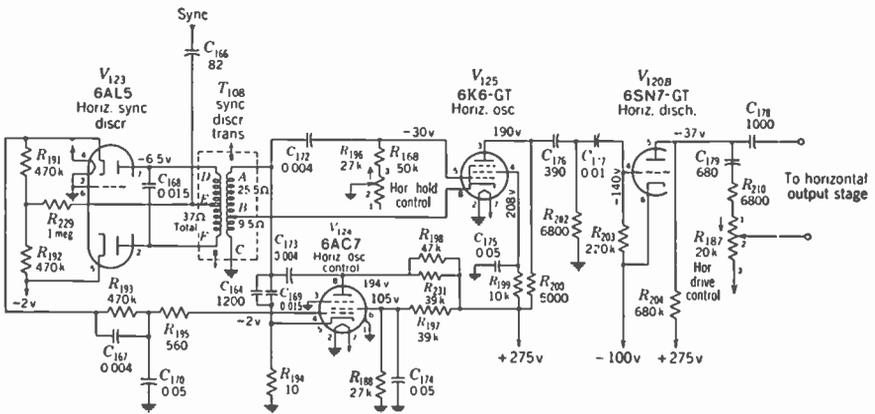


FIG. 17-31. Hartley oscillator with reactance tube. Capacitance values more than one in  $\mu\text{m}$ f and less than one in  $\mu\text{f}$ ; resistance values in ohms unless otherwise indicated. Direction of arrows at controls indicates clockwise rotation. (RCA 630TS and 8TS:30 chassis.)

produce the flyback,  $C_{179}$  discharges through the peaking resistors  $R_{210}$  and  $R_{187}$ , and through the discharge tube. The discharge path returns to the cathode of the discharge tube through the -100-volt line connected to the discharge tube cathode and the cathode circuit of the output stage. The horizontal drive control  $R_{187}$  adjusts the amplitude of the peak in the trapezoidal voltage coupled to the grid of the horizontal output stage.

The winding  $A-B-C$  on the sync discriminator transformer  $T_{108}$  is the tapped coil for the Hartley oscillator. The total capacitance of the parallel condensers  $C_{164}$  and  $C_{169}$  is in series with the 10-ohm resistor  $R_{194}$ , across terminals  $A$  and  $C$  of the coil to provide a tuned circuit resonant at 15,750 cps. The oscillator uses the electron-coupled circuit, with the screen grid as oscillator anode, bypassed by  $C_{175}$  to terminal  $C$  of the oscillator coil, while the cathode connects to the tapped coil at terminal  $B$ . Grid-leak bias for the oscillator is provided by  $C_{172}$  with  $R_{196}$  and

$R_{168}$ . The horizontal hold control  $R_{168}$  varies the resistance of the oscillator grid circuit to change the frequency slightly.

The oscillator coil is closely coupled to the center-tapped discriminator coil, applying equal sine-wave voltages in opposite polarity to the two diode plates of the discriminator, as shown in Fig. 17-32. The horizontal synchronizing pulses are coupled to the center tap, applying the same sync voltage to the two diode plates. Each diode produces d-c output voltage in the cathode circuit proportional to the peak value of the input voltage to the plate circuit. The net output at the top cathode of the 6AL5 is the difference between the voltages across  $R_{191}$  and  $R_{192}$ . When sync voltage and sine-wave voltage are phased correctly, both diodes have the same peak input voltage, resulting in equal d-c output voltages, and the net voltage output is zero. If the phase of the sine wave changes with respect to the sync voltage, more voltage will be applied to one diode and

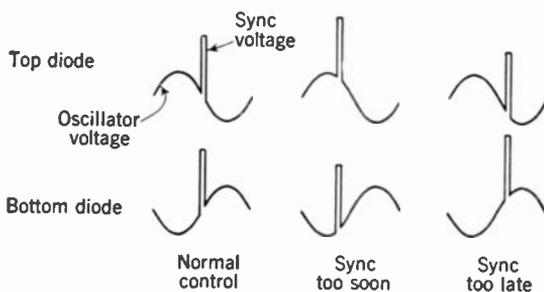


FIG. 17-32. Waveforms for the sync discriminator circuit in Fig. 17-31. (RCA *Synchro-Lock*.)

the net output will be d-c voltage of either positive or negative polarity, depending on which diode has the greater voltage. This d-c control voltage is filtered and applied in series with the fixed bias of  $-2$  volt to the control grid of the reactance tube.

The 6AC7 reactance-tube circuit is basically the same as those shown for frequency modulation in Sec. 7-5. The quadrature network for feedback voltage consists of the parallel combination of  $C_{164}$  and  $C_{169}$  in series with  $C_{173}$  and  $R_{194}$ . The voltage across  $R_{194}$  is the quadrature voltage fed back to the cathode, so that the d-c control voltage from the sync discriminator can be coupled to the control grid. The quadrature voltage leads the plate voltage by  $90^\circ$  in the cathode but its effect on plate current is the same as a  $90^\circ$  lagging angle in control-grid voltage, as the cathode and grid voltages are  $180^\circ$  out of phase. D-C control voltage of positive polarity from the discriminator increases the oscillator frequency; negative d-c control voltage decreases the frequency.

In addition to the horizontal hold control  $R_{168}$  in the oscillator grid circuit, the *Synchro-Lock* circuit has oscillator frequency and phase adjust-

ments. The variable core for coil *A-B-C* in transformer  $T_{108}$  in Fig. 17-31 is the oscillator frequency adjustment. This is adjusted to pull the picture into horizontal synchronization. Then the picture usually remains in sync through the entire rotation of the horizontal hold control and when changing channels. The oscillator phase adjustment is the variable core for coil *D-E-F* in  $T_{108}$ . This adjustment shifts the oscillator frequency slightly to make the horizontal flyback time occur during horizontal blanking. When the blanking bar is in the picture, as shown in Fig. 17-33, the phasing control is adjusted to move the bar toward the right and out of the picture. The frequency adjustment is at the top of the transformer shield can and the phase adjustment at the bottom.

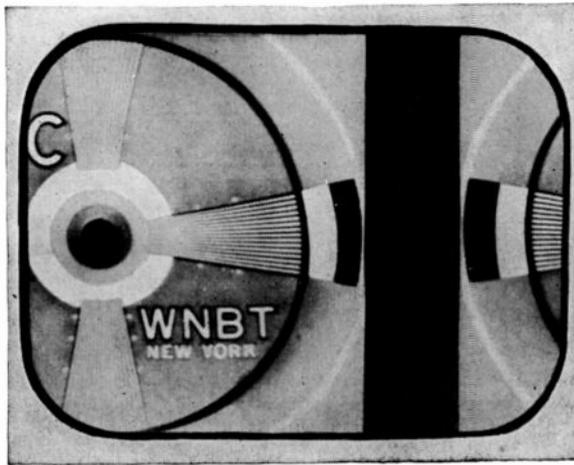


FIG. 17-33. Blanking bar in the picture, caused by incorrect phasing of the horizontal flyback with respect to horizontal blanking time. (RCA.)

**17-12. Horizontal Scanning with Automatic Frequency Control.** The a-f-c circuit provides easy adjustment of the horizontal hold control, stability of the horizontal synchronization against line voltage changes, and good noise immunity. Since the a-f-c filter does not allow the d-c control voltage to vary rapidly enough to produce abrupt changes in frequency, single lines in the picture do not tear out. The picture as a whole may move horizontally during severe noise conditions, without tearing apart. However, when synchronization is lost and the oscillator operates at the incorrect frequency, the picture tears into several diagonal segments as shown in Fig. 16-2 for the case of no horizontal hold. In normal operation, though, the a-f-c circuit can hold the picture in horizontal sync for the weakest usable signal.

*Horizontal Hold Control.* If horizontal synchronization is lost temporarily, the hold control can be readjusted to lock the picture in sync.

The a-f-c circuit generally holds the picture in horizontal sync through one-half the rotation of the horizontal hold control, or more. When the hold control is rotated, though, the horizontal centering of the picture on the raster usually shifts because of a small change in phasing between horizontal flyback time and blanking. The range through which the oscillator frequency can be varied manually by the horizontal hold control and still hold the picture locked in sync is called the *hold-in range* or *lock-in range*. The range of frequencies through which the a-f-c circuit can lock in the oscillator after sync has been temporarily interrupted is the *pull-in range*. The pull-in range is always less than the hold-in range. Therefore, the horizontal hold control should be set at the middle of its hold-in range so that the oscillator can be pulled into sync easily when changing channels.

*Phasing between Horizontal Blanking and Flyback.* While the deflection circuits are scanning the raster, the composite video signal on the kinescope grid is varying the intensity of the electron scanning beam. The blanking on the kinescope screen has the timing of the sync and blanking pulses in the transmitted signal, but the flyback is determined by the deflection circuits in the receiver. Therefore, the position where the black blanking bars appear on the kinescope screen depends upon the phase of the scanning in the receiver's deflection circuits with respect to the blanking pulses in the video signal. In a triggered system, the synchronized flyback starts during blanking time automatically because each sync pulse begins the retrace. With automatic frequency control, however, the horizontal oscillator produces scanning independently of individual synchronizing pulses. As a result, the a-f-c circuit can lock in the oscillator at the synchronizing frequency, but with the horizontal flyback out of phase with horizontal blanking. Then the black bar produced by horizontal blanking voltage is in the picture, during trace time, as shown in Fig. 17-33.

The wave shapes in Fig. 17-34 illustrate several conditions of phasing between horizontal flyback and blanking. Normal phasing of the flyback within blanking time is shown in *a*. In *b*, horizontal flyback starts just before blanking. Then some of the picture information that should be at the extreme right side of the trace in the picture is reproduced during the flyback to the left. Since picture information is reproduced during retrace and trace at the right, this side can appear folded over or under itself, usually brighter than normal. When the horizontal flyback starts too late after blanking, as in *c*, the retrace cannot be completed before picture information starts for the trace at the left side of the next line. Then the left side of the picture appears folded. If the horizontal retrace time is too long, as in *d*, the flyback can start before blanking but still continue after blanking and both the left and right sides of the picture

may be folded. In addition to the folded effect, whenever the flyback occurs during picture information time there may be a light haze in the background, pointing outward from either side of the picture, like a big spear. This is white picture information spread out by the fast flyback, instead of being reproduced correctly during trace time.

*Filtering the D-C Control Voltage.* Variations in the amount of d-c control voltage change the oscillator frequency. The time constant of the filter for the d-c control voltage determines how fast the filtered d-c control voltage can change in amplitude. Therefore, the filtering of the d-c control voltage is important in determining how fast the a-f-c circuit can lock in the oscillator at the synchronizing frequency. A typical value for the a-f-c filter time constant is about 0.01 to 0.05 sec.

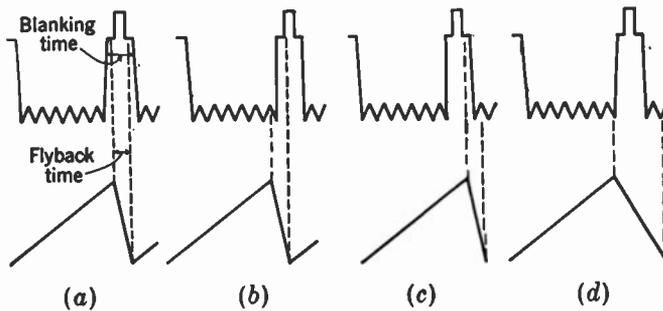


FIG. 17-34. Phasing of flyback and blanking time for horizontal scanning. (a) Flyback within blanking time. (b) Flyback starts too soon. (c) Flyback starts too late. (d) Flyback time too long.

*Hunting in the A-F-C Circuit.* When the filtered d-c control voltage cannot vary fast enough to change in amplitude after it has corrected the oscillator frequency, the a-f-c circuit tends to overcorrect the oscillator. Then the frequency will decrease below 15,750 cps after being too high and increase above 15,750 cps after being too low. Each succeeding step of overcorrection is less until finally the control voltage indicates the correct frequency and the oscillator operates at 15,750 cps. This action of the control circuit in varying the oscillator frequency within a smaller and smaller range until the correct frequency is indicated is called *hunting*. Excessive hunting in the a-f-c circuit can cause several cycles of sine-wave bend to be produced in the top half of the picture. In order to minimize hunting in the a-f-c circuit, the filter for the d-c control voltage generally includes an antihunt RC network. Referring back to the a-f-c circuit in Fig. 17-30, as an example,  $R_{428}$  and  $C_{416}$  provide a long-time constant filter for the d-c control voltage, but the relatively small condenser  $C_{415}$  in parallel with  $R_{428}$  forms an a-c voltage divider with  $C_{416}$  for rapid variations in the control voltage, allowing the control voltage across

$C_{416}$  to change fast enough to eliminate hunting. In the a-f-c circuit in Fig. 17-29,  $R_{228}$  and  $C_{186}$  form the antihunt network.

*Hum in the A-F-C Circuit.* Since this is part of the sync circuits, hum voltage in the a-f-c circuit causes bend in the picture, but not in the raster, without hum bars. This is illustrated in Fig. 17-35, which shows severe horizontal pulling in the picture caused by heater-cathode leakage in the horizontal control tube.

**17-13. Localizing Hold Troubles.** When the picture does not hold, the trouble is either no sync or the oscillator is so far off the correct frequency that the sync cannot lock in the oscillator. In order to distinguish between incorrect oscillator frequency and no sync, the frequency of the

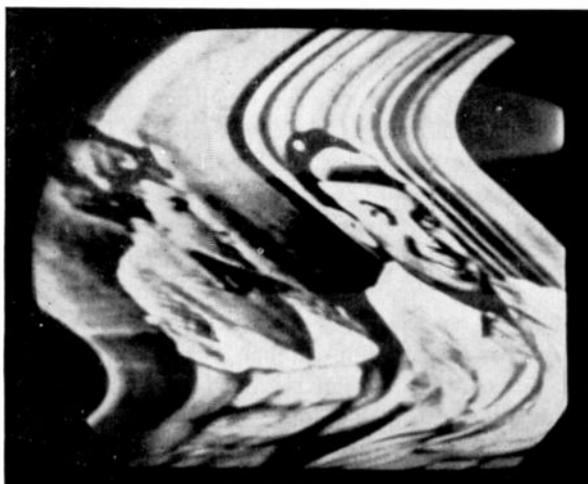


FIG. 17-35. Horizontal pulling in picture caused by heater-cathode leakage in horizontal control tube. (RCA.)

deflection oscillator may be varied manually to see if the picture can be framed momentarily.

When the horizontal hold is normal but the picture rolls vertically, as shown in Fig. 16-1, the vertical hold control is adjusted to vary the frequency of the vertical deflection oscillator. As the vertical oscillator frequency is brought closer to 60 cps the picture rolls more slowly. If the vertical hold control stops one complete frame momentarily, this shows the oscillator can operate at 60 cps. The fact that the rolling can be stopped but the picture does not lock in vertically indicates that the trouble is no vertical sync input voltage to the vertical deflection oscillator. If the vertical hold control cannot stop the picture from rolling, even for an instant, this indicates the oscillator frequency is incorrect since it cannot be adjusted to 60 cps. With the oscillator frequency lower than 60 cps, the picture rolls upward; higher than 60 cps the picture rolls

downward. The farther the vertical scanning frequency is from 60 cps, the faster the picture rolls, until multiple sections of the image appear folded over or under each other in the picture, as shown in Fig. 16-1b. At 30 cps or 20 cps, which are submultiples of 60 cps, two or three duplicate pictures can be seen one above the other. An important factor in the frequency of a blocking oscillator or multivibrator is the  $R_p C_c$  time constant in the grid circuit. Too short a time constant increases the frequency; too long a time constant decreases the frequency.

When the vertical hold is normal but the picture is in diagonal segments, as shown in Fig. 16-2, the horizontal oscillator frequency can be adjusted manually to see if the picture will hold horizontally for just an instant. In multivibrator circuits, the horizontal hold control usually varies the frequency of the horizontal oscillator. However, in circuits with a horizontal oscillator transformer, the oscillator frequency adjustment is generally in the transformer. In either case, the oscillator frequency can be varied manually. As the oscillator frequency is brought closer to 15,750 cps, the number of diagonal segments decreases. When the horizontal oscillator frequency can be adjusted to produce a complete picture for an instant, but it does not hold horizontally, this indicates the absence of d-c control voltage from the a-f-c circuit to lock in the horizontal oscillator. Then the picture slips horizontally, like the vertical rolling present without vertical sync. Multiple sections of the image may appear folded over or under each other as the picture slips to the left or right, before it tears into diagonal segments. If varying the oscillator frequency cannot change the diagonal segments into a complete picture momentarily, the trouble is incorrect oscillator frequency. When the frequency differs from 15,750 cps by 60 cps there is one diagonal black bar, which is produced by horizontal blanking. Every 60-cps difference between the oscillator frequency and 15,750 cps results in another diagonal bar. As the bars increase in number, they become thinner with less slope. The diagonal bars slope down to the left when the horizontal oscillator frequency is below 15,750 cps, or up to the left when the frequency is higher than 15,750 cps.

#### REVIEW QUESTIONS

1. What is the function of the vertical deflection oscillator? The horizontal deflection oscillator?
2. Why is the saw-tooth wave shape required for linear deflection?
3. What is the required wave shape of current in the scanning coils for magnetic deflection?
4. Give two factors that affect the linearity of the saw-tooth voltage rise in the output from a saw-tooth voltage generator.
5. Draw the grid-voltage waveform of a blocking oscillator and explain briefly how it is produced.
6. Draw the schematic diagram of a blocking oscillator and discharge tube saw-

tooth voltage generator for vertical deflection. Label the vertical hold control and the height control.

7. Why is the free-running frequency of a blocking oscillator lower than the synchronized frequency? Why must the synchronizing voltage for triggering the oscillator have positive polarity?

8. Draw the schematic diagram of cathode-coupled multivibrator saw-tooth voltage generator for horizontal deflection. Show where the control voltage from the a-f-c circuit is connected, and label the horizontal hold control.

9. Explain briefly how a cathode-coupled multivibrator operating as the vertical deflection oscillator is triggered by integrated vertical sync voltage of negative polarity.

10. Why does the frequency increase when a positive d-c voltage is added to the grid of a blocking oscillator?

11. In a cathode-coupled multivibrator, which of the two tubes always has the saw-tooth condenser in its plate circuit?

12. In a single-stage blocking oscillator and discharge tube operating as the vertical deflection oscillator, why does the picture usually roll when the height control in the plate circuit is adjusted?

13. What is the function of the peaking resistor in a trapezoidal-voltage generator circuit?

14. What is the main advantage of a horizontal a-f-c circuit, compared with triggered sync for the horizontal deflection oscillator?

15. In the horizontal a-f-c circuit shown in Fig. 17-29, what stage compares the sync and scanning frequencies? In what part of the circuit is the control voltage filtered? Where is the filtered control voltage applied?

16. What control varies the frequency of the horizontal deflection oscillator in Fig. 17-30?

17. If the vertical deflection oscillator locks in at 30 cps what will be the effect on the picture?

18. Where is the electron scanning beam on the screen of the picture tube at the time corresponding to the start of flyback? What should the kinescope grid voltage normally be at this time?

19. What is meant by "hunting" in the horizontal a-f-c circuit?

20. Referring to the horizontal a-f-c circuit in Fig. 17-29, give the function of each of the following components:  $C_{184}$ ,  $R_{224}$ ,  $R_{231}$ ,  $C_{187}$ , and  $C_{188}$ .

21. If varying the hold control in Fig. 17-29 changes the number of diagonal bars in the picture, does this indicate the control tube is operating? Why?

22. Referring to the horizontal a-f-c circuit in Fig. 17-30, give the function of each of the following components:  $C_{416}$ ,  $R_{432}$ ,  $R_{433}$ ,  $R_{436}$ ,  $C_{419}$ ,  $R_{429}$ ,  $R_{430}$ , and  $C_{417}$ .

23. If varying the hold control in Fig. 17-30 changes the number of diagonal bars in the picture, does this indicate the sync discriminator stage is operating? Why?

24. Referring to the horizontal a-f-c circuit in Fig. 17-31, give the function of each of the following components:  $C_{170}$ ,  $C_{174}$ ,  $R_{199}$ ,  $C_{175}$ ,  $R_{200}$ ,  $C_{177}$ ,  $C_{179}$ ,  $R_{204}$ , and  $R_{198}$  with  $R_{231}$ .

25. The picture is in diagonal segments, in a receiver having the horizontal a-f-c circuit in Fig. 17-31. Varying the oscillator frequency control in  $T_{108}$  produces a complete picture but it slips horizontally. In which stage or stages can the trouble be?

26. The cathode resistor is open in a cathode-coupled multivibrator operating as the vertical deflection oscillator. What will be seen on the kinescope screen?

27. The plate winding of the blocking oscillator transformer in a horizontal deflection oscillator is open. What will be seen on the kinescope screen, in a receiver using a flying high-voltage supply?

28. Referring to the horizontal a-f-c circuit in Fig. 17-29, describe briefly the effect on the picture for each of the following component troubles:
- The control-tube plate bypass condenser  $C_{185}$  is shorted.
  - The blocking oscillator plate resistor  $R_{231}$  is open.
  - The control-tube cathode bypass condenser  $C_{184}$  is shorted.
29. Referring to the horizontal a-f-c circuit in Fig. 17-30, describe briefly the effect on the picture for each of the following component troubles:
- The 6AL5 sync discriminator heater is open.
  - The a-f-c filter condenser  $C_{416}$  is shorted.
  - The oscillator plate decoupling resistor  $R_{435}$  is open.
30. Give four factors that affect the frequency of a blocking oscillator or multi-vibrator.

## CHAPTER 18

### DEFLECTION CIRCUITS

The horizontal and vertical deflection circuits produce the scanning raster on the kinescope screen, as illustrated in Fig. 18-1. For horizontal scanning, the horizontal oscillator generates 15,750-cps deflection voltage that is amplified by the horizontal output stage to provide 15,750-cps saw-tooth current in the horizontal deflection coils. These circuits produce the horizontal scanning lines. Similarly, the vertical oscillator generates 60-cps deflection voltage that is amplified by the vertical output stage to supply 60-cps saw-tooth current for the vertical deflection

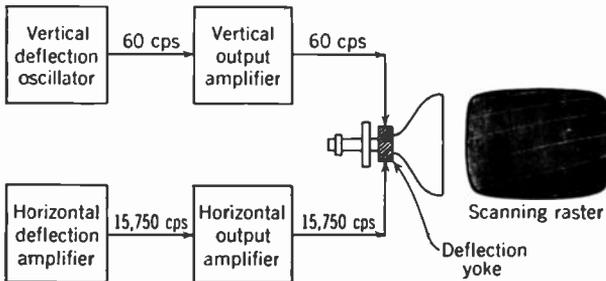


FIG. 18-1. The horizontal and vertical deflection circuits produce the scanning raster on the kinescope screen.

coils. As a result, the vertical scanning fills the screen from top to bottom with the horizontal scanning lines, to form the raster. The width of the raster depends upon the peak-to-peak amplitude of saw-tooth current in the horizontal deflection coils, while the height of the raster depends upon the saw-tooth current amplitude in the vertical deflection coils. Since the electromagnetic field associated with the current in the yoke deflects the electron beam in magnetic deflection, the current in the deflection coils must have the saw-tooth wave shape for linear scanning. Practically all television receivers use magnetic deflection because large-screen picture tubes, with a wide deflection angle and high value of anode voltage, would require too much deflection voltage for electrostatic

scanning, and the focus at the edges of the raster is better with magnetic deflection.

**18-1. Deflection Amplifiers.** The deflection amplifier for either horizontal or vertical scanning is usually a single power output stage, which has the function of amplifying the output voltage from the deflection oscillator in order to provide the amount of saw-tooth current required in the scanning coils for a full-sized raster. As illustrated by the vertical deflection amplifier circuit in Fig. 18-2, the main components of the deflection output circuit are the power output tube, the output transformer, and the deflection coils in the yoke. The output transformer is used in order to match the plate load requirements of the output tube to the lower impedance of the deflection coils.

The operation of the output stage as a power amplifier to supply saw-tooth current output for the deflection coils, with deflection voltage input from the deflection oscillator, can be summarized as follows:

1. The deflection voltage from the oscillator produces saw-tooth plate current in the amplifier.
2. The saw-tooth plate current of the deflection amplifier is the primary current of the output transformer.
3. The output transformer couples the current variations in the primary to the secondary, resulting in saw-tooth secondary current.
4. Since the deflection coils are connected across the secondary of the output transformer, they also have saw-tooth current.

The vertical output tube is usually a triode, or a beam-power tube connected as a triode with the screen grid tied to the plate. Typical vertical output tubes are the 6S4 or 6J5 triodes, one section of the twin triodes 6SN7-GT or 12BH7, and the 6K6-GT or 6V6-GT beam-power tubes triode-connected. With a triode vertical output tube, trapezoidal grid voltage from the deflection oscillator is necessary for saw-tooth plate current. The amplitude of the trapezoidal grid voltage for the vertical output tube is about 100 volts, peak to peak, including the negative spike.

The horizontal output tube is usually a beam-power pentode, like the 6BG6-G, 19BG6-G, or 6BQ6-GT which can supply more power output than the vertical amplifier. Two tubes can be connected in parallel for the horizontal output stage if necessary to supply the required amount of power. The grid voltage for the horizontal output tube can have the saw-tooth wave shape, with a peak-to-peak amplitude of about 50 to 90 volts.

High-voltage peaks are produced in the inductive output circuit of the deflection amplifier when the saw-tooth current drops sharply for the flyback. The voltage peak at the plate of the vertical output tube is 500 to 800 volts, while the horizontal output tube's plate has peaks of about 6,000 volts. Higher voltages are generated in the horizontal deflection

circuits because the faster flyback produces a greater rate of change in current.

**18-2. The Vertical Output Stage.** Figure 18-2 shows a typical circuit for the vertical output stage. The trapezoidal voltage output from the vertical deflection oscillator is coupled to the grid of the vertical output tube by the coupling condenser and the grid resistor. As a result, the input voltage drives the grid in the positive direction, producing a linear

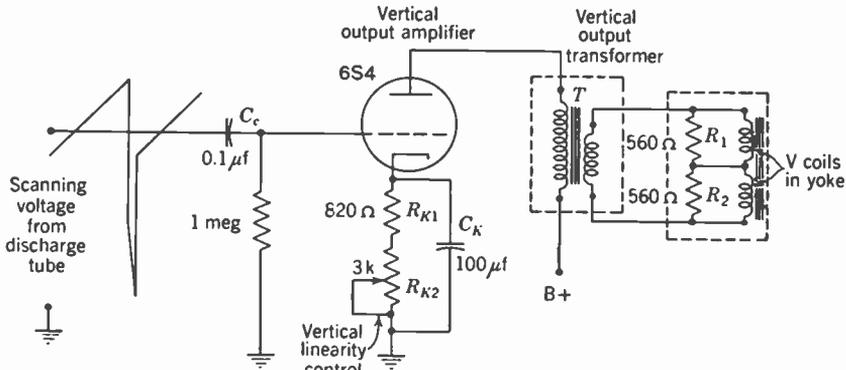


FIG. 18-2. Vertical deflection amplifier circuit.

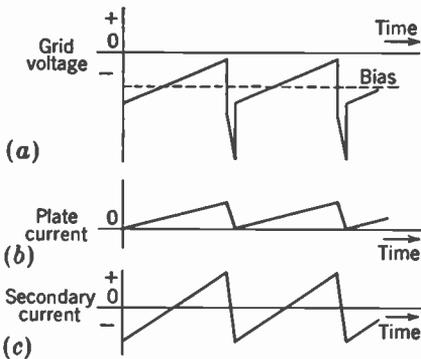


FIG. 18-3. Voltage and current wave shapes in vertical deflection amplifier circuit. (a) Trapezoidal voltage at grid of amplifier. (b) Saw-tooth plate current in primary of output transformer. (c) Saw-tooth secondary current in deflection coils.

rise of plate current in the output. Then the spike on the trapezoidal input voltage drives the grid voltage sharply negative, below cutoff, to make the plate current drop to zero for the flyback. Therefore, the output tube produces saw-tooth plate current, as long as the deflection oscillator supplies grid-driving voltage. The saw-tooth plate current through the primary of the output transformer induces voltage across the secondary that produces saw-tooth current through the scanning coils. The wave shapes in Fig. 18-3 illustrate how the saw-tooth output current corresponds to the trapezoidal input voltage at the grid of the vertical deflection amplifier.

$R_1$  and  $R_2$  in Fig. 18-2 are shunt damping resistors across the vertical deflection coils. Cathode bias for the vertical output stage, of 30 to 50 volts, is produced by  $R_{k_1}$  and  $R_{k_2}$  with the bypass condenser  $C_k$ . The cathode resistance  $R_{k_2}$  can be varied to adjust the bias and control the linearity of the saw-tooth rise in plate current.  $R_{k_1}$  provides a minimum bias for the output tube. The vertical output tube is usually a triode, but a pentode can be used with negative feedback to reduce the tubes's plate resistance and improve the linearity. Beam-power pentode tubes have the advantage of greater power sensitivity, requiring less grid-driving

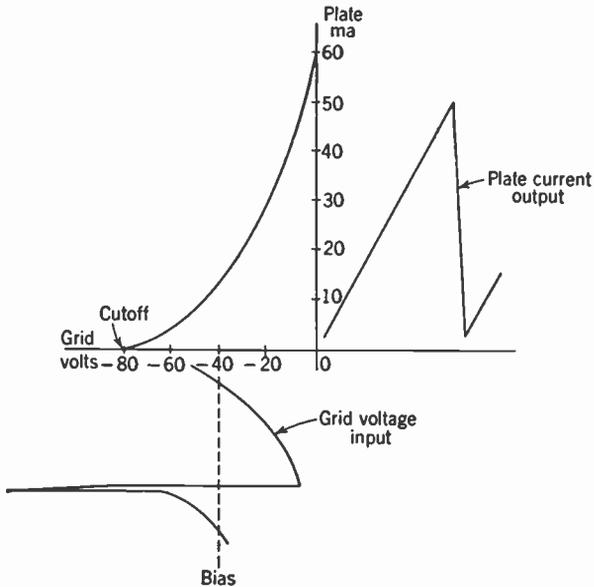


FIG. 18-4. Effect of vertical output tube's grid-plate transfer characteristic on linearity of the saw-tooth output current.

voltage for equal power output, but triodes can provide better linearity in the saw-tooth output current.

*Vertical Linearity Control.* The variable cathode resistance  $R_{k_2}$  in Fig. 18-3 adjusts the bias on the vertical deflection amplifier, in order to control the linearity of the saw-tooth output current. As illustrated in Fig. 18-4, the tube's grid-plate transfer characteristic curve is nonlinear, with curvature that is bowed inward. The saw-tooth wave tends to be bowed outward, resulting from either using too much of the  $RC$  charge curve in producing the saw-tooth voltage, or saturation in the output transformer for high values of current. Therefore, the operating characteristic of the deflection amplifier can compensate for nonlinearity in the saw-tooth wave. The tube tends to expand the amplitude values near the peak

of the wave that produce high values of plate current, corresponding to the bottom of the raster, and to compress the amplitude values at the start of the saw-tooth wave near the grid cutoff voltage, corresponding to the top of the raster. By changing the bias, the operating point is shifted to control the linearity of the saw-tooth current output. This variable bias control in the cathode circuit of the vertical deflection amplifier is called *vertical linearity*.

The vertical linearity control is mounted on the rear apron of the chassis as a servicing or installation adjustment. Linearity of the vertical scanning can be checked by observing crowding and spreading from top to bottom in the raster. For instance, the vertical scanning

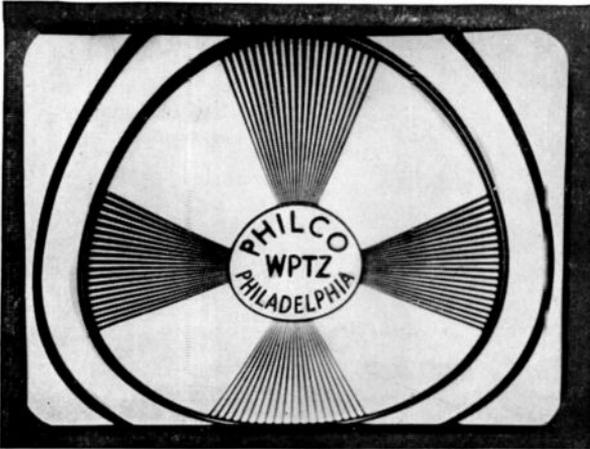


FIG. 18-5. Incorrect adjustment of vertical linearity. (Philco Corporation.)

current may rise to three-fourths its peak amplitude during the time one-half the field is scanned. Then the picture information that should fill one-half the screen vertically is stretched out to fill three-quarters of the raster. The remaining half of the picture information is crowded in one quarter of the raster at the bottom. As a result, the picture is stretched at the top and crowded at the bottom, as shown in Fig. 18-5. Notice that the top wedge is much longer than the bottom wedge of the test pattern. In a picture with people, they would appear to have short legs and long heads. The opposite type of vertical nonlinearity can also occur, with crowding at the top and stretching at the bottom. Correct adjustment of the vertical linearity is indicated by equal top and bottom wedges in the test pattern. The vertical linearity can also be set without a picture, by adjusting for uniform spacing of the scanning lines in the raster.

Since the gain of the vertical deflection amplifier varies when the bias is changed, adjustment of the vertical linearity control also changes the height of the raster. Similarly, adjustment of the height control changes

the vertical linearity. The vertical linearity control is more effective in adjusting the top of the raster. Adjusting the linearity control for less bias increases the height of the raster and stretches the top, as the amplitude values at the start of the saw-tooth wave are emphasized because of greater amplification farther from the grid cutoff voltage. Increasing the height with the height control stretches the bottom of the picture slightly, as the increased peak deflection voltage drives the grid voltage closer to

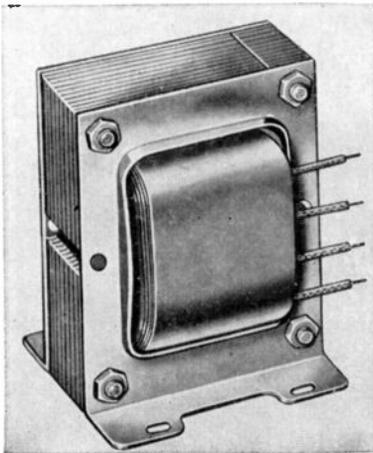


FIG. 18-6. Vertical output transformer. Base is  $2\frac{1}{2}$  in. square. (Standard Transformer Corporation.)

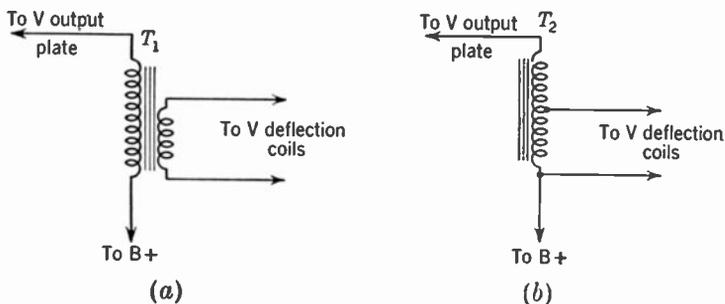


FIG. 18-7. Vertical output transformer connections. (a) Transformer with isolated secondary winding. (b) Autotransformer.

zero, resulting in greater emphasis of the amplitude variations near the peak of the saw-tooth wave just before flyback. Both the vertical linearity and height controls must be adjusted to obtain the required height with good linearity in the scanning raster.

*The Vertical Output Transformer.* A vertical output transformer is shown in Fig. 18-6. To illustrate the wiring connections, Fig. 18-7 shows the primary and secondary circuits for two types of transformers. In *a* the transformer has separate windings for inductive coupling of the

a-c deflection current from the primary to the secondary, while blocking the d-c plate voltage from the isolated secondary winding. The circuit in *b* is an autotransformer, which uses part of the primary winding as the secondary. The a-c deflection current is coupled inductively but the d-c plate voltage in the primary is also in the secondary circuit. In either case, the function of the vertical output transformer is to provide the primary impedance required as the plate load for the vertical deflection amplifier, while supplying the amount of secondary current necessary for the vertical deflection coils in the yoke. The electrical characteristics of the primary and secondary windings are indicated by the values listed in Table 18-1 for two different vertical output transformers.

TABLE 18-1. VERTICAL OUTPUT TRANSFORMER CHARACTERISTICS

Primary winding			Secondary winding	Turns ratio, primary/secondary	Remarks
D-C resistance, ohms	Impedance, ohms	Current, ma	D-C resistance, ohms		
590	19,000	15	6.9	10:1	Separate primary and secondary; for yoke with 50-mh vertical coils
1,600	27,000	12	4.4	18:1	Autotransformer; for yoke with 41-mh vertical coils

**18-3. Damping.** The inductance of the output transformer and scanning coils, with their distributed and stray capacitance, provide a tuned output circuit that can oscillate at its natural resonant frequency. Oscillations occur because the flyback on each saw-tooth wave of current produces a rapid change in current that generates a high value of induced voltage across the inductance, forcing the output circuit to oscillate at its natural resonant frequency by shock excitation. This effect of shock-excited oscillations is called *ringing*. The oscillations, at about 70 to 100 kc in the horizontal output circuit, would continue past the flyback time and produce ripples on the scanning-current waveform, as shown in Fig. 18-8. Since the scanning current produces deflection, the electron beam also oscillates back and forth in accordance with the oscillatory ripples on the saw-tooth deflection waveform. As a result, the oscillations can produce one or more white bars at the left side of the raster, as shown in Fig. 18-9. The bars are at the left because the oscillations are present immediately after flyback time. The bars are white because the electron

beam scans these areas several times during every horizontal line traced, as the oscillations make the horizontal scanning current repeat equal amplitudes at different times. Ringing occurs in the horizontal output circuit, primarily, rather than in vertical scanning, because of the fast horizontal flyback.

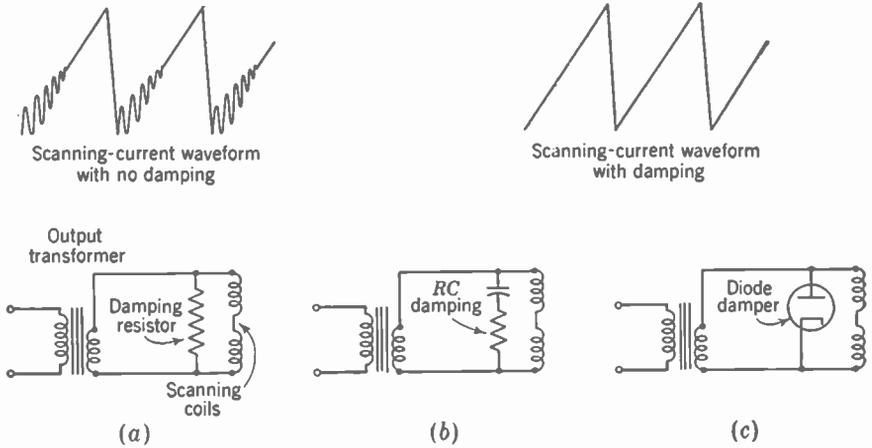


FIG. 18-8. Sine-wave oscillations in saw-tooth deflection current waveform, and methods of damping. (a) Resistor damping. (b) RC damping. (c) Diode damper tube.

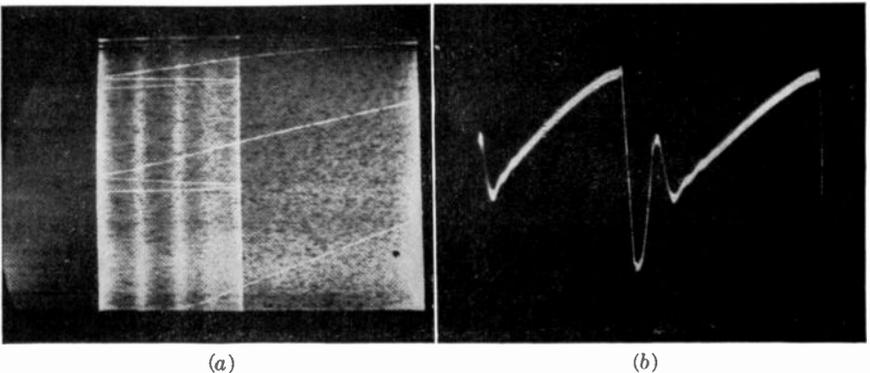


FIG. 18-9. Effect of insufficient damping in horizontal output circuit. (a) Wide white vertical bar at left side of raster caused by oscillation. (b) Oscillogram of current in horizontal deflection coils with oscillations. Oscilloscope internal sweep at 15,750/2 cps. (RCA Institutes, Inc.)

*Shock-excited Oscillations.* The action of shock-exciting a tuned circuit into oscillations is illustrated in Fig. 18-10. In *a* the switch *S* has been closed and current flows through *L* and *R*, which represent an inductive-resistive output circuit for magnetic scanning, while the distributed and stray capacitance *C* charges to the applied voltage. When the switch is opened in *b*, the voltage *E* is disconnected from the tuned circuit, remov-

ing the source of applied voltage and its internal resistance. Then the change in applied voltage can make the tuned circuit start oscillating at its natural resonant frequency. The tuned circuit oscillates without the battery because the energy stored in the electromagnetic field of the inductance and the electrostatic field of the capacitance, while the battery was connected, provide the required current and voltage. The current in the tuned circuit and the voltage across it oscillate with decreasing amplitudes until the stored energy is dissipated in the resistance of the circuit. The switch  $S$  corresponds to the action of the deflection amplifier, as it conducts during trace time to supply current to the output circuit but then is cut off at flyback time by the grid driving voltage.

The current in the coil decays toward zero when the switch is opened, but the sharp change as the current starts to decline produces a large self-induced voltage across  $L$ . Now the coil is a generator producing voltage

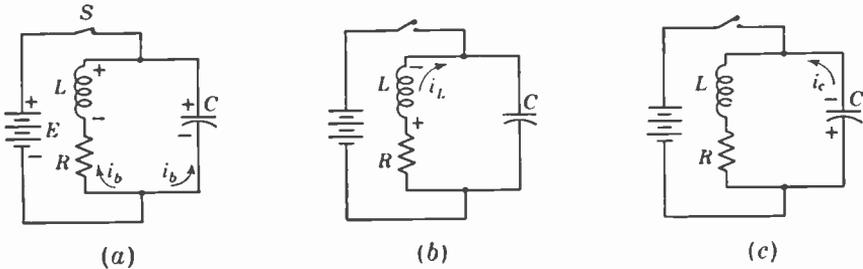


FIG. 18-10. Shock-exciting a tuned circuit into oscillation. (a) Switch is closed, and the battery supplies current and voltage for the  $LC$  circuit. (b) Switch is opened, and the self-induced voltage across  $L$  acts as the generator. (c) Reversed polarities one-half cycle later, with the voltage across  $C$  as the generator.

that keeps the current flowing in the same direction through the inductance as when the switch was closed. This is illustrated in  $b$  of Fig. 18-10. At the same time, the condenser discharges through  $L$  and  $R$ , reducing the voltage across  $C$ . The discharge current of the condenser is in the same direction as the current in the coil. A little later the condenser is completely discharged and the voltage across  $C$  is zero. The current  $i_L$  produced by the coil still is in the same direction, however, charging  $C$  to produce a voltage of opposite polarity from the original battery voltage. As the current in the coil decays to zero, the voltage across  $C$  charges to its maximum value. Then the condenser voltage becomes the generator as it discharges to produce the current  $i_c$  through  $L$  and  $R$ , as shown in  $c$ . This current through the coil is in the opposite direction from the original current from the battery. When the condenser voltage is down to zero, the coil again acts as the generator to maintain the current. The condenser then charges with its original polarity. As a result, the inductance and capacitance interchange energy, making the tuned circuit oscillate.

The waveform of the oscillations is shown in Fig. 18-11. Notice that the current and voltage are  $90^\circ$  out of phase with each other, as the current is maximum when the voltage is zero. During the first half cycle of oscillations, the current reverses from maximum in one direction to maximum in the opposite direction, while the voltage reaches its maximum negative value and returns to zero.

The  $LC$  circuit oscillates at its resonant frequency, with reduced amplitude for successive cycles until the stored energy is dissipated in the resistance. The frequency is approximately  $1/2\pi\sqrt{LC}$ , when  $R$  is small. How long the oscillations continue depends upon the  $Q$  of the circuit. The higher the series resistance  $R$ , the lower the  $Q$  and the sooner the oscillations decay to zero. A low value of resistance in parallel with the circuit has the same effect as a high series resistance in reducing the  $Q$  and damping the oscillations.

*Damping Methods.* The undesired oscillations can be eliminated from the linear rise on the saw-tooth wave by connecting a damping resistance

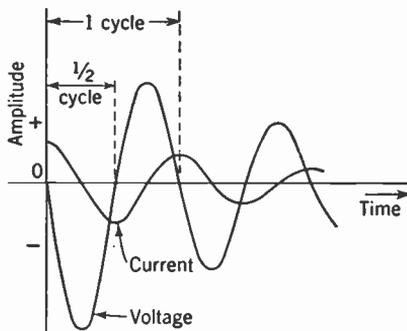


FIG. 18-11. Current and voltage waveforms in the oscillating  $LC$  circuit.

in parallel with the scanning coils as shown in Fig. 18-8a. Because of its low value, which is about 1,000 ohms, the parallel damping resistor lowers the  $Q$  of the output circuit and reduces the amount of self-induced voltage generated on the flyback. Therefore, the amplitude of oscillation is decreased and the oscillations decay to zero before the linear trace begins. Usually, two damping resistors are used in a balanced arrangement with one across each section of the scanning coils as shown in Fig. 18-2. With just a simple resistance for damping, however, some of the saw-tooth current is shunted through the parallel resistor. By the insertion of a condenser of suitable size in series with the damping resistor, as in Fig. 18-8b, the saw-tooth current can be prevented from flowing through the damping resistor so that all the scanning current is usefully employed in the deflection coils, while the high-frequency oscillations are damped out.

Either of these two methods is usually employed for damping in the



2. The damper tube operates in conjunction with the deflection amplifier to allow a fast flyback and then damp the oscillations in the output circuit after the retrace.  $V_2$  in Fig. 18-12 is the diode damper. For greater efficiency in horizontal scanning, the damped current is used to produce part of the trace at the left side of the raster while the output tube

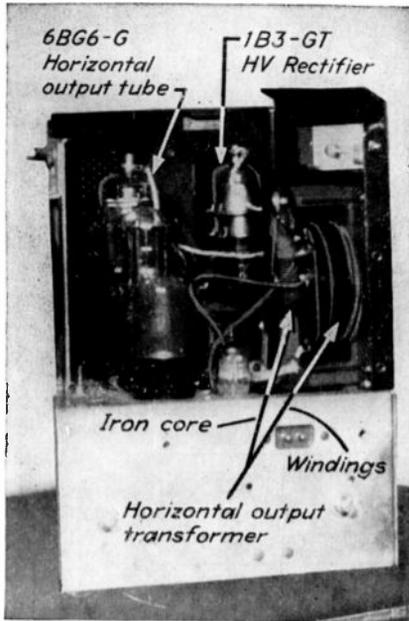


FIG. 18-13. Components of horizontal output circuit in high-voltage compartment on receiver chassis. (RCA Institutes, Inc.)

is cut off. Therefore, both the deflection amplifier and damper stages produce horizontal scanning.

3. The current through the diode damper when it conducts is allowed to charge a condenser in series with the B supply voltage and the damper, in order to produce a voltage across the condenser higher than B+.

$C_3$  in Fig. 18-12 is the B+ boost condenser. This increased supply voltage is generally called *boosted B+*, and  $C_3$  is the *B+ boost condenser*. With a boost of 100 volts added to the B+ of 300 volts, the voltage across the B+ boost condenser is 400 volts. This boosted B+ is the plate-supply voltage for the horizontal output tube in Fig. 18-12.

4. The high-voltage pulse produced by the first half cycle of oscillations, during the flyback, is stepped up, rectified, and filtered

to supply the anode voltage of 9 to 18 kv for the kinescope.  $V_3$  in Fig. 18-12 is the high-voltage rectifier.

*Operation of the Horizontal Deflection Amplifier.* The coupling condenser  $C_c$  and grid resistor  $R_g$  in Fig. 18-12 couple the saw-tooth voltage from the deflection generator to the grid of the output tube. The amplitude of the grid-driving voltage is about 75 volts, peak to peak. Grid-leak bias is developed because of grid current when the positive peak of the saw-tooth input voltage drives the grid of the output tube positive. The amount of grid-driving voltage is controlled by varying the capacitance of  $C_1$ . The cathode bias produced by  $R_k C_k$  is a safety bias to limit the plate current and protect the tube should there be no grid-voltage drive from the deflection generator and therefore no grid-leak bias. The total grid bias is about 40 volts, including the cathode bias and grid-leak bias.  $R_s$  drops the B+ voltage to the value required for the tube's screen

voltage and  $C$  is the screen bypass condenser. The plate current of the output tube  $V_1$  can flow through the transformer primary  $L_1$ ,  $L_6$ , from cathode to plate in the damper tube  $V_2$ , through  $L_2$  and  $L_5$ , the B supply, and back to the  $V_1$  cathode through  $R_k$ . The damper is conducting when plate current flows in the output tube, making a complete circuit.  $C_5$  prevents the d-c component of the output tube's plate current from flowing through the scanning coils, which would shift the horizontal centering, but couples the a-c saw-tooth wave. Since  $C_5$  has little reactance at 15,750 cps, the scanning coils are effectively connected across the secondary  $L_2$  for the horizontal scanning current.

As the linear rise on the saw-tooth input voltage drives the grid in the positive direction the amount of plate current increases, providing a saw-tooth rise of current in the primary winding  $L_1$  of the output transformer. The primary current in  $L_1$  induces voltage in  $L_2$  to produce a saw-tooth rise of current through the scanning coils in the secondary circuit. At the peak of the saw-tooth input, the grid voltage drops sharply to drive the instantaneous grid voltage more negative than the grid cutoff voltage. As a result,  $V_1$  stops supplying current to the output circuit. The output tube remains cut off until the grid-driving voltage completes its swing in the negative direction for the flyback and the linear rise in the positive direction makes the instantaneous grid voltage less negative than the cutoff voltage. Then the saw-tooth rise in grid voltage produces a saw-tooth rise of current again in the output circuit to produce the next cycle of operation.

Figure 18-14 shows the grid-plate transfer characteristic curve of the horizontal output tube, with saw-tooth grid-driving voltage and plate-current output pulses corresponding to part of the saw-tooth voltage input. The stage operates as a class C amplifier. The bias is more negative than cutoff, but the input voltage drives the tube into conduction. While the grid voltage due to the bias and driving voltage is more negative than cutoff, no plate current flows. For part of the cycle, however, the saw-tooth input voltage drives the grid voltage less negative than cutoff to produce a linear rise of plate current. As an example, when the a-c input voltage drives the grid voltage 20 volts more positive than the bias of  $-40$  volts, the instantaneous grid voltage is  $-20$  volts, allowing plate current to flow. The plate-current output, therefore, is a linear rise in plate current corresponding to about two-thirds of the cycle of saw-tooth voltage input. For approximately the first one-third of the cycle of saw-tooth input voltage, the tube remains cut off and does not supply any output. However, this part of the scanning cycle is provided by the damped current in the output circuit.

**18-5. Horizontal Scanning and Damping.** When the horizontal output tube conducts to supply current for the horizontal scanning coils, the

linear rise in current deflects the electron beam to the right side of the raster. During this time, energy is stored in the inductance and capacitance of the output circuit. Then, when the deflection amplifier tube is cut off by its grid-driving voltage, the output circuit starts oscillating. The first half cycle of oscillations is allowed to continue undamped for a fast flyback from the right side of the raster to the left side. After the retrace, though, the damper conducts to make the oscillations decay to zero. The damped current in the horizontal scanning coils deflects the electron beam from the left side of the raster toward the center. Before

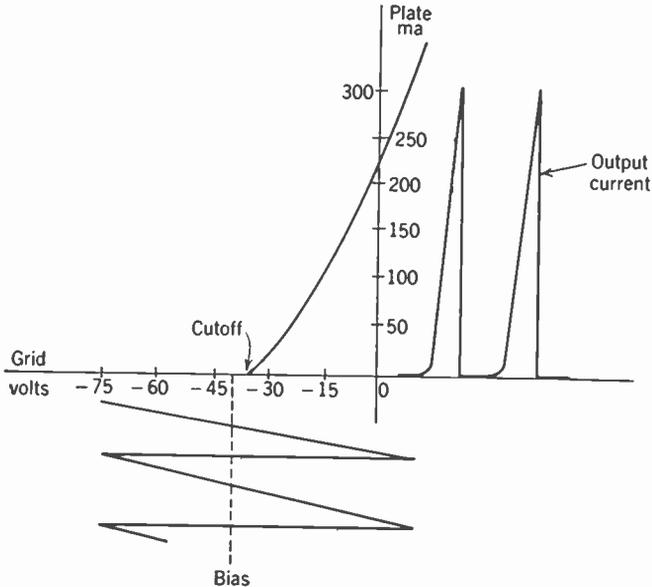


FIG. 18-14. Operation of horizontal output tube as class C amplifier. Saw-tooth plate current flows for only part of the input voltage cycle.

the damped current decays to zero, the deflection amplifier starts conducting again to finish the trace to the right edge of the raster. The use of the damped current for producing part of the horizontal trace at the left side of the raster is often called *reaction scanning*.

The operation of the diode damper in the horizontal output circuit is illustrated in Fig. 18-15, for the time when the deflection amplifier is cut off. The nonconducting output tube is now a high resistance which cannot damp the oscillating circuit, and the first half cycle of oscillations makes the damper plate negative, as shown in *a*, so that the diode cannot conduct either. As a result, the current in the deflection coils drops sharply to zero and reverses to its maximum value in the opposite direction very quickly for a fast flyback. This wave shape of current during retrace time is shown in Fig. 18-16. With a frequency of 70 kc for the

oscillations in the horizontal output circuit, the period of a half cycle is  $1/0.14 \mu\text{sec}$ , resulting in a flyback time of approximately  $7 \mu\text{sec}$ .

Since the voltage across the oscillating circuit is  $90^\circ$  out of phase with the current, as shown in Fig. 18-16, the polarity of the damper plate voltage reverses to become positive just as the current starts to decrease from its maximum negative value. This is the time for the start of the trace. Now the damper tube conducts to produce a linear decay of current, as the positive voltage across the deflection coils charges the B+ boost condenser through the diode damper. Conduction in the diode is

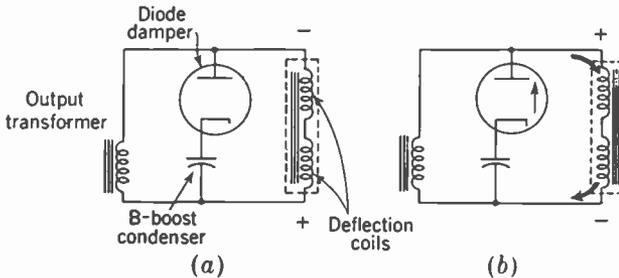


FIG. 18-15. A-c equivalent circuit of the damper tube in the horizontal output circuit. (a) Damper plate negative and nonconducting for the retrace. (b) Damper plate positive and conducting for the trace.

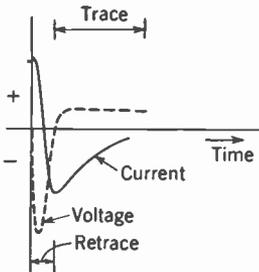


FIG. 18-16. Wave shapes of voltage across the deflection coils and current through the coils while the output tube is cut off.

illustrated in Fig. 18-15*b*. The linear decay of current in the deflection coils when the damper conducts during trace time is shown in Fig. 18-16. As the damped current declines to zero, the electron beam is deflected from the left side of the raster toward the center. With zero current in the scanning coils, the electron beam would be undeflected at the center. Before the damped current declines to zero, however, the output tube starts conducting to produce the linear rise of current in the scanning coils that completes the trace to the right side of the raster.

Figure 18-17 shows the time relations between conduction in the output tube, the flyback when both the output tube and damper are nonconducting, and conduction in the damper while the output tube is cut off. Notice that while the output tube's grid voltage is more negative than

cutoff, there is no plate current, but the damped oscillations in the output circuit provide the reaction scanning current for horizontal deflection. Just before the damped current decays to zero, the grid voltage makes the deflection amplifier conduct and the plate current supplies output for horizontal scanning. The same action occurs for every horizontal scanning line.

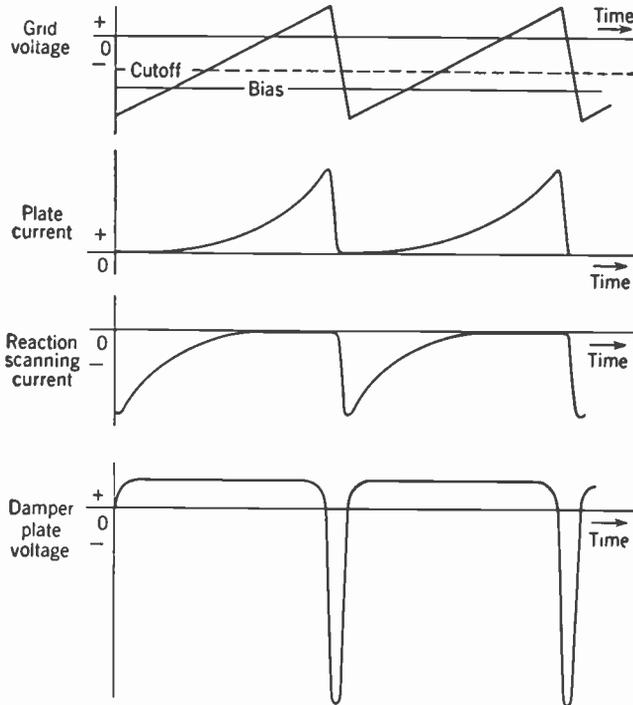


FIG. 18-17. Comparison of waveforms in the horizontal output stage, for two scanning lines.

Figure 18-18 illustrates how the current supplied by the deflection amplifier and the damped current in the output circuit combine to produce saw-tooth current in the horizontal deflection coils for the full horizontal trace from left to right across the scanning raster. Note that both components of the scanning current for the trace deflect the electron beam toward the right, since the decrease in the damped current of negative polarity varies in the same direction as the positive rise of current produced by the output tube.

In summary, then, the horizontal scanning action occurs in three steps:

1. The output tube supplies current for the horizontal scanning coils to deflect the electron beam about two-thirds the distance across the raster to the right edge.

2. The undamped oscillation in the horizontal output circuit, while the output tube and damper are not conducting, produces the rapid reversal of scanning current required for the fast flyback from right to left.

3. Immediately after the flyback, the damper conducts while the output tube is cut off, to provide the reaction scanning current for the trace at the left side, about one-third the distance across the raster.

The damper tube in the horizontal output circuit must withstand the high value of peak inverse voltage, have good cathode-to-heater insulation, and be able to conduct the peak values of scanning current. The maximum inverse voltage peak at the damper plate, which occurs in the middle of the flyback when the first half cycle of oscillations makes the damper plate negative, is 1 to 2 kv. In addition, sufficient cathode-to-heater insulation is required because the damper cathode can be positive with respect to chassis ground by the amount of the boosted B+ voltage. If the heater is connected to ground through the filament transformer, arcing may occur between cathode and heater in the tube. When the cathode and heater are tied together to eliminate arcing in the tube, a separate ungrounded filament winding is necessary for the heater of the damper. In this case, the damper filament winding may arc to ground.

Finally, the peak value of scanning current through the damper tube may be as high as 500 ma. The peak current amplitude occurs when the damper starts to conduct immediately after the first half cycle of oscillations. The electron beam is then at the left side of the raster beginning the trace toward the center as the current declines to zero. Typical tubes used for the horizontal damper stage are the 5V4-G, 6W4-GT, 25W4-GT, 6AX4-GT, and 12AX4-GT diode rectifiers and the 6AS7-G twin-triode power tube.

**18-6. Boosted B+ Voltage.** The current through the damper tube when it conducts is used to charge the B+ boost condenser, providing voltage output higher than the B supply voltage. The two components that add to produce the boosted B+ voltage are the B+ from the low-voltage power supply and the rectified deflection voltage in the horizontal

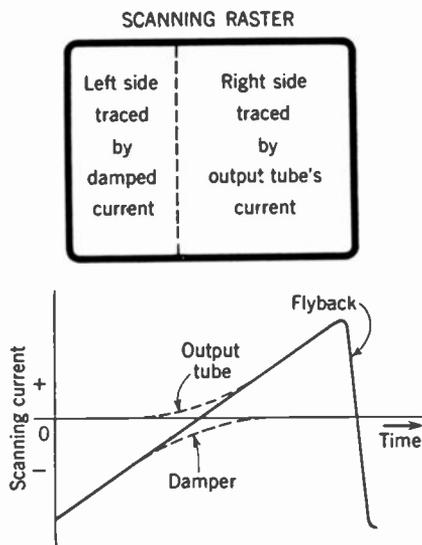


FIG. 18-18. Illustrating how the damped current and output current combine to produce horizontal scanning in the raster.

output circuit. The amount of boost added to  $B+$  by the rectified deflection voltage is generally 100 to 250 volts.

The damper tube is a half-wave rectifier. As illustrated in Fig. 18-19a, the  $B+$  voltage is applied to one side of the secondary winding  $L_2$  on the horizontal output transformer, while the other side of  $L_2$  connects to the plate of the diode damper. Therefore, the damper conducts to charge the  $B+$  boost condenser  $C_3$  in the cathode circuit of the damper to the  $B+$  voltage. The path of charging current is through  $L_2$ , and from cathode to plate in the damper, charging  $C_3$  to 300 volts, with the cathode side of  $C_3$  positive. In addition, the a-c deflection voltage across  $L_2$ , which produces saw-tooth current in the scanning coils, makes the diode plate 100 volts more positive during horizontal trace time. As a result,

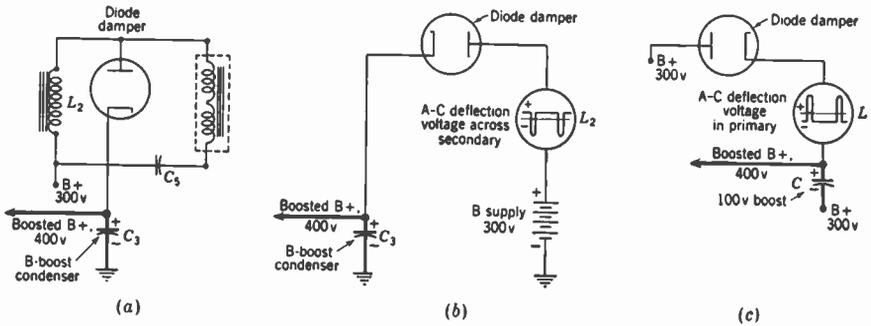


FIG. 18-19. Producing the boosted  $B+$  voltage. (a) Diode damper in secondary circuit of horizontal output transformer. Component part numbers are the same as in Fig. 18-12. (b) Equivalent circuit of *a*, with diode damper in secondary circuit. (c) Diode damper in primary circuit, where a-c deflection voltage has opposite polarity, and  $B+$  boost condenser returns to  $B+$  instead of ground.

$C_3$  charges to a voltage 100 volts higher than  $B+$ , or 400 volts. The path of charging current is the same as when the  $B$  supply charges  $C_3$ . In effect, the a-c deflection voltage is in series with the  $B$  supply voltage and the damper tube, to produce the boosted  $B+$  voltage across the  $B+$  boost condenser.

The equivalent circuits in Fig. 18-19b and c illustrate the two main types of  $B+$  boost circuits, with the diode damper either in the primary or secondary of the horizontal output transformer. In the secondary, as in *a* and *b*, the a-c deflection voltage during horizontal trace time has positive polarity and drives the diode plate positive. In the primary, the a-c deflection voltage during trace time is negative, as shown in *c*, and drives the inverted diode damper's cathode negative. In either case, the a-c deflection voltage during trace time allows conduction in the diode damper to charge the  $B+$  boost condenser. In addition, the  $B+$  voltage is applied to the plate of the damper. The  $B+$  boost condenser is in the

cathode-to-ground circuit of the damper, to provide rectified voltage that is positive with respect to ground.

The boosted B+ voltage is generally used for the horizontal output tube and additional stages, including the horizontal oscillator, vertical oscillator, and vertical amplifier, because the higher plate-supply voltage allows more power output with improved linearity. More stages drawing load current from the B+ boost condenser, however, reduce the amount of voltage boost. It is important to note that the B+ boost circuit must be operating in order to produce normal plate voltage for all stages that use the boosted B+ voltage. The fact that the horizontal deflection amplifier's plate current can return through the damper tube to the B

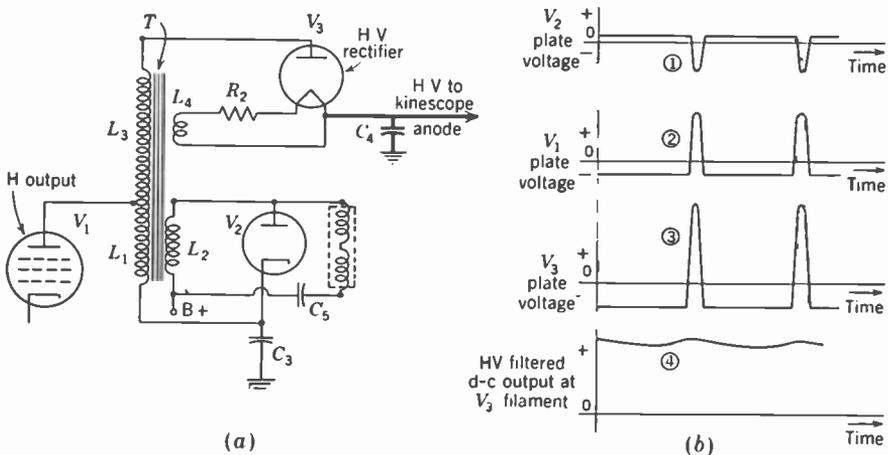


FIG. 18-20. Producing the flyback high voltage. (a) Typical circuit. Component part numbers the same as in Fig. 18-12. (b) Voltage wave shapes, with numbers corresponding to the steps described in text.

supply enables the B+ boost circuit to operate at the start, before boosted B+ voltage is available across the B+ boost condenser.

**18-7. Flyback High Voltage.** The voltage pulse produced in the horizontal output circuit during the first half cycle of oscillations for the flyback is generally used to provide the d-c high voltage required for the kinescope anode. As illustrated in Fig. 18-20, this is accomplished by stepping up the a-c deflection voltage and applying it to the plate of the high-voltage rectifier tube, with positive polarity at the plate of the rectifier for the pulse produced during retrace time. The rectifier then supplies d-c output voltage in the cathode circuit, equal to the peak value of the input pulses, which is filtered to provide the kinescope anode voltage.

Referring to the circuit in Fig. 18-20a and the voltage wave shapes in b, the operation of the flyback high-voltage supply can be analyzed as follows:

1. The first half cycle of oscillations in the output circuit produces a voltage peak of 1 to 2 kv, which is the flyback pulse. This negative voltage pulse in the secondary, shown by wave shape 1, keeps the damper from conducting.

2. The flyback pulse has positive polarity at the plate of the horizontal amplifier in the primary of the output transformer, as shown by wave shape 2. The positive voltage pulse of 3 to 6 kv in the primary is higher than the secondary voltage by the turns ratio of the transformer. Since both the damper and amplifier tubes are cut off for the retrace, the first half cycle of undamped oscillations produces the high-voltage flyback pulse. Maximum amplitude of the pulse occurs in the middle of the flyback.

3. The high-voltage primary winding  $L_3$  steps up the flyback pulse for the plate of the high-voltage rectifier. As shown by wave shape 3, this stepped-up positive flyback pulse has an amplitude of 9 to 18 kv.

4. With the stepped-up a-c deflection voltage from the primary high-voltage winding at the plate of the high-voltage rectifier, it produces positive d-c output voltage at the cathode, which is filtered to provide the steady d-c high-voltage output shown by wave shape 4.

When the high-voltage rectifier conducts, the positive flyback pulse charges the high-voltage filter condenser  $C_4$  in Fig. 18-20 to produce the d-c high-voltage output. The rectifier current can flow from cathode to plate in  $V_3$ , through  $L_3$  and  $L_1$  in the primary of the output transformer, discharging slightly the B+ boost condenser  $C_3$  and charging the filter condenser  $C_4$  in the cathode circuit of the high-voltage rectifier to approximately the peak amplitude of the flyback pulse. Since the ripple frequency is 15,750 cps for the half-wave rectifier, 500  $\mu\text{mf}$  is enough capacitance for  $C_4$  to provide the required filtering of the d-c high-voltage output. It is important to note that, with a flyback high-voltage supply, the d-c output for kinescope anode voltage cannot be produced without the a-c deflection voltage input to the plate of the high-voltage rectifier. Therefore, the horizontal deflection circuits must be operating to have brightness on the kinescope screen.

**18-8. Horizontal Deflection Controls.** The horizontal drive, linearity, and width controls are usually in the horizontal deflection amplifier stage. All three controls, which are setup adjustments on the rear apron of the chassis, must be adjusted to provide the required width with good horizontal linearity in the picture.

*Horizontal Drive Control.* Referring to the horizontal deflection amplifier grid circuit in Fig. 18-21a, the variable condenser  $C_1$  is the drive control to adjust the amount of saw-tooth voltage applied to the grid of the output stage from the deflection oscillator. Since the grid-driving voltage controls the plate current, the horizontal drive adjustment varies the

output of the deflection amplifier. The reason why the drive control varies the amplifier's grid voltage is illustrated in Fig. 18-21*b*. The drive control  $C_1$  forms a capacitive voltage divider with the coupling condenser  $C_c$  for the a-c grid voltage. In a capacitive voltage divider, the applied a-c voltage is divided according to the reactance of the condensers, just as the voltage is proportional to the resistance in a resistive divider.

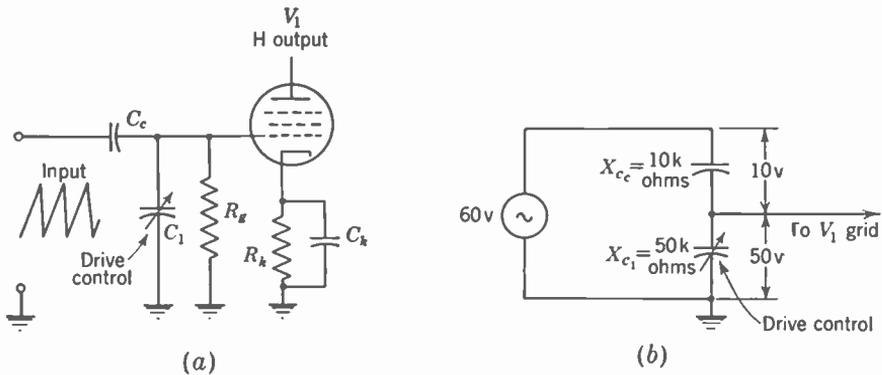


FIG. 18-21. The horizontal drive control varies the amount of saw-tooth voltage input to the grid of the deflection amplifier. (a) Circuit diagram, with component part numbers the same as in Fig. 18-12. (b) Equivalent a-c voltage divider.

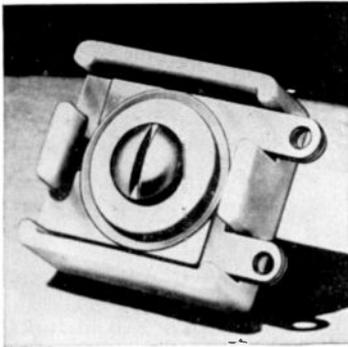


FIG. 18-22. Drive-control condenser. Capacitance range 10-200  $\mu\text{f}$ . (Arco Electronics, Inc.)

Capacitive reactance is inversely proportional to capacitance. Therefore, the smaller capacitance in the divider has more reactance and develops more a-c voltage than a larger capacitance. The maximum capacitance of the drive control is less than the capacitance of the coupling condenser, so that most of the input voltage is developed across the drive control. In Fig. 18-21*b*, the reactance of the drive control is shown for a setting in the middle of its capacitance range. With 50,000 ohms of capacitive reactance for  $C_1$ , out of the total of 60,000 ohms for  $C_1$  and  $C_c$ , the voltage across  $C_1$  is five-sixths the total applied voltage of 60 volts, which equals 50 volts. The voltage produced across the drive

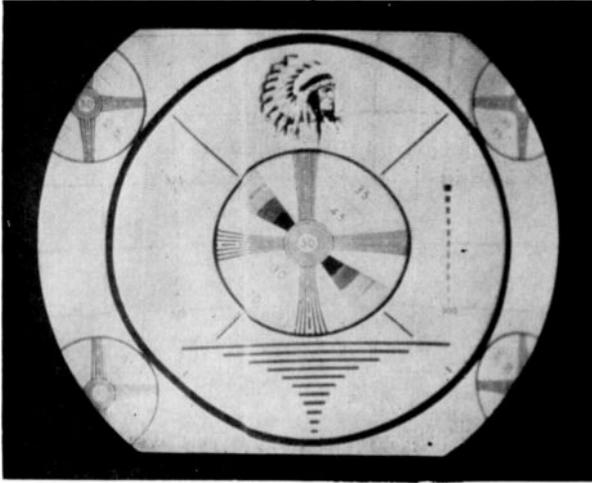


FIG. 18-23. White vertical bar near the center of raster, caused by excessive horizontal drive. (Admiral Corporation.)



FIG. 18-24. Width-control coil. Slug inside the coil varies the inductance from 3 to 16 millihenrys, approximately. (RCA.)



FIG. 18-25. Horizontal-linearity-control coil. Slug inside the coil varies the inductance from 1 to 8 millihenrys, approximately. (RCA.)

control is the grid-driving voltage for the output stage, since it is applied between the grid and cathode. The drive control is often a mica compression-type trimmer condenser, as shown in Fig. 18-22, mounted on the rear apron of the chassis.

Since the horizontal drive increases the brightness and width of the raster on the kinescope screen, the control is adjusted for maximum

drive, up to the point where the horizontal scanning is distorted. Excessive horizontal drive usually causes either a white vertical line or wrinkle near the center of the raster. Figure 18-23 shows a raster with the white line caused by too much horizontal drive. To produce maximum brightness and width without distortion, therefore, the horizontal drive control is generally adjusted by increasing the drive until the white line or wrinkle appears and then backing off the control enough to eliminate the distortion. The increased width obtained with more horizontal drive usually tends to stretch the left side of the raster.

*Width Control.* Referring to Fig. 18-12, the width-control coil  $L_5$  in parallel with part of the secondary winding  $L_2$  of the output transformer, can change the effective secondary inductance and, therefore, vary the amount of output to the horizontal deflection coils to adjust the width of the raster. A typical width-control coil is shown in Fig. 18-24. The coil is usually mounted on the back wall of the high-voltage compartment, near the horizontal output transformer but not inductively coupled to it. Increasing the inductance makes the raster wider, usually with a tendency to stretch the right side slightly.

*Horizontal Linearity.* The amount of voltage across the B+ boost condenser varies as the condenser charges and then discharges. This rise and fall of voltage can be considered an a-c ripple in the plate-supply voltage provided by the booster circuit for the horizontal deflection amplifier. Slight variations in the output tube's operating characteristic can be obtained, therefore, by shifting the phase of the ripple in plate voltage, with respect to the saw-tooth grid voltage, providing small changes of horizontal scanning linearity. This is the function of the horizontal linearity coil  $L_6$  in Fig. 18-12, which forms a filter for the a-c ripple, with the bypass condenser  $C_2$ . Figure 18-25 shows a typical horizontal-linearity-control coil. Adjusting the inductance of the horizontal linearity coil usually varies the linearity at the center of the picture.

Three examples of poor horizontal scanning linearity are illustrated in Fig. 18-26, with the corresponding nonlinear saw-tooth current in the horizontal deflection coils. To relate the current wave shape to the scanning, remember that the damped current produces the left side of the trace; the beam is at the center with zero current, when the damped current approaches zero and the output tube starts to conduct; and the output tube's current finishes the trace at the right. In *a*, the picture information is stretched at the left side and crowded at the right, as indicated by the unequal side wedges of the test pattern. The outside circle is also distorted, stretching to the left. With an actual picture, people at the left in the scene would appear too broad, while at the right they would be too thin. This type of nonlinearity can be caused by setting the horizontal drive control too high and the width control too low. The reverse

happens for the nonlinear scanning in *b*, where the left side is crowded and the right side is stretched. Either type of nonlinear horizontal scanning can cause the effect of a person at the center of the scene appearing to have one shoulder broader than the other. In *c*, the side wedges are equal but the center circle is distorted because of incorrect adjustment of the

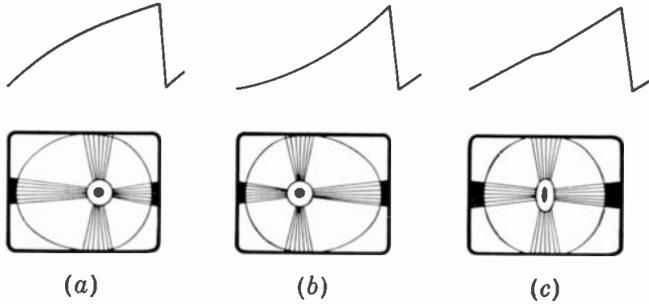


FIG. 18-26. Effects of nonlinear horizontal scanning, with corresponding distorted saw-tooth current wave shape. (a) Stretching at the left side and crowding at the right. (b) Crowding at the left side and stretching at the right. (c) Crowding at the center.

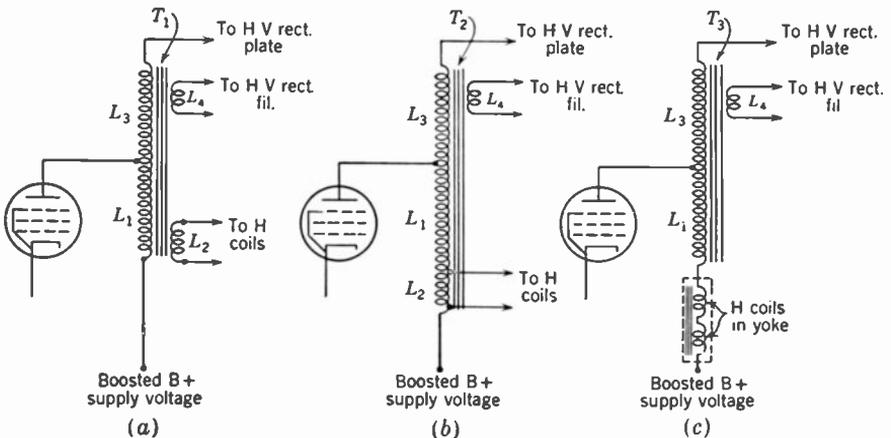


FIG. 18-27. Types of horizontal output transformer circuits. (a) Separate secondary winding. (b) Autotransformer. (c) Direct-drive, with horizontal deflection coils in primary plate circuit.

horizontal linearity control. It should be noted, though, that the width and height must have the ratio of 4:3, since incorrect aspect ratio can also result in distorted circles.

**18-9. Horizontal Output Transformers.** Figure 18-27 illustrates the three main types of horizontal output transformers, according to the way the horizontal deflection coils are connected into the output circuit. In

*a*, the transformer has a separate secondary winding for the horizontal deflection coils. The autotransformer in *b* does not isolate the secondary winding from the primary. In *c*, the deflection coils are not isolated either, but they are connected in series with the primary winding in the plate circuit of the horizontal output tube, in a circuit arrangement generally called *direct drive*. In the direct-drive horizontal output circuit the impedance of the horizontal deflection coils and the primary impedance of the transformer provide the plate load for the output tube. The horizontal deflection coils for a direct-drive horizontal output circuit have an inductance of about 30 mh, compared with approximately 8 mh for low-impedance deflection coils transformer-coupled to the primary. Therefore, less current through the horizontal scanning coils is required in the direct-drive output circuit.

The horizontal output transformer produces the desired amount of flyback high voltage and matches the horizontal output tube to the horizontal scanning coils, in order to produce the sweep width required for the deflection angle of the kinescope and the yoke. Typical values of inductance are approximately 100 mh for the high-voltage winding and about 30 mh for the primary inductance in series with the plate of the output tube. The d-c resistance of the high-voltage winding is 200 to 400 ohms, while the primary winding is about 20 to 30 ohms. Figure 18-28 shows a *universal* type of horizontal output transformer, which has multitapped windings to provide different inductance values, for replacement or conversion use. Voltage pulses at the horizontal scanning rate can be obtained for keyed a-g-c systems, from either the horizontal output transformer or a tapped width-control coil, although these connections are not shown in Fig. 18-27. The horizontal output transformer generally has a core of powdered iron to minimize mechanical vibrations, which can cause a *singing* sound at the frequency of 15,750 cps.

**18-10. Deflection Yokes.** Since the current amplitude in the scanning coils determines how much the electron beam in the kinescope is deflected, the yoke is generally rated in terms of deflection angle. Furthermore, kinescopes with the same deflection angle and high voltage require equal deflection for a full-sized raster on the screen, regardless of the screen size. As an example, a 70° yoke can fill the screen of either a 17TP4 or 24AP4 picture tube because they both have the same deflection angle of 70°.

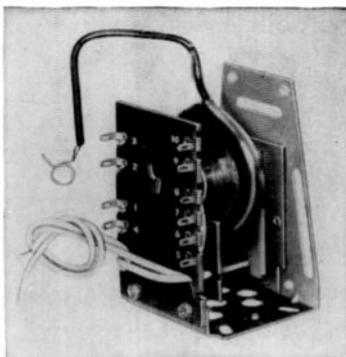


FIG. 18-28. Horizontal output transformer. (RCA.)

TABLE 18-2. DEFLECTION YOKES

Horizontal coils		Vertical coils		Kinescope deflection angle	Remarks
Inductance, mh	$R$ , ohms	Inductance, mh	$R$ , ohms		
8	14	50	65	50-57	For autotransformer or isolated secondary in horizontal output circuit. For 10- to 16-in. kinescopes
13	24	41	48	70	For autotransformer or isolated secondary in horizontal output circuit. For 16- to 24-in. kinescopes
30	55	3	5	70	For direct drive in horizontal output circuit. For 16- to 24-in. kinescopes
12	15	46	43	90	For 27- to 30-in. kinescopes

The electrical characteristics of several deflection yokes are listed in Table 18-2, to illustrate the different requirements of the vertical and horizontal coils, depending on the kinescope deflection angle and the horizontal output circuit. For any one type of deflection yoke, the

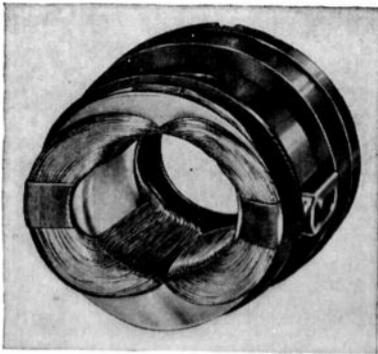


FIG. 18-29. Deflection yoke. Outside diameter of ferrite core approximately 3 in. (Standard Transformer Corporation.)

inductances of the vertical and horizontal coils are different from each other, in order to minimize cross-talk, or cross coupling. It should be noted that, although the inductance of the horizontal deflection coils is relatively low, their inductive reactance at the horizontal scanning frequency of 15,750 cps is high enough to make the horizontal scanning coils primarily an inductive load. With saw-tooth current, therefore, the voltage wave shape across the inductive horizontal deflection coils is rectangular with a sharp pulse for the flyback. Since the horizontal output tube then has an inductive plate load impedance, the plate voltage waveform is also rectangular with a sharp flyback pulse of high amplitude. The vertical coils have enough resistance compared with their inductive reactance at 60 cps to form a resistive-inductive load. Therefore, the voltage across the vertical deflection coils has the trapezoidal wave shape,

with saw-tooth current in the coils. The plate voltage of the vertical output tube has the same trapezoidal wave shape because of the inductive-resistive plate load impedance.

Figure 18-29 shows a typical deflection yoke, while the wiring connections for the terminals at the back of the yoke are illustrated in Fig. 18-30. Note that  $C$ ,  $R_1$ , and  $R_2$  are located physically across the terminals in the yoke. The small condenser  $C$ , which compensates for leakage reactance in the horizontal output transformer, is connected across one of the horizontal coils to prevent excessive ringing in the horizontal output transformer.  $R_1$  and  $R_2$  are damping resistors across the vertical coils but mainly they have the function of reducing the amplitude of horizontal oscillations coupled into the vertical coils by crosstalk from the horizontal coils, to prevent ripple at the top and bottom of the raster at the left side.

**18-11. Horizontal and Vertical Deflection Circuits.** The deflection circuits consist of the vertical oscillator that drives the vertical output circuit to produce current in the vertical deflection coils for vertical scanning, and the horizontal oscillator to drive the horizontal output circuit for horizontal scanning. Generally, the blocking oscillator and discharge tube or the multivibrator is used for the deflection oscillator stage.

Either circuit can be used for the vertical or horizontal oscillator. In the vertical deflection amplifier, its output circuit is transformer-coupled to the vertical scanning coils in the yoke, using a vertical output transformer that either is an autotransformer or has a separate secondary. Direct drive is not practicable for the vertical output circuit because too much inductance would be necessary for the vertical coils to provide the required plate load impedance at 60 cps. In many receivers, the vertical output is coupled to the kinescope grid-cathode circuit, in order to provide internal blanking during vertical retrace time. In the horizontal output circuit, direct drive, an autotransformer or a separate secondary can be used for the horizontal coils. The damper in the horizontal output circuit

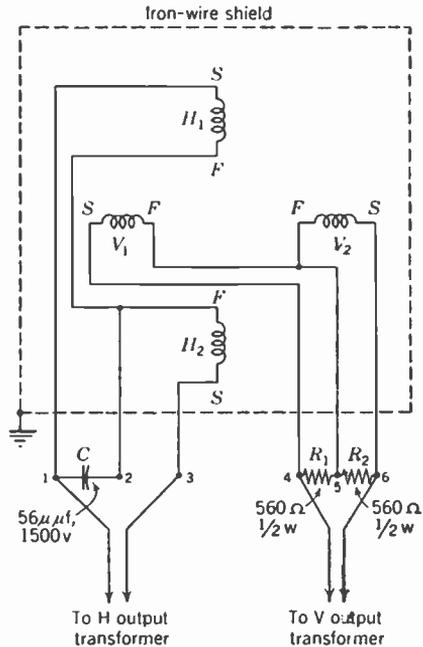


FIG. 18-30. Deflection-yoke wiring connections, with damping network connected across output terminals in yoke.  $S$  is start and  $F$  finish of winding.

provides reaction scanning for the left side of the horizontal trace and is used to produce the boosted  $B+$  voltage. With direct drive or an auto-transformer in the horizontal output circuit, the damper diode is inverted and the a-c deflection voltage is coupled to the cathode. The flyback high voltage is generally used to produce the kinescope anode voltage. In most receivers, horizontal output pulses are coupled to the a-g-c circuit in order to key the a-g-c amplifier into conduction for the horizontal sync pulses. It is important to remember that the deflection circuits produce the scanning raster with or without synchronizing voltage, but the sync is needed to lock in the deflection oscillators at the correct frequency to hold the picture vertically and horizontally. The integrated vertical sync triggers the vertical deflection oscillator to lock it in at 60 cps. In the horizontal deflection circuits, the d-c control voltage from the horizontal a-f-c circuit corrects the horizontal oscillator's frequency to make it operate at 15,750 cps.

*Typical Horizontal and Vertical Deflection Circuit.* The schematic diagram in Fig. 18-31 illustrates the main requirements of the horizontal and vertical scanning circuits to produce the raster, with typical voltage wave shapes. For vertical scanning, one triode section  $V_{401A}$  of the 6SN7-GT is a blocking oscillator and discharge tube circuit to generate the vertical deflection voltage. Vertical synchronizing voltage from the integrating circuit is coupled by  $C_{404}$  to the grid of the oscillator to synchronize the vertical scanning. The vertical hold control  $R_{405}$  varies the resistance in the grid circuit to control the free-running frequency of the oscillator. In the plate circuit, the height control  $R_{408}$  varies the amount of deflection voltage output from the vertical oscillator.  $C_{406}$  is the saw-tooth condenser, with the peaking resistor  $R_{407}$  to produce trapezoidal voltage. The 6S4 triode output tube supplies the saw-tooth output current, which is coupled to the vertical deflection coils in the yoke by the autotransformer  $T_{402}$ . The vertical linearity control  $R_{410}$  varies the output tube's cathode bias.

The horizontal deflection oscillator  $V_{405}$  uses both triode sections of the twin triode 6SN7-GT in a cathode-coupled multivibrator circuit. Since automatic frequency control is used for the horizontal oscillator, the d-c control voltage from the horizontal sync discriminator stage synchronizes the horizontal scanning. The variable inductance  $L_{401}$  in the stabilizing tuned circuit, which is the horizontal lock control, is adjusted to make the oscillator pull into horizontal synchronization. The oscillator's free-running frequency is varied by the hold control  $R_{434}$ , which changes the resistance in the grid circuit.  $C_{420}$  is the saw-tooth condenser in the plate circuit of the discharge tube, with the peaking resistor  $R_{438}$ . The deflection voltage produced by the horizontal oscillator stage is coupled to the grid of the 6BQ6-GT horizontal output tube by the coupling con-



varied to adjust the width of the raster. The capacitance of  $C_{433}$  also affects the width, since it is in parallel with  $L_{402}$ . Notice that the horizontal output circuit provides voltage for the horizontal sync discriminator stage, from terminal 8 on  $T_{404}$  with respect to ground, so that the frequency of the deflection voltage can be compared with the frequency of the horizontal synchronizing pulses to control the horizontal oscillator frequency. To produce the boosted B+ voltage in the horizontal output circuit, current through the 6W4-GT damper tube charges  $C_{428}$  in the cathode circuit of the damper to 400 volts. The boosted B+ voltage is filtered by  $C_{427}$  and  $L_{403}$ , with  $L_{403}$  variable to serve as the horizontal linearity control by varying the plate-supply voltage for  $V_{406}$ . In addition to the horizontal output tube, the horizontal and vertical oscillator stages

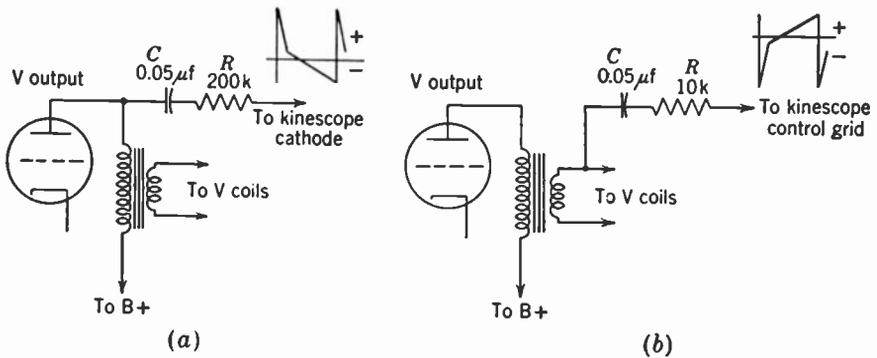


FIG. 18-32. Circuit arrangements for using vertical flyback pulses to provide internal vertical blanking. (a) From plate of vertical output tube. (b) From secondary of vertical output transformer.

use the boosted B+ as the plate-supply voltage. The boosted B+ voltage also provides the accelerating grid voltage for the kinescope. The kinescope anode voltage is obtained from the flyback high voltage produced across terminals 3 and 1 on  $T_{404}$ , which is applied to the 1B3-GT high-voltage rectifier. The high-voltage filter condenser  $C_{429}$  returns to the plate of the damper tube, instead of chassis ground, in order to increase the kinescope anode voltage by 300 volts, since the damper plate is connected to B+. Note the  $\frac{1}{4}$ -amp fuse in the B supply line to protect the components in the horizontal output circuit against excessive current. If the fuse opens, there will be no deflection and no high voltage.  $R_{439}$  and  $R_{441}$  are parasitic resistors to reduce the effects of spurious oscillations in the horizontal output circuit.

*Internal Blanking of Vertical Retrace.* In many receivers, the voltage pulse produced by the flyback in the vertical output circuit is coupled to the kinescope grid-cathode circuit, in order to provide additional blanking during vertical retrace time. This internal vertical blanking is in addi-

tion to the blanking voltage present at the kinescope grid as part of the composite video signal. The advantage of using the additional vertical blanking voltage is that the retrace lines produced during vertical flyback do not appear on the kinescope screen for any setting of the brightness control. Figure 18-32 illustrates two arrangements for the internal vertical blanking circuit. In *a*, the voltage pulse at the plate of the vertical output tube is coupled by the *RC* circuit to the kinescope cathode circuit. Since the flyback pulse at the plate of the output tube has positive polarity with respect to ground, it is coupled to the kinescope cathode, driving the control-grid voltage negative, to cut off the beam current for blanking. The coupling condenser *C* blocks the d-c voltage from the B supply, while the decoupling resistor *R* isolates the kinescope circuits

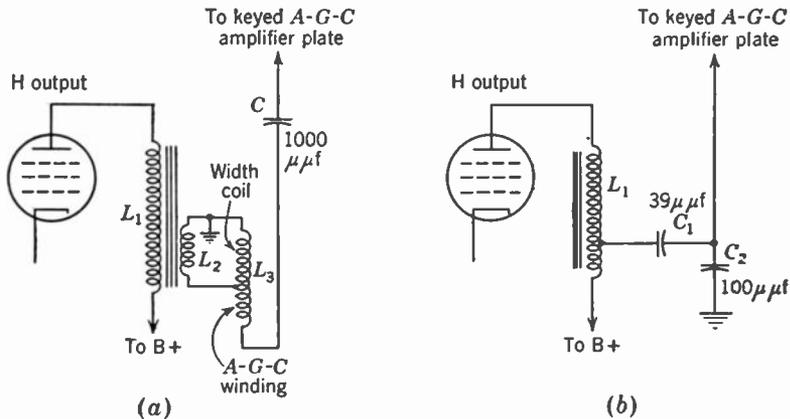


FIG. 18-33. Circuit arrangements for obtaining horizontal flyback pulses for keyed a-g-c amplifier from horizontal output circuit. (a) From a-g-c winding in secondary. (b) With capacitive voltage divider in primary.

from the vertical output stage. About 50 to 100 volts is required for blanking. In *b*, the vertical flyback pulse for blanking is taken from the secondary of the output transformer. The flyback voltage in the secondary can be obtained with negative or positive polarity for either the kinescope grid or cathode. Negative pulses for internal vertical blanking can also be taken from the peaking resistor in the vertical oscillator circuit.

**Horizontal Keying Pulses for Keyed A-G-C Amplifier.** In receivers that use a keyed or gated a-g-c circuit, the flyback pulses from the horizontal output circuits are coupled to the a-g-c amplifier to key it into conduction during the horizontal pulse time only. The polarity used for the keying pulses is positive with respect to ground, because they are applied to the plate of the a-g-c amplifier to make it conduct. The peak amplitude of the keying pulses is several hundred volts. Figure 18-33 illustrates two

arrangements for obtaining the horizontal flyback pulses for keying the a-g-c amplifier. In *a*, the separate a-g-c winding on the coil in the secondary of the horizontal output circuit provides a-c deflection voltage with flyback pulses. This is coupled by *C* to the plate of the a-g-c amplifier. In *b*, the a-c deflection voltage with flyback pulses is taken from the primary of the output transformer. Since the deflection voltage is much higher in the primary, the capacitive voltage divider consisting of *C*<sub>1</sub> and *C*<sub>2</sub> is used to reduce the pulse amplitude for the plate of the a-g-c amplifier.

**18-12. Testing Scanning Linearity with Bar Patterns.** Linearity of the vertical and horizontal scanning can be checked by producing bar patterns on the kinescope screen, with a signal generator. This method does not require any transmitted picture signal. Instead, the output from the signal generator is coupled into the receiver to supply signal voltage for the kinescope grid-cathode circuit. As the generator signal varies the kinescope grid voltage, while the deflection circuits are producing scanning, pairs of dark and light bars are formed on the raster, as shown in Fig. 18-34. Horizontal bars are produced when the frequency of the signal at the kinescope grid is less than 15,750 cps; above 15,750 cps the bars are vertical or diagonal. Since the synchronizing voltage for the deflection oscillators is usually taken from the video amplifier in the receiver, the generator signal also provides synchronization. The synchronization can hold when the synchronizing frequency is an exact multiple of the scanning frequency. Just vary the signal generator frequency to obtain the desired number of bars and adjust the receiver hold control to make the bars stay still.

Suppose that a 60-cps sine-wave signal is varying the kinescope control-grid voltage in synchronism with the vertical scanning motion at the field frequency of 60 cps. During the positive half cycle the signal makes the grid more positive, increasing the beam current and screen illumination; the negative half cycle reduces the beam current and screen illumination. Since it takes  $\frac{1}{120}$  sec for a half cycle of the 60-cps signal, the scanning beam moves approximately halfway down the screen during this time. Therefore, if the positive half cycle of the sine-wave signal coincides with the first half of the vertical scan, the top half of the picture will be brighter than the bottom half. The pattern on the screen then is a pair of horizontal bars, one bright and the other dark. When the signal generator output frequency is increased to multiples of 60 cps, additional pairs of narrower horizontal bars will be formed on the screen, as shown in Fig. 18-34*a*. The number of pairs of bars is equal to the signal generator frequency divided by the vertical scanning frequency, minus any bars that may be produced during vertical retrace time if the signal frequency is high enough to produce more than about 20 pairs of bars. As an example, a frequency of 240 cps results in four pairs of horizontal bars when the

vertical scanning frequency is 60 cps. With an audio-frequency output of 400 cps from the signal generator, six pairs of bars can be obtained by adjusting the vertical hold control to the vertical scanning frequency of  $400\frac{2}{6}$  cps.

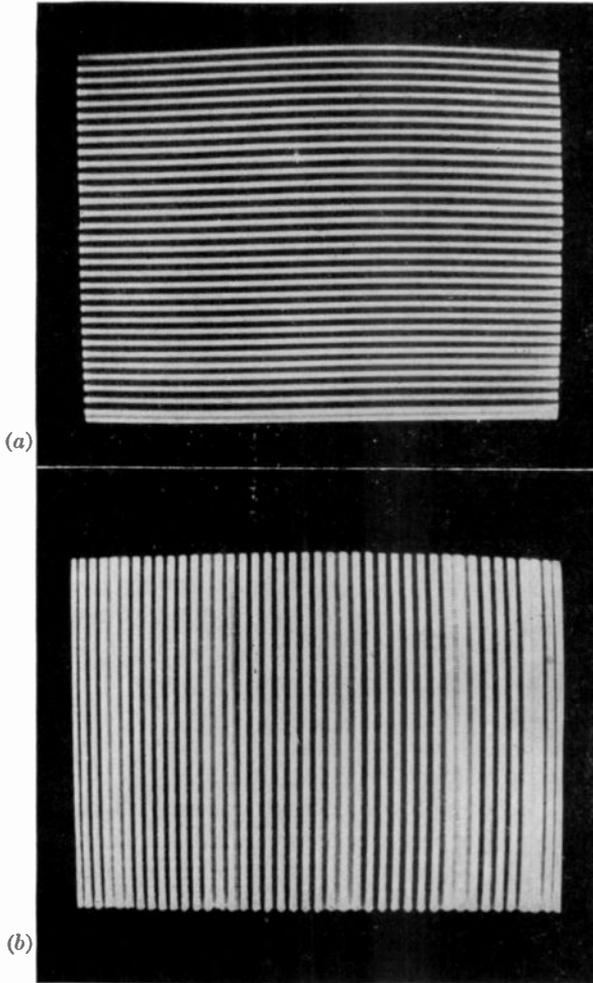


FIG. 18-34. Bar patterns on kinescope screen obtained with signal generator. (a) Horizontal bars to check vertical scanning linearity. (b) Vertical bars to check horizontal scanning linearity. (From Donald G. Fink, *Television Engineering*, McGraw-Hill Book Company, Inc.)

Vertical scanning linearity is indicated by the spacing between the parallel horizontal bars. If the vertical scanning motion is linear, the bars will be equally spaced. Otherwise, the bars will be spread out or crowded together. Adjustments can then be made with the vertical

linearity and height controls to obtain the most uniform distribution of the bars.

When the frequency of the modulating signal becomes equal to the horizontal line-scanning frequency, vertical bars are formed instead of the horizontal bars. Consider the case of a 15,750-cps sine-wave signal varying the kinescope grid voltage in phase with the horizontal scanning. During one horizontal line the screen is made brighter for approximately one-half the picture width, as the positive half cycle of the grid voltage increases the beam current; the negative half cycle makes the screen darker. The same effect occurs for succeeding horizontal lines and the result is a pair of vertical bars on the screen, one bar white and the other dark. If the frequency of the signal generator is increased to multiples of 15,750 cps, additional pairs of narrower vertical bars will be produced on the screen, as shown in Fig. 18-34*b*. Their spacing indicates linearity of the horizontal scanning motion. If necessary, the horizontal linearity, drive, and width controls can be adjusted to obtain uniform spacing of the bars and linear scanning.

The number of pairs of vertical bars is equal to the frequency of the applied grid signal divided by the horizontal scanning frequency. As an example, a frequency of 157.5 kc results in 10 pairs of vertical bars when the horizontal scanning frequency is 15,750 cps. However, all the bars may not be visible. Some are formed during the horizontal retrace time, when the signal frequency is high enough to produce more than about 10 pairs of bars. These bars formed during the flyback are wider because of the fast retrace and appear as variations of shading in the background, as can be seen in Fig. 18-36. It is possible to determine the retrace time by counting the bars visible during the trace time and comparing this with the total that should be produced.

Diagonal bars are produced when the frequency of the grid modulating voltage is higher than the horizontal line-scanning frequency but is not an exact multiple. In this case the light and dark parts of each line are regularly displaced in successive order at different positions with respect to the start of the trace, instead of lining up one under the other. The diagonal bars usually do not stay still because the signal frequency is not an exact multiple of the horizontal scanning frequency and does not synchronize the deflection oscillator.

**18-13. Hum Voltage in the Deflection Circuits.** Excessive 60- or 120-cps hum voltage in the horizontal deflection circuits produces S-shaped edges on the scanning raster, as illustrated in Fig. 18-35. When the edges of the raster are straight but the picture has the S-shaped curvature at the sides, this indicates hum in the horizontal sync, but not in the horizontal deflection circuits. In the vertical deflection amplifier excessive hum voltage causes nonlinear vertical scanning.

The S-shaped edges of the raster may be caused by either hum adding to the horizontal scanning current as in *a* of Fig. 18-35, or hum modulation of the horizontal scanning current as in *b*. The hum modulation in *b* results when the plate current of a tube in the horizontal deflection circuits is made to vary at the hum frequency, as might be caused by leakage between cathode and heater. In this case the hum frequency is 60 cps. Because of the modulation, the peak-to-peak amplitude of each saw-tooth cycle in the horizontal scanning current increases and decreases, varying the width of every horizontal line at the hum frequency. Therefore, the raster has S-shaped edges curving in opposite directions, as in *b* of Fig. 18-35, with bulge opposite bulge. One sine-wave cycle is produced along each vertical edge when the hum frequency is 60 cps, since the beam scans one field from top to bottom during  $\frac{1}{60}$  sec.

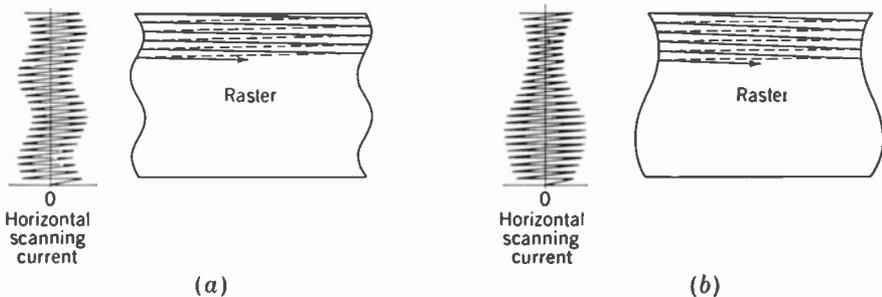


FIG. 18-35. Hum voltage in the horizontal deflection circuits. (a) Addition of 120 cps hum and horizontal scanning signal. (b) Modulation of horizontal scanning signal by 60 cps hum.

When the B supply voltage for a tube in the horizontal deflection circuits has excessive ripple, the hum voltage generally adds to the horizontal scanning voltage, as in *a* of Fig. 18-35. Every saw-tooth cycle of the saw-tooth scanning wave has the correct amplitude but each horizontal scanning line is displaced relative to the next, curving the vertical edges of the raster, with a bulge on one side opposite an indentation on the other side. The effect is comparable with varying the horizontal centering at the hum frequency. Two complete sine waves are shown on each edge in *b*, for 120-cps hum from a full-wave B supply. Since one cycle of the 120-cps hum takes  $\frac{1}{120}$  sec, which is one-half the time to scan a complete field, two complete sine-wave ripples are produced during the field period of  $\frac{1}{60}$  sec.

**18-14. Troubles in the Deflection Circuits.** Since they produce scanning, the deflection circuits cause troubles in the raster. In order to determine whether a trouble is in the raster, the picture can be removed to observe the raster alone, by either switching to an unused channel or shorting the antenna input. Raster troubles generally appear with or

without signal and are the same on all channels. Troubles in the height of the raster are caused by the vertical deflection circuits. The horizontal deflection circuits cause troubles in the width of the raster. In addition, troubles in the horizontal deflection circuits can cause no brightness, because of lack of kinescope anode voltage, when the receiver uses the flyback high-voltage power supply. It is also important to note that, with a circuit for internal blanking of the vertical retrace, a defect in the vertical output circuit can cause troubles in kinescope brightness.

*Horizontal Line Only.* Just a bright horizontal line at the center of the screen, as in Fig. 8-4b, shows there is no output from the vertical deflection circuits. Either the vertical oscillator is not producing deflection voltage or the trouble is in the vertical amplifier and its output circuit. The trouble can be localized by injecting 60-cycle a-c voltage into the grid circuit of the vertical amplifier. This is a sine-wave voltage but the amplified output should produce vertical deflection on the kinescope screen, with distorted linearity. Increased height with the injected voltage shows the vertical output stage is operating. Therefore, the trouble is in the vertical oscillator stage. No increase in height with the injected voltage means trouble in the vertical output circuit, including the amplifier tube, output transformer, and vertical deflection coils.

*No Horizontal Deflection.* Assuming the receiver has the flyback high-voltage power supply, no output from the horizontal deflection circuit results in no kinescope anode voltage and no brightness on the kinescope screen. The cause of no horizontal output is in the horizontal oscillator, the horizontal output stage, or the damper circuit. Remember that boosted B+ for plate-supply voltage cannot be produced if the damper is not conducting. In a receiver using an r-f high-voltage power supply, it should be noted that the absence of horizontal output results in a bright vertical line at the center of the screen.

*Size Troubles.* When the raster does not have enough height or width to fill the mask of the kinescope, the picture appears with black bars at the edges, corresponding to the unused screen areas not scanned and therefore not illuminated. This can be seen in Figs. 17-21 and 17-22. Also, the linearity will be distorted if the raster does not have the 4:3 aspect ratio with the incorrect height or width. Insufficient height in the raster, which cannot be corrected with the vertical linearity and height controls, indicates insufficient scanning current in the vertical deflection coils. Either there is not enough deflection voltage output from the oscillator or the gain of the output stage is low. A raster that cannot be made wide enough by adjusting the horizontal drive and width controls indicates insufficient scanning current in the horizontal scanning coils. The trouble can be in the horizontal oscillator, output stage, or damper circuit. When both the height and width of the raster are insufficient at the

same time, resulting in a small raster, this indicates low B+ voltage. It should be noted, though, that in many receivers insufficient B+ voltage results in just reduced width of the raster, while the normal height can be obtained by adjusting the height and linearity controls, because the horizontal deflection circuits require more power from the low-voltage supply. In receivers using the r-f high-voltage power supply, excessive kinescope anode voltage can also cause a small raster. A raster that is too large, with poor focus, indicates insufficient kinescope anode voltage.

**Bright Bars in the Raster.** Bright bars that are vertical, mainly at the left side of the raster, are caused by ripples in the horizontal scanning current. A bright bar at the top or bottom edge of the raster means trouble in the vertical deflection circuits. The different types of bright bars that can be produced in the raster by the deflection circuits are as follows:

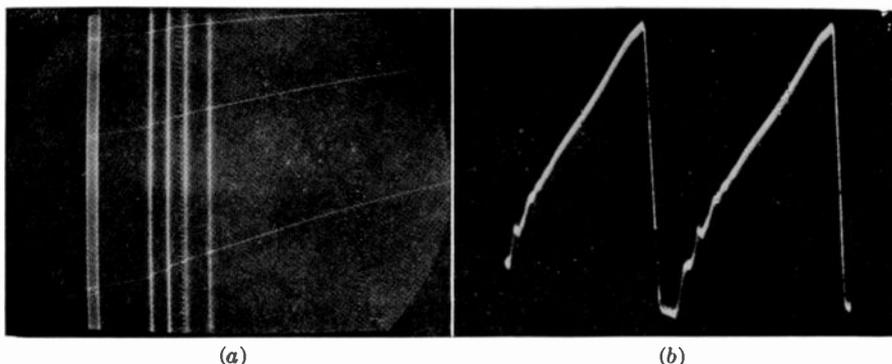


FIG. 18-36. Excessive horizontal drive. (a) Thin, bright bars at left side of raster caused by ripples in horizontal scanning current in (b). (RCA Institutes, Inc.)

1. Thin bright bars at the left, as shown in Fig. 18-36a, result from excessive ringing that cannot be completely damped, although the damping circuit may be normal. The width of the raster is not reduced in this case. Figure 18-36b shows the horizontal saw-tooth scanning current with the oscillatory ripples that produce the bright bars. Notice that the bars are bright where the electron beam scans the same area more than once for each scanning line. The excessive ringing that produces the thin bright bars can be caused by too much drive at the grid of the horizontal output stage, or troubles in the damping network across the deflection coils in the yoke. It should be noted, however, that some ringing at the left side of the raster is normal in most receivers, but the bright bars should not be obvious when the picture is on the raster.

2. A wide bright bar at the left, as illustrated in Fig. 18-37, indicates trouble in the horizontal damping circuit. Narrow bars are also present. The wide bar corresponds to the high-amplitude oscillation immediately after the flyback in the horizontal saw-tooth scanning current. Nor-

mally, the amplitude of this oscillation is low but a trouble in the damping circuit that results in low damper current causes insufficient damping. Notice that the low value of damping current results in a narrow raster with insufficient width at the left side. Too low a value of damping current can be caused by troubles such as an open B+ boost condenser or low emission in the damper tube. However, failure of the damper tube results in no horizontal output and no brightness because the deflection amplifier must return to B+ through the damper.

3. A bright bar at the top or bottom edge of the raster with crowding of the scanning lines usually is a result of nonlinear vertical scanning,

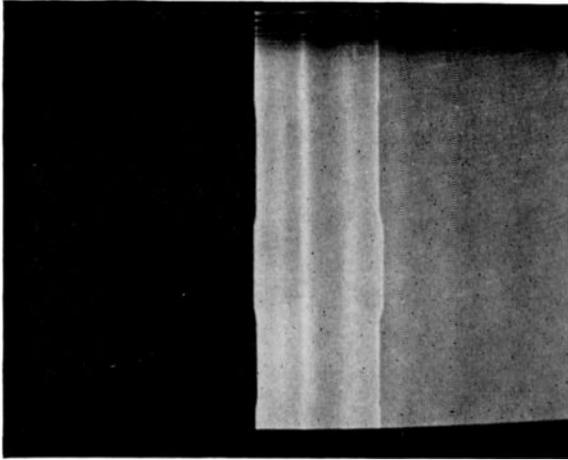


FIG. 18-37. Wide bright bar at left side of raster caused by open B+ boost condenser in cathode circuit of damper tube. (RCA.)

caused by amplitude distortion of the vertical saw-tooth deflection current. Figure 18-38 shows the effect of insufficient bias on the vertical output tube. The bright bar with crowded scanning lines at the bottom of the raster indicates insufficient rise in the saw-tooth current wave to complete the vertical trace just before the flyback. Compression at the start of the vertical saw-tooth current wave can cause a bright bar at the top edge of the raster.

4. A bright bar at the left or right edge of the raster can result from compression of the horizontal saw-tooth deflection current, at the start or finish of the horizontal trace.

*Fold-over.* The effect that results when the same area on the kinescope screen is scanned more than once during a scanning cycle is called *fold-over*. With picture information on the raster, part of the image appears folded over or under itself, as though the folded portion of the picture were wrapped around a cylinder. An example of fold-over is shown in Fig. 18-39. This illustrates the same damping trouble as in Fig. 18-37 but the

fold-over is more evident with a picture on the raster. However, there is a difference between fold-over in the raster and fold-over only in the picture. Figures 18-37 to 18-39 illustrate troubles in the deflection circuits causing fold-over in the raster, which is also evident in the picture. When

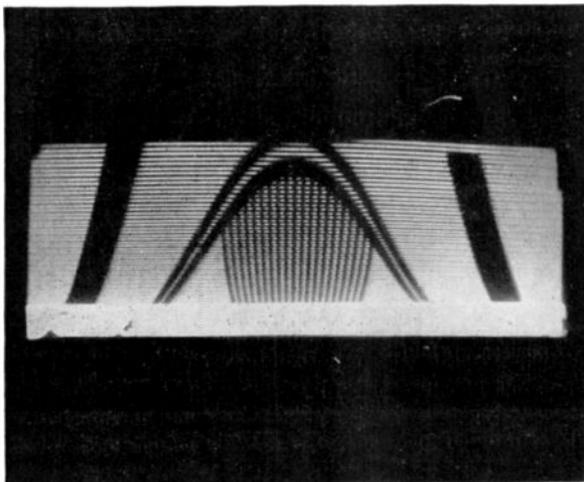


FIG. 18-38. Fold-over at bottom edge of raster and picture caused by leaky coupling condenser in grid of vertical output tube. (RCA.)

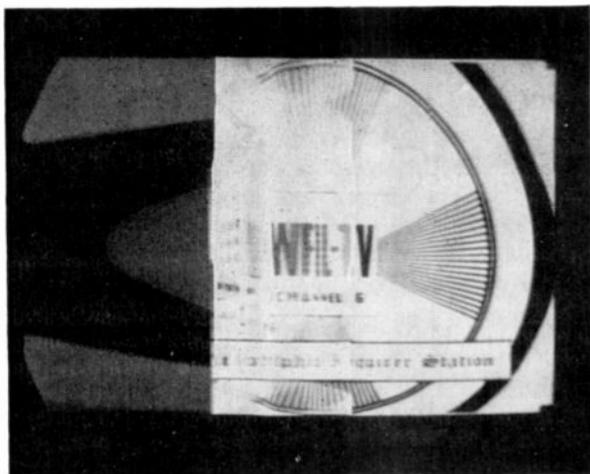


FIG. 18-39. Fold-over at left side. Same raster and horizontal damping trouble as in Fig. 18-37. (RCA.)

the fold-over effect can be seen in the picture but is not in the raster, this indicates: (1) incorrect phasing of the sync and blanking with respect to the flyback; (2) the retrace is longer than blanking time; or (3) the scanning frequency is too high.

Fold-over in the picture caused by too high a scanning frequency can be recognized by the fact that different sections of the picture are superimposed on each other. This effect can appear more readily in the vertical scanning because the vertical deflection oscillator does not have automatic frequency control. Picture information being scanned during the flyback is more common in the horizontal deflection circuits, resulting in fold-over at the left and right edges, since the tolerance for horizontal retrace within blanking time is much less than for vertical retrace. The fold-over in the picture at the left and right edges can change when different channels are selected, because blanking time is not exactly the same for all broadcast stations.

*Thin Black Lines at Left Side of Raster.* This effect is illustrated in Fig. 18-40. The thin black lines at the left side of the raster are usually

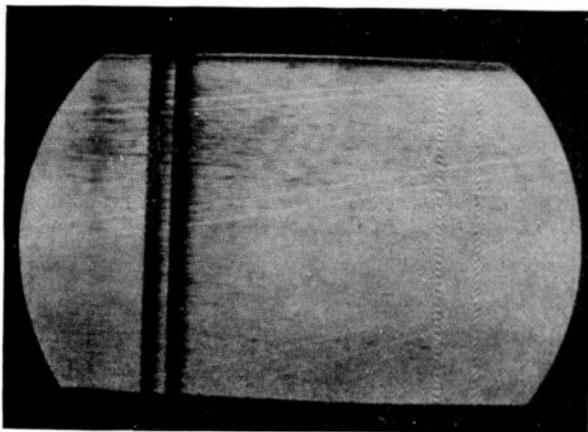


Fig. 18-40. Thin, dark lines at left side caused by Barkhausen oscillations. (RCA.)

caused by spurious oscillations produced within the horizontal output tube and radiated to the signal circuits in the receiver. The bars are black because the oscillations coupled into the signal circuits have enough amplitude to drive the kinescope grid voltage to cutoff; they are at the left side because the oscillations occur immediately after the flyback. At this time, the output circuit can drive the horizontal deflection amplifier's plate voltage negative, momentarily. While the plate voltage is negative, electrons attracted from the cathode by the screen grid are repelled from the plate. As a result, the electrons can oscillate around the screen grid. This effect is named *Barkhausen* oscillations. Their frequency is in the v-h-f range. The Barkhausen lines can be evident on any v-h-f channel, when the oscillations are converted in the front end of the receiver to a frequency within the range of the picture i-f amplifiers, but appear more often on the high-band v-h-f channels 7 to 13. To

eliminate the Barkhausen lines, the horizontal drive control can be adjusted for a different setting, a different horizontal output tube may be substituted, or a magnet can be mounted on the tube and rotated to stop the oscillations. In addition, the receiver's built-in antenna often couples the Barkhausen oscillations into the signal circuits. Taking out the internal antenna or changing its placement may remove the Barkhausen oscillations from the signal circuits, eliminating the black lines from the picture.

Black lines at the left side of the raster, which look like Barkhausen lines, can also be produced by interference in the kinescope grid-cathode circuit from harmonic frequency components of the 15,750-cps horizontal



FIG. 18-41. Trapezoidal raster caused by defective deflection yoke. Keystoning at top and bottom shown here produced by shorted vertical coil. Keystoning at sides can be caused by shorted horizontal coil. (RCA.)

scanning current. This effect is more likely to be evident on the lowest channels. The lines are at the left side, corresponding to the time when the damper suddenly changes from the condition of no current flow to maximum current. To eliminate these lines, r-f chokes of 1 to 5  $\mu$ h inductance are inserted in the cathode or plate lead of the damper tube.

*Trapezoidal Raster.* An open or shorted deflection coil in the yoke produces keystone of the raster, as shown in Fig. 18-41. Geometrical distortions of the raster shape, which include keystone, pincushion, and barrel distortion, can be caused only by the deflection yoke, in magnetic scanning.

#### REVIEW QUESTIONS

1. Describe briefly how the scanning raster is produced. What determines the height? The width?

2. How is the vertical deflection oscillator synchronized at 60 cps? How is the horizontal deflection oscillator synchronized at 15,750 cps?
3. Why can the deflection circuits produce the scanning raster with or without sync?
4. Why does the plate current of the vertical output tube have the saw-tooth wave shape?
5. What is the function of the vertical output transformer?
6. If the picture shows crowding at the top and spreading at the bottom, what control should be adjusted to correct this? In what circuit is the control?
7. Why is damping necessary in output circuits for magnetic deflection?
8. What is meant by "ringing"? What determines the frequency of ringing?
9. Give three functions of the horizontal output circuit.
10. Why can the horizontal output stage be considered a class C amplifier?
11. In horizontal deflection, describe briefly how scanning current is produced for the right side of the trace, the retrace, and the left side of the trace.
12. Why is the damper tube cut off for the flyback?
13. What determines the horizontal flyback time?
14. Show typical wave shapes of the deflection amplifier tube's grid-driving voltage, plate current, and the current in the scanning coils for the vertical output circuit.
15. What two voltages add to produce the boosted B+ voltage?
16. What stages must be operating to produce a-c input voltage for the high-voltage rectifier in a flyback high-voltage power supply?
17. What is the function of the horizontal drive control? Describe briefly how to adjust it.
18. Draw a diagram showing the horizontal deflection coils connected in a direct-drive output circuit.
19. What components of the deflection circuits are matched to the electrical characteristics of the yoke?
20. Why is a smaller saw-tooth condenser used for the horizontal deflection oscillator compared with the vertical deflection oscillator?
21. Referring to the schematic diagram in Fig. 18-31:
  - a. Which RC network provides the trapezoidal wave shape for the output voltage of the vertical deflection oscillator?
  - b. Why is the plate decoupling filter  $R_{433}C_{422}$  used in the plate circuit of the horizontal oscillator?
  - c. What stages use boosted B+ for plate-supply voltage?
  - d. State briefly the function of each of the following components:  $R_{404}$ ,  $R_{405}$ ,  $R_{406}$ ,  $R_{408}$ ,  $C_{405}$ ,  $C_{406}$ ,  $R_{407}$ ,  $C_{407A}$ ,  $R_{410}$ ,  $R_{411}$ ,  $R_{434}$ ,  $R_{436}$ ,  $C_{421}$ ,  $R_{440}$ , and  $C_{424}$ .
22. Referring to the schematic diagram in Fig. 18-31, describe what will be seen on the kinescope screen for each of the following component troubles:
  - a. Open-plate winding in vertical blocking oscillator transformer  $T_{401}$ .
  - b. Open-cathode resistor  $R_{432}$  in horizontal oscillator.
  - c. Low emission in vertical output tube 6S4.
  - d. Short across drive control  $C_{423}$ .
  - e. Leakage in coupling condenser  $C_{431}$  to the horizontal deflection coils.
  - f. Open fuse  $M_{401}$ .
23. When internal vertical blanking is used what is the polarity of the vertical flyback pulses coupled to the kinescope grid?
24. When the horizontal flyback pulses are used for keying the a-g-c circuit, what is their polarity at the plate of the keyed a-g-c amplifier?
25. Draw a block diagram illustrating the equipment setup for producing bar

patterns on the kinescope screen. What frequency will be used to obtain 10 pairs of horizontal bars?

26. The raster has one 60-cps sine-wave ripple at each side, with bulge opposite bulge. Give one possible cause of this trouble.

27. Explain briefly one method of removing the picture to observe the raster alone.

28. Explain briefly one method of testing the vertical output stage to see if it is normal, when there is no vertical deflection.

29. Give one method of determining whether the horizontal oscillator is operating, when there is no horizontal output.

30. Give two causes of insufficient width in the raster.

31. Give two causes of narrow white vertical bars at the left side of the raster.

32. What causes fold-over with a wide white bar at the left side of the raster and reduced width?

33. Give one cause of a bright bar at the bottom of the raster.

34. What is one cause of thin dark bars at the left side of the raster? How can the bars be eliminated?

35. How can you distinguish between fold-over in the raster and fold-over in the picture? Give one cause of fold-over at the left side in the picture only. Give one cause of fold-over at the left side in the raster.

## CHAPTER 19

### PICTURE I-F AMPLIFIERS

The superheterodyne television receiver uses three or four i-f stages to amplify the picture i-f signal output of the mixer stage in the front end, as illustrated in Fig. 19-1. With approximately 0.5 to 5 mv of picture i-f signal from the mixer, the i-f amplifier can supply for the detector a signal level of about 5 volts, peak amplitude. Practically all the receiver gain and selectivity for the picture signal is obtained in the i-f section. In addition, it is here that the receiver's response is given the characteristics required to compensate for the vestigial-side-band transmission. Although the sound i-f signal is amplified along with the picture

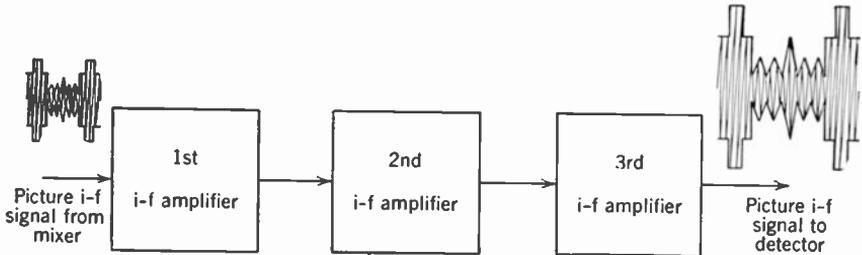


FIG. 19-1. The picture i-f stages amplify the picture signal from the mixer to supply enough signal to the video detector.

i-f signal in the common i-f section of intercarrier-sound receivers, the requirements of picture i-f amplification are the same as in split-sound receivers.

**19-1. Picture I-F Response.** Figure 19-2 shows the typical over-all i-f response curve required for the picture signal. Note the inversion of signal frequencies caused by operation of the local oscillator above the incoming signal, making all signals that are of higher frequency than the picture carrier in transmission lower than the picture carrier in the i-f section of the receiver. Table 19-1 illustrates this frequency inversion for a receiver with a picture intermediate frequency of 45.75 Mc tuned to channel 4 (66 to 72 Mc). The correct operating frequency for the local oscillator on this channel is 113 Mc. All r-f signal frequencies are mixed with the

oscillator signal in the converter stage, and the difference frequencies are developed in the output of the converter to be coupled to the i-f stages. The difference frequency between the picture carrier and the oscillator is 45.75 Mc, which is the picture carrier intermediate frequency. The frequency difference between the sound carrier and oscillator is less because

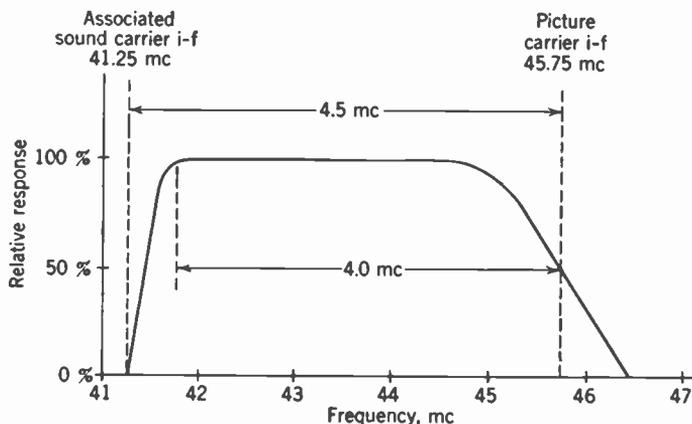


FIG. 19-2. Desired response curve of amplifier with picture i-f of 45.75 Mc.

of the higher frequency of the sound carrier, producing the sound carrier 41.25-Mc intermediate frequency. The side carrier frequency for 2-Mc video modulation is also given in Table 19-1 as an example to show that side-band frequencies transmitted above the carrier frequency are inverted by the frequency conversion to become lower side-band frequencies in the i-f stages. This inversion of the side-band frequencies occurs only because the local oscillator is operating above the incoming signal frequencies, which is the most common case because the oscillator tuning range is less than what it would be with operation below the signal frequencies.

TABLE 19-1. INVERSION OF SIGNAL FREQUENCIES IN A RECEIVER WITH 45.75-Mc PICTURE CARRIER INTERMEDIATE FREQUENCY, TUNED TO CHANNEL 1  
Local oscillator frequency = 113 Mc

	Transmitted signal frequency, Mc	Intermediate frequency, Mc
Upper edge of channel 4.....	72	41
Sound carrier of channel 4.....	71.75	41.25
An upper side carrier frequency of channel 4....	69.25	43.75
Picture carrier of channel 4.....	67.25	45.75
Lower edge of channel 4.....	66	47
Sound carrier of adjacent channel 3.....	65.75	47.25

**Bandwidth.** The i-f section must provide suitable amplification for the picture carrier and all side-band frequencies of the picture signal in order to allow reproduction of the fine detail in the picture associated with the high video frequencies. Lack of adequate amplification for the side-band frequencies produced by high-frequency video modulation is equivalent to poor high-frequency response in the video amplifier section and will produce the same loss of horizontal detail in the reproduced picture. This is illustrated by the resolution chart of Fig. 19-3, showing the continuity of r-f, i-f, and video signal frequencies. The extent to which the divisions in a vertical wedge of the test pattern can be resolved indicates the merit of

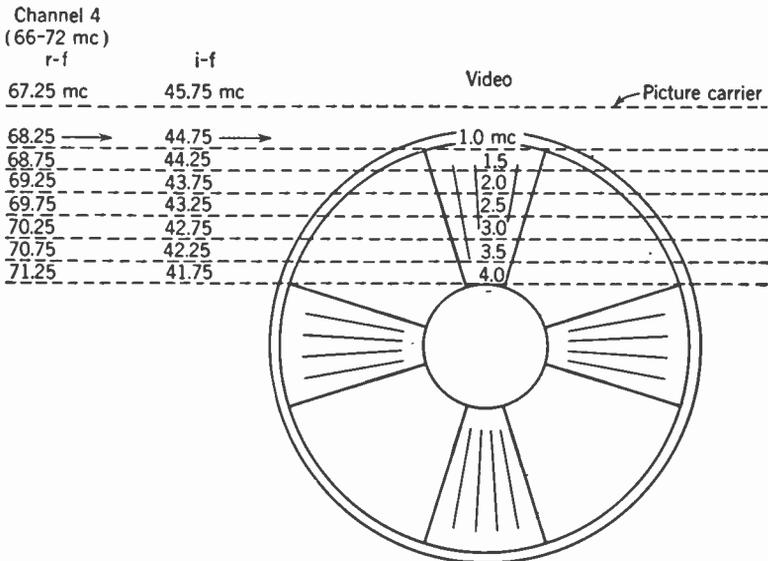


FIG. 19-3. A resolution chart to illustrate the continuity of r-f, i-f, and video signal frequencies for channel 4.

the receiver's high-frequency response and the ability to reproduce fine detail in the horizontal direction.

Optimum response in the i-f section requires a bandwidth of approximately 4 Mc between the picture carrier intermediate frequency and the lower edge of the pass band as shown in Fig. 19-2. A bandwidth of about 3 Mc may be used in some receivers, however, in order to provide more gain per stage.

**Compensation for Vestigial-side-band Transmission.** The picture signal is transmitted with vestigial-side-band transmission in order to make maximum use of the assigned channel in transmitting the high-frequency components of the signal required for good picture detail. In this quasi-single-side-band method of transmission, the unwanted lower side-band

frequencies are removed at the transmitter only for those frequencies separated from the carrier by approximately 0.75 Mc or more. Therefore, for video modulating frequencies less than 0.75 Mc, both the upper and lower side-band frequencies plus the carrier are transmitted as a normal double-side-band signal. However, for video modulating frequencies higher than 0.75 Mc the lower side-band frequencies are sufficiently separated from the picture carrier to fall outside the channel and are not transmitted. The upper side bands are transmitted for all frequencies up to about 4 Mc above the picture carrier. As a result, the lower video modulating frequencies up to 0.75 Mc are transmitted as double-side-band signals, while the higher video frequencies are transmitted as single-side-band signals.

Since all the modulation energy of a modulated carrier wave is in the side bands, the single-side-band signals for video frequencies above 0.75 Mc have only one-half the effective modulation of the lower frequency double-side-band signals. If the receiver i-f response were the same for all signal frequencies, the demodulated output from the video detector would be twice as great for video signals below 0.75 Mc as for the higher video-frequency signals. This would be a distortion of the picture signal, since different video signal frequencies would be given widely different amplitudes because of the method of transmission and not because of picture content.

In order to equalize the effect of this vestigial-side-band transmission, the over-all i-f response for the receiver is adjusted to give the picture carrier intermediate frequency approximately 50 to 60 per cent of the maximum response obtained for the side-band frequencies associated with the higher video frequencies, as shown in Fig. 19-2. With this placement of the picture carrier intermediate frequency, the two side bands of a double-side-band signal are given an average response of 50 per cent, compared with the 100 per cent response for single-side-band frequencies. Therefore, the output from the video detector will be the same for all video modulating signals having the same amplitude, regardless of whether they are transmitted with single or double side bands. When the picture carrier intermediate frequency is placed higher up on the response curve, giving it more than 50 to 60 per cent response, the effect is to emphasize the video modulating frequencies up to 0.75 Mc and to produce a relative attenuation of the higher frequencies that correspond to fine detail. If the picture carrier intermediate frequency response is too low, the lower video frequencies will be attenuated. This may cause insufficient contrast in the picture.

*The Associated Sound.* The associated sound carrier frequency is very close to the picture signal's side-band frequencies for which maximum response is desired in the picture i-f amplifier, as shown in Fig. 19-4. Since the associated sound i-f carrier is always 4.5 Mc from the picture

i-f carrier, a sharp cutoff at the sound carrier side of the picture i-f response is needed in an i-f amplifier with 4-Mc response. The sharp cutoff is obtained by using wave traps resonant at the associated sound i-f carrier frequency to reduce the sound i-f response in the picture i-f amplifier. In a split-sound receiver, which has separate amplifiers for the sound and picture i-f carrier signals, the sound traps reject the sound i-f signal to produce practically zero response, as shown in *a* of Fig. 19-4. With intercarrier receivers, however, the response of the sound i-f signal in the common i-f amplifier is approximately 2 to 5 per cent. This is the relative strength required to obtain the 4.5-Mc sound signal as the beat frequency between the picture and sound carriers in the second detector.

**19-2. The Intermediate Frequency.** A choice for the value of picture carrier intermediate frequency lies between 8.5 and 54 Mc, with the sound carrier frequency automatically placed 4.5 Mc below. The top frequency

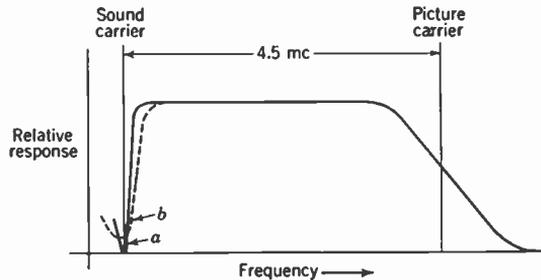


FIG. 19-4. Response for associated sound i-f carrier frequency in picture i-f amplifier. (a) Practically zero response for receivers with separate sound i-f amplifier. (b) 2 to 5 per cent response in common i-f amplifier of intercarrier-sound receivers.

limit is 54 Mc because the picture intermediate frequency must not fall in any of the television channels. The lower frequency limit for the picture carrier intermediate frequency is 8.5 Mc, because the sound intermediate frequency, which is 4.5 Mc lower, should not be in the video-frequency range. Generally, higher values are desired for the intermediate frequencies for the following reasons:

1. The higher the intermediate frequency the greater is the image-frequency rejection. The image has a frequency equal to twice the intermediate frequency plus the desired frequency and will be separated from the desired signal by a greater amount with a higher intermediate frequency.
2. Less oscillator signal is coupled to the antenna circuit from the local oscillator of the receiver when higher values are used for the intermediate frequency, because there is a greater separation between the resonant frequency of the local oscillator and the r-f tuning circuits.
3. The extremely wide bandwidth required for the picture i-f stages can more easily be obtained at higher frequencies, since the bandwidth is then

a smaller percentage of the resonant frequency of the tuned coupling circuits.

4. Filtering of the intermediate frequency signals from the desired video signal in the output of the video detector is more easily accomplished with higher values of intermediate frequency because of the greater frequency difference between the intermediate frequency and the highest video frequency.

Lower values of intermediate frequency have the advantages of allowing more gain for an i-f stage, with greater stability. Instability is caused by regenerative feedback between stages amplifying the same signal frequencies. This can result in the amplifier oscillating at its resonant frequency with or without any input signal.

In any case, the value chosen for the intermediate frequencies should not be too close to a band of frequencies assigned to other services that can produce excessive interference if these signal frequencies or their harmonics are in the i-f pass band of the television receiver. Also important is the position in the frequency spectrum of the image frequency when the receiver is tuned to different channels. If the image is in another television channel, or in the FM broadcast band of 88 to 100 Mc, these signal frequencies can produce r-f interference in the reproduced picture. Another consideration is the frequency of the local oscillator for each channel. If the local oscillator is in the frequency range of a television channel, the oscillator in a receiver operating on one channel can produce r-f interference in a nearby receiver tuned to a different channel. Consideration of these factors has led to the present RETMA recommended standard of 45.75 Mc and 41.25 Mc for the picture and sound i-f carrier frequencies in television receivers. Important advantages of these i-f values are, that for any v-h-f channel from 2-13: (1) One v-h-f channel cannot be an image of another v-h-f channel; (2) any frequency in the FM broadcast band of 88 to 108 Mc cannot be an image of a television channel; (3) the local oscillator frequency cannot be in any v-h-f television channel. The previous RETMA standard i-f values are approximately 26 Mc and 21.5 Mc.

**19-3. I-F Amplification.** A typical i-f amplifier stage using a double-tuned coupling transformer is illustrated in Fig. 19-5. The primary and secondary inductances  $L_p$  and  $L_s$  in the i-f transformer  $T$  tune the shunt capacitance  $C_{out}$  and  $C_{in}$ , each about 5 to 10  $\mu\mu\text{f}$ , to the desired frequency in the i-f range.  $L_p$  and  $L_s$  each have an individual tuning adjustment, which is generally a powdered-iron core. Sufficient bandwidth is provided by the shunt damping resistors  $R_2$  and  $R_3$ . The screen dropping resistor  $R_s$  and bypass condenser  $C_s$  supply the required screen-grid voltage for the pentode i-f amplifier tube. The plate decoupling resistor  $R_f$  isolates the i-f stage from the common B supply line and drops the B+ to

the required plate voltage.  $R_1C_1$  is a decoupling filter in the grid circuit connecting to the common a-g-c line for all stages using the a-g-c bias. In addition, cathode bias is produced by  $R_k$ , which is unbypassed in order to stabilize the input capacitance of the stage with variations in gain caused by changing the bias. Since a pentode i-f tube is used with relatively low value of plate load impedance, as in the video amplifier, the gain of the i-f stage is equal to  $g_m Z_L$ . The load impedance for the i-f signal is the tuned circuit. With damping, the impedance of the tuned i-f circuit is about 5,000 ohms. Therefore, using an i-f amplifier tube having a transconductance of 4,000  $\mu$ mhos with a bias of  $-3$  volts and 150 volts on the plate, produces a gain of 20 times in voltage for a single i-f stage.

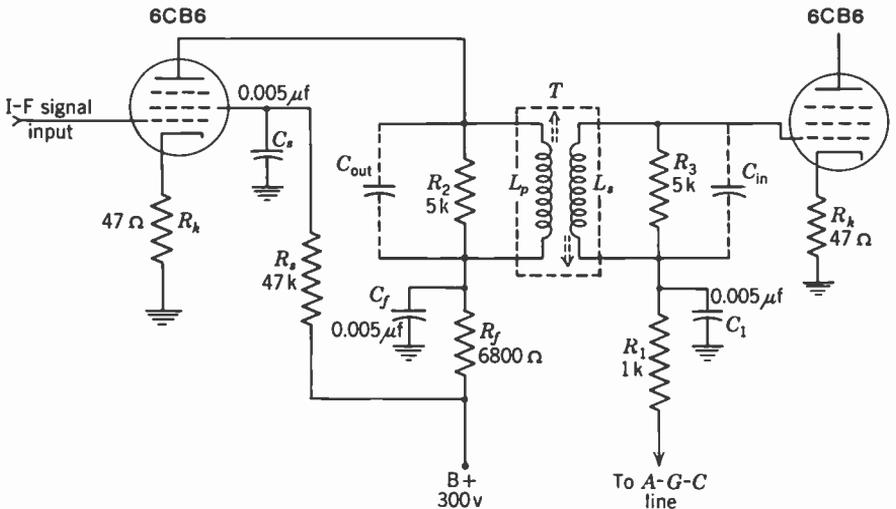


FIG. 19-5. Picture i-f amplifier stage with double-tuned transformer.

Three or four i-f amplifier stages are generally used to provide a total i-f gain of 2,000 to 20,000, depending on the amount of bias. Typical i-f amplifier tubes are the miniature glass pentodes 6AG5, 6AU6, 6BA6, 6CB6, and 6CF6, which feature high transconductance with low input and output capacitances.

*I-F Coupling Circuits.* Several basic coupling arrangements for the i-f amplifier are shown in Fig. 19-6, with their individual response curves. Triple-tuned circuits are not generally used because they are extremely critical with respect to tuning and difficult to adjust. All the circuits are shown in their equivalent form without provision for grid bias, plate voltage, or any blocking condensers that may be necessary. Only the input and output capacitances  $C_{in}$  and  $C_{out}$  are used to tune the coils to the desired frequency, without any trimmer condensers, because minimum shunt capacitance allows higher gain for a given bandwidth. Tuning is usually done by means of a variable core in the inductance.

The simplest arrangement consists of a single-tuned circuit as the i-f amplifier plate load impedance, as shown in Fig. 19-6a. The amplifier response for such an arrangement follows the impedance characteristic of a single parallel resonant circuit, with maximum gain at the resonant frequency and symmetrical response about this center frequency. Flat response over a wide range of frequencies is not feasible with such an arrangement, but the single-tuned circuit coupling is useful if the resonant frequencies for cascaded stages are staggered above and below the center

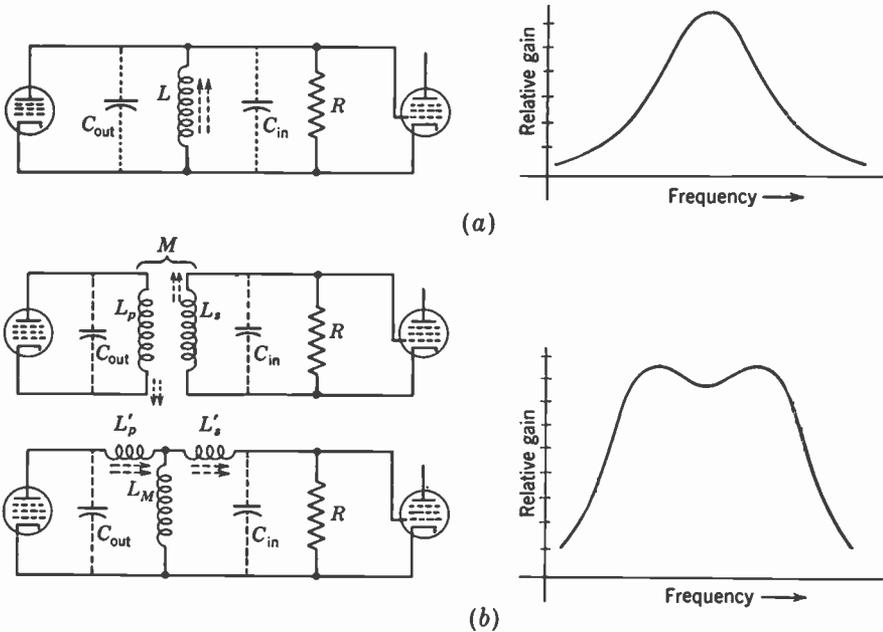


FIG. 19-6. Basic picture i-f coupling circuits. (a) Single-tuned. (b) Double-tuned with inductive coupling.

frequency of the desired over-all response. Also, the single resonant peak of such a stage can be used to compensate for the double-peak response obtained with a double-tuned coupling circuit.

Figure 19-6b shows a double-tuned coupling transformer arrangement. Both primary and secondary are tuned to the same frequency to produce a typical double-peaked response curve, with close coupling used in order to obtain greater bandwidth. Coupling is obtained by means of the mutual inductance between primary and secondary, which provides a coupling impedance that is common to both tuned circuits. The same results can be obtained with the use of the coupling coil  $L_M$ , since this can be given a value equal to the mutual inductance between primary and

secondary in the transformer, and the impedance of  $L_M$  is also common to both tuned circuits. Using the coil  $L_M$  as the coupling impedance has the advantage of facilitating control of the amount of mutual coupling, when compared to adjustment of the spacing between primary and secondary to vary the mutual inductance. For equivalence between the two types of double-tuned circuits,

$$L_M = M, \quad L'_p = L_p - M \quad \text{and} \quad L'_s = L_s - M$$

Capacitive coupling can be used between the two circuits, instead of  $L_M$ , but the gain is less than for inductive coupling.

Close coupling is used in the case of mutually coupled double-tuned circuits for the picture i-f amplifier, in order to increase the bandwidth. The effect of increased coupling is illustrated in Fig. 19-7. As the coupling is increased, the secondary current increases and has a greater effect on the primary current, producing two peaks to broaden the

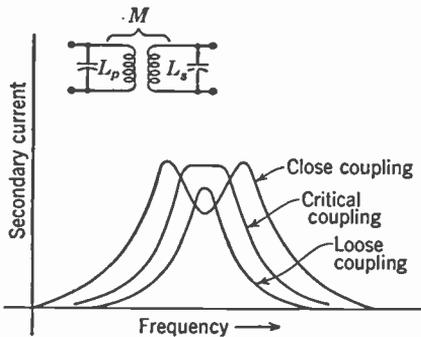


FIG. 19-7. Illustrating the increased bandwidth obtained with an increase in mutual coupling.

frequency response. The primary and secondary circuits are tuned to the same frequency but the mutual coupling results in peaks at two frequencies, above and below the resonant frequency of each circuit by itself. Note that there is only one resonant peak with narrower bandwidth when a double-tuned transformer has loose coupling or critical coupling between the primary and secondary, similar to the response of a single-tuned stage.

**Damping Resistors.** To obtain the bandwidth required for the picture i-f stages, damping resistors are usually connected in parallel with the tuned circuits, as illustrated by  $R_2$  and  $R_3$  in Fig. 19-5. Typical values for a damping resistor are 5,000 to 15,000 ohms. The effect of the shunt

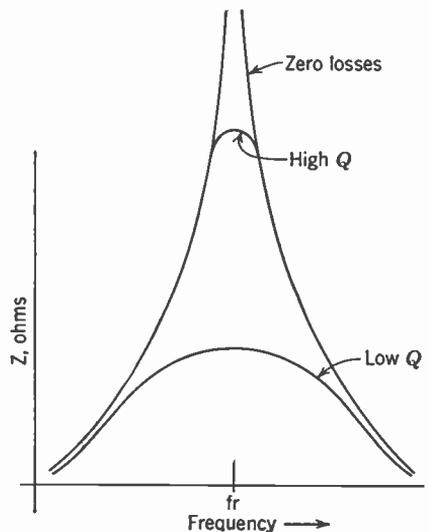


FIG. 19-8. Effect of circuit  $Q$  on impedance frequency response for a single parallel resonant circuit.

damping resistor is to lower the  $Q$  of the tuned circuit, as an equivalent resistance in series with the inductance. Lower values of shunt damping resistance are equivalent to increased resistance in series with the inductance in the tuned circuit, reducing the  $Q$  and increasing the bandwidth of the circuit. Referring to Fig. 19-8, it is shown that with lower values of  $Q$  the impedance of the tuned circuit is relatively uniform over a wider range of frequencies. It should be noted that the input resistance of the grid-cathode circuit of the i-f amplifier tube may be low enough to produce appreciable damping for the previous plate circuit at the intermediate frequencies used. The resistance of the second detector diode provides appreciable damping on the last i-f stage. Double-tuned circuits may be shunt damped at both ends or damped at one end only, with higher impedances possible for the case of single-end damping but less

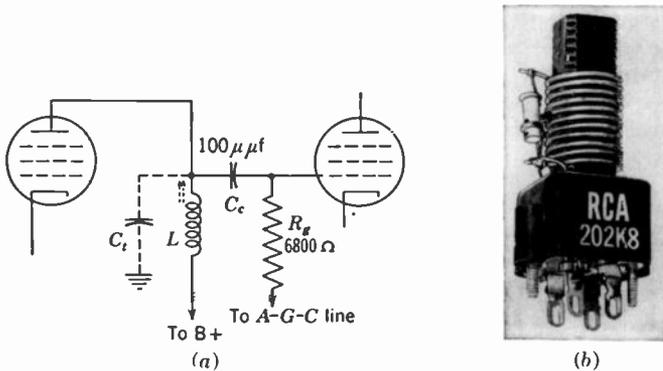


FIG. 19-9. (a) Single-tuned i-f amplifier stage with  $RC$  grid coupling circuit. (b) Single-tuned 22-Mc i-f coil with wave trap. Inductance 3.1 to 5.2 microhenrys.

critical tuning obtained with double damping. Shunt damping is generally used in preference to a damping resistor in series with the tuned circuit, because any series damping resistance has the effect of producing unequal amplitudes for the two peaks of the typical response curve obtained for a double-tuned circuit with tight coupling.

*Single-tuned I-F Stages.* Single-tuned i-f stages are commonly used in stagger-tuned i-f amplifiers, with the individual resonant frequencies staggered above and below a frequency in the middle of the i-f pass band. The coupling arrangement for a single-tuned i-f amplifier stage is illustrated in Fig. 19-9a, with the conventional  $R_p C_c$  network coupling the i-f signal voltage from the plate of the amplifier to the grid of the next stage. The single parallel resonant circuit formed by  $L$  with the total shunt capacitance  $C_i$  is the plate load impedance for the amplifier to develop the output signal voltage. Damping across the tuned circuit is provided by  $R_p$ . A typical single-tuned i-f coil is shown in Fig. 19-9b.



The single-tuned i-f amplifier in Fig. 19-10 does not require any coupling condenser because of the *bifilar* winding of the i-f coil. As shown in Fig. 19-11, the coil is wound with two conductor wires, each insulated

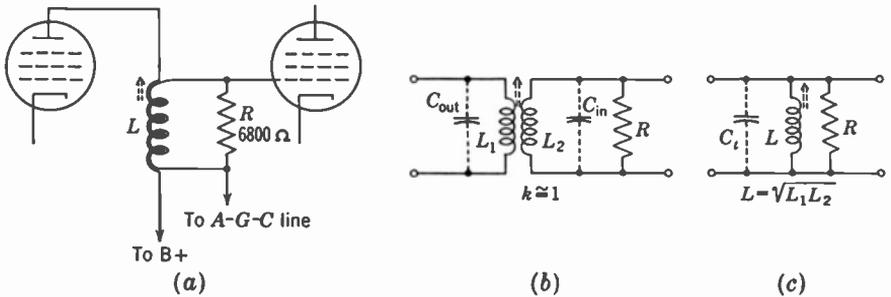


FIG. 19-10. Single-tuned i-f amplifier stage with bifilar i-f coil. (a) Amplifier circuit. (b) Circuit of bifilar coil.  $C_{out}$  is shunt capacitance of amplifier plate circuit, and  $C_{in}$  is grid input capacitance of next stage. (c) Equivalent circuit of bifilar coil.  $C_t$  is total shunt capacitance.

from the other. One conductor is the plate winding that connects to B+, while the other conductor is the grid winding. Since the i-f signal voltage inductively coupled into the grid winding is applied to the grid circuit

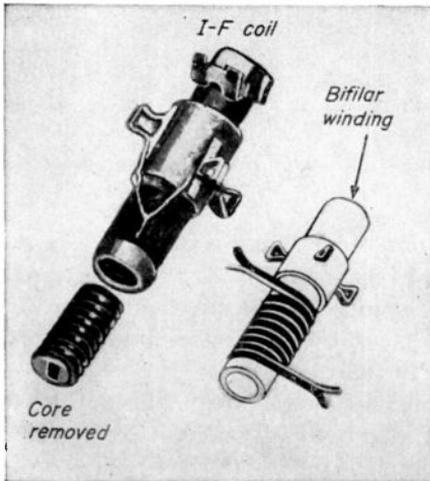


FIG. 19-11. Construction of a bifilar i-f coil. (RCA.)

of the next stage, no coupling condenser is necessary. A grid load resistor is not required either, since the grid winding provides the d-c return path from grid to cathode. However, the grid resistor  $R$  is used as a damping resistor for the tuned circuit. The advantage of the bifilar coil is that it provides the response of a single-tuned amplifier without the need for a coupling condenser to the next stage. Elimination of the coupling condenser is helpful because it tends to be charged by noise pulse voltages, causing too much negative bias trailing after noise peaks in the signal. This

effect can cause white tails to appear at the right of horizontal black streaks produced by noise pulses. Figure 19-10b and c illustrate how the two windings of the bifilar coil are equivalent to a single inductance  $L$  for the plate load tuned circuit.

*Over-all Picture I-F Response.* It should be noted that the resonant

frequency of the individual i-f coupling circuits will not generally be the i-f picture carrier frequency. Therefore, peaking the stages at this frequency will not produce the required picture i-f response because the individual resonant frequencies are usually staggered above and below the center of the i-f pass band. Although none of the individual response curves provides the required i-f response, the over-all performance of the cascaded i-f amplifiers is equal to the product of the individual response curves. These can be combined to produce the desired picture i-f response, with wave traps used when necessary to produce the required amount of attenuation for the associated and adjacent sound carrier frequencies at the edges of the picture i-f pass band.

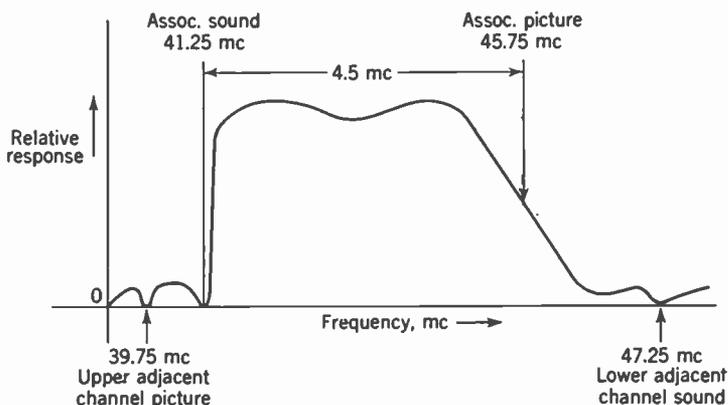


FIG. 19-12. Location of carrier frequencies for the associated sound and picture, lower adjacent channel sound, and upper adjacent channel picture, in the picture i-f response.

**19-4. Sound Traps.** A resonant circuit tuned to reject an undesired frequency is called a *wave trap*, or simply a *trap* circuit. To reject undesired sound signals, the picture i-f amplifier usually has wave traps for the associated sound and adjacent sound signals. The associated sound is the signal in the channel to which the receiver is tuned. As illustrated in Fig. 19-12, the associated sound carrier frequency is at the lower frequency edge of the picture i-f response, 4.5 Mc below the picture carrier frequency, because of inversion of the intermediate frequencies when the local oscillator operates above the r-f signal frequencies. The adjacent sound is the signal in the lower adjacent channel. When the receiver is tuned to channel 4, for instance, the sound signal of channel 3 can interfere with the picture i-f signal. The lower adjacent channel's sound i-f signal is 6 Mc higher than the associated sound i-f signal and 1.5 Mc above the picture i-f carrier, as shown in Fig. 19-12. The adjacent sound carrier signal is troublesome only for the lower channel because the sound carrier in the upper adjacent channel will produce an i-f signal completely

removed from the picture i-f response. Adjacent channels are not assigned in the same area, but in some locations there may be reception of signals from transmitters in different areas and on adjacent channels.

*Sound Bars in the Picture.* This is illustrated in Fig. 19-13. When the picture i-f circuits have a sharp sloping response at the associated sound carrier frequency and there is enough sound signal amplitude, the frequency variations of the FM signal can be converted to an appreciable amount of AM sound signal. Coupling this to the second detector results in an interfering audio signal in the video amplifier. The audio signal at the kinescope grid-cathode circuit produces the bar interference pattern shown in Fig. 19-13. Bars in the picture caused by the associated sound signal can be recognized by the fact that they vary in width and number

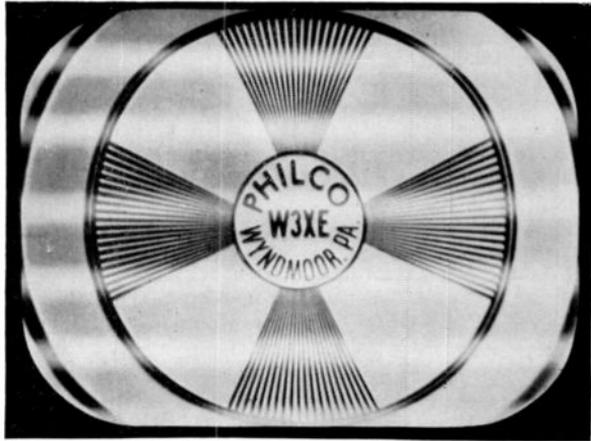


FIG. 19-13. Sound in the picture. The frequency of the audio signal interference at the kinescope grid is about 360 cps. (Philco Corporation.)

in accordance with the audio modulation as the people in the picture speak, and the bars disappear when there is no voice.

*Wave-trap Circuits.* Figure 19-14 illustrates the three main types of wave-trap circuits generally used. In *a* the parallel resonant trap circuit is connected in series with the succeeding input circuit. The trap is tuned to the rejection frequency, providing a very high resistance at the parallel resonant frequency. Most of the undesired signal voltage is developed across the trap circuit, therefore, with little voltage coupled to the next grid circuit at the rejection frequency. The trap shown in *b* is a series resonant circuit connected in shunt with the grid input circuit, with the trap tuned to the rejection frequency. At this frequency the grid input circuit is shunted by the very low resistance of the series circuit tuned to resonance. Therefore, very little voltage is developed across the grid-cathode circuit at the rejection frequency. The absorption type

of wave trap in *c* is commonly used in an i-f amplifier having a single-tuned circuit as the plate load. The absorption trap is tuned to the rejection frequency and is inductively coupled to the plate load inductance. At resonance, maximum current flows in the absorption trap circuit. As a result, enough current is coupled from the primary at the trap frequency to decrease the Q of the tuned circuit plate load impedance sharply for the rejection frequency, reducing the gain of the stage at the undesired signal frequency.

In all cases, the wave trap is tuned to the frequency to be rejected. With a sound i-f carrier of 41.25 Mc, as an example, the traps in the picture i-f circuits are tuned to 41.25 Mc to attenuate the associated sound and 47.25 Mc to reject the lower adjacent channel sound. Usually, the

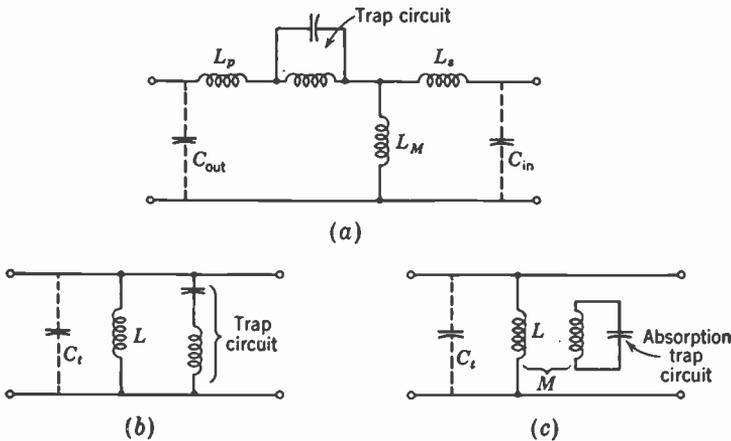


FIG. 19-14. Three types of wave-trap circuits for rejecting an undesired signal frequency. (a) Parallel resonant trap in series with load. (b) Series resonant trap in shunt with load. (c) Absorption trap inductively coupled to load.

coil of the trap circuit has an adjustable slug, which varies the inductance to tune the trap for minimum output at the rejection frequency.

**Sound Take-off.** It should be noted that, although the associated sound signal must be attenuated in the picture circuits, this is the desired signal for the sound circuits. Since maximum signal is developed in the trap itself, the trap for the sound signal can also function as the sound take-off circuit, coupling the associated sound signal to the sound i-f amplifier.

**19-5. Picture I-F Alignment.** Correct alignment of the picture i-f stages is very important in obtaining a good picture, since the composite video signal for the kinescope is the envelope of the picture carrier signal amplified in the i-f stages. In the modulated i-f signal, the side-band frequencies farthest from the picture carrier frequency contain the detailed picture information. If these are missing or attenuated, the picture will

lack horizontal detail. The intermediate frequencies close to the i-f carrier correspond to the low video frequencies that reproduce the large areas of picture information. If the i-f response for the picture carrier frequency is incorrect, this can cause weak contrast or smear in the picture. Sound bars in the picture may result from improper adjustment of the sound traps.

*Single-peaked Response.* I-F amplifier stages that have a single-peak response can be aligned by tuning for maximum output at the resonant frequency. With a signal generator supplying input at the desired frequency, resonance for each individual tuned circuit can be indicated by adjusting for maximum d-c output voltage across the second detector load resistance. Circuits that have a single-peak response are single-tuned stages, including bifilar coupling coil circuits, and double-tuned stages with loose coupling, less than critical coupling. Double-tuned picture i-f

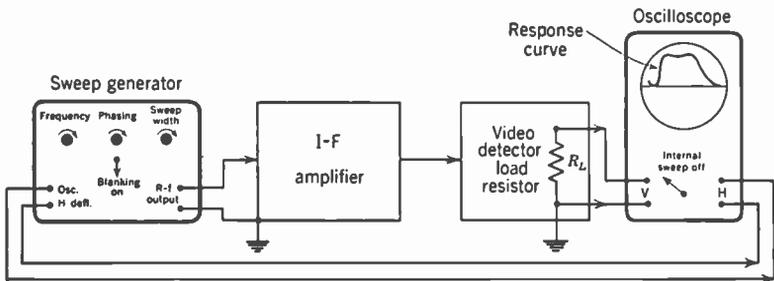


FIG. 19-15. Test equipment and connections for obtaining the visual i-f response curve.

coupling circuits are usually overcoupled, resulting in a double-peak response. They can be peak aligned if desired, however, by shunting a resistance of approximately 300 ohms temporarily across either the primary or secondary, broadening the resonance on one side, to peak the other tuned circuit at its resonant frequency. This is done for one side and then the other.

*Visual Response Curve.* The visual alignment method will be used generally to align double-tuned picture i-f stages and to check the over-all response of staggered single-tuned stages. Figure 19-15 shows the test equipment and connections used for obtaining a visual response curve. The vertical amplifier of the oscilloscope is used to obtain visual indication of the amplifier's response at different signal frequencies. Rectified i-f signal output from the video detector is connected to the vertical input binding posts on the oscilloscope, so that the amount of vertical deflection on the oscilloscope screen is a measure of the i-f output from the detector. Horizontal sweep voltage for the oscilloscope must be provided simultaneously with the vertical deflection to spread out the curve on the screen. Usually, the horizontal deflection voltage for the oscilloscope is

obtained from the signal generator used for the visual alignment. This is connected to the horizontal deflection amplifiers of the oscilloscope at the horizontal input binding posts, in place of the internal saw-tooth sweep voltage, which is switched off.

The oscilloscope need not have facilities greater than those found in the average instrument for servicing and testing. Although the signal frequencies used in the i-f alignment are above the range of the oscilloscope vertical amplifiers, which normally have a range extending up to about 100 to 500 kc, the problem is considerably simplified because the rectified signal voltage is measured. With the vertical input of the oscilloscope connected across the d-c load resistance of the detector, the deflection voltage is not the i-f signal but only a fluctuating d-c voltage whose magnitude is dependent on the amplitude of the signal input to the detector. The detector output indicates the i-f amplifier response as it varies in magnitude for the different input signal frequencies continuously provided by the sweep generator.

The sweep generator that must be used in obtaining the visual response curve is an FM signal generator, often using a reactance-tube modulator with 60-cps voltage from the power line as the modulating voltage. The sweep generator can be set to any desired center frequency in the range, and the frequency modulation produced by the reactance tube causes the instantaneous frequency of the output to swing above and below center frequency, as in a typical FM system. With a 60-cps sine-wave modulating voltage, the instantaneous frequency swings above and below center frequency at the repetition rate of 60 cps. The amount of swing, or sweep, is adjusted by varying the magnitude of the modulating voltage with the *sweep width control*. The modulating voltage itself, a 60-cps sine-wave voltage in this case, is usually taken out to binding posts on the front to serve as the horizontal deflection or time-base voltage for the oscilloscope. This is preferable to using the internal saw-tooth sweep of the oscilloscope, since the horizontal deflection voltage from the sweep generator is proportional to the amount of frequency swing produced and, therefore, produces a linear division of frequencies along the horizontal axis of the response curve. Two response curves are usually observed on the screen when a sine-wave voltage is used for horizontal deflection, one for the trace from left to right and one for the return trace. A phasing control is provided on the sweep generator for adjusting the phase of the deflection voltage, in order to obtain on the screen a single pattern consisting of the two curves superimposed on each other. In addition, a blanking switch is usually provided to blank out one trace and produce a base line on the response curve for zero reference. The two traces should be phased to be superimposed over each other before the blanking is turned on. The sweep generator used for a visual alignment must be capable of

producing the signal frequencies needed for checking the amplifier response, with a total frequency swing that is slightly greater than the bandwidth. About 8- to 10-Mc sweep width is used for picture i-f alignment. A typical commercial sweep generator is illustrated in Fig. 19-16.

Taking the visual response curve of an amplifier with a picture i-f carrier of 45.75 Mc as an example, the r-f output of the generator is connected



Fig. 19-16. TV sweep generator for alignment. R-f range 30 to 230 Mc; output up to 0.5 volt; 4.5 Mc crystal for intercarrier-sound i-f alignment; sweep width 15 Mc or 1.5 Mc; sweep rate 60 eps; marker absorption dip or oscillating pip frequency range 19 to 48 Mc. (Hickok Electrical Instrument Company.)

to the grid circuit of the amplifier under test, with the video detector output voltage connected to the vertical input of the oscilloscope. Varying the center frequency of the r-f output from the sweep generator shifts the pattern to the left or right on the screen, as the amplifier pass band is shifted toward the beginning or end of the frequency sweep. Therefore, the generator is adjusted to about 43 Mc for a pattern in the center of the screen. The response curve may appear either up or down on the oscil-

oscope screen, depending on the polarity of the video detector output and the vertical amplifiers in the oscilloscope. Also the picture i-f carrier frequency may appear either at the right or left side of the trace, depending on the sweep generator and the horizontal deflection voltage. Vertical gain on the oscilloscope is adjusted to provide suitable deflection, using as little generator output as possible. The gain of a stage can be checked roughly by noting the reduced vertical gain setting as the generator connection is moved back toward the mixer stage.

Adjusting the sweep width on the generator has the effect of spreading or crowding the pattern in the horizontal direction on the screen as the amplifier pass band is made a smaller or greater percentage of the total frequency sweep. Insufficient sweep width results in only part of the response curve appearing on the screen. The control is adjusted to about 10-Mc sweep for a pattern of convenient size horizontally. The phasing control is adjusted to obtain a single trace on the screen and the blanking is then turned on. With the sweep generator providing an i-f signal that is continuously varying in frequency through the range of 38 to 48 Mc, at the repetition rate of 60 cps, and the detector output voltage proportional to the amount of input signal, which depends on the i-f amplification, the vertical deflection voltage for the oscilloscope is a fluctuating d-c voltage that varies at the 60-cps rate in accordance with the amount of i-f gain for the different input signal frequencies. Thus the frequency response of the system is plotted on the screen of the oscilloscope in terms of relative gain vs. frequency.

The visual response curve must be marked in terms of frequency in order to measure the bandwidth and indicate important frequencies. This is accomplished by coupling very loosely to the signal input of the i-f amplifier the r-f output of a marker oscillator, which may be incorporated within the sweep generator or obtained as the unmodulated output from an r-f signal generator. Small *pips* or *birdies* appear on the screen at the marker oscillator frequency to indicate the frequency at a particular point on the pattern, as illustrated in Fig. 19-17, and the marker oscillator frequency can be varied to mark different frequencies on the response curve. The amplitude of the marker signal should be as low as possible in order to minimize its effect on the response curve and to make it easier to interpret the marker. When the marker pip is too broad, an r-f bypass condenser of about 0.01 to 0.001  $\mu\text{f}$  connected across the vertical binding posts of the oscilloscope will be helpful in making the pattern on the screen clearer. If there is difficulty in determining the position of the marker pip, close examination of the pattern will show a blank notch at about the center of the birdie, which is the exact frequency mark. This is the point at which zero beat is obtained as the fixed frequency marker is heterodyned with the varying frequency signal. Reducing the width of the frequency

sweep to make the response curve a greater part of the sweep width may make it easier to read the markers. To be of any value, the marker oscillator frequency must be accurately known, and if an r-f signal generator is used for marking it must have a very accurate frequency calibration. It is preferable that a crystal-controlled marker oscillator be used.

*Alignment Procedure.* The specific procedures that should be followed in aligning the picture i-f stages will vary for different receivers according to the manufacturer's alignment instructions. The i-f bias specified for normal gain during alignment, which is generally 3 to 5 volts, can be obtained by means of a battery bias box. This can be connected directly

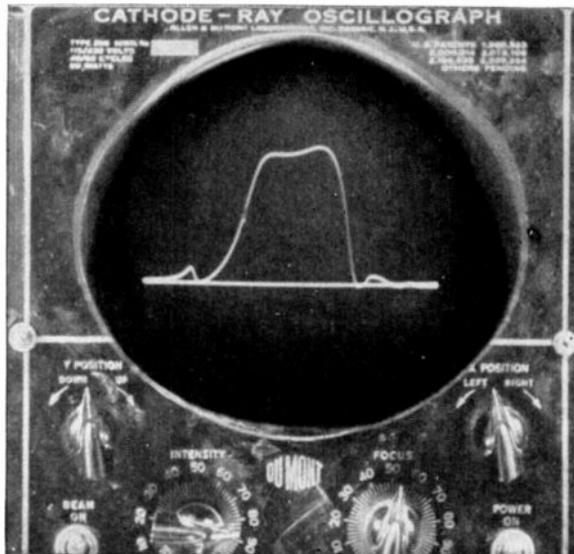


FIG. 19-17. Over-all picture i-f response curve. Marker at picture i-f carrier frequency. (From McGraw-Hill motion-picture series, *Basic Television*).

to the a-g-c line. For the over-all i-f response curve, the sweep generator is connected to the grid circuit of the mixer stage and the second detector output coupled to the vertical input of the oscilloscope. If the detector load resistor does not return to chassis ground and has no low-impedance return circuit for 60 cps, it may be necessary to take the oscilloscope vertical input signal from the plate of the first video amplifier. The receiver chassis often has jacks at the test point connections for alignment.

In the alignment procedure, the i-f wave traps are usually adjusted first, for minimum response at the rejection frequency, to provide the required response at the edges of the i-f pass band. The i-f circuits are adjusted for maximum output, indicated by the height of the curve, consistent with the required bandwidth. Adjusting the i-f tuned circuits changes the

shape of the response curve, as the height of the curve is varied for different frequencies. Marking the frequencies on the final response curve should show the picture i-f carrier frequency 50 to 60 per cent up on the sloping side, with about 1.5 Mc between the bottom and top of the slope, sufficient bandwidth across the flat top, without any dip more than about 10 to 20 per cent down, and the required response for the associated sound i-f carrier frequency. The associated sound i-f carrier has practically zero response in split-sound receivers but is about 2 to 5 per cent up in intercarrier receivers. The marker at the sound i-f carrier frequency is usually not visible because of its low amplitude.

In some cases, the alignment procedure may require individual response curves for each stage. If it is necessary to follow this stage-by-stage alignment procedure, the i-f output of the amplifier must be connected to the vertical input of the oscilloscope by means of an r-f probe detector, such as the one illustrated in Fig. 19-18. The sweep generator can be connected to the grid of the i-f stage to be aligned and the probe to the plate of the next stage, to align the i-f coupling circuit between the two stages. The probe rectifies the i-f signal for the oscilloscope and has low impedance to damp the tuned circuit to which it is connected, so that it will not alter the response of the tuned circuit being aligned.

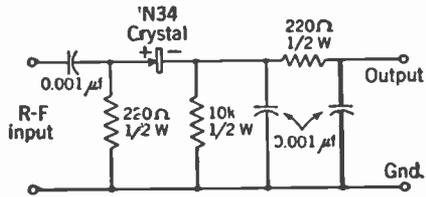


FIG. 19-18. Low-impedance r-f probe detector for alignment work.

*Peaking Staggered Single-tuned Stages.* The alignment procedure for staggered single-tuned i-f stages is relatively simple because each stage can be peaked at its resonant frequency and the response for each stage is independent of the others. The alignment can be accomplished with a d-c voltmeter and r-f signal generator. This need not be a sweep generator, since it just supplies test signal at the individual resonant frequency for each of the tuned circuits. The voltmeter, set to its lowest d-c voltage scale, can be connected across the video detector load resistor as the output indicator for all adjustments. The signal generator can remain connected to the mixer grid to supply the test signal at the different frequencies needed for each i-f stage. With the signal generator frequency set to each of the rejection frequencies, the wave traps are adjusted for minimum output on the d-c voltmeter. Then, the signal generator is set for the individual peaking frequencies, as each single-tuned i-f circuit is adjusted for maximum output at its resonant frequency, as indicated by maximum voltage on the d-c voltmeter in the detector output circuit. Battery bias is used for the alignment, instead of the a-g-c bias, so that the output indications for maximum or minimum amplitude will be sharp.

If desired, the over-all response can be checked by obtaining the visual response curve with a sweep generator and oscilloscope.

**19-6. Picture I-F Amplifier Circuits.** To provide the required bandwidth with enough gain, three or four picture i-f stages are generally used, with double-tuned coupling circuits or staggered single-tuned stages. A double-tuned stage produces more gain and bandwidth with a response curve that has a flat top and sharp slope at the sides, often called the *skirts* of the curve. However, staggered single-tuned stages have the advantages of easier alignment and less phase distortion. Wave traps are usually included in the picture i-f circuits to sharpen the skirts of the over-all i-f response curve and reject interfering signal frequencies.

*Staggered Single-tuned Amplifiers.* When cascaded amplifiers are tuned to the same frequency, the over-all bandwidth shrinks rapidly because the total gain is the product of the individual stage gains and the over-all

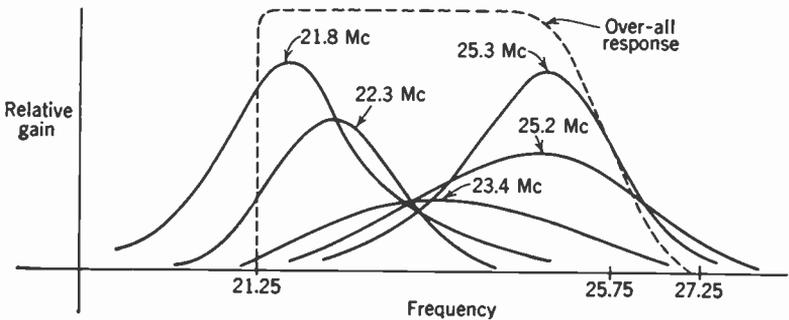


FIG. 19-19. Stagger-tuned response for i-f circuit with picture carrier frequency of 25.75 Mc. The sharp cutoff at 21.25 Mc is produced by the associated sound traps. Effect of traps on individual response curves not shown.

response curve becomes more peaked. This is undesirable for the wide-band picture i-f amplifier, but the required over-all response can be obtained by staggering the resonant frequency of individual single-tuned stages above and below a frequency in the center of the pass band. With each circuit staggered at the proper frequency and the required  $Q$  obtained by a shunt damping resistor, the over-all gain provides the required response. Figure 19-19 illustrates the response of a four-stage, staggered single-tuned picture i-f amplifier. Five individual response curves are shown, since the number of i-f coupling circuits, counting the mixer plate circuit, is one more than the number of i-f stages. Notice that, although each stage peaks at only one frequency, other signal frequencies not too far removed from resonance are also amplified but with less gain. This is why all the single-tuned stages can usually be peaked with the signal generator connected in the mixer grid circuit.

*Over-all I-F Amplifier Circuit.* Figure 19-20 shows the schematic diagram of a three-stage i-f amplifier using a double-tuned coupling circuit

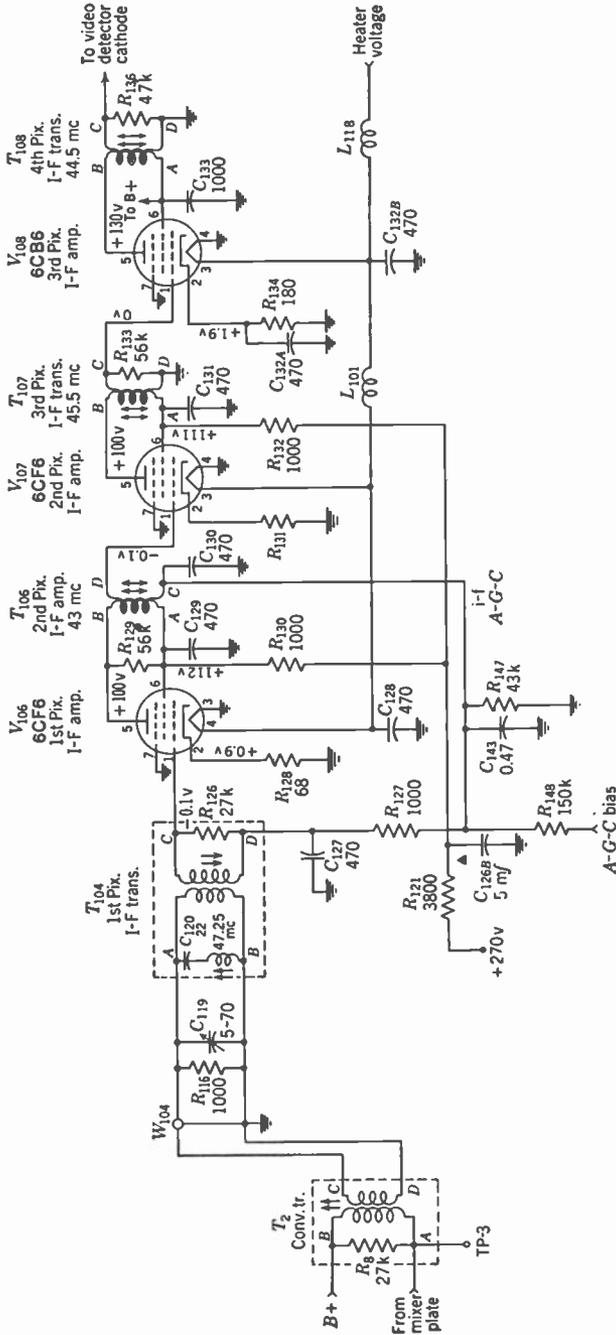


Fig. 19-20. Over-all i-f circuit with one double-tuned coupling circuit and three staggered single-tuned stages using bifilar coils. (RCA KCS 78 chassis.)

from the mixer stage and three-single tuned i-f stages with bifilar coils. Response for the amplifier, which has picture and sound i-f carrier frequencies of 45.75 Mc and 41.25 Mc, is shown in Fig. 19-21. The i-f amplifier is common to the picture and sound i-f signals, since this is an intercarrier-sound receiver, with the 4.5-Mc sound take-off circuit in the output of the second detector.

Starting at the left in the diagram, the converter transformer  $T_2$  in the plate circuit of the mixer stage is mounted on the r-f tuner subchassis. The i-f output of the mixer is link-coupled by means of a short length of 75-ohm coaxial line to  $T_{104}$  in the grid circuit of the first i-f stage  $V_{106}$ . The primary winding  $B-A$  in  $T_2$  is transformer-coupled to the winding  $C-D$ , which is connected by the cable link to the winding  $A-B$  in  $T_{104}$ . The link coupling has the advantages of minimizing radiation of the local oscillator signal from the main chassis, while allowing the r-f tuner to be located further from the i-f amplifier. Transformer coupling in  $T_{104}$  provides the i-f signal voltage across the secondary winding  $C-D$  in the grid circuit of the first i-f stage.  $R_3$  in the plate circuit of the mixer stage and  $R_{126}$  in the grid circuit of  $V_{106}$  are shunt damping resistors. The response of  $T_2$  and  $T_{104}$  with the link coupling is the same as though their primary and secondary windings were transformer-coupled, as shown by the double-tuned overcoupled response curve in Fig. 19-21a. Because of the sharp skirt selectivity with the double-tuned coupling circuit, only one wave trap is used in the i-f amplifier. This series resonant circuit across the link winding  $A-B$  in  $T_{104}$  attenuates the adjacent sound i-f carrier frequency of 47.25 Mc. The variable trimmer condenser  $C_{119}$  adjusts the slope of the i-f response curve at the low-frequency side to provide the relative gain required for the associated sound. The over-all i-f response in Fig. 19-21b is the combined response of the double-tuned coupling circuit provided by  $T_1$  and  $T_{104}$  with the response of the three single-tuned i-f stages using the bifilar coupling coils  $T_{106}$ ,  $T_{107}$ , and  $T_{108}$ .

In the plate circuit of the first i-f stage,  $T_{106}$  tunes with the shunt capacitances to form a resonant circuit peaked at 43 Mc, which is the plate load impedance for the amplifier.  $R_{129}$  is a shunt damping resistor. The required screen grid voltage is provided by  $R_{130}$ , with the bypass condenser  $C_{129}$ . No coupling condenser is needed as the bifilar coil couples the signal to the next stage. The second i-f stage peaks at 45.5 Mc.  $R_{133}$  in the grid circuit of the next stage is the shunt damping resistor. The screen dropping resistor is  $R_{132}$ , with the bypass condenser  $C_{131}$ . In these two stages, which are controlled by the a-g-c bias, the screen voltage is higher than the plate voltage, providing a more remote cutoff voltage for the control grid. In addition, the cathode resistors are unbypassed to allow some degeneration and stabilize the input capacitance with changes in a-g-c bias. The third i-f stage is resonant at 44.5 Mc, while

the shunt damping resistor is  $R_{136}$ . As the i-f signal output is coupled to the cathode of the second detector, its relatively low resistance provides additional damping across the last i-f tuned circuit. Cathode bias is provided by  $R_{134}$  with  $C_{132A}$  for the last i-f stage, which is not controlled by automatic gain control. Notice the use of the r-f decoupling chokes  $L_{101}$  and  $L_{118}$  with the bypass condensers  $C_{128}$  and  $C_{132B}$  in the heater line. I-F signal at the cathode of individual stages can be coupled by cathode-to-heater capacitance to the heater circuit, but the r-f chokes isolate each stage from the common filament line and transformer.

To align this i-f amplifier, first the single-tuned stages are peaked at their individual resonant frequencies, the wave trap set for minimum output at the rejection frequency, and the double-tuned coupling circuit in the mixer plate is then adjusted by obtaining its visual response curve. For the peak alignment, the d-c voltmeter is connected in the video

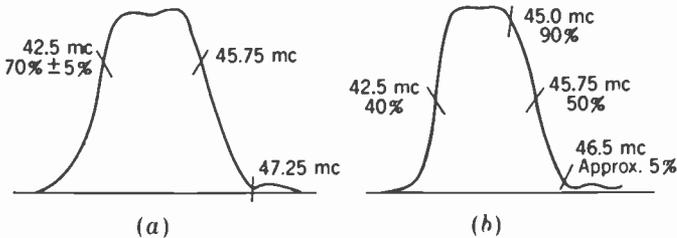


FIG. 19-21. Response curves for i-f circuit in Fig. 19-20. Relative amplitudes between *a* and *b* not to scale. (a) Double-peaked response of  $T_1$  with  $T_{104}$ . (b) Over-all i-f response.

detector output circuit, across the load resistor to ground. A battery bias of  $-5$  volts is connected to the a-g-c line. The signal generator is connected across the link at terminals *A-B* of  $T_{104}$ . Then the last i-f circuit  $T_{108}$  is peaked first, at  $44.5$  Mc,  $T_{107}$  at  $45.5$  Mc, and  $T_{106}$  at  $43.0$  Mc. The amount of signal from the generator should provide a reading of about 3 volts on the d-c voltmeter in the detector output. Next, the signal generator is set to  $47.25$  Mc and the adjacent sound trap in  $T_{104}$  is adjusted for minimum output. For aligning the double-tuned coupling circuit of  $T_2$  and  $T_{104}$ , a sweep generator is used, connected to the mixer grid, with a diode probe at plate pin 5 of the first i-f amplifier stage  $V_{106}$ , to provide rectified signal output for the oscilloscope.  $T_2$  and  $T_{104}$  are adjusted for maximum gain, indicated by the height of the curve, with the response shown in Fig. 19-21*a*. Then the oscilloscope can be connected to the video detector output circuit, without the probe, to show the over-all i-f response curve. If necessary,  $T_{106}$ ,  $T_{107}$ , and  $T_{108}$  may be retouched to provide the over-all response illustrated in Fig. 19-21*b*. The amount of signal from the sweep generator should provide a peak-to-peak ampli-

tude of 0.3 volt for the oscilloscope when it is connected in the first i-f stage, and 3 volts in the detector output circuit.

**19-7. Troubles in the Picture I-F Amplifier.** Since the picture i-f stages amplify the modulated picture signal that is detected to provide the video signal for the kinescope grid, troubles in this section of the receiver can cause no picture, weak picture, or poor picture quality, while the raster is normal. In addition, the sound is affected by trouble in an i-f amplifier common to both the picture and sound signals. The amount of i-f signal voltage can be checked by measuring the rectified d-c output voltage across the second detector load resistance.

*Tunable Smear.* Smear that moves in the picture when the fine tuning control is varied is called *tunable smear*. As the fine tuning control varies the local oscillator frequency, the i-f picture carrier frequency changes, resulting in a different response for the i-f picture signal. Tunable smear in the picture generally indicates a distorted response curve for the i-f amplifier, with incorrect relative gain and excessive phase distortion. The following effects can occur:

1. Smear and streaking in large areas or lettering in the picture. This can be caused by excessive gain for the i-f picture carrier and a sharp sloping response for the side-band frequencies close to the carrier, resulting in phase distortion for the corresponding low video signal frequencies. Excessive gain for the low video frequencies is illustrated in Fig. 11-28.

2. Weak contrast, with poor synchronization. This can be caused by low gain for the i-f picture carrier and the side-band frequencies close to it. This is illustrated in Fig. 11-11, which shows the effects of insufficient gain for the low video signal frequencies.

3. Trailing outlines in picture. This can be caused by excessive gain, with a sharp cutoff, for the side-band frequencies farthest from the i-f picture carrier, resulting in ringing for the high video signal frequencies. As illustrated in Fig. 19-22, this effect can be severe enough to give the appearance of close, regularly spaced, duplicate images, especially when the excessive response is caused by regeneration in the i-f amplifier. Multiple outlines in the picture might be mistaken for duplicate images caused by reflections of the antenna signal, but the ringing effect produced by excessive response in the i-f or video stages is the same on all stations and the outlines are regularly spaced. Compared with the video amplifier, the i-f amplifier can produce more severe distortion, with several trailing outlines, because there are more i-f stages to contribute to the phase and frequency distortion. In addition, it should be noted that trailing outlines caused by excessive response in the i-f amplifier usually move when the fine tuning control is varied.

4. Indistinct outlines at edges of objects in the picture with insufficient

detail. This results from insufficient bandwidth in the i-f response curve, with low gain for the side-band frequencies farthest from the i-f picture carrier, resulting in loss of the high video frequencies. Insufficient response for the high video frequencies is illustrated in Fig. 11-10.

5. White halo at the right of black edges in the picture. This can be caused by an excessive dip in the i-f response curve for the middle range of frequencies, with a sharp cutoff for the side-band frequencies farthest from the i-f picture carrier. The halo effect can be seen in Fig. 11-29.

The i-f amplifier can produce each of the preceding effects individually, or more than one at the same time, depending on the i-f response. Usually, all these effects of poor picture quality caused by incorrect i-f response vary with the fine tuning control. The incorrect response may

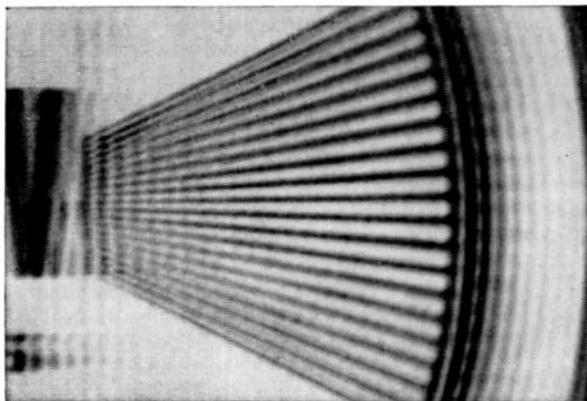


FIG. 19-22. Multiple outlines in picture caused by regeneration in the i-f amplifier. (RCA.)

be due to i-f misalignment, a defective component that alters the alignment, or regeneration due to feedback in the i-f amplifier.

*Oscillations in the I-F Amplifier.* To check whether the i-f amplifier is oscillating, the rectified i-f output can be measured across the second detector load resistor. Normally this d-c voltage in the detector output circuit is several volts with a station tuned in, which varies in amplitude as the fine tuning control is rotated but drops practically to zero when the receiver is switched to an unused channel. If the detector output voltage is comparatively high and stays the same regardless of the r-f tuning, this indicates the i-f amplifier is oscillating.

*Hum in the I-F Signal.* Hum voltage introduced in the i-f amplifier causes modulation hum, which is present only when a signal is tuned in. Therefore, the effects of the hum may be evident in the picture but not in the raster alone. Excessive hum voltage modulating the i-f signal can cause hum bars in the picture and hum in the sync.

## REVIEW QUESTIONS

1. Draw the desired i-f response curve for a bandwidth of 3.5 Mc, in an inter-carrier-sound receiver, assuming a picture intermediate frequency of 32.5 Mc. Mark the following carrier frequencies on the response curve: associated sound and picture, lower adjacent channel sound, and upper adjacent channel picture.
2. Give one possible cause of sound bars in the picture.
3. In a receiver with a picture intermediate frequency of 45.75 Mc, calculate the i-f and video signal frequencies corresponding to a side-band frequency 2 Mc higher than the picture carrier in the transmitted picture signal.
4. Give one advantage and one disadvantage of the higher i-f value of 45.75 Mc for the picture carrier, compared with 25.75 Mc.
5. What is the image for the picture carrier frequency of channel 2, 54 to 60 Mc, in a receiver with a picture intermediate frequency of 25.75 Mc?
6. Draw the schematic diagram of an i-f amplifier stage, either double-tuned or single-tuned, with a sound trap.
7. To provide more damping for wider bandwidth, should the resistance of the shunt damping resistor across an i-f tuned circuit be increased or decreased?
8. What is meant by a stagger-tuned amplifier?
9. Give one advantage and one disadvantage of an i-f amplifier with staggered single-tuned stages, compared with overcoupled double-tuned stages.
10. What is the advantage of a single-tuned i-f stage using a bifilar coil, compared with a circuit with a coupling condenser?
11. Describe briefly how an i-f amplifier with staggered single-tuned stages is aligned, using an r-f signal generator and d-c voltmeter.
12. Describe briefly how to obtain the visual response curve for the over-all gain of the i-f amplifier, with provision for marking frequencies on the curve.
13. What is the effect in the picture of insufficient bandwidth in the i-f response?
14. What is the effect in the picture of excessive relative gain for the picture carrier frequency in the i-f response?
15. What is meant by tunable smear in the picture? Give one possible cause of tunable smear.
16. Referring to the i-f amplifier circuit in Fig. 19-20 state briefly the function of the following components:  $T_{106}$ ,  $R_{127}$ ,  $R_{134}$ ,  $C_{131}$ , and  $R_{132}$ .
17. Referring to the i-f amplifier circuit in Fig. 19-20, what will be the effect on the reproduced picture and sound for each of the following component troubles:
  - a. In the screen grid circuit of  $V_{107}$ ,  $R_{132}$  opens.
  - b. In the screen grid circuit of  $V_{106}$ ,  $C_{129}$  shorts.
  - c. In the cathode circuit of  $V_{106}$ ,  $R_{128}$  opens.
  - d. In the plate circuit of  $V_{107}$ , coil  $A-B$  in  $T_{107}$  opens.
  - e. In the grid circuit of  $V_{106}$ ,  $C_{127}$  shorts.
18. Why can excessive cathode-to-heater leakage in an i-f stage cause horizontal pulling in the picture, with one pair of hum bars?
19. What is meant by an overloaded picture? How can loss of the a-g-c bias cause an overloaded picture?
20. What is one possible cause of a white halo at the right of black objects in the picture?

## CHAPTER 20

### THE R-F TUNER

The r-f amplifier, local oscillator, and mixer stages in the superheterodyne television receiver form the r-f tuning section, often called the *front end* or *tuner*. The front end has the function of selecting the picture carrier signal and the sound carrier signal of the desired channel, by converting the r-f signal frequencies of the selected station to the intermediate frequencies of the receiver.

**20-1. Operation of the R-F Tuner.** As illustrated in Fig. 20-1, the r-f stage in the front end is tuned to the frequencies of the desired channel

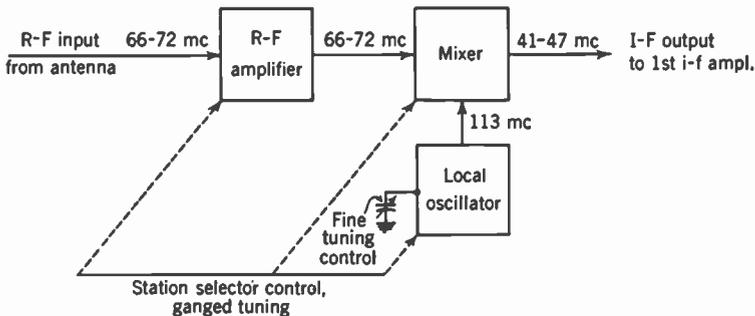


FIG. 20-1. Block diagram of r-f tuner. Oscillator frequency indicated for tuning in channel 4, in a receiver with picture i-f carrier of 45.75 Mc.

in order to amplify the r-f picture and sound carrier signals; the amplified r-f output is coupled into the grid circuit of the mixer stage to heterodyne with the local oscillator signal. The i-f output from the plate circuit of the mixer can then be amplified in the i-f section of the receiver, where more gain and selectivity can be obtained at the relatively low intermediate frequencies compared with operation at the r-f signal frequencies. As the station-selector control on the front end is varied, the r-f circuits for the r-f amplifier and mixer grid are tuned to the desired station, while the local oscillator is set to the frequency required for converting the selected channel's r-f signal frequencies in the mixer stage to the corresponding i-f signal frequencies. It is important to note that the local

oscillator tunes in the station, as the oscillator frequency determines which r-f signal frequencies are converted to intermediate frequencies in the pass band of the i-f section. The fine tuning control changes the oscillator frequency slightly to adjust the tuning exactly.

The front end is usually constructed as a complete subchassis, which is mounted on the main receiver chassis, as illustrated by the tuner shown in Fig. 20-2. This is a turret-type tuner, with coil strips on a drum

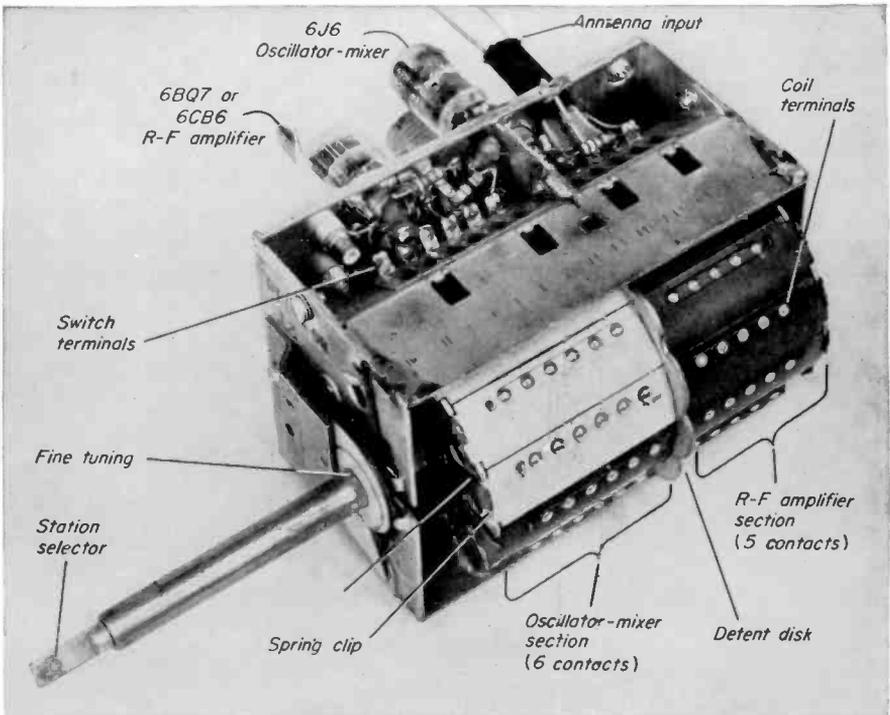


FIG. 20-2. Typical turret-type r-f tuner, with individual coil strips for each channel. Bottom shield cover and tube shields removed to show components. Schematic diagram shown in Fig. 20-18. (Standard Coil Products Co., Inc.)

rotated by the station-selector control. In switch-type tuners, the station selector usually is a rotary switch having multiple switch wafers for the r-f and oscillator sections on a common shaft for ganged tuning. A typical switch-wafer section can be seen in Fig. 20-22. Continuous tuning is also used, with variable coil or condenser sections on a common shaft. Figure 20-3 shows a continuous tuner for the u-h-f television channels, using spiral inductances.

**20-2. The R-F Amplifier Stage.** The r-f stage, or *preselector*, is optional in a superheterodyne circuit but practically all television receivers have one r-f stage to amplify the r-f picture and sound carrier signals for the

mixer stage, mainly to improve the receiver's signal-to-noise ratio. As illustrated in Fig. 20-4, double-tuned circuits are generally used for the r-f amplifier, the inductances resonating with the shunt capacitances at the desired channel frequencies. The amount of coupling between the

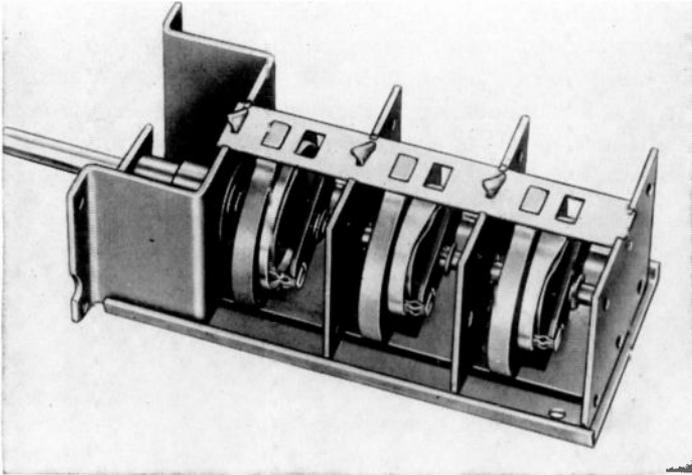


FIG. 20-3. U-H-F inductuner with three ganged spiral-inductance sections. Continuous tuning range covered in  $\frac{3}{4}$  turn of shaft. Height of unit 1.755 in. (P. R. Mallory and Co.)

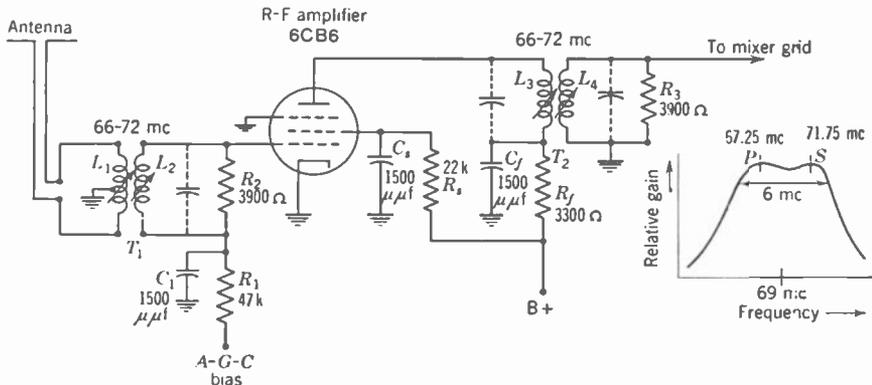


FIG. 20-4. R-F amplifier circuit with r-f response curve showing gain from antenna input to mixer grid, illustrated for channel 4. The resonant circuits are tuned to provide the same r-f response for each channel.

double-tuned circuits and the shunt damping resistance provide the required bandwidth. It should be noted, though, that the tube's input resistance from grid to cathode can provide a value of shunt damping resistance low enough for the required bandwidth at the v-h-f and u-h-f signal frequencies. A typical r-f response curve is shown in Fig. 20-4,

illustrating the relative gain of the r-f signal circuits for the frequencies in channel 4. Notice that the total bandwidth of the r-f response curve is approximately 6 Mc in order to accept both the picture and sound r-f carrier signals, and the r-f response is symmetrical about the center frequency of the channel. The r-f section must provide this response for each selected channel, with the required r-f selectivity and enough r-f gain to provide a good signal-to-noise ratio.

*Signal-to-noise Ratio.* Although most of the receiver's gain is in the i-f section, the r-f gain is of primary importance in determining the ratio of signal voltage to noise voltage generated in the receiver. The receiver noise consists mainly of *thermal noise* generated by random voltages in the

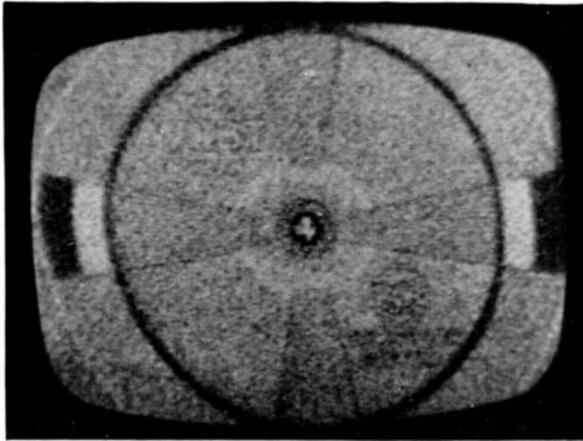


FIG. 20-5. White speckled background of snow, produced by receiver noise voltages, with weak r-f signal. (From McGraw-Hill motion-picture series *Basic Television*.)

circuit components and *shot-effect* noise caused by random fluctuations of plate current in the vacuum tubes. The amplitude of receiver noise produced by the r-f amplifier is in the order of microvolts, but this is comparable with the signal level in the r-f section. The mixer stage generates the most noise voltage. As a result, the signal level at the mixer grid is the limiting factor in the ability of the receiver to reproduce an acceptable picture with weak signal input from the antenna. Enough r-f amplification is needed, therefore, with a low noise level in the r-f amplifier, to supply adequate signal with a high signal-to-noise ratio for the grid of the mixer stage. Suitable r-f amplification can make the difference between a picture with a clean background or a picture that is speckled with snow produced by the random noise voltages generated in the receiver. Figure 20-5 shows snow in the reproduced picture.

*Oscillator Radiation.* The r-f amplifier provides the only isolation between the local oscillator and the antenna, which can radiate the oscil-

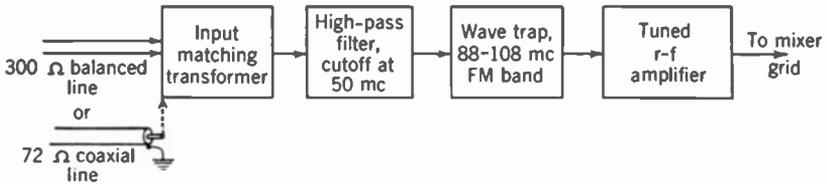
lator signal to produce r-f interference in nearby receivers. Therefore, the use of an r-f amplifier is advantageous in providing a buffer stage between the local oscillator and the antenna. Pentode r-f stages provide more isolation than triodes because of the smaller plate-to-grid capacitance. It should be noted, though, that the local oscillator output can also be radiated from the receiver chassis.

*R-F Selectivity.* Most of the receiver's selectivity to reject adjacent channel frequencies is in the i-f section, but the use of r-f amplification helps in rejecting interfering r-f signals that can produce beat frequencies within the i-f pass band of the receiver. This is especially important in rejecting image frequencies. The first-order image frequency is equal to twice the intermediate frequency plus the desired frequency, when the local oscillator operates above the r-f signal frequencies. Filters and wave traps are generally used in the r-f input circuit to reject interfering r-f signals.

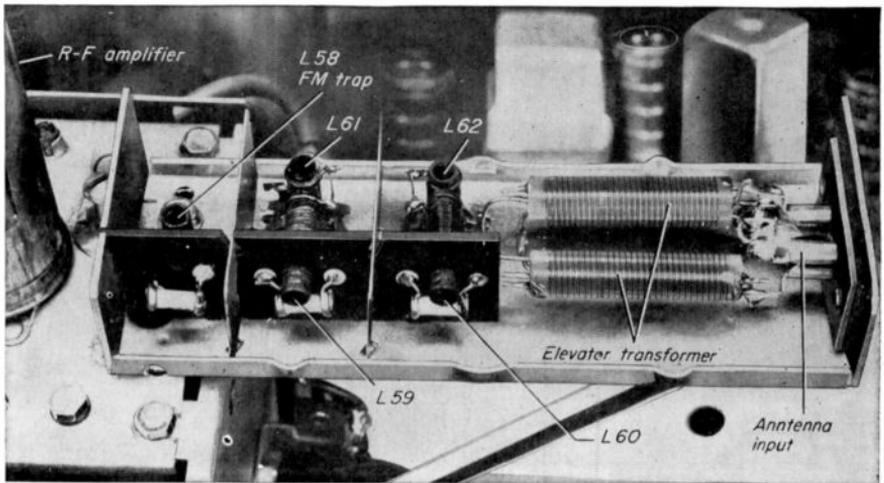
*Input Impedance.* A definite value of input impedance at the receiver's antenna terminals, which is the r-f amplifier input, must be provided so that the impedance of the transmission line from the antenna can be matched to the r-f input circuit. The impedance match prevents signal reflections in the transmission line, which could cause duplicate images or ghosts in the picture when the line is long. This receiver input impedance is usually 300 ohms, with 72 ohms another common value. The input impedance is resistive and relatively constant for all channels, although in most cases it cannot be measured as a d-c resistance.

*R-F Input Circuit.* The block diagram in Fig. 20-6a illustrates the impedance matching and filter circuit requirements of the antenna input circuit to the r-f tuner. The input transformer couples the antenna signal to the grid circuit of the r-f amplifier, while matching the impedance of the grid circuit to the impedance of the transmission line from the antenna. In order to improve the r-f selectivity and reject interfering r-f signals, the r-f input circuit usually includes a high-pass filter and wave trap. The wave trap is a resonant circuit that can be tuned to reject one frequency within the FM broadcast band of 88 to 108 Mc, since this may include image frequencies of channel 2. As an example, in a receiver tuned to channel 2 (54 to 60 Mc), with the local oscillator operating at 81 Mc for the picture i-f carrier of 25.75 Mc, the picture carrier frequency of 55.25 Mc has an image at 106.75 Mc, which is in the 88 to 108-Mc band. If a station broadcasting on this frequency produces interference in the reproduced picture, the wave trap can be adjusted for minimum interference. The high-pass filter is an *LC* band-pass network that attenuates a wide range of frequencies lower than 50 Mc, approximately, cutting off just below channel 2. Interfering frequencies in this range include diathermy equipment and some amateur transmitter bands. The desired signal fre-

quencies for channel 2 or higher are passed by the filter to provide the picture and sound signals for the grid input circuit of the r-f amplifier. The components in a typical r-f amplifier input circuit are shown in Fig. 20-6b. In addition to the filter and impedance-matching requirements, the r-f amplifier grid circuit usually has a double-tuned circuit resonated at each channel to provide maximum r-f signal.



(a)



(b)

FIG. 20-6. (a) Illustrating impedance matching and filter circuit requirements of the antenna input circuit to the r-f tuner. (b) Components in typical antenna input circuit to grid of r-f amplifier. Entire unit is in shielded case mounted to front-end chassis. Elevator transformer at right provides 300-ohm balanced, or 72-ohm unbalanced, impedance for antenna input.  $L_{61}$  and  $L_{62}$  with coupling condenser form high-pass filter 50 per cent down in voltage at 50 Mc. (RCA KRK-11 R-f unit.)

**R-F Amplifier Tubes.** Triodes or pentodes can be used for the r-f stage. The miniature glass pentode r-f amplifier tubes, such as the 6CB6, 6CF6, and 6AG5, allow more gain and their lower plate-to-grid interelectrode capacitance results in less coupling of the local oscillator output in the mixer grid circuit back to the antenna input circuit. Miniature glass triode r-f amplifier tubes, such as the 6BQ7-A, 6BK7, and 6BZ7 twin triodes or 6AN4, 6AF4, and 6J4 triodes, produce less tube noise than pen-

todes but their greater interelectrode capacitance provides less isolation between the local oscillator and antenna. Also, neutralization of the triode r-f amplifier may be necessary to prevent excessive feedback of the r-f signal from plate to grid which can make the r-f amplifier oscillate. The voltage gain of a typical r-f stage is approximately 5 to 10.

**20-3. R-F Amplifier Circuits.** Two basic types of r-f amplifier circuits are shown in Fig. 20-7. The conventional amplifier circuit in *a* where the antenna signal is applied to the control grid, with the cathode grounded, is called a *grounded cathode amplifier*. In *b*, the input signal is applied to the cathode, while the grid of the triode amplifier is grounded. This is called a *grounded grid amplifier* circuit. For both amplifiers, the r-f output signal is taken from the plate circuit.

*Grounded Cathode Amplifier.* Pentodes are generally used in the grounded cathode circuit to provide an r-f amplifier having relatively high

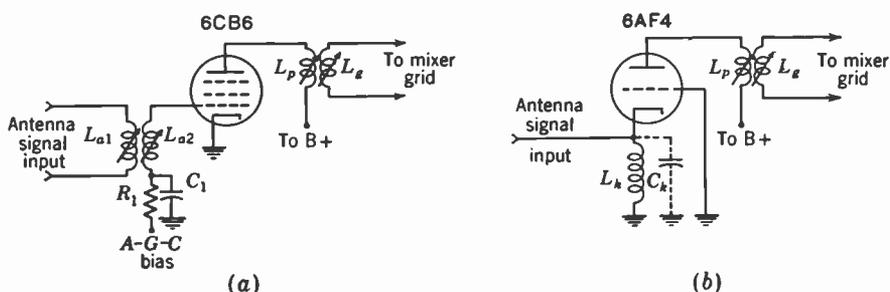


FIG. 20-7. Signal input and output for two basic types of r-f amplifiers. (a) Conventional pentode grounded cathode amplifier. (b) Triode grounded grid amplifier.

gain. With a transconductance ( $g_m$ ) of 5,000  $\mu\text{mhos}$  for the 6CB6, as an example, an impedance ( $Z_L$ ) of 2,000 ohms for the plate load tuned circuit, the voltage gain of the stage, which is  $g_m \times Z_L$ , equals 10. Bias for the amplifier is usually supplied by the a-g-c line. A cathode bias resistor may also be used but the cathode is essentially at ground potential for the r-f signal. The main disadvantage of a grounded cathode pentode r-f amplifier is the relatively high noise level of the stage, compared with a triode amplifier.

*Grounded Grid Amplifier.* Triodes are generally used in a grounded grid circuit for the r-f amplifier. Referring to Fig. 20-7*b*, note that the input signal is applied to the cathode, while the grid is grounded. If it is desired to maintain a d-c potential on the grid, it can be grounded only for r-f signal by means of a bypass condenser. The input signal coupled to the cathode is applied between cathode and grid, since the grounded grid returns to the low-potential side of the cathode load inductance  $L_k$ . The advantage of the grounded grid arrangement is that it shields the out-

put signal circuit from the input signal circuit, preventing feedback through the tube's plate-to-grid capacitance.

An important feature of the grounded grid r-f amplifier is the characteristics of the cathode input circuit. Compared with the grid input impedance of a conventional amplifier, the cathode input impedance of the grounded grid amplifier is low, because the plate signal current also flows through the cathode coil  $L_k$ , in the same direction as the input signal. As a result, the input impedance is approximately equal in value to the reciprocal of the tube's transconductance. For a  $g_m$  of 6,000  $\mu$ hos, therefore, the cathode-to-ground input impedance is approximately 166 ohms. This low impedance is in parallel with the resonant circuit formed by the cathode inductance  $L_k$  and the cathode-to-ground capacitance  $C_k$ . Assuming approximate values of 6  $\mu$ mf for  $C_k$  and 0.25- $\mu$ h inductance for  $L_k$ , as examples, this tuned circuit resonates at 130 Mc. With the low

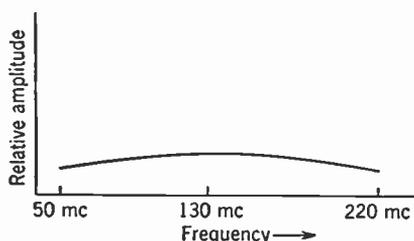


FIG. 20-8. Response of fixed-tuned cathode circuit in grounded grid amplifier.

value of 166 ohms as a shunt damping resistance, the cathode-tuned circuit has a total bandwidth of about 160 Mc, equal to  $1/2\pi RC$ . As shown in Fig. 20-8, this broad response from 50 to 220 Mc includes all 12 v-h-f television channels from 54 to 216 Mc. The grounded grid circuit is useful as a broad-band r-f amplifier, therefore, fixed-tuned for maximum response over the entire v-h-f television band, although the gain is less than in a conventional pentode tuned r-f amplifier.

**Driven Grounded Grid Amplifier.** The arrangement illustrated in Fig. 20-9, combining a grounded-grid stage driven by a grounded cathode stage in a twin-triode tube, is commonly used for the r-f amplifier in television receivers. This is generally called a *cascode* circuit. The advantage of the cascode circuit is that it has the low noise level of a triode r-f amplifier, with approximately the same gain as a pentode r-f amplifier. The cascode circuit in *a* has a-c coupling from the plate of the grounded-cathode section  $V_1$ , through the coupling condenser  $C_c$ , to the cathode of the grounded grid section  $V_2$ . The direct-coupled cascode circuit in *b* operates as an r-f amplifier similar to the circuit in *a*, with  $V_1$  the grounded cathode stage and  $V_2$  the grounded grid stage, but the circuit arrangement is slightly different because of the d-c coupling from the plate of  $V_1$  to the cathode of  $V_2$ .

Referring to the d-c coupled cascode circuit in Fig. 20-9b, two triodes form a d-c voltage divider consisting of the internal plate-to-cathode resistance ( $R_p$ ) of each tube, across the B supply. The plate circuit of  $V_2$

connects to +250 volts, but the plate of  $V_1$  obtains its plate voltage, approximately equal to one-half B+, by the voltage division across the  $R_p$  of the two tubes. Note that the d-c plate current of  $V_1$  flows through  $V_2$  to return to the B supply. Since the plate of  $V_1$  connects to the cathode of  $V_2$  directly through  $L_k$ , this cathode has a d-c potential of +125 volts with respect to chassis ground. Therefore, the  $R_2R_3$  voltage divider is used to supply +125 volts to the control grid of  $V_2$ , so that its grid-to-cathode d-c voltage is approximately zero. Otherwise,  $V_2$  would be cut off by an effective cathode bias of 125 volts, negative at the control grid, if it were connected directly to ground.  $C_0$  is an r-f bypass condenser to return the grid to ground for r-f signal while maintaining the required d-c grid voltage. When signal is received, the resultant a-g-c bias voltage

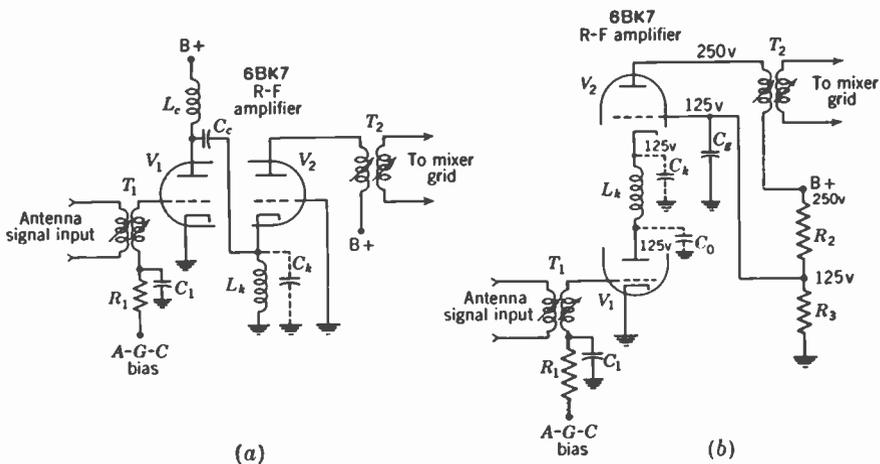


FIG. 20-9. Driven grounded grid, or cascode, r-f amplifier circuits. (a) A-C coupling between the two triode sections. (b) Direct-coupled.

applied to the grid of  $V_1$  reduces its plate current, increasing the plate voltage, which makes the grid-cathode bias voltage for  $V_2$  more negative.

The r-f signal input to the d-c coupled cascode amplifier in Fig. 20-9b is applied by  $T_1$  to the grid of grounded cathode stage  $V_1$ . Since the plate of  $V_1$  is directly coupled by  $L_k$  to the cathode of the grounded grid stage  $V_2$ , the plate load impedance of  $V_1$  is the resonant circuit formed by  $L_k$ ,  $C_k$ , and  $C_0$ , which is the output capacitance from plate to cathode of  $V_1$ . They form a  $\pi$ -type filter coupling circuit, with the resonant rise in voltage across  $C_k$  applied as the input signal between grid and cathode for the grounded grid stage  $V_2$ . The amplified r-f signal output in the plate circuit of  $V_2$  is coupled by  $T_2$  to the mixer grid. R-F tuning for each station is necessary only in the antenna input transformer  $T_1$  and the mixer grid transformer  $T_2$ , as the interstage coupling network provides a resonant response broad enough to include all 12 v-h-f channels.

The use of a triode and the low gain result in a very low noise figure for the grounded cathode input section of the cascode amplifier. However, the grounded grid section amplifies the r-f signal enough to provide a combined gain for the cascode amplifier approximately equal to the gain of a pentode r-f amplifier with the low noise figure of a triode amplifier.

**20-4. The Mixer Stage.** Figure 20-10 illustrates a typical mixer-oscillator circuit that converts the r-f signal frequencies of the selected channel to the intermediate frequencies of the receiver. The 6J6 twin triode is employed, with the triode section  $V_2$  operating as the local oscillator while the triode section  $V_1$  is the mixer stage. The input to the mixer stage includes the desired r-f signal from the r-f amplifier, coupled by the mixer grid transformer  $T_r$ , and the output from the local oscillator coupled by  $C_6$  to the mixer grid. The difference frequency between the r-f picture carrier and the oscillator frequency is the picture carrier intermediate frequency for the picture i-f amplifier, while the difference between the r-f sound carrier and the oscillator is the sound i-f carrier frequency. Note that the same oscillator signal beats with both the picture and sound carrier frequencies. Therefore, the separation between the picture and sound carrier frequencies in the i-f output of the mixer is the same 4.5 Mc as between the two r-f carriers in the standard broadcast channel.  $T_i$  is the mixer plate i-f transformer, coupling the picture and sound i-f carrier signals, with their side-band frequencies, to the grid of the first i-f stage.

It is preferable to use separate oscillator and mixer stages in the television receiver because of the high signal frequencies and wide range of the television band. The multigrid converter tubes are unstable, have too little transconductance, and are noisy. The mixer stage can use a pentode or triode r-f amplifier having high transconductance. Pentodes have less feedback of i-f signal through plate-grid capacitance, but triodes produce less tube noise. Both the r-f signal and the local oscillator output are applied to the control grid of the mixer stage, as shown in Fig. 20-10. This provides more conversion gain and less noise, compared with the method of coupling the oscillator output to a separate injector grid. Interaction between the r-f signal circuits and the oscillator is minimized by using loose coupling from the oscillator to the mixer grid, as illustrated by the 1- to 2- $\mu\mu\text{f}$  capacitance for  $C_6$  in Fig. 20-10. The coupling may be obtained by means of the capacitance between contacts on the station selector switch, a small link can be used between the oscillator and mixer sections, or a small variable condenser may be used. The oscillator injection voltage at the mixer grid is usually about 3 volts, peak value. This produces grid-leak bias of approximately 3 volts, which is a d-c voltage, negative at the mixer grid. The mixer, or first detector, operates as a grid-leak detector to convert the selected r-f signal to the required i-f sig-

nal. The combined oscillator and r-f signal waveform in the mixer grid circuit is rectified, the cathode and grid functioning as a diode detector, and the rectified voltage, which includes the desired difference frequency components for the i-f section, is amplified in the mixer plate circuit.

In frequency-conversion applications for the u-h-f television channels, the mixer stage is often a crystal-diode detector because it can operate at higher frequencies than conventional vacuum tubes. The crystal mixer produces less noise and requires less oscillator injection voltage, compared with a triode mixer, but has no gain. The conversion gain, which is the ratio of i-f signal output to r-f signal input, is about 2 for a triode mixer, operating in the v-h-f range. A crystal mixer has a conversion loss, equal to 0.25 to 0.4, approximately.

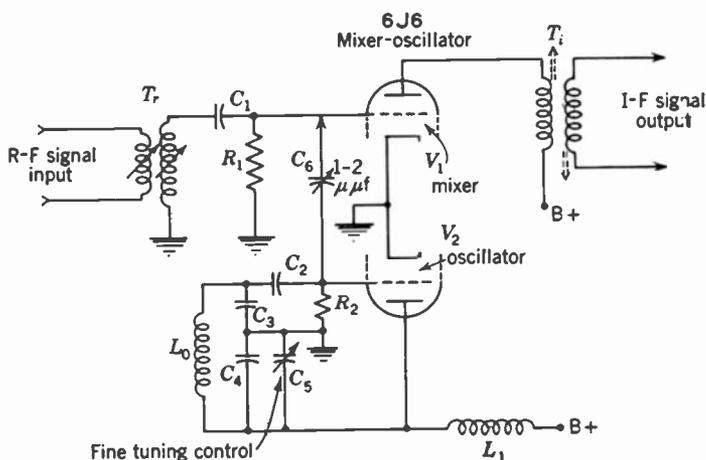


FIG. 20-10. Mixer-oscillator circuit illustrated for a single channel. The r-f transformer  $T_r$  and oscillator coil  $L_0$  are tuned for each selected channel.

**20-5. The Local Oscillator.** The function of the local oscillator in the r-f tuner is to generate an unmodulated r-f sine-wave output at the frequency required to beat with the r-f signal to produce difference frequencies equal to the receiver's intermediate frequencies. For any one station, the oscillator operates at only one frequency. The oscillator may beat either above or below the r-f signal frequencies but usually is above. This has the advantage of reducing the range of frequencies the oscillator must cover. Several typical oscillator circuits are shown in Fig. 20-11. Operation for a single channel alone is illustrated, but the station-selector control changes the resonant frequency of the oscillator's tuned circuit for each channel. Usually, the oscillator has a fine tuning control for exact adjustment of the frequency after the station selector has been tuned to the desired channel. The oscillator generates its own output, without the need for any input signal, by feeding back part of the output

in the plate circuit to the input grid-cathode circuit. Uniform output with good frequency stability through the tuning range is required. The output voltage produced by the local oscillator is about 2 to 5 volts, peak value. Since little power output is required, the local oscillator tube is usually a miniature glass triode such as the 6C4, 6AF4, or one triode section of the 6J6.

*Oscillator Circuits.* The diagram in Fig. 20-11a is the basic Hartley oscillator circuit, which is identified by a tapped inductance  $L_1$  providing, with the variable tuning condenser  $C_1$ , a single-tuned circuit common to both the plate-cathode and grid-cathode circuits. Inductive feedback

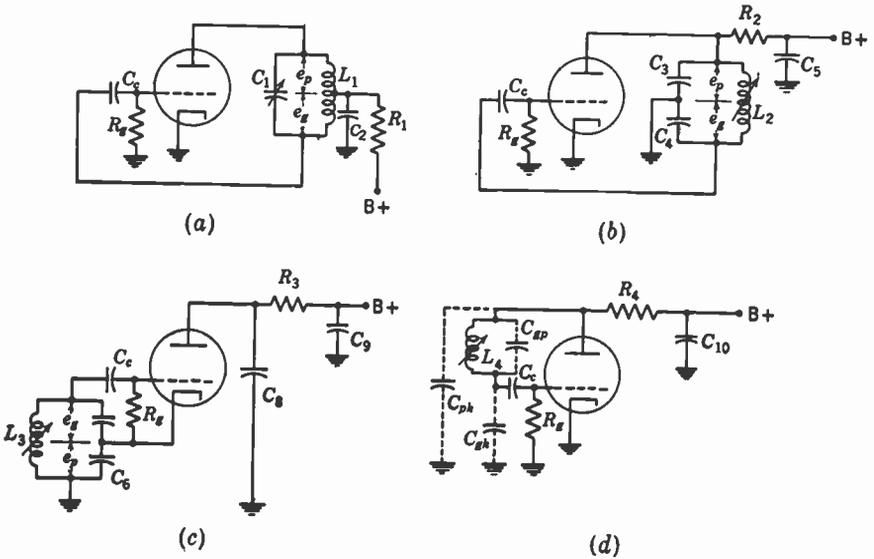


FIG. 20-11. Four typical oscillator circuits. (a) Hartley oscillator with grounded cathode. (b) Colpitts oscillator with grounded cathode. (c) Colpitts oscillator with grounded plate. (d) Ultraudion oscillator.

from plate to grid is supplied by  $L_1$ . The tap where plate-supply voltage is applied is at cathode potential for the oscillator a-c signal because this point is bypassed to the chassis ground and cathode by  $C_2$ . Therefore, the oscillator r-f voltage between the plate side of the tuned circuit and the tap is the plate signal voltage  $e_p$ . The induced grid signal voltage  $e_g$  between the grid side of the resonant circuit and the tap is the feedback voltage for the grid. This voltage drives the grid in the positive direction when the plate current increases, which is the correct phase for sustaining oscillations. When the feedback voltage drives the grid positive, resulting in grid current, grid-leak bias is produced by  $R_g C_g$ . Since the oscillator operates at the frequency of the  $L_1 C_1$  tuned circuit,  $C_1$  can be varied to tune the oscillator to the desired frequency.

The basic Colpitts oscillator circuit in Fig. 20-11*b* also has a single-tuned circuit, consisting of  $L_2$  in parallel with the series combination of  $C_3$  and  $C_4$ . The tuned circuit is common to the plate and grid circuits, as in the Hartley oscillator but capacitive feedback is obtained by means of the  $C_3C_4$  voltage divider instead of a tapped inductance. The voltage across  $C_4$  is the grid feedback voltage, with the amount of feedback determined by the ratio of  $C_4$  to  $C_3$ . Grid-leak bias is produced by  $C_cR_g$ .  $R_2$  is a plate voltage dropping resistor and  $C_5$  an r-f bypass condenser. Since  $R_2$  is in shunt with the plate signal voltage its value is generally about 15,000 ohms. An r-f choke can be used instead of  $R_2$  to isolate the oscillator tuned circuit from B+, without dropping the plate voltage. In the Colpitts oscillator, the frequency is varied by changing the inductance  $L_2$ . This may be a continuously variable inductance or coils can be switched for each channel. In the Colpitts circuit in *c* the cathode is above ground, with oscillator r-f voltage, while the plate is grounded for oscillator signal by  $C_8$ . This circuit has the advantage of allowing one side of the tuned circuit to be connected to chassis ground. The Colpitts circuit is commonly used for the local oscillator in television receivers, because the oscillator tuning inductance can be switched more easily without any tap on the coil, and the interelectrode capacitances of the tube can be made part of the oscillator-tuned circuit. The circuit in *d*, which is called an *ultraudion* oscillator, is equivalent to the Colpitts circuit in *c* but the tube capacitances resonate with the oscillator coil  $L_4$ . The grid-plate capacitance is in parallel with  $L_4$ , while the plate-cathode capacitance and grid-cathode capacitance form a capacitive voltage divider, with the oscillator voltage across  $C_{pk}$  the feedback voltage for the grid.

*Fine Tuning Control.* A typical arrangement is illustrated by the fine tuning control  $C_5$  in the Colpitts oscillator circuit in Fig. 20-10. This is usually an operating control with the station selector on the receiver's front panel. The fine tuning capacitance is across part of the oscillator-tuned circuit, so that varying the control changes the oscillator frequency within the range of about 1.5 Mc, which is small compared with the oscillator frequency. After the station selector has been tuned to the desired channel, the fine tuning control is adjusted for the best sound in split-sound receivers. The oscillator is then at the frequency that produces the correct sound i-f carrier frequency for the sound i-f stages. This automatically results in the best picture also, when the picture and sound i-f amplifiers are correctly aligned with the required 4.5-Mc separation and the receiver has the normal amount of antenna signal. With inter-carrier-sound receivers, however, the fine tuning control is adjusted for the best picture. Since the correct 4.5-Mc sound signal is produced as the difference-frequency beat between the picture and sound carriers, the sound is relatively independent of the exact local oscillator frequency.

It should be noted that continuous tuners generally do not have a separate fine tuning control, as small changes of oscillator frequency can be made with the station-selector control.

*Oscillator Stability.* The local oscillator frequency tends to change with variations in temperature, humidity, tube characteristics, and supply voltage. Since a frequency drift as great as 2 Mc is only 1 per cent change with a frequency of 200 Mc, the local oscillator in the television receiver must be extremely stable. Changes in tube capacitance, caused by warm-up or variations in supply voltage, can be minimized by using a value of external capacitance as large as possible in the oscillator-tuned circuit, in order to make the resonant frequency less dependent on small variations of the tube capacitance. Also, a voltage-regulator tube can be used to stabilize the plate voltage of the oscillator, minimizing changes in the tube's input capacitance. Temperature-compensating capacitors may be used in the oscillator-tuned circuit, to correct for the rapid drift caused by the oscillator tube warming up after the set has been turned on, and the slower change as the temperature of the circuit components increases. All components are rigidly mounted so that movement of the chassis and vibration will not alter their physical position and change the oscillator frequency. Since operation of the loudspeaker can cause vibration of the chassis the oscillator tube often has a heavy lead shield to reduce vibration, or is shock-mounted on a rubber-supported socket. Connecting leads must be short to reduce the stray capacitance. For minimum r-f losses, low-loss insulating materials are used, with ceramic tube sockets.

*Oscillator Radiation.* An important receiver problem is reduction of the signal radiated by the local oscillator, either directly from the chassis or through the antenna circuit, which produces r-f interference in nearby receivers. Figure 20-12 shows the r-f interference bar pattern commonly produced in a receiver by excessive local oscillator radiation from another receiver. When the oscillator interference is very strong it can produce amplitude distortions of the desired picture signal that cause a negative picture or black out the picture completely. With receivers that have the local oscillator operating 25.75 Mc above the r-f picture carrier frequency of the selected station, specific combinations of channels are most likely to result in oscillator interference. For instance, in a receiver with a picture i-f carrier of 25.75 Mc, the local oscillator is at 81 Mc when the station selector is tuned to channel 2. Excessive radiation of the 81-Mc local oscillator output can produce r-f interference in any receiver nearby tuned to channel 5 (76 to 82 Mc), resulting in a 3.75-Mc interfering beat frequency between the channel 5 picture carrier of 77.25 Mc and the 81-Mc r-f interference from the local oscillator. Oscillator interference can be produced also by similar combinations of channels 3 and 6, chan-

nels 7 and 11, channels 8 and 12, and channels 9 and 13. It should be noted that the receiver producing the excessive oscillator radiation does not have the interference but can interfere with another receiver nearby when the oscillator frequency is within the range of a television channel. Receivers with a picture i-f carrier of 45.75 Mc, however, produce oscillator frequencies that are not in any of the 12 v-h-f channels from channels 2 to 13.

Excessive local oscillator radiation can cause r-f interference in receivers several hundred feet away.<sup>1</sup> Oscillator radiation is minimized by operating the stage with low power output, having an r-f amplifier to act as a buffer stage between the antenna circuit and mixer stage, using higher intermediate frequencies, and by adequate shielding. The r-f tuner is

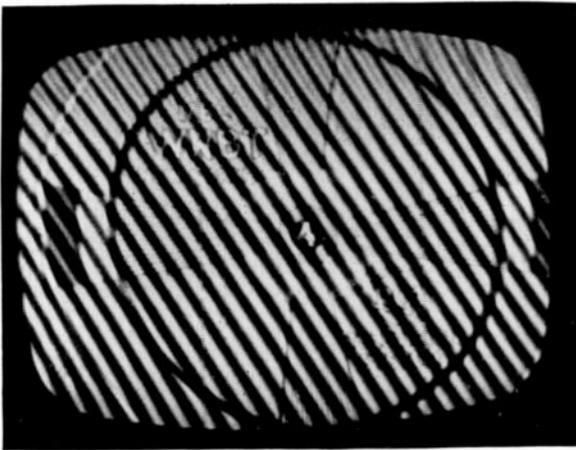


FIG. 20-12. R-F interference pattern produced by local oscillator radiation from a nearby receiver. (From McGraw-Hill motion-picture series, *Basic Television*.)

usually shielded but most of the oscillator radiation is from the receiver main chassis. To reduce oscillator radiation from the main chassis, in many receivers the mixer output is coupled to the i-f amplifier by means of a short length of coaxial line, instead of using common chassis ground-return connections.

**20-6. High-frequency Techniques.** Frequencies of about 100 Mc or more are high enough to require special attention to many factors usually ignored at lower radio frequencies. Since the r-f tuner operates with v-h-f and u-h-f signal frequencies, different techniques are often necessary for satisfactory operation of the r-f stages. The main problems at these high frequencies are obtaining sufficient amplification by the tube, pro-

<sup>1</sup> The RETMA rating for television receivers specifies that the field strength of the radiated local oscillator signal 1,000 ft away shall not exceed 25  $\mu$ v per meter with an antenna height of 30 ft.

viding values of  $Q$  for the tuned circuits high enough for adequate gain, and arranging low-impedance bypass circuits, as insufficient bypassing can result in undesired regeneration and oscillations.

*Tube Limitations.* At high frequencies, the interelectrode capacitances and the inductances of the connecting leads, from the pin connections to the electrodes in the tube, must be considered as part of the circuit. The interelectrode capacitances become more dominant in circuit operation as the external capacitance of the associated tuned circuit is decreased for higher frequencies. Also, the inductances of the connecting leads, which are negligible at lower frequencies, may have appreciable inductive reactance at high frequencies. As an example, a wire of 0.04 in. diameter and 4 in. long has an inductance of approximately  $0.1 \mu\text{h}$ . At a frequency of 1 Mc this inductance has a reactance of 0.63 ohm. However, at 100 Mc the reactance is 63 ohms, and at 500 Mc the reactance 315 ohms, which can make a connecting lead effective as an r-f choke. The inductance of the cathode lead, including the internal connection from tube pin to cathode, can have enough reactance at high signal frequencies to act as a mutual impedance for the plate and grid return circuits, resulting in degenerative feedback. For this reason, some r-f amplifier tubes have two pin connections for the cathode, one for the grid input circuit return to cathode and the other for the plate output circuit return. In tubes designed for the grounded grid amplifier circuit, the control grid may have multiple pin connections.

When a tube is used in an oscillator circuit, in order to increase the oscillating frequency the inductance and capacitance of the associated tuned circuit must be decreased. In the extreme case, with the external capacitance reduced to zero, and the inductance only a straight conductor connecting the plate and grid terminals, the upper frequency limit of the tube is reached at the *tube's resonant frequency*. At this frequency, the tube itself would oscillate as an equivalent ultraudion circuit. For high-frequency operation, therefore, the inductance of the leads within the tube and the interelectrode capacitances are decreased by reducing the physical size of the tube and the internal electrodes. The leads in high-frequency tubes are brought straight out through the glass envelope to the pin connections, without a base.

*Transit-time Effect.* In low-frequency operation of a vacuum tube it may be assumed that electrons emitted from the cathode reach the plate instantaneously, since a typical transit time is  $0.001 \mu\text{sec}$ . However, the same transit time is long enough to be one-tenth of a complete cycle when the operating frequency is 100 Mc, and the transit time becomes a proportionately greater part of the cycle for higher operating frequencies. The efficiency of the tube decreases appreciably for transit times longer than  $\frac{1}{10}$  cycle.

*Grid Input Resistance.* An important effect of the relatively long transit time at high frequencies is the reduced value of the tube's internal input resistance between grid and cathode. As an example, the 6SK7 r-f pentode for lower frequency applications has a grid input resistance of approximately 120,000 ohms at 10 Mc but only 50 ohms at 150 Mc. The 6AK5 pentode r-f amplifier for higher frequencies, which is a miniature glass tube, has a grid input resistance of approximately 300,000 ohms at 10 Mc and 2,000 ohms at 150 Mc. Since the input resistance of the tube is in parallel with the external grid-cathode circuit, the impedance of the grid circuit decreases at high frequencies. With low values of grid input impedance loading down the output circuit of the previous stage, r-f amplifier tubes must have high transconductance for adequate gain in high-frequency operation.

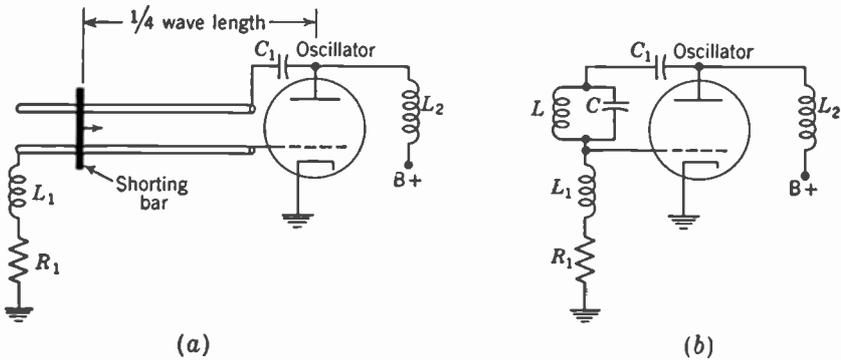


FIG. 20-13. Ultraudion oscillator circuit using a quarter-wave line as the tuned circuit. (a) Schematic diagram. (b) Equivalent tuned circuit.

*Line Sections as Resonant Circuits.* The tuned circuits in high-frequency applications are often resonant lines, instead using the lumped capacitance and inductance of condensers and coils. There is always an irreducible value of stray capacitance in the circuit, and as the operating frequency increases the inductance required to resonate with the minimum capacitance at the desired frequency is either impossible to obtain or too small to provide a value of  $Q$  high enough for the resonant circuit. With resonant lines used as tuned circuits, high  $Q$  and high impedance can be obtained at high frequencies. The physical length of the lines is feasible because the line is shorter for higher frequencies. The shorted end is a point of low impedance for the signal frequency, while the open end to which the plate and grid are connected has high impedance. Figure 20-13 shows a u-h-f oscillator using a quarter-wave transmission line as the tuned circuit between grid and plate, in an ultraudion oscillator circuit. The operating frequency of the oscillator is that frequency for which the length of

the line is an electrical quarter wave. For 500 Mc, as an example, a quarter wavelength is approximately 6 in. The shorted end is a point of low impedance, while the open end 6 in. away, where the plate and grid are connected, has high impedance for 500-Mc signal. By adjusting the physical position of the shorting bar, the resonant length of the line is changed, making it one-quarter wavelength for a different frequency, to vary the oscillator frequency. The lines are usually silver plated to reduce resistance and noise produced by the sliding contact.

In Fig. 20-14, a quarter-wave resonant line section is illustrated as the input tuned circuit for the grid of an r-f amplifier stage. The antenna input is connected to a comparatively low impedance point on the line to provide the desired input impedance, while the grid is at the point of high impedance and maximum signal voltage a quarter wavelength from the

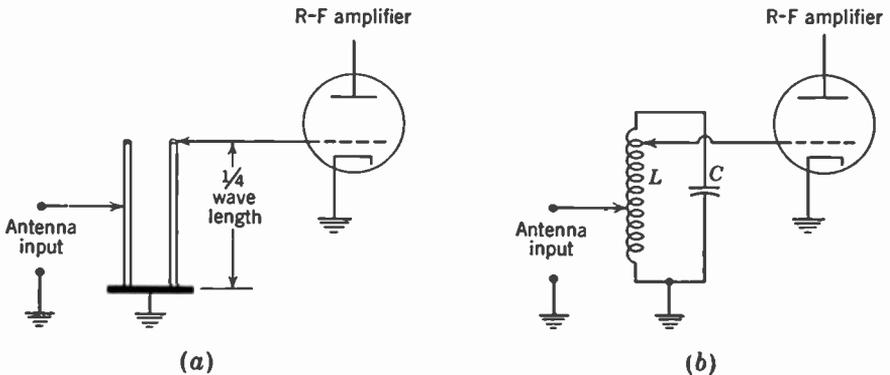


FIG. 20-14. R-F amplifier input circuit using a quarter-wave line as the tuned circuit. (a) Schematic diagram. (b) Equivalent tuned circuit.

shorted end. In this way, the line functions as a resonant circuit tuned to the signal frequency, with the tapped connection providing an impedance match between the grid input circuit and the lower impedance of the transmission line from the antenna.

**Bypassing Circuits.** Since a short length of wire can have appreciable impedance at v-h-f and u-h-f signal frequencies, all connecting leads of bypass condensers must be as short as possible. The return leads should be connected to a single ground close to the tube socket. This prevents common coupling between stages through circulating currents in the chassis ground return circuits. The bypass condensers should not be too large because they can have appreciable inductance.

**20-7. R-F Alignment.** The r-f alignment consists of setting the local oscillator to the correct frequency for tuning in each channel, and adjusting the r-f signal circuits for maximum r-f gain with the required bandwidth. Usually, the r-f alignment is done with the picture and sound

i-f stages correctly aligned, since the i-f section of the receiver can be used to help indicate proper alignment of the r-f oscillator.

*Oscillator Alignment.* R-F tuners that switch inductances for each selected channel have individual oscillator coils with a slug that can be adjusted to set the oscillator frequency for each station. These are provided in addition to the fine tuning control, as installation or servicing adjustments. The oscillator coil slugs can usually be adjusted through the front panel of the receiver, without removing the chassis. As illustrated in Fig. 20-15, taking off the front panel's channel-number plate, called the *escutcheon*, makes the oscillator adjustments at the front of the tuner available. The oscillator coils are adjusted for the correct oscillator frequency on each selected channel, within the middle range of the fine

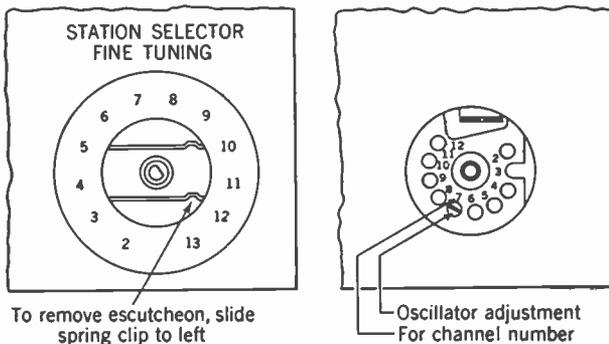


FIG. 20-15. R-F oscillator adjustments. (a) Removing escutcheon plate. (b) Location of coil slugs. Channel 13 adjustment on top of tuner. (RCA KRK-11 R-f unit.)

tuning control. Readjustment of the oscillator frequency is often necessary when the oscillator tube is replaced, because of the slightly different interelectrode capacitances. With the fine tuning control set at the middle of its range, the oscillator coils can be adjusted for each channel by one of the following methods.

1. *Using the transmitted r-f sound and picture carrier signals.* When the desired stations are broadcasting, the oscillator frequency can be set by adjusting for the best sound and picture. In split-sound receivers, adjust the oscillator for the best sound, with maximum volume and minimum background noise, which indicates balance in the FM sound detector. Since the bandwidth of the sound i-f stage is very narrow compared with the oscillator frequency, an accurate adjustment can be made. With the sound and picture i-f sections correctly aligned, this also results in the proper response for the i-f picture signal. In intercarrier-sound receivers, however, the best picture is used to indicate the correct frequency of the oscillator, since the sound output is independent of the exact oscillator frequency. The best picture is indicated by good contrast with maxi-

imum detail. In either case, it is good practice to check both the sound and picture for each channel while adjusting each oscillator coil.

2. *Injecting the r-f sound and picture carrier frequencies.* When transmitted signals are not available in the channel to be aligned, the oscillator can be set to the correct frequency by using the output from an accurate crystal-calibrated signal generator to supply the r-f picture and sound carrier frequencies. In split-sound receivers, the signal generator supplies the r-f sound carrier frequency and the oscillator is adjusted for zero output from the balanced FM sound detector at center frequency. The r-f sound carrier frequency is always 0.25 Mc below the high end of the channel. With the signal generator set to the r-f sound carrier frequency of 215.75 Mc for channel 13, as an example, it is connected to the receiver input terminals. A vacuum-tube voltmeter is connected to the audio output terminals of the FM sound detector. Then the oscillator coil can be adjusted for zero output on the meter, as the correct i-f sound carrier is produced for balance at center frequency. The correct indication is a sharp drop to zero, between positive and negative peaks, not a broad zero response caused by being entirely off frequency. In intercarrier-sound receivers, the r-f signal generator can be set at the r-f picture carrier frequency and used to mark the over-all r-f and i-f visual response curve. The r-f picture carrier frequency is always 1.25 Mc above the low end of the channel. With the r-f signal generator at 211.25 Mc for channel 13, as an example, and the over-all response curve on the oscilloscope screen for this channel, the oscillator frequency is adjusted to place the picture carrier marker at the point of 50 to 60 per cent response on the sloping side of the curve.

3. *Measuring the oscillator frequency.* First, the correct oscillator frequency must be known for each channel. This always differs from the r-f carrier frequency by the i-f carrier frequency, usually above. In a receiver which has a picture i-f carrier frequency of 45.75 Mc, as an example, tuned to channel 2 with the r-f picture carrier frequency of 55.25 Mc, the oscillator frequency equals  $55.25 + 45.75$ , or 101 Mc. Then, the oscillator adjustment for the channel is varied to produce the required frequency, as indicated by an accurate frequency-measuring instrument. The frequency of the oscillator output can be measured by (1) using an accurate frequency meter that has crystal calibration points; (2) heterodyning the oscillator output against an accurately calibrated signal generator for zero beat by means of a communications receiver that can tune to the range of oscillator frequencies; or (3) using an accurate wavemeter.

Regardless of the method used to check the oscillator frequency, each oscillator coil is adjusted individually as the station selector is set to the channel to be aligned. In tuners where the oscillator frequency is decreased for the lower channels by adding series inductances, the higher

channels must be adjusted first. It should be noted that continuous tuners usually do not have individual oscillator coil adjustments for each channel, but there are oscillator trimmer adjustments for tracking of the oscillator frequency with the r-f signal circuits from the low end to the high end of the tuning range.

*R-F Response Curve.* The r-f circuits tuned to the signal frequencies of each selected channel, including the antenna input and mixer grid transformers, are usually aligned by the visual response curve method. The procedure is similar to the method used for obtaining the i-f response curve. However, the sweep generator is connected to the antenna input terminals of the receiver and set to sweep the band of r-f signal frequencies in the channel to which the station selector is set, while the vertical input

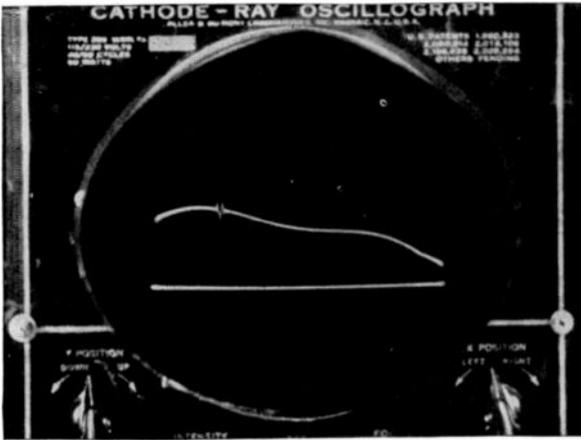


FIG. 20-15. R-F response curve for channel 12, with marker at picture r-f carrier frequency of 205.25 Mc. (From McGraw-Hill motion-picture series, *Basic Television*.)

terminals of the oscilloscope are connected to the mixer grid circuit to indicate the relative amplitude of the r-f signal frequencies. A test point is usually provided on the tuner for this oscilloscope connection, through a decoupling resistor of about 50,000 ohms to the mixer grid. The grid-leak detector action here results in rectified r-f signal for the oscilloscope. Compared with the r-f response, the r-f gain is much lower, requiring more r-f output from the sweep generator and increased gain in the oscilloscope to obtain the r-f response curve, and the width of the bottom part of the curve may be more than 10 Mc, making more sweep width necessary to see the entire r-f curve. A typical r-f response curve is shown in Fig. 20-16, with the marker at the picture r-f carrier frequency. The marker, set to the r-f signal frequencies in the selected channel, is used to indicate the position of both the picture and sound r-f carrier frequencies on the r-f response curve. They should have a relative ampli-

couples the r-f signal to the grid of the grounded cathode section  $V_{1A}$  in the 6BQ7 cascode r-f amplifier.  $R_8$  is a shunt damping resistor to provide the required r-f bandwidth.  $R_9$  is a decoupling resistor in the a-g-c bias line to the r-f amplifier, with the r-f bypass  $C_{14}$ , which is a low-inductance feed-through type of condenser. The variable condenser  $C_{20}$  is the trimmer adjustment for aligning the r-f amplifier grid circuit. The plate of the grounded cathode section is directly coupled by  $L_7$  to the cathode of the grounded grid section  $V_{1B}$ , which has the grid grounded by  $C_1$  for r-f signal frequencies.  $R_1$  forms a voltage divider with  $R_2$  to provide the d-c grid voltage required, with  $R_3$  isolating the grid from the B+ line. The inductance of  $L_7$  with the input and output capacitance results in a  $\pi$ -type filter circuit that provides the plate load impedance for  $V_{1A}$  and couples the r-f output to  $V_{1B}$  as the signal voltage developed across the cathode-to-ground capacitance of the grounded grid section. The response of this circuit is broad enough to include all 12 v-h-f channels, so that it need not be tuned for individual channels. Part of the r-f output in the plate circuit of  $V_{1A}$  is fed back to the grid circuit by  $C_9$ , in order to increase the gain for the higher frequency channels. This condenser forms an a-c voltage divider with  $C_{20}$ , the feedback voltage developed across  $C_{20}$  being applied to the grid in series with the signal voltage across  $L_2$ .  $C_9$  is often called a *neutralizing condenser*, because the feedback voltage cancels the plate signal voltage coupled back to the grid through the plate-grid capacitance in the tube.  $L_{11}$  and  $C_{19}$  in the feedback circuit form a series resonant wave trap that can be adjusted to reject r-f interference.

In the input circuit of  $V_{1B}$ , the r-f voltage produced across the cathode-ground capacitance is the input signal, since the grid is grounded for the r-f signal frequencies. This input signal is amplified in the grounded grid section, with the r-f transformer  $L_3L_4$  providing the plate load impedance. The signal voltage across the secondary  $L_4$  is applied to the mixer grid.  $C_3$  is the trimmer adjustment for aligning the r-f amplifier plate circuit, while  $C_6$  aligns the mixer grid circuit. The coupling condenser  $C_5$  is used to provide grid-leak bias for the mixer grid circuit, with the grid resistors  $R_4$  and  $R_6$ . The junction of these two resistors is the r-f test point, where a meter or oscilloscope can be connected to indicate d-c voltage across  $R_6$  without detuning the grid circuit appreciably because of the isolation by  $R_4$ . The feedback network of  $C_{18}$  in series with  $L_9$  from plate to grid of the mixer prevents regeneration at the intermediate frequencies.

The lower triode section of the 6J6 in Fig. 20-18 is an ultrasonic oscillator circuit. The capacitive voltage divider across the oscillator coil consists of the tube's plate-cathode capacitance in series with the grid-cathode capacitance, which is shunted by  $C_8$ . Since the divider is grounded at the cathode, the voltage across  $C_8$  is the grid feedback volt-

age. The oscillator coil is switched to select the desired channel. To vary the oscillator frequency slightly on any channel, the fine tuning control varies in capacitance as a dielectric disk, which can be seen in Fig. 20-2, is rotated to change its position between the grounded frame and a stator disk insulated from the frame. This produces about a 3 Mc change in oscillator frequency on the high-band v-h-f channels and about a 0.75 Mc change on the low-band channels. Grid-leak bias for the oscillator is produced by  $C_{10}$  and  $R_7$ . The oscillator plate load resistor  $R_{10}$  maintains the r-f voltage on the plate, while dropping the d-c voltage from the B+ line. Plate voltage for the mixer stage is taken from the same line, with  $R_{11}$  serving as a plate dropping resistor and  $C_{22}$  an i-f bypass condenser. Since the inductive coupling between  $L_5$  and  $L_4$

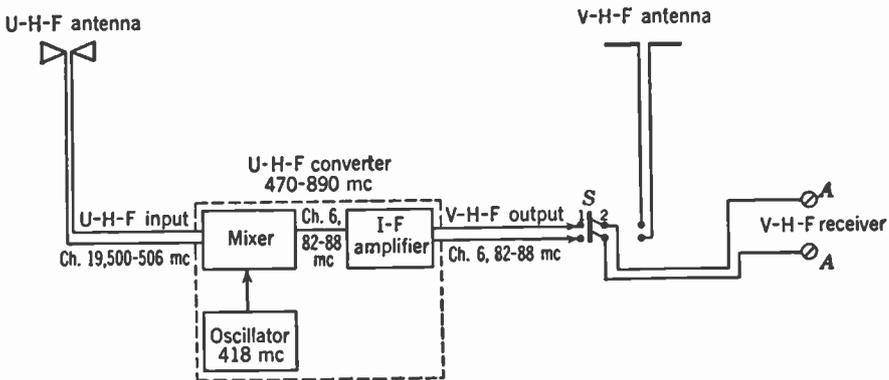


FIG. 20-19. Operation of u-h-f converter, illustrated for u-h-f channel 19 converted to v-h-f channel 6 frequencies.

injects the oscillator output into the mixer, it beats with the r-f signal frequencies of the selected channel to produce the i-f output in the mixer plate circuit.  $L_{10}$  is an i-f peaking coil, while  $T_1$  is the first i-f transformer, with link coupling to the input of the i-f amplifier section.

For the r-f alignment of this v-h-f turret tuner, the oscillator frequency is set for each channel by adjusting the slugs for each individual oscillator coil. The r-f response curve is obtained by connecting the r-f sweep generator to the antenna input, and the scope to the r-f test point in the converter grid circuit. The r-f signal circuits are aligned on channel 10, by adjusting the trimmer condensers in the grid and plate circuits of the r-f amplifier and in the mixer grid circuit for the desired r-f response curve.

**U-H-F Converters.** Television receivers with r-f tuners designed only for v-h-f channels can be adapted to receive u-h-f channels by adding a frequency converter. This is an external u-h-f tuning unit connected between the antenna and the input to the v-h-f receiver, as illustrated in Fig. 20-19. Since the u-h-f channels have the standard characteristics for

the transmitted picture and sound carrier signals, they can be received simply by converting the desired u-h-f channel frequencies to the lower frequencies of a v-h-f channel. Generally, the converter changes the selected u-h-f channel frequencies to either the channel 5 (76 to 82 Mc) or channel 6 (82 to 88 Mc) band of frequencies. The converter consists of r-f preselection circuits for the u-h-f channels, a u-h-f oscillator, u-h-f mixer, and an amplifier for the v-h-f converted output of the u-h-f mixer. The mixer is often a crystal diode. The amplifier for the u-h-f mixer output, which is called an i-f stage, is usually a cascode amplifier. The oscillator beats below the u-h-f signal frequencies so that there will be no inversion of the side-band frequencies in the mixer output.

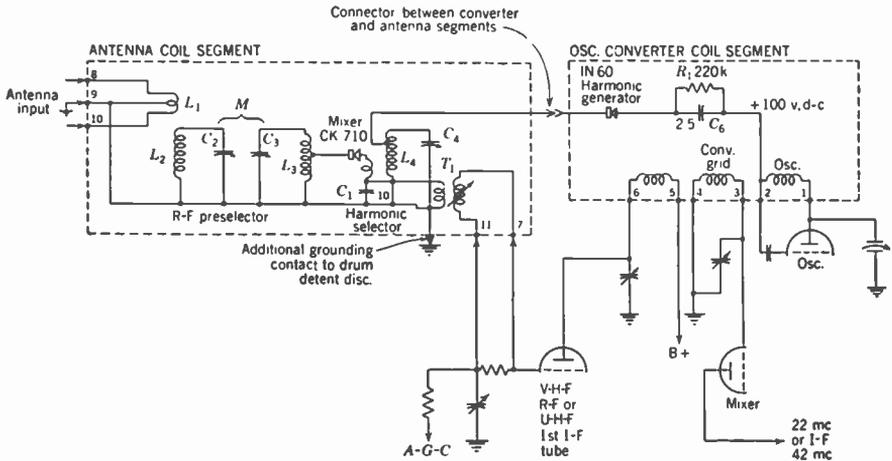


FIG. 20-20. Circuit arrangement of u-h-f conversion strips for v-h-f turret tuner in Fig. 20-18. (Standard Coil Products, Inc.)

**U-H-F Conversion Strips.** V-H-F turret tuners can be converted to receive u-h-f channels by inserting a pair of u-h-f channel strips in the coil drum, in place of a pair of unused v-h-f channel strips. Figure 20-20 shows the schematic diagram of a set of u-h-f strips. These employ two steps of frequency conversion in heterodyning the u-h-f channel frequencies down to the intermediate frequencies of the v-h-f receiver. To tune in channel 19 at 500 to 506 Mc, as an example, with a receiver having the 41- to 47-Mc i-f pass band, the u-h-f oscillator coil strip tunes the local oscillator to the fundamental frequency of 153 Mc. The local oscillator output is coupled to both the v-h-f mixer stage and the u-h-f harmonic generator, which is a crystal diode. Rectification of the oscillator output produces strong harmonics, and the harmonic selector tuned circuit on the antenna coil strip resonates at the second harmonic frequency of the oscillator at 306 Mc. This beats with the channel 19 signal frequencies of 500 to 506 Mc in the u-h-f crystal mixer, converting the signal

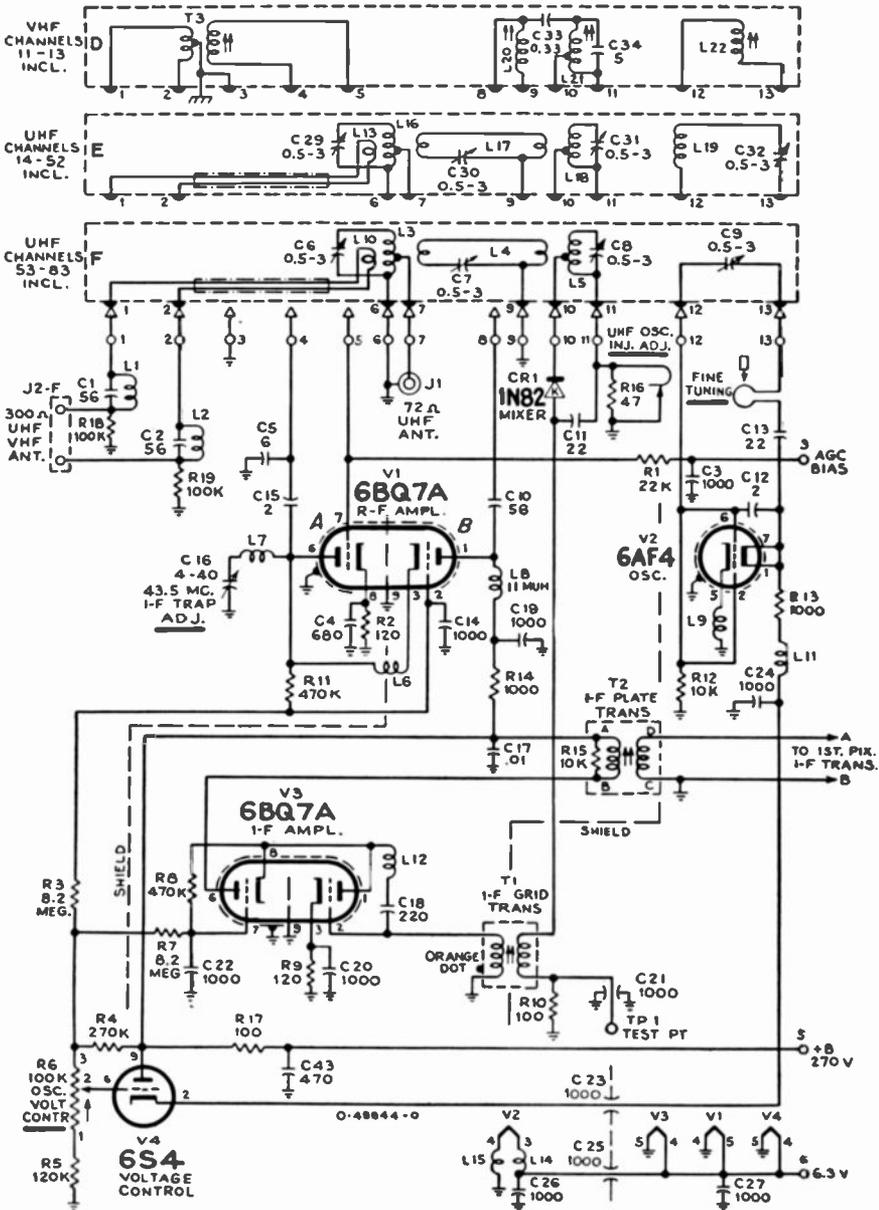


FIG. 20-21. Schematic diagram of r-f tuner for u-h-f and v-h-f channels. (RCA KKK-12 r-f unit.)

to the lower difference frequencies of 194 to 200 Mc. The u-h-f mixer output transformer  $T_1$  is tuned to 194 to 200 Mc and couples these v-h-f signal frequencies to the r-f amplifier stage, which is the same tube used for v-h-f operation. The amplified v-h-f signal at 194 to 200 Mc is coupled by the converter grid transformer to the grid circuit of the v-h-f mixer. Here the fundamental oscillator frequency of 153 Mc beats above the v-h-f signal frequencies of 194 to 200 Mc to produce the difference frequencies of 41 to 47 Mc. This is the same i-f output for the i-f section of the v-h-f receiver as in v-h-f operation, produced by double conversion of the u-h-f channel frequencies.

*U-H-F and V-H-F Tuner.* Figure 20-21 shows the schematic diagram of a turret tuner for any combination of 16 u-h-f and v-h-f channels, with six different types of tuning strips. The stages in the tuner include the cascode r-f amplifier  $V_1$  for v-h-f channels only, with the 1N82 crystal-diode mixer, 6AF4 local oscillator, and cascode i-f amplifier  $V_2$  used for both v-h-f and u-h-f operation.  $V_4$  is a voltage-regulator tube to stabilize the plate voltage of the local oscillator for improved frequency stability.

Referring to the v-h-f channel strip  $D$  in Fig. 20-21, terminals 1 and 2 connect to the antenna input, providing 300 ohms balanced input impedance. The r-f signal across the secondary of the antenna input transformer  $T_3$  is connected to the r-f amplifier stage through terminals 4 and 5 on the strip. The amplified output from  $V_1$  is connected to the crystal mixer through the strip terminals 8 and 10. The coil  $L_{25}$  between terminals 12 and 13 is the inductance for the ultraudion oscillator circuit. The oscillator output is injected into the crystal mixer by a pickup loop in the oscillator compartment, which is adjustable to vary the amount of injection voltage. With the oscillator beating above the r-f signal frequencies, the i-f output from the mixer is the i-f signal of 41 to 47 Mc. This is coupled to the cascode i-f amplifier  $V_3$  by the i-f grid transformer  $T_1$ . The i-f plate transformer  $T_2$  couples the 41- to 47-Mc i-f signal to the first i-f stage on the main chassis of the receiver.

In the u-h-f strips  $E$  and  $F$  in Fig. 20-21, resonant line sections are used for tuning to the u-h-f signal frequencies. Referring to strip  $F$ , either 300-ohm balanced antenna input impedance can be used across terminals 1 and 2, or 72-ohm coaxial cable can be connected to terminals 6 and 7. A triple-tuned circuit, with  $L_3C_6$ ,  $L_4C_7$ , and  $L_5C_8$ , couples the u-h-f signal to the crystal mixer, through terminal 10. The r-f stage is not used for the u-h-f channels because its low gain at these high signal frequencies would reduce the signal-to-noise ratio. In the oscillator circuit, the leads have enough inductance to provide the high resonant frequencies required. The fundamental oscillator frequency is high enough to beat above the u-h-f channel frequencies by 41 to 47 Mc, converting the u-h-f signal directly to the intermediate frequencies of the receiver. As in v-h-f

operation, the i-f output from the mixer is coupled by  $T_1$  to the cascode i-f amplifier  $V_3$  and its output is coupled by  $T_2$  to the first i-f stage of the receiver's i-f section. The cascode i-f amplifier on the tuner chassis is necessary because its low noise figure provides a good signal-to-noise ratio for u-h-f operation, when there is no r-f amplifier stage, although it is also used for operation on the v-h-f channels.

**20-9. Receiver Noise and Snow in the Picture.** The noise resulting from random voltages generated by the vacuum tubes and circuit components in the receiver is called *receiver noise*. The receiver noise results in video noise voltages at the kinescope grid, producing the white speckles called *snow* throughout the reproduced picture, as shown in Fig. 20-5, analogous to the continuous frying or hissing sound of receiver noise in an audio system. With a strong r-f signal providing a high value of signal-to-noise ratio, the snow in the picture is not so evident, but when the r-f signal is weak the receiver noise amplitude can be comparable with the signal voltage to produce a weak picture obscured by snow. Primarily, the amount of noise generated in the low-level r-f circuits of the front end limits the signal-to-noise ratio obtainable for the receiver, which determines how snowy the picture appears with weak r-f signal. In the front end, the r-f signal has low amplitude, and any noise introduced here is amplified just as much as the desired signal. The noise generated by the low-level r-f stages is especially important in television receivers because the noise power increases in proportion to the bandwidth of the receiver circuits, since the noise is generated by random motion of electrons and has no specific frequency. Noise voltages are generated in the i-f and video stages also but the level of the noise introduced in these stages is usually insignificant compared with the signal level.

*Thermal Agitation Noise.* The heat in a conducting material agitates the molecules of the conductor, producing a random electron motion equivalent to a small noise current flowing through the conductor. The higher the bandwidth, temperature, and resistance the greater is the generated thermal noise voltage. For a 6-Mc bandwidth, and at room temperature, a resistance of 300 ohms generates a thermal noise voltage of approximately  $5 \mu\text{v}$ . It is especially important to have little thermal agitation noise in the antenna input circuit, since the signal amplitude has its lowest level here and both the noise and the signal are amplified by the first r-f stage.

*Shot-effect Noise.* This is caused by random fluctuations in the electron flow from cathode to plate, producing noise voltages across the plate load impedance. The shot-effect noise voltage in the plate circuit increases with greater values of load impedance, bandwidth, and emission current. In addition, the more electrodes for collecting electrons in the tube, the greater is the shot effect because the increased partition of the emission

current results in a more random effect in the plate current. Triodes are less noisy than pentodes, therefore. A mixer stage is more noisy than the same tube used as an r-f amplifier because of the lower transconductance. Pentagrid converter tubes are the noisiest of all because of the low transconductance and the multigrid structure. Crystal-diode mixers produce less noise than vacuum tubes but have a conversion loss. However, when the signal frequencies are so high that no conversion gain can be obtained with a vacuum-tube mixer, the crystal mixer can provide a better signal-to-noise ratio. For a vacuum tube conducting 5-ma plate current, the shot-effect noise voltage generated across a 1,500-ohm plate load impedance with a bandwidth of 6 Mc is about 140  $\mu\text{v}$ . Shot-effect noise dominates the thermal noise and therefore must be minimized in the r-f amplifier and mixer stages to provide a good signal-to-noise ratio for the output signal to the i-f amplifier section. The first r-f stage should be a low-noise amplifier that supplies enough r-f signal to overcome the relatively high noise of the mixer stage. When the r-f signal is coupled directly to a crystal-diode mixer, without an r-f stage, as is often done in u-h-f signal circuits, the first i-f stage for the converted u-h-f signal must be a low-noise amplifier.

*Minimum R-F Signal Level.* To obtain a satisfactory noise-free picture, the signal-to-noise ratio should be about 30:1, peak r-f signal voltage to r-m-s noise voltage. The r-f amplitude required at the antenna input is several hundred microvolts or more, therefore, to obtain a noise-free picture. It should be noted, however, that the receiver noise does not include external noise voltages, such as ignition noise. Since the streaks produced by external noise voltages may be more noticeable in the picture than the snow caused by receiver noise, careful attention to the antenna system is necessary to minimize pickup of external noise. With typical external noise in city areas, an antenna input signal of 5 to 50 mv is needed for a good picture. In areas of low signal strength and little external interference, though, where the receiver sensitivity is limited by snow in the picture resulting from receiver noise, a noise-free picture can be obtained with an antenna signal of about 200 to 600  $\mu\text{v}$ , depending on the amount of noise introduced in the r-f section.

**20-10. Troubles in the R-F Tuner.** Since the r-f section of the receiver operates on both the picture and sound signals, defects in the front end cause troubles in the picture and sound. Two symptoms that usually indicate trouble in the front end are (1) the trouble appears on one or more channels but operation is normal for other channels; (2) excessive snow appears in the picture. When a trouble is evident on some channels, but not others, this indicates a defect in the r-f section, including the antenna and transmission line, because here the r-f signal frequencies and the local oscillator output are different for different channels. After the

r-f signal frequencies have been converted to the intermediate frequencies of the receiver, the signal frequencies are the same for all channels. It should be noted, however, that a defect in the r-f section can also result in trouble on all channels. Excessive snow indicates the fault is in the r-f section because the signal-to-noise ratio is determined primarily by the receiver noise generated in the front end.

*Local Oscillator Does Not Operate.* If the local oscillator does not produce the injection voltage required in the mixer to beat with the r-f signal to produce the i-f output, there will be no picture and no sound. Insufficient oscillator injection voltage can cause weak picture and sound. The oscillator may operate normally on some channels, but not on others, particularly the higher frequency channels. If the local oscillator does not operate at all, there will be no picture and no sound on all channels. Operation of the local oscillator can be checked by measuring its negative grid-leak bias with a d-c voltmeter. This should be about  $-3$  volts. Similarly, the oscillator injection voltage can be checked by measuring the negative grid-leak bias voltage in the mixer grid circuit, which is usually about  $-3$  volts also.

*Incorrect Local-oscillator Frequency.* This usually causes trouble symptoms on individual channels, while other channels are normal. Incorrect oscillator frequency can result in (1) picture but no sound, (2) no picture and no sound, or (3) picture and sound of a channel present at the wrong setting of the station selector. These trouble symptoms are often corrected by realignment of the oscillator frequency adjustments.

*Excessive Snow in the Picture.* Snow that is visible on the kinescope screen is caused primarily by shot-effect noise generated in the mixer tube. Therefore, the snow can be used as an indicator to localize a trouble between the r-f circuits preceding the mixer tube and the i-f section. Suppose that the trouble is no picture and no sound, or weak picture and sound. If the trouble occurs on some channels only, this indicates a defect in the front end, but if the picture and sound are affected on all channels the trouble can be in either the r-f section or the common i-f amplifier section. To localize between the r-f and i-f sections, turn up the contrast and volume controls to maximum and note whether the mixer-tube noise can be seen as snow on the kinescope screen and heard as a hissing sound. When an increase in receiver noise is evident, this indicates that the mixer stage is operating and all the succeeding stages are amplifying the mixer-tube noise. Therefore, if the desired signal is not going through the receiver but the mixer noise is being amplified, the trouble probably is in the circuits before the mixer tube, including the r-f amplifier, antenna, and transmission line, or in the local oscillator stage. If there is little or no increase in receiver noise evident, this indicates the trouble is in the mixer tube or succeeding i-f stages.

*Microphonic Oscillator.* The term *microphonic* is applied to the undesired mechanical vibrations of a circuit component or the electrodes in a vacuum tube, often caused by acoustic feedback from the loudspeaker at high-volume levels. All parts of the local oscillator circuit must be rigidly mounted to minimize vibrations in the components of the tuned circuit and in the tube. The local oscillator often has a heavy lead shield that fits snugly over the tube to reduce microphonics in this stage. Microphonics in the local oscillator cause slow changes of oscillator frequency at the rate of the mechanical vibrations, which can produce a sustained howl in the sound and sound bars in the picture. The microphonics may be evident only when the chassis is tapped, or at high-volume levels when the sound output from the loudspeaker can produce vibrations on the chassis. It should be noted that, in intercarrier-sound receivers, microphonics in the local oscillator have little effect on the sound, since the 4.5-Mc sound signal is independent of the exact local oscillator frequency, but oscillator microphonics can still cause sound bars in the picture.

*Cross Modulation in the R-F Amplifier.* When the modulating information on one carrier wave is transferred to another modulated carrier signal, the effect is called *cross modulation*. A very strong r-f signal in the r-f amplifier for one channel, which drives the grid voltage past cutoff, produces rectification of the signal, resulting in cross modulation with the r-f signal of other channels. For instance, if the picture signal for channel 9 is very strong in one receiver location, this cannot be rejected by the r-f signal circuits when tuned to channel 7. As a result, two pictures may be reproduced on the kinescope screen, one for channel 7 and the other corresponding to the modulation information of channel 9, transferred to the desired carrier wave by cross modulation. If the interfering signal is strong enough, it can cross-modulate more than one station. Cross modulation in the r-f amplifier is minimized by providing sufficient r-f selectivity in the antenna input circuit and using a remote-cutoff tube for the stage. Triodes give less trouble from cross modulation than pentodes.

*Mechanical Troubles.* Two common mechanical troubles in r-f tuners are dirty contacts and a faulty detent in switch-type tuners. The spring arrangement for holding the station-selector switch as it clicks into place for each channel is the *detent*. If the detent spring does not hold the station selector firmly, poor contact is made at the switch contacts and intermittent operation may result. In rotary-switch-type station selectors, the detent is part of the shaft that turns the ganged wafer switches, as illustrated in Fig. 20-22. When this type of detent is replaced, therefore, it is very important to insert the new shaft in exactly the same position as the old one. The detent must be set to the same channel as the

switch before it is installed. Otherwise, stations may be tuned in at the wrong channel setting, or some stations may appear to be tuned in best at switch positions between channels. Dirty contacts, especially on the

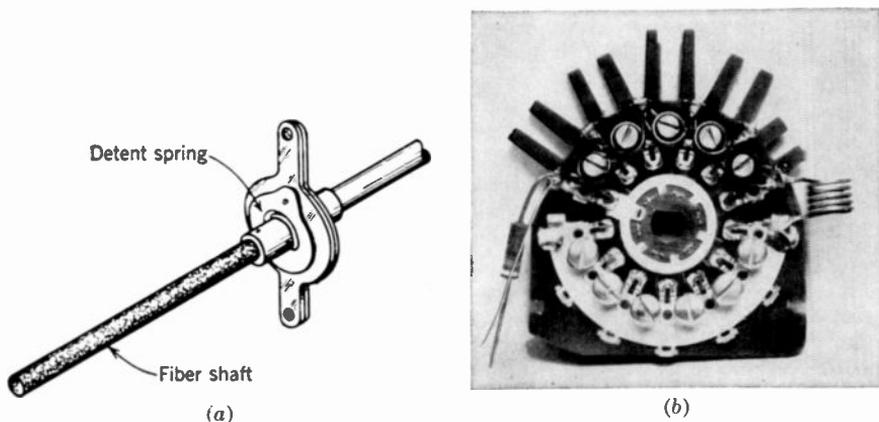


FIG. 20-22. Detent for rotary-switch-type tuner and one switch-wafer section. (a) Detent. (b) Switch wafer for oscillator section. Screws are oscillator frequency adjustments for individual channels. (From RCA KKK-2 R-f unit.)

coil strips in turret tuners, cause an intermittent trouble that often makes it necessary to jiggle the station-selector switch slightly for best results.

#### REVIEW QUESTIONS

1. What is the function of the front end in the superheterodyne television receiver? Give the specific function of each stage in the front end.
2. What is the function of the station-selector control? The fine tuning control?
3. Draw the desired r-f response curve for channel 5 (76 to 82 Mc), marking the picture and sound carrier frequencies.
4. State three requirements of the r-f amplifier stage.
5. What is the advantage of the cascode r-f amplifier, compared with a pentode stage?
6. In a receiver with an intermediate frequency of 45.75 Mc for the picture carrier, tuned to channel 5:
  - a. What is the correct operating frequency of the local oscillator, beating above the signal frequencies?
  - b. To what frequency range is the grid circuit of the mixer stage tuned?
  - c. To what frequency range is the plate circuit of the mixer tuned?
7. Describe one possible effect of local oscillator radiation interfering in a nearby receiver.
8. Explain how local oscillator radiation from a receiver with a picture carrier intermediate frequency of 25.75 Mc, tuned to channel 7, can interfere in another receiver tuned to channel 11 that has a picture carrier intermediate frequency of 45.75 Mc.
9. Draw a block diagram showing the equipment and connections needed for obtaining a visual response curve of the r-f section.

10. Explain how the local oscillator in a turret-type tuner can be aligned to the correct frequency for channel 5 by:
- Using the station's transmitted signal, in a split-sound receiver.
  - Using the station's transmitted signal, in an intercarrier-sound receiver.
  - Using a signal generator to supply r-f carrier signal, in an intercarrier-sound receiver. To what frequency will the signal generator be set?
11. a. What is the ratio of the highest operating frequency to the lowest for a local oscillator covering the v-h-f channels from 54 to 60 Mc to 210 to 216 Mc, with the local oscillator beating 45.75 Mc above the r-f picture carrier frequency? Compare this with the case of the local oscillator beating below the r-f signal frequencies.
- b. What is the ratio of highest to lowest local oscillator frequencies to cover 540 to 1,600 kc, in a radio with the oscillator beating 455 kc above the r-f signal frequencies? Compare this with the oscillator tuning ratio required to cover the u-h-f television band of 470 to 890 Mc, with the oscillator beating 45.75 Mc above the r-f picture carrier frequency.
12. Why are resonant lines often used as tuned circuits at ultra-high frequencies?
13. Referring to the schematic diagram of the r-f tuner in Fig. 20-18, give the function of:  $C_{20}$ ,  $R_8$ ,  $R_9$ ,  $C_1$ ,  $R_1$ ,  $R_2$ ,  $L_3$ ,  $L_4$ ,  $L_5$ ,  $R_7$ ,  $R_{10}$ , and  $T_1$ .
14. Referring to the u-h-f conversion strip in Fig. 20-20, to what frequency are  $L_1$  and  $L_2$  tuned for receiving channel 14 (470 to 476 Mc)?
15. Referring to the schematic diagram of the r-f tuner in Fig. 20-21, give the function of  $T_3$ ,  $L_{10}$ ,  $L_6$ ,  $R_2C_4$ ,  $L_{22}$ ,  $R_{12}$ , and  $T_1$ .
16. How can the local oscillator be checked to see if it is operating? Give one precaution to be observed when doing this.
17. Describe the effect on the reproduced picture and sound if the oscillator frequency is varied to cause:
- Sound carrier signal too high up on the i-f response curve, with the picture carrier signal having too little i-f response.
  - Picture carrier signal at 100 per cent response on the i-f curve, with practically no response for the sound i-f carrier signal.
18. For each of the following symptoms, state what the trouble probably would be and explain why:
- No picture and no sound on channel 11 only. Other channels normal.
  - No picture and no sound on all channels. There is no receiver noise evident when turning the contrast and volume controls up to maximum.
  - No picture and no sound on all channels. There is a marked increase in receiver noise when the contrast and volume controls are turned up to maximum.
  - Very snowy picture with little sound volume on channel 13 only.
  - Picture is weak on some channels but increases in contrast and brightness when the station-selector switch is moved slightly.
19. Referring to the schematic diagram of the r-f tuner in Fig. 20-18, give the effect in the picture and sound for each of the following component failures:
- Open heater in 6BQ7 r-f amplifier.
  - $C_4$  in plate circuit of r-f amplifier shorted. What effect will this have on  $R_5$ ?
  - $R_{11}$  in plate circuit of local oscillator open.
  - High-potential side of fine tuning condenser shorts to chassis. What effect will this have on the oscillator's d-c plate current and voltage?

## CHAPTER 21

### ANTENNAS AND TRANSMISSION LINES

An antenna is a conductor or group of conductors that has the function of either transmitting or receiving radiated signal in the form of electromagnetic waves. The characteristics of transmitting and receiving antennas are similar but the subject is discussed here in terms of the receiving antenna because this is the more common problem. In its elementary form, the antenna, or *aerial*, as it is often called, may be simply a length of wire, which can operate satisfactorily as the receiving antenna for the average broadcast-band receiver. For television receivers, however, the antenna problem is much more difficult. The television receiving antenna must provide sufficient signal over an extremely wide

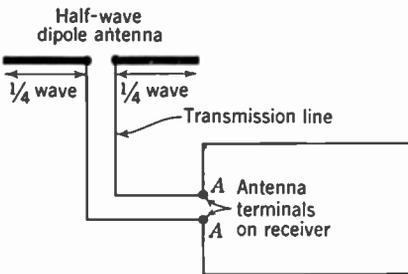


FIG. 21-1. Antenna and transmission line connected to receiver.

range of frequencies for different television channels. Also, the antenna has a relatively short physical length and the shorter the wire the smaller is the amount of signal that can be induced in the antenna. Furthermore, the problem of obtaining adequate antenna signal is aggravated by the relatively low field strengths in the v-h-f and u-h-f bands, compared with radio broadcasting in the standard broadcast band, and the television receiver needs much more antenna signal to overcome receiver noise because of the greater bandwidth. For these reasons, a television receiver usually needs an external antenna, often mounted on the roof of the house, as high as possible to obtain the maximum amount of signal with minimum external noise.

Figure 21-1 illustrates the antenna system for a television receiver, showing the antenna connected to the transmission line, which is con-

needed to the antenna input terminals on the receiver. The receiving antenna intercepts the electromagnetic wave radiated by the transmitter, inducing current and voltage in the antenna conductor to produce the antenna signal for the receiver. Note that the antenna signal includes both the picture and sound carrier signals, which are received by the same antenna. The transmission line has the function of coupling the antenna signal to the receiver. A typical antenna mounting is shown in Fig. 21-2.

**21-1. Definition of Antenna Terms. Wave Polarization.** The moving electromagnetic field, which is the radio signal, consists of two components: a magnetic field associated with the current in the antenna and the electric field associated with the potential on the antenna. The two

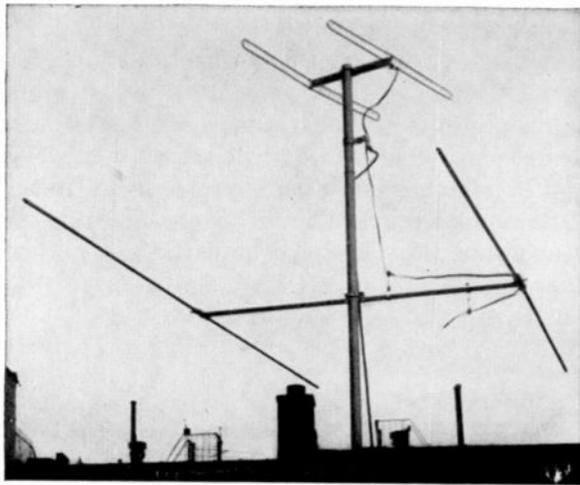


FIG. 21-2. High-low band antenna. Short folded dipole at top of mast is for channels 7 to 13; long straight dipole below is for channels 2 to 6. (From McGraw-Hill motion-picture series, *Basic Television*.)

fields are perpendicular to each other in space, and both are perpendicular to the direction of travel of the wave. When the electromagnetic wave passes through a conductor constituting the receiving antenna, it induces current in the antenna with the same variations.

The direction of polarization of the radiated electromagnetic wave is arbitrarily defined as the direction of the electric field and is determined by the physical position of the antenna in space. A horizontal antenna is horizontally polarized, since the magnetic lines of force are then in the vertical plane and the electric field is horizontal, while an antenna vertical with respect to earth is vertically polarized. Horizontal polarization is specified for transmission in the television and FM broadcast bands. Therefore the receiving antenna is mounted horizontally.

*Length of an Antenna.* The frequency, wavelength, and velocity of propagation of the radiated electromagnetic wave are related to each other by the equation

$$\lambda = \frac{\text{velocity}}{\text{frequency}} = \frac{3 \times 10^{10} \text{ cm per sec}}{f} = \frac{186,000 \text{ miles per sec}}{f} \quad (21-1)$$

where  $\lambda$  is wavelength,  $f$  is frequency, and the velocity in free space is equal to the speed of light. When the physical length of the antenna is cut to be equal to the wavelength corresponding to the signal frequency, or a submultiple of this wavelength, the antenna is resonant at that frequency and provides a resonant rise in antenna current.

The two basic antenna types are the grounded quarter-wave Marconi antenna used at lower frequencies, and the half-wave Hertz antenna shown in Fig. 21-1. The half-wave antenna usually consists of two quarter-wave elements insulated from each other, which add to provide a half wavelength. This type of antenna is called a *dipole*. The dipole operates independently of ground and, therefore, may be installed far above the earth or other absorbing bodies. Because of this, and because the physical length of a half wave is a practical size at high frequencies, the dipole is the most common type of antenna in television.

On the basis of Eq. (21-1), the formula for the length of a half wave in feet is derived as  $L = 492/f$ , where  $f$  is in megacycles, and  $L$  gives the half wavelength in feet, in terms of the electromagnetic field traveling with the speed of light. However, the resonant length of a half-wave conductor is slightly less than a half wave in free space because of the *end effect*. The antenna has capacitance that alters the current distribution at the ends of the antenna and requires foreshortening of the conductor length, in order to provide the resonant current distribution of the antenna that corresponds to the length of a half wave in free space. Wider antenna conductors and higher frequencies require more foreshortening, but it can be taken as approximately 6 per cent for the antennas used in television. Therefore, the length of the half-wave dipole is computed from the formula

$$L = \frac{462}{f} \quad (21-2)$$

where  $f$  is in megacycles and  $L$  gives the length of the half-wave dipole directly in feet. This is the actual physical length of the half-wave antenna corresponding to an electrical half wave. As illustrated in Fig. 21-1, one-half this value is used for the length of each of the two quarter-wave poles, as the small insulation distance between the two poles can be considered negligible. For a dipole tuned to 60 Mc, as an example, the length of the half wave is 7.7 ft, and each section is made 3.85 ft long.

*Antenna Impedance.* At any point on the antenna there is a definite value of current for the r-f voltage present, and the antenna has a definite impedance equal to  $E/I$  at that point. The impedance varies with the current distribution along the antenna. For a half-wave antenna the impedance is several thousand ohms at the ends and approximately 72 ohms at the center. Intermediate points have intermediate values.

*Antenna Bandwidth.* The antenna is equivalent to a resonant circuit with series resistance and reactance and as such has a definite  $Q$ , which determines its bandwidth. Since  $Q = X/R$ , the higher the antenna resistance or the lower the reactance, the lower the  $Q$ . Larger diameters for the antenna conductor have the effect of decreasing the antenna reactance, providing an antenna with lower  $Q$  and wider frequency response. Metal tubing of  $\frac{1}{4}$  to  $\frac{1}{2}$  in. diameter is generally used for television receiving antennas.

*Polar Directivity Patterns.* The variation of signal strength around an antenna is shown graphically by polar diagrams, which are circular

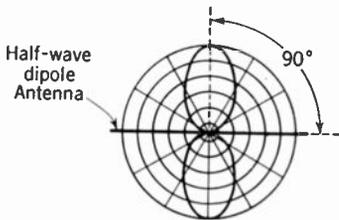


FIG. 21-3. Polar directivity pattern for half-wave dipole antenna, in horizontal plane.

charts with zero taken at the center and the circumference laid off in angular degrees as shown in Fig. 21-3. Signal strength is plotted in polar coordinates to show magnitude and direction, the angle giving the direction for which the signal strength is plotted and the length of the radial arm defining the magnitude. The polar diagram shows the directivity pattern of the antenna, as determined by the current distribution of the antenna conductor. For a transmitting antenna, the pattern shows in which direction the antenna radiates the most signal; for a receiving antenna the pattern shows the direction from which most signal is intercepted by the antenna conductor. A half-wave dipole at its fundamental resonant frequency has the directivity pattern illustrated in Fig. 21-3, often called a *figure-eight* pattern, indicating that the antenna receives best from the front and back, broadside to the antenna conductor, with little signal received from directions off the ends of the antenna.

*Antenna Gain.* This is a term used to express the increase in signal for one antenna over a standard antenna, usually a half-wave dipole having the same polarization. Antenna gain is generally used in connection with directional antenna systems and is measured in the optimum direction.

It is usually stated in decibels. An antenna with a gain of 3 db, as an example, has a power gain of 2 or voltage gain of 1.4.

*Front-to-back Ratio.* This indicates the amount of signal the antenna receives from the front compared with the signal received from the back. As an example, if the antenna intercepts  $1,000 \mu\text{v}$  of signal from a transmitter in front, but only  $500 \mu\text{v}$  for signal of the same frequency arriving from the back, the front-to-back ratio is 2 times in voltage, or 6 db.

**21-2. Ghosts.** A duplicate image of the reproduced picture, offset a little to the side of the original, as shown in Fig. 21-4, is generally called a *ghost*. Ghosts are commonly caused by multipath reception of reflected signals. Referring to Fig. 21-5, the antenna at point *C* can receive picture signal by two separate paths from the transmitter at point *A*. The path



FIG. 21-4. Reproduced picture of test pattern, with ghost to the right caused by reflected signal. (From McGraw-Hill motion-picture series, *Basic Television*.)

*ABC* for the signal reflected from the building at point *B* is longer than the direct path *AC* by 2 miles, in this case. Since the velocity of the radiated signal is 186,000 miles per sec and the reflected wave path is 2 miles longer than the direct path, the reflected wave is delayed by  $2/186,000$  sec in reaching the antenna. This is approximately equal to  $11 \mu\text{sec}$ . The electron beam scanning the screen of the picture tube requires about  $55 \mu\text{sec}$  to scan across one horizontal line. On a picture having a width of 10 in., then, it requires only  $5.5 \mu\text{sec}$  to scan 1 in. The reflected signal, delayed in reception by  $11 \mu\text{sec}$ , will produce a second image displaced from the original by 2 in. The ghost is displaced to the right, in the direction of scanning, because the reflected signal arrives at the receiving antenna later in time than the direct signal.

With multiple reflections there may be multiple ghosts. The intensity of the ghost may be nearly as strong as the original image or it may be just

noticeable, any difference in relative intensity being the result of attenuation suffered by the reflected wave in its travel. The ghost may be a positive or negative image, depending on the relative phase between the multipath r-f signals. The degree of interference produced in the picture may vary from definite ghost images to reflections that are not noticeable as duplicate images because of insufficient time delay but which cause the picture to appear fuzzy. A delay distance of about 50 ft or less can be considered negligible, since the resultant horizontal displacement on the

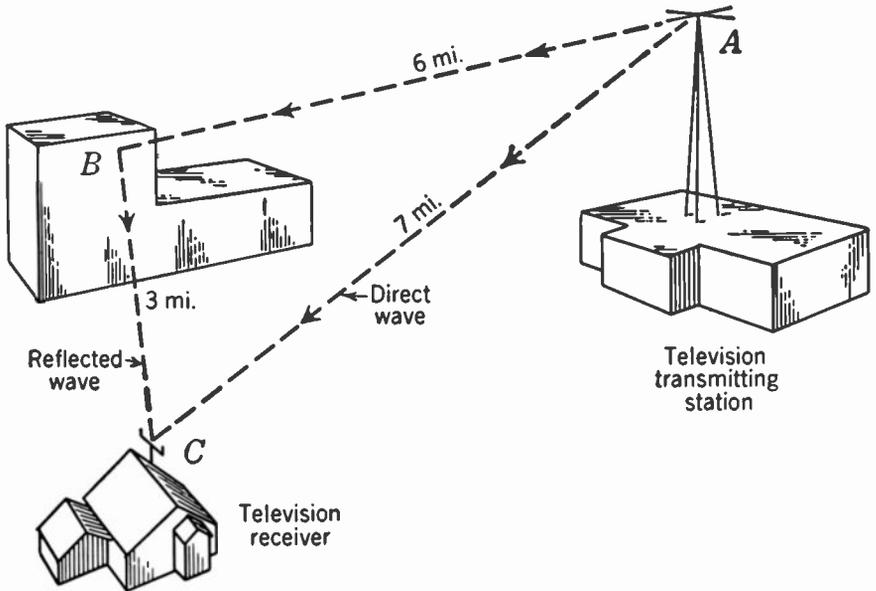


FIG. 21-5. Reception of multipath signals. The reflected wave distance  $ABC$  is 2 miles longer than the direct path  $AC$ .

screen of the picture tube is then considerably less than the width of a picture element.

For the problem of multipath reception, an antenna that has a good front-to-back ratio and a narrow forward lobe with minimum side responses can be rotated horizontally to minimize ghosts. Sometimes changing the antenna location reduces the intensity of the ghost.

**21-3. Dipole Antennas.** The half-wave dipole illustrated in Fig. 21-1 is the basis of practically all television antennas because it is efficient and easy to construct for the high frequencies in the v-h-f and u-h-f television bands. This type is often called a *Hertz, doublet, half-wave*, or simply *straight dipole* antenna. Calculation of the dipole length for a half wave can be made from the formula:  $L = 462/f$  where  $L$  is in feet and  $f$  in megacycles. For a dipole cut to be one-half wave long at 69 Mc, in the

center of channel 4, for instance, the dipole length is approximately 6.7 ft. The dipole is constructed of large-diameter conductor, often  $\frac{1}{2}$ -in. aluminum tubing for light weight. The transmission line connects to the antenna at the center, where the impedance is approximately 72 ohms. The dipole has an antenna gain of 1, since the amount of signal picked up by a tuned dipole is the reference standard to which other antenna types are compared to indicate their gain.

*Current and Voltage Distribution.* The current and voltage distribution on a half-wave antenna is shown in Fig. 21-6. With r-f excitation by the radiated electromagnetic field, current is induced in the antenna with the same variations as the applied voltage. The electron flow is not instantaneous but travels along the wire in free space with the speed of light. When the charge reaches the end of the wire, the resulting accumulation of charge at the end provides a potential for moving the charge in the opposite direction, reversing the direction of current flow and reflecting the current wave. The resultant current is zero at the ends because of the reflection, which produces two currents of equal amplitude flowing in opposite directions at the end. Farther back on the wire, the outgoing and reflected currents are not the same, since the charges causing the currents have been supplied to the antenna at different parts of the r-f cycle.

Maximum current is at the center, where the reflected current adds to the original current. The ends of the antenna in free space are points of maximum voltage and zero current. Because of capacitance of the ends, however, the current is normally not zero at the ends but has a definite value, and the antenna must be foreshortened to give the same current distribution that would be obtained for a half wave in free space.

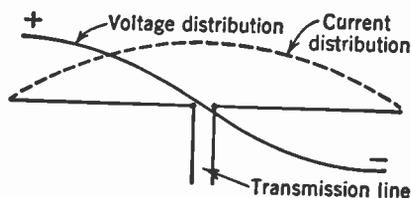


FIG. 21-6. Current and voltage distribution on half-wave antenna.

The current distribution on the horizontal half-wave antenna produces the figure-eight horizontal field pattern shown in Fig. 21-3. Signal is received best broadside to the antenna and equally well from front or back, with very little signal received from either end. Therefore, the half-wave dipole should be broadside to the transmitter, positioned for receiving maximum desired signal while eliminating undesired signals coming from directions off the ends of the antenna. Rotating the antenna for the direction of best signal pickup is generally called *orientation*.

The antenna impedance corresponding to the current and voltage distribution consists of resistance and reactance. The reactance is capacitive at frequencies below resonance or inductive above resonance. At resonance the reactance is zero and the antenna impedance is resistive.

The center of the half-wave dipole is a point of maximum current and therefore has the minimum antenna resistance value equal to about 72 ohms. At the ends, where the current is minimum, the antenna resistance is high, being several thousand ohms for a conductor diameter of  $\lambda/500$ . At intermediate points on the dipole, which are not a quarter wave from the ends, the antenna impedance has a reactive component. Wider diameter antenna conductors result in lower resistance and reactance values.

*Response off Resonance.* When the television receiving antenna is used to receive several channels, the directional characteristics and impedance of the dipole at frequencies off half-wave resonance vary as the antenna current and voltage distribution change. Referring to Fig. 21-7a, the

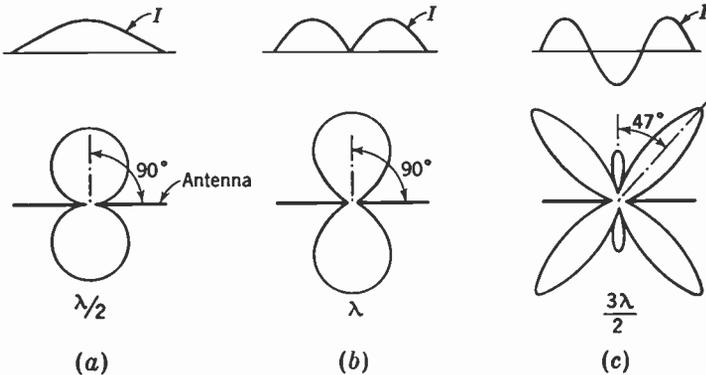


FIG. 21-7. Directional characteristics of dipole antenna connected to transmission line at center, with distribution of current  $I$  on antenna. (a) Figure-eight pattern at half-wave resonance. (b) Broadened response at twice the resonant frequency. (c) Lobes split because of operation as harmonic antenna at three times the resonant frequency.

dipole has the figure-eight directivity pattern at its fundamental resonant frequency, when the antenna length is one-half wave. Maintaining the same physical length, at double this frequency the dipole is a full-wave antenna. It still has the figure-eight directivity pattern but the impedance at the center, where the transmission line connects, is now a point of maximum antenna resistance. At three times the fundamental resonant frequency, the same physical antenna length is three half waves. The center is again a point of minimum antenna resistance, equal to about 100 ohms. However, notice that for the third harmonic frequency the directional pattern splits into four major lobes. There is a gain of 1 db in the direction of maximum reception but this is  $47^\circ$  off the ends, with little pickup from the broadside direction. At  $4\lambda$  the response of the dipole from the broadside direction is practically zero. The point of minimum

response is called a *null*. In order to utilize the figure-eight directivity pattern and prevent the null in the broadside direction, therefore, the center-fed dipole is limited to operation over a frequency range of about 2 to 1 or less.

**Folded Dipole.** As shown in Fig. 21-8, the folded dipole is constructed of two half-wave rods joined at the ends, with one rod open at the center where it connects to the transmission line. The spacing between the rods is small compared with a half wavelength, 1 to 3 in. being satisfactory for a v-h-f antenna. The half-wave folded dipole has the same directional characteristics as the half-wave straight dipole, with the same amount of signal pickup. However, the antenna resistance of the folded dipole is approximately 300 ohms, which is a convenient value for matching to 300-ohm transmission line. The center of the closed section of the half-wave folded dipole is a point of minimum voltage, allowing direct mounting at this point to a grounded metal mast without shorting the signal voltage. It should be noted that, with an operating frequency twice the fundamental half-wave resonant frequency, the full-wave folded dipole does not have the figure-eight polar pattern, receiving little signal from the broadside direction.

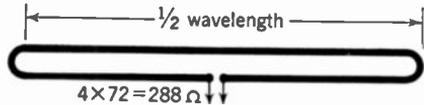


FIG. 21-8. Folded half-wave dipole antenna.

The resistance of the half-wave folded dipole at the center where it connects to the transmission line is higher than the value for a straight dipole

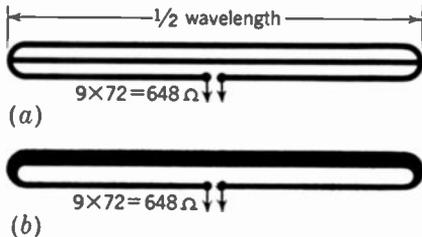


FIG. 21-9. High-impedance folded dipole antennas. (a) With additional closed conductor section. (b) With wider-diameter closed conductor section.

because only part of the total antenna current flows in the open section. As a result, the folded dipole antenna resistance equals 72 ohms multiplied by the square of the ratio of the total diameter of all conductor sections to the diameter of the open section. As an example, in Fig. 21-8 with the same diameter for the two conductor sections, the total diameter is twice the diameter of the open section. Therefore, the antenna impedance is  $4 \times 72$ , or 288, ohms, which is generally considered as approximately equal to 300 ohms. In applications where a higher resistance is desired for the folded dipole, the diameter of the closed conductor section is increased. In Fig. 21-9a, an additional closed conductor of the same diameter is added, making the total conductor diameter triple the diameter of the open section, to provide an antenna resistance of  $9 \times 72$ , or 648, ohms. The same impedance trans-

formation can be obtained by increasing the diameter of the closed conductor, as in *b*, instead of adding sections.

**Broad-band Dipoles.** A *thick dipole* antenna, which has a cross-sectional dimension approximately  $0.1\lambda$  or greater, can provide more uniform response over a wider band of frequencies, compared with a *thin dipole* conductor having negligible diameter. Figure 21-10 shows several

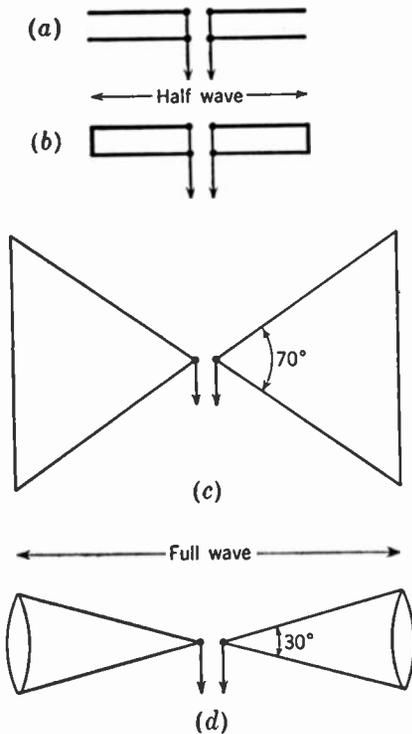


FIG. 21-10. Broad-band dipole antennas. (a) and (b) Half-wave dipoles. (c) Full-wave triangular dipole. (d) Full-wave dicone antenna.

The centers are joined in *a* and the ends can also be joined, as in *b*, since the end points of the two antennas are at the same potential. The triangular dipole in *c* and conical dipole in *d*, however, are operated as full-wave antennas because they have too low a resistance at the center at half-wave resonance. As a full-wave dipole, the antenna has a gain of approximately 3 db. The included angles shown provide an antenna resistance of approximately 300 ohms at the center, for full-wave resonance, smaller angles resulting in higher impedance. The over-all physical length can be foreshortened about 10 per cent for the triangular dipole and 25 per cent for the dicone antenna. The triangular dipole, often called a *bowtie*

types of thick broad-band dipoles. These can be constructed of wire conductors, metal sheets, wire screening, or metallic foil. The increased thickness has the following three effects on the dipole characteristics:

1. The antenna resistance decreases and the reactance is lowered even more, resulting in an antenna with lower  $Q$  that has more uniform impedance values over a wide band of frequencies.

2. The broadside response of the directional pattern is maintained over a wider frequency range, before splitting into multiple lobes.

3. More foreshortening is needed with increasing thickness to provide physical lengths equivalent to the electrical wavelengths.

Referring to Fig. 21-10, the antennas in *a* and *b* are operated as half-wave dipoles. When the conductors are separated by  $0.1\lambda$  or less they can be considered as one uniform antenna of wider cross section.

antenna, is commonly used as a broad-band receiving antenna to cover all the u-h-f channels. The nonuniform cross section, with wider diameter at the ends, improves the broad-band characteristics.

**21-4. Long-wire Antennas.** Compared with a half-wave dipole, a long-wire antenna, which is several wavelengths, has the advantages of increased signal pickup and sharper directivity. As the antenna wire is made longer in terms of the number of half waves, the directional pattern changes because of the current distribution, increasing the directivity along the line of the antenna wire itself. This is illustrated by the polar directivity patterns in Fig. 21-7.

**V Antenna.** When two long wires are combined in the form of the horizontal V antenna shown in Fig. 21-11, the major lobes of the directional pattern for each wire reinforce along the line bisecting the angle between the two wires. Therefore, the V antenna receives best along the

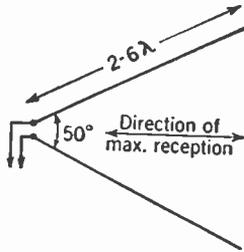


FIG. 21-11. V antenna, with wire conductors for the legs.

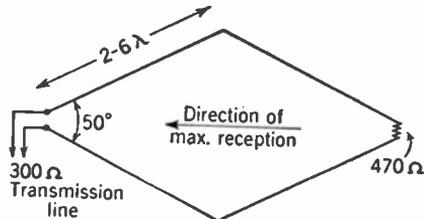


FIG. 21-12. Rhombic antenna.

line of the bisector. The greater the electrical length of the conductor legs, the smaller the included angle of the V should be for maximum antenna gain. The impedance of the V antenna is about 600 to 800 ohms.

**Rhombic Antenna.** A more efficient arrangement is the rhombic antenna shown in Fig. 21-12 which consists of two horizontal V sections. To make it unidirectional, the rhombic antenna can be terminated with a resistor of 470 ohms, for an approximate match to 300-ohm transmission line. Both the V and rhombic antennas are mounted horizontally for horizontal polarization as television antennas. Each leg should be at least two wavelengths at the lowest operating frequency, the gain and directivity of the antenna increasing with the length. The angle of  $50^\circ$  is a compromise value suitable for leg lengths of 2 to  $6\lambda$ . Longer legs should have a smaller angle. With each leg four wavelengths, the V antenna has a gain of 7 db, while the rhombic antenna gain is 10 db, approximately. At ultra-high frequencies these lengths are practicable, to provide a high-gain antenna for the u-h-f television band. The rhombic can provide a uniform value of antenna impedance over a total frequency range of 3 to 1, with high gain and sharp directivity.

**21-5. Parasitic Arrays.** When current flows in the receiving antenna it radiates part of the intercepted signal, as in a transmitting antenna. If a conductor approximately one-half wave long is placed parallel to the half-wave dipole antenna but not connected to it, as illustrated in Fig. 21-13, the free wire will intercept some of the signal radiated by the antenna. This signal is reradiated by the free wire to combine in phase with the original antenna current. As a result, part of the intercepted signal lost by radiation in the receiving antenna is recovered by using the free wire, providing increased gain and directivity. The free wire is called a *parasitic* element because it is not connected to the dipole antenna. A parasitic element placed behind the antenna is a *reflector*; a parasitic element in front of the antenna is a *director*. The dipole antenna itself, to which the transmission line is connected, is often called the *driven element*. This can be either a straight dipole or a folded dipole. A dipole antenna

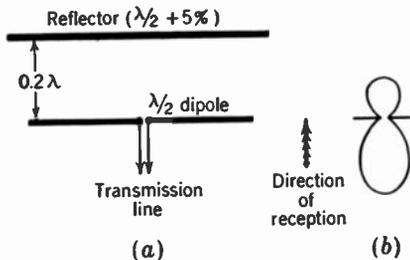


FIG. 21-13. Dipole with reflector in back of antenna. (a) Spacing of elements. (b) Directional pattern.

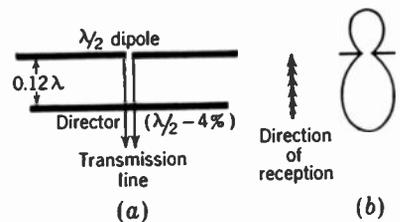


FIG. 21-14. Dipole with director in front of antenna. (a) Spacing of elements. (b) Directional pattern.

with one or more parasitic elements is a *parasitic array*. This is the most common type of television receiving antenna because it is simple to construct, can be oriented easily, provides enough gain for average signal strengths, and increases the directivity compared with a dipole alone. The main directional effect of the parasitic element is to reduce the strength of signals received from the rear of the antenna, making its response unidirectional, as illustrated in Fig. 21-13b. Therefore, an antenna with a parasitic element is useful for reducing the strength of multipath reflected signals arriving from directions behind the antenna, to eliminate ghosts in the picture.

**Dipole and Reflector.** Referring to the dipole and reflector in Fig. 21-13, the reflector is usually placed approximately  $0.2\lambda$  behind the dipole, to reinforce signals arriving from the front, and is about 5 per cent longer than the dipole. The phase of the current in the parasitic element with respect to the current in the dipole determines how the directional pattern and gain will be altered. This depends on the spacing between the ele-

ments, and the tuning of the parasitic, which is adjusted by changing its length. Closer spacing lowers the antenna impedance and narrows the frequency response. For the typical arrangement shown in Fig. 21-13 the dipole and reflector have a gain of approximately 5 db. The antenna impedance is about one-half the impedance of the antenna itself, resulting in 150 ohms for the folded dipole and reflector and 36 ohms for the straight dipole and reflector. The front-to-back ratio is approximately 4 db, resulting in about  $1\frac{1}{2}$  times more signal voltage from the front than from the back. Spacings closer than  $0.12\lambda$  lower the gain and front-to-back ratio appreciably. It is important to note that the parasitic element is effective only within the frequency range for which it is approximately tuned to half-wave resonance. Furthermore, the increase in gain and directivity with a reflector cuts off sharply at frequencies lower than the resonant frequency. For this reason, a dipole with reflector is usually cut for the lowest frequency in the range to be covered. The reflector can then be effective up to a frequency about 30 per cent higher than the resonant frequency.

*Dipole with Director.* Referring to the parasitic director element in Fig. 21-14, it is placed  $0.12\lambda$  in front of the dipole, and is about 4 per cent shorter than the driven element, to reinforce signals arriving from the front, while reducing reception from the back. At  $0.12\lambda$  spacing, the gain of the dipole with director is 5 db and the front-to-back ratio approximately 3 db. The antenna resistance is about one-fourth the value of the driven element by itself. Higher values of gain and front-to-back ratio can be obtained by using closer spacing for the director, but the antenna resistance becomes lower. The gain and directivity of the dipole with director drop off sharply at frequencies higher than the resonant frequency, which is opposite to the operation of the dipole and reflector off resonance. Practically all television antennas with one parasitic element use a reflector, instead of a director, because the director needs closer spacing for the same gain and front-to-back ratio, which reduces the antenna resistance and narrows the frequency response. For an antenna operating over a relatively narrow frequency, however, directors are combined with a dipole and reflector. A dipole with one reflector and one director, having the same spacings as for a single parasitic element, provides about 7 db gain, while the antenna resistance is approximately one-eighth the value of the driven element by itself.

*Yagi Antenna.* A dipole with one reflector and two or more directors, as illustrated in Fig. 21-15, is called a Yagi antenna. This is a compact high-gain array, with a sharp forward broadside lobe, and narrow bandwidth often used in low-signal areas to cover one television channel, or several adjacent channels. The gain of the Yagi antenna with three parasitic elements is about 10 db, with a front-to-back ratio of approxi-

mately 15 db. A high-impedance folded dipole is generally used for the driven element so that the reduced value of antenna resistance with the parasitic elements can be about 150 to 300 ohms.

*Dipole with Corner Reflector.* The antenna in Fig. 21-16 has a reflector constructed as a corner conducting sheet behind the half-wave dipole. The corner reflector can be either a solid metal sheet or a grid consisting of wires or wire screening, provided that the spacing between grid wires is  $0.1\lambda$  or less. The dipole antenna, insulated from the parasitic reflector, is mounted along the line bisecting the  $90^\circ$  corner. Maximum signal is received from the front along this line. Compared with a straight-rod reflector, the corner reflector provides increased gain over a wider band of frequencies. The gain is approximately 10 db, front-to-back ratio 20 db, and the antenna resistance about two-thirds the resistance of the dipole

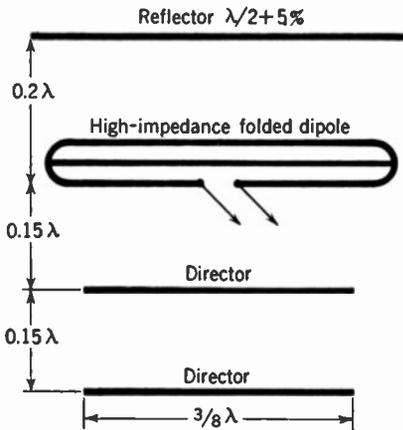


FIG. 21-15. Yagi antenna, consisting of dipole with one reflector and two directors.

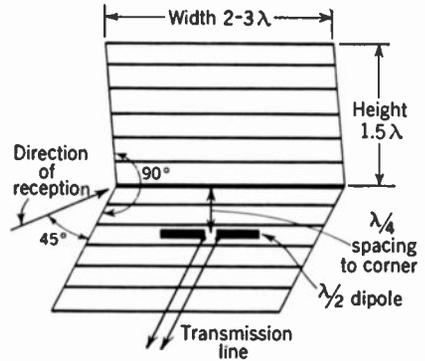


FIG. 21-16. Dipole with corner reflector.

by itself, for the quarter-wave spacing of the dipole to the corner reflector shown in Fig. 21-16. Increasing the dipole spacing up to one-half wave increases the gain and antenna resistance, but the polar pattern splits into multiple lobes. Angles of less than  $90^\circ$  for the corner reflector increase the gain but result in very low value of antenna resistance. It should be noted that the length and width of the corner reflector are not critical but need only be long compared with the spacing to the dipole in front, to provide the effect of an infinite conducting sheet to intercept maximum signal for the driven antenna. The reflector can also be in the form of a plane sheet, corresponding to a  $180^\circ$  angle but this provides less gain than the  $90^\circ$  corner reflector. With spacings of 0.1 to  $0.4\lambda$  between the plane reflector and dipole, the average antenna gain is approximately 5 db and the antenna resistance 45 to 90 ohms. At ultra-high frequencies, the

physical dimensions are small enough to make the dipole with corner reflector or plane reflector a useful high-gain antenna for the v-h-f television band.

**21-6. Multiband Antennas.** All the television broadcast channels can be considered in three bands: 54 to 88 Mc for the low-band v-h-f channels, 174 to 216 Mc for the high-band v-h-f channels, and 470 to 896 Mc for the u-h-f channels. The main problem in using one dipole for both v-h-f bands is maintaining the broadside response, as the directional pattern of a low-band dipole cut for a half wavelength in the 54- to 88-Mc band splits into side lobes at the third and fourth harmonic frequencies in the 176- to 216-Mc band. A high-band dipole cut for a half wavelength in the 176- to 216-Mc band is not suitable for the 54- to 88-Mc band because of insufficient signal pickup at the lower frequencies. As a result, dual-band antennas for both v-h-f bands generally use either separate dipoles for each band or a dipole for the 54- to 88-Mc band modified to provide broadside unidirectional response in the 174- to 216-Mc band also. For the u-h-f band, a v-h-f antenna can operate as a long-wire antenna or be used as a reflector sheet behind a broad-band dipole added in front. It should be noted that higher gain is necessary for a u-h-f antenna to provide the same signal strength as on the v-h-f channels because the reference-tuned dipole is shorter at ultra-high frequencies.

*Conical Dipole.* The v-h-f dual-band antenna illustrated in Fig. 21-17 is generally called a *conical*, *forked*, or *fan* dipole. The conical-dipole antenna consists of two half-wave dipoles inclined about  $30^\circ$  from the horizontal plane, similar to a section of a cone, and usually a horizontal dipole in the middle. All the dipoles are tilted inward toward the wave front of the arriving signal at an angle approximately  $30^\circ$  from the broadside direction, resulting in a total included angle of  $120^\circ$  between the two conical sections. Smaller values of included angle reduce the amount of signal intercepted at low frequencies, as the distance across the front is decreased. Either a straight reflector or conical reflector can be used behind the conical dipole with approximately the same results. Cut for a half wavelength at channel 2, the conical dipole with reflector is commonly used as a receiving antenna to cover both v-h-f bands. For the 54- to 88-Mc band, the antenna is a conical-type half-wave dipole with a parasitic reflector, providing a directivity pattern that has the desired unidirectional broadside response with a relatively uniform antenna resistance of about 150 ohms. However, minor side lobes are usually present because the antenna is not a symmetrical cone with respect to the ends. For the 174- to 216-Mc band, the tilting of the dipole rods shifts the direction of the split lobes produced at the third and fourth harmonic frequencies, so that they combine to produce a main forward lobe in the broadside direction.

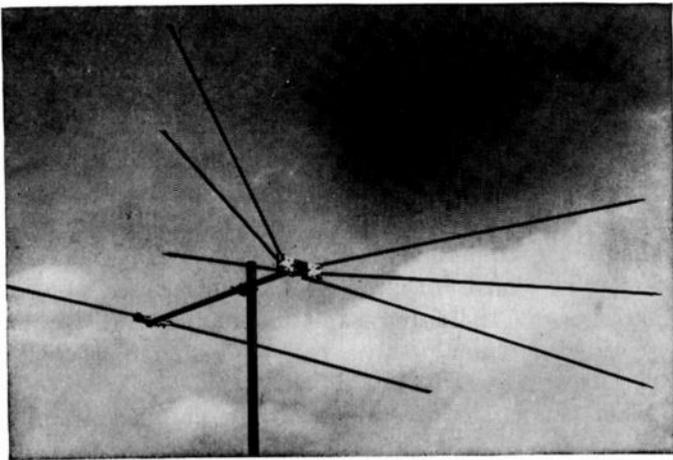
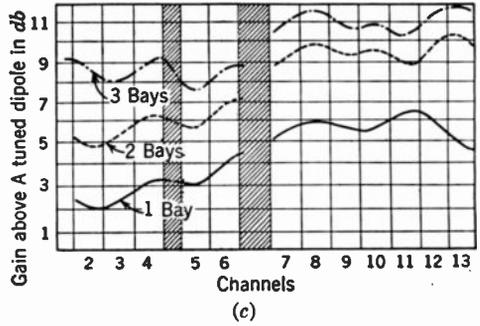
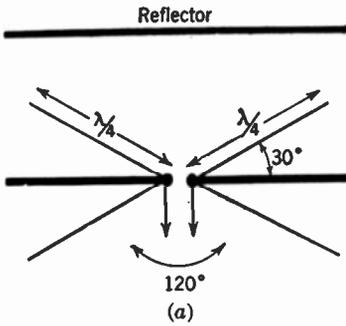


FIG. 21-17. Conical dipole with reflector. (a) Spacing of elements. (b) Typical antenna, single bay. (c) Gain-frequency response curves for channels 2 to 13. (Channel Master Corporation.)

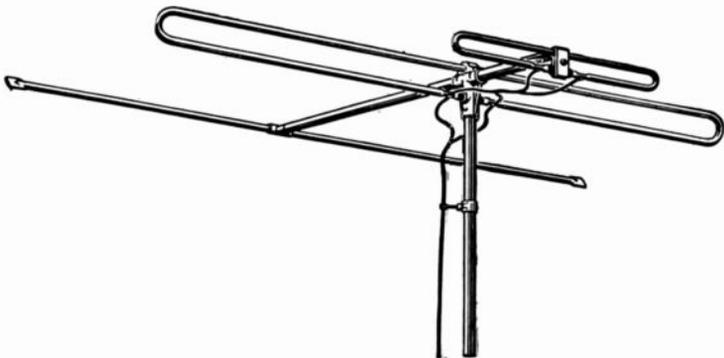


FIG. 21-18. In-line dual-band dipole antenna. (American Phenolic Corporation.)

*In-line Antenna.* As shown in Fig. 21-18, this v-h-f dual-band antenna consists of a half-wave folded dipole with reflector for the 54- to 88-Mc band, in line behind the shorter half-wave folded dipole for the 176- to 216-Mc band. The distance between the two folded dipoles is approximately one-quarter wavelength at the high-band dipole frequency. This is the length of line connecting the short dipole to the long dipole, where the transmission line to the receiver is connected. For the low-band channels, the long folded dipole with reflector supplies antenna signal to the transmission line, as the short dipole has little pickup at these low frequencies.

For the high-band channels, the short folded dipole supplies signal to the transmission line, with the long folded dipole operating as a reflector. The directivity pattern of the in-line antenna is relatively uniform on all v-h-f channels, with a unidirectional broadside response that has practically none of the undesired side-response lobes. The average antenna gain on the low-band channels is approximately 2 db, and 5 db on the high-band channels. The antenna resistance is about 150 ohms.

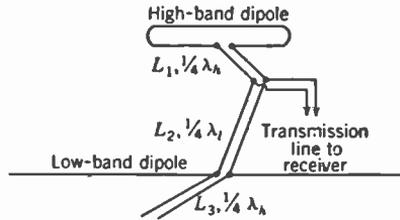


FIG. 21-19. High-low-band dipoles connected to common transmission line. Reflectors omitted for clarity.  $\lambda_h$  is high-band wavelength;  $\lambda_l$  low-band wavelength.

The average antenna gain on the low-band channels is approximately 2 db, and 5 db on the high-band channels. The antenna resistance is about 150 ohms.

*High-Low Antenna.* The v-h-f dual-band antenna shown previously in Fig. 21-2 uses a separate dipole and reflector for the 54- to 88-Mc band, mounted above the dipole and reflector for the 176- to 216-Mc band. This allows separate orientation of each dipole, which can be an advantage when stations in the two bands are in different directions. A folded dipole or straight dipole can be used for either antenna or both antennas. The short dipole is usually mounted at the top of the mast for mechanical considerations, although it can be in a separate location if this provides more signal. The distance above the long dipole is not critical, but approximately the same half wavelength as the short dipole is a suitable spacing. Best results are obtained when separate transmission lines are connected to each antenna, with a double-pole double-throw switch at the receiver to select either one, but the two dipoles can be connected to a common transmission line as shown in Fig. 21-19. Each dipole is connected by a quarter wavelength section of line to the common junction where the transmission line to the receiver is connected. For the low-band channels, the short dipole has little signal pickup, while the long dipole supplies antenna signal to the transmission line. For the high-band channels, both dipoles supply antenna signal. However the open quarter-wave stub  $L_3$  acts as a short circuit for high-band signals picked

up by the long dipole that would cancel the antenna signal from the short dipole. As a result, broadside unidirectional response is maintained on both v-h-f bands. The gain of the high-low antenna is approximately 3 db on the low band and 6 db on the high band. The antenna resistance is the same as for each section by itself.

**Dual-frequency Dipoles.** Figure 21-20 shows several methods of adding short-dipole rods to a low-band dipole antenna in order to maintain the broadside response at its third and fourth harmonic frequencies. The short rods are mounted in the same vertical plane as the long dipole but inclined about  $45^\circ$  from the horizontal. In order to be effective in providing broadside response for the desired harmonic frequency, the rods

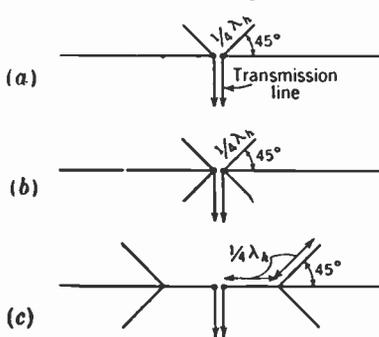


FIG. 21-20. Dual-frequency dipoles.  $\lambda_h$  is high-frequency wavelength. (a) Straight dipole elements added at center for third harmonic. (b) Double-dipole elements added at center for third harmonic. (c) Double-dipole elements added quarter wavelength from center for fourth harmonic.

are connected where the current is minimum and impedance maximum at harmonic resonance on the long dipole. In *a*, the short dipole is connected at the center of the long dipole, in order to be effective in providing broadside response as a half-wave antenna at the fourth harmonic frequency. At even harmonic frequencies the center has minimum current, equivalent to the ends, as the values repeat every half wavelength. The double-dipole rods in *b*, generally called *whiskers* or *wings*, also are connected at the center to be effective at the fourth harmonic frequency. In *c*, however, the whiskers are connected a quarter wavelength from the center for operation at the third harmonic frequency. As a result, the short dipole operates as a full-wave antenna, allowing gain up to 3 db with broadside response. It should be noted that, in order to obtain unidirectional response at the harmonic frequency, a short reflector must be added.

**V-H-F and U-H-F Antenna.** The antenna shown in Fig. 21-21 combines a v-h-f conical dipole with reflector and u-h-f triangular dipole to cover all the television channels in the v-h-f and u-h-f bands. For the low-band channels 2 to 5, the antenna is a conical dipole with parasitic reflector, while for the high-band channels 7 to 13 it operates as a V-type antenna with large-diameter conductors, just like the fan antenna illustrated in Fig. 21-17. On the 470- to 890-Mc band, the conical elements form a sheet reflector for the small triangular dipole, covering the u-h-f channels 14 to 83 with an average antenna gain of 5 to 6 db.

**21-7. Stacked Arrays.** In order to increase the antenna gain and directivity, two or more antennas of the same type can be mounted close to each other and connected to a common transmission line. The individual antennas are called *bays* and the combined unit is a *stacked array*. Mounting the antennas one above the other, as illustrated in Fig. 21-22, is vertical stacking. In general, stacking can provide an additional gain up to 3 db for each antenna bay added, since the larger the area of the array the greater is the amount of signal that can be intercepted. To utilize the signal, however, the individual bays must be phased correctly

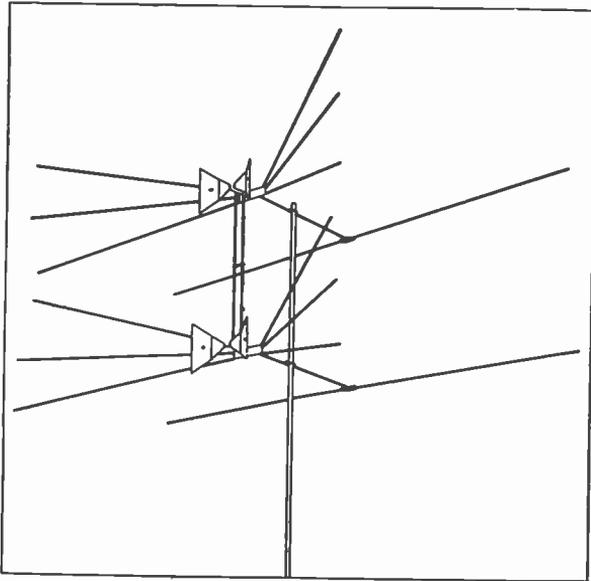


FIG. 21-21. V-H-F fan antenna with u-h-f triangular dipole. (Channel Master Corporation.)

with respect to the common junction for the transmission line, so that the antenna signals can add to provide the required antenna gain, with the desired directional response. Rods used for interconnecting the bays in the stacked array are called *phasing rods*. In phasing separate antennas to a common junction for the transmission line, the following points should be noted:

1. A difference of one-quarter wavelength in the distance the antenna signal must travel is equivalent to a phase-angle difference of  $90^\circ$ .
2. A difference of one-half wavelength is equivalent to a phase-angle difference of  $180^\circ$ .
3. Reversing the connections is equivalent to a phase-angle difference of  $180^\circ$ .

*Vertical Stacking.* This is often done with television receiving antennas in weak signal areas in order to increase the gain and to sharpen the directivity in the vertical plane, which reduces pickup of external noise from sources usually below the antenna. Figure 21-22 illustrates vertical stacking and how the antenna bays are usually connected to the common transmission line. Folded dipoles are shown, without reflectors for simplicity, but the same principles apply for any other antenna. The half-wave spacing generally used between stacked antenna bays is convenient for interconnecting them with two quarter-wave sections. Because of the symmetry, the antenna signals from both bays travel the same distance to point *X* and are in phase at the common junction for the transmission line to the receiver. The impedance here is one-half the impedance of either section, as the two lines are in parallel. With more than two antennas in the array, they can be grouped in adjacent pairs, and the

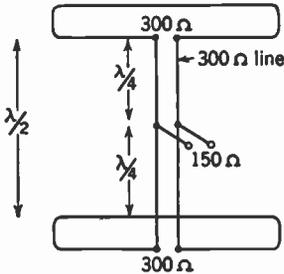


FIG. 21-22. Two vertically stacked antenna bays connected to common transmission line.

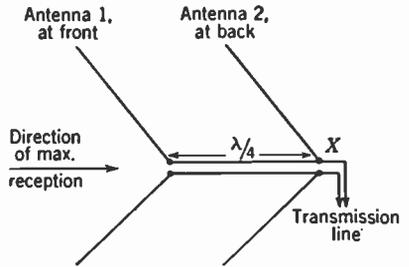


FIG. 21-23. Double-V antennas mounted one behind the other in horizontal end-fire array.

pairs are then interconnected to the common transmission line. The directivity pattern of the vertically stacked array has the same broadside response as the individual antennas.

*Unidirectional End-fire Array.* Referring to Fig. 21-23, two antennas are stacked one behind the other a quarter wavelength apart and connected to the common transmission line at point *X* to form an array that has the unidirectional response shown, without any parasitic reflector or director. The arrangement is called an *end-fire array* because it has maximum response off the end of the line of the array. The end-fire directivity results from making the current in the individual antennas out of phase with each other. With  $90^\circ$  phasing the response is unidirectional off the end farthest from the transmission line. The antenna to which the transmission line connects is then the back of the array. When receiving from the front, antenna 1 in Fig. 21-23 intercepts the signal a quarter cycle sooner than antenna 2, but the quarter-wave line connecting the two antennas delivers this signal at point *X* in the same phase as the

signal intercepted by antenna 2. The array provides a gain of 3 db, therefore, for signal arriving from the front. Signal from the back, however, is intercepted by antenna 1 a quarter cycle later than antenna 2. With the additional quarter wavelength of the connecting line, the signal delivered by antenna 1 arrives at point X 180° out of phase with the signal from antenna 2. The two signals cancel, therefore, resulting in minimum reception from the back.

**21-8. Transmission Lines.** The transmission line has the function of delivering the antenna signal to the receiver with minimum loss. The line itself should not pick up any stray signal and for this reason should be either balanced or shielded or both. A line is balanced when each of the two conductors has the same capacitance to ground. Then the in-phase signal currents in the two conductors, due to stray pickup by the line, are canceled with a balanced receiver input circuit. Otherwise, the interfering signal pickup by a long line may combine with the desired signal from the antenna to produce time-delay distortions, similar to multipath reflected signals, causing ghosts in the picture. A shielded line is completely enclosed with a metallic braid that is grounded to serve as a shield for the inner conductor. The main types of transmission line generally used for television receiver antenna installations are the concentric or *coaxial* line and the two-wire parallel conductor, as illustrated in Fig. 21-24. Parallel-wire line constructed in the form of a plastic ribbon as shown in *a* and *b*, is generally called *twin lead*, either flat or tubular. The type in *c* is *open-wire* line. These are balanced lines but not shielded. The coaxial line in *d* is shielded but unbalanced. Shielded lines generally have higher losses than the equivalent unshielded line, as shown in Table 21-1, which lists the attenuation and characteristic impedance for the common transmission lines. The attenuation is caused by  $I^2R$  losses in the a-c resistance of the line, reducing the amplitude of antenna signal delivered by the line to the receiver. The longer the line and the higher the frequency, the greater the attenuation. The characteristic impedance of the line results from the uniform spacing between the two conductors and is the same regardless of the length of line.

*Flat Twin Lead.* This is the most popular type of line because it has low losses, is available in 300 ohms characteristic impedance, costs less, and is flexible for ease of handling. The plastic ribbon is a low-loss dielectric, such as polyethylene, with the parallel conductors embedded in the plastic, as illustrated in Fig. 21-24*a*. The 300-ohm twin-lead spacing is approximately  $\frac{3}{8}$  in. between two conductors with No. 20 B & S wire size; 75- and 150-ohm twin lead have closer spacing and therefore more attenuation. Since most television receivers have a balanced input impedance of 300 ohms, the 300-ohm twin lead is convenient for matching the transmission line to the receiver. Because twin lead is unshielded it

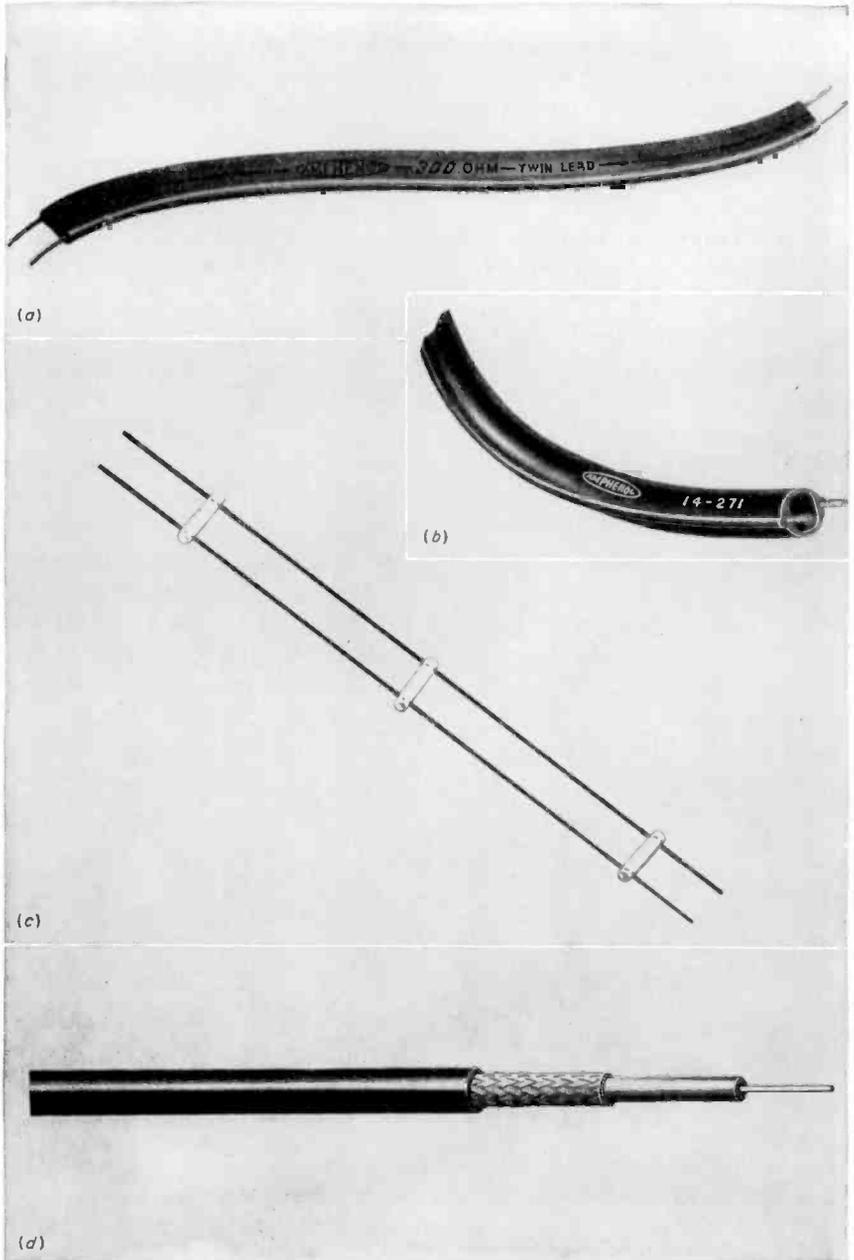


FIG. 21-24. Common types of transmission line. (a) 300-ohm flat, or ribbon, twin lead. (b) 300-ohm tubular twin lead. (*American Phenolic Corporation.*) (c) 450-ohm open-wire line. (*Gonsel Company.*) (d) 72-ohm coaxial line. (*American Phenolic Corporation.*)

should not be run close to power lines, in order to avoid pickup of 60-cps hum voltage, and should be kept away from large metal structures which can alter the balance of the line. Because of unbalance, there may be stray pickup of interfering noise voltages or signals from a nearby unshielded transmission line. The losses of flat twin lead are much greater when the line is wet, increasing from 1.2 to 7.3 db per 100 ft at 100 Mc. Taking a 6-db loss as an example, which is a voltage ratio of  $\frac{1}{2}$ , this means that the signal-voltage output at the end of 100 ft of line is one-half the signal-voltage input; for 200 ft the signal-voltage output is one-fourth the input.

*Tubular Twin Lead.* As illustrated in Fig. 21-24b, in this type of line the two parallel conductors are embedded in the polyethylene plastic formed as hollow tubing, with air the dielectric for most of the inside area. The inside of the tubing is protected from rain, snow, salt, or dirt; as a result, deposits on the outside of the line do not lower the leakage resistance as much as in flat twin lead because of the greater distance between the two conductors through the plastic. Therefore, the tubular twin lead has much less attenuation than flat twin lead when wet, as indicated in Table 21-1. The top end of the tubular line should be sealed, either by heating the plastic dielectric or with water-repellent plastic tape, to prevent rain from wetting the inside; at the bottom end, a  $\frac{1}{4}$ -in. hole cut into the line will provide drainage for condensation that collects inside.

*Open-wire Line.* As shown in Fig. 21-24c, this line is constructed with low-loss insulating spacers between the bare-wire parallel conductors. The open-wire line has the least attenuation because the dielectric is air. The characteristic impedance of open-wire line equals  $276 \log \chi$ , where  $\chi$  is the ratio of the spacing between conductors to the radius of either conductor. One-inch spacing between centers of conductors of No. 12 B & S wire size provides a characteristic impedance of approximately 380 ohms.

*Coaxial Line.* As illustrated in Fig. 21-24d, this line consists of a center conductor in a solid dielectric completely enclosed by a metallic covering, which is often a flexible copper braid. An insulating jacket, usually of vinylite plastic, is molded over the entire line as a protective outside coating. Relatively immune to stray pickup because the outer conductor acts as a grounded shield, the coaxial line can be used in noisy locations or where stray pickup of interfering signals by the line is a problem, when unshielded lines would not be suitable. The coaxial line is unbalanced, with the inner conductor supplying antenna signal and the outer conductor connected to the receiver chassis. Compared with twin lead, coaxial line is stronger and unaffected by moisture but costs more and has higher losses. Coaxial line generally has a characteristic impedance less than 100 ohms, as indicated by the four commercial types of coaxial line listed in Table 21-1. Those with lower losses have larger diameter

and cost more. The RG-59/U coaxial line has an outside diameter of approximately  $\frac{1}{4}$  in.

TABLE 21-1. TRANSMISSION LINES

Type	Characteristic impedance, ohms	Attenuation, db per 100 ft, at 100 Mc
Flat twin lead . . . . .	300*	Dry, 1.2; wet, 7.3
Tubular twin lead . . . . .	300	Dry, 1.1; wet, 2.5
Open-wire line . . . . .	350-600	0.2
Coaxial line:		
RG-8U . . . . .	52	2.1
RG-11U . . . . .	75	1.9
RG-58U . . . . .	53	4.1
RG-59U . . . . .	72	3.7
Shielded pair† . . . . .	225	3.4

\* Flat twin lead is also available in 75- and 150-ohm characteristic impedance.

† Shielded pair has two coaxial conductors in common shield. Is balanced and shielded.

**21-9. Characteristic Impedance.** When the transmission line has a length comparable with a wavelength of the signal frequency carried by the line, the line has important properties other than its resistance. The small amount of inductance of the conductor and the small capacitance between conductors are distributed over the entire length of the line, giving it a distributed inductance and capacitance which can make the line a reactive load, equivalent to lumped reactance in an ordinary circuit. Since the amount of inductance and capacitance is characteristic of the line, depending on its physical construction and the dielectric, the line's characteristic impedance can be defined as

$$Z_0 = \sqrt{\frac{L}{C}}$$

$Z_0$  is the characteristic impedance,  $L$  the inductance per unit length, and  $C$  the shunt capacitance per unit length. The closer the conductor spacing, the greater the capacitance and the smaller the characteristic impedance of the line.

The characteristic impedance is an important property of the line. If the line is terminated with a resistive load equal to the characteristic impedance, all the energy traveling down the line will be dissipated in the load. Maximum power transfer is accomplished and no energy is reflected back into the line. Termination of the line in its characteristic impedance makes it in effect an infinitely long line because of the continuity of the line and load impedance, and there are no reflections.

If the end of the transmission line is open or short or has any other termination not equal to the characteristic impedance, the current and voltage waves on the line will be reflected from the end of the line, and standing waves are set up on the line just as on an antenna. The ratio of current or voltage at a maximum point to the value at a minimum point is defined as the *standing-wave ratio*. This determines the extent to which reflections are present in the line. Standing waves along the line allow it to radiate energy into space, reducing the efficiency of the transmission line. Most important for television receivers, the reflections in the line are equivalent to multipath signals, and if the reflection is over a path more than about 50 ft long it may impair the quality of the picture in the same way as multipath signals picked up by the antenna.

The ratio between the characteristic impedance of the line and the terminating impedance is numerically equal to the standing-wave ratio. This determines whether the transmission line is classified as a tuned line or a nonresonant line. If the line is terminated in its characteristic impedance the standing-wave ratio will be 1, and the line is nonresonant, without standing waves. The nonresonant line is used for television receivers because the length of such a line is not critical. If the line is terminated in a resistance smaller or larger than its characteristic impedance, the standing-wave ratio will be greater than 1 and the length of line required for optimum operation becomes critical. This is why altering the length or capacitance of the line can change the amount of input signal voltage when the transmission line is not matched to the receiver input.

**21-10. Transmission-line Sections as Resonant Circuits.** When the transmission line is not terminated in its characteristic impedance, the values of current and voltage change along the line, the magnitudes varying with a wave motion that is the same as for an antenna. Therefore the impedance for different points on the line varies from maximum at the point of highest voltage on the line to minimum at the point of highest current, as the impedance at any point equals the ratio of voltage to current. This is illustrated in Fig. 21-25 showing transmission lines being used as resonant circuits. Since the action of the line in such an application depends on the existence of reflections, the lines are not terminated in their characteristic impedance but are either shorted or open at the end in order to produce the maximum standing-wave ratio and the highest  $Q$  for the equivalent resonant circuit.

In analyzing the action of the transmission-line sections it should be noted that an open end must be a point of maximum voltage, minimum current, and maximum impedance. Conversely, a shorted end must be a point of maximum current, minimum voltage, and minimum impedance. For each length equal to a quarter wave back from the end of the line the

voltage and current values are reversed from maximum to minimum or vice versa. Intermediate values of impedance are obtained for points along the line between the maximum and minimum points.

In Fig. 21-25 it is shown that a quarter-wave section shorted at the end is equivalent to a parallel-tuned circuit at the generator side because there is a very high impedance across these terminals at the resonant frequency; for a length shorter than a quarter wave the line is equivalent to an inductance. Conversely, the open quarter-wave section provides a very low impedance at the generator side of the line; a length less than a quarter wave makes the line appear as a capacitance. The half-wave sections, however, repeat the impedance at the end of the line to furnish the same impedance at the generator side, with a phase reversal of the voltage and current.

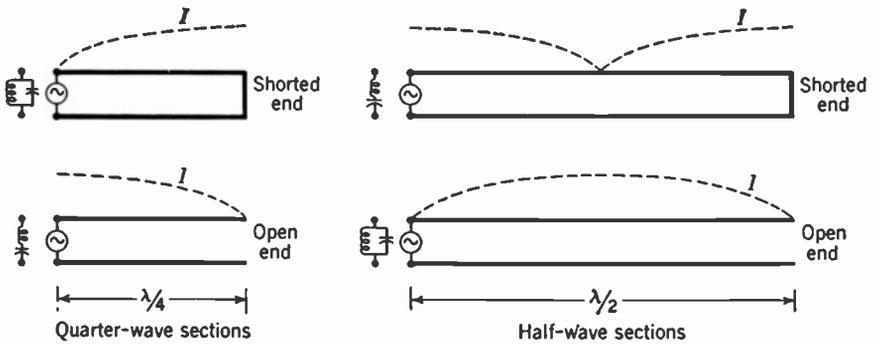


FIG. 21-25. Transmission-line sections as resonant circuits.

Transmission-line sections are often called *stubs*. These can be used for impedance matching, as an equivalent series-resonant circuit for shorting an interfering r-f signal, or to phase antenna signals correctly in multi-element antennas. For phasing sections, a quarter wave produces a  $90^\circ$  change in phase angle between the signal at the input and output ends, while a half-wave section shifts the phase by  $180^\circ$ . To reduce interference, an open stub one-quarter wavelength at the interfering signal frequency can be used. One side is connected across the antenna input terminals on the receiver, while the end of the quarter-wave stub is left open to produce a short at the receiver input one-quarter wave back from the open end. The same results can be obtained with a half-wave stub shorted at the end.

**21-11. Impedance Matching.** In order to obtain maximum efficiency and eliminate reflections in the antenna system that may produce ghosts or reduce the detail in the picture, the impedance of the antenna, transmission line, and receiver input circuit should be matched. Matching impedances means making the impedance of the load circuit equal to the

impedance of the generator producing signal for the load. This is the condition for maximum transfer of power from the generator to its load because no energy is reflected from the load back to the generator. In the receiver antenna system, the transmission line is the load for the antenna and the receiver input circuit is the load for the transmission line. The effect of an impedance mismatch at either end of the transmission line, therefore, is a loss in power transfer and signal level. However, it should be noted that only one end of the transmission line need be connected to its characteristic impedance to eliminate traveling reflections in the line, since there cannot be any reflections from the matched end.

Usual practice is to terminate the transmission line in its characteristic impedance at the receiver end, because the receiver input impedance is designed to be approximately constant for all channels. Therefore, the transmission line used should have a characteristic impedance equal to the receiver input impedance. An impedance match at the receiver end of the line is maintained throughout the television band, as a result, to provide maximum transfer of signal from the transmission line to the receiver and eliminate reflections in the line. Matching the value of the antenna impedance at resonance to the characteristic impedance of the transmission line usually is not critical, since the antenna impedance may vary over a wide range for different television channels as the electrical length of the antenna changes with the operating frequency. An impedance mismatch of 2.5 to 1 results in a 1-db loss of signal. When it is necessary to match impedances, matching sections of transmission line can be employed, or resistance networks are suitable when there is enough signal.

*Quarter-wave Matching Section.* When a quarter-wave section of transmission line with a characteristic impedance  $Z_0$  is neither shorted nor open at the end but has an impedance  $Z_1$  connected across one end, the impedance at the other end  $Z_2$  is

$$Z_2 = \frac{Z_0^2}{Z_1}$$

or

$$Z_0 = \sqrt{Z_1 Z_2}$$

$Z_0$  is the geometric mean of  $Z_1$  and  $Z_2$ . Therefore, if a quarter-wave section of line having an impedance equal to  $\sqrt{Z_1 Z_2}$  is used to couple two unequal impedances  $Z_1$  and  $Z_2$ , the section will provide an impedance match at both ends. Referring to Fig. 21-26, a 35-ohm antenna is used with 300-ohm line, and the matching section has an impedance equal to  $\sqrt{35 \times 300}$ , or approximately 100 ohms. This quarter-wave matching transformer, often called a *Q section*, can be used for matching the antenna impedance to the transmission line.

The length of the quarter-wave section is calculated for the desired frequency from the formula

$$L \text{ (ft)} = \frac{246V}{f(\text{Mc})} \quad (21-3)$$

$V$  is the *velocity factor*, which depends on the velocity of propagation along the line. This is less than the speed of light because of the reduced velocity constant of solid dielectric materials. The velocity factor can be taken as approximately 1 for open-wire lines, 0.85 for twin-lead, and 0.66 for solid-dielectric coaxial line. As an example, a quarter-wave matching stub at 67 Mc using twin lead would have a length of 2.9 ft. The quarter-wave matching section has the advantage of producing an impedance

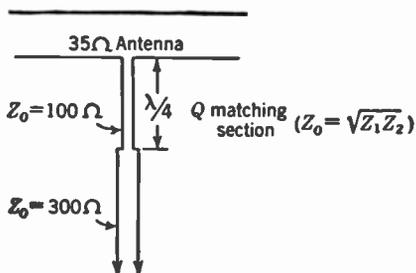


FIG. 21-26. Matching antenna to transmission line by quarter-wave matching section.

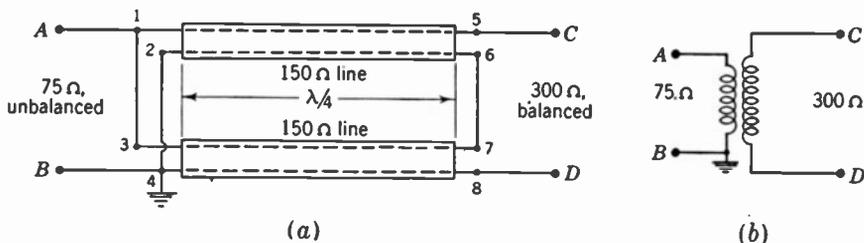


FIG. 21-27. Balancing unit to match between  $75\text{-ohm}$  unbalanced impedance and  $300\text{-ohm}$  balanced impedance. (a) With two  $150\text{-ohm}$  line sections. (b) Equivalent circuit.

match with very little attenuation of the signal, but the section can provide a match only for frequencies at which it is approximately resonant. However, the stub functions in the same way when its length is  $\frac{3}{4}\lambda$  or any odd multiple of a quarter wave, so that it can be cut to serve for both the low- and high-frequency television bands.

*Balancing Unit.* Two  $Q$  sections can be combined to make the balancing and impedance-transforming unit illustrated in Fig. 21-27, which is called a *balun*, for matching between balanced and unbalanced impedances. At one end, the two  $150\Omega$  quarter-wave transmission-line sections are connected in parallel, resulting in  $75\Omega$  impedance across points A and B. Either a or b can be grounded to provide an unbalanced impedance at

the ungrounded point with respect to ground. At the other end the two  $150\Omega$  quarter-wave sections are connected in series to produce  $300\Omega$  impedance between points  $C$  and  $D$ . The quarter wavelength of line isolates the ground point from  $C$  or  $D$ , allowing a balanced impedance with respect to ground. Either side of the balun can be used for input or output. A useful application is matching 72-ohm coaxial line to a 300-ohm receiver input, with the line connected to  $A$  and  $B$ , while  $C$  and  $D$  connect to the antenna terminals on the receiver. The balun can be constructed more compactly in the form of the *elevator transformer* shown in Fig 21-28,

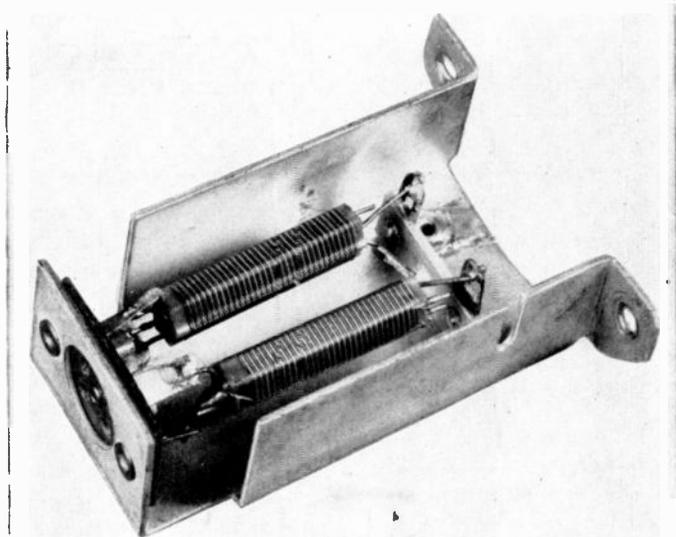


FIG. 21-28. Elevator transformer to match between 72-ohm unbalanced impedance and 300-ohm balanced impedance. (RCA.)

which consists of two bifilar coils equivalent to quarter-wave sections of 150-ohm transmission line.

**Resistance Attenuator Pads.** In cases where excessive antenna signal causes overloading, the signal can be attenuated without introducing any impedance mismatch by using the resistance networks, called *pads*, illustrated in Fig. 21-29. The H pad in  $a$  is for a 300-ohm balanced input and output impedances, and the T pad in  $b$  for a 72-ohm unbalanced arrangement. The input terminals of the pad are connected to the transmission line, while the output terminals connect to the antenna terminals, with the entire pad mounted nearby at the back of the receiver. Carbon resistors of the smallest wattage are used; wire-wound resistors are not suitable because of their inductance. The amount of attenuation

required is usually about 6 db, 10 db, or 20 db, which correspond to voltage-loss ratios of  $\frac{1}{2}$ ,  $\frac{1}{3}$ , and  $\frac{1}{10}$ , respectively.

**Resistance Matching Pad.** When it is desired to attenuate the signal and match a 72 $\Omega$  coaxial line to 300 $\Omega$  receiver input, the resistance pad shown in Fig. 21-30 can be used. The balanced L pad provides an impedance match in both directions, from the 72-ohm transmission line into the pad and from the 300-ohm receiver input circuit into the pad. Regarding the transmission line as a generator supplying signal, it is terminated at points 1 and 2 in a resistance of 82 $\Omega$  shunted by 540 $\Omega$ , equal to approximately 72 $\Omega$ . Looking from the receiver into the pad, the impedance

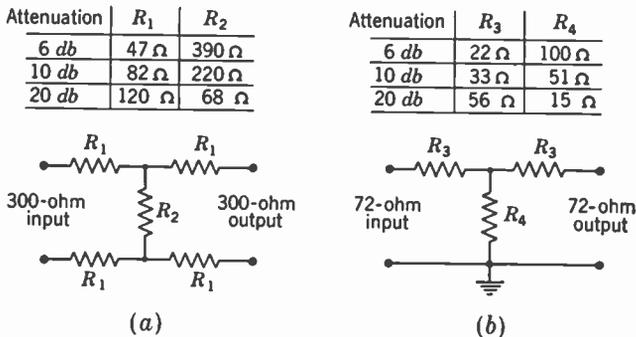


FIG. 21-29. Resistance pads with values for 6 db, 10 db, or 20 db attenuation. (a) Balanced H pad. (b) Unbalanced T pad.

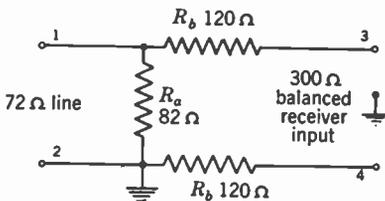


FIG. 21-30. Balanced L pad for matching between 72-ohm unbalanced impedance and 300-ohm balanced impedance.

across the receiver input terminals at points 3 and 4 is 240 ohms in series with the parallel combination of 82 $\Omega$  and 72 $\Omega$ , equal to approximately 300 ohms. Matching from the receiver side into the pad avoids detuning the r-f input circuit, which can change the r-f gain and bandwidth. The use of a resistance matching pad has the advantage of allowing an impedance match that does not vary with frequency. However, the attenuation inserted by the pad in Fig. 21-30 is approximately 11 db. Without the pad, the attenuation caused by the 4:1 mismatch is only 2 db, but the reflections on a long transmission line may result in ghosts in the reproduced picture. To match from a 72 $\Omega$  unbalanced impedance to a 300 balanced impedance without attenuation of the signal, the balun or an equivalent transformer is generally used.

**21-12. Antenna Installation.** The receiver results in producing a picture of good quality can be no better than the signal input from the antenna. The antenna installation, therefore, should be planned carefully in terms of the following three major considerations:

1. Selection, installation, and orientation of a suitable antenna, which will provide enough signal for a snow-free picture, without objectionable ghosts, and be sturdy enough to withstand the wind and ice loading on outdoor antennas.

2. Selection and installation of a suitable transmission line that delivers the antenna signal to the receiver without appreciable attenuation and will be strong enough to last long without breaking.

3. Proper matching of the antenna, transmission line, and receiver impedances to provide maximum signal and eliminate reflections on the line.

For conditions of weak signal when the picture is recognizable but does not have enough contrast, it may be desired to install a *booster*, which is a separate unit containing an r-f preamplifier. With the transmission line from the antenna connected to the booster and its output connected to the receiver input terminals, the booster provides an additional r-f amplifier stage for the receiver.

*Antenna Selection.* In strong signal areas where there are no serious reflections, a simple half-wave dipole located indoors or mounted in the receiver cabinet as a built-in antenna may be suitable, if there are no objectionable ghosts. For most cases, though, an outdoor antenna mounted on the roof provides much better results. In locations of average signal strength within about 25 miles of the transmitter, usually typical of suburban areas just outside the city center for the transmitters, the dipole with reflector has adequate gain and directivity. When channels in different bands must be received, a multiband antenna is necessary. A typical antenna installation in a suburban area is illustrated in Fig. 21-31. To obtain more signal, an array of two antenna bays stacked vertically is commonly used. In fringe areas far from the transmitter, where the field strength is very low, arrays of three or four bays may be used. It should be noted that locations in crowded city areas close to the transmitter but surrounded by tall buildings can have very weak signal with severe reflections.

*Antenna Mounting.* The main requirements for a typical outdoor installation are to mount the antenna near a line of sight, broadside to the transmitter, if possible, and as high as possible. Increased height often results in more antenna signal and may also reduce pickup of external noise and interference. The antenna should be at least 6 ft away from other antennas and large metal objects. It is important to note that changing the antenna placement only a few feet either horizontally or ver-

tically sometimes can make a big difference in the amount of antenna signal because of standing waves of the radio signal in areas where there are large conductors nearby, such as a location inside a steel building or between buildings.

The physical arrangement usually consists of the antenna itself mounted on a sturdy rustproof metal pole 4 to 8 ft high or longer, which is the *antenna mast*. The antenna conductors are insulated from the mast. The mast is clamped in a metal *antenna mounting bracket* that is secured to hold the entire antenna installation. For installations on the roof of an

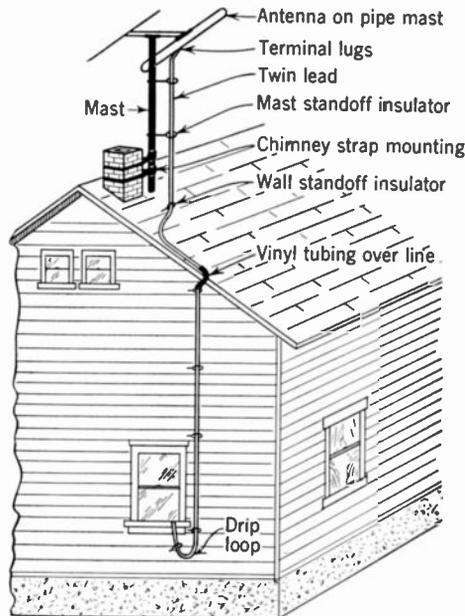


FIG. 21-31. An antenna installation.

apartment house, a pipe-strap type of antenna mounting bracket is usually fastened to a brick wall by screw bolts anchored into lead sleeves inserted in holes drilled in the brick. Clearance of 7 ft or more between the antenna and the roof may be required by local regulations. On private houses, the antenna is often installed by means of metal chimney straps, which encircle the chimney to hold the antenna bracket secure without the need for drilling holes in the brick; or a pipe-strap type of bracket can be fastened by screws into a strong wood beam near the roof. In areas of average signal strength without excessive reflections, the specific location of the antenna mounting usually can be decided on the basis of strong support for the bracket, accessibility, and the most direct transmission-line run. When the mast is more than 10 ft high, approx-

imately, it should be supported by guy wires equally spaced around the mast and anchored as far away as possible. Antenna mountings higher than about 40 ft, which may be necessary in fringe areas, usually require the erection of a steel supporting tower.

It should be noted that in private houses an outdoor type of antenna can often be mounted in the attic with good results. In apartment houses, an antenna on the window frame may be adequate in some locations, if there are no restrictions against window antennas.

*Antenna Orientation.* Before the antenna is clamped tight in its mounting it is rotated to the position that results in the best picture and sound. For suburban locations and fringe areas, this is usually broadside to the transmitter location. All the antennas in one neighborhood generally face the same way. In city areas, however, the antenna may receive more signal over a reflected path from a high building nearby. When several channels must be received in different directions, it is helpful to use an *antenna rotator*, which is a motor-driven arrangement to turn the antenna mast from a remote-control switch at the receiver location.

*Selecting the Transmission Line.* In most cases, 300-ohm twin-lead ribbon line is used for the run from the antenna to the receiver, which generally has 300-ohm input impedance. For long runs, heavy-duty ribbon line should be used for greater strength. Where the picture quality is affected seriously by wet weather, the tubular twin lead is preferable. When there is a problem of pickup of noise and interference by the line, shielded cable should be used. In locations where the transmission line must be run down a corner of an apartment house parallel to many other lines, for instance, shielded line is preferable in order to minimize pickup of interference from the other lines. For long line runs that must be strong and have minimum attenuation, the open-wire type of line can be used.

*The Transmission-line Run.* The run should be as short as possible and any excess line should be cut off in order to minimize attenuation of the signal. If it becomes necessary to add line, it can be spliced, keeping the same spacing between conductors at the connection. The line from the antenna should be run down the mast, vertical to the antenna for at least a quarter wavelength. Horizontal transmission-line runs should be avoided to minimize signal pickup by the line. Standoff insulators support the line at the top and bottom of the mast and about every 6 ft along the run. It is important to clear metal rain gutters or other metal structures and keep the line at least 6 in. away from power and telephone lines. With ribbon line, it is generally twisted about one turn per foot to improve the electrical balance and strengthen the line mechanically so that it does not flap too much in the wind. Where it rubs against the side of the house, the line can be covered with a short length of protective plastic tubing.

The line is preferably run down the back or side of the house to the point of entry near the receiver. For apartment houses, usually a hole is drilled in the window frame to bring the line into the house. On private houses the line can enter the house in the basement and then be routed to the receiver location, or with wood framing the line can enter near the receiver through a hole drilled in an outside wall. Where it enters the house, the line should be given about 4 in. of slack to form a vertical loop, called a *drip loop*, which allows water to drain off. The outside of the entry hole is covered with calking compound or plastic wood for waterproofing.

*Grounding.* To prevent the accumulation of static charge on the mast, it can be grounded by connecting a heavy ground wire to a cold-water pipe or to a metal stake driven into the earth. In some areas, local regulations require grounding the mast. When the antenna is a high point in the area a lightning arrestor should be installed along the transmission-line run. The lightning arrestor provides a high-resistance discharge path that prevents static charge from accumulating on the antenna; and, in case of lightning striking the antenna, the arrestor is a short circuit to ground that protects the receiver. Indoors, the lightning arrestor is usually mounted by a grounding strap on a cold-water pipe. For use outdoors, a weatherproof type of arrestor can be installed on a grounded mast.

*Impedance-matching Applications.* It is best to use transmission line having a characteristic impedance the same as the input impedance of the receiver, since then a match at one end of the line is approximately maintained for all channels. This generally means using 300-ohm line in order to match the receiver input. At the antenna, it is preferable not to match to a higher value of line impedance when the antenna is used for several channels and is cut for the lowest frequency. The antenna impedance then increases with frequency and the mismatch allows a broader frequency response in terms of antenna signal delivered at the receiver. If the antenna is used for only one channel, however, more signal can be obtained by matching the antenna to the transmission line, using a quarter-wave matching section. When 72-ohm coaxial cable is used in order to have a shielded transmission line, it can be matched to 300-ohm receiver input by means of a balun.

When the antenna is used for several channels, it is possible to increase the signal strength considerably for a weak station by tuning out the antenna's reactance for a closer impedance match on this particular channel. A procedure that can be used conveniently with unshielded twin lead is as follows: With the weakest channel tuned in, slide your hand along the line for a few feet back from the receiver input, while watching the picture. If the picture strength is increased at some points and

decreased a quarter wavelength back on the line, this indicates a reactive line termination. Find the spot where the hand capacitance produces the strongest picture and connect a small capacitance of about 2 to 20  $\mu\mu\text{f}$  across the line at this point. The capacitance can be a small trimmer condenser, a piece of metal foil wrapped around the transmission line, or a section of line less than a quarter wavelength open at the end to form a capacitive stub. Reception on all channels should be checked after this is done because the capacitance can affect more than one channel. When the hand capacitance decreases the picture strength at all positions on the line, this indicates the impedance match is correct.

**Indoor Antennas.** Where the signal is strong and there are no excessive reflections, half-wave dipole structures constructed for convenience in mounting indoors can be used, generally with good results on one, two, or three channels not too widely separated in frequency. Some common types include telescoping rods in a dipole assembly of adjustable length, which is usually placed on the receiver cabinet, a wide-band triangular dipole made of metal foil or screening, and a folded dipole formed from a half-wave section of 300-ohm twin lead as shown in Fig. 21-32. The

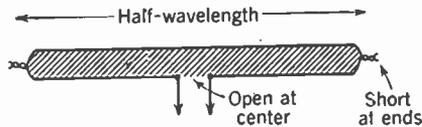


FIG. 21-32. Folded dipole antenna made of half-wave section of 300-ohm twin lead.

exact location of an indoor antenna is usually determined by trial and error to find the spot that provides the most signal.

**Built-in Antennas.** Practically all television receivers have an antenna mounted in the receiver cabinet, which can provide adequate results in strong signal locations without ghosts. The built-in antenna is generally a folded dipole made of twin lead mounted around the cabinet frame, or a short triangular dipole constructed of metal foil mounted on the underside of the cabinet top. When the antenna is too short because of the restricted physical length it can be made electrically longer by inserting series inductance. If the built-in antenna is used, the location and orientation of the receiver cabinet can make a big difference in the amount of antenna signal. When it is not being used, the built-in antenna should be disconnected from the receiver input terminals.

**21-13. Multiple Installations.** When multiple receiver outlets from a common antenna system are needed, there are three main requirements:

1. The amount of antenna signal available for each receiver must be at least several hundred microvolts for a good picture without excessive snow. In addition, the signal from the antenna distribution system must be much greater than the amount of signal picked up directly without the

antenna. Otherwise, there can be ghosts in the picture caused by duplicate signals.

2. Isolation should be provided between receivers to attenuate the local oscillator signal, which can produce beat-frequency interference patterns consisting of diagonal bars in other receivers on the common distribution line.

3. The impedance of the transmission line should be matched at the receiver end. This is important with a long line, to prevent ghosts caused by reflections on the line, or when the impedance match is necessary for maximum signal to provide a satisfactory picture.

Considering these requirements, it may be useful to note that several receivers can simply be connected in parallel to one transmission line, with adequate results if there is enough antenna signal, local oscillator interference is no problem, and the transmission line is less than about

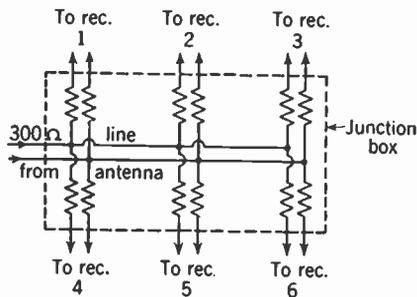


FIG. 21-33. Parallel distribution system for feeding signal from one antenna to multiple receivers from common junction box. All resistors are 750 ohms for six receivers with 300-ohm input impedance.

50 ft long. If the multiple installation can use only one line at a time, a switch with positions for each line gives good results. Double-pole switches are available for two or more positions in the form of a knife switch, rotary wafer switch, or toggle switches in a junction box. *Distribution transformers*, or *set couplers*, are available for connecting two, three, or four receivers simultaneously to one line. These usually provide an impedance match but without decoupling local oscillator signal from individual sets. For an installation that also isolates each receiver a distribution line with resistance pads can be used, or a more elaborate amplifier distribution system may be necessary.

*Distribution Line with Resistance Pads.* The distribution system illustrated in Fig. 21-33 is economical and requires little maintenance, but a strong antenna signal is needed because of the attenuation produced by the resistance pads. This is a parallel distribution system from 300-ohm line to six receivers having 300 ohms input impedance. The pads isolate each receiver from the distribution line. The value of 750 ohms for all

the resistors provides six 1,800-ohm parallel paths, including the 300-ohm input impedance for each receiver, to match the 300-ohm transmission line. The antenna signal available for each receiver is  $300/1,800$ , or one-sixth, the total signal. With fewer receivers, the resistors are 600, 450, 300, or 150 ohms for five, four, three, or two sets, respectively. More than six receivers should not be connected to one antenna, even if enough signal is available, because direct pickup can then be as much as the antenna signal from the distribution line. In order to maintain the impedance match when the individual receivers are either on or off, it may be necessary to have a switch and 300-ohm dummy load resistor across each branch line to replace the receiver input impedance when the receiver is not operating.

*Amplifier Distribution System.* Where there is not enough antenna signal to operate several receivers, or when many receivers must be fed

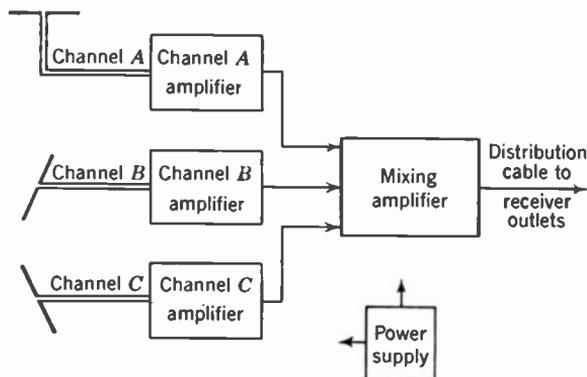


FIG. 21-34. Illustrating a master antenna distribution system using amplifiers.

from a single antenna system, a more elaborate master antenna distribution system using r-f amplifiers is necessary. If individual antennas must be used for different channels, the amplifiers also provide a means of mixing the signals for common distribution. As illustrated in Fig. 21-34, a separate amplifier can be used for each antenna and all signals coupled to a mixing amplifier where a common load impedance provides output for all the channels. The gain usually can be adjusted for each individual channel, to provide uniform output of several millivolts on all channels. A coaxial-cable distribution line couples the output to as many as 30 receiver outlets on the line. Additional distribution cables can be used from the output amplifier for more receivers if necessary. At each outlet a resistance pad isolates the receiver and matches its input impedance to the distribution line.

*Community Television.* In fringe areas where reception of the nearest television broadcast stations requires elaborate and costly antenna struc-

tures, a master antenna and distribution system is often used to supply signal by coaxial cable to subscribers who pay for this private service. Arrays of high-gain receiving antennas are usually mounted on tall towers located at a high point in the terrain of the community, to pick up antenna signal from distant stations. The signals are amplified and combined in a distribution amplifier system at the antenna site. If desired, high-band channels can be converted to low-band frequencies in order to have less attenuation in the distribution line. A coaxial-cable line feeds the receivers in the community.

**21-14. Troubles in the Antenna System.** Since the antenna and transmission line are part of the r-f signal circuits, defects in this section of the receiver can cause troubles on individual channels, while other channels are normal. Also, the trouble symptom of weak picture with excessive snow, either on some channels or on all channels, is often caused by insufficient antenna signal. It is important to note that, when the transmission line from the antenna is open, reception may be normal on some channels, with the open line functioning as the antenna, while there is insufficient signal only on individual channels. The transmission line can be checked for continuity by an ohmmeter connected across the receiver end. With a folded dipole antenna, the line should have practically zero resistance to indicate continuity; with antennas that do not have a closed circuit, the line can be temporarily shorted or a resistor connected across the antenna end to check continuity. A transmission line that is intermittently open will produce flashing in the picture, especially in windy weather. Flashing can also be caused by static discharge of the antenna when there is no lightning arrestor. A transmission line that flaps in the wind can cause the picture intensity to fade in and out. For the u-h-f channels, vibration of the antenna also may produce fading of the picture. Clean tight contacts at the antenna are especially important for the u-h-f channels to avoid intermittent connections and loss of signal caused by leakage.

#### REVIEW QUESTIONS

1. What is the length of a half-wave dipole antenna at 54 Mc?
2. What is the definition of antenna gain?
3. A ghost caused by multipath reception by the antenna is displaced 2 in. to the right from the main image, on a raster 20 in. wide. What is the difference in length between the direct and reflected signal paths?
4. Give two advantages of a dipole with parasitic reflector, compared with a dipole alone.
5. Give two differences between a dipole with reflector and dipole with director.
6. Give two advantages and one disadvantage of a Yagi antenna.
7. Describe briefly how the directional pattern of a dipole with parasitic reflector, cut for channel 2, changes from channel 2 to channel 6 to channel 13.
8. Describe briefly five types of television receiving antennas, giving one important feature of each.

9. Draw a diagram illustrating vertical stacking of two antenna bays each using the double-V end-fire array. Show the transmission-line connections.

10. Why is a transmission line terminated in its characteristic impedance?

11. Compare the following transmission lines, giving the characteristic impedance and one additional feature of each: flat twin lead, tubular twin lead, coaxial line.

12. What is the characteristic impedance of a quarter-wave matching section coupling a 14-ohm antenna to 300-ohm transmission line? Calculate the length of the matching section, cut for channel 4, using flat twin lead.

13. Draw the diagram of an H pad connected between 300-ohm twin lead and 300-ohm balanced receiver input, giving resistance values to attenuate the signal voltage by one-half.

14. Draw the diagram of a parallel distribution system, with values of the resistance pad, for four 300-ohm receivers supplied from a 300-ohm transmission line, terminating the line in its characteristic impedance.

15. Give one advantage and one disadvantage of a master antenna distribution system using amplifiers.

16. Describe briefly at least five main parts of an antenna installation job.

17. Why does a weak picture with snow indicate that the trouble may be insufficient antenna signal? Give one other cause of this trouble symptom.

## CHAPTER 22

### THE FM SOUND SIGNAL

Since the sound associated with the television program is transmitted as an FM signal, the television receiver includes FM receiver circuits for the sound signal. After conversion to the intermediate frequencies the sound and picture signals are separated and the FM sound signal is coupled to the sound i-f amplifier. The fact that the sound carrier signal is frequency-modulated does not alter the heterodyning process in producing the sound i-f signal, as the original frequency swing is maintained about a lower center frequency equal to the i-f sound carrier. The i-f section for the FM sound signal is then the same as in a typical FM receiver.

**22-1. FM Receivers.** Figure 22-1 illustrates the i-f circuits of an FM receiver. The frequency-modulated i-f signal is obtained from the r-f tuner in FM receivers for the FM broadcast band of 88 to 108 Mc. In

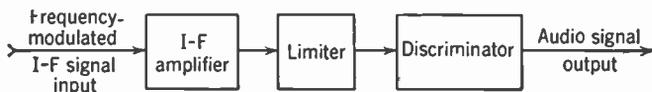


FIG. 22-1. I-F section of an FM receiver.

television receivers the frequency-modulated i-f signal is the associated sound for the selected channel. This FM signal for the sound i-f section is obtained from the sound take-off circuit in the plate of the mixer stage in the front end or from a common i-f stage, for split-sound television receivers. In intercarrier-sound television receivers, the 4.5-Mc sound take-off circuit in the video detector or video amplifier section couples the 4.5-Mc frequency-modulated sound signal to the 4.5-Mc sound i-f section. The main requirements of the i-f section for an FM signal are (1) the bandwidth of the i-f circuits must be broad enough for the relatively wide-band FM sound signal; (2) amplitude-rejection circuits are included in the i-f circuits to reduce the effects of AM interference in the FM signal, as illustrated by the limiter stage in Fig. 22-1; (3) an FM detector is necessary to convert the frequency variations in the FM signal to the desired audio signal voltage for the audio amplifier, as illustrated

by the discriminator stage in Fig. 22-1. In addition, the audio section should be capable of amplifying a high-fidelity signal, with a wider audio-frequency range, greater volume range, and less noise, compared with AM receivers for the standard broadcast band.

*I-F Bandwidth.* In receivers for the FM broadcast band, the maximum frequency swing is  $\pm 75$  kc, requiring an i-f bandwidth of at least 150 kc. The maximum frequency swing for the FM sound signal in television is  $\pm 25$  kc, which requires an i-f bandwidth of 50 kc or more. While these values may seem high compared with the 4- to 10-kc bandwidth in receivers for the AM standard broadcast band, the required bandwidth is easily attained because it is a relatively small percentage of the intermediate frequency. In receivers for the FM broadcast band, the intermediate frequency generally used is 10.7 Mc, and the bandwidth of 150 kc is only 1.4 per cent of center frequency. For the 4.5-Mc sound i-f carrier in intercarrier-sound receivers, a bandwidth of 50 kc is 0.9 per cent of center frequency. In AM radio receivers a bandwidth of 5 kc is 1.1 per cent of the 455-kc intermediate frequency. In all three cases, therefore, the bandwidth is only a small percentage of the center resonant frequency.

*AM Rejection.* Eliminating amplitude modulation of the FM signal is possible in the FM receiver because the desired signal is a variation in frequency and limiting the amplitude does not distort the FM signal variations. Referring to Fig. 22-1, the last i-f amplifier is operated as a limiter stage to reject amplitude modulation in the FM signal. The limiter is similar to the preceding i-f stages but the d-c operating potentials in the tube enable the stage to function as a saturated amplifier. Constant amplitude of the i-f output signal is maintained, as a result, over a wide range of variations in the amplitude of the input signal. Since the greatest effect of an interfering signal is to change the amplitude of the desired signal, the AM rejection eliminates interference that changes the amplitude of the FM signal, without altering the signal variations in frequency. The ability of an FM receiver to reduce the effects of noise and interference in the signal, therefore, depends upon the use of AM rejection circuits in the receiver. Eliminating the undesired amplitude variations is accomplished by using a limiter, or with an FM detector circuit that is insensitive to amplitude modulation in the FM signal.

*FM Detection.* Following the limiter in Fig. 22-1 is the discriminator stage, which is a balanced double-diode detector having the functions of rectifying and filtering the FM signal to recover the desired audio voltage. The discriminator is the second detector in the FM receiver, capable of converting the frequency changes of the FM signal into corresponding variations in audio signal voltage. Any amplitude variations caused by

undesired amplitude modulation are removed by the limiter, which supplies an FM signal of constant amplitude for the FM detector. The discriminator then detects the constant-amplitude FM signal to produce the desired audio voltage without any interfering AM signals. Instead of the limiter-discriminator combination, however, many FM receivers use an FM detector that does not produce audio output voltage for undesired amplitude variations in the FM signal and, therefore, a limiter stage is not necessary.

*Audio Section.* The audio output of the FM detector is amplified in the audio amplifier with enough gain to drive the loudspeaker, as in a typical AM radio receiver. If the high-fidelity characteristics of the FM sound signal are to be used to advantage, however, the audio system of the FM receiver should be of better than average quality. It should be capable of reproducing the audio modulation range of 50 to 15,000 cps, which is used in FM transmission for both television sound and commercial FM broadcasting. In many cases, the high-fidelity requirements of the audio system are not completely fulfilled because of economy and space requirements, especially in table-model receivers. However, all other advantages of the FM system may still be retained.

**22-2. Basic Requirements for FM Detection.** The requirements for FM detection are much the same as in the second detector of an AM receiver. A rectifier, usually a diode, is needed to provide a rectified signal that varies in magnitude with the amplitude variations of the a-c signal voltage applied to the diode, exactly as in AM detection. The only additional requirement for FM detection is a circuit arrangement that enables the frequency variations of the FM signal to appear as variations in amplitude of the signal applied to the detector for rectification. This can be accomplished by means of tuned *LC* circuits, since these are resonant at one frequency and the output across the tuned circuit varies in magnitude for different input signal frequencies.

*Slope Detection.* The sloping side of the i-f response curve can provide a varying amplitude response for different signal frequencies, as illustrated in Fig. 22-2, and the resultant amplitude variations of the i-f signal can be coupled to a rectifier to detect the FM signal. Reviewing the characteristics of an FM signal, it should be remembered that the amount of frequency departure from the carrier center frequency varies with the amplitude of the audio modulating voltage during the audio cycle. When the carrier frequency of the FM signal falls on a sloping side of the i-f response curve, the frequency variations of the carrier signal are converted to equivalent amplitude variations because of the unequal responses above and below the carrier frequency. Therefore, the i-f output is made to vary in amplitude at the audio rate, in addition to its continuously changing frequency, and the resultant amplitude variations can be

coupled to an AM detector to recover the audio voltage. With adequate filtering of the rectifier output, the voltage across the diode load resistor is the desired audio voltage because it varies in amplitude with the amount of input voltage, which in turn, varies with the amount of frequency departure. The frequency of the audio output voltage is the same as the original modulating audio signal, since both are equal to the rate at which the FM signal goes through its frequency swing.

Although slope detection is seldom used because it cannot employ AM rejection circuits, the principle is important because it illustrates the two basic requirements of FM detection: (1) converting the FM signal into equivalent amplitude variations in the i-f signal input to a diode rectifier

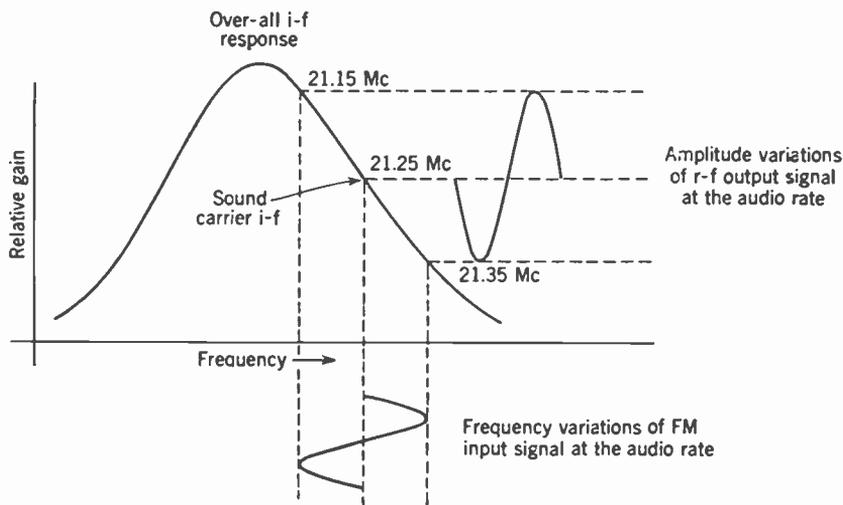


FIG. 22-2. Conversion of frequency variations to amplitude variations by slope detection of an FM signal. Both the input and output are r-f signals, although only the audio variations are illustrated.

and (2) rectifying the audio variations in i-f signal amplitude. In addition, slope detection is the reason why an FM signal can produce audio output in an AM receiver. As an example, the FM sound in the AM picture channel of the television receiver can be detected in this way, producing an interfering audio signal in the video amplifier and the resultant interference pattern of sound bars in the picture.

**22-3. Triple-tuned Discriminator.** A simple type of discriminator circuit that can be used for FM detection is shown in Fig. 22-3. Two diodes are used, with balanced push-pull input for the i-f signal coupled from a limiter stage, and a balanced cathode load circuit for the detector output. Each diode is a rectifier. When positive signal voltage is applied between plate and cathode of  $V_1$ , plate current flows from cathode to

plate, through the transformer secondary, and back to the cathode through the diode load resistor  $R_1$ .  $C_1$  acts as an r-f filter, just as in the ordinary AM detector. With positive signal voltage applied to  $V_1$ , therefore, the rectified current produces an  $IR$  drop across  $R_1$ , the cathode side being positive as indicated in the diagram. In the same way, when signal voltage is applied to  $V_2$ , plate current flows from cathode to plate, through the secondary  $L_2$ , and back to cathode through  $R_2$ . This produces an  $IR$  drop across  $R_2$  with the cathode side positive, resulting in a voltage across  $R_2$  of opposite polarity from the rectified voltage across  $R_1$ . Assuming that the i-f signal voltages applied to the two diodes are of equal magnitude, for example, 5 volts, the voltages across  $R_1$  and  $R_2$  will be equal, with a value close to 5 volts for each. However, the output voltage is taken from the cathode load resistors between chassis ground and the point marked  $A$  in the diagram, including the voltages across both  $R_1$  and  $R_2$ . These are in series opposition, and the net output voltage available at  $A$  with respect to the chassis ground is equal to the difference between the two rectified voltages. If the voltages across the diode load resistors are equal, the net output voltage will be zero. When equal i-f signal voltages are applied to the two diode rectifiers, therefore, the rectified output of each is the same and the net output voltage is zero.

*Audio Output Voltage.* If the signal voltages applied to the two diodes vary, for example, the plate voltage for  $V_1$  increasing from 5 volts to 7 volts while the plate signal for  $V_2$  decreases correspondingly to 3 volts, then  $V_1$  will conduct more plate current, producing a larger  $IR$  drop across  $R_1$ . Diode 2 conducts less current to produce a smaller voltage drop across  $R_2$  than was obtained with 5 volts applied to the diodes. The voltage across  $R_1$  rises to a value close to 7 volts while the voltage drop across  $R_2$  decreases to approximately 3 volts. The voltages are unbalanced now, and the net output voltage at point  $A$  is the difference between 7 and 3 volts, or 4 volts, positive with respect to ground because the positive voltage is greater. When the applied signal voltages are reversed so that 7 volts is applied to  $V_2$  and 3 volts to  $V_1$ , the voltage drop across  $R_2$  is 7 volts with only 3 volts across  $R_1$ . The circuit is again unbalanced with an output voltage of 4 volts, which is now negative with respect to ground.

In summary, then, when equal signal voltages are applied to both diodes, the voltages across  $R_1$  and  $R_2$  are equal and the net output voltage is zero. As the plate signal voltage for  $V_1$  increases, while decreasing for  $V_2$ , a positive output voltage is obtained. Conversely, when the plate voltage for  $V_2$  increases while decreasing for  $V_1$  the net output voltage is negative. Therefore, if the FM signal alternately increases the plate signal voltage for one diode while decreasing the signal applied to the other diode, in step with the frequency variations in the FM signal pro-

duced by modulation, the output voltage of the discriminator will be a reproduction of the original audio modulation.

*I-F Input Voltage.* The frequency swings of the FM signal are converted to corresponding amplitude variations in the signal applied to the two diodes by means of the triple-tuned coupling circuit for the i-f signal.

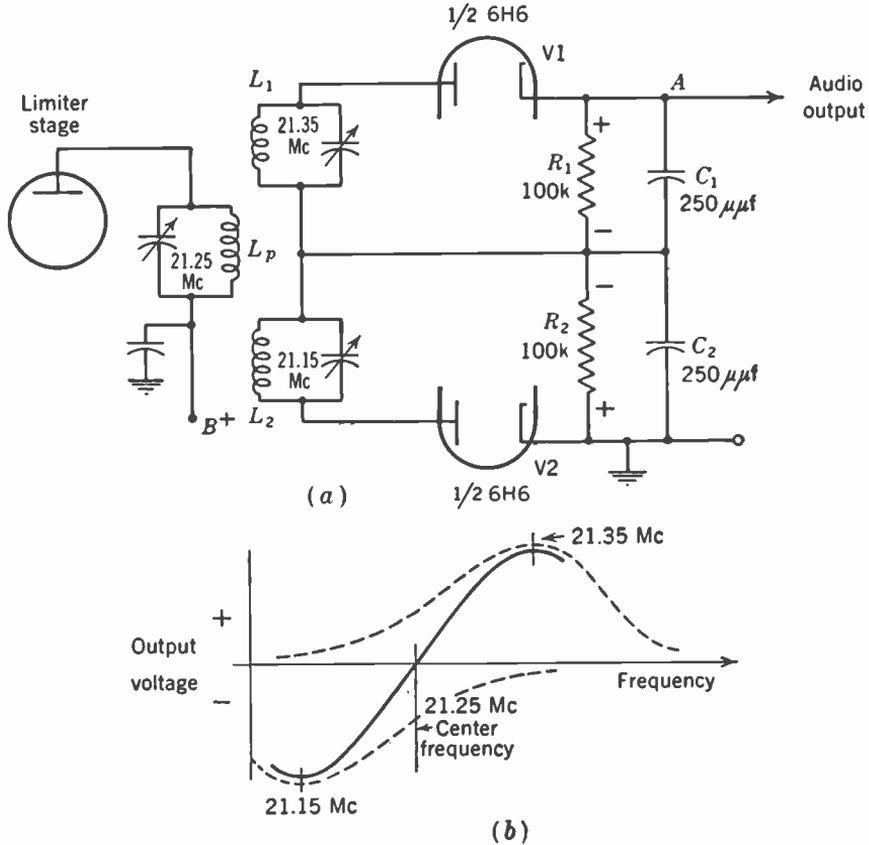


FIG. 22-3. Triple-tuned discriminator for an i-f center frequency of 21.25 Mc. (a) Circuit. (b) Response curve.

As noted in Fig. 22-3 for a sound intermediate frequency of 21.25 Mc, the primary circuit is tuned to this center frequency, with one secondary being tuned 100 kc above center frequency and the other secondary tuned below center by the same amount. Consider the case of FM signal output from the limiter when the instantaneous frequency is at rest or center frequency. The primary of the triple-tuned coupling circuit is resonant at this frequency for maximum output, but each of the secondary circuits is resonant at a frequency 100 kc removed from the center frequency.

Neither secondary tuned circuit is resonant at this instantaneous signal frequency, but both are off resonance by the same amount. The signal voltage developed across both secondary tuned circuits will be relatively small and have the same magnitude. Therefore, equal signal voltage is coupled to both diodes when the instantaneous frequency of the FM signal is at center frequency, the rectified output voltages are the same, and the net output voltage is zero.

When the instantaneous frequency of the FM signal deviates above center frequency, the secondary tuned circuit  $L_1C_1$  develops a greater signal voltage because the signal frequency is closer to the resonant frequency of this tuned circuit. The secondary voltage developed across  $L_2C_2$  is now less than at center frequency because the signal frequency is further removed from its resonant frequency. With more signal voltage applied to  $V_1$  and less to  $V_2$  for a frequency departure above rest frequency, diode 1 provides a greater rectified output voltage than diode 2 and a net output voltage of positive polarity is obtained. When the instantaneous frequency of the FM signal deviates below center frequency, more signal voltage is developed across the tuned circuit  $L_2C_2$  and a negative output voltage is obtained. Thus, with the primary tuned to center frequency and having a frequency response broad enough to provide uniform response for the total frequency swing of the FM signal, the two secondary circuits alternately provide more or less i-f signal voltage for the two diode rectifiers, and the amplitude variations correspond to the frequency variations in the FM signal.

*Discriminator Response Curve.* The typical S-shaped discriminator response curve is shown in Fig. 22-3b for this balanced detector. The output at center frequency is zero. For frequencies above center, the output voltage is positive and increases progressively for an increasing swing away from center frequency. Similarly, the output voltage is negative when the input signal is below center frequency. Note that the composite response curve is the net resultant of the sloping responses of the two individual secondary tuned circuits, with each providing output of opposite polarity because of the balanced arrangement of the diodes. The positive and negative peaks on the discriminator response curve occur at the resonant frequency for each of the two secondary tuned circuits. The separation between peaks should be great enough to accommodate the total frequency swing over a linear portion of the discriminator response curve, which is linear for about one-half the distance between peaks.

The primary can be tuned to center frequency and the two secondary circuits tuned above and below center frequency, with adequate separation between peaks, to provide linear detection for faithful recovery of the audio modulation. However, tuning adjustments may be difficult

because of mutual coupling between the tuned circuits in this triple-tuned arrangement.

**22-4. Phase-shift Discriminator.** Figure 22-4 shows the circuit of a balanced discriminator using only two tuned circuits, both of which are tuned to center frequency. This is the most frequently used discriminator circuit, generally called the *phase-shift*, or *Foster-Seeley*, discriminator. A twin diode with separate cathodes such as the 6H6 or 6AL5 is used for the rectifier. The cathode resistors  $R_1$  and  $R_2$  are equal, as are the r-f

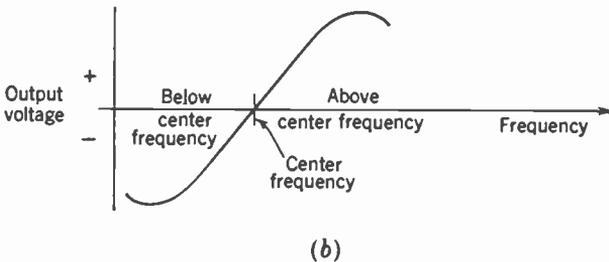
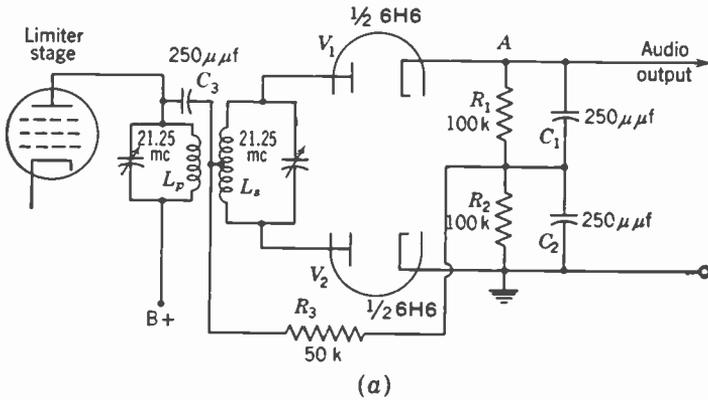


FIG. 22-4. Phase-shift discriminator for an i-f center frequency of 21.25 Mc. (a) Circuit. (b) Response curve.

bypass condensers  $C_1$  and  $C_2$ .  $L_p$  and  $L_s$  are the primary and secondary of the i-f transformer used to couple the signal inductively from the limiter stage. Both primary and secondary are tuned to center frequency. In addition to the transformer coupling, signal voltage from the primary is coupled to the center tap of the secondary by means of the coupling condenser  $C_3$ , which has negligible reactance at the signal frequency. When the plate circuit of the limiter stage is shunt fed the primary voltage can be connected directly to the secondary center tap, omitting  $C_3$ , since its only function is to couple signal and isolate the plate-supply voltage from the discriminator circuit. The load resistor  $R_3$  provides a d-c return

for diode plate current while maintaining a load impedance across the primary. If the secondary center tap were returned directly to the cathode circuit, at the junction of  $C_1$  and  $C_2$ , the primary circuit would be shorted to ground through r-f bypass condenser  $C_2$ . An r-f choke can be used instead of the load resistor  $R_3$ . Or, the load impedance  $R_3$  can be eliminated and the secondary center tap returned directly to the junction of  $R_1$  and  $R_2$ , but only one r-f bypass condenser is connected across both cathode resistors so that the primary signal will not be shorted to ground.

*Balanced Detection.* The two diode rectifiers in the phase-shift discriminator are balanced in exactly the same way as in the triple-tuned discriminator. When equal signal voltage is applied to the two diodes, the rectified output is the same for each cathode load resistor and the net output voltage across the two equal series opposing voltages is zero. As the applied voltage for each of the diodes alternately increases, while decreasing for the other, an output voltage is obtained of either positive or negative polarity.

*Coupling Circuit.* As the FM signal varies in frequency, the i-f signal voltage applied to the diodes is made to vary in amplitude by means of the coupling arrangement between the limiter stage and discriminator. Note that signal from the limiter stage is simultaneously coupled to the discriminator in two ways: one by induction across the i-f transformer, and the other by the direct coupling obtained with  $C_3$ , which is independent of the mutual induction between primary and secondary of the i-f transformer. The voltage across the secondary, induced by the primary current of the discriminator transformer, is applied to the diode plates in push-pull by means of the center-tap return to cathode. The directly coupled voltage is the primary voltage itself and is applied in parallel to the two diodes, both of which are connected to the primary at the same point. Therefore, the i-f signal voltage for the two diodes is the resultant sum of an induced secondary voltage that is applied in push-pull and the primary voltage applied in parallel to the two diodes.

The secondary voltage  $E_s$  is  $90^\circ$  out of phase with the primary voltage  $E_p$  at the resonant frequency and varies in phase above and below  $90^\circ$  as the applied signal varies in frequency above and below the resonant frequency. This phase relation results because the secondary is tuned to resonance, as explained more fully in the next paragraph. Since the primary and secondary of the coupling transformer are tuned to the i-f carrier frequency, which is center frequency for the FM signal, the secondary and primary voltages are  $90^\circ$  out of phase with each other at center frequency. Both these voltages are simultaneously applied to the diode rectifiers, with the phase relations illustrated by the vector diagrams in Fig. 22-5. In *a* is shown the case for resonance, with the FM signal at center frequency. When the secondary tuned circuit is resonant,  $E_p$  is

90° out of phase with  $E_s$ , which is divided into two equal voltages of opposite polarity,  $E_{s1}$  and  $E_{s2}$ , by the center-tap connection. The voltage  $e_1$  is the resultant of its two components  $E_{s1}$  and  $E_p$ , and is the combined signal voltage applied to diode 1 for rectification. Similarly, the voltage  $e_2$  is the resultant of  $E_{s2}$  and  $E_p$  and is the voltage applied to diode 2. The two applied voltages  $e_1$  and  $e_2$  are equal in this case because of the identical phase angle between their components. With equal amplitudes of signal voltage applied to the two diodes, the rectified output of each is the same and the net output voltage from the balanced cathode circuit is zero. In the vector diagram of *b*, the i-f signal is above center frequency and  $E_s$  is less than 90° out of phase with  $E_p$  because the secondary is tuned below the signal frequency. As a result,  $e_1$  increases as the component voltages of  $e_1$  come closer to being in phase, while  $e_2$  decreases.

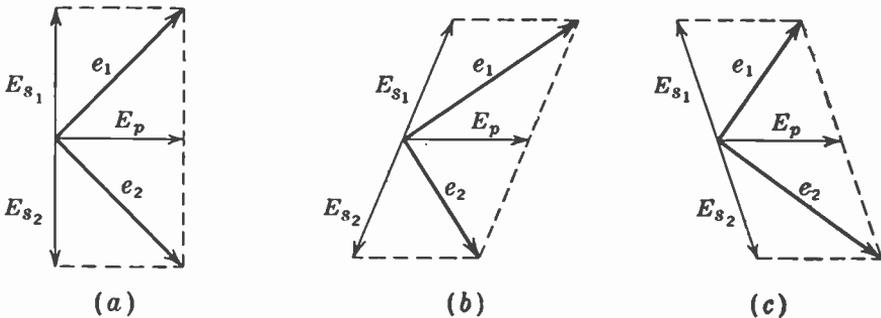


FIG. 22-5. Vector diagrams for the phase-shift discriminator as the FM signal varies above and below the i-f center frequency. (a) At center frequency; resonance; 90° phase angle between  $E_p$  and  $E_{s1}$ ;  $e_1 = e_2$ . (b) Above center frequency; above resonance; 70° angle between  $E_p$  and  $E_{s1}$ ;  $e_1 > e_2$ . (c) Below center frequency; below resonance; 110° angle between  $E_p$  and  $E_{s1}$ ;  $e_1 < e_2$ .

More signal is applied to  $V_1$  than to  $V_2$  and a positive output voltage is obtained. For the case of a signal frequency below center frequency,  $E_s$  is more than 90° out of phase with  $E_p$ , as shown in the vector diagram of *c*. The signal voltage  $e_2$  is then greater than  $e_1$  and the rectified output voltage is negative.

**Phase Relations.** Usually, the voltage across a transformer secondary is assumed to be 180° out of phase, or in phase, with the primary voltage, depending on the direction of winding and the connections made. However, this is a simplification that does not take into account the effect of the secondary load. When the secondary is resonant as it is here, the voltage across the secondary caused by induction is 90° out of phase with the primary voltage, and this 90° phase relation varies as the frequency of the input signal varies from the resonant frequency.

The first point to be noted is that the voltage induced in the secondary is 90° out of phase with the current through the primary coil. Consider

the sine wave of current  $I_p$  shown in Fig. 22-6 flowing through the primary coil. As this current changes in value, voltage is induced across the secondary coil because of the changes in flux caused by the varying primary current. This induced voltage  $e_i$  is greatest when the most lines of force per second are cutting the secondary. It is seen in Fig. 22-6 that at point  $a$  the primary current  $I_p$  is changing rapidly, since it is rising rapidly. Therefore, the magnitude of the induced voltage is large here. At point  $b$  the primary current is at its maximum value but is not changing. Therefore there is no change of flux and no induced voltage. Plotting values of induced voltages for changes in primary current, Fig. 22-6 shows that the induced voltage is  $90^\circ$  out of phase with the current through the primary coil, since  $E_i$  has its maximum values for minimum values of  $I_p$ . By convention, windings are taken so that the induced voltage is of negative polarity and  $E_i$  is shown lagging  $I_p$  by  $90^\circ$ .

The induced voltage  $E_i$  is not the secondary output voltage but must be considered as a generator in series with the secondary coil, as shown in

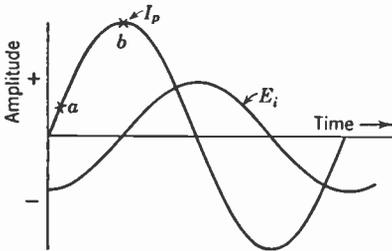


FIG. 22-6. The induced voltage  $E_i$  varies in magnitude with the change in primary current and therefore is  $90^\circ$  out of phase with  $I_p$ .

the equivalent secondary circuit of Fig. 22-7, since any current that flows in the secondary because of this induced voltage must flow through the coil. The inductance of the secondary coil is  $L_s$ ,  $C$  is the capacitance across the secondary inductance, and  $R_s$  is the small resistance of the coil.  $E_s$  is the output voltage across the terminals of the secondary coil and is the same as the voltage  $E_c$  across the condenser. This voltage is applied in parallel with any load connected across the secondary and is the secondary voltage applied to the two diodes of the balanced discriminator.

Phase relations in the discriminator transformer can be followed now, analyzing the equivalent secondary circuit as an a-c circuit with  $E_i$  the generator voltage in series with the secondary coil. In the vector diagrams of Fig. 22-7, phase relations are shown for the case of a signal frequency at center and below or above center frequency. In all cases, the primary voltage  $E_p$  is made the reference vector and leads the primary current by  $90^\circ$  because the voltage across an inductance necessarily leads the current through it by  $90^\circ$ . The i-f transformer is tuned to the i-f carrier frequency, making the primary and secondary resonant at center frequency of the FM signal. When the FM signal is at center frequency,

the secondary is resonant and  $X_L$  equals  $X_C$ . Therefore, the secondary current  $I_s$  is in phase with the generator voltage  $E_i$ , as shown in *a* of Fig. 22-7, because there is no net reactance in the circuit. The voltage across  $C$  lags the secondary current by  $90^\circ$ , since the condenser voltage lags its current by  $90^\circ$ . This voltage is  $E_c$  or  $E_s$  and is the voltage applied to the diode plates in push-pull. As shown in the vector diagram, the secondary output voltage  $E_s$  leads the primary voltage  $E_p$  by  $90^\circ$ . When the secondary is center-tapped, the two secondary voltages  $E_{s1}$  and  $E_{s2}$  are provided, each being equal to  $E_s/2$  and of opposite polarity, with one leading

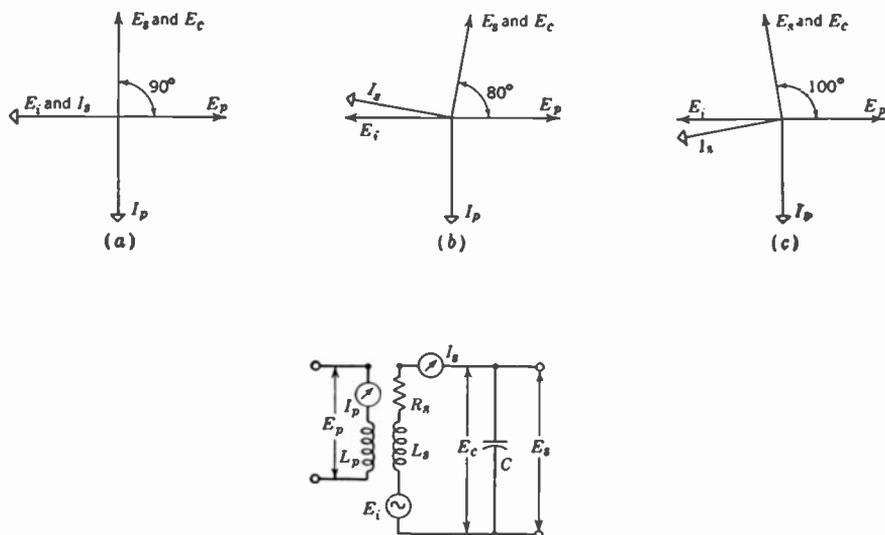


FIG. 22-7. Equivalent circuit and phase relations for the discriminator transformer in a phase-shift discriminator. (a) At center frequency  $I_s$  is in phase with  $E_i$  and  $E_s$  leads  $E_p$  by  $90^\circ$ . (b) Above center frequency  $I_s$  lags  $E_i$  by  $10^\circ$  and  $E_s$  leads  $E_p$  by  $80^\circ$ . (c) Below center frequency  $I_s$  leads  $E_i$  by  $10^\circ$  and  $E_s$  leads  $E_p$  by  $100^\circ$ .

and the other lagging  $E_p$  by  $90^\circ$ .  $E_{s1}$  is the voltage applied to one diode from one side of the secondary coil, and  $E_{s2}$  is the voltage applied to the other diode plate from the opposite side of the coil.  $E_p$  is applied in parallel to both diode plates, as noted previously.

When the FM signal deviates above center frequency, the instantaneous signal frequency is above the resonant frequency of the secondary and  $X_L$  is greater than  $X_C$ . Assuming that the amount of net inductive reactance produces a phase angle of  $10^\circ$ , as an example, the secondary current  $I_s$  lags the generator voltage  $E_i$  by  $10^\circ$ . The voltage across the condenser, which is the same as  $E_s$ , always lags  $I_s$  by  $90^\circ$ , regardless of the phase relation between  $E_i$  and  $I_s$ . Therefore,  $E_s$  now leads  $E_p$  by  $80^\circ$ , as shown in Fig. 22-7b. When the FM signal deviates the same amount

below center frequency, the signal frequency is lower than the resonant frequency and  $X_c$  is greater than  $X_L$ . The phase angle is  $10^\circ$  again, but this time  $I_s$  leads  $E_i$  because the net reactance is capacitive, and it is seen in *c* that  $E_s$  now leads  $E_p$  by  $100^\circ$ .

As the FM signal swings above and below center frequency, therefore, the phase angle of the secondary voltage with respect to the primary voltage varies above and below  $90^\circ$ . When the instantaneous frequency departure from resonance is  $(1/2Q)f_r$ , the net reactance equals the resistance and the phase angle is  $45^\circ$ . Because of the varying phase angle the directly coupled and transformer-coupled voltages combine to provide for the two diodes of the balanced detector a signal voltage that varies in magnitude as the FM signal varies in frequency. The signal is then detected to produce the desired audio voltage in the discriminator output.

*Response.* The phase-shift discriminator response curve shown in Fig. 22-4b is very much the same as for the triple-tuned discriminator. The net output voltage is zero at center frequency. For signal frequency variations above center frequency, a positive output voltage is obtained which increases in magnitude with an increase in the amount of frequency departure. When the instantaneous frequency of the FM signal deviates below center, the output voltage is negative. The useful frequency range must be broad enough to accommodate the maximum frequency swing over the linear portion of the response curve, which is approximately two-thirds of the distance between the peaks. The separation between peaks is increased with higher values of center frequency, increased coupling between primary and secondary, and lower values of  $Q$  for the secondary circuit.

The manner in which a balanced discriminator circuit recovers the desired audio voltage from the FM signal can be seen by reviewing the action of the phase-shift discriminator as the transmitted signal is modulated. When the audio modulating voltage is at its zero value the transmitter output is an r-f signal of constant frequency, which is the center frequency. Tuned in at the receiver, the signal is converted to the intermediate frequency, amplified in the i-f section, and coupled to the discriminator for rectification. With the signal at center frequency the i-f transformer coupling the signal to the discriminator is resonant and the secondary voltage is  $90^\circ$  out of phase with the primary voltage. The signal voltages applied to the two diodes of the discriminator are equal and the output is zero. As the transmitter is modulated, the audio modulating voltage produces frequency departure in the r-f output swing above and below center frequency. Although this r-f signal is converted in the receiver to the lower intermediate frequencies, the variations about center frequency are maintained and the i-f signal has the same amount of frequency departure. Therefore, when the r-f signal departs above center

frequency on the positive half cycle of the audio modulating voltage, the i-f input to the discriminator is above the resonant frequency of the tuned coupling transformer.  $E_s$  is less than  $90^\circ$  out of phase with  $E_p$ , the applied signal voltage for one diode is greater than the other, and the discriminator output is unbalanced to produce a positive output voltage. Following the operation through a positive half cycle, the amount of frequency departure is small when the audio voltage is a little greater than zero, and the discriminator output voltage is small. As the audio modulating voltage progressively increases in magnitude to its peak value, the amount of frequency departure increases to its maximum deviation and the discriminator output voltage increases to its peak value. While the audio modulating voltage declines from its peak value toward zero, the frequency departure decreases and the discriminator output also declines to zero. Similarly, on the negative half cycle of audio modulating voltage the FM signal varies below center frequency, the signal frequency is below resonance and a negative output voltage is obtained which varies in amplitude with the amount of frequency departure. As a result, the audio voltage modulating the r-f carrier is reproduced at the output of the discriminator as a changing d-c voltage that varies with the frequency changes in the FM signal.

*Effect of Interfering Amplitude Modulation.* A balanced discriminator circuit detects amplitude modulation in the i-f signal in addition to the FM detection, because the circuit is balanced only at center frequency. This can be illustrated with some numerical examples. Assume that at the center frequency the FM signal has an amplitude such that 3 volts is applied to each of the two diodes of the discriminator. Each diode provides approximately 3 volts in the output, and since the two rectified voltages are in series opposition the net output voltage is zero. If the amplitude of the FM signal increases to provide 6 volts of i-f signal for the diodes, the rectified output of each diode is 6 volts and the net output is still zero. The detection is balanced at center frequency so that the net output voltage is zero regardless of the amplitude of the FM signal.

For signal frequencies other than center frequency, however, any variation in amplitude of the i-f signal is reproduced in the output of the discriminator. Assume that the i-f signal amplitude provides 5 volts for one diode and 1 volt for the other diode when the FM signal deviates from center frequency by 25 kc. The rectified output is close to 5 volts and 1 volt, respectively, for each of the diodes, and the net output voltage is 4 volts of either positive or negative polarity, depending on which diode has the greater signal voltage. If the i-f signal has the same amount of frequency swing but increases in amplitude to twice its original value, the applied signal voltage for each of the diodes will be twice as great, the signal voltage for each rectifier increasing to 10 and 2 volts, respectively. The dis-

criminator output voltage will then be 8 volts instead of 4 volts, producing an audio output that corresponds to the amplitude variation of the i-f signal in addition to the FM signal deviation.

It should be noted that the ratio between the two diode signal voltages is independent of the i-f signal amplitude and depends only on the amount of frequency departure. This follows from the coupling arrangement used in the balanced discriminator circuit, which proportions the amount of signal voltage for the two diodes according to the amount that the instantaneous signal frequencies differs from the resonant center frequency. In the triple-tuned discriminator the proportion of signal voltage applied to each diode depends upon the difference between the resonant frequency of the diode's tuned circuit and the instantaneous signal frequency. In the phase-shift discriminator the voltage division between the two diodes depends on the phase angle of the secondary voltage with respect to the primary voltage, which varies from  $90^\circ$  according to the difference between the instantaneous frequency and the center resonant frequency. Therefore, the ratio between the two diode voltages is a true measure of the desired audio variations in signal frequency, independent of any amplitude variations in the FM signal.

**22-5. The Limiter.** Since the discriminator responds to amplitude variations in the FM signal that might be present because of interference, the discriminator audio output would include interference effects such as static, ignition noise, and beat-frequency whistles. Therefore, in FM receivers using a balanced discriminator, it is usually preceded by a limiter stage, which has the function of removing all amplitude variations from the FM signal coupled to the discriminator. The limiter in an FM receiver is essentially a class C amplifier in which the plate and screen voltages are so low that effective saturation is reached with small values of signal voltage applied to the control grid. A typical circuit for a single limiter stage is shown in Fig. 22-8a. A sharp-cutoff pentode tube is used and the plate and screen voltages are lowered to 50 volts in order to reduce the amount of negative grid voltage required for plate-current cutoff to a value of about 3 volts. Grid-leak bias is used so that with varying signal amplitudes the bias can automatically adjust itself to a value that allows just the positive tip of the signal swing to drive the grid positive and cause grid current to flow.

The manner in which the limiter functions is illustrated in *b* of Fig. 22-8. Suppose that a signal having a peak amplitude greater than the cutoff bias is impressed on the grid of the tube. Grid-leak bias voltage will be developed with a magnitude approximately equal to the peak value of the signal swing. Grid current flows for a very small part of the positive half cycle at the tip of the positive signal swing. Plate current flows for almost the entire positive half cycle, as indicated by the shaded area in the illustra-

tion. When the amplitude of the input grid signal increases, a greater negative bias voltage is developed, but the grid cutoff voltage remains the same and the average plate current changes very little. Therefore, the amount of plate-current flow in the limiter stage is approximately constant for all signals having an amplitude great enough to develop a grid-leak bias voltage that is greater than the cutoff voltage. The frequency variations in the FM signal are maintained in the output, since the plate-current pulses are produced at the grid signal frequency and excite the plate-tuned circuit at the instantaneous signal frequency. With a relatively uniform value of average plate current, the output voltage across the tuned circuit is constant.

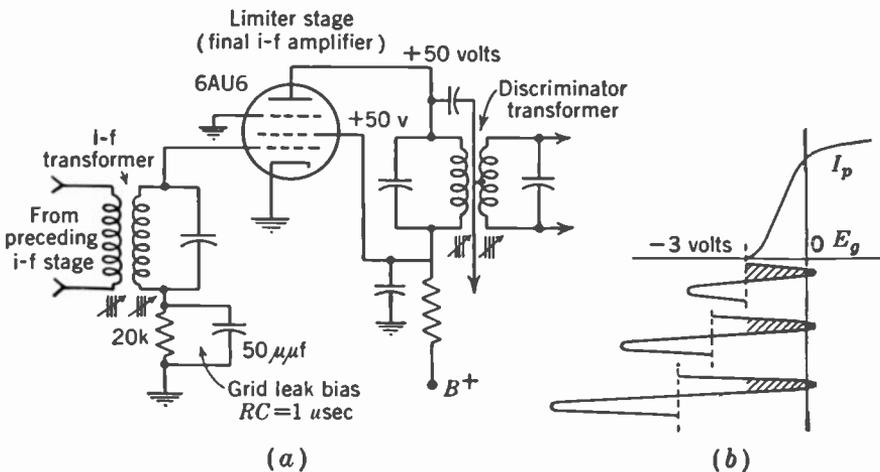


FIG. 22-8. Amplitude limiter for an FM receiver. Input and output circuits tuned to the i-f center frequency of the FM signal. (a) Circuit. (b) Constant plate-current characteristic for different grid-voltage amplitudes.

When the peak amplitude of the grid signal is less than the cutoff voltage the limiting action fails because the stage is practically a class A amplifier for such signals, and the average plate current varies as the grid-leak bias changes with varying signal amplitudes. For this reason, it is required that the stages preceding the limiter have enough gain to provide sufficient signal voltage to saturate the limiter with the smallest useful antenna signal. Assuming an antenna signal of  $50 \mu\text{v}$ , a gain of 60,000 is needed to produce 3 volts at the limiter grid for adequate limiting if this is the grid cutoff voltage. This is the threshold value of signal voltage required for the limiting action. For any signal voltage greater than this value the average plate current is approximately the same and the output voltage is constant. Although a substantial reduction of amplitude variations in the FM signal is obtained with the use of one limiter stage,

two limiters are often used in cascade because of the great improvement of limiting action that can be obtained. Almost perfect limiting is possible when two limiters are used and each is adjusted so that it limits best for signal amplitudes that are not limited as well in the other stage. A small amount of gain is usually obtained in the limiter stage, since the constant output voltage can still be greater than the input signal. The gain varies for different input levels but is about 2 to 5.

For satisfactory limiting action, the time constant of the grid-leak resistor and condenser should be much larger than the period of one cycle of the radio frequency input signal. This is necessary to maintain the grid bias constant as long as the signal amplitude is constant. At the same time, the  $RC$  product should be small compared to the period of one cycle of the highest audio frequency. If it is not, the peaks of amplitude variations occurring at an audio rate will determine the bias and smaller signal amplitudes will not affect the plate current at all. As a result, the interfering AM signal is not limited and can be heard in the audio output. The bias should follow any amplitude variations of the signal in order to keep the peak positive signal swing clamped at zero grid voltage if the plate current is to be constant. Time-constant values of 1 to 4  $\mu\text{sec}$  are commonly used, the smaller values being more effective in rejecting ignition noise, which produces sharp pulses of very short duration.

**22-6. Ratio Detector.** This is an FM detector circuit insensitive to amplitude variations in the FM signal. Because no limiter stage is necessary, allowing fewer i-f amplifiers, the ratio detector circuit is often used in FM receivers.

*Circuit Arrangement.* Referring to the ratio detector diagram in Fig. 22-9, the input coupling transformer has the same function as in the phase-shift discriminator. Both the primary and secondary tuned circuits of the input coupling transformer are resonant at the i-f center frequency. The secondary is center-tapped to produce equal voltages of opposite polarity for the diode rectifiers, while the primary voltage is applied in parallel to both diodes. In the ratio detector circuit, however, one diode is reversed so that the two half-wave rectifiers are in series with the secondary voltage in the input circuit for charging the stabilizing voltage source  $E_3$  in the output side of the circuit. Audio output signal voltage is taken from the audio take-off point  $A$  at the junction of the two diode load condensers  $C_1$  and  $C_2$ , with respect to the center tap on the stabilizing voltage.

*Stabilizing Voltage.* In order to make the ratio detector insensitive to AM interference effects in the audio output, the total voltage  $E_3$  equal to the diode output voltages  $e_1 + e_2$  must be stabilized so that it cannot vary at the audio-frequency rate. Then audio output is obtained at point  $A$  only when the ratio between  $e_1$  and  $e_2$  changes, as their sum voltage

$E_3$  remains fixed by the stabilizing voltage source. Since the ratio of the two diode output voltages depends only on the amount of frequency change in the i-f input, audio output is obtained only for the FM signal variations. The 6-volt battery in Fig. 22-9 illustrates the stabilizing voltage source connected across both diodes, although a large condenser actually is used so that the amount of stabilizing voltage will be determined by the i-f signal level.

The polarity of the stabilizing voltage makes one diode plate negative and the other diode cathode positive, with the amount of bias on each diode equal to one-half the stabilizing voltage. The i-f signal applied to

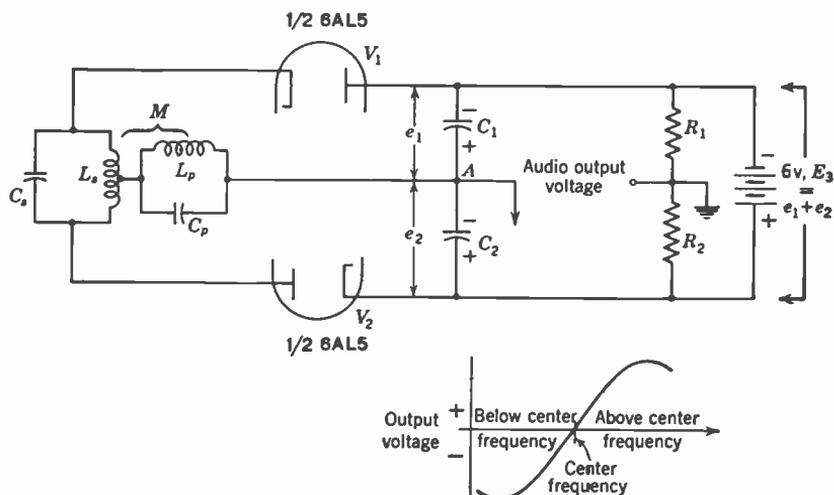


FIG. 22-9. Basic ratio detector circuit and its response curve. Just as in the phase-shift discriminator, the primary and secondary of the coupling transformer are tuned to the i-f center frequency of the FM signal.

each diode must have enough amplitude to overcome the bias, therefore, to produce plate current. It is important to note that the stabilizing voltage source prevents variations caused by amplitude changes in the total secondary signal voltage across both diodes, but it is the proportion of combined primary and secondary signal voltage applied to each diode that produces the audio output.

**Audio Output Signal.** As each diode conducts it produces the rectified output voltage  $e_1$  or  $e_2$  across the diode load condenser  $C_1$  or  $C_2$ , approximately equal to the peak value of the i-f signal applied to each rectifier. At center frequency the input transformer proportions the i-f signal voltage equally for the two diodes, resulting in equal voltages across  $C_1$  and  $C_2$ . The output voltage at the audio take-off point  $A$  is zero, therefore, since  $e_1$  and  $e_2$  are equal and have opposite polarity at point  $A$  with respect to chassis ground. When the FM signal input is above center

frequency and the diode  $V_1$  has more i-f signal input than  $V_2$ , the rectified diode voltage  $e_1$  is greater than  $e_2$ . This makes point  $A$  more positive, producing audio output voltage of positive polarity. Below center frequency, the diode  $V_2$  has more i-f signal input and  $e_2$  is greater than  $e_1$ , resulting in audio output voltage of negative polarity at point  $A$ . This is illustrated by the ratio detector response curve in Fig. 22-9, which is the same as the discriminator response curve.

An important characteristic of the ratio detector output circuit can be illustrated by numerical examples. Assume that the frequency deviation above center frequency increases  $e_1$  by 1 volt. This makes point  $A$  1 volt more positive. At the same time,  $e_2$  decreases by 1 volt. This makes point  $A$  1 volt less negative, which is the same change as 1 volt more positive. The two rectified diode voltages  $e_1$  and  $e_2$ , then, produce the identical voltage change of 1 volt in the positive direction at point  $A$ . Since the audio signal is taken from point  $A$ , the amount of output is the same as though only one diode were supplying audio voltage corresponding to the frequency variations in the FM signal. As a result, the audio voltage output of the ratio detector is one-half the output of a discriminator, where the audio signal voltages from the two diodes are combined in series with each other at the audio take-off point. The output in the ratio detector must be taken from the junction of the two diode loads because there is no audio signal voltage across the stabilizing voltage source.

*Typical Circuit.* If a battery were used for the stabilizing voltage, the diodes would operate only with a signal at least great enough to overcome the battery bias on each diode. A large condenser is used instead, as illustrated by  $C_3$  in the schematic diagram of a typical ratio detector circuit in Fig. 22-10.  $C_3$  charges through the two diodes in series, automatically providing the desired amount of stabilizing voltage for the i-f signal level. The capacitance of  $C_3$ , which is generally called the *stabilizing condenser*, is large enough to prevent the stabilizing voltage from varying at the audio-frequency rate. A discharge time constant of about 0.1 sec is required, as provided by  $R_1$  in series with  $R_2$  across  $C_3$ . Since the voltage across the stabilizing condenser is proportional to the amount of i-f signal input, the negative side of  $C_3$  is a convenient source of a-v-c voltage, equal to one-half the stabilizing voltage, to control the gain of preceding i-f and r-f stage, if this is desired. Whether automatic voltage control is used or not, though, the d-c voltage at this point indicates the i-f signal amplitude into the ratio detector.

The tertiary winding  $L_t$  of the ratio detector transformer is used to couple the primary signal voltage in parallel to the two diodes.  $L_t$  is wound directly over the primary winding for very close coupling, so that the phase of the primary signal voltage across  $L_p$  and  $L_t$  is substantially

the same. The construction of a ratio detector transformer with the tertiary winding is illustrated in Fig. 22-11. This arrangement with the tertiary winding, instead of direct coupling to the secondary center tap, is commonly used with the ratio detector circuit in order to match the high-impedance primary and the relatively low impedance secondary, which is loaded by the conduction in the diodes. The resistor  $R_3$  in series with  $L_t$  limits the peak diode current, to improve the balance of the dynamic input capacitance of the diodes at high signal levels. Because of the  $90^\circ$  phase relation between the voltages across the secondary and across  $L_t$  at resonance, the circuit provides detection of the FM signal in the same manner as the phase-shift discriminator. Both the primary

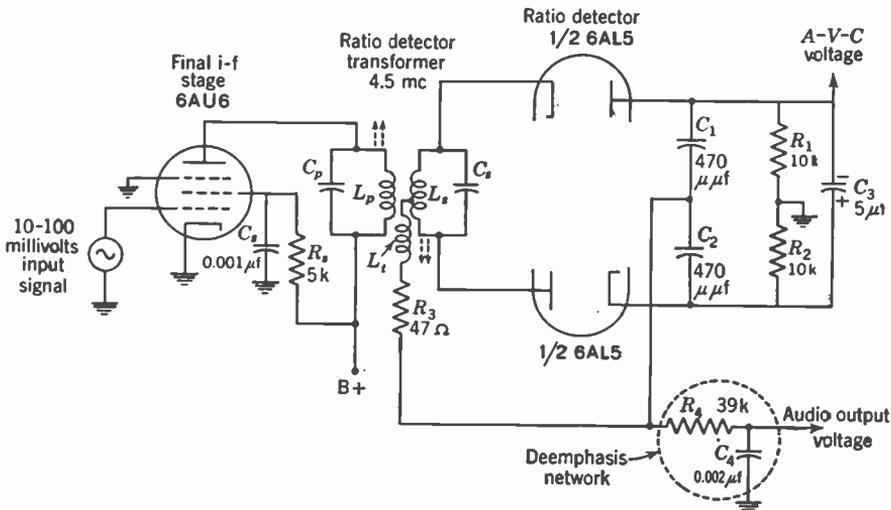


FIG. 22-10. Schematic diagram of ratio detector for i-f center frequency of 4.5 Mc, with balanced output circuit.

and secondary of the ratio detector transformer are tuned to the i-f center frequency, which is generally 10.7 Mc for FM broadcast receivers or 4.5 Mc in intercarrier-sound television receivers. The audio output voltage of the ratio detector is taken from the junction of the diode load condensers  $C_1$  and  $C_2$ , in series with the audio deemphasis network, to supply the desired audio signal for the first audio amplifier.

*The Stabilizing Voltage and AM Rejection.* Referring to Fig. 22-10 the secondary voltage across  $L_s$  charges the stabilizing condenser  $C_3$  in series with the two diodes, each with an internal plate resistance of about 200 ohms.  $C_3$  can discharge through  $R_1$  and  $R_2$ , providing the discharge time constant of 0.1 sec, which is long compared with the lowest audio frequency. As a result,  $C_3$  charges to a value of stabilizing voltage determined by the carrier level and the stabilizing voltage cannot change

appreciably when AM interference varies the i-f signal amplitude at the audio-frequency rate.

For upward modulation, when the i-f signal amplitude increases, the greater secondary voltage produces more current through the diodes. When the i-f signal amplitude decreases, with downward modulation, there is less current through the diodes. The difference between the applied voltage and the stabilizing voltage is across the diodes. They serve effectively as a variable load resistance across the secondary, therefore, their resistance changing with the amount of diode current. More current is equivalent to a lower diode resistance. Since the resistance of the diodes is in parallel with the secondary tuned circuit, its operating  $Q$  depends upon the amount of the shunt load resistance provided by the diodes. A higher value of shunt resistance produces less loading, allowing a higher  $Q$  for the tuned circuit; less shunt resistance lowers the  $Q$ .

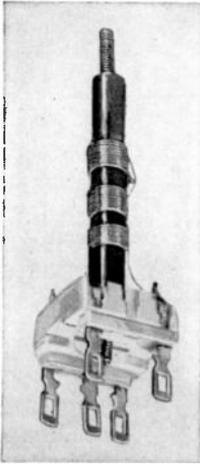


FIG. 22-11. Construction of ratio detector transformer. (Meisner Mfg. Div.)

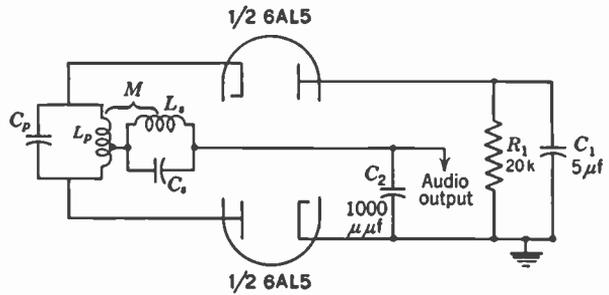


FIG. 22-12. Single-ended ratio detector circuit.

It is important to note now that the  $Q$  of the secondary tuned circuit determines its phase sensitivity. This defines the amount of phase shift from  $90^\circ$  produced by a given amount of frequency variation from center frequency. Reduced phase sensitivity causes smaller changes in the ratio of the signal voltages applied to the two diodes when the FM signal varies from center frequency, producing less audio output. The phase sensitivity of the secondary tuned circuit is proportional to its  $Q$ , resulting in less audio output for a lower  $Q$ . The amount of audio output produced by the frequency modulation will be reduced if the i-f signal amplitude increases, therefore, as the conduction in the diodes increases to lower the shunt resistance across the secondary tuned circuit, reducing its  $Q$  and phase sensitivity. If the i-f signal amplitude decreases, the  $Q$  and phase sensitivity will increase, resulting in more audio output. In sum-

mary, then, the effect of the variable diode load resistance compensates for AM variations in i-f signal, by decreasing the amount of audio output for an increase in carrier amplitude and increasing the audio output for a decrease in carrier amplitude. By using the proper value of  $Q$  for the ratio detector transformer and suitable values of load resistance, the variation in phase sensitivity can be made to compensate for the change in carrier amplitude, resulting in no effect in the audio output, as long as the stabilizing voltage remains constant. In general, the ratio detector can reject a higher percentage of upward modulation than downward modulation, because of the limitation of the diodes being biased out of conduction by the stabilizing voltage if the AM interference makes the carrier amplitude fall too low. To provide for rejection of upward modulation and increase the amount of downward modulation that can be rejected without the diodes cutting off, a high  $Q$  secondary tuned circuit is used with relatively low resistances for  $R_1$  and  $R_2$ , so that the operating  $Q$  is much lower than the unloaded  $Q$ .

*Single-ended Ratio Detector.* Figure 22-12 illustrates a ratio detector circuit that has unbalanced output. The input circuit, which is shown in its equivalent form, is the same as in the balanced circuit. However, only the one diode load condenser  $C_2$  is used in the audio output circuit and the stabilizing voltage across  $R_1$  and  $C_1$  is not center-tapped. It is not necessary to ground the center point of the stabilizing voltage source, since the only effect of the different ground connection is to change the d-c level of the audio output signal with respect to chassis ground. The same audio signal variations are obtained from the audio take-off point, about a d-c voltage axis equal to one-half the stabilizing voltage, instead of varying around the zero axis of chassis ground. It should be noted, though, that the audio output always has a zero d-c level with respect to the mid-point of the stabilizing voltage source. With the unbalanced arrangement, the entire stabilizing voltage is available for automatic volume control. Only the one diode load condenser  $C_2$  is necessary in the output circuit to obtain the audio signal because it serves as the load for both diodes. Since each diode charges  $C_2$  in proportion to the i-f signal input for each rectifier, and in opposite polarity, the voltage across the condenser is the same audio output signal as in the balanced ratio detector. The capacitance of the one audio output condenser  $C_2$  is doubled in the single-ended circuit because it replaces two diode condensers that are effectively in parallel for signal voltage in the balanced circuit. The unbalanced circuit is often used because fewer components are necessary, but the AM rejection for low-frequency amplitude variations is better in the balanced ratio detector circuit.

*Ratio Detector Receiver Characteristics.* Referring to Fig. 22-10, note that only 10- to 100-mv signal is applied to the grid of the i-f stage driv-

ing the ratio detector. With a gain of about 100, the i-f output voltage will be enough to drive the diodes. Since no limiter is required, less i-f gain is needed as there is no fixed threshold voltage that must be exceeded, the ratio detector automatically adjusting itself to the i-f signal level. Fewer i-f stages are necessary therefore, making the ratio detector receiver more economical, compared with the limiter-discriminator arrangement. The ratio detector receiver is relatively quiet between stations and easy to tune.

**22-7. Gated-beam Tube FM Detector and Limiter.** This arrangement of an FM detector circuit that can also reject AM makes use of the electron-beam characteristics in the gated-beam tube types 6BN6 and 12BN6. The circuit is illustrated in Fig. 22-13. The construction of

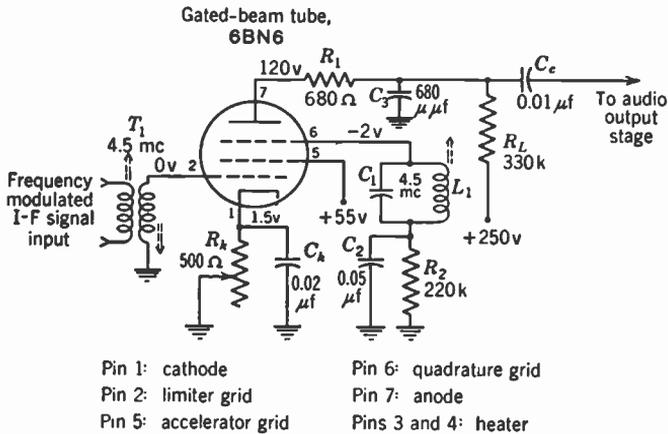


FIG. 22-13. Gated-beam tube circuit for limiting and FM detection. I-F center frequency is 4.5 Mc. Connections shown for 7-pin miniature glass tube 6BN6. (From Zenith K Series chassis.)

the gated-beam tube is different from conventional amplifier tubes, in that the electrons emitted from the cathode form a vertical beam in passing through the grids to the anode. In normal operation, plate voltage of 100 to 150 volts is applied to the anode and the positive d-c voltage for the accelerating grid is 40 to 100 volts, while the limiter grid and quadrature grid serve as control grids.

**Grid-voltage Control.** Plate current is cut off by the limiter grid when its potential is approximately  $-2$  volts. When the limiter grid voltage is zero or 1 to 2 volts positive, however, maximum plate current of about 3 ma flows, and the plate current is constant for limiter grid voltages up to  $+30$  volts. The constant plate current results because the beam current is limited in passing through the aperture in the accelerator structure. With wide variations of amplitude in the signal input, therefore, the limiter grid voltage is effective in limiting the plate current to a constant

value, eliminating AM interference in the output. Because of its sharp cutoff and the constant plate-current characteristic, the limiter grid effectively produces an on-off control, resulting in square-wave pulses of plate current at the frequency of the input signal. Similarly, the quadrature grid produces plate-current cutoff with a negative potential of several volts, and plate-current saturation when its grid potential is about +2 volts or more. Since a relatively small change in voltage for either grid changes the plate current from cutoff to saturation, each grid acts as a voltage-controlled gate. Either gate can cut off the plate current, resulting in no output. Therefore, both gates must be open at the same time to have output plate current.

*Quadrature Grid.* With signal voltage on the limiter grid, signal current flows in the quadrature grid circuit, by space charge coupling between the electron beam and the quadrature grid. This makes the quadrature grid current approximately  $90^\circ$  out of phase with the anode beam current and the limiter grid voltage.

*Circuit Operation.* Referring to Fig. 22-13, the frequency-modulated signal from the last i-f transformer  $T_1$  is coupled to the limiter grid of the 6BN6 gated-beam tube. The operating point for best limiting action is adjusted between cutoff and saturation by the cathode bias produced by  $R_kC_k$ . The i-f signal with a peak amplitude of 2 to 6 volts then drives the tube alternately into conduction and cutoff, resulting in square-wave variations of beam current with the instantaneous frequency of the input signal. Notice that the limiting of the i-f signal is independent of any  $RC$  time constant in the grid circuit, providing good rejection of impulse noise. To the quadrature grid is connected the parallel resonant circuit  $L_1C_1$  tuned to the i-f center frequency. Since this grid extracts energy from the electron beam by space charge coupling, the resonant circuit is excited to produce the i-f signal voltage on the quadrature grid. The signal voltage variations on the quadrature grid lag the input signal voltage on the limiter grid by  $90^\circ$ , however, at the i-f center frequency when the tuned circuit is resonant.  $R_2C_2$  provides bias for the quadrature grid, in addition to the cathode bias.

It is important to note now that both grid-controlled gates must be open at the same time to produce plate current. At center frequency, when the quadrature grid voltage lags the limiter grid voltage by  $90^\circ$ , the square-wave plate-current pulses have about one-half the width that would be obtained if both grids were open simultaneously. Plate current starts with the delayed opening of the quadrature grid gate and stops with closing of the limiter grid gate. Above center frequency, the phase of the quadrature grid voltage changes to produce narrower plate-current pulses, resulting in a lower value of average plate current. Below center frequency the phase of the quadrature grid voltage permits wider

pulses and a higher average value of plate current. As a result, the variations of frequency in the FM input signal produce corresponding variations in the average plate current. The  $IR$  voltage drop across the plate load resistor, therefore, is the desired audio output signal, which is coupled by  $C_c$  to the audio amplifier.  $R_1$  is an isolating resistor to prevent the audio output circuit from bypassing the i-f signal voltage at the anode, which is coupled back to reinforce the quadrature grid voltage.  $C_3$  is the i-f bypass condenser in the audio output circuit, integrating the plate signal current variations to provide the desired audio output voltage. Notice that only a single power-amplifier stage is used, which drives the loudspeaker, as the audio signal of about 10 to 15 volts r-m-s from the 6BN6 is enough to drive the output stage without the need for an audio voltage amplifier stage.

**22-8. Audio Deemphasis.** As was discussed in Sec. 7-10, it is standard practice in FM broadcasting to preemphasize the audio modulating voltage at the transmitter, increasing the relative amplitudes of the higher audio frequencies in order to improve the signal-to-noise ratio for these normally weak signals. This would be a tone distortion of the audio signal, producing a treble boost, if the relative response for audio frequencies were not corrected by deemphasis in the receiver. Although the deemphasis can be accomplished in any part of the audio section up to the loudspeaker, a deemphasis network is often connected in the FM detector output circuit, as shown in Fig. 22-10, consisting of a series resistance and shunt capacitance. Since the reactance for the shunt capacitance is smaller for the higher audio frequencies, the audio voltage output decreases with an increase in audio frequency. The time constant for preemphasis is standardized at  $75 \mu\text{sec}$  for FM commercial broadcast stations and the FM associated sound in television broadcasting. Therefore, the time constant required for the deemphasis in the receiver is also  $75 \mu\text{sec}$ .

**22-9. Intercarrier Sound.** Practically all television receivers use the intercarrier system of converting the associated sound into a frequency-modulated sound i-f signal with the center frequency of 4.5 Mc, equal to the difference between the transmitted sound and picture carrier frequencies. This is illustrated in Fig. 22-14. Instead of using a separate amplifier section for the i-f sound signal from the mixer stage in the r-f tuner, the picture and sound i-f signals are amplified together in the common i-f amplifier section. The FM sound signal beats with the picture carrier in the second detector, producing the difference frequency of 4.5 Mc that is used as the sound i-f carrier frequency in the receiver. The frequency difference between the sound and picture carriers is standardized at 4.5 Mc for any channel and is held accurately within close tolerances at the transmitter. Therefore, intercarrier-sound receiv-

ers do not depend upon any precise local oscillator frequency for reception of the associated sound signal, since the difference between the sound and picture carrier frequencies is still 4.5 Mc even when the receiver's local oscillator drifts in frequency. This is an important advantage over the split-sound system with a separate sound i-f amplifier section, where it is difficult to maintain the local oscillator frequency within the narrow limits required to heterodyne the r-f sound signal down exactly to the intermediate frequencies within the relatively narrow sound i-f pass band. Since the higher the oscillator frequency, the more difficult is the problem

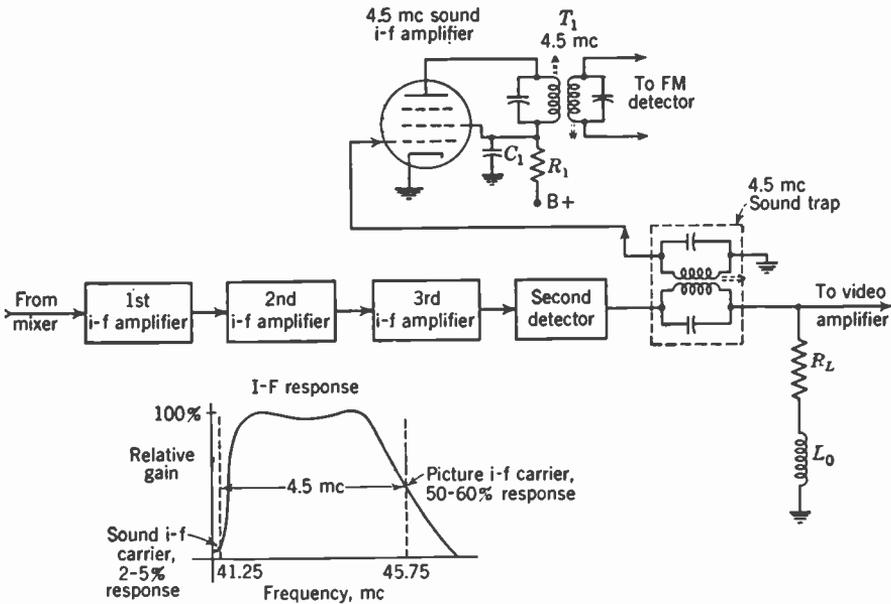


FIG. 22-14. The intercarrier-sound system, showing the common picture and sound amplifier with its response curve, 4.5-Mc sound take-off circuit in output of second detector, and 4.5-Mc sound i-f amplifier driving the FM detector.

of drift, the intercarrier-sound system is especially advantageous for the u-h-f channels.

*Circuit Arrangement.* Referring to Fig. 22-14, the intercarrier-sound receiver uses one common i-f amplifier section with a bandwidth great enough to pass both the picture and sound i-f signals obtained from the mixer or first detector. This sound i-f signal is identical with the sound i-f output in a split-sound receiver. However, it is amplified in the same stages as the picture i-f signal, in the common i-f amplifier of the intercarrier-sound receiver. The 4.5-Mc sound signal is obtained in the output of the second detector stage, where the sound i-f signal can heterodyne with the picture carrier. This is also the video detector for the

video amplifier. Therefore, the second detector's output includes the composite video signal required for the video amplifier and the 4.5-Mc sound signal for the 4.5-Mc sound i-f section of the receiver. The 4.5-Mc output from the second detector can be considered the second sound i-f carrier frequency, to distinguish this from the i-f output of the mixer or first detector stage. With respect to the sound signal, the intercarrier receiver is in effect a double superheterodyne, beating the transmitted r-f sound carrier signal down to the 4.5-Mc i-f sound signal in two steps. The first frequency conversion occurs in the first detector, which produces an i-f sound signal equal to the difference between the local oscillator and r-f sound carrier frequencies. This is then converted to the lower second sound i-f carrier frequency at 4.5 Mc in the second detector, with the picture carrier serving effectively as a local oscillator signal that can beat with the sound signal.

The fact that the sound is frequency-modulated does not alter the frequency conversion in either case, so that the output of the second detector is the desired FM sound signal with the i-f center frequency of 4.5 Mc. In any intercarrier-sound receiver, the second sound i-f carrier is always 4.5 Mc, regardless of the frequencies for the first sound i-f carrier and picture i-f carrier. It should be noted that in some intercarrier-sound receivers the 4.5-Mc sound signal is amplified in the video section, along with the composite video signal, and the 4.5-Mc sound take-off circuit is in the video output. Also, some receivers have a separate second detector for producing the 4.5-Mc sound signal, which is called the *sound converter* stage. This arrangement has the advantage of preventing beat frequencies produced in the sound conversion from interfering with the video signal for the picture.

In order to produce the 4.5-Mc sound signal, the first sound i-f carrier must have much less amplitude than the picture carrier in the input to the second detector. This is necessary because, when two signal voltages of slightly different frequencies are mixed, the resultant wave varies in amplitude at the difference frequency and has the original modulation of the weaker of the two input signals. To preserve the modulation of the sound signal, therefore, the amplitude of the first sound i-f carrier is made one-tenth to one-twenty-fifth of the i-f picture carrier amplitude. The sound i-f carrier is then weaker even when the picture carrier amplitude is very low with maximum white picture information. As shown in Fig. 22-14, the over-all response curve of the common i-f amplifier section provides the required response for the first sound i-f signal, with its relative gain about 2 to 5 per cent up from the zero base line. Higher gain results in more sound output, but the first sound i-f carrier amplitude must be very much less than the picture i-f carrier in order to minimize the AM distortion produced in the 4.5-Mc frequency-modulated sound

signal by the amplitude-modulated picture signal. Since the sound is an FM signal, the problem of AM picture signal in the 4.5-Mc sound i-f signal is solved by the use of the FM detector to recover the desired audio signal, while the AM rejection circuits reduce amplitude distortions. Sound signal interference in the picture is minimized by making the response for the first sound i-f carrier very low and broad enough to prevent slope detection with the resultant sound bars in the picture, while a 4.5-Mc trap in the video amplifier eliminates the 4.5-Mc beat pattern.

It should be noted that the 4.5-Mc sound signal is still an r-f FM signal with variations about the i-f center frequency of 4.5 Mc that correspond to the desired audio signal. The maximum frequency swing is  $\pm 25$  kc, as in the r-f and first i-f sound signal. When the 4.5-Mc sound signal is coupled into an FM detector, the audio voltage recovered in the detector output can then be coupled to an audio amplifier for driving the loudspeaker. The FM detector circuit must be aligned exactly at 4.5 Mc and be stable in tuning, as a slight change in the 4.5-Mc sound i-f alignment cannot be corrected by the oscillator fine tuning control. Rejection of amplitude modulation in the 4.5-Mc frequency-modulated signal is important in order to minimize AM distortion in the audio output.

*Intercarrier Buzz.* Since the 4.5-Mc beat depends upon the picture carrier, excessive amplitude variations of the AM picture carrier signal can be transferred to the 4.5-Mc sound i-f signal by cross modulation in the second detector. This is especially evident with a predominantly white picture, when the picture carrier amplitude varies between the extremes of the white level close to zero amplitude and the 75 per cent pedestal level for horizontal and vertical blanking pulses. Then the 4.5-Mc sound i-f signal has severe amplitude variations corresponding to the blanking and sync pulses. Without enough AM rejection, these extreme amplitude variations can produce a 15,750-cycle hiss in the audio output, corresponding to the horizontal blanking rate, and 60-cycle buzz at the vertical blanking rate. Since the 60-cycle buzz produced by voltage variations of the vertical blanking and sync pulses in the 4.5-Mc sound i-f signal is usually more evident, it is generally called *intercarrier buzz*. The buzz is minimized by keeping the relative amplitude of the first sound i-f carrier very low to reduce amplitude modulation of the 4.5-Mc sound i-f signal by the picture carrier signal and providing sufficient AM rejection in the 4.5-Mc sound i-f circuits, with the FM detector exactly aligned for balance at the 4.5-Mc center frequency.

Another possible source of intercarrier buzz is the video amplifier if the 4.5-Mc sound i-f signal is amplified in this section of the receiver before being coupled to the 4.5-Mc sound i-f section. When a sharp-cutoff tube is used for the video amplifier, the amplitude of negative

synchronizing pulse voltage at the grid may be great enough to cut off the tube with excessive composite video signal input. If the video amplifier stage is cut off during each vertical sync pulse, or for the vertical blanking time, there is no sound signal momentarily at the 60-cycle rate causing severe 60-cps buzz in the audio output. This condition is avoided when the 4.5-Mc sound i-f signal is taken from the second detector, rather than the video amplifier output.

If the amplitude of the transmitted picture carrier signal is reduced to zero for maximum white picture information, intercarrier buzz will result because the 4.5-Mc sound i-f signal is interrupted when either the picture carrier or the sound carrier is missing. Then the sound signal would vary at the 60-cycle rate as it is off during the time of maximum white picture information without any picture carrier and on for vertical blanking time. Even if the picture carrier does not fall all the way to zero for maximum white, excessive amplitude modulation of the 4.5-Mc sound i-f signal occurs with a white level much less than 10 per cent of the peak picture carrier amplitude. Broadcast stations generally monitor the modulated picture carrier signal to keep the downward peaks of modulation at the level of  $12.5 \pm 2.5$  per cent to minimize buzz in intercarrier-sound receivers.

*Oscillator Fine Tuning.* The main advantage of the intercarrier-sound system is the fact that the problem of frequency drift in the local oscillator is largely eliminated with respect to tuning in the associated sound signal, as the 4.5-Mc sound i-f signal is obtained regardless of the exact local oscillator frequency. With the second sound i-f signal now dependent on the 4.5-Mc difference between the transmitted picture and sound carrier frequencies, oscillator tuning is not so critical because of the wide band pass of the picture signal circuits. Experimental viewing tests have shown that variations in local oscillator tuning up to 0.5 to 1.0 Mc may be permissible without distorting the picture to the point where the receiver must be retuned. Therefore, the oscillator fine tuning control is not absolutely necessary on intercarrier-sound receivers, because the associated sound signal is automatically tuned in when the picture is obtained for each channel. The fine tuning control is usually included on intercarrier-sound receivers, however, in order to vary the tuning for the picture. An r-f interfering signal can often be tuned out with the fine tuning control, while maintaining the sound and picture. In intercarrier-sound receivers, therefore, the fine tuning control should be adjusted for the best picture.

Referring to the i-f response curve in Fig. 22-14 it can be seen that a change in oscillator frequency that decreases the i-f picture carrier frequency, moving it higher up on the response curve, results in more picture contrast, but there is less detail as the full bandwidth of the i-f

response is not utilized. A change in oscillator frequency that moves the i-f picture carrier down on the response curve allows less contrast but with more detail in the picture. The sound volume becomes louder in this direction as the first sound i-f carrier amplitude is increased but the intercarrier buzz also increases. In addition, sound bars appear in the picture because of slope detection. The best way to adjust the fine tuning control on intercarrier-sound receivers, usually, is to move it to the point where sound bars appear in the picture and then back off a little for a good contrast range with full detail. The contrast control can then be adjusted for the desired contrast in the picture and the volume control used for the sound level.

Another advantage of the intercarrier-sound system is that microphonics in the local oscillator do not affect the sound output. Although the microphonics vary the oscillator frequency, the picture and sound i-f carrier frequencies are varied by the same amount and there is no change in the 4.5-Mc sound i-f signal.

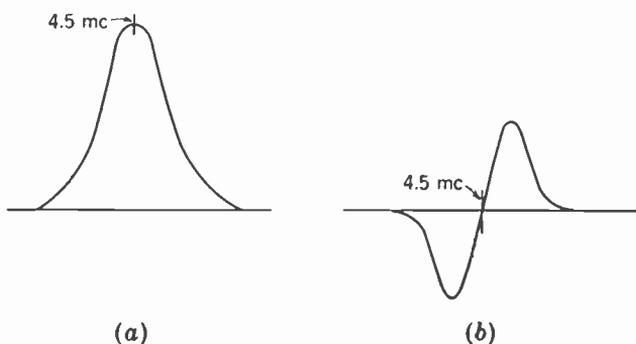


FIG. 22-15. Response curves for sound i-f section with center frequency of 4.5 Mc. (a) Single-peaked response of sound i-f stages. (b) S curve of discriminator or ratio detector.

**22-10. Sound I-F Alignment.** The alignment of the FM sound section consists of peaking the sound i-f stages for maximum output and balancing the discriminator or ratio detector secondary circuit for zero output at the i-f center frequency. Figure 22-15 shows typical response curves for the sound i-f stages and the discriminator or ratio detector.

*Discriminator Alignment.* Adjustment of the discriminator is essentially a problem of tuning the input coupling circuits of the discriminator transformer to the required frequency. This can be done without a visual response curve, using just a high-impedance d-c voltmeter and a conventional r-f signal generator to supply unmodulated output voltage at a single frequency. The d-c voltmeter is connected across the audio output of the discriminator. A d-c voltmeter is used because any output

from the discriminator is a steady voltage when the input signal is not frequency-modulated, since there are no frequency variations to produce variations in the output. The meter should have an impedance of 20,000 ohms per volt or higher to avoid detuning the discriminator. The signal generator, accurately set at the resonant frequency of the circuit being tuned, is connected to the control-grid circuit of the last i-f stage before the discriminator.

The phase-shift discriminator is aligned by tuning the primary and secondary circuits of the discriminator transfer to the i-f center frequency. With the signal generator supplying output voltage to the grid of the preceding stage, at the i-f center frequency, and the d-c voltmeter connected across the audio output terminals:

1. Tune the primary of the discriminator transformer for maximum output.
2. Tune the secondary of the discriminator transformer for a sharp drop to zero.

Since it will be possible to produce either a positive or negative output voltage when adjusting the secondary, it should be tuned for zero indication at the balance point where the output voltage starts to swing from one polarity to the other.

When the signal generator frequency is varied manually above and below center frequency, the d-c output voltage should vary from zero at center frequency to a maximum value at both sides of center frequency, with opposite polarities below and above center. The response should be symmetrical about center frequency with equal output voltages produced for the same amount of frequency change below or above center frequency. The two points of maximum output voltage, corresponding to the two peaks on the discriminator response curve, should have the required frequency separation. The actual polarity of the output voltage for a frequency departure above or below center does not matter. It is only required that the output voltages be of opposite polarity for frequency departures above and below center frequency.

*Ratio Detector Alignment.* In the ratio detector, the primary and secondary of the input transformer are also resonant at center frequency. However, the d-c voltmeter must be connected across the stabilizing condenser voltage to read rectified i-f signal and then moved to the audio take-off point to indicate zero balance. With the signal generator supplying output voltage to the grid of the last i-f stage at the i-f center frequency:

1. Tune the primary of the ratio detector transformer for maximum output on the d-c voltmeter across the stabilizing condenser voltage.
2. Move the d-c voltmeter to the audio take-off point and tune the secondary for balance.

In a ratio detector with an output circuit balanced to ground, the secondary is tuned for a sharp drop to zero, as in balancing the discriminator secondary. When the ratio detector has a single-ended output circuit, however, the audio output has a d-c level equal to one-half the stabilizing voltage. Therefore, usual practice is to insert two balancing resistors temporarily, as shown in Fig. 22-16, converting the single-ended arrangement to a balanced output circuit, so that the ratio detector secondary can be aligned for balance at zero with the d-c voltmeter.

*Visual Response Curve.* To observe the bandwidth, linearity, and symmetry of the discriminator or ratio detector, its visual response curve can

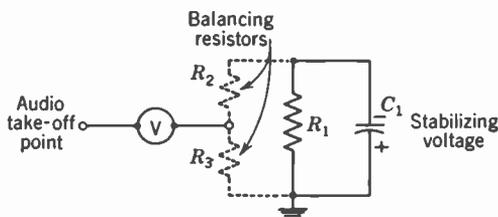


FIG. 22-16. Temporary resistor connections to balance audio take-off point to ground for aligning secondary of single-ended ratio detector with a d-c voltmeter.

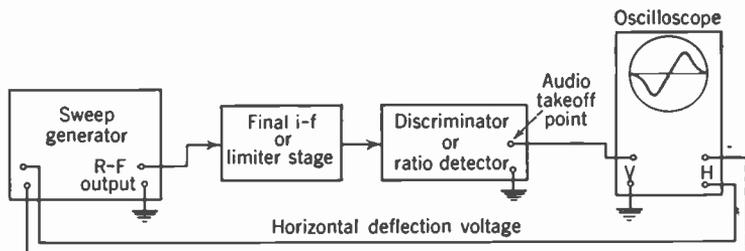


FIG. 22-17. Test equipment and connections to obtain visual response curve of discriminator or ratio detector.

be obtained, using a sweep generator and oscilloscope. The sweep generator is connected to the grid of the last i-f stage and set to a center frequency equal to the i-f carrier frequency, with a sweep width of about 1 Mc, while the vertical input of the oscilloscope connects to the audio take-off point of the detector. Horizontal deflection voltage for the oscilloscope is taken from the generator, with the oscilloscope's internal deflection generator turned off. The pattern on the oscilloscope screen showing the visual response is usually the S-shaped curve illustrated in Fig. 22-17, for either the discriminator or ratio detector, with the sweep generator's blanking turned on to provide the zero reference line. An accurate marker generator is necessary to indicate the intermediate frequencies on the curve. The i-f center frequency should be at the point

where the curve crosses the zero reference line. Note that the marker indication will disappear at this crossover point. For this reason it may be helpful to use an AM marker, noting that the amplitude modulation on the entire response curve disappears when the marker frequency is at the crossover point.

The bandwidth between the two peaks on the response curve above and below center frequency is about 300 to 500 kc in split-sound receivers, in order to allow for slight changes in local oscillator tuning, but can be less in intercarrier-sound receivers. The secondary tuned circuit of the input coupling transformer is adjusted for balance by making the crossover point equal to the i-f center frequency. The primary is effective in producing the symmetrical response above and below center frequency. Bandwidth and linearity of the response curve depend on the coupling and Q of the tuned circuits in the transformer.

*Aligning Gated-beam Tube FM Detector and Limiter.* In a receiver using the gated-beam tube for FM detection and AM rejection, the quadrature tuned circuit is aligned for maximum audio output, with FM signal input at the i-f center frequency, and the cathode resistor is adjusted for minimum audio output with AM signal input at the i-f center frequency.

*Sound I-F Stages.* The sound i-f stages usually have a symmetrical single-peaked response at the i-f center frequency, as shown in Fig. 22-15a. This response can be obtained by peaking the tuned circuits in the sound i-f stages for maximum output at center frequency, from the last i-f stage backward. The signal generator, which is set for unmodulated r-f signal at the i-f center frequency, is connected for input signal to the grid of the first i-f stage, while the d-c voltmeter is connected to measure rectified i-f signal output. In a limiter-discriminator circuit, the d-c voltmeter is connected across the grid resistor of the limiter, using the grid-leak bias to indicate the amount of rectified i-f signal. With a ratio detector, the d-c voltmeter is connected across the stabilizing condenser, since this d-c voltage indicates the amount of rectified i-f signal. The visual response curve of the sound i-f stages can be obtained by using a sweep generator to supply frequency-modulated signal at the i-f center frequency to the first i-f stage, while the vertical input terminals of the oscilloscope are connected in place of the d-c voltmeter.

*Over-all Sound I-F Alignment.* In aligning the sound i-f section, the FM detector is adjusted first. With a discriminator-limiter circuit, after the secondary of the discriminator transformer has been adjusted for balance and the primary for symmetry, the i-f stages are peaked for maximum output across the grid resistor of the limiter. In a ratio detector receiver, the secondary of the input transformer is balanced for zero output at the audio take-off point. Then the primary of the ratio

detector transformer and all the preceding sound i-f stages can be peaked for maximum output voltage across the stabilizing condenser.

If a crystal-controlled marker accurate within 1 kc is not available, it may be preferable to align the sound i-f section by using the associated sound signal transmitted by the broadcast station. Any channel can be used, and the i-f alignment will then be the same for all stations. Since this procedure uses frequency-modulated sound signal input for the alignment, the following points should be noted:

1. With FM signal input, the d-c output voltage is still zero at the audio take-off point, assuming a balanced output circuit, when the secondary of the discriminator or ratio detector transformer is tuned to the i-f center frequency. Zero balance also results in a sharp drop in background noise. The center frequency is automatically at the correct i-f value of 4.5 Mc in intercarrier-sound receivers, but in split-sound receivers the center frequency of the sound i-f signal changes with the setting of the local oscillator fine tuning control. In this case the required sound i-f center frequency can be obtained approximately by setting the fine tuning control for the best picture.

2. After the secondary has been balanced at the correct i-f center frequency, the primary of the detector transformer and all the preceding sound i-f circuits can be peaked for maximum audio output.

**22-11. Complete Sound I-F Circuit.** Figure 22-18 shows the schematic diagram of the complete sound i-f section for the frequency-modulated 4.5-Mc second sound i-f signal from the second detector in an intercarrier-sound receiver. Two 4.5-Mc sound i-f stages are used with the ratio detector, which provides the desired audio output signal for the first audio amplifier stage.

The 4.5-Mc sound i-f signal is taken from a 4.5-Mc trap circuit in the output of the second detector and coupled to the control grid of the first sound i-f stage  $V_{101}$ . This is a single-tuned amplifier, with the 4.5-Mc i-f transformer  $T_{101}$  providing the plate load impedance. The amplified output is coupled by  $C_{103}R_{103}$  to the grid of the ratio detector driver stage  $V_{102}$ . This is also a 4.5-Mc sound i-f amplifier, with the ratio detector transformer  $T_{102}$  providing the plate load impedance tuned to 4.5 Mc. The single-peaked over-all i-f response curve for this circuit is shown in Fig. 22-15, with the ratio detector response curve. The ratio detector has a balanced output circuit.  $C_{106}$  and  $C_{107}$  are the two diode load condensers and the stabilized voltage across the stabilizing condenser  $C_3$  is center-tapped to chassis ground at the junction of the discharge resistors  $R_{107}$  and  $R_{108}$ .

Audio output voltage from the take-off point at the junction of  $C_{106}$  and  $C_{107}$  is coupled through the deemphasis circuit consisting of  $R_{104}$  and  $C_{108}$  to the TV-PHONO switch. On the TV position, terminals 5 and 6 com-

plete the heater circuits to ground for the ratio detector and kinescope, while terminals 2 and 3 connect the audio output of the ratio detector to the volume control  $R_{110A}$ . On PHONO position there is no heater voltage for the ratio detector and kinescope, disabling these tubes, and the audio input voltage from the PHONO jack  $J_{101}$  is coupled to the volume control. The tap at terminal 4 of the volume control allows the tone compensating circuit  $R_{111}$  and  $C_{112}$  to increase the low-frequency response at low volume settings. The audio amplifier section is conventional, with the desired level of audio voltage from the volume control coupled by  $C_{111}$  to the first audio amplifier, which uses the 6AV6 duodiode high- $\mu$  triode tube, followed by a 6K6 audio power output stage to drive the loudspeaker.

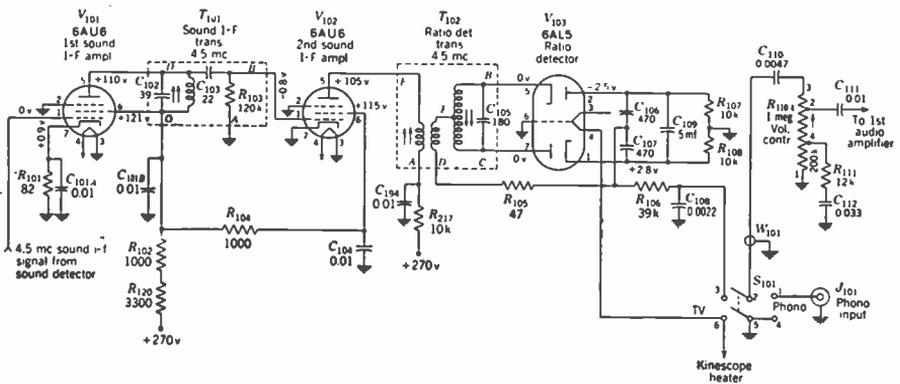


Fig. 22-18. Sound i-f section of intercarrier-sound receiver with ratio detector. Resistance value in ohms, capacitance values less than one in  $\mu\text{f}$  and above one in  $\mu\mu\text{f}$ , unless otherwise noted. (RCA KCS 78H chassis.)

The sound i-f section can be aligned easily without the need for obtaining the visual response curve. Set the signal generator for unmodulated output at 4.5 Mc and connect to the first sound i-f amplifier grid, pin 1 of  $V_{101}$ , where the generator remains for the entire alignment. Connect the d-c voltmeter to the audio take-off point at the junction of  $C_{106}$ ,  $C_{107}$ , and  $R_{106}$  and adjust the ratio detector transformer  $T_{102}$  secondary bottom core for zero. With the secondary balanced, the d-c voltmeter is connected to the 6AL5 pin 2 to indicate stabilizing voltage, so that the  $T_{102}$  primary top core and the single-tuned circuit of  $T_{101}$  can be adjusted for maximum output. The signal generator level should be set for 6 volts on the d-c voltmeter, which is one-half the total stabilizing voltage, for typical operating conditions. In retouching, the final adjustment should be on the ratio detector secondary for zero balance at the audio take-off point.

**22-12. Troubles in the Sound I-F Section.** A trouble only in the sound while the picture is normal, on all channels, indicates a defect in the

sound section after the sound take-off circuit, including the audio amplifier and the sound i-f circuits with the FM detector. The trouble is only in the sound because this part of the receiver does not affect the picture or raster. Since the i-f signal and audio signal are the same for any station, the trouble appears on all channels. In order to localize between the sound i-f circuits and the audio amplifier, audio voltage can be injected into the grid circuit of the first audio stage to check the audio section. If the injected voltage produces normal audio output, the trouble is in the sound i-f section.

*Checking FM Detector Output.* As in an AM detector, the output of the ratio detector or discriminator can be measured with a d-c voltmeter to check whether there is i-f signal input. Note that no B supply voltage is applied to the ratio detector or discriminator. The method of connecting the d-c voltmeter to check the output of a ratio detector or discriminator

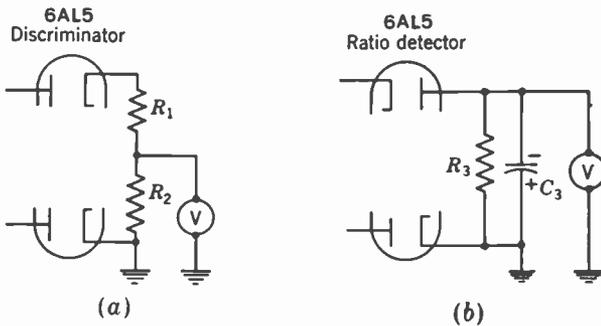


FIG. 22-19. Meter connections to check d-c voltage output of FM detector. (a) Discriminator. (b) Ratio detector.

is illustrated in Fig. 22-19. In *a* the high-impedance d-c voltmeter is connected to the junction of the two cathode load resistors in the discriminator output circuit, to read rectified i-f signal output of one diode. This avoids the possibility of reading zero balance voltage at the audio take-off point. In *b*, the meter is connected to the negative side of the stabilizing condenser in the ratio detector output circuit. With the opposite side of the condenser grounded, in an unbalanced output circuit, the d-c voltmeter reads the total stabilizing voltage produced by rectified i-f signal. When the output circuit is balanced, the meter reads one-half the stabilizing voltage. The normal amount of d-c output voltage with i-f signal input is about 5 to 15 volts.

*Checking Grid-leak Bias.* I-F amplifier stages for an FM signal often employ grid-leak bias, since any amplitude distortions do not affect the desired signal. Therefore, the negative bias voltage across the grid-leak resistor can be measured with a d-c voltmeter to check for rectified signal. The presence of the grid-leak bias indicates the preceding stages are

operating to supply i-f signal input, with a peak value approximately equal to the bias voltage.

#### REVIEW QUESTIONS

1. What are two important differences between an FM receiver and an AM receiver?
2. Give numerical values for the beat frequencies in the output of the mixer showing that the amount of frequency swing of a 41.25-Mc i-f signal in the receiver is the same as the FM sound signal transmitted in channel 2 (54 to 60 Mc).
3. Give three requirements for detecting an FM signal.
4. What is meant by slope detection?
5. Why is AM rejection an essential requirement of an FM receiver?
6. Draw the schematic diagram of a phase-shift discriminator circuit for an i-f center frequency of 41.25 Mc and briefly describe how the circuit operates. Where is the audio take-off point?
7. Draw the schematic diagram of a balanced ratio detector circuit for an i-f center frequency of 4.5 Mc. Indicate the audio take-off point and the stabilizing voltage.
8. Draw the schematic diagram of a grid-leak bias limiter stage for an i-f carrier frequency of 41.25 Mc and explain briefly how limiting is accomplished.
9. Draw a diagram showing how a d-c voltmeter is connected to measure the grid-leak bias on an i-f amplifier stage for the FM sound signal. Why does the presence of the grid-leak bias indicate i-f signal input?
10. In the gated-beam tube limiter and FM detector circuit, where is the i-f signal applied and where is the audio output voltage produced?
11. How is the second sound i-f signal of 4.5 Mc produced in intercarrier sound receivers?
12. Give two reasons why most television receivers use intercarrier sound. Give one disadvantage of intercarrier sound.
13. When aligning by the voltmeter method, what is the required indication when tuning the secondary in a phase-shift discriminator and in a ratio detector? Where is the meter connected in both cases?
14. Show the connections of the d-c voltmeter and balancing resistors for aligning the secondary in a single-ended ratio detector.
15. In a sound i-f circuit consisting of two i-f stages and a ratio detector, describe how to align the entire section:
  - a. By the d-c voltmeter method with a signal generator.
  - b. Using the transmitted associated sound signal.
  - c. By the visual response curve method. Show the required response curves and marker frequencies.
16. Explain how the FM sound signal can produce horizontal sound bars in the picture.
17. Referring to the schematic diagram in Fig. 22-18 give the function of each of the following components:  $R_{101}$  and  $C_{101A}$ ,  $T_{101}$ ,  $C_{101B}$ ,  $C_{103}$  and  $R_{103}$ ,  $R_{104}$ ,  $C_{104}$ ,  $R_{217}$ ,  $T_{102}$ ,  $C_{109}$ , and  $C_{111}$ .
18. The trouble symptom is no sound on any channel with normal picture.
  - a. Why does this indicate trouble in the sound i-f section or audio amplifier?
  - b. Assuming the defect is in the first sound i-f stage, explain how the trouble can be localized to this point by using a d-c voltmeter, in a sound i-f circuit with a ratio detector and two i-f stages having grid-leak bias.

19. Referring to the schematic diagram in Fig. 22-18 describe briefly the effect of the following individual troubles:

- a. TV-PHONO switch on PHONO instead of TV.
- b.  $C_{102}$  in the first sound i-f transformer  $T_{101}$  shorts. What will be the indication on a d-c voltmeter measuring the stabilizing voltage in the ratio detector output circuit and the grid-leak bias voltage on  $V_{102}$ ?
- c.  $R_{217}$  in the plate circuit of  $V_{102}$  opens.
- d.  $R_{104}$  opens.

## CHAPTER 23

### RECEIVER SERVICING

The servicing of television receivers includes antenna installation when necessary, adjustment of the installation or setup controls for the picture tube and receiver chassis, trouble shooting defective circuits, and possible alignment of the r-f and i-f circuits. The trouble shooting is simplified to some extent because in most cases the trouble can be localized by noting how it affects the three receiver indicators—raster, picture, and sound. This applies to all television receivers, since they follow the same general pattern. However, different models vary in details and the manufacturer's schematic diagram with service notes for the receiver will be helpful, especially for alignment.

**23-1. Receiver Installation.** Since the merit of the television receiver is judged from the reproduced picture, the installation must be well done for good results. The antenna and transmission line must supply the receiver enough signal for a good picture with minimum ghosts, noise, and interference. Details of the antenna installation are described in Sec. 21-12. If the receiver's built-in antenna is used, the antenna signal will depend on the location of the receiver. In general, the receiver location should be away from windows where bright light will shine directly on the screen, although some illumination in the room is desirable. Adequate ventilation at the back of the cabinet is necessary because an appreciable amount of heat is produced in the chassis.

Figure 23-1 illustrates a typical arrangement of the receiver installation adjustments on the back of the chassis. Notice that the power input plug at the left is open, as the safety interlock arrangement disconnects the power cord when the back panel is removed. Therefore, a substitute cord with a female plug may be necessary to operate the receiver for making the setup adjustments with the back panel off. A mirror may be needed to view the kinescope screen from the back of the set. In making the setup adjustments, the first requirement is to have brightness and a raster on the kinescope screen. Therefore, the ion-trap magnet is adjusted first in order to direct the electron beam to the screen. This

adjustment is made to obtain the brightest raster. With double-magnet ion traps, the smaller one is toward the front. The position of the ion-trap magnet should be within  $\frac{3}{4}$  in. of the slant cut in a slashed-field type of electron gun or the bend in a bent gun.

With magnetic focusing, the position of the focus coil is set for good focus in the entire raster, as judged by the sharpness of the scanning lines, without any shadowed corners. Best results usually are obtained with the focus magnet about  $\frac{3}{8}$  in. behind the deflection yoke, without

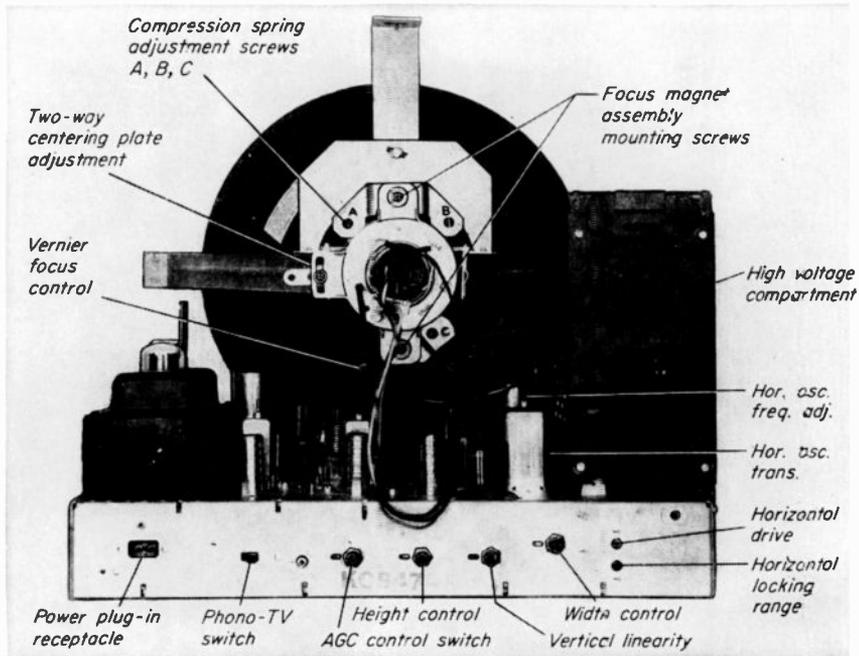
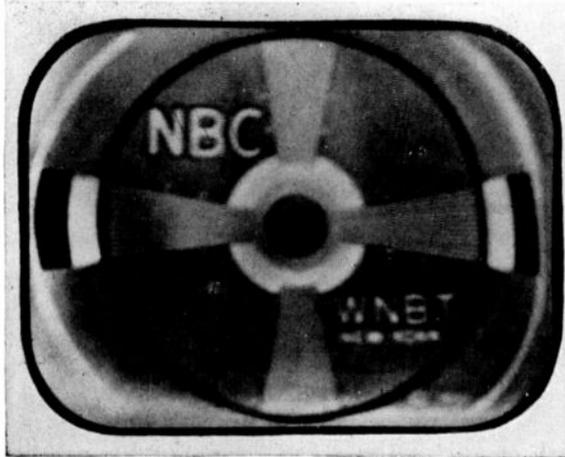


FIG. 23-1. Set-up adjustments at back of receiver chassis. (RCA KCS 47 chassis.)

any tilt, and the kinescope neck centered. Then the fine focus control can be varied for exact focus. It is usually better to check the horizontal detail for sharpest focus, since the spot generally is elliptical rather than circular. Adjust for best focus in the center area of the picture, although the focus will usually be poorer at the edges. Some receivers have a *dynamic focusing* arrangement, which combines scanning current with the direct current in the focus coil in order to correct for the defocusing with deflection. Poor focus in the entire raster is shown in Fig. 23-2a. The shadowed corner in *b* usually means incorrect adjustment of the focus magnet or the ion-trap magnet or both. When all corners are shadowed, this indicates the deflection yoke is too far back from the neck

of the kinescope, allowing the deflected electron beam to strike the sides of the tube.

A tilted raster, as illustrated in Fig. 23-3, is corrected by rotating the deflection yoke slightly in its housing. The raster is centered and the



(a)



(b)

FIG. 23-2. Effects of incorrect focusing adjustments. (a) Scanning lines out of focus. (b) Shadowed corner. (RCA.)

height and width controls adjusted to make the raster fill the screen with the correct aspect ratio of 4:3. With insufficient raster size the screen has blank areas at the sides; more blank area on one side than on the opposite side means the raster is off center, as shown in Fig. 23-4. Check

that the screen is filled with a picture on all channels, since the picture with blanking at the edges is smaller than the unblanked raster and the blanking may vary slightly for different stations. The vertical linearity and height controls are set for the required height with good vertical linearity. The horizontal linearity and width controls are used to obtain the required width and horizontal linearity, after the drive control has been set for maximum drive without a vertical white bar or wrinkle. Linearity is best checked with a picture on the screen, preferably a test pattern or bar patterns. With signal input, the a-g-c level and horizontal a-f-c adjustments can then be set.

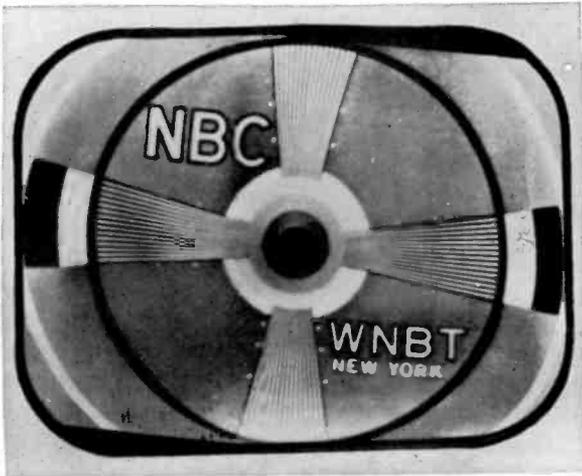
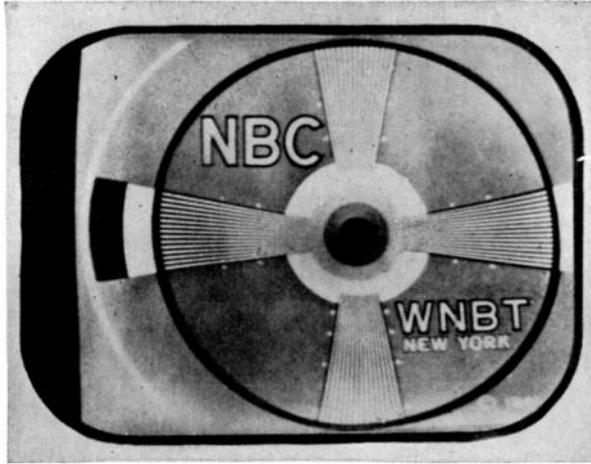


FIG. 23-3. Raster tilted with respect to mask because deflection yoke is tilted. (RCA.)

**23-2. Types of Ghosts.** Although reception of multipath signals by the antenna is the most common cause of ghosts, duplicate images in the picture can also be caused by excessive response in the receiver for the high video frequencies, reflections in a long transmission line due to an impedance mismatch at both ends, or direct pickup of r-f signal by a long transmission line or by the front end.

**Built-in Ghosts.** Multiple images produced by the receiver or by the transmission line are generally called *built-in ghosts*. These can be recognized by the fact that usually there are three, four, or more images uniformly spaced to the right. Excessive response for the high video frequencies in the i-f amplifier or video amplifier causes built-in ghosts on all channels. If the excessive response is in the picture i-f amplifier, varying the fine tuning will affect the ghosts. Ghosts caused by reflections of the antenna signal in a long transmission line change when hand capacitance is added by holding the line. The time delay for reflected

signals on a long transmission line can be calculated on the basis of approximately  $1 \mu\text{sec}$  per 800 ft, noting that a signal reflection travels twice the length of the line, from the receiver input terminals to the antenna and back.



(a)



(b)

FIG. 23-4. Raster off center, (a) horizontally and (b) vertically. (RCA.)

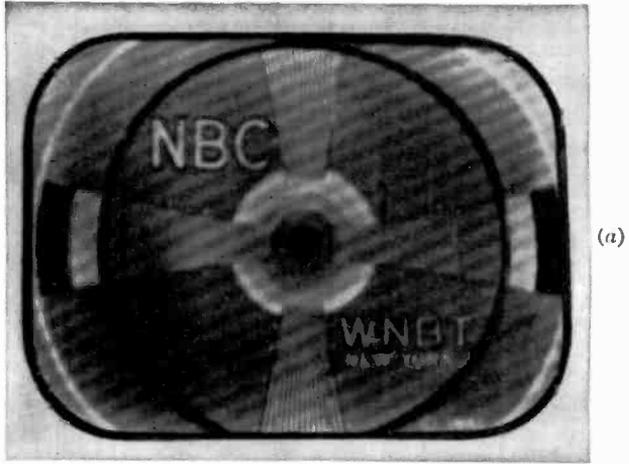
*Leading and Trailing Ghosts.* A ghost due to multipath reception by the antenna is usually to the right of the main image, producing a *trailing ghost*. However, a *leading ghost*, to the left of the main image, can result when the reflected signal is stronger than the direct signal. Another cause of leading ghosts is direct pickup of signal, especially in strong sig-

nal areas, which can provide a picture before the antenna signal delivered by a long transmission line. With direct pickup, the ghost will vary when people walk near the receiver.

*Minimizing Ghosts.* For the problem of multipath reception, an antenna that has a good front-to-back ratio and a narrow forward lobe, with minimum side responses, can be oriented carefully to minimize the ghost. Sometimes, changing the antenna location reduces the intensity of the ghost. Or trying vertical polarization, especially with indoor antennas, may help. It should be noted that multiband antennas generally have a broader directional response and lower front-to-back ratio for channels off the antenna's resonant frequency. Separate antennas cut for the individual channels may be necessary, therefore, to obtain sharp forward lobes in order to eliminate ghosts. Minimizing ghosts caused by direct pickup is a problem of reducing the stray pickup by shielding the transmission line and r-f tuner, if necessary, and supplying more signal from the antenna. The antenna system and the r-f amplifier can be checked for the possibility of a trouble that causes weak antenna signal.

**23-3. R-F Interference.** Interfering r-f signals in the television receiver are heterodyned to produce video frequencies in the output of the second detector that cause interference patterns on the kinescope screen. Typical examples are shown in Fig. 23-5.

*Carrier-wave Interference Pattern.* The uniform diagonal bars in Fig. 23-5a are caused by an unmodulated carrier wave (c-w) signal. The bars are of uniform thickness because there is no modulation of the c-w interference. Usually, the bars shift slowly from one diagonal position, through the vertical, and then to the opposite diagonal, as the interfering carrier wave drifts in frequency. The number of bars and their thickness depend on the beat frequency produced by the c-w interference. As an example, suppose that the frequency of the interfering carrier wave is 1 Mc higher than the picture carrier frequency of the selected channel. The interfering signal will beat with the local oscillator to produce a frequency differing from the picture i-f carrier frequency by 1 Mc, which is within the receiver's i-f range. Then the output of the second detector will include a 1-Mc beat interference in addition to the video signal. With an interfering video frequency of 1 Mc in the kinescope grid-cathode circuit, the interference pattern superimposed on the picture contains approximately 50 pairs of dark and light vertical or diagonal bars. A higher beat frequency produces more bars, which are then thinner. A lower beat frequency results in fewer bars that are thicker. If the beat frequency resulting from the r-f interference is less than 15,750 cps, it will produce uniform horizontal bars. The uniform-bar pattern can be caused by any interfering c-w transmitter but most often the



(a)



(b)



(c)

FIG. 23-5. R-F interference effects in the picture. (a) Uniform diagonal bars caused by unmodulated r-f interference. (b) Horizontal bars caused by amplitude modulation. (c) Herringbone-weave effect caused by frequency modulation. (RCA.)

interference source is local oscillator radiation from a nearby television receiver.

*AM Interference Pattern.* An interfering r-f carrier wave that varies in amplitude with audio modulation produces in the second detector output audio interference in the video signal. This results in a horizontal-bar pattern, as illustrated in Fig. 23-5b, but it should be noted that the number, width, and intensity of the bars will vary with the audio modulation. An example of this effect is sound bars in the picture when the sloping side response of the picture i-f amplifier produces amplitude variations of the associated sound signal corresponding to the audio modulation.

*FM Interference Pattern.* An interfering r-f carrier wave varying in frequency with audio modulation produces in the second detector output interfering beats that vary in frequency with the audio modulation. With a center frequency high enough to produce a fine-line interference pattern, the frequency modulation adds a herringbone weave to the vertical or diagonal bars. When the beat frequencies are too low for a diagonal-line pattern, instead of horizontal bars the FM interference produces a watery effect through the entire picture, which then looks as if it were covered with a shimmering silk gauze. FM interference can be caused by FM broadcast signals, which are generally image frequencies of the selected channel, and by harmonics of the sound i-f signal in the receiver coupled back to the front end.

*Strong R-F Interference.* In addition to the bar patterns in the picture, the light values are altered by the r-f interference when it has enough amplitude to raise the white level of the picture carrier signal closer to the black level. A very strong interfering signal, therefore, can produce a negative picture or black out the picture completely.<sup>1</sup>

*How Interference Enters the Receiver.* An interfering r-f signal can enter the receiver through the antenna and transmission line, direct pickup by the chassis, or from the power line. R-f interference at the antenna input of the receiver must go through the r-f tuner and, therefore, usually appears on specific channels. Direct pickup by the chassis of i-f interference results in the same interference on all channels. If the r-f interference is from the power line, reversing the plug and adding hand capacitance by holding the line should affect the interference. In addition, it should be noted that the receiver itself can produce interference patterns in the picture. Common examples are sound bars in the picture and 4.5-Mc beat in the picture, while c-w interference can be produced by radiation from the power oscillator if the receiver has an r-f high-voltage power supply.

<sup>1</sup> Details of this effect are described in E. W. Herold, Local Oscillator Radiation and Its Effect on Television Picture Contrast, *RCA Review*, March, 1946.

*Local Oscillator Interference.* Radiation of the local oscillator signal from a nearby receiver produces c-w interference on specific channels, usually resulting in the diagonal-bar pattern of Fig. 23-5a, although the picture can be reversed or blacked out completely when the local oscillator radiation is strong. As an example, a receiver with a picture i-f carrier frequency of 25.75 Mc has a local oscillator frequency of 81 Mc when tuned to channel 2. This r-f interference is in channel 5, 76 to 82 Mc, for any receiver regardless of its intermediate frequencies. In this way, a 25.75-Mc i-f receiver can produce local oscillator radiation in other receivers by the combinations of v-h-f channels shown in Table 23-1. Notice that the oscillator produces r-f interference in the channel higher than the selected station by the amount of the intermediate frequency. When set for the higher channel of the combination, the oscillator is also the image frequency of the lower channel, but only in a receiver with the same intermediate frequency. Receivers with a picture i-f carrier of 45.75 Mc do not have a local oscillator frequency in any v-h-f channel.

TABLE 23-1. LOCAL OSCILLATOR INTERFERENCE

Oscillator setting, channel number	Oscillator frequency,* Mc	R-F interference
2	81	Channel 5, 76-82 Mc
3	87	Channel 6, 82-88 Mc
7	201	Channel 11, 198-204 Mc
8	207	Channel 12, 204-210 Mc
9	213	Channel 13, 210-216 Mc

\* For receiver with 25.75-Mc i-f picture carrier.

However, the 45.75-Mc i-f value does allow local oscillator interference in the u-h-f channels, either directly or as image frequencies. The image frequencies are essentially eliminated with an r-f amplifier stage in the receiver, since the image is outside the selected channel frequencies, but interference in the desired channel cannot be filtered out because this would remove the desired signal. Therefore, local oscillator interference in the desired channel can be eliminated only by preventing the oscillator radiation from entering the receiver or by changing the interfering frequency. The change in i-f and oscillator frequencies must be made in the receiver that produces the oscillator radiation, not in the receiver having the interference.

*Diathermy Interference.* Diathermy machines and other medical or industrial equipment usually produce the r-f interference pattern shown in Fig. 23-6. There may be two dark bands across the screen instead of the one shown, and the bars will be darker if the interference is stronger. This r-f interference pattern is produced because diathermy equipment is

effectively a transmitter, often with poor frequency stability and strong 60- or 120-cps modulation from its power supply. Referring to Fig. 23-6, the diagonal lines at the bottom indicate c-w interference, and the weaving effect is caused by frequency variations. The single band shows a strong 60-cps component in the interference; with 120-cps modulation, there would be two bands. The frequency ranges of r-f equipment for industrial, scientific, and medical use are 13.553 to 13.566 Mc, 26.957 to 27.282 Mc, and 40.659 to 40.700 Mc.<sup>1</sup> Since these frequencies are below channel 2, diathermy interference that enters the receiver through the antenna can be eliminated by inserting a high-pass filter in series with the transmission line at the receiver input. The front end in most receivers



Fig. 23-6. Diathermy interference. (Philco Corporation.)

has a high-pass filter cutting off at about 50 Mc but an external filter can be used if necessary.

*FM Broadcast Interference.* An FM interference pattern, as illustrated in Fig. 23-5c, that appears only when the receiver is tuned to channel 2 is usually caused by an FM broadcast station received as an image frequency. This is indicated by the fact that carrier frequencies between 102 and 108 Mc in the FM broadcast band of 88 to 108 Mc are images of channel 2, 54 to 60 Mc, in a receiver with a 25.75-Mc picture i-f carrier. Receivers having the 45.75-Mc picture i-f value do not have any image frequency in the FM broadcast band. FM broadcast interference generally enters the receiver by the antenna. Also, its frequency is not in the television channels. Therefore, a trap in the antenna input circuit can be tuned to reject the interfering frequency and eliminate the interference.

<sup>1</sup> Assigned by FCC July 1, 1952.

*Sound I-F Harmonics.* Harmonics of the associated sound i-f signal in the receiver can cause an FM interference pattern, as illustrated in Fig. 23-5c. The i-f harmonics are usually generated in the last sound i-f stage or in the FM detector. They can be coupled back to the receiver's r-f input by radiation or insufficient bypassing and decoupling. A receiver tuned to channel 6, 82 to 88 Mc, as an example, can have FM interference in the picture from the 85-Mc fourth harmonic of its own 21.25-Mc sound i-f carrier. Other interference combinations are possible, if the harmonics of the receiver's sound i-f carrier frequency are in a television channel or are image frequencies. Sound i-f harmonic interference in the picture can be identified by noting that it varies with the audio modulation of the associated sound signal and will disappear if a sound i-f tube is removed.

*Picture I-F Harmonics.* Harmonics of the picture i-f carrier signal, usually generated in the second detector, also can be coupled back to the front end and cause interference in the picture. This may appear as diagonal lines, depending on the beat frequency, with a grainy effect due to amplitude modulation by the high-frequency components of the video signal. A receiver tuned to channel 5, as an example, can have AM interference from the 77.25-Mc third i-f harmonic of its own 25.75-Mc picture i-f carrier. The frequency of the beat interference will vary widely with a slight change in the oscillator fine tuning control.

*Co-channel Interference.* Stations broadcasting on the same channel are separated by 150 miles or more, but in fringe-area locations between cities co-channel stations can interfere with each other when the ratio of the desired signal to the interference is less than 45 db, approximately. If the interfering signal is strong enough, its picture will be superimposed on the desired picture. In addition, there is usually a bar pattern resulting from the beat between the two picture carrier frequencies. The beat is an audio frequency that produces a horizontal bar pattern, generally called *venetian-blind* effect, which is similar to sound bars in the picture. The remedy for co-channel interference is a more directional antenna, especially with respect to the front-to-back ratio, as interfering co-channel stations are often in opposite directions.

*Adjacent Channel Interference.* Stations that are adjacent in frequency, as well as in number, are separated by 55 miles or more but in some locations adjacent channels in different cities can be received. Adjacent channel interference in the picture usually is from either the picture carrier signal of the upper adjacent channel or the sound carrier signal of the lower adjacent channel. When the picture signal of the upper adjacent channel is strong enough, the side bands corresponding to low video frequencies can beat with the desired picture carrier, producing picture information of the interfering station superimposed on the desired

picture. Most noticeable is the vertical bar produced by horizontal blanking, as it usually drifts from side to side because of the slight difference in horizontal phasing between the two signals. This is generally called the *windshield-wiper* effect. The sound carrier frequency of the lower adjacent channel differs from the desired picture carrier by 1.5 Mc. This beat frequency can produce a diagonal bar pattern of about 100 pairs of black-and-white lines superimposed on the desired picture, with a herringbone weave caused by the frequency modulation of the interfering sound signal. The remedy for adjacent channel interference is a more directional antenna, as for co-channel interference, but in addition the i-f selectivity can be improved by wave traps tuned to reject the adjacent channel intermediate frequencies because they are outside the required i-f pass band.

*Interchannel Interference.* Channels that are not adjacent can interfere with each other, resulting in two pictures, if one is strong enough to produce cross modulation in the r-f amplifier. In addition, specific combinations of channels can interfere with each other by double conversion in the front end.

*Radio Broadcast Interference in Television Receivers.* If the high-pass filter in the front end has insufficient attenuation for the radio broadcast band of 535 to 1,605 kc, a strong local radio station can produce cross modulation, resulting in a diagonal bar pattern on the kinescope screen, corresponding to the frequency of the carrier wave.

*Television Receiver Interference in Radio Receivers.* Harmonics of the 15,750-cps horizontal deflection current in the television receiver can cause beat-frequency whistles in nearby radio receivers. The whistles appear every 15 kc with maximum points every 70 to 100 kc, approximately, on the dial of the radio receiver, only when the television receiver is on, producing horizontal deflection. The interference is often coupled to the radio receiver through the power line, from the damper tube's filament winding in the power transformer. The radio receiver picks up the interference because of the loop antenna's capacitance to ground. This interference can be reduced by inserting an r-f filter in series with the power cord, preferably close to the television receiver chassis. Inductive coupling between the radio's loop antenna and the television receiver's deflection yoke and horizontal output transformer, which is effective up to about 30 ft, can be minimized by shielding the output transformer, grounding all the mounting hardware for the kinescope and deflection yoke, and shortening exposed leads in the deflection circuits. Sixty-cycle buzz at 15-kc intervals on the radio dial is caused by the vertical blanking pulses in video signal coupled from the television receiver; this can be minimized by shortening exposed leads in the video amplifier circuits.

**23-4. External Noise Interference in the Picture.** Noise pulse voltages from automobile ignition systems and motors produce short horizontal black streaks in the picture, as illustrated in Fig. 23-7. The streaks are black as the interfering noise voltage raises the carrier amplitude during the interval of the pulse width, which is shorter than the horizontal line-scanning time. Often the black streaks have a *white tail* smeared to the right, because of phase distortion in the receiver or noise setup in the *RC* grid coupling circuits. The picture may also skip frames vertically and tear apart horizontally if the noise is strong enough to interfere with synchronization. Similar effects in the picture can be caused by elevator motors, neon signs, cash registers, or any device that produces sparking, which generates r-f energy in the v-h-f range modulated at the

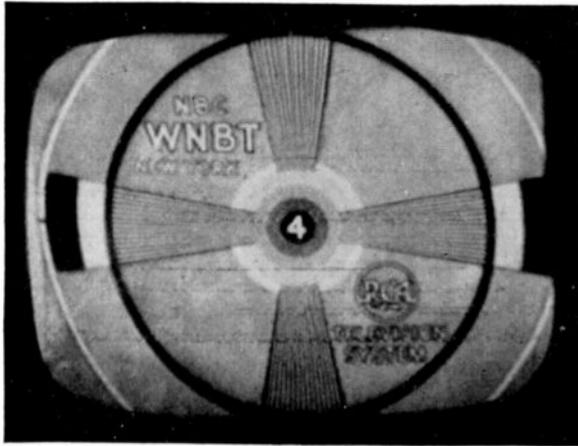


FIG. 23-7. Short, horizontal black streaks caused by auto-ignition interference. (From McGraw-Hill Motion-picture Series, *Basic Television*.)

sparkling rate. However, a device operated from the 60-cps power line produces noise streaks in a cluster that stays still like a horizontal hum bar, while ignition noise streaks occur at random throughout the picture.

External noise interference can enter the receiver as pickup by the antenna or by an unshielded transmission line, direct pickup by the chassis, or through the a-c power line. Minimizing the interference effect in the picture is a problem of supplying more antenna signal or reducing the noise pickup, or doing both, to provide a suitable signal-to-noise ratio. This may be accomplished by using a high-gain directive antenna, with vertical stacking. Moving the antenna out of the noise field, by either increasing the antenna height or just finding an antenna placement farther from the noise source, is often helpful. It may be necessary to use shielded transmission line to prevent noise pickup by the line. To reduce

pickup from the power line, a low-pass filter can be used, consisting of 1-mh r-f chokes in series in each side of the line and 0.01- $\mu$ f bypass condensers across the line. Connect the filter at a point closest to the interference source in order to minimize radiation from the power line.

**23-5. Sound in the Picture.** The associated sound signal can produce horizontal sound bars corresponding to the audio modulation in the picture, as shown in Fig. 19-13, or the fine beat pattern illustrated in Fig. 23-8 caused by the 4.5-Mc beat between the picture carrier and the sound carrier frequencies.

*Sound Bars.* Slope detection converts the modulation of the FM sound signal to audio voltage in the second detector output, which is coupled to the kinescope grid-cathode circuit to produce the horizontal sound bars. The sound bars can be recognized as they vary with the audio modulation

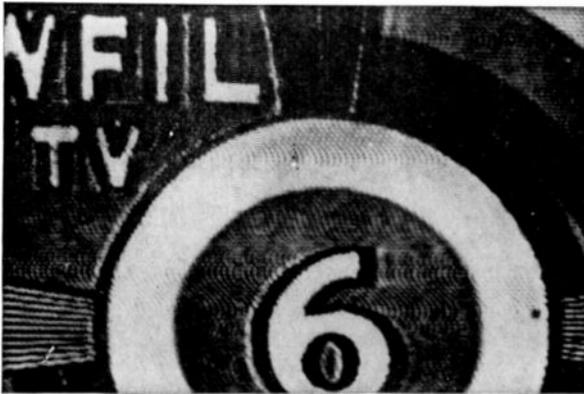


FIG. 23-8. 4.5-Mc beat in the picture. (RCA.)

and disappear when there is no voice. The cause of sound bars is incorrect response for the i-f sound signal in the picture i-f amplifier. This can often be corrected by tuning the associated sound traps in the i-f amplifier to eliminate the sound bars. When the sound bars are present only at high volume levels, this indicates microphonics caused by the vibrating loudspeaker. The microphonics may be in the local oscillator, r-f, i-f, or video amplifier. Sound bars only at high volume levels can also be the result of insufficient filtering of the plate-supply voltage in the audio output stage.

*4.5-Mc Beat.* The 4.5-Mc output of the second detector has the FM sound signal variations. Coupled to the kinescope grid-cathode circuit, the 4.5-Mc signal produces a fine beat pattern in the picture consisting of about 225 pairs of thin black-and-white lines with small "wiggles" like a fine herringbone weave. This is illustrated in Fig. 23-8. The fine lines are caused by the 4.5-Mc carrier, while the herringbone weave is the

result of the frequency variations in the FM sound signal. The 4.5-Mc beat produced by the associated sound signal can be recognized by observing the pattern closely to see that the herringbone effect disappears when there is no voice, leaving just the straight-line pattern corresponding to the c-w interference of the 4.5-Mc carrier without modulation.

Excessive 4.5-Mc sound signal at the kinescope is the result of insufficient rejection in the video amplifier circuits. The 4.5-Mc traps in the video amplifier can be tuned for minimum 4.5-Mc signal at the kinescope grid-cathode circuit and minimum beat interference in the picture. If necessary, a 4.5-Mc trap can be added to the video amplifier, as illustrated in Fig. 23-9. The parallel resonant trap circuit is inserted in

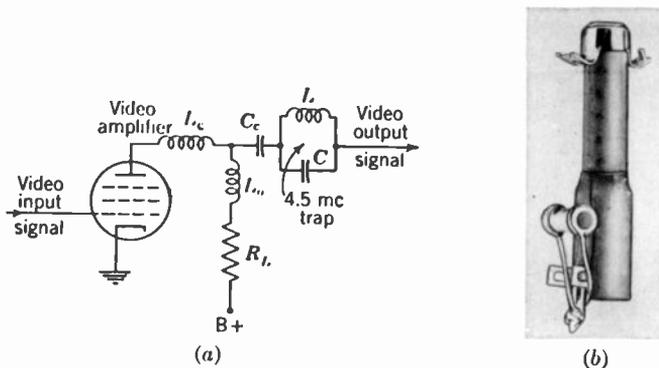


FIG. 23-9. 4.5-Mc parallel resonant trap circuit in the video amplifier. (a) Trap connected in series with output circuit. (b) Photo of typical trap. (RCA.)

series with  $C_c$  and  $L_c$ , on either side of the coupling condenser or peaking coil.

**23-6. Localizing Hum Troubles.** Table 23-2 indicates how different symptoms of hum on the kinescope screen can be distinguished to localize the trouble. Hum in the picture means the symptom is present only when a picture is on the screen, but disappears with just a raster. The picture can be removed by switching to an unused channel or shorting the antenna terminals to see the raster alone. In order to see whether bend is in the picture or raster, shift the raster off center horizontally and increase the brightness to see the side of the raster and edge of the picture where horizontal blanking begins. It should be noted that 60-cycle hum produces one pair of hum bars or one sine-wave bend from top to bottom on the screen, while 120-cycle hum results in two pairs of bars or two cycles of bend. Heater-to-cathode leakage in a tube introduces 60-cycle hum but 120-cycle hum is caused by excessive ripple in the B supply voltage from a full-wave power supply. Hum caused by ripple in the B supply voltage often causes multiple symptoms, including hum in the sound.

TABLE 23-2. LOCALIZING HUM TROUBLES

Symptom	Cause	Source
Hum bend in picture but not in raster	Hum in H sync	H sync circuits, if no hum bars R-F, i-f, and video circuits, if hum bars also
Hum bend in raster and picture	Hum in H deflection	H oscillator, amplifier, or damper
Hum bars in picture but not in raster	Hum in picture signal	R-F, i-f, and local oscillator. Can have bend also
Hum bars in raster and picture	Hum in video signal	Video amplifier. Can have bend also

**23-7. Buzz in the Sound.** Compared with hum, buzz is a raspy sound at the 60-cycle rate of the vertical blanking and sync pulses that is not as smooth as 60-cycle sine-wave hum and has a lower tone than 120-cycle hum. In addition, hum is often accompanied by visible symptoms in the picture or raster, but buzz in the sound generally does not affect the picture. The buzz is usually caused by picture signal in the sound or by vertical deflection pulses coupled into the audio circuits.

*Intercarrier Buzz.* As noted in Sec. 22-9, this type of buzz is produced in intercarrier-sound receivers because of excessive amplitude modulation of the 4.5-Mc sound signal by the vertical blanking pulses and insufficient AM rejection in the FM sound circuits.

*Cross-modulation Buzz.* Overload in any common stage for the picture and sound signals can produce cross modulation that allows the vertical blanking pulses to produce buzz in the sound. The cross-modulation buzz can be introduced in the r-f amplifier, the common i-f amplifier stages, and in the video amplifier of intercarrier-sound receivers where this section is used to amplify the 4.5-Mc second sound i-f signal. It may be helpful to note that the buzz varies with picture content, showing it is caused by picture signal in the sound, and the buzz is louder on strong signals, which indicates overload.

*High-voltage Buzz.* Because of the poor regulation of a flyback high-voltage power supply, the kinescope anode voltage varies at the frequency of the vertical blanking pulses. The anode voltage is high with minimum anode current when the beam is cut off during vertical blanking time. With active scanning of white picture information, the average beam current is greater and the kinescope anode voltage drops. Since the high-voltage filter condenser cannot eliminate the 60-cps ripple, the kinescope anode voltage has variations at the vertical blanking rate. As a result, any stray capacitive coupling from the kinescope anode circuit to the audio input circuit produces buzz in the sound corresponding to the 60-cps vertical blanking pulses. In addition to the fact that it depends upon

the amount of white information in the scene, high-voltage buzz varies with the setting of the brightness control and disappears when the kinescope anode voltage is temporarily disabled.

*Vertical-deflection Buzz.* Any coupling between the vertical scanning circuits and the audio section can allow the vertical deflection pulses generated in the receiver to produce 60-cps buzz in the sound. The vertical deflection buzz is present on all channels, independent of picture information, varies in tone with rotation of the vertical hold control, and disappears when the vertical oscillator is temporarily disabled.

*Buzzing Transformers.* Mechanical vibrations in the vertical output transformer, vertical oscillator transformer, or the power-supply transformer and filter choke can produce 60-cycle buzz. When the tone varies with rotation of the vertical hold control the buzz is in the vertical deflection circuits. Vibration of the horizontal output transformer causes a very high pitched 15,750-cps singing sound.

**23-8. Localizing Receiver Troubles.** The key to quick efficient trouble shooting in the television receiver is the ability to use the raster, picture, and sound as indicators to localize the trouble to a particular section. This procedure is illustrated by the trouble-shooting charts in Tables 23-3 to 23-6. The analysis is based on a typical receiver circuit using intercarrier sound with the 4.5-Mc sound take-off circuit in the video detector output. Table 23-3 illustrates troubles that affect more than just one indicator. As examples, the low-voltage power supply can produce troubles in the raster and signal; the circuits common to the picture and sound signals affect both the picture and sound.

Table 23-4 lists common raster troubles. These are also evident in the picture, but the fact that the trouble is in the raster circuits can be determined by removing the picture to see the trouble symptom in the raster alone. This table also includes troubles evident in the picture but caused by a defect in the raster circuits. For instance, crowding and spreading in the picture are the result of nonlinear scanning. Table 23-5 indicates common types of trouble that are in the picture but not in the raster or in the sound. Therefore, the trouble is in a circuit for the picture alone. Sync troubles are considered as troubles in the picture, rather than the raster, when the picture can be made to hold by varying the deflection oscillator frequency. This indicates the oscillator operates normally but there is trouble in the synchronizing pulses obtained from the composite video signal. If the picture does not hold at any setting of the frequency control, the trouble is incorrect frequency of the deflection oscillator, which is a defect in the raster circuits. Troubles in the sound alone must be in the 4.5-Mc second sound i-f circuits or the audio section, as indicated in Table 23-6, when the trouble is the same on every channel.

TABLE 23-3. RECEIVER TROUBLES

Trouble Indication	Analysis
Set completely dead. No raster, picture, or sound	Check power line into receiver, safety interlock switch, and power-line fuse if any. Possible failure in low-voltage power supply. If tube filaments light, receiver has input power
No sound and no picture but raster is normal (see Fig. 8-4c)	Power supplies and deflection circuits operating normally to produce raster. Trouble in signal circuit common to picture and sound. Check r-f tuner, common i-f amplifier, second detector. Check a-g-c circuit for excessive negative bias cutting off signal stage. Check video amplifier directly coupled to a-g-c tube
Trouble on one channel, but operation normal on other channels	Check front end. Incorrect local oscillator frequency. Dirty contacts on station-selector switch. Poor antenna signal on one channel (see Sec. 20-10)

TABLE 23-4. RASTER TROUBLES

Trouble Indication	Analysis
No brightness at all on screen. Sound is normal.	No high voltage; check high-voltage circuits and horizontal deflection circuits with flyback supply; check fuse in horizontal output circuit. Kinescope cut off by excessive negative bias. Check kinescope, filament, cathode, control grid, and accelerating grid voltages. Check ion-trap magnet adjustment
Insufficient brightness	Insufficient high voltage. Excessive negative bias on kinescope. Check kinescope low-voltage operating potentials. Check ion-trap magnet adjustment. Low emission in kinescope. Try filament voltage booster or new kinescope. Dust on kinescope faceplate can reduce brightness appreciably
Raster too small, horizontally and vertically	Low B supply voltage. Check line voltage and low-voltage rectifier (see Sec. 10-7)
Raster too big. Cannot be reduced to normal size with height and width controls.	Kinescope anode voltage too low
Raster too narrow. Insufficient width (see Fig. 17-22)	Insufficient horizontal deflection amplitude; check horizontal oscillator, amplifier, and damper stages. Low B supply voltage (see Sec. 18-14)
Raster too small vertically. Insufficient height (see Fig. 17-21)	Insufficient vertical deflection amplitude; check vertical oscillator and output stages (see Sec. 18-14)
Only a bright horizontal line on screen (see Fig. 8-4b)	No vertical deflection; check vertical oscillator and output stages (see Sec. 18-14).
Raster tilted with respect to mask (see Fig. 23-3)	Rotate deflection yoke
Raster off center, vertically or horizontally (see Fig. 23-4)	Check centering adjustments and tilt of focus magnet

TABLE 23-4. RASTER TROUBLES (Continued)

Trouble Indication	Analysis
Trapezoidal raster (keystoning) (see Fig. 18-41)	Defective deflection yoke
Pincushion or barrel distortion of raster	Distortion caused by deflection yoke or misadjustment of correction magnets
Shadowed corner in raster (see Fig. 23-2b)	Check focusing magnet and ion-trap magnet
Entire picture out of focus (see Fig. 23-2a)	Check focusing magnet and ion-trap magnet; check focusing voltage or current adjustment when provided; focus control should be effective on both sides of best focus; if focus control has no effect, picture tube may be gassy
Blooming with defocusing at high brightness levels	Poor high-voltage regulation; weak high-voltage rectifier. Kinescope grid driven positive; check kinescope bias
White vertical bars at left side of raster (see Fig. 18-36)	Ringling in horizontal output circuits (see Sec. 18-14)
Wide bright bar at left side of raster, with fold-over and reduced width (see Fig. 18-37)	Insufficient damping in horizontal output circuit; check damper tube and circuit (see Sec. 18-14)
White vertical bar at left or right side of raster	Crowded nonlinear horizontal scanning
White vertical bar in middle of raster (see Fig. 18-23)	Excessive horizontal drive; readjust drive control; check input voltage and bias on horizontal amplifier
White horizontal bar at top or bottom of raster (see Fig. 18-38)	Crowded nonlinear vertical scanning; check height and vertical linearity controls; trouble in vertical oscillator or amplifier (see Sec. 18-14)
Fold-over in picture at left or right edges of raster	Horizontal retrace time too long, or incorrect phasing of retrace with respect to blanking (see Sec. 18-14)
Crowding and spreading of picture at left and right sides of raster (see Fig. 18-26)	Nonlinear horizontal scanning; check width, drive, and linearity controls. Weak horizontal oscillator, amplifier, or damper
Crowding and spreading of picture at top and bottom of raster (see Fig. 18-5)	Nonlinear vertical scanning; check height and linearity controls; weak vertical oscillator or amplifier
Bend at sides of raster, with both sides having same curve (see Fig. 18-35a)	Additive hum in horizontal scanning circuits displacing scanning lines at hum frequency; two sine waves along each edge indicate 120-cps hum from low-voltage supply
Bend at sides of raster, with both sides having opposite curve (see Fig. 18-35b)	60-cps or 120-cps hum modulation of horizontal scanning signal; check heater-cathode leakage in horizontal oscillator, amplifier, or damper tube for 60-cps hum
Thin black vertical bar at left side (see Fig. 18-40)	Barkhausen oscillations in horizontal output tube; reduce horizontal drive or change tube. Can also be harmonics of damper current pulse (see Sec. 18-14)

TABLE 23-4. RASTER TROUBLES (Continued)

Trouble Indication	Analysis
Wide dark vertical bars just barely visible at left side, with slight ripple at top	Crosstalk in deflection yoke coupling damped oscillation in horizontal output circuit into vertical scanning coils; check for open damping resistors across vertical coils
Reversed direction of scanning, right to left or bottom to top	Reverse connections to horizontal or vertical deflection coils
Raster jitters vertically or horizontally	Electrical centering control noisy, causing noise in direct current through scanning coils of deflection yoke; when raster is steady but picture jitters, trouble is in sync
Two short full-width pictures—one above the other—separated by black horizontal bar	Vertical oscillator frequency one-half correct value; black bar is produced by vertical blanking; check for increased $R_p C_c$ in grid circuit of vertical oscillator
Bottom half of picture superimposed on top half. Part of picture looks inverted vertically	Vertical oscillator frequency twice correct value; check for reduced $R_p C_c$ in grid circuit of vertical oscillator
Two narrow full-height pictures side by side separated by black vertical bar	Horizontal oscillator frequency one-half correct value; black bar is produced by horizontal blanking; check for increased $R_p C_c$ in grid-circuit of horizontal oscillator; check horizontal a-f-c circuit
Right half of picture superimposed on left half. Part of picture looks inverted horizontally	Horizontal oscillator frequency twice correct value; check for reduced $R_p C_c$ in grid circuit horizontal oscillator; check horizontal a-f-c circuit

TABLE 23-5. PICTURE TROUBLES

Trouble Indication	Analysis
No picture. Normal raster and normal sound (see Fig. 8-4c)	Trouble in circuit only for picture—video amplifier in most receivers
Weak picture with low contrast (see Fig. 11-25)	Insufficient video signal amplitude for kinescope grid; check for low gain in video amplifier, i-f amplifier, or r-f section. Presence of snow indicates weak r-f signal
Snowy picture with little contrast (see Fig. 20-5)	Insufficient antenna signal; weak r-f amplifier (see Secs. 20-9 and 20-10)
Very dark picture or reversed picture, out of sync (see Fig. 15-8)	Overloaded picture; excessive antenna signal, or insufficient grid bias; check a-g-c circuit (see Sec. 15-6)
Ghosts in picture (see Fig. 21-4)	Change antenna orientation and try more directional antenna. Can be built-in ghost or result of direct pickup (see Sec. 23-2)
Horizontal sound bars in picture (see Fig. 19-13)	Incorrect alignment of sound traps in picture i-f amplifier. Check adjustment of oscillator fine tuning control. Microphonics at high-volume levels (see Sec. 23-5)

TABLE 23-5. PICTURE TROUBLES (*Continued*)

Trouble Indication	Analysis
4.5-Mc beat in picture (see Fig. 23-8)	Incorrect adjustment of 4.5-Mc trap in video amplifier. Check adjustment of oscillator fine tuning control (see Sec. 23-5)
Coarse herringbone pattern in picture (see Fig. 23-6)	Diathermy interference (see Sec. 23-3)
Diagonal bars in picture (see Fig. 23-5a)	R-F interference (see Sec. 23-3)
Shimmering effect in picture (see Fig. 23-5c)	FM interference (see Sec. 23-3)
Horizontal black streaks in picture (see Fig. 23-7)	Ignition noise or motor noise (see Sec. 23-4)
Streaks in picture with antenna disconnected	Arcing in high-voltage supply. Intermittent coupling condenser in i-f or video amplifier
Picture intensity fades in and out	Check for loose transmission line that flaps in the wind (see Sec. 21-14)
Picture background does not follow changes in scene brightness. Retrace lines show on dark scenes if receiver does not have internal vertical blanking	Check d-c restorer circuit (see Sec. 13-7)
Picture rolls vertically and slips horizontally	Insufficient vertical and horizontal sync; check common sync circuits; check video signal for normal sync (see Secs. 16-9 and 16-10)
Picture rolls vertically. Can be stopped but does not hold (see Fig. 16-1). Horizontal hold normal	Insufficient vertical sync; check vertical sync circuits (see Secs. 16-10 and 17-13)
Picture slips horizontally and tears into diagonal segments. Can be stopped but does not hold (see Fig. 16-2). Vertical hold normal	Insufficient horizontal sync or trouble in horizontal a-f-c circuit (see Secs. 16-10 and 17-13)
Bend in picture but not in raster (see Fig. 16-28)	Hum in horizontal sync, or weak horizontal sync (see Sec. 16-10)
Bend only at top of picture (see Fig. 16-26)	Weak horizontal sync just after vertical blanking (see Sec. 16-10).
Scalloped effect with ripples at vertical edges in picture; or picture completely torn apart with scalloped sides	Horizontal a-f-c circuit oscillating. Check a-f-c and oscillator frequency adjustments; a-f-c filter time constant too short
Picture does not hold horizontally when changing channels	Insufficient pull-in range for horizontal a-f-c circuit. Check a-f-c adjustments; a-f-c filter time constant too long
Picture jitters vertically or horizontally but not raster	Intermittent vertical or horizontal synchronization. Check oscillator transformer. Horizontal jitter in picture can also be caused by a-f-c circuit

TABLE 23-5. PICTURE TROUBLES (Continued)

Trouble Indication	Analysis
Fold-over in picture only, at left or right	Incorrect phasing between horizontal retrace and blanking (see Sec. 17-12)
Poor interlacing. Moire or fishtailing effect in side wedge of test pattern. Vertical resolution poor. Wide black spaces between white scanning lines caused by pairing (see Fig. 4-9)	Check vertical integrator circuit; camera signal variations in vertical sync; horizontal deflection pulses in receiver coupled into vertical sync circuits (see Sec. 16-10)
Picture not clear, with insufficient detail but focus is good. Vertical edges between black and white not distinct. Less resolution in top and bottom wedges of test pattern than in side wedges (see Fig. 11-10)	Insufficient response for high video frequencies. Bandwidth too narrow in r-f and i-f response; insufficient high-frequency response in video amplifier; video detector or amplifier load resistance too high (see Secs. 11-12 and 19-7)
Reversed white outlines trailing to the right of black edges in picture (see Fig. 11-29)	Excessive response for high video frequencies in i-f or video amplifier; video detector or amplifier load resistance too low (see Secs. 11-12 and 19-7)
Large-area smear or streaking in the picture (see Fig. 11-28)	Time-delay distortion for low video frequencies. Picture i-f carrier too high on response curve, check i-f alignment and local oscillator tuning. Video detector or amplifier load resistance too high (see Secs. 11-12 and 19-7)
One or two pairs of wide black-and-white bars across the picture	Hum in the video signal. One pair is 60-cps hum; two pairs 120 cps. Bars in raster additive hum introduced in video amplifier. Bars only in picture is modulation hum from r-f or i-f section (see Sec. 11-11)

TABLE 23-6. SOUND TROUBLES

Trouble Indication	Analysis
No sound. Picture normal	Check 4.5-Mc sound i-f section and audio amplifier
Weak sound	Check 4.5-Mc sound i-f section and audio amplifier
Distorted sound	Check FM detector alignment and audio section With ghost in picture, multipath reception can cause garbled sound
Hum in sound only	Hum just with signal is modulation hum introduced in 4.5-Mc sound i-f section. Hum with or without signal is additive hum from audio section
Buzz in sound	Check AM rejection circuit and FM detector alignment for balance. If picture is black and out of sync, buzz caused by overload in common stage for picture and sound. Can be high-voltage buzz or vertical deflection buzz (see Sec. 23-7)

**23-9. Test Equipment.** The following items of test equipment are used in the servicing of television receivers:

*Volt-ohmmeter.* Its input resistance should be 20,000 ohms per volt or more on the d-c ranges so that the meter will not be an appreciable load on the circuit being tested. The ohmmeter ranges should extend to 20 megohms or higher. The usual d-c voltage ranges up to about 600 volts are adequate for measuring B+ and bias voltages. To measure the kinescope anode voltage a high-voltage probe, with a built-in voltage divider, can be used. The a-c voltage ranges can be used to measure filament voltages and audio signal. Many multimeters have an r-f detector probe for r-f voltage measurements, which can be used to measure a-c picture signal and video signal voltages.

*Oscilloscope.* This is used for visual response curves of the r-f and i-f circuits, to check wave shapes when trouble shooting the sync and deflection circuits and for checking the video signal voltage in the detector and video amplifier. A conventional oscilloscope with frequency response up to 100 to 500 kc in the vertical amplifier and an internal sweep frequency range of up to 15,750 cps, or more, is suitable. For visual response curves, good low-frequency response is most important to show the rectified sweep generator signal, which is a d-c voltage varying at 60 cps. The repetition rate of the sync, deflection, or video signal voltages is either 60 cps at the vertical field rate or 15,750 cps at the horizontal line rate. However, if it is desired to check the sharp changes in the sync and deflection voltages or observe the full frequency range of the video signal, the high-frequency response of the oscilloscope should extend to 4 Mc or more. Figure 23-10 shows a typical oscilloscope. Its frequency response is from 20 cps to 100 kc for the vertical amplifier and the frequency range of the internal sweep for horizontal deflection is 2 to 30,000 cps. The Y input is for the vertical amplifier; the X input is for the horizontal amplifier when the internal sweep generator is not used. Provision is included for voltage calibration so that the waveform on the screen can be measured in peak-to-peak voltage amplitude.

*Sweep Generator.* Center frequency for the FM sweep generator should be adjustable over the range of about 20 to 45 Mc so that it can be used for visual alignment of the receiver's i-f amplifiers. It is desirable that the sweep generator also cover the frequency range of the television channels for use in aligning r-f stages. The sweep width or frequency swing of the generator's FM output should be adjustable to about 10 Mc or more. A frequency swing of about 10 Mc is necessary to obtain the complete i-f response curve. For the sound i-f response a sweep width of about 1 Mc is enough. Crystal-calibration facilities for accurate marking of frequencies on the visual response curve may be included in the sweep generator, or a crystal-calibrated r-f signal generator can be used as a marker generator.

*Signal Generator.* This is the common test oscillator often used in trouble shooting and aligning radio receivers. It usually provides unmodulated r-f output over a wide frequency range, 400-cps audio output, or r-f output amplitude-modulated with the 400-cps audio voltage. For television applications, the r-f range should extend at least up to about



FIG. 23-10. Typical 5-in. oscilloscope. Frequency response of vertical amplifiers is from d-c to 100 kc within 1 db. Internal-sweep frequency range 2 to 30,000 cps. (Allen B. DuMont Laboratories, Inc.)

50 Mc for the intermediate frequencies. Crystal-controlled calibration of the frequencies is required if the generator is used as a marker.

Signal injection through the television receiver circuit can be accomplished with the signal generator, as in radio receivers, working back toward the antenna stage by stage to localize a defective stage. The signal generator can also be used for producing bar patterns on the kinescope screen to test scanning linearity without a picture. These techniques are described in Sec. 23-11.

*Additional Equipment.* Several other items of test equipment can be helpful in specific applications. A square-wave generator can be used for

quick and accurate checks on the frequency and phase angle response of the video amplifier. A video sweep generator with a range from very low frequencies up to 4 Mc can be used to obtain the visual response curve of the video amplifier. This provision is included in some r-f sweep generators. A low-capacitance r-f detector probe that can be easily moved to different



FIG. 23-11. Crosshatch generator for producing bar patterns to test and adjust scanning linearity. (Hickok Electrical Instrument Company.)

circuits is helpful when working in the signal frequency circuits. The probe produces d-c output corresponding to the a-c input. Multi-meters with provisions for measuring r-f voltage include the probe for r-f and i-f signal measurements. An auxiliary input lead with an r-f detector probe for the oscilloscope enables it to provide visual response curves for individual r-f, i-f, or video stages, independent of the video detector. A crystal-controlled frequency meter or a sensitive and accurate wave-

meter can be used to measure the frequency of the local oscillator during alignment and can also be used for calibrating the signal generator. The *crosshatch* generator shown in Fig. 23-11 produces on the kinescope screen a vertical and horizontal bar or dot pattern simultaneously, which is convenient for testing and adjusting the scanning linearity accurately.

**23-10. Calibrating the Oscilloscope.** In addition to being thoroughly familiar with operation of the oscilloscope, the serviceman should be able to read the screen pattern in terms of voltage, amplitude, and frequency, because quantitative measurements can be helpful. Fundamentally, the calibration procedure compares the amplitude or frequency of the signal voltage being measured to a known voltage, using the oscilloscope screen pattern to indicate the comparison.

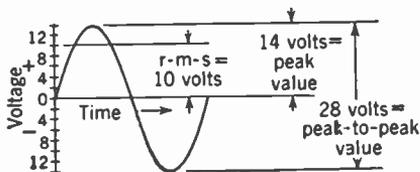


FIG. 23-12. The peak-to-peak sine-wave voltage is 2.8 times the r-m-s value.

**Voltage Measurements.** As an example of voltage calibration, suppose that a saw-tooth deflecting voltage connected to the vertical binding posts of the oscilloscope for measurement produces a vertical deflection on the oscilloscope screen of 2 in. peak to peak. No horizontal deflection is used for the oscilloscope, so there will be just an easily measured vertical line on the screen. The setting of the oscilloscope vertical gain must not be changed, because the amount of vertical deflection varies with the gain of the vertical amplifiers. The signal voltage being measured can then be disconnected and sine-wave voltage from the 60-cps power line with a value that produces a vertical deflection of 1 in. peak to peak can be connected in its place. Assume that the sine-wave voltage required for the 1-in. deflection is 10 r-m-s volts as measured on a conventional a-c voltmeter. As shown in Fig. 23-12, the peak-to-peak voltage swing for this sine-wave voltage is 28 volts. Therefore, it is known that a vertical deflection of 1 in. corresponds to 28 volts peak to peak. Therefore, the saw-tooth voltage producing a deflection of 2 in. with the same setting of the vertical gain control has a peak-to-peak value of 56 volts.

The voltage calibration applies over the frequency range of the oscilloscope vertical amplifiers. However, the calibration applies only for the one setting of the vertical gain control. In many cases the voltage amplitudes measured require different adjustments of the oscilloscope gain in order to obtain screen patterns of reasonable size. Consequently, it will prove convenient to calibrate the oscilloscope for about 5 different settings of the vertical gain control, keeping the data in graphical or tabular form, so that voltage amplitudes can be read directly from the screen pattern.

*Current Measurements.* Although the oscilloscope is a voltage-operated device, current waveforms can be observed and peak-to-peak measurements made the same way. This can be accomplished by inserting a known resistance in series with the circuit in such a manner that the current to be measured flows through the resistor. The voltage across this resistor is then coupled to the vertical binding posts of the oscilloscope so that its wave shape can be observed and peak-to-peak amplitude measured as usual. With the voltage measured, the current can be calculated as  $E/R$ , where  $R$  is the known value of resistance. The current amplitudes are then in terms of peak-to-peak values, like the voltage amplitudes. The current wave shape is identical with the observed voltage wave shape, since current and voltage are in phase with each other in the resistive element and there is no change in wave shape.

The value of known resistance to be inserted for the current measurement depends on the circuit. It should be as small as possible so that it will not change the current in the circuit appreciably but must be large enough to provide sufficient voltage for the oscilloscope input. About 1 volt is generally sufficient, and the known resistance can have a value that is 5 to 10 per cent or less of the circuit resistance. In high-current circuits such as the horizontal scanning coils a resistance of 10 ohms is enough.

*Frequency Comparisons.* In order to check frequencies with the oscilloscope, the unknown signal voltage is compared to a voltage source the frequency of which is known. Suppose that a deflection signal from the television receiver is coupled to the vertical input of the oscilloscope, and the oscilloscope sweep frequency and synchronizing controls are adjusted to produce five complete cycles on the screen. Use as little synchronizing voltage as possible in locking the picture into a stationary pattern on the screen. Do not change the settings of these sweep controls. The signal from the receiver can then be disconnected, and the voltage output of an audio signal generator having an accurate frequency calibration is connected to the oscilloscope input. Adjust the signal-generator frequency until five complete cycles are obtained. The unknown frequency is then equal to the signal-generator frequency, which we will take as 15,000 cps as an example. If it is more convenient, the signal-generator frequency can be adjusted to produce one complete cycle in the screen pattern. This will occur when the generator frequency is 3,000 cps in this case, and the unknown frequency is  $5 \times 3,000$ , or 15,000 cps, since it produces five complete cycles with the same setting of the oscilloscope sweep frequency. The gain control settings do not matter in the frequency calibration and can be adjusted for a screen pattern of convenient size. The results can be very accurate when no synchronizing is used for the oscilloscope sweep and the signal-generator frequency is correct.

The primary problem in making the frequency calibration is obtaining a

suitable source for comparison. The audio-frequency signal generator is useful when its frequency is known accurately and is convenient for checking frequencies in the horizontal deflection circuits of the receiver. For the vertical deflection circuits, the 60-cps power-line voltage is a convenient source with accurately known frequency. In addition, the 60-cps voltage from the power line can be used to check the frequency calibration of the audio signal generator. This can be done with or without the oscilloscope internal sweep. When the internal sweep is not used the 60-cps power-line voltage is connected directly to the horizontal amplifier binding posts and serves as the horizontal deflecting voltage for the oscilloscope. The signal-generator output is connected to the vertical input of the oscilloscope and its frequency is varied, noting the resultant Lissajous patterns on the oscilloscope screen.

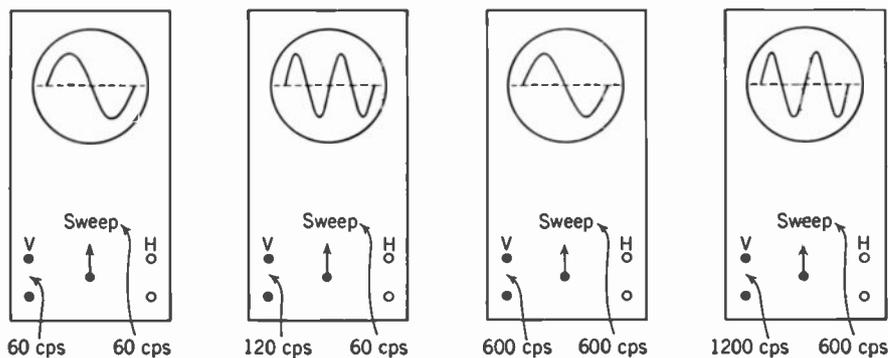


FIG. 23-13. Comparing frequencies with the oscilloscope.

When the oscilloscope internal sweep is used, the 60-cps reference voltage is connected to the vertical input to calibrate the oscilloscope sweep frequency and is then removed so that the signal-generator output can be coupled to the vertical amplifiers and checked against the oscilloscope sweep frequency. The oscilloscope sync control must be at zero during the frequency calibration. As an illustration of this procedure, the 60-cps line voltage is connected to the vertical input and the oscilloscope sweep frequency adjusted for one complete cycle on the screen. With the sweep control settings fixed, the signal-generator voltage is coupled to the oscilloscope instead of the 60-cps line voltage. The signal-generator frequency is then varied. At the point where one complete cycle is produced on the screen the generator frequency is 60 cps. Two complete cycles indicate 120 cps. Similarly, a pattern with three complete cycles indicates a frequency of 180 cps, and this can be done up to 600 cps, when 10 cycles are produced. Without synchronization it will usually be impossible to lock in the screen pattern to make it stationary, but the calibra-

tion is taken for the points where the pattern just barely drifts across the screen.

At this point, when it becomes difficult to count the cycles in the screen pattern, the oscilloscope sweep controls can be readjusted to produce one cycle on the screen. The oscilloscope sweep frequency is now 600 cps and, therefore, the generator frequency can be checked at intervals of 600 cps. A pattern of two cycles now indicates a frequency of 1,200 cps; three cycles correspond to 1,800 cps, etc. This is illustrated in Fig. 23-13. It is necessary to calibrate the sweep frequency of the oscilloscope in this way, or some similar manner, because the frequency range of the controls is only approximate and the sweep frequency usually does not vary linearly over the range.

**23-11. Trouble-shooting Techniques.** Once the trouble has been localized to a section of the receiver, the defective component can be located by tube substitutions, signal tracing, voltage measurements, and resistance checks. Tubes are the most common source of trouble and should be checked first, preferably by substitution of a tube known to be good or substituting the suspected tube in a chassis that is operating normally. When the same tube type is used in different sections of the receiver, switching the tubes will often indicate whether the suspected tube is bad. In the local oscillator stage, it may be necessary to try several tubes because one tube may not work well where another of the same type will. This may also happen in the horizontal oscillator control stage. Microphonic or noisy tubes can be checked by tapping them lightly and noting any indication in the picture and sound. The power-supply rectifiers and the output amplifier stages are common sources of trouble, since the components subject to high voltages or high currents are more likely to fail. The position of the components and their connecting leads, called *lead dress*, is usually critical in the r-f and i-f signal circuits and in the high-voltage supply. Replacement parts should match the original component, in value and position, especially in coupling and bypass circuits for the signal. The critical lead dress is usually described in the manufacturer's service notes.

*Signal Injection.* Figure 23-14 illustrates how a conventional signal generator can be used to inject test signal into the picture circuits when there is a raster but no picture. Each amplifier stage can be checked, working back from the kinescope toward the antenna input, to localize a defective stage that cannot pass signal. The 400-cps audio output of the signal generator is coupled into the video amplifier circuit to see if the signal is amplified to produce horizontal bars on the kinescope screen. In the i-f amplifier the modulated r-f test signal is injected at the i-f picture carrier frequency. If the i-f and video sections are operating, the amplitude-modulated test signal will be detected to provide 400-cps video

signal that produces bars. In the r-f section, the modulated r-f test signal at the r-f picture carrier frequency of the selected channel can be injected into the mixer grid circuit and antenna input. If the local oscillator stage, i-f section, and video amplifier are operating, the 400-cps modulation will produce horizontal bars. The test signal usually cannot be substituted for the local oscillator because of insufficient r-f output from the signal generator.

In the vertical deflection circuits, 60-cps filament voltage can be injected into the grid circuit of the vertical amplifier to check operation of the output stage when there is no vertical scanning. Vertical deflection with the injected voltage indicates the output stage is operating and, therefore, the trouble is no deflection voltage from the vertical oscillator.

*Signal Tracing with the Oscilloscope.* With picture signal input to the receiver, the path of the signal can be traced by observing the composite

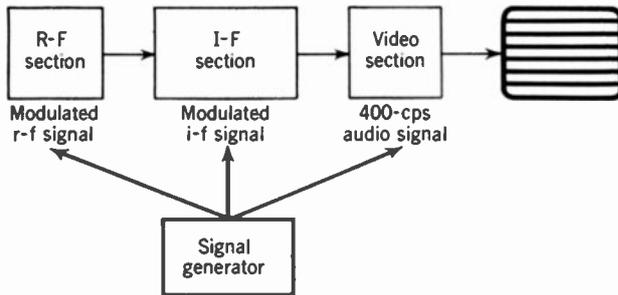


FIG. 23-14. Using generator to inject test signal in picture circuits.

video signal on the oscilloscope screen, from the video detector output to the kinescope grid input. No r-f probe is needed to see the signal after it has been rectified in the detector. The oscilloscope can also be used to trace the sync signals from the composite video signal input for the sync separator circuits to the sync output for the deflection oscillators. Observing the sync wave shapes can be helpful in cases of hum in the sync, poor interlacing, and horizontal a-f-c troubles. In the deflection circuits, the deflection voltage wave shapes can be observed to check linearity and amplitude for horizontal and vertical scanning.

To see the desired wave shape, the vertical input of the oscilloscope is connected to the circuit being checked and the oscilloscope's internal sweep is used. Setting the internal sweep frequency to 30 cps provides the waveform for two cycles of vertical scanning fields; at 7,875 cps the internal sweep frequency provides the waveform for two cycles of the horizontal scanning lines.

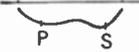
*Signal Tracing with the Voltmeter.* A stage that has grid-leak bias can be checked by measuring the negative bias with the d-c voltmeter. In an

oscillator, the grid-leak bias indicates feedback voltage is being generated, which means the oscillator is operating; no bias indicates no feedback and the oscillator is not operating. Therefore, operation of the local oscillator, vertical deflection oscillator, or horizontal deflection oscillator can be checked by measuring the negative grid-leak bias with the d-c voltmeter. In an amplifier with grid-leak bias, the negative d-c bias voltage is produced by the a-c input signal from the previous stage. The presence of the grid-leak bias, therefore, indicates a-c signal input. The amplifier stages in the receiver that usually have grid-leak bias are the horizontal output, sound i-f, and mixer stages. When the horizontal output stage does not have its normal grid-leak bias, this indicates there is no deflection voltage output from the horizontal oscillator. If the normal grid-leak bias is not present in a sound i-f amplifier, there is no FM signal input from the preceding stage. No grid-leak bias in the mixer stage means the oscillator injection voltage is not present. The d-c voltmeter can also be used to measure the rectified signal output of the video detector and FM sound detector. No d-c voltage output from the detector means no i-f signal input.

**23-12. Receiver Alignment.** When the picture is poor and adjustment of the fine tuning control has a marked effect on the quality, this indicates the possibility of receiver misalignment. Since varying the local oscillator frequency with the fine tuning control changes the i-f signal frequencies, this is equivalent to checking the i-f response curve in terms of the reproduced picture. Typical results of misalignment are (1) low sensitivity, causing a weak picture of poor quality with horizontal pulling; (2) built-in ghosts that vary in spacing when the oscillator frequency is changed; (3) smear that varies with the fine tuning control. To check whether the receiver is misaligned, the visual response curve can be obtained. The procedures can be reviewed by referring to Sec. 19-5 for the picture i-f section, Sec. 22-10 for the sound i-f section, and Sec. 20-7 for the r-f tuner. In addition, the main points to be noted when using the alignment equipment to obtain the visual response curve are summarized in Table 23-7, for the r-f and picture i-f stages. If realignment is necessary, the specific alignment instructions in the manufacturer's service notes should be followed.

Referring to Table 23-7, typical curves are shown with the picture and sound carrier frequencies marked for the i-f response, the r-f amplifier response, and the over-all r-f and i-f response from the antenna input to the video detector output. The curves can be as shown or upside down, depending on the polarity of the rectified signal input to the oscilloscope. Which side of the curve corresponds to the sound or picture should be checked with the marker oscillator. The negative bias on the amplifiers should be at its normal value for the response curve because the tuning

TABLE 23-7. VISUAL RESPONSE CURVES

Test equipment	I-F response curve 	R-F response curve 	Over-all r-f and i-f response curve 
Sweep generator connected to.....	Mixer grid	Antenna input (match sweep cable to receiver impedance)	Antenna input (match sweep cable to receiver impedance)
Marker oscillator connected to.....	I-F circuit	R-F circuit	I-F circuit
Oscilloscope connected to	Video detector load resistance	Mixer grid circuit	Video detector load resistance
Sweep generator frequency at.....	I-F	R-F, same as selected channel	R-F, same as selected channel
Marker frequency at...	I-F	R-F, same as selected channel	I-F

will vary with a change of input capacitance caused by different bias voltages. Manufacturer's service notes should be consulted for the required bias, which is 1 to 6 volts. This is supplied by a battery bias box connected to the a-g-c line to replace the a-g-c bias. Make sure that the low-impedance cable of the sweep generator does not short the bias when connected to a grid circuit on the a-g-c line. A series blocking condenser must be inserted if the cable does not have one. The local oscillator output can be prevented from interfering with the i-f response curve by setting the receiver to an unused channel at the high end of the band, or by disabling the oscillator. This can be checked by varying the fine tuning control and noting that there is no effect on the response curve. However, the local oscillator must be operating to obtain the over-all r-f and i-f response curve. For the r-f response curve alone, removing the first i-f tube minimizes spurious responses.

In setting the sweep generator controls, it is important to remember that, when connected into the r-f circuits, the sweep generator must be set to the same frequencies as the selected channel on the receiver. In the i-f circuits, the sweep generator remains at the intermediate frequencies of the receiver. Similarly, when the marker oscillator is injected into the r-f circuits, the marker frequency must be changed for each channel, but in the i-f circuits the marker remains at the intermediate frequencies. For this reason it is preferable to mark the over-all response curve with intermediate frequencies, instead of the r-f input frequencies. The r-f output control on the sweep generator is set for no more than the required amount of sweep signal, since the limiting action produced by overloading in the

amplifiers causes incorrect response curves. When the horizontal deflection voltage input for the oscilloscope is taken from the sweep generator, this is connected to the horizontal input terminals and the oscilloscope's internal sweep is turned off. The sweep generator's phasing control is used to superimpose the two traces on the oscilloscope screen. This must be done before the blanking is turned on, when the sweep generator has internal blanking for one-half cycle in order to provide a zero base line for the curve. Otherwise, two marker pips will be observed, moving in opposite directions on the response curve as the marker generator frequency is varied. The sweep width control is set for about 10 Mc. Insufficient sweep width results in only part of the response curve showing on the oscilloscope screen. The vertical input of the oscilloscope is usually connected either to the video detector load resistance for rectified i-f sweep signal or to the mixer grid for rectified r-f sweep signal. If these points do not provide a load impedance to chassis ground for the 60-cps rectified sweep signal, however, it may be necessary to connect into the video amplifier or the mixer plate circuit in order to obtain the response curve.

The curve should represent the actual response of the receiver circuits, without distortion introduced by the test equipment. Otherwise, the receiver seems to be normal with the test equipment but, when the transmitted signal is received and the equipment disconnected, the alignment is incorrect. For best results in aligning the receiver circuits, careful attention is necessary to several factors in addition to providing normal bias and eliminating spurious effects from the local oscillator. Because of the high signal frequencies, the r-f output cable from the sweep generator must be shielded, with no more than about 1 in. unshielded to the connection point. The cable shield should be grounded close to the signal lead connection. A 6-in. ground return may prove troublesome in the r-f and i-f circuits, causing dips or peaks in the response curve. The oscilloscope cable should also be shielded to prevent stray pickup. All adjustments must be made with low-loss nonmetallic alignment tools to avoid detuning the circuits being adjusted. Good grounding is necessary for the test equipment and receiver chassis; if the visual response curve changes when the equipment is touched, this indicates the need for better grounding.

The sweep generator cable is usually terminated in a resistance equal to its characteristic impedance, which is generally 50 or 72 ohms, in order to prevent radiation of the signal and provide uniform output voltage for all frequencies in the sweep width. This is especially important in a sweep generator with a wide frequency swing because an incorrect response curve will be obtained if the generator's output is not constant for all frequencies. Varying the position of the r-f output attenuator control

should vary only the size of the response curve, without changing its shape. When the sweep generator cable is connected to the antenna terminals of the receiver, for r-f or over-all alignment, the impedance of the cable should be matched to the receiver input impedance by means of a resistance pad. Otherwise, the cable connection will detune the antenna input circuit and it cannot be aligned correctly. In general, the sweep generator must be matched when the cable is connected to a circuit to be aligned. When the connection precedes the circuits being aligned, the generator can supply the required input signal without the need for impedance matching.

Do not overload the amplifiers by coupling in too much signal. Overloading produces a flat top or sharp dips in the visual response curve, independent of the tuning, which makes the final results incorrect. As a precaution, use less output from the signal generator and a higher setting of the vertical gain control on the oscilloscope to obtain the desired size for the response curve. When there is insufficient signal from the generator, the receiver noise is evident in the response curve as rapid vertical variations, generally called "grass" or "hash." The manufacturer's alignment instructions may state the amount of signal required, in terms of the peak-to-peak voltage amplitude of the response curve on the oscilloscope screen, which is about 3 volts at the detector output. To be sure there is no overloading, the signal generator output can be varied and the amplitude of the response curve should also change. Similarly, the reading on a d-c voltmeter used as the indicator for peak alignment should vary with a change of input signal.

It is important to note that incorrect results will be obtained also if the amplifier is oscillating. This can be checked by noting that there is no output when the generator signal is removed. As a final precaution, it is important that the marker signal be injected without changing the response curve. This can be done by using very loose coupling, often depending on just stray coupling for injecting the marker signal. The isolation can be checked by noting that removing the marker signal does not change the response curve. If the marker cannot be injected without distortion, note the marker points on the oscilloscope screen and then turn off the marker generator to observe the undistorted response curve for the alignment. When the marker beat on the oscilloscope screen is too broad it can be made narrower by connecting a 0.01- $\mu$ f r-f bypass condenser across the vertical input binding posts and using a decoupling resistor of 5,000 to 15,000 ohms in series with the oscilloscope lead.

**23-13. Improving Fringe-area Reception.** In addition to using a high-gain antenna and the possible need for a booster, in fringe areas where the signal is very weak the picture can be improved by increasing the receiver gain. It is assumed that a low-loss transmission line is used, matched to

the receiver impedance and the antenna impedance at the frequencies of the weak channels, in order to deliver maximum antenna signal. The receiver gain can be increased by (1) making the r-f response favor the r-f picture carrier signal for a weak channel, (2) modifying the i-f response to increase the gain for the i-f picture carrier signal on all channels, and (3) using less grid bias.

In general, the r-f alignment adjustments can be set for best results on the weak channel to be received, instead of compromising on adjustments that are suitable for all channels. The r-f amplifier's response curve often has a tilt, with more gain at either the high- or low-frequency side. By aligning the r-f amplifier to make the response tilt upward for the r-f picture carrier, more r-f gain is obtained for the picture signal at the expense of the sound signal. Usually, however, adequate reception of the FM sound signal is less of a problem than obtaining a good picture with weak signal input. It may also be possible to narrow the r-f bandwidth for one or two particular channels, resulting in a higher response curve and more gain. In addition, the adjustment of oscillator injection voltage can be made for best results on the weak channel to be received. When the r-f amplifier is connected to the a-g-c line, more r-f gain can be obtained by returning the grid circuit to chassis ground, instead of using the a-g-c bias. Then, the r-f stage can operate with the grid-leak bias produced by the weak input signal, which is less than the a-g-c bias and allows more gain. It is important to note that these modifications apply only to the case where only one or a few channels are to be received and they are all weak, since the changes may cause overload and cross modulation on a strong station. For a weak station, though, the picture can be improved perceptibly because the slight increase in r-f gain raises the signal level without adding receiver noise.

In the i-f amplifier, the receiver gain can be increased more than in the r-f amplifier. The increased i-f gain usually does not improve the signal-to-noise ratio, however, and more gain for the i-f picture carrier usually results in less detail and more smear in the picture. Still, with a very weak picture, increasing the i-f gain can improve the picture contrast. When the i-f amplifier is aligned to make the i-f picture carrier frequency have more than 50 per cent response, an appreciable increase in contrast is obtained. Moving the picture carrier up on the i-f response curve tends to produce smear in the picture, caused by excessive response for the low video frequencies, but about 80 to 90 per cent response may give improved results. In addition, the bandwidth of the i-f response curve can be made as narrow as 2 to 3 Mc, at the half-power points on the skirts, in order to increase the amplitude of the response curve, allowing much more i-f gain. The reduced i-f bandwidth results in loss of detail in the picture, but with excessive snow in the picture the fine detail is

not visible anyway. If the i-f bandwidth is reduced, it is important to note that in intercarrier-sound receivers enough response must be provided for the sound i-f carrier frequency to beat with the picture carrier to produce the 4.5-Mc second sound i-f signal in the video detector. The i-f gain can also be increased by using tubes with a higher value of transconductance, reducing the a-g-c bias, taking one or two i-f tubes off the a-g-c line, or grounding the a-g-c line completely. When the i-f bias is reduced, though, the receiver should be checked at all settings of the contrast control for the possibility of buzz in the sound and an overloaded picture.

In the video section, more gain can be obtained for a very weak signal by using tubes with a higher value of transconductance, reducing the bias, and increasing the load resistances. Higher values of load resistance in the video detector and video amplifier result in less detail and more smear, as the high-frequency response is reduced with excessive gain for the low video frequencies, but an increase of about 25 per cent can provide suitable results. If the i-f bandwidth is reduced, the frequency response of the video section might as well be limited to the same bandwidth also so that more gain for the video signal can be obtained, without wasting bandwidth for frequencies not being supplied by the i-f section. When the bias is reduced it may be necessary to disable the a-g-c in a receiver where the a-g-c rectifier is directly coupled to the video amplifier. The picture should be checked at all settings of the contrast control.

**23-14. Receiver Circuits.** Practically all television receivers have a 17- to 27-in. rectangular kinescope, using magnetic deflection. The anode voltage of about 9 to 18 kv is generally produced by a flyback high-voltage supply, sometimes with two high-voltage rectifiers in a voltage-doubler circuit. The low-voltage supply in larger receivers often has two rectifiers, which may be in parallel for a greater value of load current, or operated separately to supply B+ voltage for different sections. Often one low-voltage rectifier is for the signal circuits while the other supplies higher B+ voltage for the deflection circuits. In this case, it is important to note that trouble in one low-voltage supply can affect the raster without disturbing the sound, or vice versa. The power used by a television receiver from the 60-cps a-c power line is 150 to 300 watts.

In the deflection circuits, reaction scanning for horizontal deflection is the usual practice, with the damper providing boosted B+ voltage. Internal blanking of the vertical retrace lines is usually provided by coupling the vertical output pulses to the kinescope grid-cathode circuit. Note that troubles in the vertical output circuit can affect kinescope brightness through the vertical blanking circuit. Either the blocking oscillator or multivibrator circuit is generally used for the horizontal and

vertical deflection oscillator, with automatic frequency control always applied to the horizontal oscillator.

In the signal circuits, practically all receivers use intercarrier sound, with a 4.5-Mc second sound i-f section for the associated sound signal. A picture a-g-c circuit is generally included, consisting of either a diode a-g-c rectifier to rectify the picture i-f signal in a simple a-g-c circuit, or a keyed a-g-c amplifier circuit in the video section. Since the a-g-c bias is applied to i-f and r-f amplifier stages common to the picture and sound signals, troubles in the a-g-c circuit affect both the picture and sound. Note that, with a keyed a-g-c stage directly coupled to the video amplifier, defects in the video circuits can cause troubles in the a-g-c bias. Most receivers have three i-f amplifier stages for the picture signal, although some chassis are available with four i-f stages. In the r-f section, the front end usually has a cascode r-f amplifier stage for minimum receiver noise. The local oscillator usually operates above the r-f signal frequencies, in order to reduce the oscillator tuning range required, producing in the mixer output the i-f picture and sound carrier frequencies of 45.75 Mc and 41.25 Mc or 25.75 Mc and 21.25 Mc, respectively. Reception of the u-h-f television channels is provided by u-h-f coil strips in turret tuners, or continuous tuning through the u-h-f television band.

*Typical Circuit.* General practice in receiver circuits is indicated by the schematic diagram given in Fig. 23-15 to illustrate how the circuits for all the individual receiver sections fit together. The front end is a turret tuner using the 6BZ7 twin-triode  $V_{101}$  for the cascode r-f amplifier and the 6J6 twin-triode  $V_{102}$  as oscillator-mixer. This is essentially the same r-f tuner described in detail in Sec. 20-3. The i-f output at plate pin 2 of the mixer stage  $V_{102A}$  is coupled to the common i-f amplifier of the intercarrier-sound receiver, with i-f picture and sound carrier frequencies of 25.75 Mc and 21.25 Mc, respectively, while the 4.5-Mc second sound i-f signal in the detector output is coupled to the 4.5-Mc sound i-f section.

In addition to the double-tuned i-f coupling circuit in the mixer output, the common i-f amplifier uses the three single-tuned stages  $V_{301}$ ,  $V_{302}$ , and  $V_{303}$ , which have single-peaked response at the staggered resonant frequencies indicated in the diagram for  $T_{301}$ ,  $T_{302}$ , and  $T_{303}$ . In the grid circuit of  $V_{302}$ , the wave-trap adjustment  $L_{310}$  attenuates the associated sound i-f carrier at 21.25 Mc for the required 2 to 5 per cent response.  $T_{301}$  and  $T_{302}$  are bifilar i-f coils but the last i-f stage is impedance-coupled to the detector by  $T_{303}$ . In the plate circuit of the mixer, the i-f coupling circuit to the grid of the first i-f stage consists of the double-tuned circuit formed by  $L_{103}$  and  $L_{309}$ , which have a common impedance to ground at their junction. This mutual impedance is provided by  $L_{110}$  in combination with two wave traps,  $C_{124}$  being just a d-c blocking condenser. The wave-trap adjustment  $L_{108}$  rejects the upper adjacent

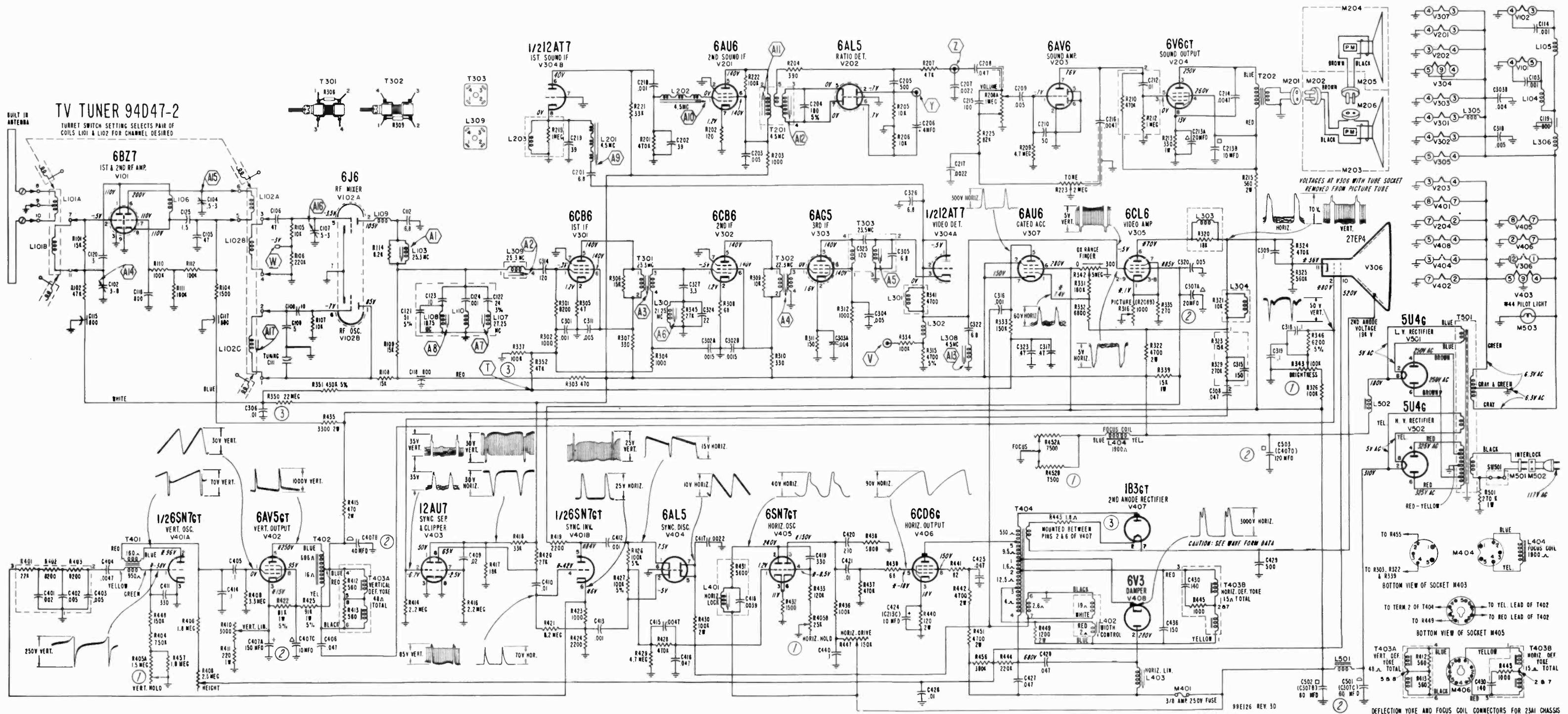


FIG. 23-15. Complete television receiver schematic diagram. Alignment adjustments indicated by letter A. Letters V, W, Y, Z and T indicate alignment connections and test points. Numerical symbols in parts of circuit indicate production changes. Voltages marked with asterisk vary widely with setting of controls. Resistance in ohms, capacitance values less than 1 in  $\mu$ f and more than 1  $\mu$ fd, unless otherwise noted. (Admiral Series 23 chassis.)

channel's i-f picture carrier at 19.75 Mc;  $L_{107}$  is tuned to reject the lower adjacent channel's i-f sound carrier at 27.25 Mc. The entire mutual coupling network is parallel resonant at 23.5 Mc. Since  $L_{103}$  and  $L_{309}$  are tuned to 25.3 Mc, while the mutual coupling impedance is maximum at 23.5 Mc, the interstage coupling circuit has a double-peaked response 5 Mc wide with one peak at 25.3 Mc and the other at 23.5 Mc. Combining this with the response of the three stagger-tuned stages results in the over-all i-f response curve shown in Fig. 23-16.

For the detector, one triode section  $V_{304A}$  of the twin-triode tube 12AT7 is used as a diode with the plate tied to the grid. The i-f signal is applied to the cathode to provide video signal output of positive picture phase and negative sync polarity. The plate load resistor is  $R_{315}$ , while  $L_{301}$  and  $L_{302}$  are peaking coils. The video signal at their junction is directly

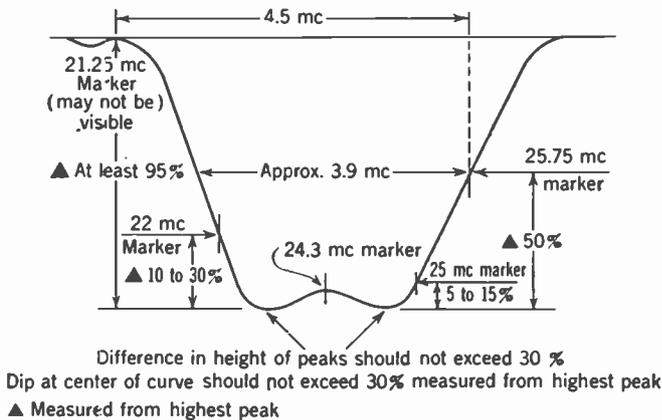


FIG. 23-16. I-F response curve for receiver in Fig. 23-15. (Admiral Corporation.)

coupled to the control grid of the 6CL6 pentode video amplifier. In the grid circuit,  $L_{308}$  with  $C_{322}$  form a wave trap to minimize 4.5-Mc inter-carrier beat in the picture. The video amplifier's output is d-c coupled to the kinescope cathode to provide video signal of the correct picture phase with the required d-c component. The 4,700-ohm resistor  $R_{322}$  is the plate load resistor for the video amplifier. At the junction of  $R_{322}$  and  $L_{304}$ , composite video signal voltage is coupled to the sync circuits and to the control grid of the 6AU6 gated a-g-c stage  $V_{307}$ . The negative a-g-c bias voltage at plate pin 5 of  $V_{307}$  controls the gain of the first and second i-f stages and the r-f amplifier. This keyed a-g-c circuit is described in detail in Sec. 15-4.

Referring back to the second detector  $V_{304A}$  in Fig. 23-15, the output includes the 4.5-Mc frequency-modulated second sound i-f signal produced as the beat between the associated FM sound signal and the pic-

ture carrier in the detector. The 4.5-Mc sound signal is coupled from the plate circuit of the detector through the  $6.8\text{-}\mu\text{f}$  condenser  $C_{201}$  to the cathode of the first 4.5-Mc sound i-f stage  $V_{304B}$ . The small capacitance of  $C_{201}$  allows the 4.5-Mc sound signal to be coupled while providing higher reactance to attenuate the lower frequencies of the video signal. Two 4.5-Mc sound i-f amplifier stages are used, followed by the 6AL5 ratio detector for the FM sound signal.  $C_{206}$  is the stabilizing condenser in the output circuit, while  $C_{205}$  is the load for the diodes to produce audio output voltage. The audio output is coupled by  $C_{208}$  and  $C_{209}$  to the two-stage audio amplifier, which drives two 10-in. loudspeakers.  $C_{208}$  prevents direct current from flowing through the volume control, which would make it noisy.

For the sync circuits, composite video signal is taken from the video amplifier plate circuit through the isolating resistors  $R_{323}$  and  $R_{329}$  and the coupling condenser  $C_{308}$  to the grid pin 2 of the 12AU7 sync separator and clipper  $V_{403}$ . The separated sync output at plate pin 6 of  $V_{403}$  is coupled to the grid of the sync inverter stage  $V_{401B}$ , which produces push-pull sync output voltage in the plate and cathode circuits for the 6AL5 sync discriminator. These sync separator circuits are described in detail in Sec. 16-8.

The cathode pin 6 of the sync inverter stage  $V_{401B}$  is connected to the three-section integrator in the grid circuit of the vertical oscillator  $V_{401A}$  for vertical sync. The blocking oscillator circuit is used, with  $C_{414}$  the saw-tooth condenser. Saw-tooth deflection voltage drives the grid of the pentode vertical output stage  $V_{402}$ , with the plate circuit coupled by the autotransformer  $T_{402}$  to the vertical scanning coils. Note that  $C_{406}$  in the output circuit couples vertical deflection pulses of negative polarity to the kinescope grid for internal blanking of vertical retrace lines. For horizontal scanning, the multivibrator  $V_{405}$  is used as the horizontal oscillator, with automatic frequency control provided by the 6AL5 sync discriminator. This circuit is described in Sec. 17-10. The saw-tooth condenser for the horizontal oscillator is  $C_{420}$ , with the peaking resistor  $R_{438}$  producing a spike, which is not indicated in the diagram, for the grid-voltage wave shape driving the 6CD6 horizontal output stage. The horizontal output transformer  $T_{404}$  is an autotransformer, supplying scanning current for the horizontal deflection coils and high-voltage input for the 1B3-GT in the flyback supply. Terminal 7 of the transformer also supplies horizontal deflection pulses of positive polarity for the plate of the keyed a-g-c stage and for the 6AL5 sync discriminator. The inverted diode damper produces boosted B+ voltage at the junction of  $C_{427}$  and  $C_{428}$ .

Two 5U4-G full-wave rectifiers are used for the low-voltage power supply,  $V_{502}$  supplying higher B+ voltage than  $V_{501}$ , primarily for the

deflection circuits. Distribution of the B+ voltage in this receiver is shown in the simplified diagram given in Fig. 10-3.

*Service Hints.* A few points are summarized here briefly from the manufacturer's service data for the receiver in Fig. 23-15, in order to illustrate some specific procedures that can be important in servicing.

1. The sockets for the electromagnetic focus coil and deflection yoke must be connected for normal B+ voltage during alignment.

2. The waveform at the plate of the horizontal output tube can be taken by clipping the lead from the oscilloscope over the insulation on the lead to the 6CD6-G top cap. The shape of the observed waveform corresponds to the plate voltage but is much lower, depending on the degree of a-c coupling through the insulation. The oscilloscope cannot be connected directly to the plate of the horizontal output tube because of the high-voltage pulses.

3. If the focus control focuses only at the extreme end of its rotation, this may be caused by a weak 5U4-G rectifier tube V<sub>501</sub>.

4. Insufficient picture width may be caused by a weak low-voltage rectifier tube V<sub>501</sub>, horizontal output tube, or damper tube. Insufficient width can also be caused by incorrect setting of the horizontal drive control R<sub>447</sub>, which determines the plate-supply voltage for the horizontal oscillator. In addition, moving the focus coil too close to the deflection yoke can cause a reduction of width; the spacing should be at least  $\frac{3}{8}$  in.

5. Incorrect setting of the a-g-c control R<sub>432</sub> can cause the video amplifier V<sub>305</sub> to block on a strong signal, resulting in no picture, while the sound is normal.

6. To reduce snow in the picture, the a-g-c bias voltage for the r-f amplifier V<sub>101</sub> can be reduced to 1.5 to 2.0 volts by means of a voltage divider.

*Tube Layout.* Figure 21-17 shows the location of tubes for the receiver in Fig. 21-15. Although the tube layout will vary in different receivers, the low-voltage power supply is usually at the back of the chassis for better ventilation; the high-voltage supply is generally in an enclosed cage on one side of the chassis, with the horizontal oscillator, output, and damper tubes either in the cage or next to it; on the opposite side of the chassis are the signal stages, usually lined up in successive order to allow short connecting leads from one amplifier to the next.

*Over-all Receiver Gain.* The picture amplifier circuits in a television receiver produce a total gain of one and one-half million times in voltage, approximately, from the antenna input terminals to the kinescope grid-cathode circuit. This is the product of the gain values for the individual stages, as each amplifier multiplies the signal amplitude by an amount equal to the gain of the stage. Individual gain values may be approxi-

mately 10 for the r-f section, 15,000 for signal frequencies having 100 per cent response in the i-f amplifier section,  $\frac{1}{2}$  for the diode video detector, and 20 for the video amplifier, producing the over-all gain of

$$10 \times 15,000 \times \frac{1}{2} \times 20 = 1,500,000$$

Since the kinescope requires approximately 75 volts of video signal, a minimum r-f signal level of about  $50 \mu\text{v}$  is needed at the antenna terminals. With an internal receiver noise voltage of approximately  $10 \mu\text{v}$  in the r-f amplifier input circuit, the signal-to-noise ratio for the  $50\text{-}\mu\text{v}$

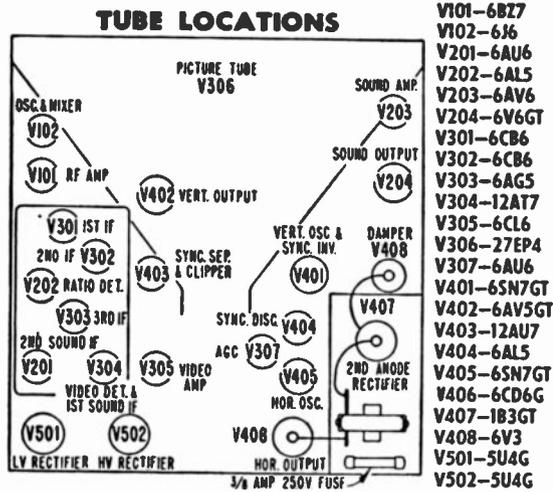


Fig. 23-17. Top view of chassis with tube locations for receiver in Fig. 23-15. (Admiral Corporation.)

signal is 5:1. More signal is necessary to provide a higher signal-to-noise ratio for a picture without appreciable snow.

#### REVIEW QUESTIONS

1. Describe briefly how to adjust the ion-trap magnet, focus coil magnet, and deflection yoke on a picture tube.
2. Describe briefly how to adjust the height and width of the raster for full size, with the correct aspect ratio and good linearity.
3. Why is the size of the picture slightly smaller than the raster without picture information?
4. Give two effects of varying the horizontal drive control, in addition to its effect on width of the raster.
5. Give two possible causes of ghosts and a remedy for each.
6. Describe briefly how local oscillator radiation produces interference in the reproduced picture. Give three possible effects of the interference in the picture.
7. When the receiver in Fig. 23-15 is tuned to channel 2, its local oscillator radiation can cause r-f interference in what channel?
8. Describe two effects of diathermy interference in the reproduced picture.

9. What is the distinguishing feature of FM interference in the picture?
10. What is an image frequency of the picture carrier in channel 2 for a receiver with a picture i-f carrier of 26 Mc? With a picture i-f carrier of 45.75 Mc?
11. How does a television receiver cause whistles in nearby radio receivers for the standard broadcast band?
12. How can horizontal streaks in the picture be distinguished between ignition interference and motor noise?
13. How can you recognize sound bars in the picture?
14. How can you recognize 4.5-Mc intercarrier beat in the picture?
15. For the receiver circuit in Fig. 23-15, there is one pair of hum bars in the raster and picture with hum bend, but no hum in the sound. Localize the source of hum voltage.
16. Give two methods of localizing buzz in the sound between the signal circuits and the vertical deflection circuits.
17. For the receiver circuit in Fig. 23-15 localize each of the following troubles:
  - a. Set completely dead. No raster, picture, or sound.
  - b. No picture and no sound with normal raster.
  - c. No raster and no picture with normal sound.
  - d. No sound, with normal raster and picture.
  - e. Insufficient width in raster.
  - f. Only a thin bright horizontal line in center of screen.
  - g. Bright bar at bottom of raster.
  - h. 4.5-Mc beat pattern in picture.
  - i. Picture does not hold horizontally. Vertical hold is normal.
  - j. Hum just in the sound, only with signal input.
18. Referring to the receiver circuit in Fig. 23-15, where would a d-c voltmeter be connected to measure:
  - a. I-F signal output from the common i-f amplifier.
  - b. I-F signal output from the 4.5-Mc sound i-f amplifier.
  - c. Grid-leak bias on the horizontal oscillator stage.
19. When the normal deflection voltage wave shape is obtained with the oscilloscope at the grid of the vertical output tube, for what circuits can normal operation be assumed?
20. What is the required internal sweep frequency of the oscilloscope for two cycles of the vertical deflection voltage wave shape? For two cycles of the horizontal deflection voltage wave shape?
21. If 6.3 volts r-m-s produces 1 in. vertical deflection on the oscilloscope screen, what is the peak-to-peak amplitude of saw-tooth voltage producing 2 in. of deflection?
22. Describe briefly how to obtain the over-all r-f and i-f response curve of the receiver in Fig. 23-15. Note two precautions to be observed in setting the controls on the equipment and two precautions in obtaining the correct response curve.
23. Give two methods of increasing the picture contrast for fringe-area reception.
24. Referring to the receiver circuit in Fig. 23-15 state briefly the function of each of the following components:
  - $C_{115}$ ,  $C_{111}$ ,  $C_{116}$ , and  $R_{111}$  in the front end.
  - $R_{305}$ ,  $R_{307}$ ,  $C_{311}$ , and  $R_{304}$  in the common i-f amplifier.
  - $R_{321}$  and  $R_{316}$  in the video amplifier.
  - $R_{342}$  in the a-g-c circuit.
  - $L_{202}$ ,  $T_{201}$ ,  $R_{209}$ , and  $C_{214}$  in the sound i-f and audio notions.
  - $R_{420}$ ,  $C_{410}$ ,  $R_{419}$ ,  $R_{423}$ , and  $R_{424}$  in the sync separation circuits.
  - $C_{416}$ ,  $L_{401}$ ,  $C_{419}$ ,  $R_{436}$ ,  $R_{447}$ , and  $C_{420}$  in the horizontal a-f-c and oscillator circuits.

$R_{440}$ ,  $C_{424}$ ,  $R_{442}$ ,  $C_{425}$ , and fuse  $M_{401}$  in the horizontal output circuit.

$C_{403}$ ,  $T_{401}$ ,  $C_{404}$ ,  $C_{414}$ ,  $R_{408}$ ,  $R_{410}$ , and  $R_{422}$  in the vertical deflection circuits.

$L_{404}$ ,  $L_{502}$ ,  $R_{452}$ ,  $C_{503}$  in the low-voltage power supply.

25. For the receiver circuit in Fig. 23-15, give the effect on the raster, picture, and sound for each of the following component troubles:

a. Heater-to-cathode leakage in the second sound i-f tube  $V_{201}$ .

b. Weak 5U4 rectifier tube  $V_{502}$ .

c. Shorted condenser  $C_{425}$  in screen-grid circuit of  $V_{406}$ .

d. Weak r-f amplifier tube 6BZ7.

e. Open heater in video amplifier tube  $V_{306}$ .

f. Open heater in gated a-g-c tube  $V_{307}$ .

g. Weak vertical output tube  $V_{402}$ .

h. Weak damper tube  $V_{407}$ .

i. Open peaking coil  $L_{303}$  in video amplifier.

j.  $C_{201}$  shorted in cathode circuit of first sound i-f stage  $V_{304B}$ .

k. Weak 6V6 audio output tube  $V_{204}$ .

l. Fuse  $M_{401}$  in horizontal output circuit open.

m. Leaky condenser  $C_{406}$  in grid of vertical output tube.

26. Approximately how much signal voltage is necessary at the antenna input terminals of a television receiver for a picture with good contrast and without snow?

## CHAPTER 24

### COLOR TELEVISION

Color increases the contrast and improves the quality of the reproduced image, adding an apparent perception of depth to the reproduction in natural colors to provide a picture much more pleasing than a black-and-white (monochrome) reproduction, which is a variation in shades of white only. The three primary colors used in television are red, green, and blue. Practically all other colors and white, gray and black can be obtained by proper mixtures of these three colors. When the image is scanned at the broadcast station, video signals corresponding to the desired picture information are obtained for the red, green, and blue colors in the scene by means of optical color filters. Picture information for the red content of the image is in the red video signal, while the green picture information is in the green video signal and the blue parts of the picture produce the blue video signal. At the color television receiver, the red, green, and blue video signals are used for reproduction of the picture in its natural colors as mixtures of red, green, and blue by means of a tricolor kinescope arrangement. A typical color television picture is shown in Plate I.

In color television broadcasting, however, the red, green, and blue video signals are combined to produce a color signal and a monochrome signal for transmission to the receiver. The monochrome signal is produced by adding the color video signals in the proportions necessary to indicate only the brightness variations in the picture information. For this reason, it is the *luminance signal*. Conversion to the luminance signal is necessary to enable conventional monochrome receivers to reproduce in black and white pictures that are televised in color. The color signal conveying the red, green, and blue information is the *chrominance signal*. Color television receivers utilize both the luminance and chrominance signals in order to recover the red, green, and blue video voltages needed for reproducing the picture with color. In summary, then, color television consists of transmitting a luminance signal essentially the same as the monochrome signal in black-and-white television broadcasting, with the addition of the chrominance signal for color.

**24-1. Color Addition.** In color television, reproduction of the many different colors in the scene is based upon the addition of primary colors.

The process is *additive* because individual color images are produced by the kinescope and combined in an arrangement that allows the eye to integrate the individual colors. The additive effect can be obtained by superimposing the color images, as illustrated in Fig. 24-1. In this arrangement the screen of each color kinescope has a phosphor that produces a red, green, or blue image. By means of an optical projection system, the three primary color images are projected onto one common viewing screen. Since the phosphor of each kinescope is a light source producing a color image, and the color images are viewed together, the observer sees the picture on the viewing screen in all its natural colors as additive mixtures of the three primary colors.

*Additive Color Mixtures.* Practically all colors, as well as white, gray, and black, can be obtained by combining individual colors. This is illustrated by several additive color mixtures of red, green, and blue shown in Plate V. There are three individual circles in red, green, and

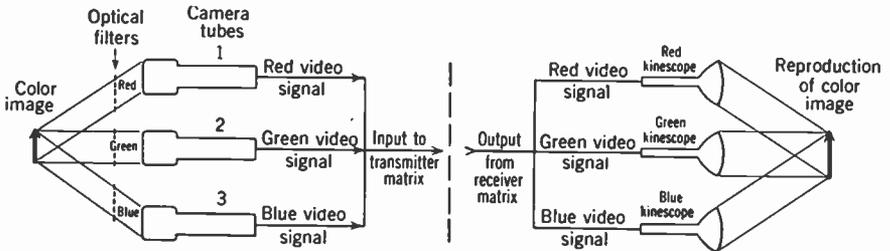


FIG. 24-1. Televising a scene in color.

blue here, which partially overlap. Where the circles are superimposed, the color shown is the mixture obtained by adding the individual colors. At the center, all three color circles overlap, resulting in white. Therefore, the white area at the center is a mixture of red, green, and blue in the proper proportions. Notice that where only green and blue add, the resultant color is a greenish-blue mixture generally called *cyan*. The red-purple color shown by the addition of red and blue is called *magenta*; more blue with less red produces purple. Yellow is an additive color mixture of approximately the same amounts of red and green; more red with less green produces orange.

*Primary Colors and Complementary Colors.* Colors that can be combined to form different color mixtures are primary colors, the only requirement being that none can be matched by any mixture of the other primaries. Red, green, and blue are the primary colors used in television because they produce a very wide range of color mixtures when added. Therefore red, green, and blue are *additive primaries*.

The color that produces white light when added to a primary is called its *complementary color*. For instance, cyan added to red produces white

light. Therefore cyan is the complement of the red primary. The fact that cyan plus red equals white follows from the fact that cyan is a mixture of blue and green, so that the combination of cyan and red actually includes all three additive primaries. Similarly, magenta is the complement of green, while yellow is the complement of blue. Sometimes the complementary colors, cyan, magenta and yellow are indicated as minus-red, minus-green and minus-blue, respectively, as each is equal to white light minus the corresponding primary. The complementary colors are also known as the *subtractive primaries*. In a reproduction process such as color photography, where color mixtures are obtained by subtracting individual colors from white light by the use of color filters, cyan, magenta, and yellow are the subtractive primary colors used to filter out red, green, and blue.

**24-2. Color Resolution.** Experimental evidence indicates that the color fidelity required for a suitable picture reproduction depends upon the relative size of the color area. For color areas corresponding to the main parts of the picture information—which are large enough to be resolved easily by the eye—mixtures of the three primaries, red, green, and blue, are necessary for a full-fidelity color reproduction. Color areas of small size, however, can be matched with only two colors, preferably orange and cyan. For very small areas requiring maximum visual acuity, the eye perceives brightness or luminance variations only, without hue. As a result, it is unnecessary to show the finest detail of the picture reproduction in color.

In order to conserve bandwidth in the restricted 6-Mc transmission channels, color television broadcasting provides a picture in color only to the extent necessary to satisfy the color resolution requirements of the average observer. Utilizing the fact that the eye requires less color for smaller areas of picture information, the color television picture can be considered in the following three parts:

1. Full three-color reproduction as additive mixtures of the red, green, and blue primaries used for the large areas of picture information.
2. Two-color reproduction as mixtures of cyan and orange used for smaller areas of picture information.
3. Monochrome reproduction in black and white used for the smallest details of picture information, which cannot be perceived in color by the eye.

The color characteristics of the color television picture are illustrated by the NTSC<sup>1</sup> flag shown in Plate VII. It should be noted that with a

<sup>1</sup>The National Television Systems Committee (NTSC) of the Radio, Electronics, and Television Manufacturers Association (RETMA) prepared the standard specifications approved by the FCC, December, 1953, for commercial color television broadcasting.

televised image the entire picture appears in color because full three-color reproduction is provided for the largest areas in the scene. In a televised picture of an automobile, for instance, the entire body of the car would be in full color; narrow vertical strips on the frame might be reproduced less exactly in two colors, while the detail of the outline of the car where it joins the background would be in black-and-white monochrome. The background would be in full natural colors.

**24-3. Definition of Color Television Terms.** The most important terms used in color television are defined as follows:

*White.* For practical purposes, white light can be considered a mixture of the red, green, and blue primary colors in the proper proportions. The reference white for color television is a mixture containing 30 per cent red, 59 per cent green and 11 per cent blue, which combine to produce a bluish white like daylight.

*Hue.* The color itself is its hue. Green leaves have a green hue; a red apple has a red hue; the color of any object is distinguished primarily by its hue. Different hues result from different wavelengths of the light producing the visual sensation in the eye. The red hues correspond to the longest wavelengths, while violet and blue light have the shortest wavelengths, with green and yellow between the two extremes. Plate VI shows the range of hues visible to the eye and their corresponding wavelengths.

*Saturation.* This indicates how little the color is diluted by white light, distinguishing between vivid and weak shades of the same hue. A weak blue color, for instance, has little saturation, while vivid blue is highly saturated. The more a color differs from white the greater is its saturation. Saturation is also indicated by the terms *purity* and *chroma*. High purity and chroma correspond to high saturation and vivid color.

*Luminance.* This indicates the amount of light intensity, which is perceived by the eye as brightness. In addition to the more familiar luminance variations in black-and-white monochrome, different colors are perceived with different brightness values by the eye. As illustrated in the relative luminosity curve in Plate VI, the blue-green and orange-red hues have maximum brightness.

*Chrominance.* This term is used to indicate both hue and saturation of a color. *Chromaticity* is also used for chrominance.

*Registration.* This refers to the positioning of individual color images to make the resultant picture of color mixtures have the correct color. With superimposed images, as an example, if each does not exactly overlay the others incorrect color mixtures will be produced because the primary colors will be in the wrong position with respect to the original colors in the picture.

*Compatibility.* Color television is compatible with black-and-white

television because essentially the same scanning standards are used and the luminance signal enables a monochrome receiver to reproduce in black and white a picture televised in color. In addition, color television receivers can use a monochrome signal to reproduce the picture in black and white. Color television broadcasting uses the same 6-Mc broadcast channels as monochrome transmission.

**24-4. Color Television Broadcasting.** The camera receives red, green, and blue light corresponding to the color information in the scene, to produce the primary red, green, and blue color video signals labeled here as *R*, *G*, and *B*. Figure 24-2 illustrates the color video signal voltages that would be obtained in scanning one horizontal line across the color bars indicated. The red, green, and blue video voltages are then combined in order to encode the primary color voltage signals as brightness and chrominance signals for transmission to the receiver. This is illustrated in Fig. 24-3.

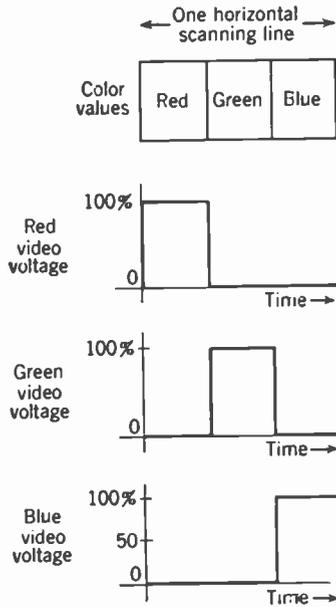


FIG. 24-2. Red, green, and blue color video signal voltages corresponding to color-bar pattern.

**Matrix Section.** The matrix is essentially a resistive voltage-divider circuit that proportions the primary color signals as required to produce the brightness and chrominance signals. With the red, green, and blue color video voltages as input to the matrix, the three video signal output combinations formed are the following:

1. Luminance signal, designated the *Y* signal, which contains the brightness variations of the picture information.
2. A color video signal, designated the *Q* signal, which corresponds to either green or purple picture information.
3. A color video signal, designated the *I* signal, which corresponds to either orange or cyan picture information.

The *I* and *Q* signals together contain the color information for the chrominance signal.

**Color Subcarrier.** The *I* and *Q* signals are transmitted to the receiver as the side bands of a 3.58-Mc subcarrier wave. A subcarrier is a relatively low-frequency carrier wave, within the range of modulation frequencies, which, in turn, modulates the main carrier wave. As an exam-

ple, the picture carrier wave at 67.25 Mc for channel 4 can be modulated by the 3.58-Mc video-frequency subcarrier. Note that the color subcarrier has the same video frequency of 3.58 Mc for all stations, although the assigned picture carrier wave is different for each channel. Only the side bands of the color subcarrier are transmitted, as the 3.58-Mc subcarrier frequency is suppressed.

**Chrominance Modulation.** Referring to Fig. 24-3, the output from the 3.58-Mc color-subcarrier oscillator is coupled to the *I* and *Q* modulators, which also have *I* and *Q* video signal input from the matrix. The signals are so applied that the 3.58-Mc subcarrier can be modulated by both the *I* and *Q* signals, forming a *modulation product* consisting of the upper and lower side bands produced by the modulation, in addition to the subcarrier. However, the *I* and *Q* modulator circuits are balanced to cancel

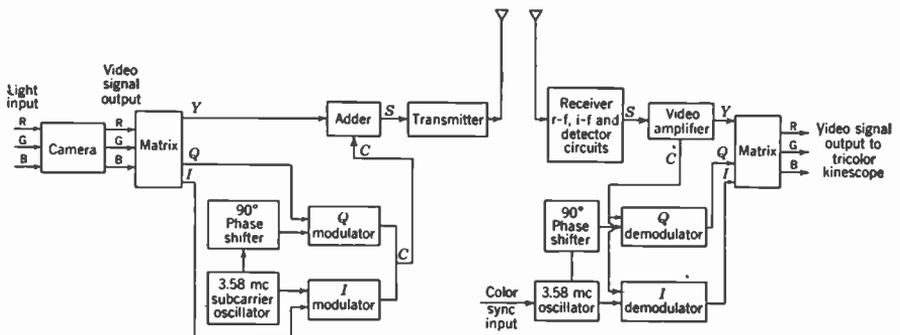


FIG. 24-3. Functions of the transmitter and receiver in the color television system.

out the subcarrier itself, so that in the output of the modulator only the modulation product is present as the side carrier frequencies of the 3.58-Mc subcarrier.

The 3.58-Mc input to the *Q* modulator in Fig. 24-3 is shifted by  $90^\circ$ . This is the reason for the designation *Q* signal, since the *Q* modulation product is *in quadrature*, or  $90^\circ$  out of phase, with the *I* signal. Modulating the subcarrier in these two different ways makes it possible to broadcast the color information of the *I* and *Q* signals simultaneously, without loss of identity. Each can be recovered as a separate signal, provided that correct timing is provided between the modulators at the transmitter and the demodulators for the *I* and *Q* signals at the receiver. The combined *I* and *Q* modulation product, generally designated the chrominance signal, is the color video signal transmitted to the receiver as the modulation product of the 3.58-Mc suppressed subcarrier.

**Total Video Signal.** As shown in Fig. 24-3, the chrominance signal *C* containing the color information and the *Y* signal with the luminance information are both coupled to the adder section, or colorplexer, which

combines or sums up the  $Y$  and  $C$  signals to form the total video signal transmitted to the receiver, here indicated as  $S$ . This total video signal is transmitted by amplitude modulation of the broadcast station's assigned picture carrier, with vestigial-side-band transmission, negative polarity of transmission, and 4.5-Mc separation of the picture carrier frequency from the sound carrier frequency, in accordance with the standard characteristics of the 6-Mc television broadcast channel. The  $S$  signal is a composite video signal including deflection sync and blanking pulses, in addition to color synchronizing signals to time the color information correctly.

*Color Sync.* Because the chrominance signal includes the  $I$  and  $Q$  color information as two-phase modulation, without the subcarrier itself, a sample of the 3.58-Mc subcarrier must be transmitted to indicate the correct frequency and phase for the 3.58-Mc signal reinserted in the receiver for the  $I$  and  $Q$  demodulators. This color synchronization is accomplished by transmitting eight to ten cycles of the unmodulated 3.58-Mc r-f subcarrier on the back porch of each horizontal blanking pulse.

*Decoding the Primary Color Signals.* To recover the red, green, and blue video signals for the color kinescope in the receiver, essentially the same functions must be performed as at the transmitter, but the decoding at the receiver is in reverse order. Starting with the receiving antenna, the modulated picture carrier signal of the selected channel is amplified in the r-f and i-f stages and rectified in the video detector to provide the total video signal  $S$ . Following the detector, the video circuits divide into two separate paths, one for monochrome and the other for chrominance. Referring to Fig. 24-3, the  $Y$  signal in the output of the video amplifier corresponds to only the luminance component of the  $S$  signal input, since there is no detection of the subcarrier chrominance signal in the video amplifier. To recover the  $I$  and  $Q$  signals in the total video signal  $S$ , this is also coupled to the  $I$  and  $Q$  demodulators in the chrominance section of the receiver. The 3.58-Mc local oscillator reinserts the subcarrier frequency needed in the demodulators to detect the  $I$  and  $Q$  signals. Notice that the subcarrier signal for the  $Q$  demodulator is  $90^\circ$  out of phase with the oscillator signal for the  $I$  demodulator, just as in the modulation process at the transmitter. Therefore, the  $Q$  demodulator produces output corresponding to the original  $Q$  signal, while the  $I$  demodulator also produces the original  $I$  signal. The  $I$  and  $Q$  color signals from the demodulators, with the  $Y$  signal from the video amplifier, are then combined in the receiver's matrix unit. The receiver matrix forms the original red, green, and blue primary color video signals for the kinescope reproduction of the picture in color.

The color picture can be reproduced by a tricolor kinescope, which has a screen that produces red, green, and blue light; or three individual color

kinescopes can be used with an optical arrangement to combine the separate color images as one resultant color picture. With separate color kinescopes, the red video signal is coupled to the control grid of the kinescope having the red-phosphor screen, the green video is for the green kinescope, and the blue video for the blue kinescope. In a kinescope having a tricolor screen and three electron guns to produce an electron beam for each of the color phosphors, the required color video signal voltage is applied to each gun. For kinescopes having a tricolor screen but only one electron gun, the color video voltages are gated so that the signal applied to the control grid corresponds to the phosphor color being excited by the electron beam.

**24-5. Y Signal.** The luminance, or  $Y$ , video signal, which contains the brightness variations of the picture information, is formed in the matrix at the transmitter by adding the red, green, and blue primary color video signals in the proportions

$$Y = 0.30R + 0.59G + 0.11B$$

These proportions correspond to the respective contributions to luminance of these three primary colors, as can be seen by referring back to the relative brightness curve in Plate VI for different hues. Figure 24-4 illustrates how the  $Y$  signal voltage in  $d$  is formed in the matrix from the specified proportions of the primary color voltages shown in  $a$ ,  $b$ , and  $c$ , corresponding to a standard color-bar pattern. Notice that the bars include the primary colors, their complementary mixtures of two primaries, and white for all three primaries. White includes all three primary colors and, therefore, has the maximum relative amplitude of unity, equal to 30 per cent red, 59 per cent green, and 11 per cent blue. The yellow color bar, as another example, produces red and green but no blue video voltage. Adding 30 per cent of the red voltage and 59 per cent of the green voltage results in luminance video signal voltage having the relative amplitude of 0.89 for yellow.

For white picture information in the scene the specified proportions for the  $Y$  signal produce a reference white designated *Illuminant C*, which is a bluish white like daylight. The choice of this reference white improves the black-and-white picture reproduction of a scene televised in color, since the luminance signal then corresponds to the relative brightness values of the colors. The  $Y$  signal has the full video-frequency bandwidth of 4.2 Mc, as illustrated in Fig. 24-5a, for reproducing the luminance information of the smallest picture details in black and white.

**24-6. Q Signal.** The color video signal designated as the  $Q$  signal is formed from the primary color signals in the matrix at the transmitter, as follows:

$$Q = 0.48(R - Y) + 0.41(B - Y)$$

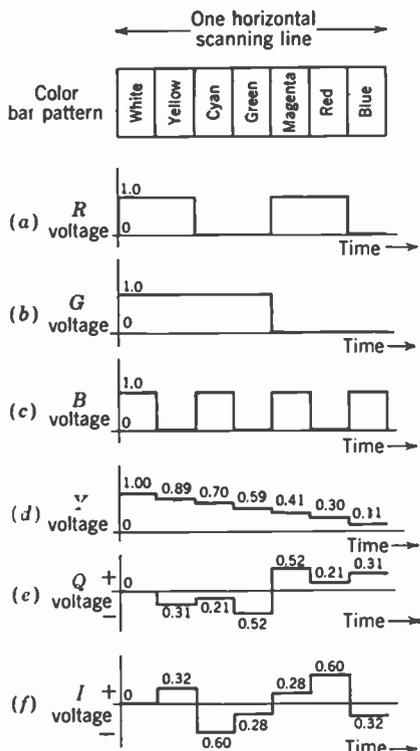


FIG. 24-4. Video voltage waveforms of  $Y$ ,  $Q$ , and  $I$  signals formed from primary color video voltages, corresponding to color-bar pattern. (a) Red video signal; (b) green video signal; (c) blue video signal; (d) luminance signal  $Y$ ; (e)  $Q$  color video signal; (f)  $I$  color video signal.

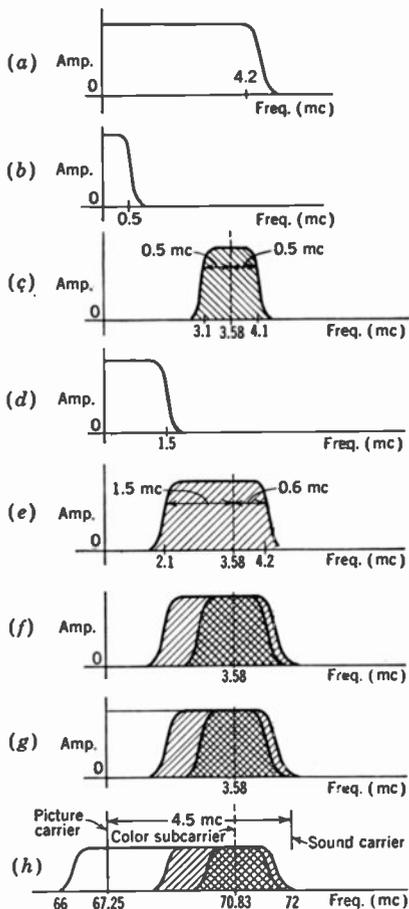


FIG. 24-5. Bandwidth requirements of luminance and chrominance signal frequencies. (a)  $Y$  video signal; (b)  $Q$  video signal; (c)  $Q$  chrominance signal as side bands of 3.58-Mc suppressed subcarrier; (d)  $I$  video signal; (e)  $I$  chrominance signal; (f) total chrominance signal—crosshatch indicates where  $I$  and  $Q$  frequencies overlap; (g) colorplexed video signal  $S$  with  $Y$ ,  $I$ , and  $Q$ ; (h) side-band frequencies of colorplexed signal transmitted as amplitude modulation of channel 4 picture carrier.

Since the  $Y$  signal consists of the red, green, and blue video signals, the  $Q$  signal can also be expressed as

$$Q = 0.21R - 0.52G + 0.31B$$

It should be noted that the minus sign, for subtraction of either luminance or color values, merely indicates the addition of video signal voltages of negative polarity.  $R - Y$  and  $B - Y$  are color-difference voltages, specifying how much the color differs from white, which indicates saturation.

Opposite polarities of the  $Q$  video signal represent the complementary colors purple and green. Considering the primary color components, positive  $Q$  signal voltage contains minus-green, or magenta, with red and blue, which combine to produce magenta or purple hues. Negative  $Q$  signal voltage is primarily green. Comparative values of the  $Q$  video signal for the standard color-bar pattern are shown in Fig. 24-4e. Relative voltage values for each bar are indicated according to the red, blue, and green components. For magenta, as an example, which is a reddish-purple hue combining unit amounts of red and blue primary color voltages, the  $Q$  video voltage consists of 21 per cent red plus 31 per cent blue. This equals 0.52 for the magenta component in the  $Q$  signal. Note that the  $Q$  signal has its peak positive and negative amplitudes for the magenta and green bars, indicating that the positive values of  $Q$  voltage represent mainly purple hues and the negative ones mainly green. Plate IV shows a reproduced picture with the  $Q$  signal only.

The range of video frequencies from 0 to 0.5 Mc is used for the  $Q$  signal, corresponding to the large areas of picture information. This is illustrated in Fig. 24-5b, showing the  $Q$  video signal frequencies. In  $c$ , the  $Q$  signal frequencies are shown as sidebands of the 3.58-Mc suppressed subcarrier, which is the way the  $Q$  signal is transmitted in the broadcast channel. Both the upper and lower sidebands of the  $Q$  signal, extending approximately 0.5-Mc above and below the 3.58-Mc subcarrier frequency, are included.

**24-7. I Signal.** The color video signal designated the  $I$  signal is formed from the primary color signals in the matrix at the transmitter, as follows:

$$I = 0.74(R - Y) - 0.27(B - Y)$$

or, in terms of the primary color signals

$$I = 0.60R - 0.28G - 0.32B$$

Opposite polarities of the  $I$  video signal represent the complementary colors orange and cyan. Considering the primary color components, positive  $I$  signal contains primarily red combined with minus-blue, or

yellow, which add to produce orange. Negative  $I$  signal combines green plus blue to form cyan, with minus-red, which is cyan.

The bar pattern is shown in Fig. 24-4f with relative  $I$  voltage values for each bar indicated according to the red, blue, and green components. Taking the cyan bar, as an example, in terms of the component blue and green primary color voltages it produces 28 per cent negative green and 32 per cent negative blue voltage, which equals  $-0.60$  for the cyan component in the  $I$  signal. Note that the  $I$  signal has its peak positive and negative amplitudes for the red and cyan bars indicating that positive values of  $I$  voltage correspond mainly to red-orange hues and negative  $I$  voltage is mainly cyan. Plate III shows the reproduced picture with the  $I$  signal only.

The  $I$  signal is transmitted with the range of video frequencies from 0 to 1.5 Mc. This bandwidth corresponds to large areas of picture information, as for the  $Q$  signal, but in addition includes color information for smaller areas corresponding to video frequencies of 0.5 to 1.5 Mc. Compared to the  $Q$  signal, more bandwidth is used for the  $I$  signal because it represents the most important color information in the scene. The dominant objects in most scenes, such as people's clothing and the flesh tones, have orange-cyan color information primarily. Also, only the orange and cyan color information can be resolved by the eye for small color details. Figure 24-5d shows the video frequencies for the  $I$  signal, while in  $e$  the side-band frequencies of the  $I$  signal are shown as the modulation product of the 3.58-Mc suppressed subcarrier. Note that vestigial-side-band transmission with respect to the 3.58-Mc subcarrier is used for the  $I$  signal, only part of the upper-side-band frequencies being transmitted. The lower side band of the  $I$  signal includes the full 1.5-Mc bandwidth, while the upper side band extends approximately 0.5-Mc above the subcarrier frequency. Upper-side-band frequencies more than 0.5 Mc above the 3.58 Mc subcarrier cannot be transmitted because they would interfere with the sound carrier signal. Therefore, the 1.5-Mc bandwidth for the  $I$  signal can be considered in two parts: (1) modulating frequencies from 0 to 0.5 Mc, approximately, are transmitted with double side bands, as for the  $Q$  signal; (2) modulating frequencies of 0.5 to 1.5 Mc in the  $I$  signal are transmitted with the lower side band only.

Referring to  $f$  in Fig. 24-5, this shows the video frequencies transmitted for both the  $I$  and  $Q$  signals, as side bands of the 3.58-Mc subcarrier. The crosshatched area extending 0.5 Mc above and below 3.58-Mc indicates where the frequencies utilized for the  $I$  and  $Q$  signals overlap. Therefore, both  $I$  and  $Q$  color information are provided for the video frequency range from 0 to 0.5 Mc. The  $I$  signal alone can be utilized for color information corresponding to video frequencies in the range of 0.5 to 1.5 Mc.

The video frequencies for the colorplexed total video signal including

$Y$ ,  $I$ , and  $Q$  are shown in Fig. 24-5*g*. The corresponding r-f signal frequencies transmitted to the receiver are illustrated in *h*, using channel 4, 66 to 72 Mc, as an example.

**24-8. The Chrominance Signal.** The modulated 3.58-Mc color subcarrier, with the combined modulation products formed in the  $I$  and  $Q$  modulators contains all the color information in terms of  $I$  and  $Q$  color signals. Since the phase of the 3.58-Mc subcarrier is shifted  $90^\circ$  for the  $Q$  modulator, the  $Q$  component is  $90^\circ$  out of phase with the  $I$  component in the chrominance signal. Therefore, the chrominance signal consists of the  $I$  and  $Q$  signals in quadrature.

When two signal voltages  $90^\circ$  out of phase with each other are combined, the resultant is the vector sum of the quadrature components, just as in combining reactive and resistive a-c voltages. The amplitudes of the two quadrature components determine the magnitude of the resultant voltage, and the phase angle of the resultant varies with the relative value of one quadrature voltage compared to the other. In terms of the  $I$  and  $Q$  quadrature voltages, the instantaneous value of the resultant chrominance signal varies in amplitude and phase with varying  $I$  and  $Q$  signals. The amplitude of the resultant chrominance signal indicates the saturation of the color information in the picture, while the phase of the resultant chrominance signal corresponds to the hue. This is illustrated in Fig. 24-6.

*Color Values of the Chrominance Signal.* Taking the chrominance signal for yellow in Fig. 24-6*a*, as an example, the relative voltages of 0.32 for  $I$  and  $-0.31$  for  $Q$ , compared to 1 for white, are obtained with equal amounts of saturated red and green but no blue. Combining  $I$  and  $Q$  vectorially results in the chrominance vector  $C$  shown for yellow, with a phase angle between  $I$  and  $-Q$ . Less  $I$  and  $-Q$  voltage, but with the same proportions, provides the same phase angle for the resultant yellow chrominance signal, but with less amplitude corresponding to less saturation. The different hues and saturation values that can be obtained with the four main combinations of positive or negative  $I$  and  $Q$  voltages are illustrated in Fig. 24-6*a*, *b*, *c*, and *d*.

Summarizing the chrominance values in all four quadrants, the color wheel in Fig. 24-7 indicates any possible hue and saturation for the chrominance signal, approximately, according to its instantaneous phase and amplitude. Notice that opposite polarities of chrominance signal voltage correspond to complementary colors. For instance, the phase angles for yellow and blue differ by exactly  $180^\circ$ . In terms of the reproduction of color information by the kinescope, positive signal voltage can be considered as adding the color, while negative signal voltage is equivalent to subtracting from the white reproduced on the kinescope screen as the sum of the primary colors. Similarly, the chrominance signal's phase

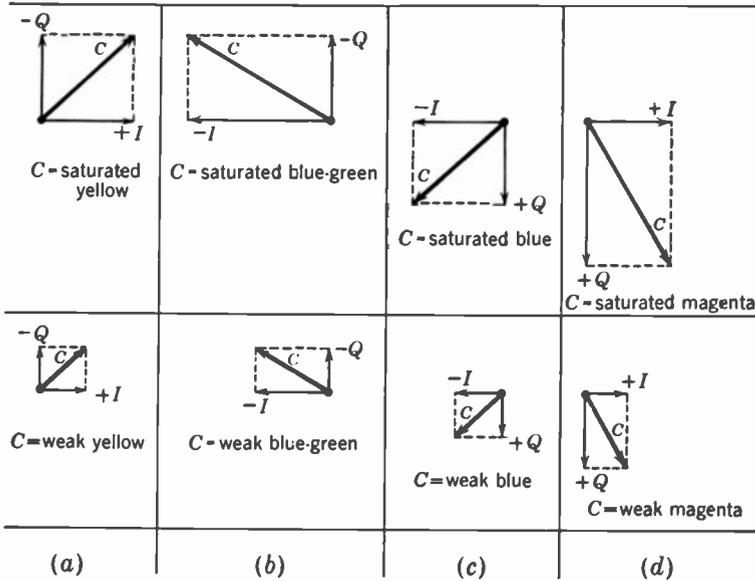


FIG. 24-6. Vector addition of  $I$  and  $Q$  signal voltages for four different hues. The phase angle of the resultant signal  $C$  indicates the hue while its length corresponds to saturation. (a) Yellow; (b) blue-green; (c) blue; (d) red-purple.

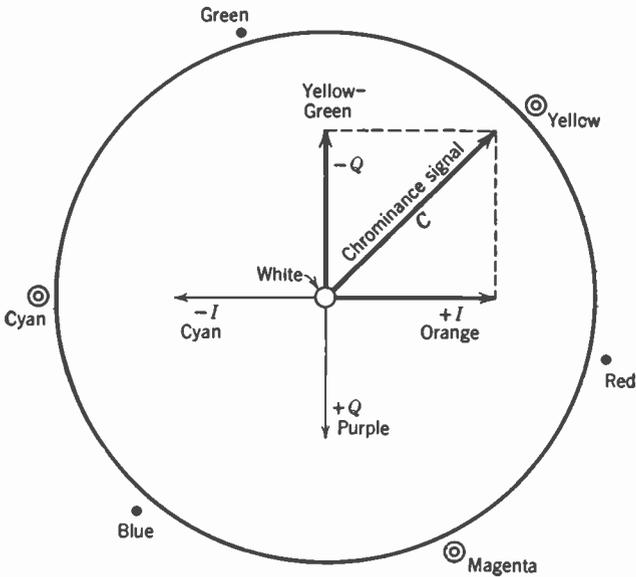


FIG. 24-7. Color wheel showing approximate chrominance values for different combinations of  $I$  and  $Q$  signal voltage. The  $-Q$  voltage is shown upward to conform to color positions on the standard chromaticity diagram in Fig. 24-38, which gives exact chrominance values.

angle can be considered a specific hue in terms of the polarity and amplitude of the component primary signal voltages for the kinescope.

In summary, then, the hue and saturation of the picture information are in the chrominance signal, while the luminance is in the  $Y$  signal. Therefore the transmitted signal includes three specifications of the picture information, converted from the original red, green, and blue signals, as follows:

1. Hue is transmitted as the instantaneous phase angle of the modulated 3.58-Mc color subcarrier.
2. Saturation is transmitted as the instantaneous amplitude of the modulated 3.58-Mc color subcarrier.
3. Luminance is transmitted as the instantaneous amplitude of the  $Y$  signal.

At the receiver, the modulated 3.58-Mc subcarrier chrominance signal is detected to recover the  $I$  and  $Q$  signals, which can be combined with the  $Y$  signal to produce the original red, green, and blue primary color voltages for the color kinescope.

*Resolution of the Chrominance Information.* The chrominance signal includes both  $I$  and  $Q$  information for video frequencies from 0 to 0.5 Mc. In terms of the corresponding size of the picture information, the video frequency of 0.5 Mc represents an area  $\frac{1}{25}$  of the width of the picture, as the entire visible trace of a horizontal line corresponds to approximately 20 kc. Picture information in the vertical direction corresponds to video frequencies less than 20 kc. As a result, for picture information corresponding to the background and smaller areas up to  $\frac{1}{25}$  of the picture width both the  $I$  and  $Q$  signals can be obtained from the chrominance subcarrier to be combined with the  $Y$  signal for reproducing the image in red, green, and blue. However, only the  $I$  signal has color information for smaller areas of picture information corresponding to 0.5 to 1.5 Mc. The video frequency of 1.5 Mc corresponds to an area approximately  $\frac{1}{150}$  the width of the picture. These smaller details of picture information, corresponding to areas  $\frac{1}{25}$  to  $\frac{1}{150}$  the picture width can be reproduced only in orange and cyan. The smallest details of picture information corresponding to video frequencies of 1.5 to 4.2 Mc, which usually represent outlines of the objects in the picture, are reproduced by the  $Y$  signal alone as brightness variations only, without any color.

**24-9. Matrix Circuits.** The matrix has the function of adding several input voltages in the desired proportions to form new combinations of output voltage. At the transmitter, the matrix has three input voltages corresponding to the red, green, and blue primary colors in the televised scene and forms the  $Y$ ,  $Q$ , and  $I$  output signals for transmission. In the receiver, the  $Y$ ,  $Q$ , and  $I$  signals are coupled into matrix circuits to form the original  $R$ ,  $G$ , and  $B$  primary color video signals for the tricolor

kinescope. The required proportions of the component voltages in the matrix can be obtained by means of resistive voltage dividers. Adding the proportioned voltages is accomplished by combining them across a common load resistance. Subtraction of a signal voltage is done by adding it in negative polarity to a signal voltage of positive polarity.

Typical matrix circuits are illustrated in Fig. 24-8. In *a*, the resistive voltage divider proportions the *R*, *G*, and *B* primary color video signals, in accordance with the resistances of  $R_1$ ,  $R_2$ , and  $R_3$  compared to the total resistance, to produce the *Y* signal output across the common load

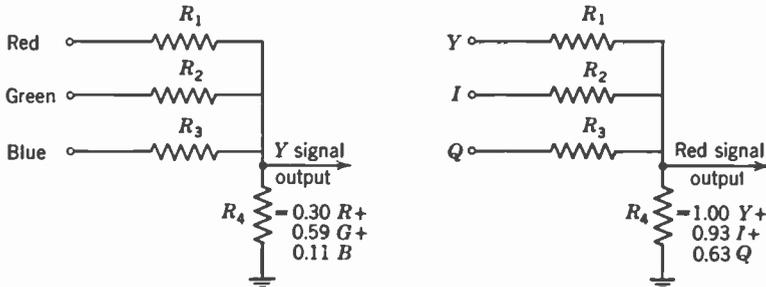


FIG. 24-8. Resistive matrix circuits: (a) for forming *Y* signal output; (b) for forming red signal output.

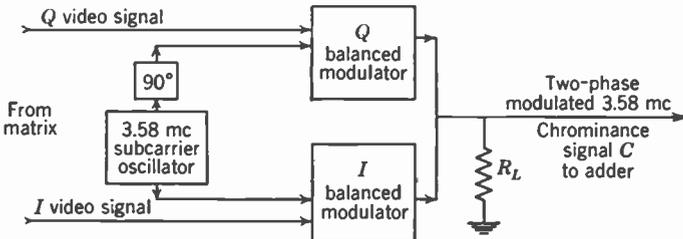


FIG. 24-9. Two-phase modulation of the suppressed color subcarrier by the *I* and *Q* video voltages produces the modulated chrominance signal.

resistor  $R_L$ . Similarly, different resistance values for separate voltage dividers can be used for proportioning *R*, *G*, and *B* in the amounts necessary to produce the *Q* and *I* signals at the transmitter. At the receiver, the voltage-divider arrangement can be used for the matrix circuits that form the primary color voltages from the *Y*, *I*, and *Q* signals. This is illustrated in *b* for a matrix forming *R* signal output.

**24-10. Suppressed Carrier Modulation of the Color Subcarrier.** The *I* and *Q* color-difference voltages obtained from the output of the matrix are coupled into the *I* and *Q* modulators at the transmitter, as illustrated in Fig. 24-9, in order to produce two-phase modulation of the 3.58-Mc color subcarrier, with the subcarrier suppressed. This has the advantage of reducing the amplitude of the 3.58-Mc beat between the

color subcarrier and the transmitted picture carrier, in order to eliminate the fine-dot pattern in the reproduced picture corresponding to the 3.58-Mc beat.

**Balanced Modulator.** The balanced modulator circuit in Fig. 24-10 illustrates how suppressed carrier modulation can be produced. This circuit consists essentially of two amplifiers having push-pull output and two input signals, one applied in parallel and the other in push-pull. With a balanced output circuit, the input voltage applied in parallel is canceled in the output circuit, since the signal current produces variations of opposite phase in the two primary sections of the output transformer. Push-pull input voltage, however, produces push-pull plate currents that reinforce each other in the output transformer. Referring to Fig. 24-10, the  $I$  modulating voltage is applied as push-pull input, while the 3.58-Mc color subcarrier is applied in parallel. Since both the  $I$  voltage and the

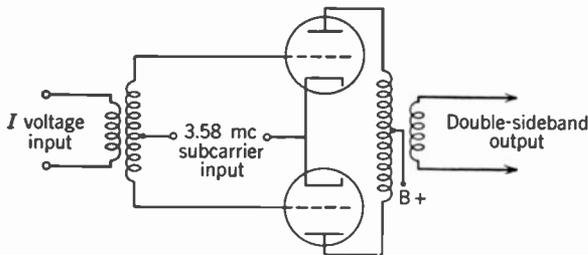


FIG. 24-10. Balanced modulator circuit for suppressed carrier modulation of the 3.58-Mc color subcarrier by the  $I$  video signal.

3.58-Mc subcarrier voltage control the plate current, the push-pull output current has the amplitude variations of the subcarrier produced by the  $I$  signal. The resultant output, therefore, is the modulation product consisting of double side bands above and below 3.58 Mc, but without the suppressed 3.58-Mc subcarrier frequency, which is canceled in the balanced output circuit. Although illustrated here for the  $I$  signal, the modulation product for the  $Q$  signal is obtained with another balanced modulator. The  $Q$  signal voltage is the push-pull input, with the parallel input being 3.58-Mc subcarrier voltage shifted by  $90^\circ$  from the phase of the subcarrier input to the  $I$  modulator, in order to produce the  $Q$  modulation product in quadrature with the  $I$  modulation product. The  $I$  and  $Q$  modulation products are then combined to produce the two-phase modulated, suppressed subcarrier, 3.58-Mc chrominance subcarrier signal.

**24-11. Chrominance Interlace.** The video signal frequencies corresponding to the televised scene are multiples of the 30-cps frame frequency with strong components at harmonics of the 15,750-cps horizontal line frequency, when there is no motion in the scene, since the same information is scanned at the frame rate. Therefore, all the picture infor-

mation is conveyed by these discrete harmonic frequencies at multiples of 30 cps, and there is no picture information for frequencies between the harmonics of 30 cps. If desired, another still picture could be televised simultaneously by timing the scanning to be offset by one-half the frame period, without any interference between the two signals. Both pictures would provide video signal frequencies at multiples of 30 cps, but the frequencies for one picture would be at 30, 60, 90 cps, etc., while the other picture had corresponding frequencies interleaved at the unused frequencies of 45, 75, 105 cps, etc. This interleaving of the frequency components of two different signals is called *frequency interlace*.

Figure 24-11 illustrates how an interfering signal frequency that is an odd multiple of one-half the line frequency interlaces the interference effect on successive scanning lines.

If the interfering signal has positive polarity at the start of line 1 to produce an increase of light from the kinescope screen, the interference produces a decrease of light at the start of the adjacent line 2 in the next scanning field. In addition, successive pairs of interlaced lines have opposite interference effects. Furthermore, in the next frame whatever interference effect that remains will be opposite in the two fields because the signal frequency is automatically an odd multiple of one-half the frame frequency when it is an odd multiple of one-half the line frequency. This space integration of opposite interference effects in the reproduction on the kinescope screen, which can be considered as chrominance interlace, minimizes the visibility of the chrominance signal in the luminance picture. Especially important is elimination of the fine-dot pattern caused by the 3.58-Mc beat between picture carrier and color subcarrier.

24-12. **Color Subcarrier Frequency.** Another consideration besides the chrominance interlace is the beat between the sound carrier and the color subcarrier at the difference frequency of approximately 0.9 Mc. This is objectionable because the relatively low beat frequency produces a coarse interference pattern in the reproduced picture. The 0.9 Mc beat is easily visible as a herringbone pattern in the colored picture areas when the sound carrier is modulated. Experiments have

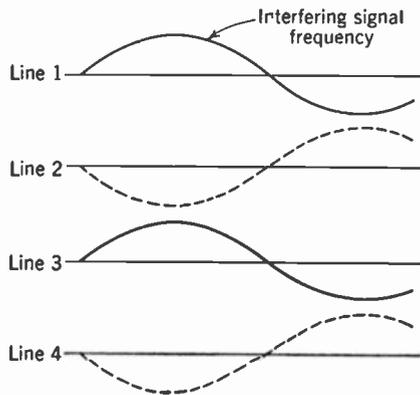


FIG. 24-11. Chrominance interlace for a frequency exactly an odd multiple of one-half the horizontal line-scanning frequency. Solid lines represent first and odd vertical scanning fields; dotted lines second and even fields. Missing part of sine wave corresponds to horizontal fly-back time.

shown that this beat interference is much less objectionable if its frequency is an odd multiple of one-half the horizontal-line frequency, because of the resultant interlace. However, the sound carrier frequency must be 4.5 Mc from the picture carrier to provide the 4.5-Mc sound intermediate frequency for intercarrier-sound receivers. It is important to remember now that the frequency interlace principle is accomplished in terms of the horizontal-line scanning period. Therefore, in order to provide the desired interlacing for frequencies resulting from the beat between the color subcarrier and the sound carrier, the required 4.5-Mc separation between sound and picture carriers is defined as the 286th harmonic of the horizontal-line frequency. This is an even harmonic, resulting in no frequency interlace for 4.5 Mc, but since this beat usually has low amplitude in the picture signal circuits, the fine pattern is not objectionable in the reproduced picture. Making 4.5 Mc an even harmonic, with the color subcarrier an odd harmonic, automatically makes the beat frequency of exactly 0.920455 Mc between the color subcarrier and sound carrier an odd multiple of one-half the line frequency and results in reduction of the 0.9-Mc beat interference by interlace.

Since 4.5 Mc is specifically the 286th harmonic of the line frequency, the horizontal-line scanning frequency becomes

$$H = \frac{4.5 \text{ Mc}}{286} = 15,734.26 \text{ cps}$$

With 525 horizontal lines per frame and  $262\frac{1}{2}$  lines per field, the vertical-field scanning frequency becomes

$$V = \frac{15,734.26 \text{ cps}}{262.5} = 59.94 \text{ cps}$$

The color subcarrier frequency is chosen as the 455th harmonic of one-half the horizontal-line scanning frequency, which becomes

$$C = 455 \left( \frac{15,734.26 \text{ cps}}{2} \right) = 3.579545 \text{ Mc}$$

Although the horizontal-line scanning frequency and vertical-field scanning frequency are slightly different from the values of black-and-white television, the difference is small enough to be within the permissible tolerance for variations from 15,750 and 60 cps. Since the deflection sync transmitted to the receiver has the same horizontal and vertical frequencies as the scanning frequencies, the slight change presents no problem in synchronizing the horizontal and vertical deflection in either monochrome or color receivers. The 3.58-Mc subcarrier frequency is made an odd multiple of one-half the line-scanning frequency by using the

subcarrier signal to lock in the synchronizing signal generator at the transmitter. As a result of the interlace principle employed, the chrominance signal can be transmitted in the same 6-Mc channel as the luminance and sound signals, with substantial freedom from interference by the 3.58-Mc color subcarrier signal.

**24-13. Color Synchronization.** Figure 24-12 shows the details of the 3.58-Mc color sync burst transmitted as part of the total composite video signal for color synchronization in color receivers. Specifically, the color burst synchronizes the phase of the 3.58-Mc local oscillator that reinserts the subcarrier needed for the *I* and *Q* demodulators in the receiver to detect the color-difference signals. The color synchronization is necessary to establish the correct hues corresponding to the phase of the chrominance signal. Then the color sync circuits in the receiver can hold the hue values steady. The burst is 8 to 10 cycles of the 3.58-Mc subcarrier on the back porch of each horizontal sync pulse. There is no color sync during vertical blanking time, in order to minimize the effect of the color burst in the sync circuits of monochrome receivers. Notice that the average value of the color burst coincides with the pedestal level of the composite video signal voltage. This prevents the horizontal sync circuits from interpreting the color burst as an increase in signal voltage, so that the color burst will not be mistaken for a horizontal sync pulse. In color receivers, the color sync is recovered by using a gating circuit that keys in the signal immediately after the horizontal sync pulses, coincident with the burst time. There is no color sync for monochrome transmission, so that the color receiver can recognize the monochrome signal by the absence of the color burst and disable the chrominance section.

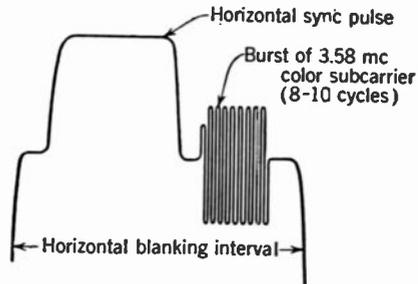


FIG. 24-12. The color sync burst.

There is no color sync for monochrome transmission, so that the color receiver can recognize the monochrome signal by the absence of the color burst and disable the chrominance section.

**Reference Color Phase.** Figure 24-13 illustrates how the hues of the modulated chrominance subcarrier signal are determined by its varying phase angle with respect to the constant phase angle of the color sync burst. However, the angles are specified with respect to the horizontal *B - Y* axis as the conventional phase angle of  $0^\circ$ . Therefore, the color reference burst is at  $180^\circ$ , the phase of the *Q* signal voltage is at  $33^\circ$ , while the phase of the *I* signal voltage in quadrature with the *Q* axis leads the *Q* voltage by  $90^\circ$ . The hue of the chrominance signal corresponds to its phase angle as it varies between 0 and  $360^\circ$  with varying amounts of *I* and *Q* signal voltages. The color corresponding to the burst phase is yellow. The hue for zero phase angle is blue. Note, as an example of

how the hue of the chrominance signal is determined by its phase angle with respect to the sync-burst phase, that when yellow<sup>1</sup> picture information is scanned at the transmitter, the phase angle of the sync burst is made the same as the phase of the modulated chrominance subcarrier signal. A chrominance signal with a phase angle  $180^\circ$  from the reference burst will then be blue (a primary hue differs from its complementary color by exactly  $180^\circ$ ). For the example shown in Fig. 24-13, the chrominance  $C$  has a phase angle of approximately  $60^\circ$ , or a  $120^\circ$  lag from burst phase, making the hue of the chrominance signal magenta.

**24-14. Choice of Chrominance Axes and Bandwidth.** It is interesting to note the reasons why transmission of the chrominance information in terms of the standard reference values indicated in Fig. 24-13 represents

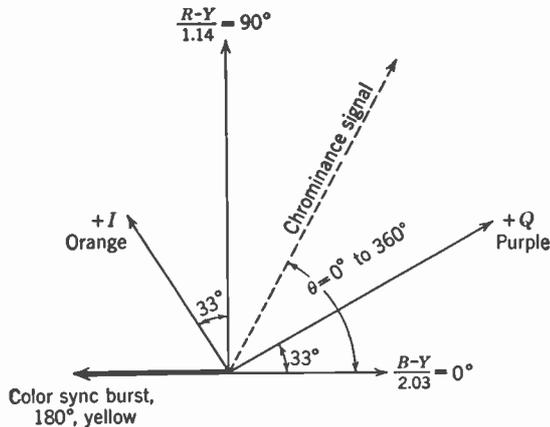


FIG. 24-13. Color reference phase of the sync burst.

the optimum use of the 6-Mc broadcast channel in the color television system. To begin with, transmitting the color information in terms of how much it differs from white is preferable to transmitting the primary colors, because with color-difference voltages, the chrominance signal reduces to zero for white picture information. This minimizes the dot interference pattern in white areas of the picture caused by the 3.58-Mc beat between the chrominance signal and the r-f picture carrier. For picture information reproduced in three colors by means of two color-difference voltages and the luminance signal, the  $R - Y$  and  $B - Y$  signal voltages would be suitable. However, if only one color signal voltage can be used, the  $I$  signal is preferable for color reproduction. As the eye is more sensitive to orange-cyan color variations in small details, the axis for the  $I$  signal has been chosen to correspond to colors between reddish-orange and cyan.

<sup>1</sup>The color values given here are approximate. Saturated yellow with exactly equal amounts of red and green actually corresponds to a phase angle of  $13^\circ$  delay from the burst phase.

The orange-cyan axis is established by making the phase angle for  $+I$  lag the reference burst corresponding to yellow by  $57^\circ$ . Since the  $Q$  signal is automatically  $90^\circ$  from the  $I$  signal, the  $Q$  axis will correspond to colors between purple and yellow-green and the phase angle for the  $+Q$  signal will lead the  $B - Y$  zero axis by  $33^\circ$ .

The transmitted chrominance signal, then, consists of the quadrature  $I$  and  $Q$  color-difference signals shifted by  $33^\circ$  with respect to the  $B - Y$  axis. At the receiver, however, the detected color video signals depend upon the phase of the 3.58-Mc subcarrier voltage reinserted in the detector, from the receiver's color reference oscillator. With the same  $33^\circ$  phase shift from the zero axis for the 3.58-Mc color subcarrier reinserted in the color demodulators, the detectors recover the original  $I$  and  $Q$  color video signals from the modulated chrominance signal. Without the  $33^\circ$  phase shifted in the 3.58-Mc reinserted carrier voltage, the detectors produce  $B - Y$  and  $R - Y$  instead of  $Q$  and  $I$ , respectively. This is often done in receivers that do not utilize the color detail of 0.5 Mc to 1.5 Mc in the  $I$  signal, because the larger areas of picture information corresponding to video frequencies up to 0.5 Mc can be reproduced in red, green, and blue with simpler circuits by using the  $B - Y$  and  $R - Y$  color-difference voltages. It should also be noted that different values of phase angle can be used for the 3.58-Mc subcarrier reinserted in the color demodulators and the desired color values obtained by the required proportions in the receiver matrix.

**24-15. Composite Video Signal Waveform.** Formation of the total video signal, combining the luminance signal  $Y$  and chrominance signal  $C$ , is illustrated in Fig. 24-14 in successive steps from the primary red, green, and blue color video voltages to the colorplexed total video signal transmitted to the receiver. Starting with the primary color video signals in  $a$ ,  $b$ , and  $c$ , the voltages are shown for the time of scanning one horizontal line containing the color bars indicated. The relative voltage amplitudes are indicated in terms of 100 per cent color video voltage for a fully saturated color. The luminance, or  $Y$ , signal in  $d$  shows the brightness component for each color bar. The  $I$  and  $Q$  signals in  $e$  and  $f$  have the relative voltage values indicated according to the proportions of the primary colors in the color-difference voltages. Note that the  $I$  and  $Q$  voltages can have positive or negative values. The  $Y$ ,  $I$ , and  $Q$  video voltages are the output signals from the matrix at the transmitter, formed from the  $R$ ,  $G$ , and  $B$  video input signals.

The  $I$  and  $Q$  video voltages from the matrix are then coupled to the  $I$  and  $Q$  modulators, with the 3.58-Mc subcarrier voltage, to produce the modulated chrominance signal  $C$ . Figure 24-14g illustrates the 3.58-Mc subcarrier amplitude-modulated by the  $I$  and  $Q$  signals. The amplitude of the 3.58-Mc modulated signal is obtained by vector addition of the quad-

perature  $I$  and  $Q$  relative voltage amplitudes. As an example, for the red color bar,  $I$  is 0.6,  $Q$  is 0.21, and the resultant vector  $C$  is  $\sqrt{0.6^2 + 0.21^2}$  or  $\sqrt{0.40}$ , approximately, which equals 0.63. Although not indicated in the figure, the phase angle for the 3.58-Mc chrominance signal would vary

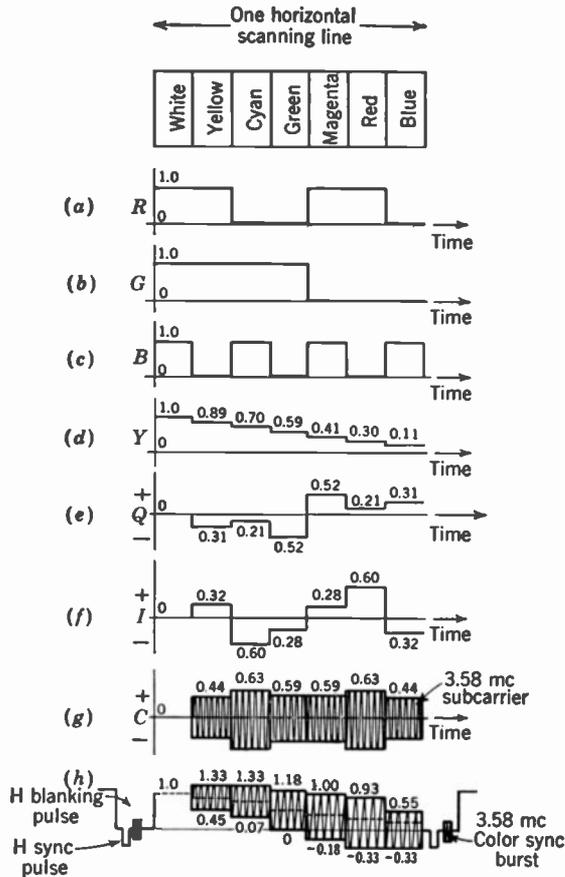


Fig. 24-14a-h. Construction of colorplexed composite video signal from  $Y$ ,  $I$ , and  $Q$  voltages. (a) Red video; (b) green video; (c) blue video; (d) luminance video; (e)  $Q$  video; (f)  $I$  video; (g) 3.58-Mc subcarrier chrominance signal  $C$  modulated by  $I$  and  $Q$  in quadrature; (h) colorplexed total video signal  $S$ , including  $Y$ ,  $I$ , and  $Q$  with standard blanking, deflection sync, and color sync burst.

for the different color bars in accordance with each hue. Complementary colors such as cyan and red have the same peak amplitude but opposite phase angles differing by  $180^\circ$  in the 3.58-Mc subcarrier signal.

In the adder section at the transmitter the modulated 3.58-Mc chrominance signal can then be combined with the luminance signal, with color sync and deflection sync and blanking added, to produce the complete

colorplexed video signal  $S$  shown in Fig. 24-14*h*. Although only horizontal sync and blanking pulses are shown here, corresponding to one line of signal, the total video signal voltage  $S$  includes the standard blanking and synchronizing voltage waveform for horizontal and vertical scanning, as in monochrome television. Notice that adding the  $Y$  video signal has the effect of shifting the average-value axis of the modulated 3.58-Mc color subcarrier signal from zero to a new level equal to the luminance voltage value of the corresponding color. Therefore, the variations in

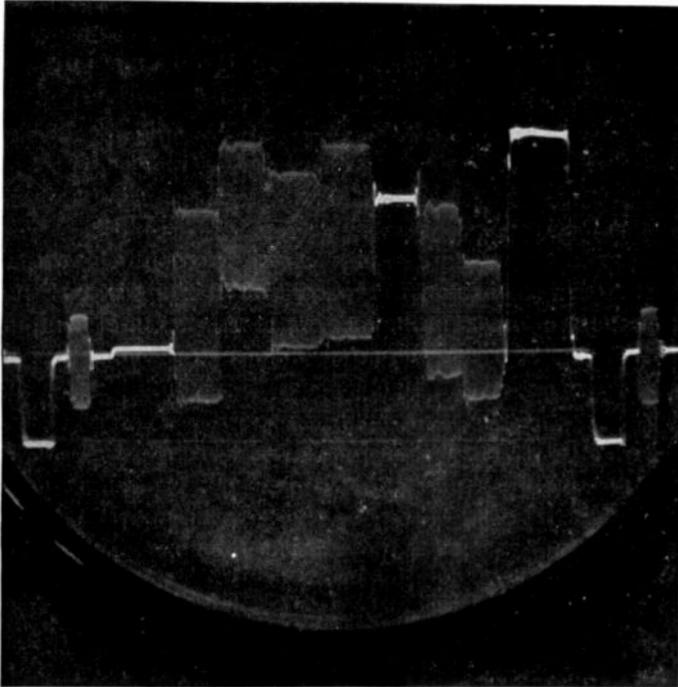


FIG. 24-14*i*. Oscillogram of colorplexed total composite video signal  $S$ . From left to right: horizontal sync pulse; 3.58-Mc color sync burst; black, red, yellow, green, cyan, gray, magenta, blue, and white bars. (Philco Corporation.)

average voltage of the  $S$  signal represent the luminance variations, while the instantaneous variations of the 3.58-Mc subcarrier voltage correspond to the chrominance information. Figure 24-14*i* shows an oscillogram of the  $S$  signal.

The colorplexed video signal is transmitted to the receiver as the envelope of the amplitude-modulated picture carrier wave in the broadcast stations' assigned channel, along with the associated FM sound signal on a separate carrier separated by 4.5 Mc from the picture carrier frequency, as in monochrome transmission. Vestigial-side-band transmission is used for the picture carrier wave in the standard 6-Mc broadcast

channel, modulated by the total video voltage with negative polarity of transmission. Maximum white information produces minimum carrier amplitude of 10 to 15 per cent and the blanking level is  $75 \pm 2.5$  per cent of the peak carrier amplitude produced by the tip of sync. The *black setup* interval is specified as  $7.5 \pm 2.5$  per cent, which means that maximum black values in the picture produce signal voltage 5 to 10 per cent less than the blanking level. Standardizing the setup helps in providing suitable d-c restorer action for the video signal. Correct reinsertion of the d-c component and linear response for the video signal are especially important in color television, as a change in relative amplitudes varies the color information.

In addition to the fact that the minimum carrier level of 10 to 15 per cent for maximum white is desirable for intercarrier-sound receivers, which need the picture carrier to beat with the sound carrier to produce the 4.5-Mc intercarrier sound, modulation of the carrier to less than 10 per cent causes serious amplitude and phase distortion. The modulation percentage is adjusted so that maximum white signal has the minimum carrier amplitude of 10 to 15 per cent. Furthermore, the chrominance signal is compressed before modulation so that saturated colors are transmitted at 75 per cent of full amplitude. Then the overshoots of 33 per cent below white level for fully saturated yellow and cyan, shown in Fig. 24-14*h*, and the 33 per cent overshoots above blanking level for saturated red and blue are practically eliminated in the colorplexed signal transmitted to the receiver.

At the receiver, the amplitude-modulated picture carrier signal is rectified in the video detector to recover the colorplexed total composite video signal. In monochrome receivers, the 3.58-Mc color subcarrier in the video signal has practically no effect because there are no color demodulators to detect the chrominance signal; the video-frequency bandwidth is usually less than 3.58-Mc, attenuating the color subcarrier; and the chrominance interlace minimizes 3.58-Mc interference by cancellation. However, the required luminance signal is present as the variations in average level of the 3.58-Mc color signal. In color receivers, these variations in average level provide the luminance signal for the *Y* video amplifier, while the color demodulators detect the 3.58-Mc chrominance signal to recover its modulation information corresponding to the color video signals.

**24-16. Color-receiver Requirements.** Figure 24-15 illustrates the circuits needed in a color television receiver in block diagram form, with signal waveshapes. Starting at the antenna, the requirements are essentially the same as in a monochrome receiver for the modulated r-f picture carrier signal and the associated sound carrier signal. The r-f tuner selects the desired station and converts the r-f signal frequencies

to the receiver's i-f signal frequencies. In the picture i-f section, the sound takeoff circuit couples the sound carrier signal to the sound i-f section, which usually has a separate sound converter to provide the 4.5-Mc intercarrier-sound signal. The picture i-f amplifier, with an overall bandwidth of 4.1 Mc, has four or five stages to provide enough signal for the video detector. Up to the video detector the signal waveform contains the colorplexed total video signal voltage  $S$  as the amplitude-modulation envelope of the r-f picture carrier. With a simple diode detector having sufficient bandwidth, as in monochrome receivers, the video detector demodulates the AM carrier to produce in its output the total video signal voltage  $S$ , containing the luminance signal  $Y$  and the 3.58-Mc modulated chrominance subcarrier signal  $C$ . The main difference between color and monochrome receivers in the r-f and i-f circuits is that in color receivers the r-f side-band frequencies corresponding to the high video frequencies around 3.58-Mc are essential for the color information in the picture, while in monochrome receivers the high video frequencies are less critical, since they provide only outline detail in the reproduced picture. Similar functions in the color television receiver include the circuits needed for intercarrier sound, deflection sync for vertical and horizontal scanning with horizontal a-f-c, flyback high voltage supplied from the horizontal output circuit, and the low voltage power supply for  $B^+$  voltages, as in monochrome receivers.

After one video amplifier stage with a bandwidth of 4.1-Mc functioning as a preamplifier for the  $S$  signal, the video circuits are divided into two parts—one for chrominance and the other for luminance. The  $Y$  video amplifier is for the luminance signal only. Usually, its bandwidth is restricted to 3.5 Mc, approximately, in order to minimize interference from the 3.58-Mc subcarrier signal. Then the amplified  $Y$  signal is coupled to the color matrix. The  $Y$  video amplifier includes a delay line, which allows the luminance signal to arrive at the color matrix at the same time as the color video signals. The delay line is usually about one foot of coaxial line for approximately 1  $\mu$ sec delay. The delay in the amplifiers results from their reactance. Circuits with more bandwidth have less time delay for the signal, as the proportion of reactance to resistance is smaller than in narrow-band circuits. Therefore, the wide-band  $Y$  amplifier needs additional delay, compared to the narrow-band chrominance circuits. Similarly, additional delay may be necessary for the  $I$  signal compared to the  $Q$  signal.

In the chrominance video circuits, the  $C$  amplifier's bandwidth is restricted to 2 to 4 Mc, approximately, in order to filter out low-frequency components of the  $Y$  signal and restrict the amplification to the side-band frequencies of the 3.58-Mc chrominance subcarrier signal. With total video signal input, the output voltage of the  $C$  amplifier will therefore

consist mainly of the chrominance modulation corresponding to the  $I$  and  $Q$  color signals. This chrominance subcarrier signal is coupled in parallel to both the  $I$  and  $Q$  demodulators. The demodulators also have input from the receiver's 3.58-Mc local oscillator, which is the reinserted subcarrier. As at the transmitter, the 3.58-Mc subcarrier voltage is shifted  $90^\circ$  for the  $Q$  signal. Note that the unmodulated continuous wave 3.58-Mc oscillator output for the  $I$  detector is called  $ICW$  and for the  $Q$  detector  $QCW$ . The detected output of the  $I$  demodulator is the  $I$  signal

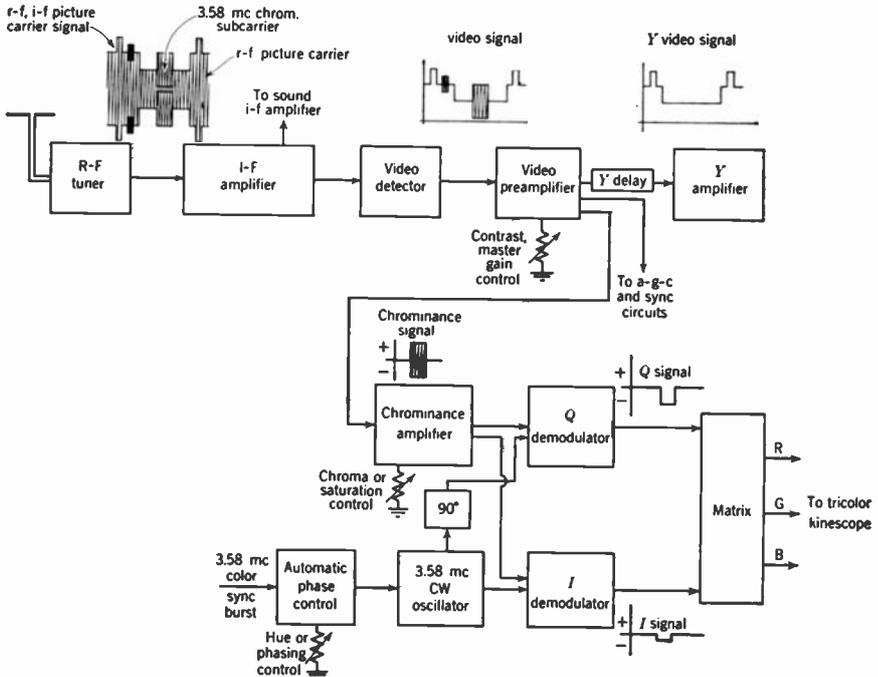


FIG. 24-15. Main sections of color television receiver. Signal wave shapes shown for vertical green bar on gray background.

having the video frequencies of the original  $I$  modulating voltage, from 0 to 1.5 Mc. Similarly the output of the  $Q$  demodulator is the  $Q$  video voltage, with a bandwidth from 0 to 0.5 Mc. The continuity of the chrominance signal frequencies in the receiver, from the r-f side-band frequencies of the AM picture carrier to the detected  $I$  and  $Q$  video signal frequencies is illustrated in Table 24-1.

With the  $I$  and  $Q$  voltages as input signals to the matrix, plus the  $Y$  signal from the luminance amplifier, the matrix circuits combine the  $I$ ,  $Q$ , and  $Y$  signals in the proportions necessary to produce the primary red, green, and blue color video voltages, which are coupled to the tricolor

TABLE 24-1. CONTINUITY OF CHROMINANCE SIGNAL FREQUENCIES FOR CHANNEL 4 (66 TO 72 Mc)

Circuit	Color subcarrier frequency, Mc	I Signal frequencies, Mc	Q signal frequencies, Mc
R-f	70.83	69.33-71.33	70.33-71.33
I-F*	42.17	41.67-43.67	42.67-43.67
Chrominance, before color demodulators	3.58	2.08-4.08	3.08-4.08
Color video, after color demodulators	0	0-1.5	0-0.5

\* Intermediate frequencies for receiver having 45.75-Mc i-f picture carrier, with local oscillator above r-f signal frequencies inverting upper r-f side-band frequencies to lower i-f side-band frequencies.

kinescope. An increase in the positive voltage for the green control grid, as an example, increases the beam current to produce more illumination from the green phosphor. Similarly, the red and blue video signals control the amount of red and blue light from the kinescope. The luminance signal also controls the screen illumination, but only in the proportions corresponding to the percentage of red, green, and blue that are found in white. Therefore, the variations in the  $Y$  signal reproduce shades of white and gray. Cutoff voltage results in no electron beam and no illumination from the kinescope screen, which corresponds to black, as in monochrome reproduction. For the signal illustrated in Fig. 24-15, green primary video color voltage is supplied for the green bar at the center of each horizontal scanning line. As a result, the green phosphor reproduces the green picture information down the center of the screen. During the remainder of the time, the  $Y$  signal provides all three colors in the necessary proportions for white, with the  $Y$  amplitude determining the shade of gray.

Color synchronization is supplied by the 3.58-Mc burst in the total video signal  $S$ . The video signal is coupled into a 3.58-Mc burst amplifier, which is gated by horizontal deflection pulses to produce output only during the burst interval, thereby separating the color sync burst. The amplified color sync burst voltage then controls the automatic-phase-control circuit used to hold the 3.58-Mc local subcarrier oscillator at the correct reference phase.

The contrast control in Fig. 24-15 is the usual contrast control on the front panel, as in monochrome receivers, to vary the over-all video gain. It is considered the *master gain control* because varying the output from the video preamplifier changes the signal level in both the luminance and chrominance circuits by the same amount, adjusting the contrast in the

reproduced picture without changing the hue or saturation. The 3.58-Mc chrominance signal level is varied by adjusting the color or saturation control, to adjust the ratio of chrominance signal to luminance signal. The color gain in the chrominance amplifier is generally a front-panel control, labeled either *color*, *saturation*, or *chroma*, to vary the intensity of the colors in the reproduced picture. The hue of the colors in the picture reproduction is controlled by the phase of the local 3.58-Mc oscillator signal. The adjustment for oscillator phase is a front-panel control labeled *hue* or *phasing*.

*Wide-band Color Decoder with Matrix for I and Q.* Figure 24-16 shows more details of the chrominance circuits in a receiver having a matrix for the *I* and *Q* signals. This arrangement utilizes the full bandwidth of the

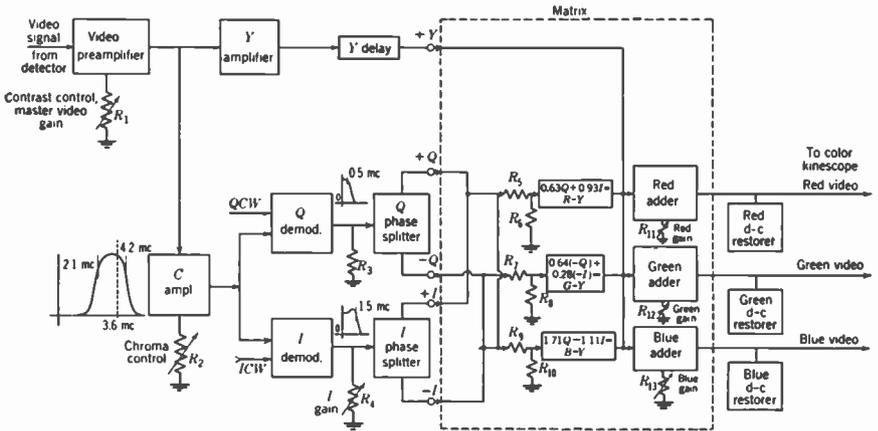


FIG. 24-16. Chrominance circuits for receiver having wide-band color decoder with matrix for *I* and *Q* color voltages.

chrominance signal. Note that the chrominance amplifier's band-pass curve in Fig. 24-16 includes the vestigial side bands of the *I* signal. Because frequencies in the *I* signal with only one side band produce quadrature crosstalk in the *Q* signal, the *Q* demodulator's output circuit must have a 0- to 0.5-Mc band-pass filter to exclude 0.5- to 1.5-Mc *I* frequencies from the *Q* circuits. Also, the *I* demodulator output circuit has a 0 to 1.5 Mc band-pass filter with a rising characteristic for the higher video frequencies transmitted with only one side-band, to compensate for the vestigial-side-band transmission of the *I* signal.

The matrix section in Fig. 24-16 combines five input voltages:  $+Q$ ,  $-Q$ ,  $+I$ ,  $-I$ , and  $+Y$ , in the proportions necessary to produce the red, green, and blue video signal in the output. Notice that  $+I$  and  $+Q$  signals, corresponding to orange and purple, are used to form red, which is between orange and purple. Similarly the green adder needs  $-I$  and  $-Q$

signals, which are cyan and yellow-green, while  $-I$  and  $+Q$  signals are used to form blue, between cyan and purple. Once the color-difference voltages are obtained, each is simply added to the  $Y$  signal to produce the primary red, green, and blue color signals for the kinescope. The d-c restorer stages are necessary for each of the three color video signals, since the d-c level must be correct independently for each color—otherwise, the color values of the picture information would vary with the background.

*Narrow-band Color Decoder with Matrix for  $R - Y$  and  $B - Y$ .* Figure 24-17 illustrates the narrow-band chrominance circuits of a receiver

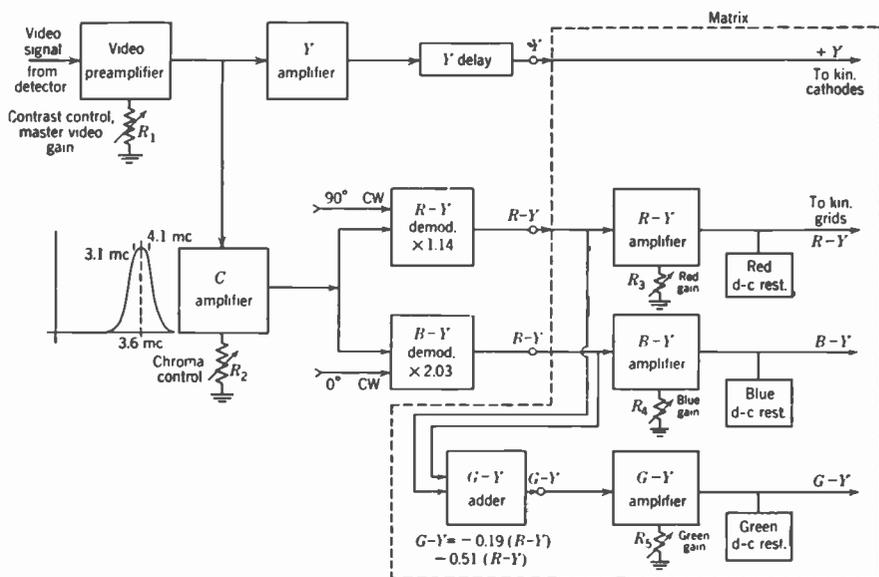


FIG. 24-17. Chrominance circuit for receiver having narrow-band color decoder with matrix for  $R - Y$  and  $B - Y$  color voltages.

that uses only the color information characterized by double-side-band transmission on the 3.58-Mc chrominance subcarrier. As a result, color video frequencies up to 0.5 Mc are utilized for three-color reproduction in red, green, and blue for large areas of picture information but there is no color for the smaller details corresponding to video frequencies of 0.5 to 1.5 Mc. In this arrangement, the demodulators recover the  $R - Y$  and  $B - Y$  modulation voltages on the 3.58-Mc subcarrier, instead of detecting the  $I$  and  $Q$  signals. There is no need for recovering the  $I$  signal, since the orange-cyan information for 0.5- to 1.5-Mc details cannot be reproduced with the restricted bandwidth. Fewer adder stages are necessary in the matrix, the color channels have the same bandwidth,

eliminating the need for equalizing the color time delays, and the reduced bandwidth allows more gain per stage. However, there is less color detail because of the limited bandwidth.

Detecting  $R - Y$  and  $B - Y$ , instead of  $I$  and  $Q$ , is accomplished by providing the required phase for the 3.58-Mc local-oscillator CW input to the demodulators. Referring back to the color phase reference in Fig. 24-13, notice that  $I$  and  $Q$  in the transmitted signal are displaced  $33^\circ$  from the  $R - Y$  and  $B - Y$  axes, respectively, in quadrature. Therefore, if the oscillator 3.58-Mc c-w output reinserted in the receiver is supplied to the demodulators in quadrature but not shifted by  $33^\circ$ , the detected signals correspond to hues along the  $R - Y$  and  $B - Y$  axes, instead of  $I$  and  $Q$  colors. For video frequencies up to 0.5 Mc, the end result is the same because the color-difference voltages are converted to the primary red, green, and blue video signals in either case. The multiplying factors indicated for the demodulators in Fig. 24-17 are necessary because the transmitted  $I$  and  $Q$  signals contain  $R - Y$  reduced by the factor of  $1/1.14$  and  $B - Y$  reduced by  $1/2.03$ . With the amplitude correction, the demodulator having c-w input of zero phase, or  $-180^\circ$  with respect to the reference burst phase, produces  $B - Y$  in the output. The quadrature c-w input then demodulates  $R - Y$ . To produce  $G - Y$ , the detected color-difference voltages are combined in the green adder. With  $R - Y$ ,  $B - Y$ , and  $G - Y$ , the primary red, green, and blue video are obtained by adding the  $Y$  signal to each.

*Color Signals for the Kinescope.* The color signals for the kinescope can be the red, green, and blue video signals, or the color kinescope can function as part of the matrix by adding the  $Y$  signal to the  $R - Y$ ,  $G - Y$ , and  $B - Y$  signals. Either type of operation can be used with the full-bandwidth matrix for  $I$  and  $Q$ , or the restricted-bandwidth matrix for  $R - Y$  and  $B - Y$ . When the primary color signals are coupled to the kinescope, the red, green, and blue amplifier stages must provide full 4.1-Mc bandwidth, since the combined signals produce the luminance information for fine details. With  $R - Y$ ,  $G - Y$ , and  $B - Y$  amplifier stages, their bandwidth can be limited to 0.5 Mc because the  $Y$  signal coupled to the kinescope provides the full bandwidth for luminance detail. In this arrangement  $R - Y$ ,  $G - Y$ , and  $B - Y$  are coupled to individual red, green, and blue kinescope control grids and the  $Y$  signal is coupled to all three cathodes. Since the color-difference voltage on the control grid and the  $Y$  signal on the cathode both control the amount of beam current, the net result in the reproduced picture is equivalent to adding the two signals. However, the kinescope is inherently nonlinear with enough signal input to operate over the entire characteristic curve, so that the reproduced colors are not so exact as the matrix values that can be obtained with linear amplifiers.

**Color Balance.** Because the red, green, and blue kinescope phosphors produce unequal amounts of light output—in the approximate ratio of 0.3:0.7:1, respectively—with the same value of beam current, compensation for the different kinescope operating characteristics must be provided to achieve color balance in the picture reproduction. Less negative control-grid bias and more positive accelerating grid voltage is necessary for red to produce color balance in the raster, without any video signal. For color balance in the picture, more a-c video signal is needed for red, in the approximate ratio of 1:0.8:0.7 for the red, green, and blue primary video signals, respectively.

**24-17. Chrominance Band-pass Amplifier Circuit.** A typical circuit for amplifying the modulated 3.58-Mc chrominance signal is illustrated

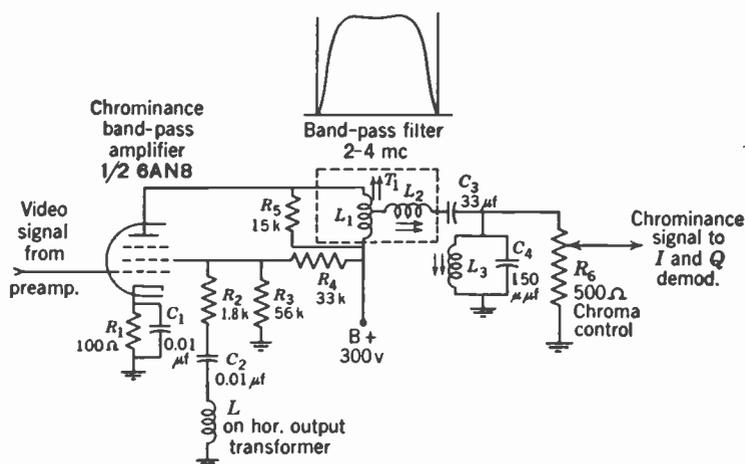


FIG. 24-18. Chrominance band-pass amplifier circuit.

in Fig. 24-18. With signal input from the video preamplifier, the amplified output of the chrominance amplifier is the 3.58-Mc suppressed sub-carrier signal's side-band frequencies, since the band-pass amplifier's plate load impedance is a tuned circuit resonant for this band of frequencies. Because the bandwidth is a very large percentage of the resonant frequency, the band-pass filter in the plate circuit is necessary. The broad frequency response is centered at about 3 Mc to include the side-band frequencies produced by the *I* and *Q* modulation. In a narrow bandwidth receiver for *R - Y* and *B - Y*, instead of *I* and *Q*, the chrominance amplifier's bandwidth can be restricted to the band of 3.1 to 4.1 Mc, approximately, centered at 3.6 Mc. Note that the screen grid is unbypassed in order to enable negative flyback-voltage pulses to be applied from the horizontal output transformer. This is done to cut off the chrominance amplifier during horizontal blanking time so that the

3.58-Mc color sync burst will not affect the d-c restorer in the blue video circuits, as burst phase corresponds to minus-blue. During active horizontal scanning time, however, the amplifier conducts to provide the 3.58-Mc chrominance signal for the demodulators.

**24-18. Chrominance Detector Circuits.** In order to detect the modulation information of a two-phase modulated carrier signal, the carrier must be reinserted by a local oscillator in the receiver to supply a reference carrier phase for the demodulator. Carrier reinsertion is necessary because the envelope of the suppressed carrier signal does not correspond to the original modulating voltage. Since the reinserted carrier has the same frequency and phase as the suppressed subcarrier, the demodulator is called a *synchronous detector*.

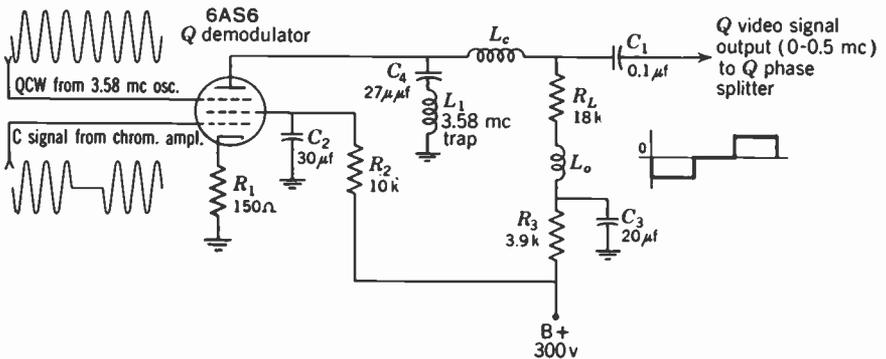


FIG. 24-19. Suppressor-grid synchronous detector circuit for *I* or *Q* demodulator stage. Wave shapes illustrated for yellow, white, and blue information.

The two-phase modulated chrominance signal contains amplitude modulation of the 3.58-Mc subcarrier in quadrature by the *I* and *Q* signals. To detect the chrominance information, the receiver needs synchronous detector circuits that rectify the 3.58-Mc modulated subcarrier to produce a video signal output proportional to its instantaneous amplitude, but the detector must respond only to a specific phase angle in the chrominance signal. Two synchronous detectors are needed, therefore, one for the *I* signal and the other for the *Q* signal. With the required filters in the detector output, the *Q* demodulator produces the desired *Q* signal, while ignoring the *I* signal's modulation, and the *I* demodulator produces *I* signal output without the *Q* signal. It should be noted that the demodulators can be used for detecting either *I* and *Q* signals or the narrow bandwidth *B - Y* and *R - Y* signals, depending on the relation of the local subcarrier phase to the reference-burst phase.

**Suppressor-grid Synchronous Detector.** The demodulator arrangement in Fig. 24-19 uses the suppressor grid of a pentode mixer tube to inject the 3.58-Mc local-oscillator output for synchronous detection of the

chrominance signal coupled to the control grid. Using the 6AS6 pentode, a typical circuit is shown for the *Q* demodulator, although a similar circuit arrangement can be used for the *I* demodulator. The 6AS6 has a fine-mesh structure for grid 3, enabling the suppressor to function as a control grid when it is biased negatively. As a result, the plate current is controlled by two voltages: the chrominance signal on grid 1 and the reinserted carrier signal from the 3.58-Mc oscillator on grid 3. The output voltage across the plate load resistor, therefore, contains a component proportional to the product of the two input voltages. For this reason, the circuit is often called a *product demodulator*. It should be noted, though, that only the difference frequencies for chrominance signal side bands having the same phase as the reinserted carrier produce output, as the quadrature components cancel. As a result, the *I* demodulator

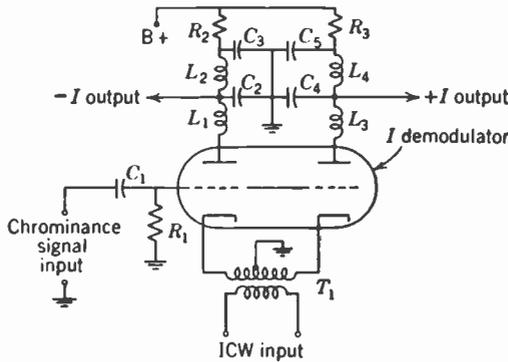


FIG. 24-20. Pulsed-envelope synchronous detector circuit for *I* or *Q* demodulator stage.

produces *I* signal output without *Q* signal, because the reinserted carrier has the same phase as the original suppressed 3.58-Mc subcarrier used for the *I* modulation. Similarly, the *Q* demodulator has the correct phase of reinserted 3.58-Mc carrier, because of the 90° phase shift with respect to ICW, providing output only for the quadrature chrominance signal produced by *Q* modulation.

Referring to the schematic diagram in Fig. 24-19, note that the demodulator's output circuit is essentially a video amplifier circuit with high-frequency compensation.  $R_L$  is the plate load resistor, with  $L_o$  and  $L_c$  the peaking coils for compensation up to 0.5 Mc, approximately, which is the upper limit of the *Q* video signal frequencies. In the *I* demodulator, the video output circuit is compensated for response up to 1.5 Mc, approximately. In a narrow-bandwidth receiver for  $R - Y$  and  $B - Y$ , both demodulators would have equal response up to 0.5 Mc.

*Pulsed-envelope Synchronous Detector.* Figure 24-20 illustrates a circuit arrangement in which the chrominance signal is applied to the two

control grids in parallel, with a twin triode, while the reinserted 3.58-Mc carrier voltage is applied in push-pull to each cathode. With sufficient carrier injection for each triode, the plate-current variations produced by the control-grid voltage, which result in envelope detection, are keyed on and off by the reinserted carrier voltage at the cathode. When the oscillator injection voltage is about four to ten times the amplitude of the chrominance signal, the output voltage across the plate load resistor varies with the phase angle difference between the chrominance and carrier voltages. Therefore, the circuit can be used as a synchronous detector for the chrominance signal to recover either the  $Q$  or  $I$  voltage. The same principle can also be applied with diode rectifiers. The balanced output for the  $I$  demodulator circuit in Fig. 24-20 provides both

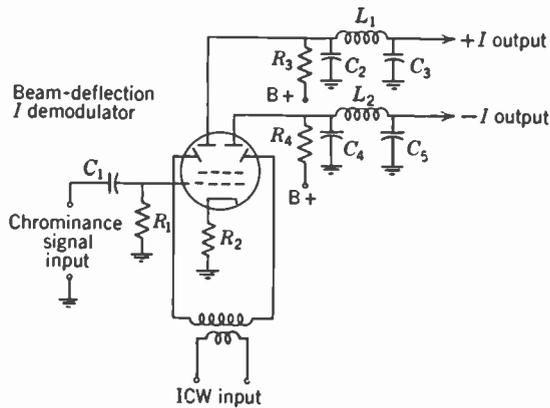


FIG. 24-21. Synchronous detector circuit using beam-deflection tube for  $I$  or  $Q$  demodulator stage.

$+I$  and  $-I$ , eliminating the need for an  $I$  phase-inverter stage in the matrix.

*Synchronous Detector with Beam-deflection Tube.* Figure 24-21 illustrates how a beam-deflection tube is used as a synchronous detector. The chrominance signal is coupled to the control grid and the reinserted oscillator signal is applied in push-pull to the pair of balanced deflectors in the tube. As a result, the electron beam varies in intensity according to the chrominance signal voltage on the control grid, while the beam is deflected back and forth between the two anodes by the deflectors. As the deflector voltage allows the beam to approach each anode, it provides more plate current. The output voltage across the plate load resistor for each anode, therefore, varies in accordance with the product of the two input voltages. With ICW phase for the reinserted carrier, then, the output is  $I$  signal voltage. In addition to the fact that the balanced output circuit provides both  $+I$  and  $-I$  signals, so that no  $I$  phase inverter

is necessary, the beam-deflection demodulator has relatively high signal output, with high input and output impedances.

**24-19. Color Synchronizing Circuits.** The reference phase for the hue of the chrominance signal is transmitted to the receiver as a burst of 8 to 10 cycles of the 3.58-Mc subcarrier on the back porch of each horizontal synchronizing pulse, as illustrated in Fig. 24-22*a*. In *b*, the phase angles of the component color voltages in the chrominance signal are indicated vectorially. In *c*, these phase angles are shown in terms of one sine wave. The sine wave at the top indicates the zero phase of  $B - Y$ . When the instantaneous value of the chrominance signal has this phase, the synchronous detectors produce output voltage corresponding to blue. Similarly, the phase angles indicated below correspond to the hue of  $Q$ ,  $R - Y$ ,  $I$ , and the sync burst, going counterclockwise in the vector

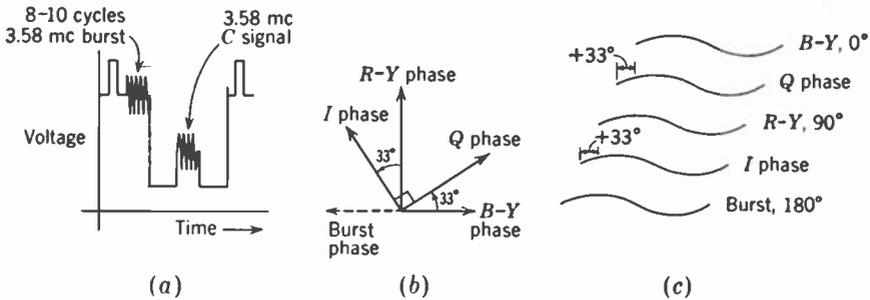


FIG. 24-22. The color sync burst and hue values in terms of its reference phase angle. (a) Burst consisting of 8 to 10 cycles of 3.58-Mc subcarrier; (b) phase angles of vectors for chrominance signals and burst phase; (c) phase angles of one sine wave for chrominance signals and burst phase.

diagram for increasing leading angles. Intermediate phase angles correspond to intermediate hues. The hues are established in the receiver by locking the phase of the 3.58-Mc local c-w oscillator with respect to the phase of the transmitted color sync burst.

**Burst Separator and Amplifier.** In order to utilize the phasing information in the color sync burst, it must first be separated from the composite video signal. As illustrated in Fig. 24-23, the color sync separation is accomplished in a 3.58-Mc amplifier, gated into conduction by pulses from the horizontal output circuit. The input voltage is composite video signal from the preamplifier, but because of the gating the amplified output contains only the 3.58-Mc burst signal present during horizontal blanking time on the back porch of the horizontal sync pulses. In the cathode circuit  $R_3$  and  $R_1$  form a d-c voltage divider for the 300-volt  $B^+$  voltage, with a return to chassis ground through the low-resistance coil  $L$ . The positive voltage applied to the cathode makes the control grid negative and cuts off plate current in the tube. However, during horizontal fly-

back time the negative voltage pulse supplied by the coil winding  $L$  on the horizontal output transformer drives the cathode negative, making the control grid positive, to gate the tube into conduction. This gating pulse must usually be delayed and shaped to make it coincident with the color sync burst, so that only color sync voltage output is produced. The 3.58-Mc color sync burst voltage is then coupled to the automatic-phase-control circuit to provide the phase reference for controlling the receiver's 3.58-Mc oscillator.

*Automatic-phase-control Circuit.* The a-p-c circuit for color synchronization shown in Fig. 24-24 is essentially the same as the sine-wave type of automatic frequency control, generally called Synchro-Lock, which is used for horizontal synchronization in many monochrome receivers. In

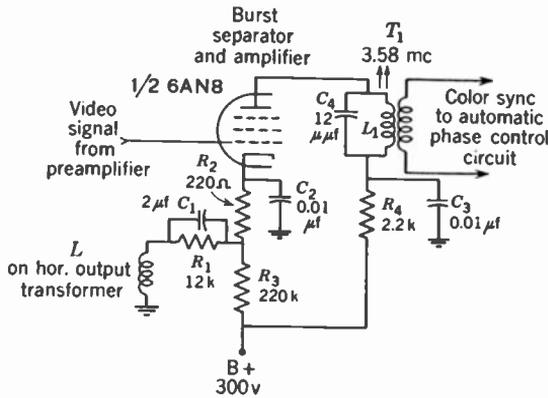


FIG. 24-23. Color sync burst separator and amplifier circuit.

the color sync circuit, however, the frequency of the 3.58-Mc local oscillator is crystal-controlled. Its phase is varied by the reactance control tube. The phase of the reinserted c-w subcarrier voltage must be controlled exactly, within  $\pm 5$  degrees, because this determines the fidelity of hue values in the color reproduction.

Referring to Fig. 24-24, both the burst sync voltage and the oscillator signal are coupled to the phase detector. This is a balanced discriminator circuit, using the grid-cathode circuit of the 6U8 triode sections  $V_{129B}$  and  $V_{130A}$  as two diode rectifiers. The control  $R_{287}$  balances the two diode rectifier circuits. The phase detector compares the phase of the burst voltage to the phase of the oscillator voltage and produces d-c voltage output proportional to the phase difference between the two input voltages. When the two input voltages are  $90^\circ$  out of phase, the d-c output voltage is zero. A difference in phase, however, results in a d-c correction voltage of either positive or negative polarity to indicate incorrect phase in the oscillator. The d-c correction voltage is applied to the



reactance tube  $V_{131A}$ , in order to control the amount of reactance in its plate-to-cathode circuit. Since the reactance tube's plate circuit is in parallel with the crystal in the oscillator grid circuit, the d-c control voltage enables the reactance tube to vary the phase of the oscillator output. The control circuit shifts the phase of the oscillator until the phase detector produces zero output voltage indicating  $90^\circ$  phase for its two input voltages.

The phasing control  $R_{327}$  is adjusted to make the oscillator voltage correspond to the phase of  $I$ , producing ICW for the  $I$  demodulator. The ICW phase is  $(90 - 33^\circ)$  lagging the burst phase. Providing an additional  $90^\circ$  lag in phase by means of the quadrature c-w amplifier  $V_{116B}$ , results in oscillator voltage of QCW phase in the output of the QCW amplifier. It should be noted that without the  $33^\circ$  angle, the oscillator output voltage has the phase of  $R - Y$  and the quadrature phase angle corresponds to  $B - Y$ . The phasing control, often labeled the *hue*

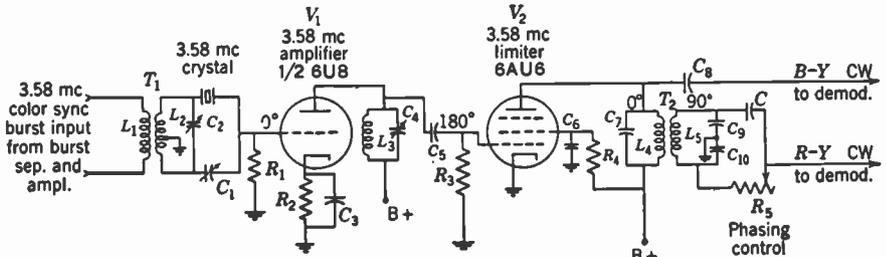


FIG. 24-25. Crystal-filter color sync circuit.

*control*, is usually available on the receiver's front panel to be adjusted for the correct hue of a known object. Proper flesh tones, as an example, provide a good guide in adjusting the phasing control for correct hues.

The color phasing amplifier  $V_{130B}$  and quadrature c-w amplifier  $V_{116B}$  in Fig. 24-24 are fixed-tuned amplifiers peaked at 3.58 Mc. The plate load impedance is a tuned transformer, with both the primary and secondary resonant at 3.58 Mc. The input and amplified output voltage is the oscillator c-w voltage having just the single frequency of 3.58 Mc. In such an amplifier stage, the voltage across the resonant plate circuit is  $180^\circ$  out of phase with the grid input voltage, but the signal voltage output from the tuned secondary is  $90^\circ$  out of phase with the primary plate signal voltage, leading or lagging depending on the transformer connections.

*Crystal-filter Color Sync Circuit.* A simpler arrangement for color synchronization utilizes the quartz crystal filter or ringing circuit illustrated in Fig. 24-25. The separated color sync burst from the burst separator and amplifier stage shock-excites the 3.58-Mc crystal, making it oscillate for the duration of each horizontal line interval. The resultant

3.58-Mc oscillations are amplified by  $V_1$  and limited in  $V_2$  to provide the reinserted carrier voltage. There is no 3.58-Mc local oscillator, as the ringing crystal supplies the required c-w signal, in the phase controlled by the sync burst. The transformer at the plate of the limiter is used to provide c-w signals of zero phase and  $90^\circ$  for the  $B - Y$  and  $R - Y$  demodulators. Note the phasing control  $R_5$  for hue adjustments in the limiter's output circuit. If desired, the phase can be shifted  $33^\circ$  to provide ICW and QCW output.

The crystal-filter color sync circuit requires good limiting of the 3.58-Mc output. If the 3.58-Mc carrier for the demodulators is not constant, the changes of amplitude in the reference wave will cause variations of color shading in the picture. The limiting must also be effective in rejecting impulse noise to prevent sudden amplitude changes in the 3.58-Mc carrier, which cause color streaks in the picture. The use of the crystal in a bridge circuit, with  $C_1$  adjustable to balance the bridge, has the advantage of providing extremely narrow bandwidth to minimize thermal noise.

*Color Killer Circuit.* The color killer circuit disables the chrominance signal circuits when a monochrome signal is being received. This is done by having the color killer stage cut off the chrominance amplifier when there is no color sync burst. Referring to Fig. 24-24, note that the phase detector in the color sync circuits produces negative d-c bias voltage for the control grid of the color killer stage  $V_{119A}$ . This is not the d-c control voltage for the oscillator control tube, which is obtained from the center of the load resistance between the two diodes, but is negative d-c output voltage from one discriminator diode. With color sync voltage input to the phase detector, the negative bias voltage is enough to cut off plate current in the color killer stage. Therefore, it does not cut off the chrominance amplifier. In the crystal-filter color sync circuit of Fig. 24-25, the grid-leak bias voltage from the limiter  $V_2$  can be used for the negative d-c voltage to cut off the color killer stage.

Without any burst voltage in the color sync circuits, however, there is no negative bias for the color killer and it conducts plate current. The plate voltage required for conduction is obtained by positive horizontal deflection pulses from the winding  $C-D$  on the output transformer in Fig. 24-24. When the color killer stage conducts, it produces negative voltage in the plate circuit that cuts off the chrominance amplifier. The negative voltage is produced by plate current charging  $C_{252}$  during conduction, which then discharges through  $R_{296}$ . This voltage is applied as negative bias to the control grid of the chrominance amplifier to cut it off for monochrome broadcasts.

**24-20. Color Kinescopes.** A color kinescope consists of a vacuumed glass or metal envelope containing an electron gun that produces a narrow

beam of electrons to strike the fluorescent screen on the faceplate, as in conventional black-and-white kinescopes. However, the screen phosphor in a color kinescope produces red, green, or blue light when energized by the incident electron beam. The screen phosphor generally used for green light is Willemite, containing zinc, silicon dioxide, and manganese; the red phosphor contains zinc phosphate with manganese; the blue phosphor includes zinc sulfide, silver, and manganese oxide. Red screen phosphors are relatively inefficient, producing 30 per cent of the light output of a blue phosphor, while the green light is about 70 per cent compared to blue. The amount of light output from the color kinescope screen is varied by the color video signal applied between control grid and cathode, as in conventional kinescopes.

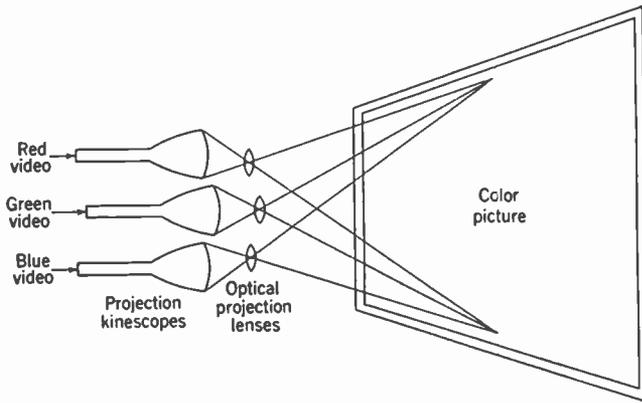


FIG. 24-26. Optical projection arrangement for three color kinescopes.

To produce color mixtures in the reproduced picture, some means must be provided for combining the red, green, and blue light from the individual screen phosphors in such a way that the eye can average the effect of the three primary colors to see the resultant additive mixtures. The basic methods of providing the required color mixtures in the kinescope reproduction are the following:

1. **Optical projection.** Three separate color kinescopes can be used, with their individual red, green, and blue color images projected onto a common viewing screen, in an arrangement commonly called the *trineoscope*. This method uses relatively simple color kinescopes, but the optical registration must be precise and the light efficiency is low, as in any optical projection system. Projection lenses can be used, as illustrated in Fig. 24-26, or spherical mirrors for greater light efficiency.

2. **Dichroic mirrors.** A dichroic mirror reflects most of the incident light of one color but transmits light of other colors. For instance, with white light incident at an angle of about  $45^\circ$  a red-reflector dichroic

mirror can reflect only the red component of the light, while the green and blue light pass through. The selective characteristic is obtained by coating glass with very thin layers having different indices of refraction. Because of their efficiency and selectivity, two dichroic mirrors are used in the color television camera for splitting light from the scene into its component primary colors for the red, green, and blue camera tubes. Similarly, at the receiver, three separate red, green, and blue kinescope images can be superimposed by dichroic mirrors for direct viewing without optical projection.

3. Three-gun kinescope with three-color screen. This type of kinescope, provides direct view on a color screen equivalent in size to 12- to 21-in. monochrome kinescopes. Three individual electron guns in the tube control the light output from the red, green and blue color phosphors on one screen. As a result, individual control for each color is obtained with one kinescope. However, a three-gun kinescope inherently presents the problem of registration of the three electron beams and their corresponding color phosphors.

4. Single-gun kinescope with three-color screen. In this type of kinescope one electron beam excites the three color phosphors on one screen, using the sequence necessary to produce the required color mixtures. Additional processing of the simultaneous color signals for the one electron gun is necessary, but registration problems can be minimized. In addition, proper color balance is more easily obtained for neutral gray tones when only one electron gun is used instead of three guns having slightly different operating characteristics. Improper color balance results in white and gray values tinged with color. Correct gray scale for all brightness levels is important for a color kinescope in reproducing black-and-white pictures for monochrome broadcasts.

*Three-gun Tricolor Kinescopes.* Figure 24-27 illustrates the construction and operating requirements of a typical three-gun tricolor kinescope. Included are three electron guns to produce individual electron beams accelerated to the color-phosphor screen, through small holes in the shadow mask. The screen consists of small closely-spaced phosphor dots for red, green, and blue, accurately deposited in interlaced positions on the glass supporting plate. Figure 24-28 shows the arrangement of the dot triangles, called *trios*. Note that the blue dot is on top, as determined by the tube mounting. The number of dot trios should be as high as possible for maximum resolution, about 200,000 trios or 600,000 dots being typical for a 15-in. tube. The shadow mask, with the same number of holes as there are dot trios, allows each electron beam to excite only its corresponding phosphor color in the dot trio. For colors requiring more than just one primary, two or three of the dots in each trio are excited by their respective electron beams.

The electron guns are positioned  $120^\circ$  apart within the one glass neck, and mounted parallel to the center axis of the tube. Each gun consists of a heater, cathode, control grid 1, accelerating grid 2, focusing grid 3, grid 4—called the *convergence* electrode, and internally connected to a

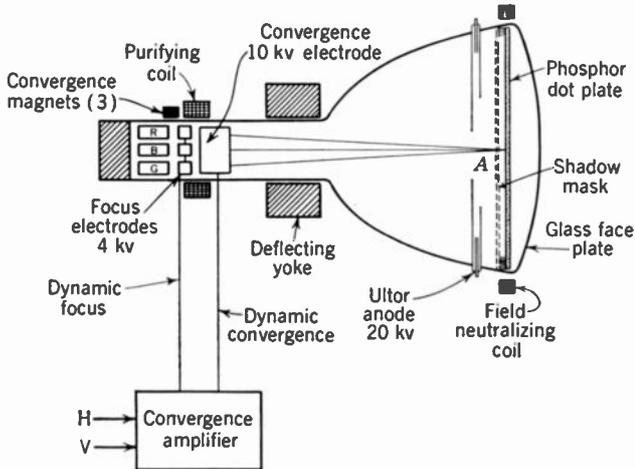


FIG. 24-27. Construction and operating requirements of tricolor kinescope. (RCA.)

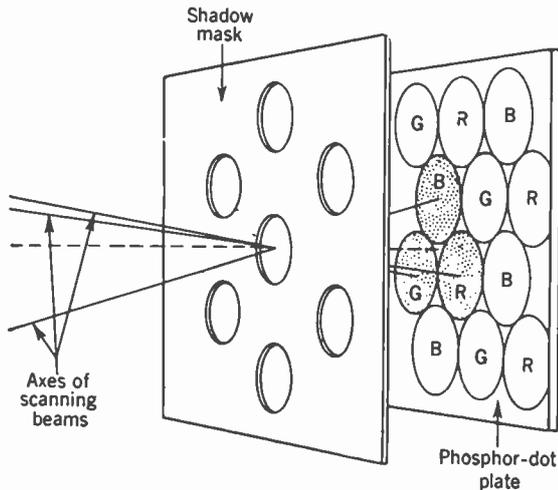


FIG. 24-28. Relationship of shadow mask to phosphor-dot trios. (RCA.)

larger electrode common to all three guns—and the internal wall coating serving as the anode. Convergence refers to the process of making the three separate electron beams meet at the point of one dot trio, for correct registration of the individual colors.

Each gun utilizes electrostatic focusing for its individual electron beam, but the focusing grids are connected internally so that only one source of focusing voltage is necessary. The three beams are made to scan the screen by electromagnetic deflection, using one common deflection yoke mounted externally at the neck of the tube. As the three beams scan the phosphor plate horizontally and vertically in the standard scanning pattern, the dot trios are excited in accordance with the color video signals applied simultaneously to the control-grid-cathode circuit of all three

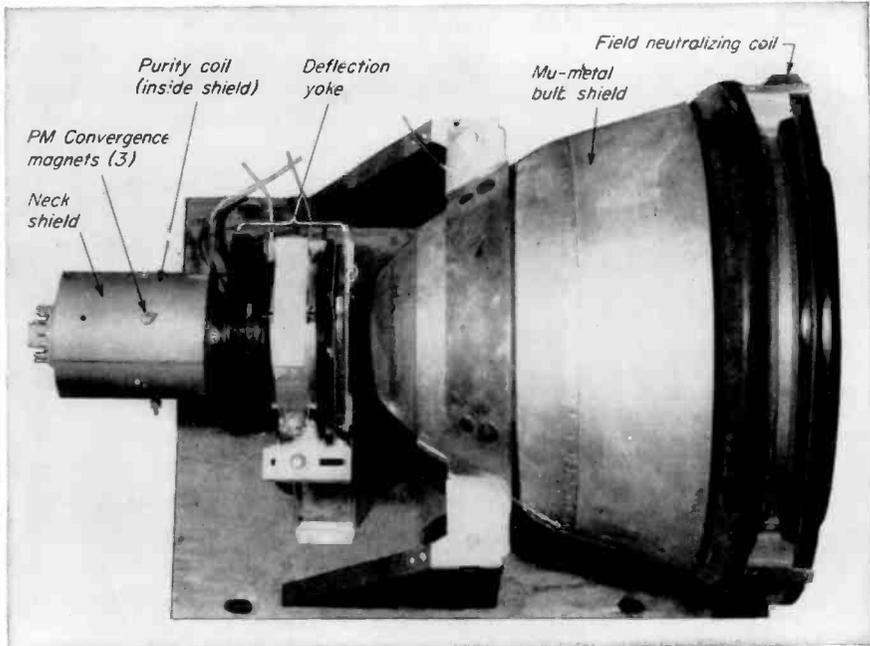


FIG. 24-29. Color kinescope mounted with auxiliary operating units. (CBS-Columbia.)

guns. The color purifying coil produces a uniform transverse magnetic field to align the three electron beams parallel to the center axis of the kinescope. By rotating the coil and varying the amount of direct current, the purity coil is adjusted for a pure primary red color from the dot trios, near the center of the screen, without contamination from the remaining two colors. When the red is pure, blue and green are normally also correct. Usually, there is also a large coil called the *field neutralizing coil* around the bulb at the front to be adjusted for color purity at the edges of the screen. Three adjustable permanent magnets are mounted around the neck of the kinescope for aligning the three beams with respect to each other to obtain proper convergence. In

addition, the d-c voltage applied to the convergence electrode is adjusted for *static* or *d-c convergence*. Correction voltages obtained from the horizontal and vertical deflection circuits must also be applied to the convergence electrode to obtain *dynamic convergence*. Convergence is indicated by white dots reproduced on the screen for each trio. It is also necessary to provide *dynamic focusing*. There is no ion-trap magnet as the screen has an aluminized backing. Figure 24-29 shows the auxiliary units mounted on a tricolor kinescope.

*Single-gun Tricolor Kinescope.* The picture tube shown in Fig. 24-30 has just one electron gun but produces three colors by exciting strips of red, green, and blue phosphors on the kinescope screen.

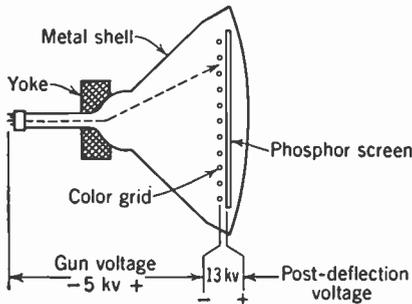


FIG. 24-30. Single-gun, line-screen color kinescope. The *chromatron* is this type.

The electron beam accelerated toward the phosphor screen must pass through the wire mesh that functions as the color grid. Because the strips are arranged in groups of three colors, the voltage applied to the color grid determines which color strip is excited by the electron beam. Figure 24-31 illustrates how color switching voltage applied to the color grid controls the color

produced by the phosphor strips. Zero voltage on the color grid results in no displacement of the beam, which then strikes the middle of the phosphor-strip trios. Positive voltage applied to the color grid allows the beam to excite a different color in the strip trios, and negative voltage displaces the beam in the opposite direction to excite the third

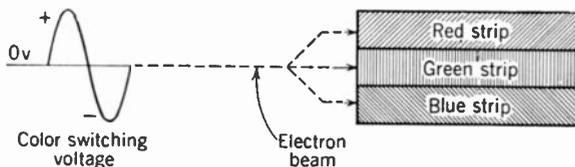


FIG. 24-31. Arrangement of phosphor-strip trios on line-screen kinescope, with switching voltage applied to color grid.

color. Notice that for one sine-wave cycle of switching voltage, the middle phosphor strip is excited by the electron beam twice as often as either of the other two strips. The line-screen tube is used with the phosphor strips horizontal, approximately parallel to the horizontal scanning lines of the raster. The phosphor-strip trios can be arranged with green at the center, as in Fig. 24-31, or with red at the center in order

to make the most inefficient phosphor have the double excitation. There are about 450 strips for the middle color and one-half that number for each of the other two colors, in a 21-in. rectangular tube.

It is important to note that with a single-gun kinescope, the color video signal for the control grid must be gated at the same rate as the electron beam is switched to the different phosphor colors. When the electron beam is on the green phosphor, for instance, the grid signal voltage must correspond to green picture information only. The transmitted color signal, however, includes the information for all three primary colors simultaneously. Therefore, gating or keying circuits are necessary to convert the simultaneous color video signal into the required sequence of individual primary color video signals for the kinescope grid, as illustrated in Fig. 24-32. The frequency used for switching the color grid

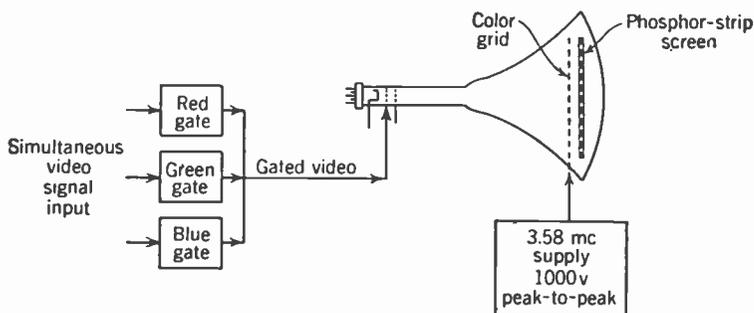


FIG. 24-32. Requirements of gated video signal for control grid and switching voltage for color grid of single-gun, line-screen kinescope.

and gating the video signal is the color subcarrier frequency of 3.58 Mc, since it is available in the color receiver. This frequency has the dual advantages of being high enough to reduce the size of the dot structure resulting from the sequential operation and of providing the benefits of dot interlace because it is an odd multiple of one-half the horizontal line frequency, which makes the dot structure less noticeable in the picture. The frequency for gating the color video signal corresponding to the center phosphor strip must be twice 3.58 Mc.

**24-21. Typical Color Television Receiver.** Essentially, the color television receiver consists of the same circuits as a monochrome receiver for the picture and sound, with the addition of the chrominance video section for the color information and auxiliary circuits for the color kinescope. For best results on u-h-f as well as v-h-f channels, the receiver generally uses 45.75 and 41.25 Mc for the i-f picture and sound carrier frequencies, respectively, and the 4.5 Mc intercarrier sound. To illustrate how the circuits fit together, Fig. 24-33 shows in block-diagram



form the stages in a typical color receiver utilizing the *I* and *Q* chrominance signals with a three-gun tricolor kinescope.

*R-F and I-F Sections.* Starting at the left in the figure, the r-f tuner selects the antenna signal for any u-h-f or v-h-f channel, converting the r-f signal frequencies of the picture and sound signals to the i-f signal frequencies of the receiver. The r-f signal circuits, including the antenna and transmission line, are the same as in monochrome receivers—except that here uniform response for each selected channel is more important to prevent attenuating the 3.58-Mc color subcarrier modulation on the r-f picture carrier signal. The station selector and fine tuning are provided as front-panel controls, as in monochrome receivers. However, adjustment of the fine tuning control is more important because a slight mistuning of the r-f local oscillator results in no i-f signal for the 3.58-Mc chrominance subcarrier, which has a corresponding intermediate frequency of 42.17 Mc at the edge of the i-f pass band near the i-f sound carrier frequency. This is illustrated by the i-f response curve in Fig. 24-34a. A crystal-diode detector produces the desired video signal output corresponding to the envelope of the amplitude-modulated i-f picture carrier signal input.

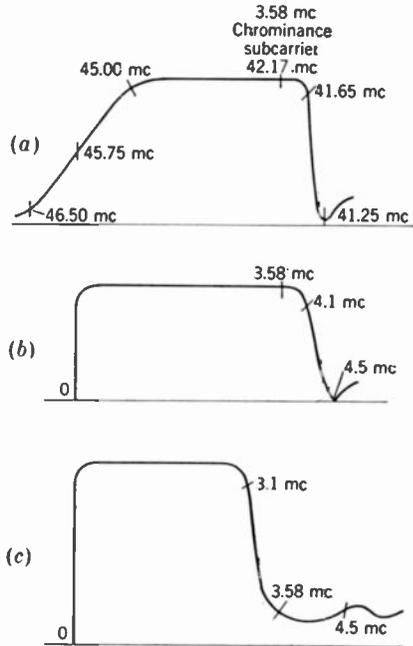


FIG. 24-34. Response curves for receiver illustrated in Fig. 24-33: (a) i-f amplifier, (b) video preamplifier, and (c) *Y* video amplifier.

*Sound Section.* Notice that a crystal-diode detector is also used in a separate sound-converter stage to produce the 4.5-Mc intercarrier-sound signal. Both the picture and sound i-f carrier signals are coupled into the sound converter to produce the 4.5-Mc intercarrier beat. Then the 4.5-Mc signal for the associated FM sound signal is detected by a ratio detector to produce the required audio output for the loudspeaker. A separate sound converter is used, instead of taking the 4.5-Mc beat from the video detector, in order to minimize the amplitude of the 0.9-Mc beat between the 3.6-Mc chrominance subcarrier frequency and the 4.5-Mc sound carrier frequency in the video circuits.

*Video Preamplifier.* The video detector output amplified in the video

preamplifier, or first video amplifier stage, is the total composite video signal containing the luminance signal variations, the blanking and deflection sync pulses for horizontal and vertical scanning, and the 3.58-Mc chrominance information including the burst for color sync. The frequency response required in the video preamplifier to include the 3.58-Mc chrominance signal is shown in Fig. 24-34b. Since the composite video signal from the video preamplifier includes all the required information, it is coupled to the following circuits:

1. Keyed a-g-c amplifier, which produces a-g-c bias for the r-f and i-f amplifiers as in monochrome receivers.
2. Sync separator and amplifier section which provide horizontal and vertical synchronizing signals for the horizontal and vertical deflection circuits, as in monochrome receivers, with automatic frequency control for the horizontal deflection oscillator. The horizontal and vertical output circuits supply saw-tooth scanning current for the deflection yoke on the neck of the color kinescope to produce the scanning raster.
3. Luminance video amplifier which supplies the  $Y$  signal required for addition with the color-difference voltages to produce the red, green, and blue video signals.
4. Chrominance amplifier, which supplies the 3.58-Mc color subcarrier signal for the chrominance section of the receiver.
5. Burst separator and amplifier, which supplies the 3.58-Mc reference burst for the color synchronizing circuits.

*Chrominance Section.* The input to the chrominance section includes the two-phase modulated 3.58-Mc color subcarrier from the video preamplifier; the output consists of the desired red, green, and blue video signals for the tricolor kinescope. First the input signal is amplified in the chrominance band-pass amplifier. A typical chrominance amplifier circuit can be seen by referring back to Fig. 24-18. The 3.58-Mc modulated chrominance signal from the band-pass amplifier is then coupled to the two synchronous detectors for the  $I$  and  $Q$  signals. Each synchronous detector has two input voltages: chrominance signal from the band-pass amplifier and c-w output from the local 3.58-Mc carrier oscillator. The c-w oscillator output for the  $Q$  demodulator is shifted  $90^\circ$  in phase by the quadrature amplifier. As a result, the  $Q$  demodulator's output is the  $Q$  color-difference voltage, while the  $I$  demodulator supplies  $I$  color-difference voltage.

Figure 24-35 shows the demodulator and amplifier circuits used in this receiver for the  $I$  and  $Q$  video signals. Compensation for the vestigial-side-band transmission of the  $I$  signal is accomplished in the  $I$  amplifier circuits by increasing the gain for the 0.5- to 1.5-Mc  $I$  video frequencies transmitted with single side bands, compared to the 0- to 0.5-Mc  $I$  video frequencies transmitted with double side bands. After sufficient ampli-

fication and phase inversion, both plus and minus polarities of the *I* and *Q* voltages are coupled into the matrix circuit, to be combined with the *Y* signal from the luminance amplifier. The *Y* video amplifier's frequency response is usually limited to approximately 3 Mc, as shown in Fig. 24-34c. This minimizes the 3.58-Mc dot pattern in the picture and avoids cross-talk between the luminance and chrominance signals produced by non-linear operation in the video output stage, which would cause incorrect color values in the matrixing.

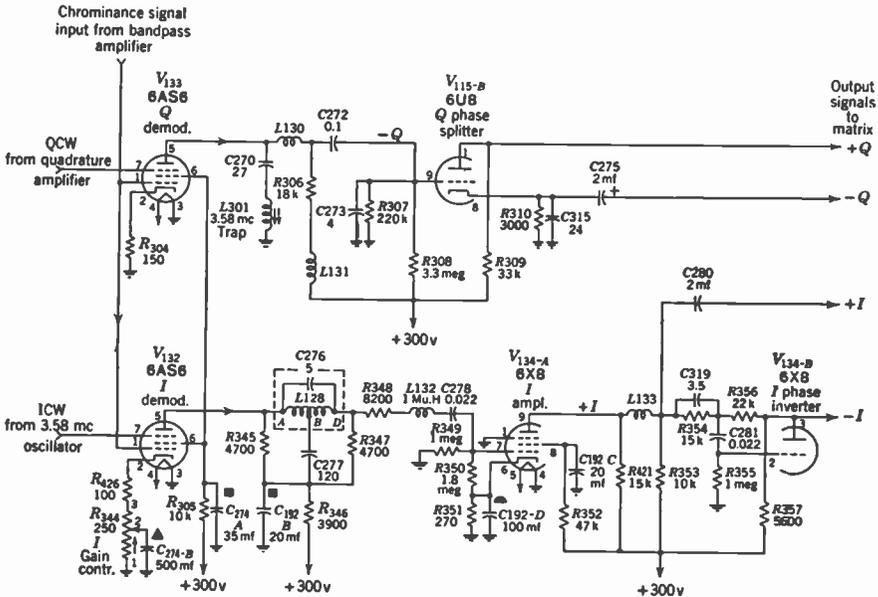


FIG. 24-35. *I* and *Q* video signal circuits in receiver illustrated in Fig. 24-33. (RCA.)

**Matrix Circuit.** Figure 24-36 shows the resistive voltage-divider circuits for the passive matrix network in the input to the color adder stages. Notice that the *I* and *Q* signals have the polarities corresponding to the colors needed to form each primary. For instance, the green adder stage has  $-Q$  input, which is yellow-green, and  $-I$  signal, which is the blue-green cyan. With proper proportions of *I*, *Q*, and *Y* voltage, each adder stage produces in the output circuit its respective primary color video signal. The three color output stages then amplify the primary video signals enough to drive the control grids of the three-gun tricolor kinescope. The three d-c restorers provide automatic background control for each of the primary colors. The primary color adder and output stages have a 3-Mc bandwidth so that the *Y* video signal can reproduce the fine detail of the picture in black and white.

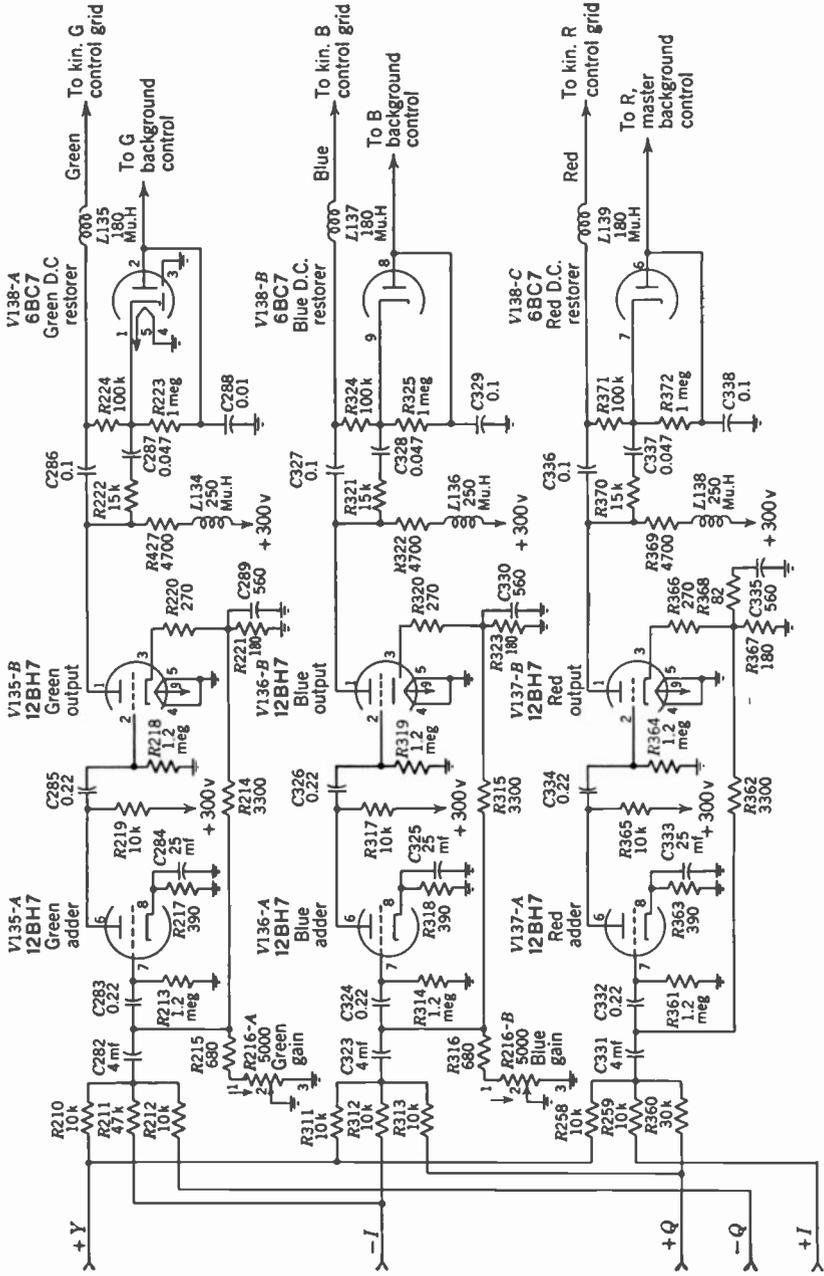


FIG. 24-36. Matrix section in receiver illustrated in Fig. 24-33. (RCA.)

*Color Sync Section.* For color synchronization, video signal input containing the 3.58-Mc color burst is taken from the video preamplifier and coupled to the burst amplifier. As shown by the typical circuit in Fig. 24-23, the burst amplifier is tuned to 3.58 Mc and gated by horizontal deflection voltage to produce output only for the 3.58-Mc color sync burst following each horizontal sync pulse. The color sync burst has the reference phase required for synchronizing the 3.58-Mc local subcarrier oscillator. With the 3.58-Mc color sync burst voltage and carrier voltage from the 3.58-Mc local oscillator coupled into the phase detector, it controls the phase of the c-w oscillator, through the reactance tube, to establish the correct hues in the picture. The schematic diagram of this a-p-c circuit is in Fig. 24-24. The phase detector is also used to supply negative d-c bias voltage to cut off the color killer stage when there is no color sync burst and, therefore, no color transmission.

*Kinescope Convergence, Focus, and Anode Voltage.* The color receiver uses reaction scanning for horizontal deflection, with the horizontal output circuit supplying flyback high voltage for the kinescope anode. However, the tricolor kinescope has additional high-voltage requirements. A voltage doubler supplies 20 kv for the anode of the tricolor kinescope, with a load current of about 750  $\mu$  for the highlights in the picture, equal to the total beam current for the three guns. A voltage regulator is necessary to stabilize the anode voltage for different brightness levels, as the amount of high voltage affects convergence. In addition, the voltage divider in the high voltage supply also provides about 10 kv for the convergence electrode in the kinescope. Proper convergence also requires an a-c correction voltage provided by the convergence amplifier stage, which varies the instantaneous value of convergence voltage to converge the three beams for any deflection angle. For this *dynamic convergence*, the amplifier combines 60-cps voltage from the cathode of the vertical output stage with 15,750-cps voltage from the horizontal deflection amplifier's cathode circuit. A separate focus voltage rectifier obtains high voltage a-c input from the horizontal output transformer to produce about 4-kv d-c voltage output for electrostatic focusing of the kinescope.

*Low-voltage Supply.* In the low-voltage power supply, two selenium rectifiers are used in a voltage doubler circuit to produce about 400 volts  $B^+$ , with a load current of about 400 ma. The low-voltage supply provides direct current through the horizontal and vertical scanning coils in the yoke, which can be adjusted by the horizontal and vertical centering controls for exact positioning of the scanning raster. Direct current is also supplied for the color purity coil and field neutralizing coil. The accelerating grid of each electron gun in the tricolor kinescope uses +400 volts, approximately, from the low voltage, which can be adjusted for

each electron gun. The amount of 60 cps a-c input power to the receiver is about 500 watts.

*Color Receiver Schematic Diagram.* Figure 24-37 shows the complete schematic diagram of a color television receiver, for reference purposes. This is essentially the same circuit as the receiver illustrated in block-diagram form in Fig. 24-33, both using a wide-band matrix for  $I$  and  $Q$  signals, with a three-gun tricolor kinescope. Note, though, that in Fig. 24-37 the hue control varies the phase of the burst signal obtained from the video preamplifier, instead of varying the 3.58-Mc oscillator phase.

**24-22. Color Controls.** Table 24-2 lists the additional controls for color in a television receiver, in order to summarize their functions in producing a good color picture. Although the usual controls for a monochrome receiver, such as station selector, fine tuning, volume, contrast, and deflection adjustments are omitted in this table, it should be noted that correct setting of all the controls is important. This is especially true of the fine tuning control, which must be set exactly to provide the chrominance signal for the picture i-f amplifier. Otherwise, there is no color. The fine tuning control should be set for color in the picture, with the saturation control turned up and minimum 0.9 Mc beat, which is evident as a coarse herringbone pattern in the picture. Of the color controls, usually the hue or phasing control and the saturation control are on the front panel of the receiver as operating controls. When switching between stations broadcasting in color, readjustment of the fine tuning, hue, and saturation controls will usually be necessary. The remainder of the color controls are on the chassis, either at the rear or available from the front through a hinged cover, or at the side of the cabinet.

Correct adjustment of the color controls is also necessary for a good black-and-white picture when a monochrome broadcast is received. As indicated in Table 24-2, the following adjustments are set with a monochrome picture for good black-and-white reproduction without color: screen-grid and d-c bias voltages for the red, green, and blue guns in the kinescope; green and blue gain controls. Convergence is checked with a *dot generator*, which is similar to a bar generator or crosshatch generator used for checking scanning linearity. Its purpose is to supply video signal that produces black and white rectangular dots on the kinescope screen. About ten rows of horizontal and vertical dots are produced, so that each dot is much larger than the phosphor-dot trios. For checking colors in the reproduced picture, a *color-bar generator* can be used. A typical generator produces video signal for ten vertical color bars differing in hue from orange to green by  $30^\circ$  spacing in phase angles, corresponding approximately to the  $R - Y$ ,  $B - Y$ ,  $I$ , and  $Q$  axes and intermediate colors.

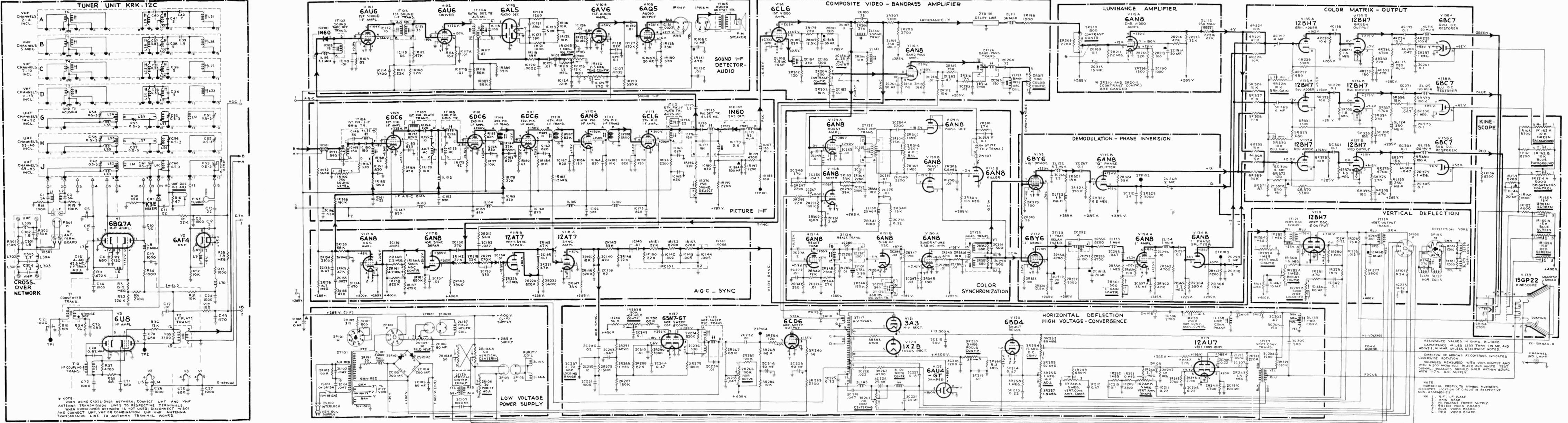


FIG. 24-37. Complete schematic diagram of color television receiver, using wide-band matrix for I and Q signals, with three-gun tricolor kinescope. (RCA CT-100 chassis.)

TABLE 24-2. COLOR CONTROLS

Control	Circuit	Function	Adjustment
Saturation or chroma	Chrominance band-pass amplifier	Vary level of 3.58-Mc chrominance signal	Vividness of colors; no color at zero setting
Hue or phasing	3.58-Mc oscillator a-p-c or sync burst circuit	Vary phase between 3.58-Mc c-w oscillator and sync burst	Correct hues, as determined by flesh tones or hue of known object
A-F-C balance	Color phase detector in a-p-c circuit	Balance diode circuits of phase detector	Zero d-c output at balance
Deflection yoke on kinescope neck	Horizontal and vertical deflection circuits	Produce scanning raster	Rotation tilts raster; movement along tube neck affects purity
Purity coil on kinescope neck	Low-voltage power supply	Align three electron beams	Pure red at center area of raster
Purity adj.	Low-voltage power supply	Vary direct current in purity coil	Pure red at center area of raster
Field-neutralizing coil around kinescope faceplate	Low-voltage power supply	Corrects beam misalignment caused by external fields	Pure red at edge area of raster
Field-neut. adj.	Low-voltage power supply	Vary direct current in field-neutralizing coil	Pure red at edge area of raster
Convergence magnet(s)	Permanent magnets	Align each electron beam with respect to other two for proper convergence	Vary d-c convergence voltage and position magnets to make dot trios produce white, in center area of raster
D-C convergence	High-voltage supply	Adjust d-c voltage applied to convergence electrode in kinescope	
Horizontal dynamic convergence	Horizontal output circuit	Vary d-c convergence and focus voltages at horizontal line rate	Convergence at left and right sides of raster
Vertical dynamic convergence	Vertical output circuit	Vary d-c convergence and focus voltages at vertical field rate	Convergence at top and bottom of raster
Focus	Focus voltage rectifier in high-voltage supply	Vary focusing grid voltage for kinescope	Sharp focus in raster and picture. Affects convergence also

TABLE 24-2. COLOR CONTROLS (Continued)

Control	Circuit	Function	Adjustment
High-voltage	Shunt regulator in high-voltage supply	Stabilize kinescope anode voltage	Constant high voltage for all brightness levels. Affects convergence also
Red-screen, green-screen, and blue-screen	Low-voltage power supply	Vary kinescope screen-grid voltage for red, green, and blue electron guns	Low-level white raster, without picture
Green-gain and blue-gain	Green adder and blue adder stages	Provide enough green and blue video signal, compared to red	High-contrast black-and-white picture without color
Red (master), green, and blue background	Kinescope grid-cathode circuit	Vary d-c bias voltage for red, green, and blue electron guns	Black-and-white picture without color tracking from low brightness to high brightness
I Gain	I signal amplifier and phase splitter	Provide enough I signal compared to Q signal	Correct hues in color bars, primarily red and yellow

**24-23. Color Troubles.** Considering only symptoms that affect the color, while the sound, raster, and monochrome picture are normal, color troubles can be considered in the following categories: no color, no color hold, incorrect colors, or undesired color interference. These color troubles can usually be localized to either the chrominance section of the receiver or the color kinescope.

**No Color.** This trouble symptom can be established definitely by checking on different channels to see that a normal monochrome picture is obtained on all stations but there is no color for a program known to be televised in color. Before assuming the trouble is in the color circuits, the setting of the fine tuning control should be checked, as misadjustment can result in no chrominance signal for the picture i-f amplifier. The fine tuning should be adjusted by varying the control until sound bars are visible and then adjusting for minimum 920-ke beat interference. Also, the adjustment of the saturation control should be checked, as the zero setting results in no color. In the chrominance section of the receiver, no color can be caused by the following conditions: (a) Inoperative chrominance band-pass amplifier, resulting in no signal for the synchronous detectors. Excessive bias from the color killer stage can cut off the chrominance amplifier. (b) Inoperative 3.58-Mc subcarrier oscillator, resulting in no detection of the chrominance signal, or insufficient oscil-

lator output. Notice that these troubles cut off all the color signal voltages, resulting in no color. A trouble in only one color causes incorrect hues.

*No Color Hold.* No color sync results in changing colors or in horizontal color bands that drift vertically. Since the lack of color sync means that the 3.58-Mc subcarrier oscillator's phase is continuously changing, this causes variations of hue corresponding to the color bands. The further the crystal-controlled 3.58-Mc subcarrier oscillator is from the correct phase, the greater the number of color bands. The loss of color hold can be caused by an inoperative burst amplifier, resulting in no color sync voltage, or a defect in the automatic-phase-control circuit. If the crystal allows the oscillator to lock in at a frequency differing from the subcarrier by multiples of 15,750 cps, stationary vertical color bars appear. This trouble of incorrect oscillator frequency is caused by a defective crystal.

*Incorrect Colors.* Troubles causing incorrect color values can be localized to either the tricolor kinescope or the chrominance circuits by noting the appearance of the raster alone and a monochrome picture. When there is color in the raster, with the contrast control at minimum, the trouble is either in the color purity or the red, green, and blue screen-grid adjustments, assuming the use of a three-gun tricolor kinescope. Non-uniform coloring in the raster results from incorrect purity. Uniform red, green, or blue in the raster indicates too much screen-grid voltage for that color. When the raster is white but a monochrome picture has color with the contrast control turned up, then the trouble is either in the convergence and focus adjustments or the chrominance and video signal circuits. Color fringing at the edges of objects in the picture indicates incorrect convergence or focus. Color throughout the entire monochrome picture is caused by lack of balance in the a-c signal drive for the three electron guns.

When the monochrome picture is reproduced correctly in black and white but color pictures have the wrong hues—assuming the hue control is set properly for the correct reference phase—this indicates (a) no quadrature phase between the ICW and QCW oscillator voltages for the demodulators or (b) incorrect relative gain for the color signal voltages. The quadrature amplifier for the QCW signal must be aligned exactly at 3.58 Mc to produce the required 90° phase. Incorrect relative gain for the color-difference voltages in the synchronous detectors or in the color adder stages can be caused by a circuit defect or misadjustment of the color gain controls. If there is no *I* signal at all because of an inoperative *I* demodulator, the picture is reproduced without orange and cyan color values. This effect can be seen in Plate IV. Without the *Q* signal, as shown in Plate III, purple and yellow-green are missing from the picture

**Color Hum Bars.** Hum voltage in the chrominance video circuits produces horizontal pairs of hum bars in the picture, as in monochrome receivers, but the bars have the colors associated with the stage where the hum is injected. For instance, 60-cycle hum voltage caused by cathode-heater leakage in the  $Q$  demodulator will produce one pair of hum bars with the  $Q$  colors. As shown in Plate II, one-half cycle of the hum voltage produces magenta-to-purple colors and the opposite polarity in the next half-cycle produces yellow-green.

**Color Snow.** When the chrominance band-pass amplifier is not cut off for a monochrome picture, the stage amplifies luminance signal-voltage variations of approximately 2 to 4 Mc, which are detected in the color demodulators, resulting in random speckles of color in the picture. Because of the relatively coarse grain, the color snow is also called *confetti*.

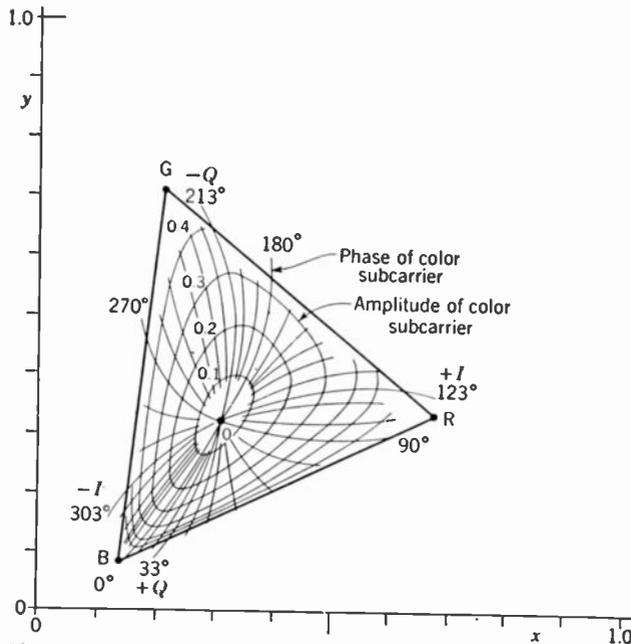


FIG. 24-38. Chromaticity diagram showing color subcarrier phase and amplitude as a function of hue and saturation. Refer to color plate VIII for actual hue and saturation values. (From "Colorimetry in Color Television," Part III, by F. J. Bingley, Proceedings of the IRE, Second Color Television Issue, January, 1954.)

**24-24. CIE<sup>1</sup> Chromaticity Diagram.** Although different systems of colorimetry can be used to specify color values numerically, the internationally standardized chromaticity diagram illustrated in Fig. 24-38 is used to indicate the color specifications in color television. To read the

<sup>1</sup> Abbreviation for the Commission Internationale de L'Eclairage (CIE), also called the International Commission on Illumination (ICI).

chromaticity diagram, consider the horizontal  $x$  axis and vertical  $y$  axis as the rectangular coordinates of a graph. The points marked  $R$ ,  $G$ , and  $B$  on the triangle indicate the primaries specified for color television. For red,  $x = 0.67$ ,  $y = 0.33$ ; for green,  $x = 0.21$ ,  $y = 0.71$ ; for blue,  $x = 0.14$ ,  $y = 0.8$ . The white at the center is the reference white illuminant  $C$ , with coordinates  $x = 0.31$ ,  $y = 0.32$ .

Actual chrominance values are reproduced in color Plate VIII. This horseshoe-shaped curve is called the *spectrum locus* because it represents all hues produced by splitting white light into its component *spectral colors*. The hue is indicated by the numerical values on the curve, which give the *dominant wavelength* in millimicron units ( $10^{-9}$  cm). These points are saturated values of the hues represented by a straight line through the white point at the center to intersect the horseshoe curve. Opposite ends of the straight line intersect the curve at points that are complementary hues. Along any one line of constant hue, values closer to white at the center have less saturation. It should be noted that the straight line closing the bottom end of the horseshoe curve corresponds to the range of *nonspectral* purple colors. These colors do not occur in the spectrum of natural white light but can be produced by additive mixtures of red and blue. Referring to Fig. 24-38, the color values in the  $I$  signal correspond to the line joining the orange point  $x = 0.55$ ,  $y = 0.42$  and the cyan point  $x = 0.16$ ,  $y = 0.26$ , through illuminant  $C$ . Colors on the  $I$  axis, then, are orange hues of varying saturation up to white, or the complementary cyan hues of varying saturation. The axis for the  $Q$  signal joins the yellow-green point  $x = 0.4$ ,  $y = 0.55$ , through white, to the purple point at  $x = 0.24$ ,  $y = 0.12$ . By suitable combinations of  $I$  and  $Q$ , all the chrominance values in the  $RGB$  triangle can be reproduced in the color television system. This is illustrated in Fig. 24-38, which shows how the phase and amplitude of the 3.58-Mc modulated chrominance signal correspond to chromaticity values. The concentric rings around the white point at  $C$  correspond to varying saturations for different amplitudes of the modulated color subcarrier signal, while the curved lines from point  $C$  indicate the phase angle and corresponding hue.

## REVIEW QUESTIONS

1. What is meant by color addition?
2. Name the three primary colors used in color television. What are their complementary colors?
3. What color corresponds to white light minus cyan? White light minus magenta?
4. Why is the picture information of the primary color video voltages converted into a luminance signal and chrominance signal for television broadcasting? What are the two components of the chrominance signal?
5. Give two advantages of using orange and cyan for a two-color picture reproduction.

6. Define hue, saturation, and chrominance.
7. State the percentages of red, green, and blue video voltages in the luminance signal.
8. What is the video frequency bandwidth of the transmitted luminance signal?
9. What parts of the reproduced picture are reproduced in black and white by the luminance signal? What parts are reproduced in full color as mixtures of red, green, and blue? What parts are reproduced as mixtures of orange and cyan?
10. What are the colors corresponding to the  $+Q$ ,  $-Q$ ,  $+I$  and  $-I$  signals?
11. What is the video frequency bandwidth for the transmitted  $Q$  signal? For the  $I$  signal?
12. How is the 3.58-Mc modulated chrominance signal transmitted to the receiver?
13. What is the function of the 3.58-Mc burst transmitted on the back porch of each horizontal sync pulse?
14. Why is the  $Q$  signal made to modulate the 3.58-Mc color subcarrier in quadrature with the  $I$  signal's modulation?
15. How does the 3.58-Mc color subcarrier signal indicate hue and saturation of the color information in the picture?
16. Why is the chrominance signal transmitted with the subcarrier suppressed?
17. Show the components in a resistive matrix circuit for forming the  $Y$  signal.
18. Give two reasons why the color subcarrier frequency is made exactly 3.579545 Mc.
19. Referring to the chrominance vector shown for yellow in the color wheel in Fig. 24-7, what is its phase angle compared to zero for  $B - Y$ ? Compared to the color sync reference burst?
20. Show the waveforms of  $R$ ,  $G$ ,  $B$ ,  $I$ ,  $Q$ , and  $C$  signals corresponding to one horizontal line of information for a pattern consisting of three vertical bars of equal width, which are white, saturated red, and cyan from left to right.
21. Give the specific functions in terms of the color picture reproduction for the following circuits in a color television receiver: (a) r-f local oscillator in the front end, (b) i-f amplifier, (c) video detector, (d) video preamplifier, (e) a-g-c amplifier, (f) chrominance band-pass amplifier, (g) color burst amplifier, (h) color killer, (i) 3.58-Mc c-w oscillator, (j) quadrature c-w amplifier, (k)  $I$  demodulator, (l)  $Y$  video amplifier, (m) matrix section, and (n) red output stage.
22. Describe briefly the operation of the hue and saturation controls.
23. Give one advantage and one disadvantage of a receiver using the  $I$  and  $Q$  color video signals, compared to the  $R - Y$  and  $B - Y$  signals.
24. Referring to the schematic diagram in Fig. 24-24, give the function of  $R_{235}$ ,  $C_{235}$ ,  $T_{122}$ , and  $T_{123}$ . To what frequency is  $T_{122}$  aligned?
25. What is meant by convergence in a tricolor kinescope? Why does focusing affect convergence?
26. What is the function of the color grid in the single-gun, line-screen type of kinescope?
27. What will be the effect in the picture if each of the following stages does not operate: (a) chrominance band-pass amplifier, (b) reactance tube in a-p-c circuit, (c) c-w crystal oscillator, (d) QCW amplifier, (e)  $I$  phase inverter, and (f) red adder.
28. How can the reproduced picture help distinguish between trouble in the 3.58-Mc modulated chrominance signal circuits and trouble in the  $I$  or  $Q$  video signal circuits?
29. What is the effect in the picture when there is no color synchronization in the receiver?
30. Referring to the receiver circuit diagram in Fig. 24-37, list six controls that affect convergence, four controls for a-c color video signal gain, and six controls that affect color balance for a white raster.

# APPENDIX

## TELEVISION BROADCAST CHANNELS WITH SOUND AND PICTURE CARRIER FREQUENCIES

Channel number	Frequency band, Mc	Picture carrier frequency, Mc	Sound carrier frequency, Mc	Channel number	Frequency band, Mc	Picture carrier frequency, Mc	Sound carrier frequency, Mc
2	54-60	55.25	59.75	43	644-650	645.25	649.75
3	60-66	61.25	65.75	44	650-656	651.25	655.75
4	66-72	67.25	71.75	45	656-662	657.25	661.75
5	76-82	77.25	81.75	46	662-668	663.25	667.75
6	82-88	83.25	87.75	47	668-674	669.25	673.75
7	174-180	175.25	179.75	48	674-680	675.25	679.75
8	180-186	181.25	185.75	49	680-686	681.25	685.75
9	186-192	187.25	191.75	50	686-692	687.25	691.75
10	192-198	193.25	197.75	51	692-698	693.25	697.75
11	198-204	199.25	203.75	52	698-704	699.25	703.75
12	204-210	205.25	209.75	53	704-710	705.25	709.75
13	210-216	211.25	215.75	54	710-716	711.25	715.75
14	470-476	471.25	475.75	55	716-722	717.25	721.75
15	476-482	477.25	481.75	56	722-728	723.25	727.75
16	482-488	483.25	487.75	57	728-734	729.25	733.75
17	488-494	489.25	493.75	58	734-740	735.25	739.75
18	494-500	495.25	499.75	59	740-746	741.25	745.75
19	500-506	501.25	505.75	60	746-752	747.25	751.75
20	506-512	507.25	511.75	61	752-758	753.25	757.75
21	512-518	513.25	517.75	62	758-764	759.25	763.75
22	518-524	519.25	523.75	63	764-770	765.25	769.75
23	524-530	525.25	529.75	64	770-776	771.25	775.75
24	530-536	531.25	535.75	65	776-782	777.25	781.75
25	536-542	537.25	541.75	66	782-788	783.25	787.75
26	542-548	543.25	547.75	67	788-794	789.25	793.75
27	548-554	549.25	553.75	68	794-800	795.25	799.75
28	554-560	555.25	559.75	69	800-806	801.25	805.75
29	560-566	561.25	565.75	70	806-812	807.25	811.75
30	566-572	567.25	571.75	71	812-818	813.25	817.75
31	572-578	573.25	577.75	72	818-824	819.25	823.75
32	578-584	579.25	583.75	73	824-830	825.25	829.75
33	584-590	585.25	589.75	74	830-836	831.25	835.75
34	590-596	591.25	595.75	75	836-842	837.25	841.75
35	596-602	597.25	601.75	76	842-848	843.25	847.75
36	602-608	603.25	607.75	77	848-854	849.25	853.75
37	608-614	609.25	613.75	78	854-860	855.25	859.75
38	614-620	615.25	619.75	79	860-866	861.25	865.75
39	620-626	621.25	625.75	80	866-872	867.25	871.75
40	626-632	627.25	631.75	81	872-878	873.25	877.75
41	632-638	633.25	637.75	82	878-884	879.25	883.75
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