

TECHNICAL TELEVISION

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Under the Editorship of Dr. Irving L. Kosow

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EDITOR'S PREFACE

This book represents a novel approach not only to the subject of television techniques but also to the presentation of technical material at the nonengineering level. Each chapter contains a summary which introduces the reader to the material which will be studied and summarizes the various techniques which will be presented in detail. Each of the techniques studied is presented in a simple, straightforward manner. Each circuit is analyzed from the point of view of its design to accomplish a specific purpose; its advantages and disadvantages are weighed, and new circuits and techniques are introduced until finally the reader is presented with the techniques which are presently used in commercial practice. In this process, which proceeds from the simple to the complex, the reader gains an insight into television design considerations from both a qualitative and quantitative point of view.

Intended for the technical institute and junior college level, perhaps the most outstanding and unique feature of this work is its deceptively simple mathematical approach and quantitative treatment of theoretical material hitherto regarded as beyond the scope of the technician. The author gets to the heart of the matter using simple and clear derivations requiring little mathematical knowledge beyond elementary trigonometry and little electrical theory beyond direct- and alternating-current circuits.

The work is unusually thorough in its treatment of the subject of television, including standards, transmitters, antennas, and all of the various techniques employed in a complete receiver. Each chapter follows the block diagram of the receiver so that the reader sees the over-all organization as well as the details. A chapter is devoted to a detailed analysis of complete receivers, summarizing all of the techniques studied in the previous chapters. The work is completely modern in its lucid and extensive treat-

ment of color television, including detailed analyses of modern color receivers.

A further indication of the author's thoroughness and wide experience as a teacher is the unusual number of illustrations, problems, and questions available for the work's use as a text for a specialized course in television.

IRVING KOSOW

PREFACE

This book assumes some familiarity with radio fundamentals as well as some knowledge of algebra. In the very few instances where mathematics above an elementary level has been used to clarify a point, the corresponding calculation can be neglected without detracting from continuity or usefulness.

The basic principles of television and the composition of the complex television signal are studied in Chap. 1. Chapter 2 is devoted to the transmitter and image pickup tubes. Chapter 3 treats the important problem of wave propagation and antenna. Chapters 4 and 5 explain the special wideband techniques used in television. Chapters 6 through 20 contain a thorough analysis of every stage of a receiver. Chapter 21 constitutes an application of the preceding chapters and describes complete television receivers. Chapter 22 is a detailed analysis of the problems associated with color television, with particular emphasis on the NTSC system. Finally, Chap. 23 analyzes a complete color receiver representative of contemporary designs.

The problems and questions at the end of each chapter have been set at the same general level as the book. They are hopefully meant to help clarify thinking, give an idea of practical numerical values, and deepen understanding of the basic principles underlying television techniques.

A. V. J. MARTIN



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TECHNICAL TELEVISION



Chapter 1

INTRODUCTION

The standard used in a television system is all-important, because it determines precisely the design requirements for the transmitter as well as for the receiver. Standards have been developed on a logical basis, starting with some form of scanning for picture analysis and proceeding through various choices or engineering compromises to fill in the details precisely.

Standards vary from country to country. Although they have a number of common characteristics, they differ so much in other respects that in most cases a receiver designed for a given standard will not perform properly on another standard.

The American NTSC standard will be used throughout this book; it is the main subject of this chapter dealing with standards.

1-1. Over-all organization

Reduced to its simpler representation, a complete television system is composed of a transmitter and a receiver. Electromagnetic waves form the link between transmitter and receiver. The planning of this book follows logically from the general design of a television system, with heavy emphasis on the receiver.

Chapter 1 gives the fundamentals of television systems and explains the various considerations leading to the choice of a standard. The standard is all-important for transmitter as well as for receiver design.

Chapter 2 is devoted to the transmitter. Some familiarity with transmitter functions and circuits is necessary for a thorough understanding of the whole system.

Chapter 3 covers the important subject of antennas, especially with regard to the receiving installation.

It is shown in Chap. 1 how a wide band is necessary for television transmission. Wideband amplification is accordingly dealt with in Chaps. 4 and 5.

The block diagram of a typical television receiver is given at the beginning of Chap. 3, and it is shown how a receiver may be divided into separate sections. Each of these sections is now taken up in logical sequence so as to follow the signal through the receiver: Chap. 6, r-f amplification; Chap. 7, frequency changing; Chap. 8, i-f amplification; Chap. 9, detection; Chap. 10, v-f amplification. These five chapters (Chaps. 6–10) take care of the picture receiver proper. Since the signal a receiver provides to modulate the picture tube must contain a d-c component, Chap. 11 is devoted to the d-c component and its related topic, sync separation.

The sound receiver is analyzed completely in one chapter (Chap. 12) because of its relative simplicity and because the reader is assumed to be at least moderately familiar with radio reception. Having covered the video and sync signals, it remains to see how these can be used to produce a picture; this is described in Chap. 13. Modern picture tubes use electromagnetic deflection and this is the subject of Chap. 14.

Two sweep sections are needed for proper operation of a picture tube, each producing a sawtooth waveform of suitable frequency. Sawtooth generators are studied in Chap. 15, followed by the vertical sweep section (Chap. 16) and the horizontal sweep section (Chap. 17).

With picture receiver, sound receiver, sync separation, vertical and horizontal sweep sections, the television receiver is now complete. However, proper voltages and currents have to be fed to the various sections by the power supply (Chap. 18) and the high voltage necessary for picture tube operation imposes problems of its own (Chap. 19).

Also, auxiliary circuits are commonly used to improve performance or facilitate operation; these are studied in Chap. 20.

To sum up, and to illustrate how the various sections are interconnected, Chap. 21 analyzes two representative designs.

Since we are now familiar with monochrome television, color television is considered, with Chap. 22 introducing the NTSC system and surveying the design of the color circuits.

Finally, Chap. 23 analyzes a representative color receiver to clarify ideas and link together the various sections introduced in the preceding chapter.

All this, however, is a few hundred pages away so let us lose no time in starting where we should—that is, at the beginning.

1-2. Picture analysis

In the present state of the art, simultaneous radio transmission of all the elements making up a televised picture is impossible. This difficulty is

eliminated by subdividing the picture into a number of elementary components, or points, whose characteristics are transmitted sequentially by the radio wave. Evidently, the larger the number of points, the better will be the quality of the reproduced picture. Picture definition or resolution corresponds precisely to the number of points transmitted and received.

The process by which all the elementary points of an image are sequentially and logically sorted is the *scanning* of the picture. This is done in the image pickup tube at the transmitter end of the television chain. Very briefly, the principle of an image pickup tube, shown in Fig. 1-1, is as follows: An optical system L projects the picture S onto a photosensitive surface M . This surface is actually a mosaic-like pattern of a large number of microscopic, independent photoelectric cells. Each cell assumes an electric charge of magnitude corresponding to the amount of light it receives, that is, to the brilliance of the corresponding point in the picture. The mosaic is enclosed in an evacuated glass envelope, which also contains an electron gun K .

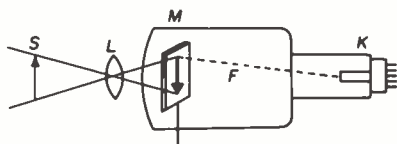


Figure 1-1. Elementary pick-up tube.

The electron gun projects a beam of electrons F on the mosaic. Electric or magnetic fields act upon this beam to move it all over the surface of the mosaic, where it collects the electric charges corresponding to the various brilliances of the picture points.

In this way, an electric current or camera signal is obtained whose intensity reproduces faithfully the luminosity variations of the picture. This video current is then amplified and used to modulate an r-f carrier.

1-3. Scanning

Theoretically, the sequential sorting of all the elementary points could be taken in any order. For example, it is perfectly possible (and, in fact, done in special applications) to start with the point in the center of the picture and proceed outward along a spiral path. However, almost all commercial television systems use a horizontal line scanning technique. The mosaic is scanned by the electron beam exactly in the same fashion as a printed page is read. The beam starts (as shown in Fig. 1-2) at the top left at point A , moves horizontally across to point B , returns quickly to point C a little below starting point A , starts on a new horizontal line, and so on.

The displacement of the beam is graphically represented by the sawtooth curve shown in Fig. 1-3. The ramp portion AB is the actual scanning of the line at constant speed, whereas BC is the quick return or retrace to the left of the picture (Fig. 1-2). However, the sawtooth wave of Fig. 1-3 is insufficient

in itself to move the beam over the whole surface of the mosaic. It can only move it from left to right, and the beam would always retrace the same line.

This difficulty in horizontal scanning is overcome by superimposing another movement of the beam, this time along the vertical axis. The vertical scanning is much slower than the horizontal scanning, so that during the time taken to move from left to right, the beam has only moved a very short distance downward. Its second horizontal line will then be traced just below

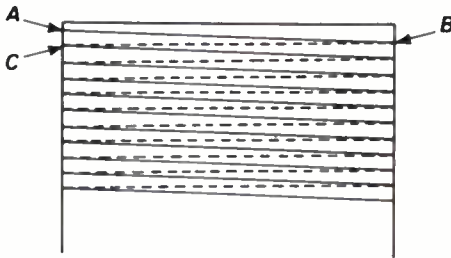


Figure 1-2. Horizontal scanning lines.

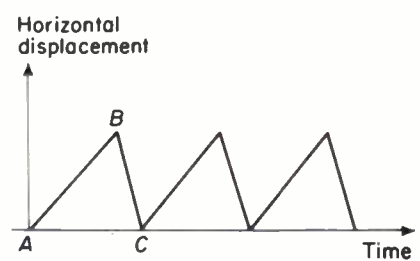


Figure 1-3. Sawtooth displacement of the beam producing horizontal scanning.

the first one, and so on. In this way, the complete scanning is done along a pattern of close, parallel, horizontal lines. In order for these lines to be equally spaced, the vertical scanning must proceed at a constant rate, that is, it must be linear. Furthermore, when the beam has reached the lower right-hand corner of the picture, it must be quickly returned to its starting point, at the upper left-hand corner, so as to start analyzing the picture again.

All this is easily obtained by a vertical displacement represented by the

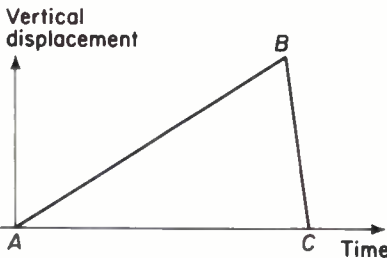


Figure 1-4. Vertical scanning is the result of a sawtooth motion.

sawtooth wave shown in Fig. 1-4. As before, part AB is the active scanning part of the cycle, part BC is the quick return to the top of the picture. Notice that in the graphical representation, the top of the picture (points A and C) corresponds to the lowest part of the diagram, whereas the bottom or lower portion of the picture (point B) corresponds to the highest part of the diagram.

The period of the vertical sawtooth is much longer than the period of the horizontal sawtooth. The ratio of periods determines how many horizontal lines there will be in the scanning.

For example, if the vertical sawtooth period is 33,333 microseconds (μsec) and the horizontal sawtooth period is 63.5 microseconds, there will be $33,333:63.5 = 525$ horizontal lines on the picture. Actually, these lines are not exactly horizontal because of the superposed vertical displacement, but this is of no consequence.

The pattern of horizontal scanning lines constitutes the *raster*. The larger the number of lines, the finer the raster and the better the picture. Hence, the advantage of using a high horizontal sawtooth frequency is obvious.

With regard to the vertical frequency, one of its periods corresponds to the scanning of the whole picture. Because of the time constant or persistency of the human eye, it is unnecessary to scan many pictures per second, a fact relied upon in making motion picture films. The adopted vertical frequency is 30 cps.

1-4. Interlace

The simple scanning described above, however, produces an unpleasant flicker. To eliminate this, interlaced scanning is used. In this system, odd lines are scanned first, followed by even lines. This is easier than it would seem at first

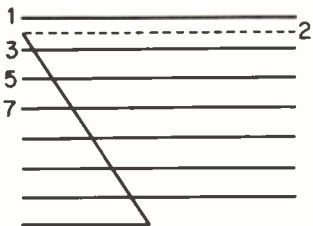


Figure 1-5. Adding a half-line at the end of the picture produces interlace.

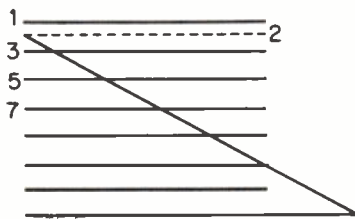


Figure 1-6. Interlace could be obtained by using a longer last line.

sight and is accomplished by making the last scanning line 50 per cent shorter than normal. Consequently, the vertical return will start half a line too soon, and so will the beginning of the first subsequent horizontal line. This line will then fall halfway between two lines of the preceding scan, and this will also hold for all the lines of the scan, thus producing the interlaced raster. This is shown by the simplified diagram of Fig. 1-5. Notice that a line 50 per cent too long, as shown in Fig. 1-6, would produce the same results. Processes other than those shown may also be used to obtain the interlace. In any case, a picture is analyzed in two scanning periods: one for the odd lines, and one for the even lines. Each pattern is called a *field*. One odd and one even field together constitute a complete picture.

If the number of complete pictures is still 30 per second, then the field frequency is 60 cps, and each field contains half the number of lines making up a picture. An exact graphical representation of interlaced scanning is not too easy since it would involve several hundred lines. For the sake of illustration, Fig. 1-7 shows a simplified case where the picture contains only ten lines.

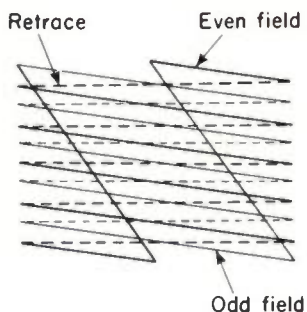


Figure 1-7. Interlace pattern.

1-5. Number of lines

The actual number of lines is a characteristic of the television standard in use. The standard in the United States is 525 lines per picture. This odd number of lines, when divided by 2, gives 262.5 lines per field, so that the extra half-line *automatically* produces the interlace. For proper interlacing there must be exactly 525—not 524 or 526—lines per picture. This is ensured by deriving the line and picture frequencies from a common oscillator. Assume that this oscillator provides the line frequency, that is,

$$525 \text{ lines/picture} \times 30 \text{ pictures/sec} = 15,750 \text{ lines/sec or cps}$$

Synchronized frequency dividers are used, with dividing ratios of 3, 5, 5, and 7. The resulting frequencies are

$$15,750:3 = 5,250 \text{ cps}$$

$$5,250:5 = 1,050$$

$$1,050:5 = 210$$

$$210:7 = 30$$

The last frequency controls the picture repetition rate, so that each picture will contain exactly 525 lines.

Since the total dividing ratio must be odd, no single divider may be even—hence the use of *odd* dividing factors, namely 3, 5, and 7.

The nominal 15,750 cps for line frequency and 30 cps for picture frequency (or 60 cps for field frequency) are standard values in this country, the latter corresponding to the nominal frequency of the power line supplying the transmitter and the receiver.

1-6. Frequency bandwidth

The resolving power of the human eye is limited. It is generally accepted that the eye cannot distinguish between two points if their angular separation is not at least one minute of arc. This is true irrespective of direction, so that there

is little point in providing a picture with greater detail in one direction, say horizontal, than in the other, say vertical.

When the vertical resolution is equal to the horizontal resolution, the picture is said to have *balanced resolutions*.

Now, along any vertical of the picture there are just as many points as scanning lines in the raster, so that the vertical definition is 525 points per height of the picture. The width of the picture being approximately $\frac{4}{5}$ of the height, a balanced definition will be obtained for $525 \times \frac{4}{5} = 700$ points along any horizontal. Consequently the entire picture will be made of $525 \times 700 = 367,500$ elementary points, and since there are 30 pictures per second, there will be $367,500 \times 30 = 11,025,000$ points per second.

The total number of points limits the maximum detail that the picture can contain. The most detailed picture would be one containing points alternately white and black. If such a pattern is translated into an electric signal, this signal will have the same periodicity as the points. With the positive alternation representing a white point and the negative alternation the following black point, one complete cycle of the signal will correspond to two points, so that the signal frequency will be $11,025,000:2 = 5,512,500$ cps. Since the pattern chosen was the maximum detail pattern, this frequency is the maximum signal frequency. It is quite high, and when it is used to modulate a carrier frequency, this carrier should be at least several times higher. Furthermore, the congestion of the frequency spectrum is such that there is simply no space available for television unless metric wavelengths are used. This is why television uses carrier wavelengths in the metric band, that is, from 10 to 1 meter (30 to 300 mc), or even in the decimetric band, that is, from 10 to 1 decimeter (300 to 3000 mc).

At the other extreme of definition, the less detailed pattern occurs when the whole picture has a uniform brilliance. The corresponding electric signal exhibits no variation, and has a frequency zero. In other words, it is a d-c signal. Thus, the complete range or bandwidth of all possible picture detail covers from 0 to some 5.5 mc—a very wide band indeed. If this is compared with a standard audio bandwidth of 5 kc, it can be seen that a single television channel occupies as much frequency space as $5500:5 = 1100$ radio channels.

Also, if both sidebands were transmitted, a television station would occupy $5.5 \times 2 = 11$ mc. To reduce the necessary frequency range, television transmitters suppress a good part of one sideband, using single sideband transmission.

The preceding calculations are only approximate because they neglect a number of secondary factors. Most important is the time taken by both sawtooth waves to return to the starting point. This return or retrace time is not actually zero, and the result is a loss of useful scanning time with corresponding reduction in the definition and hence in bandwidth. The standard allows, in fact, for a maximum modulating frequency of 4.5 mc and not 5.5 mc.

1-7. Reception

Television receivers at present use a cathode-ray tube to reproduce the picture. Briefly, a cathode-ray or picture tube is an evacuated glass envelope containing, as shown in Fig. 1-8, an electron gun K and a fluorescent screen S . The

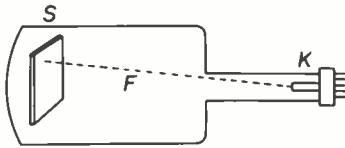


Figure 1-8. Elementary receiving tube.

electron beam F hits the screen, which luminesces at the impact point. The luminous area is the *spot*, and its brilliance increases with the electron beam intensity. By modulating the beam current, one can in fact modulate the spot luminosity and reproduce all shades of brilliance. Electronic or magnetic fields are used to move the electron beam over the surface

of the screen, producing scanning.

A cathode-ray tube can be used to reproduce a picture, but two conditions are necessary:

1. The beam must be modulated in such a way that its current variation reproduces the signal obtained when the picture is analyzed at the transmitter.
2. The scanning of the receiving screen must proceed in an exactly synchronous fashion with the scanning of the pickup tube.

The first condition is met with usual radio techniques. At the transmitter, the video signal coming from the pickup tube modulates a carrier frequency. In the receiver, this carrier frequency is detected and the resulting video signal is applied to the modulating electrode of the picture tube.

The second condition is indirectly satisfied. In the receiver, two sawtooth generators are provided for vertical and horizontal scanings. They must be synchronized with the scanning generators, or time bases, in the transmitter to ensure that

1. the frequencies are identical;
2. the sawtooth starts at the same time in transmitter and receiver.

In order to accomplish the second condition, two special signals, called *synchronization pulses*, are transmitted along with the video intelligence. One is the *vertical synchronization signal*, the other the *horizontal synchronization signal*.

1-8. Synchronization signals

The return of the scanning spot or return trace would be visible on the screen for both horizontal and vertical retraces. To avoid this, the modulation signal is kept to a level corresponding to black during the sawtooth retrace.

The resulting signal is shown in Fig. 1-9 for three consecutive lines. *AB* is the visible part of the line, traced during the displacement from left to right. *BC* is the return from right to left, during which the signal is kept at black level, thus extinguishing the spot and rendering the retrace invisible. The same process is used for the vertical scanning.

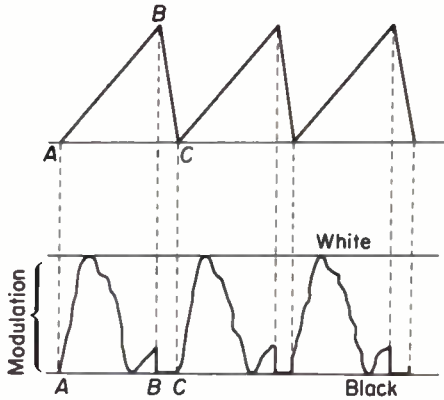


Figure 1-9. Modulation waveform and time relationship with scanning waveform.

It is apparent in Fig. 1-9 that the part *BC* of the scanning cycle is not used for the picture modulation. It is during this unused interval that the synchronizing signal has been inserted. The method customarily adopted is shown in Fig. 1-10. The synchronizing signal is a short rectangular pulse which has been added to the video signal below the black level. This peculiar arrangement has been chosen because in the receiver synchronizing pulses will have to be separated from the video signal and applied to the sawtooth generator. With the method shown, this is very easily done with a simple amplitude limiter adjusted to pass the pulses and clip off the picture modulation.

In the United States standard, the picture modulation occupies 75 per cent of the total video signal, and the remaining 25 per cent is reserved for the synchronizing pulses.

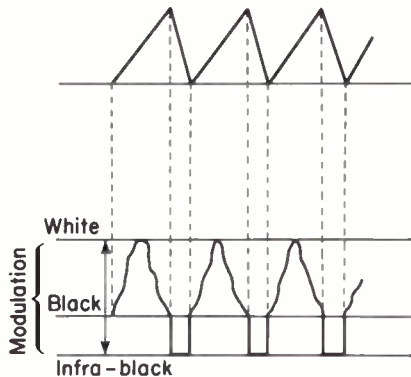


Figure 1-10. A synchronizing signal may be inserted during retrace.

The complete video signal, i.e., modulation plus pulses, is applied to the carrier in negative phase. As shown in Fig. 1-11, the bottom of the pulse corresponds to 100 per cent modulation and a maximum of radiated power. The black level is at 75 per cent modulation level, and the white level is at 0 per cent modulation level.

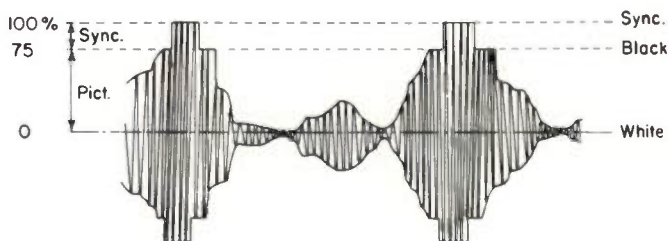


Figure 1-11. Reference levels in standard modulation.

1-9. Horizontal synchronizing signal

The horizontal synchronizing signal occurs during each horizontal return and is a pulse of duration shorter than the retrace time. Figure 1-12 shows the actual durations, as set by the standard, for a line period. The last part of the

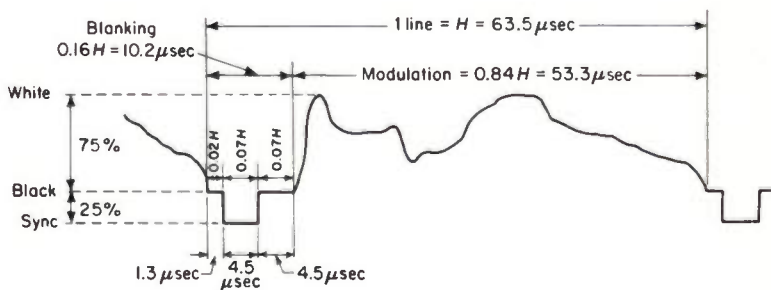


Figure 1-12. Standard durations for a line period.

modulating signal corresponding to the preceding line is also shown. It is followed by a short plateau at black level, called the *front porch*. Then there is the rectangular synchronizing pulse, followed by another plateau at black level called the *back porch*. The modulating signal producing the visible part of the line starts after the back porch.

Since the line frequency is 15,750 cps, the line duration is

$$H = 1 : 15,750 = 0.0000635 \text{ sec or } 63.5 \text{ microseconds}$$

The return takes 16 per cent of this time—that is, 10.2 microseconds. This interval is also called the *horizontal blanking*.

The front porch occupies 2 per cent of H , that is, 1.3 microseconds.

The synchronizing pulse proper has a width of 7 per cent of H , which means a duration of 4.5 microseconds.

The back porch takes 7 per cent of H , or 4.5 microseconds.

The modulation for the visible part of the line occupies the remaining 84 per cent of H , or 53.3 microseconds.

1-10. Vertical synchronizing signal

At the end of each field, of total duration $V = 16,667 \mu s$, a vertical blanking of 15 lines, that is, $15H$ or $950 \mu s$ is provided. During this period, horizontal synchronizing pulses continue to be provided so as to keep the horizontal

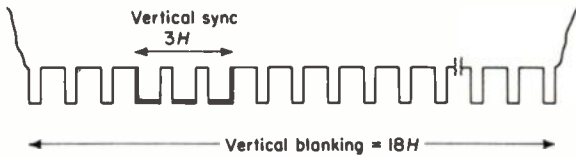


Figure 1-13. Vertical blanking and vertical synchronization signal.

generator in synchronism. Also, a vertical synchronizing pulse has to be inserted in the 25 per cent part of the amplitude reserved for the synchronizing signal. This, then, creates a problem, since the amplitude limiter will effectively separate both vertical and horizontal pulses from the modulation but will not be able to discriminate between them and apply the proper pulse to the proper oscillator. In other words, the vertical pulse somehow must be different from the horizontal pulse so that they can be sorted out and still occupy the same amplitude range.

The adopted method is to make the vertical pulse of much longer duration than the horizontal pulse, and to use a sorting system working on a duration basis. Such a system is actually quite simple and easily provided by RC time-constant circuits.

The vertical synchronizing pulse has a duration of $3H$ or $190 \mu s$. It does not occur exactly at the end of the field, but after an interval of three black lines. Moreover, horizontal synchronizing pulses are provided even during the vertical synchronizing signal, and the interval at the end of a field would look somewhat like the waveshape shown in Fig. 1-13.

This is not the end of the story, because now we must consider interlace. Since each field must contain exactly 262.5 lines, interlace is critically dependent on exact synchronization of the field.

Precisely because of interlace, there is a small but noticeable difference between the two fields. In one, the vertical pulse occurs one line duration H after a horizontal pulse; in the other, the vertical pulse occurs half a line duration $H/2$ after a horizontal pulse.

This small discrepancy is sufficient to disturb the sensitive pulse-sorting circuits, with consequent loss of interlace. But the discrepancy cannot be removed, since it is the very mechanism which produces interlace.

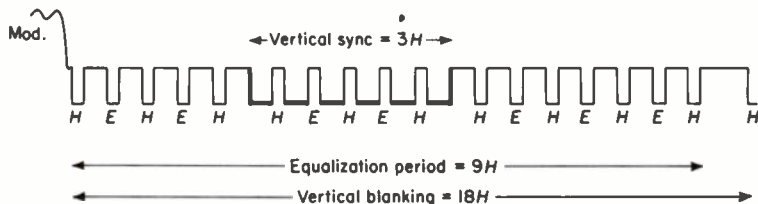


Figure 1-14. Vertical blanking with standard durations.

The problem is solved by “swamping” the irregularity so that it goes unnoticed. Extra horizontal pulses are inserted halfway between the normal line pulses, before, during, and after the vertical pulse. These half-line pulses start three lines before and end three lines after the vertical pulse. They tend to equalize the pulse distribution at the end of the two fields and are accordingly called *equalizing pulses*.

The resulting vertical blanking is shown in Fig. 1-14, where the principal durations have been indicated. The normal horizontal pulses are labeled H and the equalizing pulses E . To keep the drawing to a reasonable length, the end part of the vertical blanking has not been shown. It contains black lines with the normal horizontal pulses. It should be pointed out that some of the figures given are design center values. The standard provides for some margin in many figures.

1-11. D-c component

The complete video signal embodies three well-defined levels; namely, the white level, the black level, and the synchronizing pulses' tip level. This video signal is met in television receivers in either phase, that is, with the pulses on top and the modulation below (negative phase) or vice versa (positive phase). The last phase will be taken as an example, and in this case it is convenient to use the synchronizing pulse tip as a zero reference level, since it is the lower portion of the signal. The normal pattern of the video signal is shown at the top of Fig. 1-15. The pulse tips are aligned at zero. The black level is constant, and so is the white level. If the zero level corresponds to ground level, the video signal is entirely positive, which means that it has a positive

average value or d-c component. This d-c component represents the average brilliance of the picture. If the picture is of uniform shade, the difference in d-c component is the only way to know whether the picture is uniformly white, grey, or black.

Unfortunately, during amplification any coupling capacitor may remove the d-c component—d-c does not pass through capacitors. The video signal then takes the shape shown at the bottom of Fig. 1-15. It distributes itself above and below the zero level in such a way that it encloses equal positive and

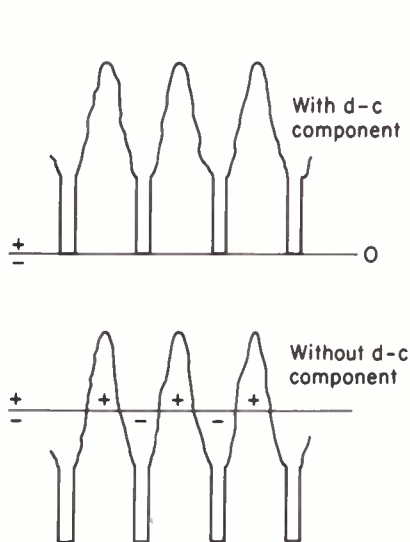


Figure 1-15. Effect of the d-c component on modulation.

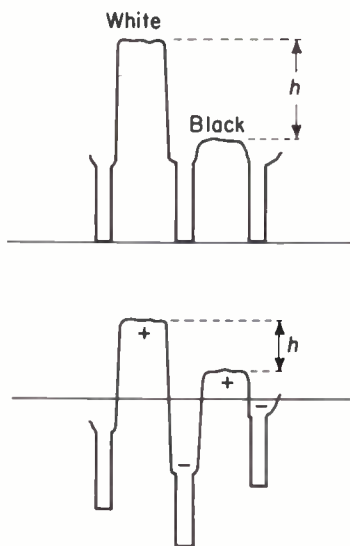


Figure 1-16. Variation of modulation signal with (top) and without (bottom) d-c component.

negative areas as indicated, which is a way of saying it has zero d-c component or zero average value.

That this may have important results for the picture is evidenced in Fig. 1-16. Above, a succession of a white line and a black line is shown when the d-c component is retained. The pulse tips are aligned, and the difference h between black and white levels is correct. Below, the d-c component has been suppressed, say by coupling the video signal through a capacitor. The signal then is distributed above and below the zero line. The pulse tips are no longer aligned and the pulses could not be easily separated by an amplitude limiter. Also, the difference h between black and white levels is incorrect. If the brilliance control is adjusted for the white line to appear white, the black line will appear grey. If the brilliance control is adjusted for the black line to appear black, the white line will appear grey.

The simplest way to retain the d-c component would be by using d-c couplings. However, this is not always possible, and the alternative method is to reinsert the d-c component, when needed, with the help of d-c restoration circuits.

1-12. Sound

The sound accompanying the televised picture is broadcast separately by an independent sound transmitter. The sound carrier is placed just outside the picture sideband, 4.5 mc away from the picture carrier. Whereas amplitude modulation is used for the picture, frequency modulation is employed for the sound, with a maximum frequency deviation of 25 kc and a pre-emphasis of 75 microseconds.

Several systems have been proposed to avoid the necessity of a second transmitter and a separate wavelength for the sound. Essentially, the idea is somehow to introduce the sound modulation in the video signal. For example, the horizontal synchronization pulses could be width-modulated by the sound.

This system has been attempted experimentally and does work. Other schemes have been proposed, but the double-transmitter system is nevertheless in universal use.

1-13. Channel

A complete television channel is shown in Fig. 1-17. It occupies a total bandwidth of 6 mc and includes the picture and sound transmitters. The picture

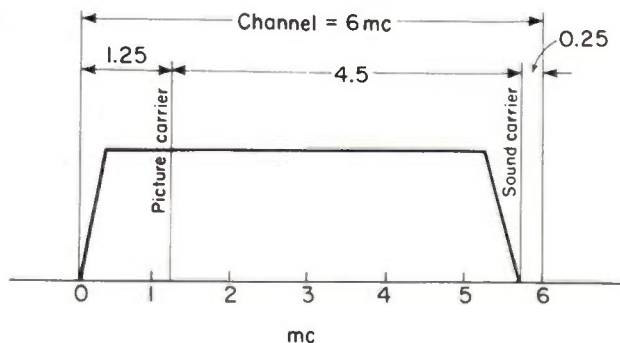


Figure 1-17. Complete television channel.

carrier frequency is set at 1.25 mc from the low end of the channel. The lower picture modulation sideband is progressively attenuated starting at 0.75 mc from the carrier and practically disappears at the edge of the channel. This

scheme is called the *vestigial sideband system*. The idea is that the upper frequency sideband is entirely transmitted and contains all the picture information, hence the possibility of suppressing the lower sideband. Complete suppression is not technically feasible, and the partial suppression adopted results from a compromise between bandwidth economy and ease of design.

The upper sideband contains all video frequencies unattenuated up to 4 mc, with progressive attenuation beyond 4 mc. The upper sideband has practically disappeared 4.5 mc away from the picture carrier, where the sound carrier is located. The sidebands of the FM sound modulation are too small to be apparent on the diagram. Above the sound transmission is an empty band or guardband of 0.25 mc to provide channel separation.

Eighty-two channels have been assigned to television. Twelve, numbered from 2 to 13, are in the very-high-frequency or VHF band. Seventy, from 14 to 83, lie in the ultra-high-frequency or UHF band, which is continuous from 470 to 890 mc.

1-14. VHF channels

The distribution of the VHF channels is given by Table 1-1, where the frequencies are in megacycles (mc)

Table 1-1. VHF Channels

Channel number	Channel range (mc)	Picture carrier frequency (mc)	Sound carrier frequency (mc)
2	54-60	55.25	59.75
3	60-66	61.25	65.75
4	66-72	67.25	71.75
5	76-82	77.25	81.75
6	82-88	83.25	87.75
7	174-180	175.25	179.75
8	180-186	181.25	185.75
9	186-192	187.25	191.75
10	192-198	193.25	197.75
11	198-204	199.25	203.75
12	204-210	205.25	209.75
13	210-216	211.25	215.75

1-15. UHF channels

The distribution of the UHF channels is given by Table 1-2, which indicates only the frequency limits of the channel. The sound carrier is in each case 0.25 mc below the upper limit, and the picture carrier is 1.25 mc above the lower limit.

Table 1-2. UHF Channels

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

1-16. Service area

Radio waves can be vertically or horizontally polarized. In the United States horizontal polarization is used, and the receiving antenna should lie in the horizontal plane.

The power radiated by the transmitters varies between wide limits, depending on the service area to be covered. The sound and picture transmitters are matched in radiated power so that both signals have approximately the same intensity at the receiver location.

A peak field intensity of 5 millivolts per meter is considered as necessary for good reception in cities; 0.5 millivolt per meter is assumed to be satisfactory in rural areas. Under good conditions, field intensities ten times lower still provide good reception.

Roughly, the primary service area of a transmitter, or area of consistent good reception, is limited by the visual horizon, that is, the horizon seen from the transmitting antenna. With the usual antenna installations, this is at most a 50-mile radius. Geographical conditions may strongly modify these

approximate indications, and practical results frequently show large discrepancies with theoretical calculations.

In any case, the preceding rough conclusions should not be taken to mean that no good reception is possible beyond 50 miles. They simply indicate that in all probability, with the exception of especially unfavorable local conditions, good reception is to be expected with a normal receiving installation within the service area of the transmitter.

Problems

1. Scanning the picture could be accomplished by using vertical lines instead of horizontal lines. Compare the two systems, and justify the adoption of the second.

2. Calculate the possible numbers of lines in a television picture using interlace, between 500 and 1000 lines.

3. Evaluate the bandwidth of the French television system using 819 lines and 50 fields per second.

4. Calculate the size of an elementary point on the screen of a picture tube 40 cm wide for a television system using 405 lines and 50 fields per second.

5. Why should the vertical synchronizing pulse be interrupted by line pulses?

6. What is the role of equalizing pulses before, during, and after the vertical synchronizing pulses?

7. Could an interlace of order 3 be used? Discuss the advantages and drawbacks.

8. Explain why pre-emphasis is used for the sound.

9. Propose and discuss a television system in which a single carrier is used for sound and picture modulations.

10. It is proposed to transmit sound and picture modulation on the same carrier, using amplitude modulation for one signal and frequency modulation for the other. Is the scheme feasible? What would be the advantages and drawbacks?

11. The visible part of a picture on a screen measures 30 by 40 cm. If the brilliance is advanced, the blanked-off parts around the visible part also become visible. Calculate their dimensions.

12. In the United States standards, frequency modulation is used for sound and amplitude modulation for picture. Compare this system with the opposite arrangement and discuss practical aspects.

13. A number of foreign television standards use amplitude modulation for sound. Discuss the advantages and drawbacks of using frequency modulation.

14. Propose two systems to obtain interlace other than the odd-number-of-lines system, and explain their operation.

15. The British television system uses 405 lines and 50 fields per second. Calculate the necessary r-f bandwidth if both sidebands are transmitted.

16. Submit a block diagram for spiral scanning; explain the operation, paying special attention to synchronization needs.

17. Derive a general expression for the number of elementary points in a picture, being given the number of lines per picture N_l , the width-to-height ratio r , the horizontal percentage of blanking u , the field frequency F_v , the vertical percentage of blanking v . Assume equal horizontal and vertical definitions. Apply your expression to the United States standard.

18. How many elementary points are there on a television picture for the following standards: (a) 405 lines, 50 fields per second (British); (b) 525 lines, 60 fields per second (United States); (c) 625 lines, 50 fields per second (Germany); (d) 819 lines, 50 fields per second (France)?

19. A high-definition television system uses 1015 lines and 50 fields per second. Horizontal blanking is 18 per cent, vertical blanking is 10 per cent. Evaluate: (a) number of points per picture; (b) video bandwidth; (c) frequency dividing ratios in the synchronization generator.

20. Assume the resolving power of the eye is 1 minute of arc. A television receiver has a screen 50 cm wide. The minimum viewing distance is defined as that distance at which the eye can just resolve two neighboring picture points. Evaluate the minimum viewing distance for the following standards: (a) 405 lines, 50 fields per second; (b) 625 lines, 60 fields per second; (c) 819 lines, 50 fields per second.

Assume in all cases 18 per cent horizontal blanking and 10 per cent vertical blanking. Vertical and horizontal definitions are supposed to be equal.

21. Discuss the relative merits of positive vs. negative modulation.

Chapter 2

TRANSMITTER

From the visual scene in front of a camera to the r-f signal transmitted, a long chain of operations takes place. These operations are performed by a complex system which is loosely called a *television transmitter*.

Briefly, a television transmitter transforms a pattern of light and shadows into a signal suitable for reception by a television receiver. It also has to take care of the sound portion accompanying the picture.

The most original device in the entire transmitter is undoubtedly the *light-to-signal transducer*, which performs the transformation of light variations into electric signal variations.

2-1. Television station

A television station embodies two separate transmitters, one for the picture and one for the sound. Their carrier frequencies are separated by only 4.5 mc so that both fit in the same channel. In the receiver, the antenna and some stages can be used to amplify both picture and sound, with some resulting economy. With the intercarrier system, practically all stages of the entire picture receiver are also used for sound.

Similarly, at the station a single antenna can be used for both transmitters through a special circuit called a *diplexer*. This has evident economic advantages.

Except for the antenna, however, the sound and picture transmitters are entirely independent; a block diagram of a typical television station might look somewhat as shown in Fig. 2-1. The upper part of the figure contains the picture transmitter, from camera to antenna. The lower part of the figure contains the sound transmitter, from microphone to antenna.

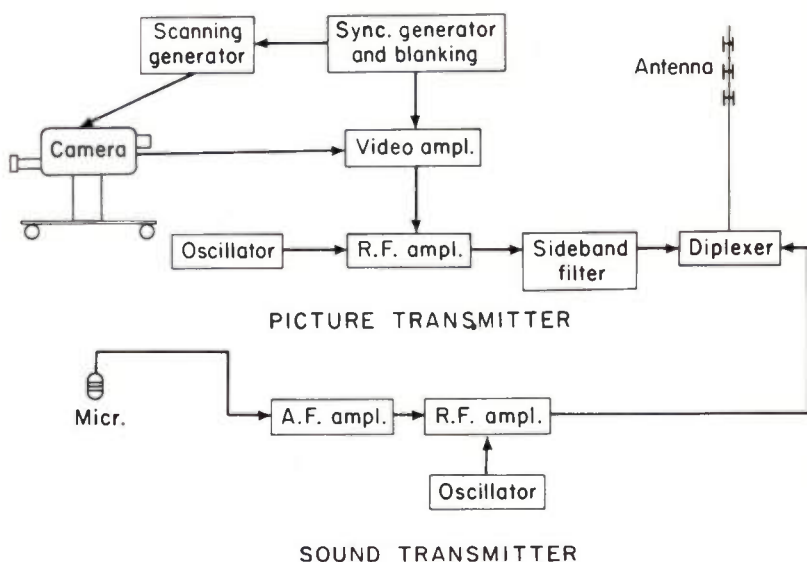


Figure 2-1. Block diagram of a television station.

2-2. Sound transmitter

The sound transmitter embodies a quartz-controlled oscillator, generally working on a rather low frequency, say a few megacycles. Frequency modulation is applied either to the oscillator itself or to a succeeding stage. It generally has a small frequency deviation. Frequency multiplying stages are then used to obtain the desired carrier frequency. At the same time, they multiply the frequency deviation and bring it up to the standard value of 25 kc. Power amplifiers follow, and their output is applied to the radiating antenna through the diplexing circuit.

The general technique of the sound transmitter is almost identical with the technique used for FM radio broadcasting stations. The main difference is that FM radio stations have a frequency deviation of 75 kc—three times larger than that of a sound transmitter.

2-3. Picture transmitter

A special timing generator provides the synchronizing and blanking signals. It controls the scanning stages of the camera, at the output of which appears the picture signal. Synchronization pulses and blanking are added, and the resulting video signal is amplified.

A quartz-controlled oscillator, followed by doublers and/or frequency multipliers, provides the picture carrier, which is amplitude-modulated by the

video signal and applied to power amplifiers. A sideband filter eliminates a good part of the lower sideband, and the remaining signal is applied to the antenna through the diplexer.

To make the best use of the available power, antenna systems have a strong vertical directivity (Fig. 2-2) and concentrate the radiated power near the horizontal.

Except in special cases, the horizontal radiation is omnidirectional. That is, the horizontal directivity diagram is a circle (Fig. 2-3) and the antenna radiates equally in all directions around the station.

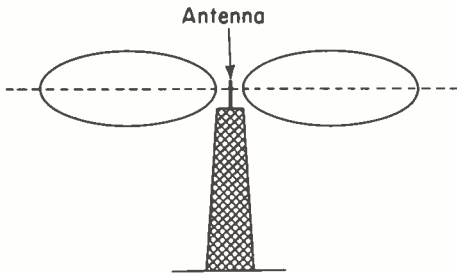


Figure 2-2. The transmitter aerial concentrates the radiated power near the horizontal.

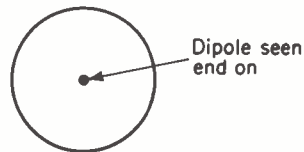


Figure 2-3. When the power is radiated equally in all directions, radiation is omnidirectional.

The vertical concentration of the radiation corresponds to an antenna gain which may exceed ten. This results in an increase in the apparent radiated power. It is taken into account in the effective radiated power, or ERP, which is the transmitter power multiplied by the antenna gain.

2-4. Synchronization generator

The block diagram of a sync generator is shown in Fig. 2-4. A master oscillator delivers a stable frequency equal to twice the horizontal line frequency F_h .

This frequency is divided by 2, and the resulting line frequency is applied to two circuits providing respectively the horizontal sync and the horizontal blanking.

In another chain, the master frequency $2F_h$ is applied to a cascade of frequency dividers, the output of which is the *field frequency* F_v . This vertical frequency drives two circuits providing the vertical sync and blanking. Finally, the horizontal and vertical sync and blanking signals are added together in a mixer stage to provide the complete signal.

The frequency divider chain has a total dividing ratio equal to the number of lines per picture, that is,

$$2F_h/F_v$$

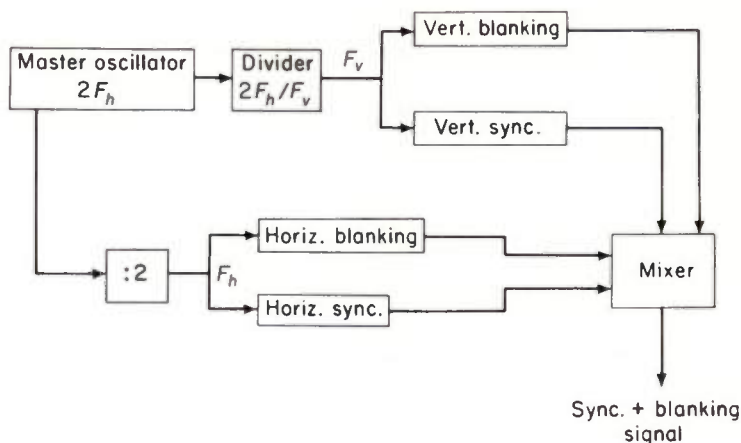


Figure 2-4. Block diagram of a synchronization generator.

A frequency divider stage is securely synchronized if the dividing ratio does not exceed 13. The available ratios are then 3, 5, 7, 9, 11, and 13. Thus, for the United States standard, the master oscillator does not work on 15,750 cps (as previously stated for simplicity) but on twice the line frequency, that is,

$$2 \times 15,750 = 31,500 \text{ cps}$$

The first stage divides by 3:

$$31,500 : 3 = 10,500 \text{ cps}$$

The second stage divides by 5:

$$10,500 : 5 = 2100 \text{ cps}$$

The next one also divides by 5:

$$2100 : 5 = 420 \text{ cps}$$

The last one divides by 7:

$$420 : 7 = 60 \text{ cps}$$

which is the field or vertical frequency.

The total dividing ratio is

$$3 \times 5 \times 5 \times 7 = 525$$

which is, of course, the number of lines per picture.

2-5. Scanning generator

The scanning or sweep generators in the camera deflect the electron beam of the image pickup tube. The camera sweep generators use a technique similar

to that found in the corresponding part of a receiver; however, it is more refined because of the excellent sweep linearity required and expected. Linearity is somewhat facilitated by the reduced deflection angle, but in any case economical restrictions are by far less stringent in a camera than in a receiver. A properly designed and adjusted camera sweep generator will provide a non-linearity of less than 1 per cent.

2-6. Camera

Essentially, the camera contains a pickup tube, the associated deflection and focusing circuits, a video preamplifier to bring the video signal to a standard level, the optical system in front of the pickup tube (lens turret or zoom lens), and an electronic viewfinder showing the picture taken as it appears on the receiving screen. In addition, there are several auxiliary circuits for verification, remote control, signalization, intercommunication, etc.

The camera is connected to the remainder of the equipment through a multiconductor cable.

The most important element in the camera is the image pickup tube, which will be discussed in detail later in this chapter.

2-7. Video amplifier

The signal at the output of the camera has an amplitude of approximately one volt. The function of the video amplifier is to bring it to a level sufficient to ensure 100 per cent modulation of the r-f carrier. This means a signal amplitude of several tens or several hundreds of volts, depending on whether the modulation system is of a low-level or high-level variety, as described below.

Theoretically, any modulation process could be used. In application, however, most designs employ grid modulation, which requires much less power than plate modulation.

Figure 2-5 shows the simplified diagram of a grid-modulated r-f stage, which is of the usual symmetrical variety. Without going into detail it can be seen that the video signal is applied to the grid of the modulating tube and appears amplified at the anode across the load impedance Z . This load is a complex combination of resistive and reactive elements and is designed to produce constant

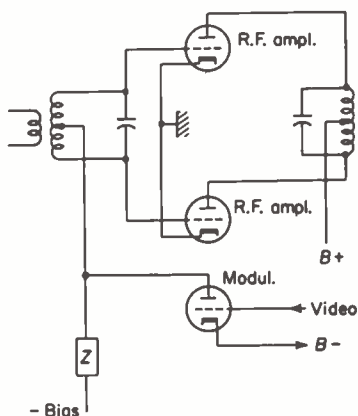


Figure 2-5. Example of grid modulation.

amplification over the entire video bandwidth. The modulating signal is then applied to both grids of the pair of r-f amplifying tubes through a center tap on the grid-tuned circuit.

If the r-f stage is one of the first in the r-f chain, the r-f signal has a relatively small amplitude and can be modulated by a video voltage of, say, 100 volts. This is called *low-level modulation*. If the r-f stage is near the output of the transmitter and handles considerable r-f power, a video signal of, say, 1000 volts is necessary to fully modulate the carrier. This is called *high-level modulation*.

2-8. D-c component

Some pickup tubes directly provide a convenient reference level for the d-c component because their output signal is set at black level during the return trace. Other tubes do not provide such a level. The d-c component may then be set arbitrarily at some convenient and sensible value, or a separate circuit is used for this purpose. For example, a photoelectric cell is used to measure the general illumination level of the televised picture and provides the requested reference value at which the level is set.

Frequently parasitic signals appear in the video output during retrace. They may have a large amplitude—larger in fact than the amplitude of the desired video signal. A convenient way to suppress these parasitic signals is to cut off one of the video amplifier stages during retrace, using a rectangular pulse derived from the blanking. Another scheme as shown in Fig. 2-6 adds a rectangular pulse to the video signal during retrace, which reduces the parasitic signals far below their original level. Then, a clipper stage working at a black level (level x in Fig. 2-6) will suppress the parasitic signals entirely and provide a clean video signal with a well-defined black level.

The sync pulses are added, and the complete video signal is amplified to reach modulation level. In the process, capacitive coupling is used between stages, and the d-c component is lost. It is then necessary to restore the d-c level at the modulating point. The same problem arises in receivers as well.

Two circuits may be used. One is in fact a peak detector, which employs a simple diode and is economical in design; it is used mainly in receivers. The other is a switching arrangement, more precise but also more complex, primarily used in transmitters. Its simplified diagram is shown in Fig. 2-7. The black-level reference pulse is applied to the grid of triode V_1 , which has equal anode and cathode loads and is working as a paraphase or split-phase amplifier. The plate and cathode signals, therefore, have equal amplitude but opposite phase. These signals simultaneously switch on the two diodes of V_2 , oppositely connected. The internal resistances of the two diodes and the resistors R_1 and R_2 constitute a Wheatstone bridge, so that the voltage at B is equal to the voltage at A .

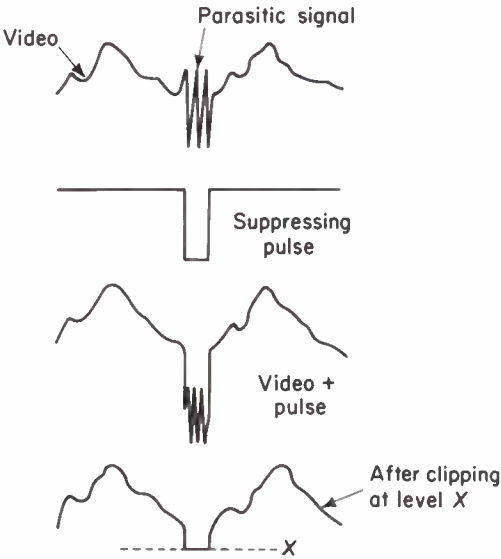


Figure 2-6. Parasitic signals produced during retrace can easily be suppressed.

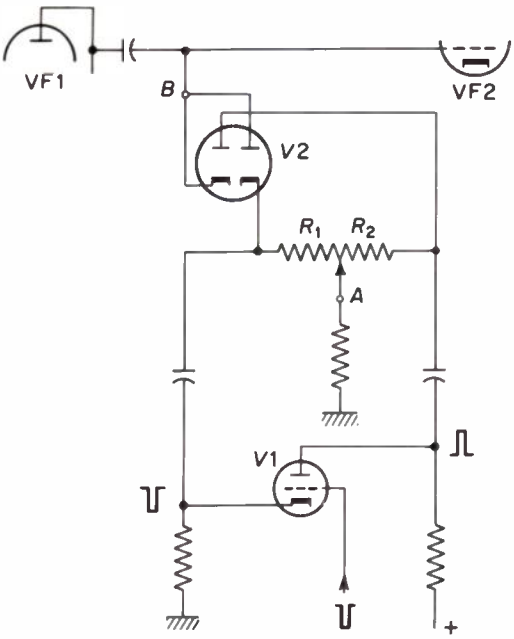


Figure 2-7. Double diode reference level switch.

The voltage at B sets the reference level for the signal, and it can be modified at will by adjusting the slider A of the potentiometer R_1-R_2 .

The unavoidable curvature of the tube characteristics may, during amplification, modify the ratio of picture signal to sync signal. For this reason it is better to use the black level as reference and not the sync pulse tip level. The alignment on black level may be made during the back porch at the black level following the sync pulse.

If the picture content varies and causes a change in the sync pulse amplitude, the black level will stay unchanged. The sync tips then will fall at a higher or lower level, according to their amplitude variation. This is of no great practical importance, because a simple clipper stage will clip all sync tips at the same value and re-establish uniformity.

2-9. Oscillator

The block labeled *oscillator* contains a complex of several stages. It embodies not only the oscillator proper but also frequency multipliers and, eventually, one or two power amplifiers.

The oscillator is controlled by a quartz crystal in one of the conventional designs. The frequency stability is excellent, especially because the quartz is placed in a temperature-controlled oven. The drift can be kept below 10^{-6} without much difficulty.

As previously mentioned, the crystal is generally cut for a frequency of a few megacycles. It can be made to oscillate on an overtone or mechanical harmonic, which economizes one multiplying stage. To obtain maximum stability, the quartz oscillator is required to deliver only a small output. If necessary, a power amplifier is used after the oscillator to drive the frequency multipliers.

A frequency multiplier is essentially an amplifier which has high distortion and consequently a large amount of harmonics in the output. An easy way to obtain such a characteristic is to overbias the tube and bring it into Class C operation.

A tank circuit in the anode is tuned to the desired harmonic frequency. Since the efficiency decreases when the harmonic order increases, the usual frequency multiplication factors are 2 or 3.

2-10. R-f amplifiers

R-f power amplifiers generally belong in the symmetrical category. A simplified diagram of such a stage is shown in Fig. 2-8. Its main peculiarity is the neutrodyning arrangement, obtained through capacitors C_1 and C_2 .

The anode-to-grid capacitance of a triode is rather high, which introduces some coupling between the plate and grid circuits. This coupling may cause

oscillation or instability. To counteract this, one method is to apply to the grid an r-f voltage of equal magnitude but opposite in phase to the one fed through the anode-to-grid capacitance. The two signals will cancel at the grid. In the circuit shown, opposite phase is secured by obtaining the canceling voltage from the anode of the opposite tube. Equal magnitude is obtained by adjusting the value of the neutrodyning capacitor C_1 or C_2 .

The above technique is applied to both tubes of the stage, resulting in cross-neutrodyning.

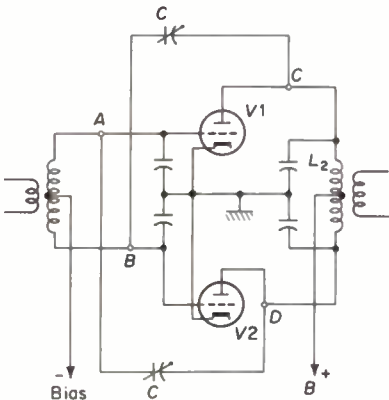


Figure 2-8. Symmetrical r-f stage with neutrodyning circuits.

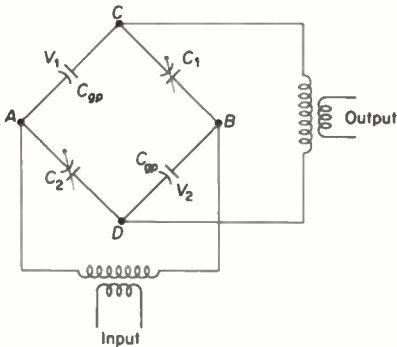


Figure 2-9. Bridge equivalent of the neutrodyning circuit.

The above circuit may be redrawn as shown in Fig. 2-9, where the grid-to-plate capacitances C_{gp} and the neutrodyning capacitors C_1 and C_2 are seen to constitute a bridge network. This illustrates the fundamental principle underlying neutrodyning.

To avoid the necessity of neutrodyning, a screen may be introduced between grid and plate to reduce the feedback capacitance. The tube then becomes a screen-grid tetrode or pentode. At high power, the manufacturing difficulties make such tubes rather costly.

Another arrangement is the grounded-grid circuit of Fig. 2-10. The input is on the cathodes and the grids are grounded and act as a shield between cathode and anode. This shielding effect will be true only as long as the grids are properly grounded. At very high frequencies, however, the grid lead inductance L is sufficient to prevent the grids from being held at ground

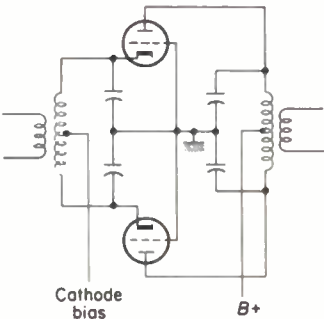


Figure 2-10. Grounded grid symmetrical amplifier.

potential. The remedy is to use a capacitor C in series with the inductor L so as to get a tuned series resonant circuit at the r-f frequency. At its resonant frequency, such a circuit acts as a short and effectively grounds the grid as shown in Fig. 2-11. Bias voltage is applied to the grids through decoupling resistors or r-f chokes.

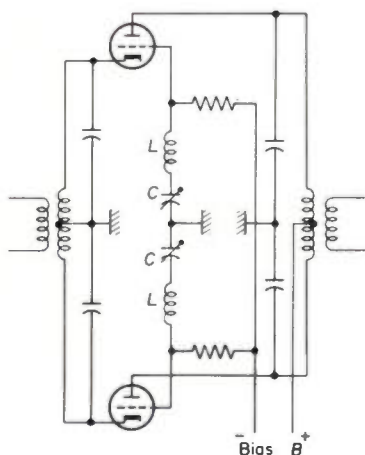


Figure 2-11. Tuned grid decoupling circuit.

out that such an arrangement is not practical and that the actual r-f spectrum (shown in Fig. 2-13) still contains part of the lower sideband. The suppressed part of this sideband may be eliminated by tuned circuits, called

2-11. Sideband filter

If it were possible to cut the r-f band sharply at the carrier frequency (as shown in Fig. 2-12), true single sideband transmission would result. It has been pointed

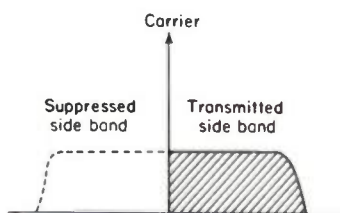


Figure 2-12. Single sideband transmission.

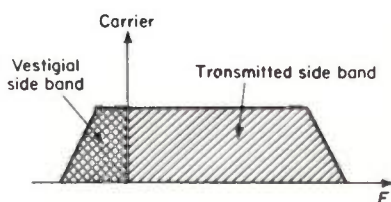


Figure 2-13. Vestigial sideband system.

sideband filters, connected to the output of the transmitter and dissipating as heat the r-f power contained in the suppressed part.

If low-level modulation is used, the r-f amplifiers following modulation may be mistuned toward the desired upper sideband deliberately and they may provide some attenuation of the unwanted frequencies. However, because of the video bandwidth, r-f tuned circuits must have poor selectivity, and the amount of attenuation they can provide is insufficient. It is still necessary, therefore, to use a sideband filter.

Such a filter is shown in Fig. 2-14, where R_a is the antenna radiation resistance and R_d a resistor dissipating the energy contained in the unwanted frequencies. The series resonant circuit L_2-C_2 is tuned on a frequency f_2 slightly lower than the carrier frequency f_0 . At this frequency f_2 , the impedance of

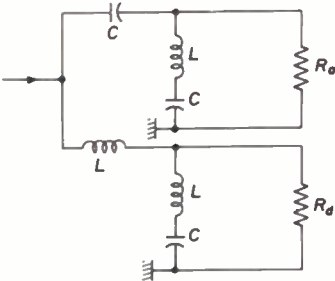


Figure 2-14. Sideband filter.

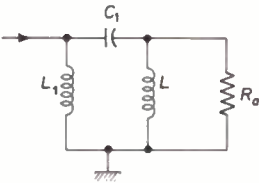


Figure 2-15. Equivalent circuit of the sideband filter for the transmitted band.

L_2-C_2 is zero and R_d is short-circuited. At the same time, L_3-C_3 appears as an inductive reactance L . The circuit then reduces to that shown in Fig. 2-15. All the available power is fed to the antenna represented by R_a .

The series resonant circuit L_3-C_3 (shown in Fig. 2-14) is tuned on a frequency f_3 corresponding to the desired cutoff. At this frequency, the circuit

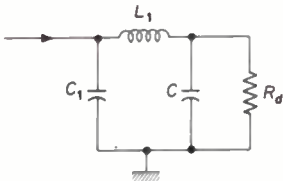


Figure 2-16. Equivalent circuit of the sideband filter for the suppressed band.

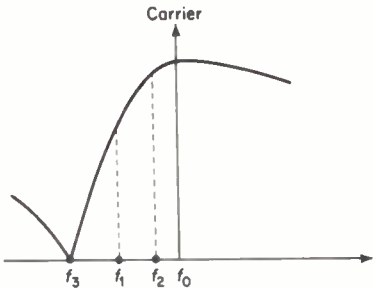


Figure 2-17. Practical response curve of a sideband filter.

impedance is zero, the antenna is effectively shorted out, and the circuit reduces to that shown in Fig. 2-16, where all the energy is dissipated by resistor R_d . The L_2-C_2 circuit appears as a capacitive reactance C at this frequency.

Notice the fundamental difference between the two resulting equivalent circuits of Figs. 2-15 and 2-16. The first (Fig. 2-15) is a high-pass filter circuit, the second (Fig. 2-16) a low-pass filter circuit. As a result, the high frequencies

of the band are more easily transmitted to the antenna, and the low frequencies are more easily transmitted to the dissipating resistor, R_d .

Resistor R_d should be able to dissipate sufficient power. A good part of the total energy, however, is in the carrier. The greatest part of the energy contained in the sidebands, furthermore, is concentrated in frequencies near the carrier. All in all, then, the dissipating resistor receives only a small amount of the total power—at most a few per cent.

R_d has a value equal to the antenna equivalent effective resistance R_a , and the filter is designed for an input impedance R_a . In actual practice, the tuned circuits of the theoretical diagram are replaced by suitable lengths of coaxial line of characteristic impedance R_a .

The frequency response of the filter shown in Fig. 2-14 is represented in Fig. 2-17. The two resonant frequencies f_2 and f_3 have been indicated, as well as the resonant frequency f_1 of L_1 - C_1 . Notice that the response rises up again below f_3 . This rise may make it necessary to cascade several filters of this type to obtain the required attenuation of the lower frequencies.

2-12. Antennas

The simplest antenna is the *dipole* shown in Fig. 2-18, also called *half-wave* antenna because its length is half the wavelength. It radiates a wave whose polarization is parallel to the dipole. That is, if the dipole is horizontal, the radiated wave is horizontally polarized.



Figure 2-18. The simplest antenna is the dipole or half-wave antenna.

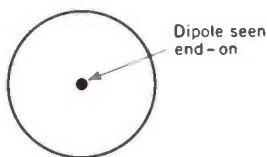


Figure 2-19. Omnidirectional radiation.

When one looks into the end of a dipole (as shown in Fig. 2-19), the energy appears to be radiated equally in all directions. The corresponding radiation diagram is a circle, so that in this plane the dipole is omnidirectional. With rare exceptions, such a radiation pattern is what is wanted in a horizontal plane for a television transmitter. A simple vertical dipole would then be suitable, but the presence of a tower or support would modify the radiation pattern. Moreover, it is customary to use high-gain antennas and several dipoles are necessary.

When one looks broadside at the dipole (as shown in Fig. 2-20), the radiation diagram is not omnidirectional but has an "8" shape. This means that for horizontal polarization the radiation would be zero along the axis of the

dipole, which is unsuitable for a transmitter. The radiation pattern can be considerably improved by using two dipoles set 90 degrees apart (as shown in Fig. 2-21) and fed with signals having a phase difference of 90 degrees. The corresponding horizontal radiation diagram, although cloverleaf in shape, provides good over-all coverage.

In the vertical plane, owing to the limited service radius, all the radiated power should be concentrated in or near the horizontal. This is accomplished by superposing several stacked antennas, for example, the crossed dipoles of

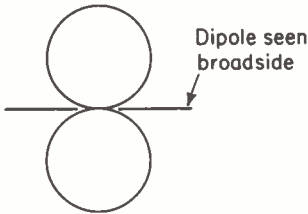


Figure 2-20. Figure 8 directional diagram.

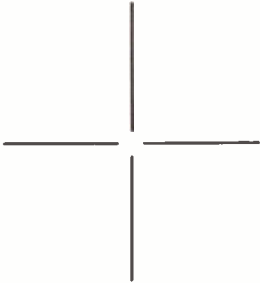


Figure 2-21. Two dipoles may be crossed at 90 degrees.

Fig. 2-21. The concentration of the radiation, and hence the useful gain of the antenna, increases with the number of elements so stacked. Generally, the elements have a vertical separation of one wavelength and consequently are fed in phase.

2-13. Conical antenna

Because of the width of a television channel, the antenna must be a wideband affair, and the bandwidth of a dipole may be increased by increasing its cross section—either by folding it or by making it thicker.

The cross-section increase may be progressive and linear (as shown in Fig. 2-22), leading to the conical dipole. Such an antenna has a generatrix of total length 0.73 wavelength. The radiation resistance is a function of the open angle and reaches 300 ohms at 29 degrees. The radiation directivity diagram is similar to the pattern of a simple dipole (shown in Fig. 2-20), except that the energy is more concentrated in the direction perpendicular to the axis of the cone. The “8” takes a more elongated shape, therefore, and this effect increases when the cone opening angle increases.

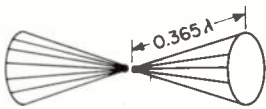


Figure 2-22. Double cone antenna.

2-14. Turnstile

The cross-section increase of a dipole may also be constant rather than tapered. Carrying this idea to an extreme, the dipole is made up (as shown in Fig. 2-23) of two metallic plates half a wavelength wide and a quarter wavelength long and fed at the center. The four inside corners, *A*, *B*, *C*, *D*, can be grounded. Current flow takes place everywhere in a direction parallel to the small side, that is, horizontally for the antenna position shown in Fig. 2-23.

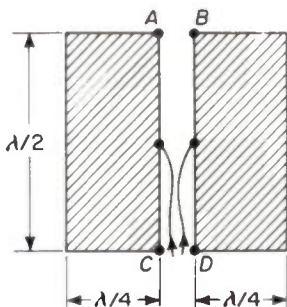


Figure 2-23. Metallic plates dipole.

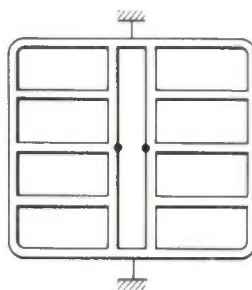


Figure 2-24. Metallic rods can replace a continuous surface.

This indicates that there is no major inconvenience or loss in replacing the continuous sheet by a sufficient number of horizontal conductors as shown in Fig. 2-24. The new arrangement has the advantages of being mechanically rugged and offering a greatly reduced area to the winds. Finally, as shown in Fig. 2-25, the horizontal conductors near the center are made shorter so as to obtain a concentration of current flow in the outside conductors and a better directivity—almost twice as good as that of the standard dipole.

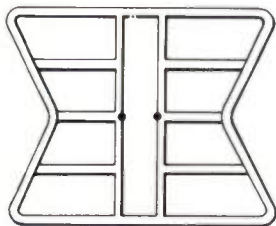


Figure 2-25. Actual shape of a turnstile antenna.

As before, two antennas of the type shown in Fig. 2-25 should be crossed at 90 degrees to obtain full horizontal coverage and fed 90 degrees apart in phase. Such an antenna is called a *turnstile* because of its appearance, and several turnstiles are superposed to concentrate the vertical directivity around the horizontal.

The turnstile antenna offers several advantages. The bandwidth reaches 25 per cent of the resonant frequency. The stacking is easy, and if n turnstiles are superposed, the gain in directivity reaches $1.2n$ with respect to the standard dipole. The feeding system is simple and easily installed.

Mechanically, the turnstile antenna is very rugged. The mast or tower is in the middle of the turnstile. Each element may be solidly fixed, welded or soldered to the mast at the four inside ground corners. The turnstile is made of soldered tubes; it has great rigidity and little wind surface. No insulating part is necessary.

Many other types of transmitting antennas may be and are used. The general requirements for transmitting antennas are always large gain, wide bandwidth, concentration of the radiated energy near the horizontal, omnidirectional horizontal directivity radiation pattern (except in special cases), and mechanical ruggedness.

2-15. Diplexer

The *diplexer* is a coupling element connecting two separate transmitters to a common antenna and at the same time preventing any intercoupling between the transmitters.

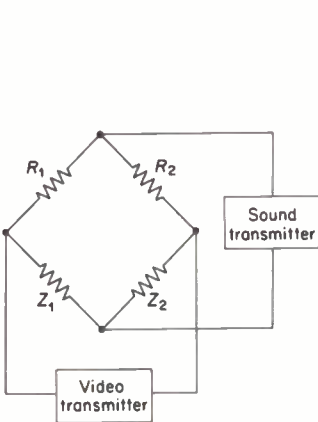


Figure 2-26. Symmetrical diplexer.

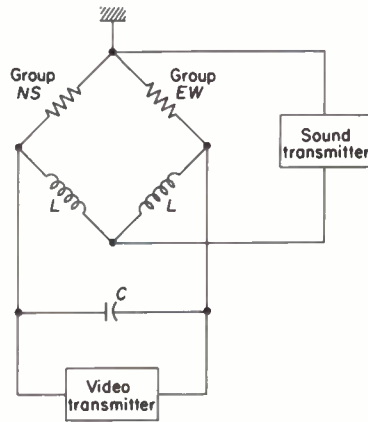


Figure 2-27. Practical circuit for a diplexer.

A simple common diplexer is the bridge circuit shown in Fig. 2-26. Resistors R_1 and R_2 represent the two halves of the antenna radiation resistance. Z_1 and Z_2 are two equal impedances to complete and balance the bridge. Because of bridge properties—or more simply in this case, because of symmetry—each transmitter creates equal voltages at the two points where the other transmitter is connected. Hence, no current flow through one transmitter is caused by the other transmitter.

In actual practice a diplexer looks more like the circuit shown in Fig. 2-27. The antenna is divided into two groups of elements having equal impedances—for example, the group north-south and the group east-west. The other two

impedances of the bridge are two equal inductances L . The two inductances in series are connected across a capacitance C so as to constitute an anti-resonant circuit tuned on the picture frequency. All the power from the picture or video transmitter is thus applied directly to the antenna. The power from the sound or audio transmitter has to go through the inductances to reach the antenna. The impedance of the inductors is very low, and practically all the sound power reaches the antenna.

Notice that the video transmitter feeds the two antenna halves in series, whereas the audio transmitter drives them in parallel.

In the actual arrangement, inductors and capacitors are replaced by suitable lengths of coaxial line. Also, an extra length of a quarter wavelength is added for one of the antenna groups to provide the 90-degree phase shift necessary for omnidirectional radiation.

2-16. Image pickup tubes

The variations in brilliance which constitute the picture to be televised are translated into electric current variations with the help of a light-current transducer. It is worth noting that one of the oldest systems, the flying-spot scanner, is still in use nowadays. In this process, the pickup tube is a simple photoelectric cell. A number of light-current transducers have been designed and developed, from the image dissector to the vidicon. A few of them will be rapidly described to illustrate their principles of operation.

2-17. Flying spot

The basis for the flying-spot system is shown in Fig. 2-28. The luminous spot produced on the screen of a scanned cathode-ray tube is projected on the televised subject through an optical system. One or several photoelectric cells collect the reflected light and translate it into the video signal.

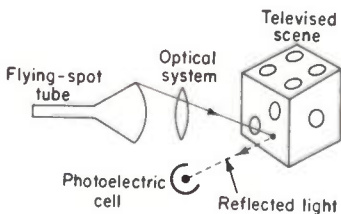


Figure 2-28. Flying-spot scanning system.

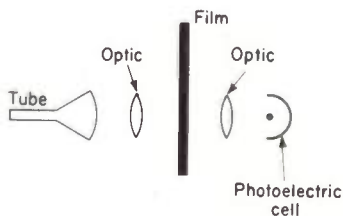


Figure 2-29. Flying-spot film scanning.

As such, the process is hardly usable for live pickup because of its lack of sensitivity. It is, however, still in use in the modified version shown in Fig. 2-29

for fixed (slides) or semi-fixed (motion picture film) subjects. In this arrangement, the photoelectric cell directly receives the light coming from the cathode-ray tube through the slide or film.

2-18. Image dissector

Farnsworth's image dissector camera tube is made (as shown in Fig. 2-30) of a large plane photoemissive cathode, an accelerating anode, and a signal anode, in an evacuated glass envelope.

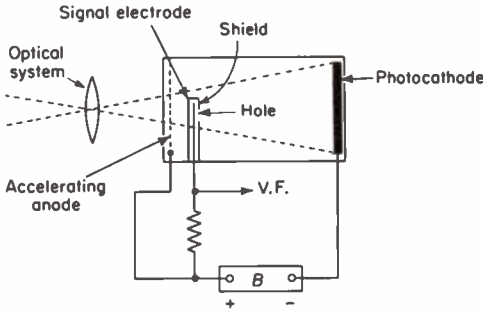


Figure 2-30. Farnsworth dissector.

The optical image of the televised subject is projected onto the photocathode, every point of which emits electrons in a quantity proportional to the amount of light which strikes the point. The optical image is thus transformed into an electrical image. The accelerating anode is positive with respect to the cathode, and all emitted electrons are attracted simultaneously with the result that the total electrical image moves instantaneously down the tube. A focus coil wound around the tube keeps the electrical image correctly focused in the anode plane. The image dissector is considered to be an instantaneous scanner because no energy storage is involved.

The signal anode is surrounded by a small tube shield in which a small hole has been drilled. The only electrons to reach the signal anode and give a signal current are those passing through the hole and originating at a given point on the photocathode.

The output signal is taken on the signal electrode.

To analyze the picture, each point of the photocathode must sequentially determine the output signal. This is easily done by moving the total electrical image simultaneously as an electron cloud across the tube with the help of external deflection coils. Each point of the photocathode is then used only during the short time that the scanning makes the electrons it emits pass through the shield hole.

Since the optical picture is projected through the length of the tube, its path should not be obstructed. For this reason, the accelerating anode is made of fine metallic gauze and the signal electrode is small.

The dissector has a rather low sensitivity and needs a highly illuminated subject such as can be obtained with movie projectors. In order to increase the signal current an electron multiplier is incorporated within the tube structure.

2-19. Electron multiplier

When a high-speed electron hits a specially prepared surface, the impact may be sufficient to knock several electrons out of the surface. These extracted electrons are said to be *secondary* and the phenomenon is called *secondary emission*.

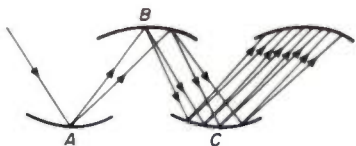


Figure 2-31. Secondary emission electron multiplier.

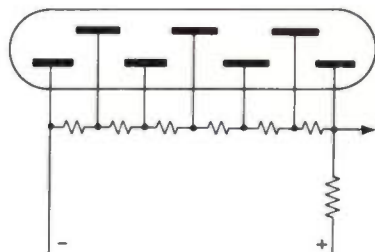


Figure 2-32. Voltage supply for a multistage electron multiplier.

The best secondary emitters are compounds similar to those used in photocathodes. They are based on cesium, silver, antimony, or their oxides.

The number of secondary electrons produced for each primary electron is called the *multiplying ratio*. It reaches 10 to 15 for some compounds. It depends markedly on the speed of the primary electrons, so that the best multiplication is obtained for a given range of accelerating voltage acting on the primary electrons.

Secondary emission is used in electron multipliers (shown in Fig. 2-31). A primary electron hits a first surface *A* and knocks off n secondary electrons onto a second surface *B*, where each of them ejects n electrons. The total number of secondary electrons produced on *B* is then $n \times n = n^2$, and they are accelerated toward a third surface, *C*, where they create n^3 new electrons, and so on. A number of multiplying stages can be cascaded in this fashion. Since the multiplication follows a geometric progression, very large gains can be obtained with electron multipliers.

The cascaded electrodes, *A*, *B*, *C*, etc., are connected to increasing voltages and have geometrical shapes that create the right electric field to guide

secondary electrons along the right path. A single voltage source may be used across a voltage divider (Fig. 2-32) to provide the required electrode voltages.

It is not necessary that the secondary emission surface be a good conductor. The phenomenon occurs, for example, in cathode-ray tubes when the electron beam impinges on the screen. This may pose an interesting practical problem. The secondary electrons from the screen phosphor are collected by a conductive coating, inside the conical part of the tube and connected to the final accelerating anode. Frequently, the coating uses aquadag, a graphitic conductive substance.

Since the screen is an insulator, it cannot compensate for a lack or an excess of electrons. Nevertheless, the total current must be zero. Two possible conditions may occur, as follows:

In the first case, a strong negative charge accumulates on the screen and repels all primary incoming electrons. If there are no primary electrons, there are no secondary electrons and the total current is zero.

In the second case, the screen takes a potential which is slightly positive with respect to final anode. This potential adjusts itself so that the total number of secondary electrons effectively leaving the screen is equal to the number of incoming primary electrons. The total current again is zero.

The existence of a charge on the screen of a cathode-ray tube can be verified by pressing a finger against the screen. The charge will locally leak away by capacity and the scanning raster will be distorted. This distortion does not occur with all tubes.

2-20. *Multiplying dissector*

The sensitivity of the image dissector can be greatly increased by including an electron multiplier in the device. Its mechanical arrangement is such that this is easily done. The multiplier is placed inside the small tube which shields the signal anode.

The last electrode of the multiplier is connected to a load across which is taken the amplified video output signal.

The electron multiplication principle is used in other pickup tubes; however, the physical arrangement is different.

2-21. *Iconoscope*

Zworykin's iconoscope uses a discontinuous photosensitive surface made up of a large number of small, elementary photocells. Such a surface is called a *mosaic*, as shown in Fig. 2-33, and is deposited on a thin plate of glass or mica

in the form of a great number of droplets or globules of a photosensitive material. The other side of the glass or mica plate carries a continuous metallic film which is the output electrode. Each droplet is part of a capacitor having the glass or mica as dielectric and the conductive backing as common electrode.

The iconoscope camera tube shown in Fig. 2-34 contains in an evacuated envelope a mosaic *M*, a collector ring *C* to collect secondary electrons, and an electron gun *K* whose electron beam *F* impinges on the droplet side of the mica or glass plate. The collector *C* is connected to the final accelerating anode of the electron gun.

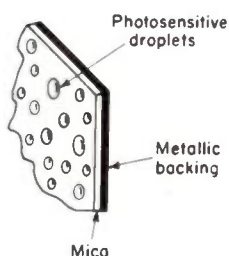


Figure 2-33. Photo-sensitive mosaic.

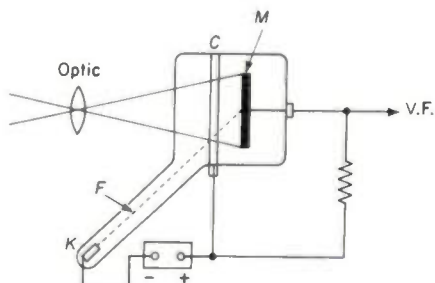


Figure 2-34. Principle of operation of the iconoscope.

An optical lens projects the image on the photosensitive mosaic. The electron gun is placed at an oblique angle so as not to be in the path of the optical rays. Each photosensitive droplet receives an amount of light proportional to the brilliance of the corresponding point in the televised scene and accordingly emits some electrons which are attracted by the collector ring. The small capacitor having this droplet as front electrode thus takes a charge, positive on the droplet side. The charge pattern produced on the whole mosaic area makes up an electrical image corresponding to the projected optical picture.

It is now necessary to collect these elementary charges in scanning order to obtain the video signal. The electron beam is deflected over the mosaic surface by a set of scanning coils. Every time it sweeps over a droplet, it replenishes it with electrons and in the process discharges the corresponding capacitor. The discharge current flows through a resistance connected to the common conductive backing and develops the output video signal across it. The output circuit is comprised of the conductive backing, the load resistance, the collector ring, and the mosaic. The droplets are actually very small, and the focalized electron beam on the mosaic sometimes covers as many as a hundred.

The iconoscope exhibits a fundamental difference compared to the image

dissector. In the image dissector, each point of the photosensitive surface is active only during the time that it is being analyzed. In other words, of the electrons produced, only those emitted during the short time corresponding to the scanning of a point will actually be used.

In the iconoscope, the charge produced by photoemission builds up continuously on each droplet and is collected only once during each scanning period. This accumulation process is important because it has a direct bearing on the magnitude of signal produced—that is, in fact, on sensitivity. It may mean an improvement of 100,000 times or more compared to the dissector sensitivity.

However, the iconoscope suffers from a major disadvantage. When the electron beam hits the mosaic, a fraction of the secondary electrons does not reach the collector ring but falls back on more positive parts of the mosaic, thus creating spurious signals called *shadows*. The difficulty is that the distribution of these stray electrons depends on the charge pattern on the mosaic, that is, on the picture content. Shadows may be corrected for by suitable circuitry, but a trained operator is needed to continuously adjust the settings for best results. An additional minor inconvenience results because the electron gun is set at an angle with respect to the mosaic, producing a scanning raster wider at the top than at the bottom. This *keystone* distortion must be corrected by preset circuits.

2-22. Image iconoscope

The image iconoscope shown in Fig. 2-35 is an outgrowth of the iconoscope which puts to good use the principle of transport of electronic image already embodied in the image dissector camera tube.

A lens *O* projects the optical picture onto a transparent photocathode *P*. The other side of this cathode carries the photosensitive layer and an electronic picture is emitted as in the iconoscope. A suitable electric field throws the electronic picture onto the mosaic *M*, where each impinging electron knocks off several secondary electrons. The charge pattern created on the mosaic reproduces the optical scene electrically, but secondary emission has intervened in the process and brought in a sensitivity improvement by a factor on the order of 5. Of course, there is no need for the mosaic to be photosensitive, but it must have a high multiplication ratio.

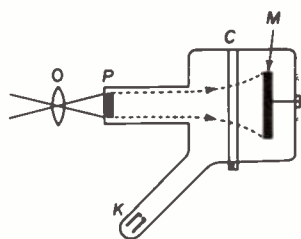


Figure 2-35. Image iconoscope.

A collector ring *C* and an electron gun *K* complete the tube, in which the analysis of the mosaic and the extraction of the output signal proceed along practically the same lines as for the iconoscope.

The transport of the electronic picture between photocathode and mosaic takes place with some enlargement in size.

The sensitivity is about ten times better than that of the standard iconoscope, in large part owing to secondary emission magnification, and to a lesser extent, to a better use of the photocathode.

The main practical advantage of the image iconoscope is that the photocathode is small and is placed at the front end of the tube. This allows the use of economical, short-focal-length lenses. For a given maximum stop, the lens is smaller and less expensive, and the depth of field is larger. Nevertheless, the image iconoscope suffers from shadows just as does its predecessor, the iconoscope.

2-23. Orthiconoscope

The orthiconoscope, or more commonly, orthicon, is also derived from the iconoscope (see Fig. 2-36). A lens O projects the optical picture onto a

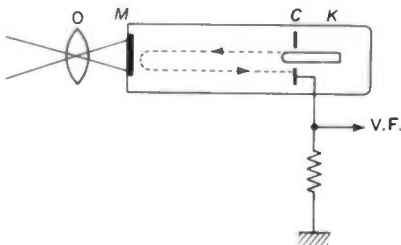


Figure 2-36. Principle of the orthiconoscope or orthicon.

photosensitive mosaic M . This mosaic now has a transparent conductive backing, so that the picture can be projected through it and the mosaic placed against the front end of the tube, with the same advantages as for the image iconoscope. Moreover, the electron gun K can be placed behind the mosaic in the axis of the tube, eliminating keystone distortion.

The main innovation, however, is that the electron beam is accelerated by a low voltage, and hence contains only slow electrons. Also, the mosaic is at cathode potential so that a decelerating field exists in the space before the mosaic which slows down the electrons to the point where they reach the mosaic with zero velocity.

In the absence of a charge on the mosaic, conditions are such that the electron beam just grazes the mosaic before being turned back by the reverse field. It then travels back along the tube and lands somewhere in the vicinity of the electron gun, where it is collected by a special electrode C . The video frequency output signal VF is taken across a load resistance connected to the collecting electrode.

Now, if a mosaic element has emitted electrons because of the illumination of the corresponding point in the televised scene, it assumes a positive charge and attracts electrons when the scanning beam happens to sweep past it. The returning beam has lost some electrons and the resulting current variation through the load resistance creates the video output signal.

The focusing and deflection of the beam originally created some problems. These have been solved through the use of a set of coils around the tube (not shown in Fig. 2-36).

Because of its principles of operation, the orthicon is free from shadows. It has one minor drawback, however. If intense illumination accidentally falls on the mosaic, the high positive charge produced may be such that the secondary emission ratio exceeds unity. The mosaic then keeps a high positive charge and the tube blanks off. This can be remedied by cutting off the tube for a short time.

Practically speaking, a worth-while improvement is that the orthicon does not use a real mosaic as does the iconoscope. The photosensitive surface is continuous and is made of an extremely thin film of activated silver. The transverse resistance of such a film is very high and is sufficient to provide suitable insulation between adjoining points.

2-24. Image orthicon

The image orthicon combines several of the features already mentioned—transparent photocathode, electronic picture transport, continuous mosaic, slow electron beam and an electron multiplier.

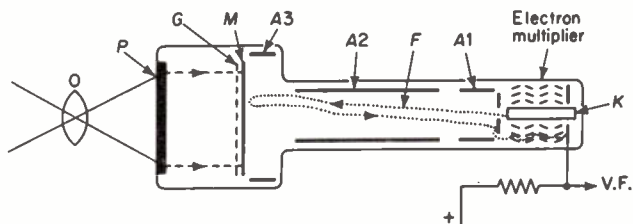


Figure 2-37. Construction of an image orthicon.

It is a highly complex tube, as shown in Fig. 2-37. A lens *O* projects the optical picture onto a transparent photocathode *P*, where it is transformed into an electronic image. The photocathode is at -300 volts with respect to the mosaic *M*, so that the electric field between them projects the electron image toward the mosaic. On impact, secondary electrons are produced and collected by a special electrode *G* of fine metallic gauze, placed very near the mosaic and set at a potential of 1 volt. The electric charge pattern thus produced on the mosaic leaks to the other side of the supporting plate, which is made of very thin, conducting glass. The charge leaks through the plate in less than one scanning period and there is no loss involved in the process.

As for the orthicon, the mosaic is, in fact, a continuous film and the transverse resistance of the supporting glass plate is high enough to provide adequate insulation.

The final result, on the far side of the mosaic plate, is an electronic picture, which is analyzed by a slow electron beam F just as in the orthiconoscope. This beam is first accelerated by anodes A_1 and A_2 , then, in the vicinity of the mosaic, it is decelerated by anode A_3 . It is focused and deflected by a set of external coils.

The beam returns back toward the electron gun K as in the orthicon camera tube, but this time the collecting electrode surrounding the cathode is the first stage of an electron multiplier whose output provides the video signal.

The sensitivity of the image orthicon is much better than that of the simple orthicon, mainly because of secondary emission gain on the mosaic and because of the electron multiplier. The improvement in sensitivity is on the order of 100–500.

As opposed to the orthicon, the image orthicon does not blank off in the event of intense illumination. Its drawback in this respect is more localized. If a strongly illuminated small area appears in the picture, the corresponding mosaic area takes on a strong positive charge and some of the secondary electrons fall back on the mosaic. This phenomenon is akin to shadow production in the iconoscope, but now the metallic grid above the mosaic limits the effect to a small zone immediately around the bright spot. This zone receives extra electrons and consequently becomes “blackier.” The result is that the bright area is surrounded by a dark halo.

This is not too important, particularly because the image orthicon permits such a wide range of contrast that the limit in this direction is set rather by the remaining associated equipment.

2-25. Vidicon

All the pickup tubes considered in the previous sections use the principle of photoemission, whereby electrons are produced when light falls on a photo-sensitive material. A new class of light-current transducers uses photoconductivity. This phenomenon occurs, for example, with selenium or antimony. When light falls on a piece of selenium, the conductivity or resistance of the metal varies in proportion to the light intensity.

The typical photoconductive pickup tube is the *vidicon* shown in Fig. 2-38. An evacuated envelope contains a signal plate P , a fine wire gauze G , and an electron gun K .

The signal plate P is a very thin, conducting metallic film; it is transparent and deposited inside the front end of the tube. On top of this film is laid another thin layer of photoconductive material, which constitutes the photo-sensitive element. As before, this layer is so thin that the transversal resistance is high enough to ensure sufficient insulation between neighboring points.

The electron gun K projects a beam of slow electrons on the photoconducting layer. Grid G is at a potential higher than the potential of the signal plate

so that the space between them contains a decelerating field which slows down the electrons to the point where they reach zero velocity on the photosensitive surface and turn back, in a manner similar to the image orthicon.

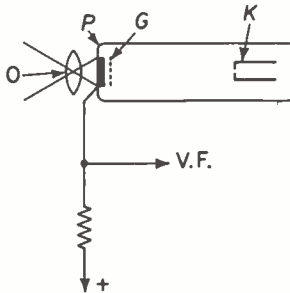


Figure 2-38. Principle of the vidicon.

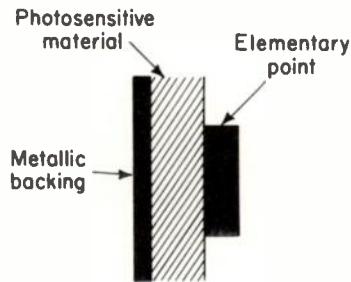


Figure 2-39. Elementary point surface in a vidicon.

With no light falling on the signal plate, conditions are such that the electron beam just grazes the photosensitive surface and keeps it uniformly at cathode potential. As opposed to the image orthicon, no use is made of the returning beam, which falls on a collecting electrode surrounding the electron gun. The beam is focussed and deflected by external coils.

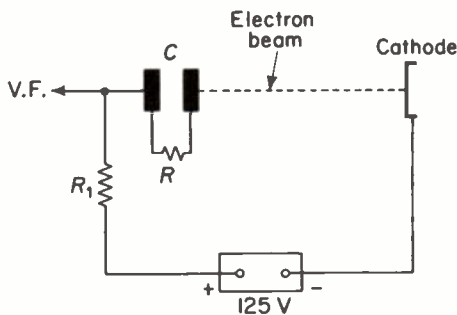


Figure 2-40. Electrical equivalent circuit of a vidicon.

Since the transversal resistance of the sensitive layer is very large, each point may be considered as insulated from the adjacent points and the equivalent of a mosaic is obtained. Every point, as shown in Fig. 2-39, is then equivalent to one globule of the iconoscope mosaic, except that it is not separated from the common metallic backing by an insulating layer of mica or glass, but by a resistive layer of photosensitive material.

The equivalent circuit of such an arrangement is shown in Fig. 2-40. The area covered by an elementary point constitutes a small capacitor C with a

metallic backing. The dielectric of this capacitor is leaky since it is made of the photoresistive material and is represented by the leakage resistance R .

The photosensitive layer has a thickness on the order of only 1 or 2 microns, and in darkness its resistance per square centimeter ranges between 10 and 100 megohms.

The left-hand electrode of the capacitor is the metallic common film backing, which is connected to the 125 volt supply. The right-hand electrode is the internal surface of the photosensitive layer. Once every scanning period this surface is swept over by the electron beam and brought to the cathode potential, which is 0 volts. Every time this happens, the capacitor is charged up to $125 - 0 = 125$ volts. When no light falls on the sensitive surface, the leakage resistance R is very high, the time constant RC is long, and the capacitor C practically does not discharge between sweeps. When the electron beam again sweeps over the area, it meets a surface still at zero potential. There is no charge and no charge current as a result.

If the sensitive surface is now illuminated, photosensitivity comes into play and resistance R decreases by a large amount, thus reducing drastically the time constant RC . The capacitor C discharges through R between successive sweeps, and the lower R is, or the more intense the illumination, the more complete the discharge.

The potential of the right-hand electrode goes up toward 125 volts and reaches a positive value which depends on the time constant, hence on the illumination, by the time the electron beam sweeps over it again and brings it back to zero. The resulting charge current is a function of the illumination level and it flows through load resistance R_1 across which is taken the video output signal VF.

The right-hand electrode takes a positive potential which is always small compared with 125 volts. The discharge is maintained at the beginning of the exponential discharge curve and is nearly linear.

The same process comes into play for all elementary picture points and when they are swept over in scanning order, the video voltage developed across R_1 is a faithful electrical image of the light distribution of the televised scene.

Compared to other pickup tubes, the vidicon is simple and rugged. It can be made quite small, with resulting advantages in weight, volume, and ease of design of the external coils. The sensitivity is excellent and compares favorably with the sensitivity of the image orthicon; it could eventually be increased by incorporating an electron multiplier in the design. The optical lenses may be standard models for motion picture cameras and are not expensive.

2-26. Telecinema

The use of motion picture film for television, or *telecinema*, poses peculiar problems. Standard film projectors employ a discontinuous motion, the film

moving at the rate of 24 pictures per second. With the light intensities presently in use, this rate produces an intolerable flicker. Flicker is eliminated by cutting off the light for a short time in the middle of the projection of each picture, thus artificially increasing the rate to 48 pictures per second.

Television, however, uses 30 pictures and 60 fields per second. It is not possible to increase the film speed to these values because the increased 25 per cent difference would be obvious in the picture and sound. Furthermore, the vertical blanking time in television is on the order of $\frac{1}{60}$ second, and the film would have to be pulled down one picture in this short time, which causes serious mechanical problems.

A simple film-scanning system is based on the remanence or memory of accumulation tubes, of which the iconoscope is a simple example. If the memory is at least equal to one scanning period, it is only necessary to impress the picture on the pickup tube once every scan. This is done during vertical retrace, when the tube is blanked off.

The picture is flashed on the mosaic through a special mechanical obturator in the film projector and stays inscribed on the tube, because of remanence, during the following scan. High illumination levels can be used in the film projector, and good results are obtained with this system.

Other arrangements have been devised. Some use continuous motion of the film, which is very satisfactory from a mechanical point of view, and compensate for this motion by an opposite motion of the scanning spot. This is easily arranged with flying-spot scanners.

The simplest of all processes, however, uses a sensitive tube such as a vidicon, which has enough memory to keep the picture for more than a scanning period. The film is then normally projected by a standard projector, and the pickup tube is scanned in the usual way.

Problems

1. Make a block diagram of the carrier generator for the sound transmitter of Channel 3. Accuracy of frequency is obtained by resorting to quartz crystal control. The quartz should oscillate on a frequency lower than 5 mc.

2. In the above problem, assume the stage immediately following the quartz oscillator is frequency-modulated. What should be the maximum frequency of deviation at this stage for the total frequency deviation of the carrier to be 25 kc?

3. Explain how vertical directivity may provide effective antenna gain.

4. Design a synchronization generator in block form for a television standard of 819 lines, 50 fields per second.

5. Propose a simple design for an electronic viewfinder to be enclosed in a camera.

6. Discuss the relative merits of three different modulation systems as applied to a picture transmitter.
7. Compare high- and low-level picture modulation.
8. The picture carrier for Channel 13 is obtained by frequency multiplication, starting with a quartz oscillator on 2.608024 mc. If the quartz has a frequency stable to ± 50 cps, what is the maximum variation in carrier frequency?
9. Make a block diagram and explain the operation of a generator for the picture carrier for Channel 11. The frequency must be controlled by a quartz working below 5 mc.
10. Design a sideband filter using coaxial lines. Give the line dimensions for Channel 11.
11. From the purely practical point of view of ease of production and installation and mechanical ruggedness, compare the dipole, the conical, and the turnstile antennas.
12. Propose a diplexer circuit other than the standard arrangement given in the text. Discuss its design and operation.
13. Design a diplexer circuit using coaxial lines. Calculate the line dimensions for Channel 8.
14. Make a mechanical diagram of a flying-spot scanner for fixed slides. Add the auxiliary electronic equipment in block form and explain its operation.
15. An electron multiplier has a secondary emission ratio $n = 3.684$. How many stages are necessary to provide a gain of one million?
16. What is a typical size for the individual photosensitive droplets in an iconoscope having a mosaic 4 by 6 cm? How many droplets are there on the mosaic?
17. Explain the large increase in sensitivity between image dissector and iconoscope. Evaluate its numerical value.
18. Make a block diagram of a circuit to correct for keystone distortion in an iconoscope and explain its operation.
19. Explain why the orthiconoscope produces no shadow in comparison to the iconoscope.
20. Barring accumulation or remanence process, design a film-scanning arrangement in block form. The film should advance at standard film rate and the vertical frequency of the television system is 60 cps.

Chapter 3

ANTENNAS

Although it does not properly belong to the television receiver, the antenna is highly important because it determines to a large extent the ultimate quality of the picture received.

The short wavelengths used in television make for peculiar properties of propagation and, with the necessity of a wide band, determine practical antenna designs. This, in turn, has some repercussions on the type of lead-in cable and on the input circuits of the receiver.

3-1. Block diagram of a receiver

A television receiver, as shown in Fig. 3-1, consists essentially of a picture, video, or vision receiver—which provides the modulating voltage for the picture tube and the synchronization signals—the horizontal and vertical scanning generators or time bases, a sound receiver, a general power supply and a high-voltage power supply for the picture tube, and finally, the receiving antenna, which is not part of the receiver proper but which has such a large effect on the final performance that it merits as much attention as any of the preceding parts.

The picture receiver, as shown in Fig. 3-2, embodies several stages and follows the same general lines as a radio receiver. It generally has an r-f amplifier, a frequency changer, an i-f amplifier, a detector, and a video amplifier. The amplified frequencies are rather high, by both radio r-f and i-f standards, and the bandwidth is much wider. This last point, as will be seen later, reduces the available gain per tube with the consequence that, for sufficient over-all amplification, a television receiver needs a large number of tubes. The picture receiver would typically include an r-f amplifying tube, a frequency changer

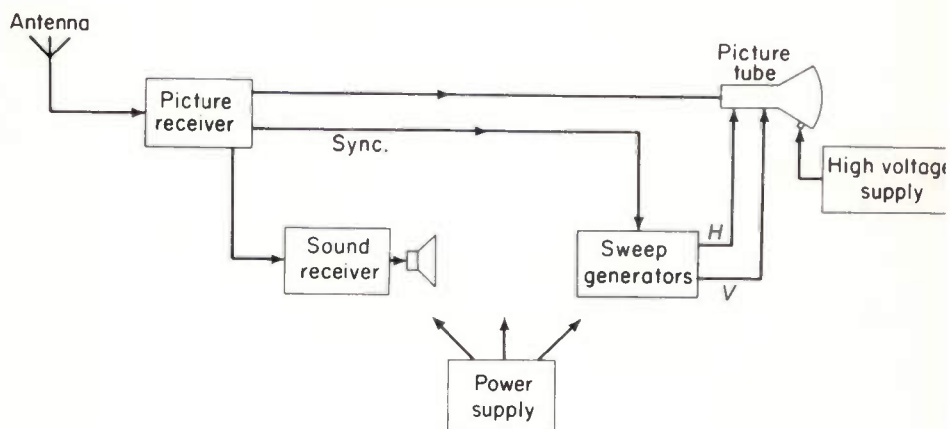


Figure 3-1. Block diagram of a television receiver.

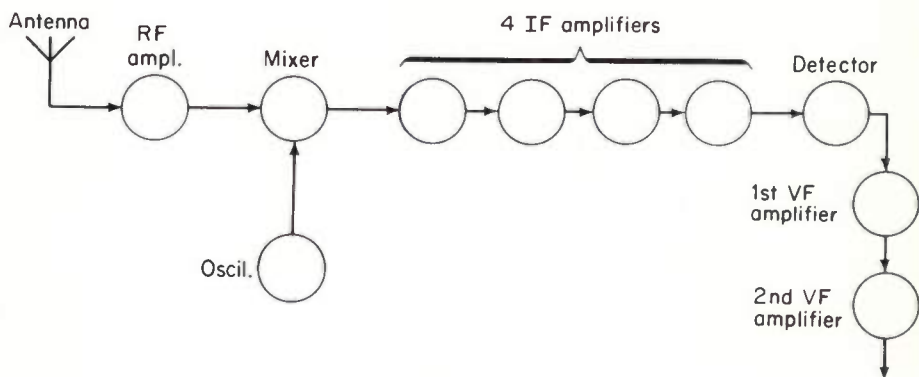


Figure 3-2. Typical stages of a picture receiver.

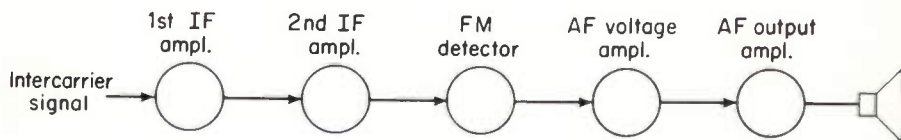


Figure 3-3. Typical stages of a sound receiver.

with separate oscillator, four i-f amplifying tubes, a detector, and two video amplifying tubes.

The exact number of tubes varies somewhat from one design to another because of different performance requirements or economic considerations. The sound receiver, as shown in Fig. 3-3, is generally of the intercarrier variety. It receives its input signal from the detection or video stages of the picture receiver, and typically is comprised of two i-f amplifying tubes, an FM detector, a voltage a-f amplifying tube, and an a-f output power tube.

Almost all receivers in use have electromagnetically deflected picture tubes and power is required for the deflection coils. Each time base—vertical and horizontal—will then contain a sawtooth oscillator and a power amplifier. The horizontal time base is almost universally accompanied by an auxiliary circuit designed to provide picture tube high voltages ranging typically from 10 to 30 kilovolts.

The general power supply provides a voltage of 100 to 400 v and can deliver a current of several hundred milliamperes. A number of circuits are used, from the half-wave a-c—d-c type to the full-wave transformer type.

A number of auxiliary circuits will also be encountered—horizontal damping, automatic gain control, sync separation, d-c restoration, etc.

The antenna is sometimes justly considered as part of the receiver although it is more precisely a part of the receiving installation, and a very important part at that. In fact, a later chapter will be devoted to each block of the general block diagram given in Fig. 3-1.

3-2. Wave propagation

The wavelengths used in television belong in the metric and decimetric classes. In other words, they range from 0.1 to 10 meters. Radiation in these bands exhibits quasi-optical properties; in particular, the waves propagate along straight lines and can be reflected by obstructions. These peculiarities become more pronounced as the wavelength decreases and the frequency increases.

Long-range radio communications rely on the wave reflecting properties of the ionized layers of the high atmosphere. Under normal conditions, this reflecting mechanism fails for wavelengths shorter than, say, 10 meters, so that television waves normally do not propagate beyond the radio horizon. However, it may occur that the reflecting properties still hold good for much shorter wavelengths, and this explains why television transmissions are sometimes received at great distances. Such spurious and erratic propagation is at best unreliable and the service area of a television station is considered to be limited to the radio horizon or to line-of-sight transmission.

It is customary, then, to consider only direct wave coverage—that is, as a first approximation, optical coverage. Since this increases with the height of

the antenna, the usual practice is to set the antenna atop a high tower or mast, which is itself on top of the highest building, mountain, or hill available.

3-3. Optical coverage

The optical radius of coverage of an antenna is easily computed with reference to Fig. 3-4. R is the earth radius, h the height of the antenna, x the optical radius. The optical radius x is tangent to earth at the horizon, and there perpendicular to the earth radius R . In the resulting right triangle one may write

$$(R + h)^2 = R^2 + x^2$$

$$\text{or } R^2 + h^2 + 2hR = R^2 + x^2$$

Figure 3-4. Computation of the radius of coverage.

Since in all practical cases h will be much smaller than R , it is justified to neglect h^2

with respect to R^2 . The equation then becomes

$$2Rh = x^2$$

The earth radius is approximately

$$R = 6300 \text{ km}$$

so that

$$x = (12,600h)^{1/2}$$

with x and h in kilometers. In more practical units, with x in kilometers and h in meters, the expression becomes

$$x = 3.6h^{1/2}$$

approximately. Alternatively, the metric units can be replaced by the customary American units, which gives the expression

$$x = 1.22h^{1/2}$$

where x is in miles and h in feet.

The resulting value gives the optical radius for the transmitter only. If the height h_1 of the receiving antenna (shown in Fig. 3-5) is considered, its own optical radius x_1 must be added to the transmitter radius to obtain the maximum total coverage p :

$$p = x + x_1$$

or using the preceding results,

$$p = 3.6(h^{1/2} + h_1^{1/2})$$

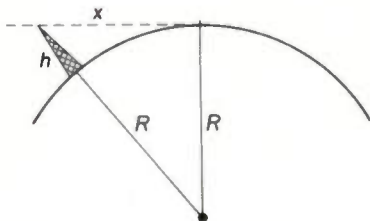


Figure 3-5. Total coverage, including receiving antenna.

in kilometers and meters, or

$$p = 1.22(h^{1/2} + h_1^{1/2})$$

in miles and feet.

The actual radio horizon is larger than the optical horizon because of propagation peculiarities. It is obtained with sufficient accuracy for practical needs by adding 15 per cent to the preceding results.

The radio horizon r for a transmitting antenna would then be

$$r = 4.14h^{1/2}$$

in kilometers and meters, or

$$r = 1.4h^{1/2}$$

in miles and feet.

3-4. Field intensity

Theoretically, the electric field intensity E in volts per meter at a distance d meters from the antenna is given by the classical expression

$$E = \frac{7W^{1/2}}{d}$$

where W is the effective radiated power (ERP) in watts.

For example, a 40-kw ERP transmitter will provide at 50 miles, or approximately 80 km, a field intensity

$$E = \frac{7(40,000)^{1/2}}{80,000} = 0.0175 \text{ v/m}$$

This value is far above any measured practical value. The discrepancy is due to the fact that the theoretical expression has been computed for an antenna isolated in space and does not take into account ground effects, absorption, diffraction, reflection, etc.

3-5. Reflected wave

For the short wavelengths used in television, the earth behaves as a mirror which reflects incident waves. Propagation between transmitter and emitter, shown in Fig. 3-6, can then take place along two different paths. The direct path produces what is called the *direct wave*. The indirect path hits the ground and is reflected, bringing in the *reflected wave*. As for any reflected wave, there is a 180-degree phase shift of the reflected wave when it hits the ground. Moreover, since this wave is propagated along a longer path than the direct wave, it will incur a phase delay with respect to the direct wave.

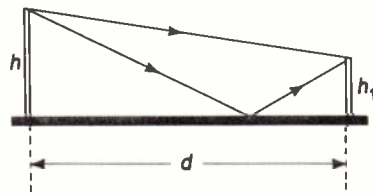


Figure 3-6. Reflection on the ground.

The total field at the receiving point will result from the combination of both waves. It can be calculated on the basis of the geometry of Fig. 3-6 and is

$$E = \frac{88hh_1W^{1/2}}{d^2\lambda} \text{ v/m}$$

with h = transmitting antenna height in meters,

h_1 = receiving antenna height in meters,

d = distance in meters,

λ = wavelength in meters,

W = effective radiated power in watts.

For example, using $h = 300$ m, $h_1 = 25$ m, $d = 80,000$ m, $\lambda = 2$ m, and $W = 40,000$ w, the field would be

$$E = \frac{88 \times 300 \times 25 \times 200}{64 \times 10^8 \times 2} = 0.01 \text{ v/m}$$

which is again much larger than any practical result because the only auxiliary effect considered was ground effect, all the others again being neglected.

The preceding expressions, then, have no great practical value except that they indicate the possible effects of varying the parameters. To increase the received signal, for example, at a given distance and on a given frequency, one can only raise the antennas or increase the transmitter power.

The last possibility is not too attractive economically, particularly since quadrupling the power would only double the field intensity.

At the receiving end, the only possible solution according to the above equations is to increase the height of the receiving antenna.

As to the other factors, decreasing the wavelength would increase the field. However, the efficiency of the receiving dipole decreases with the wavelength, and the two effects will cancel out as far as the signal at the receiver input is concerned.

The field strength also decreases with the square of the distance. This is only valid inside the radio horizon. Beyond the radio horizon the decrease is much more rapid, the received signal being only that part of the radiation which is diffracted at the point of tangency with the ground.

3-6. Multiple pictures

Television wavelengths are easily reflected when they hit an obstruction whose dimensions are at least on the order of a wavelength, whether it is conductive or not. In the latter case, it is sufficient that the dielectric constant of the material (concrete, brick, etc.) be different from that of air for reflection to occur, particularly if the angle of incidence is small.

Besides the direct wave and the ground reflected wave, there is also the possibility, as shown in Fig. 3-7, of a wave's arriving from a different direction after having been reflected by an obstruction. At first one might think that this extra wave would only combine with the direct wave and produce an increase or decrease of the total signal according to the phase relation. Unfortunately, matters are worse—the result is a loss of definition of the picture or even the superposition of two pictures on the screen. The second picture, usually weaker than the normal one and shifted laterally, is the echo or ghost picture.

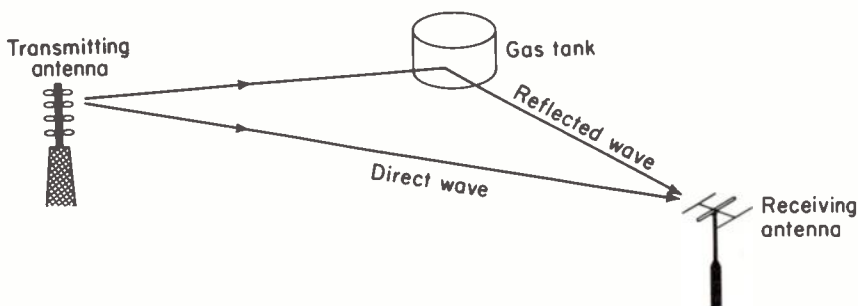


Figure 3-7. Reflection on an obstruction.

Even worse, if several reflections occur on several obstructions, an equal number of ghosts will appear and practically ruin the quality of the picture.

The basic reason for the appearance of a ghost is that the reflected wave travels a longer distance than the direct wave, and the same signal reaches the receiver twice—first when the direct wave arrives, and second when the reflected wave arrives. It is not difficult to link the cause and effect mathematically. The velocity of radio waves in air is 300,000 km/sec. If the reflected path is 300 m longer than the direct path, the reflected wave will arrive

$$\frac{300}{3 \times 10^8} = 10^{-6} \text{ sec}$$

or 1 microsecond after the direct wave. During this time, the spot will have moved to the right across the screen at the rate of 1 cm/microsecond, approximately, for a large-size picture tube. The ghost will then appear 1 cm to the right of the real picture. Since reflection is rarely complete, the reflected wave is weaker than the direct wave and the ghost is weaker than the main picture.

For smaller path differences, the picture and its ghost are separated by a smaller distance and sometimes cannot be distinguished. For example, if the difference is only 30 m, the two pictures will be 1 mm apart and cannot be separated. The result is a drastic loss of apparent definition.

It is not an easy matter to eliminate troublesome ghosts. The only point of attack is the receiving antenna, and the usual technique is to improve its directivity so as to reduce the ghost with respect to the desired signal.

3-7. Dipole

The receiving antenna, of course, should be designed to provide a maximum signal to the input of the receiver. A common way to obtain this result is to establish the antenna so that it resonates at the desired frequency. It is then said to be *tuned*. The simplest of tuned aerials is the half-wave dipole shown in Fig. 3-8, which is frequently used in television.

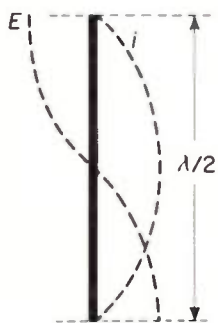


Figure 3-8. Voltage and current distribution in a dipole.

When the dipole is excited by an incoming wave on its resonant frequency, stationary waves appear in the antenna and the voltage and current distribution follows a sine law. Voltage and current exhibit a phase difference of 90 degrees. Both ends of the dipole are insulated, and there can be no current flow there. The current is zero, the voltage is maximum. At the center of the dipole the opposite condition prevails; there is a maximum of current and a zero or node voltage.

The sine distribution of voltage and current along the dipole is also indicated in Fig. 3-8.

If the impedance of the dipole is calculated on the basis of the ratio of voltage over current, it is seen to vary from zero at the center to infinity at the ends. This would be true only if there were no losses and if the antenna did not have radiating properties.

The length of the dipole is theoretically half the wavelength. Now the wavelength in space is given by

$$\lambda = \frac{300}{F}$$

where λ = wavelength in meters,

F = frequency in mc,

or equally well by

$$\lambda = \frac{984}{F}$$

with λ in feet and F in mc. The length of the dipole would then be

$$l = \frac{150}{F} \quad \text{or} \quad l = \frac{492}{F}$$

The first expression gives l in meters, the second gives l in feet.

Actually, an antenna is never standing by itself in free space. There are always some roofs, trees, buildings, or means of support for the antenna in the

vicinity, to say nothing of the connecting cable or feeder. Furthermore, television antennas are made of large-diameter wire or tubing so as to be wideband, and all these considerations introduce a correcting factor in the calculations of the length of the antenna. For a practical dipole, the resonant frequency is always slightly larger than twice the length of the antenna.

The correcting factor varies somewhat with the length-to-diameter ratio of the dipole. For practical purposes, it is sufficient to assume a constant correction of 5 per cent. The practical length of the dipole, then, is given by

$$l = \frac{142}{F}$$

with l in meters and F in mc, or

$$l = \frac{467}{F}$$

with l in feet and F in mc.

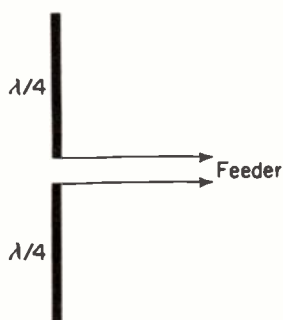


Figure 3-9. Connection of the feeder or lead-in line.

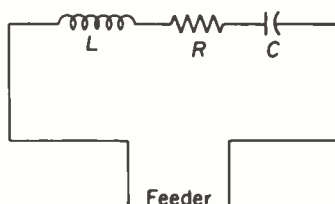


Figure 3-10. Resonant circuit equivalent to the antenna.

The exact length of the dipole is not too critical because of the wide bandwidth for which it is established.

The dipole must necessarily be connected to the transmitter to receive energy or to the receiver to deliver energy. This is done by opening the dipole at its center, transforming it into two quarter-wave stubs, and connecting the center end of each stub to the feeder or lead-in cable, as shown in Fig. 3-9. Since the dipole is symmetrical, the feeder should also be symmetrical. Viewed from the feeder, the resonant antenna appears exactly as a resonant circuit having inductance, resistance, and capacitance, as shown in Fig. 3-10. The resistance is called the *radiation resistance* because it corresponds to the power lost due to radiation by the dipole. It has a constant theoretical value of 73.2 ohms for all dipoles, variously rounded off to 70, 72, 73, or 75 ohms.

At resonance, the inductive reactance of L and the capacitive reactance of C exactly cancel each other, and the antenna appears as a pure 73.2 ohms resistance.

3-8. Sensitivity

The sensitivity of an antenna may be gaged by the amount of voltage appearing at its output for a given field intensity.

An equivalent definition is that of *effective height*, which has nothing to do with the height of the antenna above ground but depends on the type and dimensions of the antenna.

The voltage V at the antenna terminals is equal to the field intensity E multiplied by the effective height H :

$$V = EH$$

with V in volts, E in volts per meter, H in meters. For the dipole, the effective height is

$$H = \frac{\lambda}{\pi}$$

where λ is the resonant wavelength in meters.

3-9. Bandwidth

The same antenna is used to receive the picture and sound carriers, or more exactly, the entire television channel. It must therefore have a minimum bandwidth of 6 mc and should be tuned on the center frequency of the channel, which is the geometric mean of the extreme frequencies. For practical purposes the arithmetic mean gives sufficient accuracy.

Bandwidth depends mainly on the length-to-diameter ratio of the dipole wire, rod, or tubing. The smaller this ratio, the wider the bandwidth. A single channel dipole, to be satisfactory in this respect, should have a length-to-diameter ratio of 100 at most.

A secondary effect of increasing the conductor diameter is a slight reduction of the radiation resistance. For the above-mentioned ratio of length to diameter of 100, the resistance drops to 70 ohms, approximately.

3-10. Directivity

It has been stated previously that the directivity diagram of a dipole is omnidirectional in the plane perpendicular to the dipole, which means that the antenna theoretically exhibits the same sensitivity in all directions, as shown in Fig. 3-11. If the dipole were horizontal, it would be equally sensitive to waves coming almost parallel to ground or to waves falling vertically from the sky.

In the plane containing the dipole, the directivity diagram follows a cosine law and has the shape shown in Fig. 3-12. In other words, the sensitivity is maximum in both directions along a line perpendicular to the dipole at its center.

For the usual horizontal polarization of the transmitted waves, the dipole is horizontal, so that Fig. 3-11 represents the directivity in the vertical plane or vertical directivity, and Fig. 3-12 gives the horizontal directivity.

Once again these curves are purely theoretical and apply to a dipole isolated in free space. The presence of a more or less perfect ground (mast, feeder, roof, building, earth, etc.) below the antenna modifies the directivity diagrams drastically. In particular, the vertical directivity takes the complicated shape shown in Fig. 3-13, where the simple circle has been replaced by a number of lobes whose importance and angular distribution depend on the effective height of the dipole above ground. The main lobe is the first one; the others are secondary lobes.

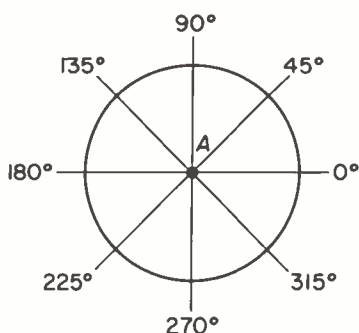


Figure 3-11. The dipole has an omnidirectional diagram in the plane perpendicular to the dipole.

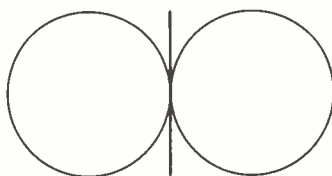


Figure 3-12. Horizontal directivity follows a cosine law.

In television, the incoming wave generally arrives at an angle very near the horizontal, and in this case it pays to favor the first lobe and reduce the secondary lobes. Theoretically again, this would be obtained by setting the dipole a quarter wavelength or less above a ground plane. This is not convenient in practice. It will be found experimentally that, in the great majority of cases, the higher the antenna the better the results.

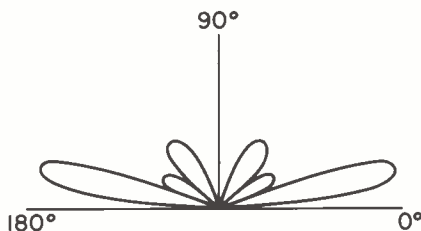


Figure 3-13. Actual directivity diagram showing the various lobes.

It will be noticed in Fig. 3-13 that between lobes there exist directions of zero or very low sensitivity. This is sometimes the reason for very poor performance when the direction of arrival of the wave coincides with a direction of low antenna sensitivity. In this case, moving the antenna *up* or *down*

changes its distance to ground and shifts the lobe pattern, thus improving reception.

3-11. Lead-in cable

The energy collected by the antenna is very small and must be transferred to the receiver with a minimum of losses by the lead-in cable or line. With symmetrical antennas such as the dipole, symmetrical or twin-lead lines are preferred for best results. However, coaxial cable, although not a symmetrical

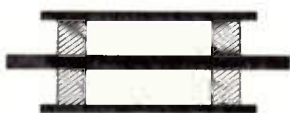


Figure 3-14. Bead centered co-axial cable.



Figure 3-15. Solid dielectric co-axial cable.

line, is sometimes used. It is made of a solid or stranded conductor running along the axis of a tubular external conductor. The inner conductor is kept centered by insulating beads, as shown in Fig. 3-14, or even the entire space between the two conductors may be filled by a good dielectric (Fig. 3-15). The outer conductor is a braided sheath which acts as a perfect shield for the inner cable and makes it immune to external parasitic fields. With a good insulator such as polyethylene and a large cable, the losses may be as low as 2 decibels per 100 feet at 100 mc. Coaxial cable, although costly, may be superior to symmetrical lines when the outside part of the lead-in is long or the signal is weak, when there are interferences due to parasitic fields acting directly on the line, or when the weather is particularly bad or irregular.

In addition to losses per unit length, another important line parameter for television use is the characteristic impedance. A line terminated by a load equal to its characteristic impedance will deliver a maximum power to the load and there will be no reflections at the load due to mismatch. Coaxial cables are made for a range of characteristic impedances commonly comprised between a few tens and a few hundred ohms. The more or less standard RG59/U has a characteristic impedance of 72 ohms, which directly matches a dipole impedance at resonance.

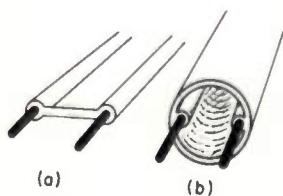


Figure 3-16. Ribbon lead-in lines.

The second type of common lead-in is the parallel twin-lead or ribbon line shown in Fig. 3-16(a), sometimes found in the tubular form of Fig. 3-16(b). It uses two conductors

held at constant spacing by a molded-on dielectric and is a symmetrical line. It performs well and is relatively inexpensive and easy to use. It can also be made to cover a wide range of characteristic impedances, the standard value being 300 ohms. In its flat ribbon form, or twin-lead, it is by far the most common type of lead-in. One good reason for this is that the input impedance of receivers has been standardized at 300 ohms symmetrical and provides a very good match to the line impedance.

3-12. Losses in the lead-in line

In all cases where signal strength is limited, losses in the lead-in line are an important consideration and should be reduced to a minimum. On open lines of the ribbon type, for example, losses increase drastically when the line is wet,

Table 3-1. Line losses

Type of line	100 mc		700 mc	
	Dry	Wet	Dry	Wet
300-ohm ribbon.....	1.2	7.3	3.6	26.5
300-ohm tubular.....	1.2	2.5	3.6	8.2
72-ohm coaxial RG59/U	3.7	3.7	11.7	11.7
72-ohm coaxial RG11/U	1.9	1.9	6.2	6.2

that is, under rain. This is particularly true at UHF frequencies. Table 3-1 gives the losses for the common types of lead-in lines at a representative VHF frequency of 100 mc and at a typical UHF frequency of 700 mc. The losses indicated are in decibels per 100 ft of line.

For all lines, losses vary very roughly as the square root of frequency as a first approximation. The table shows that the standard ribbon, despite low losses when dry, produces six to seven times the loss when wet. The tubular line is a parallel line where the two conductors are not embedded in a flat ribbon, but are held inside a tube of good insulating material at two diametrically opposite positions, as shown Fig. 3-16(b). In this way, the space between conductors is protected against moisture if both ends of the tubular line are properly sealed, with the result that rainy weather only doubles the losses. In this respect, a tubular line is much better than a flat ribbon line for a 300-ohm impedance antenna, especially at UHF frequencies.

Coaxial cable is just as good wet as it is dry, which is not surprising since the inner conductor is completely protected from weather. Losses are higher than losses in parallel lines in dry weather, and the impedance of 72 ohms requires an antenna of the same impedance for best results.

Taking economic considerations into account, the most common line is the flat 300-ohm ribbon followed by the 300-ohm tubular line.

3-13. Reflections

If the input impedance to the receiver does not match the line characteristic impedance, all the energy is not transferred to the receiver. A part of it is reflected at the impedance mismatch and is sent back up the line to the line-antenna junction. If there is another mismatch there, part of the reflected energy will be reflected again and travel down the line, reaching the receiver after traveling twice the length of the line. This involves a time delay proportional to the length of the line and may create ghosts or, more properly in this case, echo pictures. The displacement due to these reflections may be calculated in the following manner.

Let l be the length of the line in meters. The wave propagates in the line with a velocity v lower than in free space and which depends on the line. A typical value is 200 m/microsecond.

The delay t of the echo will be

$$t = \frac{2l}{v}$$

or in this case,

$$t = \frac{l}{100} \text{ microseconds}$$

Again assuming a large screen and a spot velocity on the screen of 10 millimeters per microsecond, the distance d between the picture and the echo on the screen will be

$$d = 10t = \frac{l}{10}$$

with d in millimeters and l in meters.

For example, a line 100 ft long, that is, $100 \times 0.3048 = 30$ m approximately, would produce an echo $30/10 = 3$ mm or roughly $\frac{1}{8}$ in. to the right of the picture (1 in. = 25.4 mm).

The reflections at the ends of the line are not limited to two, but carry on to infinity. Actually, the attenuation at each reflection is large, and successive reflections appear on the screen as regularly spaced echos which become weaker and weaker.

In principle it would be sufficient for one of the impedance matches, at either antenna or receiver, to be good for the echo to disappear. It is difficult, however, to obtain consistent good matching over a large bandwidth, and it is necessary to provide as good a match as possible at both ends of the line if reflections are to be avoided entirely.

3-14. Line connection

The connection of a balanced line such as the twin-lead to a symmetrical antenna such as the dipole is simple. Each conductor of the line is connected

to one of the quarter-wave stubs. The line should lead away perpendicularly to the antenna for at least a quarter wavelength to avoid disturbing the antenna characteristics.

A coaxial cable is not symmetrical but nevertheless is frequently used as though it were. The inner and outer conductor are connected as shown in Fig. 3-17 respectively to each half of the dipole, exactly as a twin-lead line. The outside braid is not grounded as it normally is, but carries an r-f current. It is always protected by a thick layer of external insulating material.

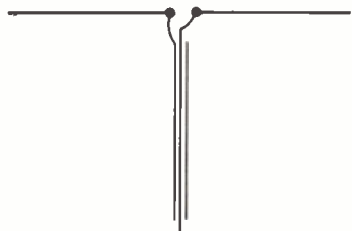


Figure 3-17. Symmetrical connection of a coaxial cable.

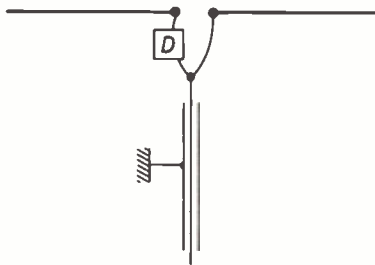


Figure 3-18. A phase inverter *D* may be used in one side of the dipole.

If the outer conductor must be grounded and both halves of the dipole used, only the inner conductor is available to carry the signal. Since the signals in each half-dipole are in opposite phases, one of them must undergo a phase shift of 180 degrees if they are to add in the common conductor. A phase-shifting circuit *D* (Fig. 3-18) is used to accomplish this. The simplest arrangement is a half-wavelength extra length of line, which introduces a 180 degrees phase shift in the connection to one of the halves of the dipole.

Phase inverters or impedance adapters add to the complexity of the installation and do not work too well over wide bands. They are not common in ordinary antenna installations.

At the receiver input, the correct connection for a balanced line is symmetrical and is shown in Fig. 3-19. The first tuned circuit of the receiver is shown, and the primary coil provides a transformer effect which matches the receiver input to the line and provides a voltage step-up. The center tap on the primary coil is generally grounded, although sometimes there is no tap and the diagram looks like that shown in Fig. 3-20.



Figure 3-19. Symmetrical input for a balanced line.

This type of input is also used when the line is a coaxial cable with the two conductors carrying r-f. In strict orthodoxy, coaxial cable properly used, with the outer conductor grounded, should be connected as shown in Fig. 3-21 to a

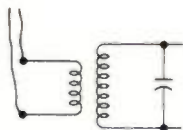


Figure 3-20. Symmetrical input with no center tap on the primary.

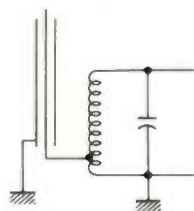


Figure 3-21. Correct connection of a coaxial cable.

nonsymmetrical input. In this case, it is easier and less expensive to replace the input transformer to the receiver by an autotransformer as shown.

3-15. Parasitic elements

The simple dipole has insufficient sensitivity and directivity for most practical cases. To improve these characteristics, a more complicated type of antenna must be used, and a very common one is the Yagi antenna, where the dipole is

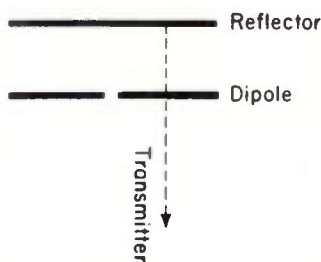


Figure 3-22. Reflector behind the dipole.

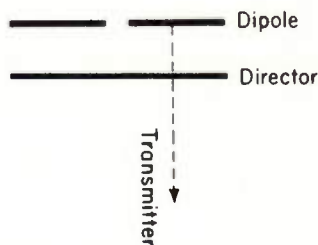


Figure 3-23. Director in front of the dipole.

accompanied by unconnected parasitic elements parallel to the dipole and excited by it because they are near enough for substantial coupling to exist.

If the directivity is increased in the direction parasitic element-dipole, the additional element is a reflector and should be placed, as shown in Fig. 3-22, behind the dipole with respect to the transmitter.

If the directivity is increased in the direction dipole-parasitic element, the additional element is a director and should be placed, as shown in Fig. 3-23, between the dipole and the transmitter.

Each arrangement constitutes a two-element Yagi. Nothing forbids the simultaneous use of a director in front of the dipole and a reflector behind it.

The resulting antenna shown in Fig. 3-24 is a three-element Yagi. The process can be carried on by adding new parasitic elements farther away from the dipole, for example, as shown in Fig. 3-25, where a second director has been added in front of the first one, giving a four-element Yagi. There are still more complicated arrangements, with several reflectors and an even larger number of directors.

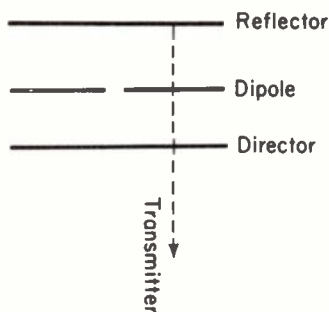


Figure 3-24. Three-element Yagi antenna.

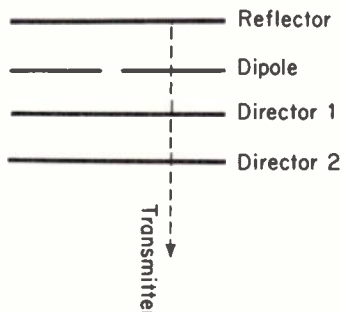


Figure 3-25. Four-element Yagi antenna.

Generally speaking, the effect of a parasitic element depends on its spacing from the dipole and on its tuning—that is, on its length. Reflectors are longer and directors are shorter than the dipole in the common Yagis.

With a number of parasitic elements and two variables—spacing and length—for each of them, the over-all characteristics of a Yagi antenna may be adjusted to cover a very wide range of cases. For example, the simple two-element Yagi may be adjusted to provide all intermediary directional diagrams between the two extremes shown in Fig. 3-26. Diagram *a* corresponds to a large spacing, diagram *b* to a small spacing. The dotted circle gives the standard dipole diagram for comparison.

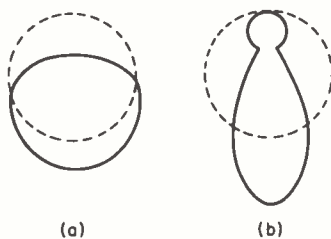


Figure 3-26. Typical directional diagrams.

Two undesirable effects are noticed when a Yagi is designed for improved gain and directivity: the bandwidth decreases and so does the radiation resistance in the center of the dipole.

3-16. Bandwidth limitation

The decrease in bandwidth limits the improvement in gain and directivity that a Yagi can provide. This is especially true if several stations are to be received

and the antenna is designed to cover a whole band or part of a band including several channels.

The gain falls off quickly enough at the edges of the band, so that the sensible technique would be to choose a Yagi providing maximum or near-maximum gain on the weak stations and to sacrifice some gain on the strong local transmitters. This is not always possible or economically sound.

In the case of a single channel antenna, the Yagi may be a high-gain, high-directivity model. Limitations in this case are reached not so much because of reduced bandwidth but because of diminishing returns. Adding another parasitic element to an already complex Yagi hardly improves performances. Ten elements seem to constitute a practical maximum in this respect.

3-17. *Folded dipole*

The decrease in the dipole radiation resistance due to the addition of parasitic elements can be compensated for by replacing the dipole with a folded dipole.



Figure 3-27. Two dipoles can be paralleled.



Figure 3-28. The lead-in line is connected to the center of one dipole.

It will be remembered that if the diameter of the conductor of a dipole is increased, the bandwidth increases but the r-f power P collected by the dipole does not change. Instead of increasing the dipole diameter, an equivalent effect is obtained if we place two identical dipoles near each other and connect them in parallel by joining their extremities as shown in Fig. 3-27. The collected power is still the same, and each dipole carries a current which is half of the current which would flow through a simple dipole. The lead-in line is easily connected by opening one of the dipoles at its center as shown in Fig.



Figure 3-29. Folded dipole antenna.

3-28. Such a configuration is a *folded dipole*. The usual arrangement does not use separate dipoles connected at the ends, but rather a single length of conductor folded to obtain an identical form, as shown in Fig. 3-29.

As for the dipole, and for the same reasons, the over-all length of the folded dipole around the two branches is not λ , but 0.94λ . This would be the length of the single piece of conductor shown in Fig. 3-29 before folding.

3-18. Folded-dipole impedance

Let I be the current through a simple dipole and Z_d its impedance. The r-f power P it collects is

$$P = Z_d I^2$$

In the case of a folded dipole, the r-f power P is the same, but the current in each branch is half the dipole value, or $I/2$. Let Z_f be the folded-dipole impedance. Then

$$P = Z_f \left(\frac{I}{2}\right)^2$$

or

$$P = Z_f \frac{I^2}{4}$$

whence, for the same power,

$$Z_f = 4Z_d$$

The radiation resistance of the folded dipole is four times that of the simple dipole. Since $Z_d = 73.2$, then

$$Z_f = 4 \times 73.2 = 292.8 \text{ ohms}$$

commonly rounded off to 300 ohms.

This indicates that, in any antenna, replacing the simple dipole by a folded dipole will increase the antenna impedance fourfold. In general, moreover, the response of a folded dipole is more uniform over a frequency range as compared to a simple dipole.

3-19. Special cases

The same reasoning can be extended to the double folded dipole of Fig. 3-30. The current through each dipole is one-third of the current which would flow through a single dipole, and the impedance of the double folded dipole is nine times that of the simple dipole, that is, approximately 660 ohms.

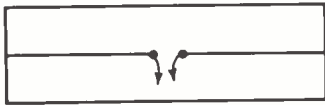


Figure 3-30. Double folded dipole.



Figure 3-31. The two dipoles may have different diameters.

There is another way to adjust the impedance of a folded dipole. The current will be divided into two equal parts only when the two dipoles have identical diameters. For example, if the open dipole (to which the line is connected) has a smaller diameter than the continuous dipole shown in Fig. 3-31, it will carry less current and the impedance at its center will increase.

Thus, the impedance of a folded dipole having members of unequal diameters is a function of the ratio of the diameters. Because of the close proximity of the two dipoles and of the resulting coupling, the impedance also varies with the spacing.

The only case when the spacing need not be considered is when the diameters are equal. In this case, there is no need for the spacing even to be constant nor for the dipoles to be parallel.

3-20. Arrays

The sensitivity of a Yagi antenna does not increase in proportion to its complexity, that is, to the number of parasitic elements, whether they are reflectors or directors. The practical limit of improvement over a simple dipole seems to lie at about 12 db.

If more sensitivity is needed, either other types of antennas must be resorted to, or several Yagi arrangements must be grouped together in a complex antenna array.

Two possibilities are available: the Yagis may be placed side by side (juxtaposed) or stacked one above the other (superposed). Juxtaposition and superposition may be combined in high-gain arrays. Whatever the method of combination, the gain increases approximately 3 db every time the number of Yagis is doubled. This might seem a small result for such a large effort, but in borderline cases where a few decibels make the difference between a usable picture and no picture, the improvement is well worthwhile.

For example, the gain of a given Yagi is 8 db. It increases to 11 db if two identical Yagis are combined and to 14 db if four Yagis are used.

Because of mechanical considerations, arrays of more than four Yagis are somewhat uncommon. Also, because of horizontal polarization, it is easier and more common to stack the Yagis than to juxtapose them. What is more, directivity increases with the number of Yagis in the plane of combination, so that superposition increases the vertical directivity and favors waves arriving at a small angle over the horizontal.

3-21. Superposition

The case of two superposed Yagis is shown in Fig. 3-32. For purposes of clarity in the drawing, only the folded dipole of each Yagi has been shown; the parasitic elements are omitted.

Since the two folded dipoles are at the same distance from the transmitter, the signals they receive are in phase. The illustration shows how these two signals may be added together simply by arranging the connections in such a way that they travel the same distance to the receiver: a line links the two

dipoles, and the lead-in is connected to the middle of this line. The usual spacing between Yagis is $\lambda/2$ or λ .

When they are spaced $\lambda/2$ apart, Fig. 3-33 shows another way of combining the two signals so that they add. The lead-in is connected to the lower dipole,

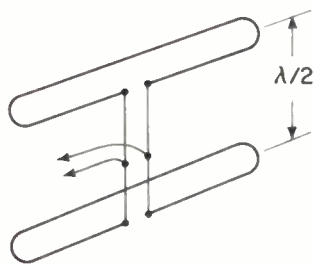


Figure 3-32. Superposing two antennas.

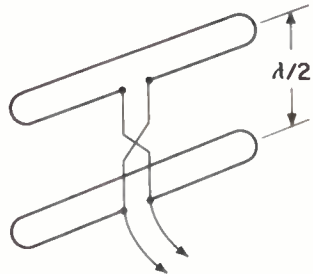


Figure 3-33. Variant for the line connection.

which is linked to the upper dipole by a half-wave line. The signal coming from the upper Yagi and reaching the lower Yagi undergoes a 180-degree phase shift due to the length of the $\lambda/2$ line, plus a 180-degree phase shift due to the line's being twisted, e.g., its two conductors are crossed once.

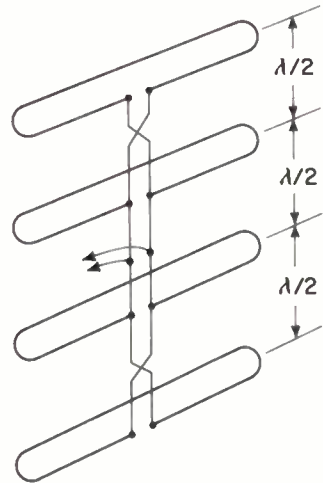


Figure 3-34. Superposing four antennas.

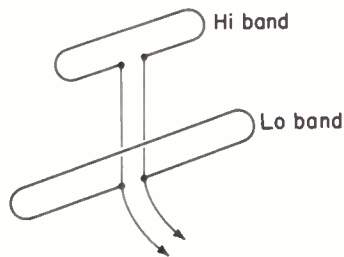


Figure 3-35. Combining antennas for two frequency bands.

When four Yagis are to be superposed, the two preceding arrangements may be combined. They result in the array shown in Fig. 3-34, which has a gain of 6 db with respect to a single Yagi and a rather sharp vertical directivity.

A variant of the superposition principle is found when two different Yagis are combined, for example, one for Channels 1 to 6 and one for Channels 7 to 13 as shown in Fig. 3-35. A common lead-in may be used as shown. The interaction between the two Yagis will be minimum if the spacing is equal to half the resonant wavelength of the low band Yagi.

3-22. Miscellaneous antennas

Although Yagis are the most common type of antennas, at least for the VHF bands, a large number of antenna designs have been proposed and a few familiar arrangements will be rapidly described. A number of them are popular for the UHF band, either because of wide bandwidth or of some other desirable characteristic.

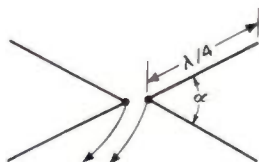


Figure 3-36. Double V antenna.

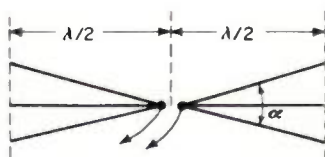


Figure 3-37. Fan antenna.

The double V antenna shown in Fig. 3-36 may be considered as the result of paralleling two divergent dipoles. It has four members $\lambda/4$ long. The divergence angle α is on the order of 30–45 degrees and is not critical. The impedance varies between 150 and 200 ohms according to the divergence angle.

The fan antenna of Fig. 3-37 is an extension of the double V principle. The impedance varies from 300–350 ohms when the divergence angle α varies from 30 to 45 degrees. The total length is $\lambda/2$ as shown. A variant is illustrated in Fig. 3-38; it has a total length of λ , and its impedance is double that of the shorter fan. That is, it varies between 600 and 700 ohms when the divergence angle α varies between 30 and 45 degrees. The gain is greater for the longer fan.

The conical antenna derives from the double V through revolution around the axis as shown in Fig. 3-39. Each cone has a generatrix 0.365λ long, and the divergence angle α determines the impedance. Some representative values are 1000 ohms for 10 degrees, 600 ohms for 20 degrees, and 300 ohms for 30 degrees. The conical antenna has a wide band, on the order of 30 per cent of the resonant frequency. Mechanically, the cones may be sheet metal, gauze, or wires as shown to reduce wind surface.

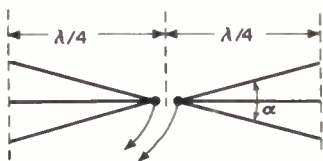


Figure 3-38. Long fan antenna.

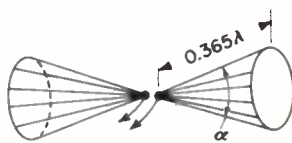


Figure 3-39. Double conical antenna.

The butterfly antenna of Fig. 3-40 may also be made of metal sheet or gauze. The dimensions are indicated on the figure. As such, this antenna is *bidirectional*, which means that it is equally sensitive to waves coming from the back as to waves coming from the front. This can be easily remedied with

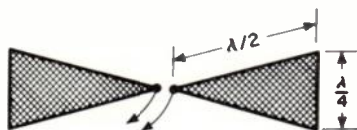


Figure 3-40. Butterfly or bow-tie antenna.

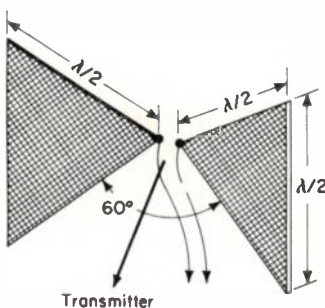


Figure 3-41. Angular butterfly antenna.

the variant of Fig. 3-41. Each half of the antenna is an equilateral triangle $\lambda/2$ on each side, and the two halves converge at an angle of 60 degrees, whose bisectrix should be directed toward the transmitter. The butterfly antenna has an interesting peculiarity; its gain increases in almost direct proportion to the received frequency above the resonant frequency.

The corner antenna of Fig. 3-42 uses a number of reflectors arranged in a corner array which can be considered as a rough approximation of a parabolic reflector. The spacing between reflectors is small with respect to the wavelength, say 0.1λ , so that for all practical purposes the array of reflectors behaves as a continuous sheet. The active element is a simple or folded dipole placed in the axis of the reflector and $\lambda/2$ in front of the apex. The bandwidth is not very large, and the gain is on the order of 10 db with respect to the standard dipole. The forward sensitivity exceeds the backward sensitivity by some 35 db. Secondary lobes are practically negligible.

The individual reflectors in the array may be either insulated from each other or connected together by their centers, which are neutral points.

The helix antenna of Fig. 3-43 is somewhat more difficult to design mechanically, and its length of 1.5λ limits its use to high frequencies unless it becomes cumbersome. On the other hand, it has a wide bandwidth and important gain. Representative dimensions are indicated on the drawing and are not critical.

The ground plane behind the helix may be made of metal sheet, gauze, or wire. It is of 0.8λ on each side if square, and λ in diameter if circular. The helix is fixed at its center by an insulating post which may be part of the connecting arrangement. It is supported by insulating rods.

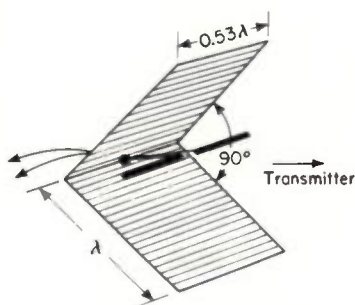


Figure 3-42. Corner antenna.

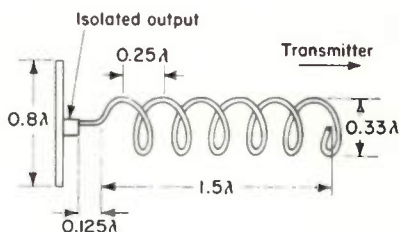


Figure 3-43. Helical antenna.

The best lead-in is a coaxial cable whose inner conductor is connected to the helix at the insulating center post and whose outer conductor is connected to the ground plane.

The impedance of the helix antenna is on the order of 125 ohms.

3-23. Rhombic antenna

When the received field intensity is very weak, either because of an unfavorable geographical position or because the transmitter is too far away, it may become

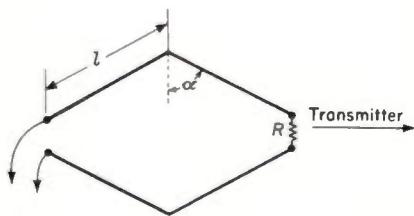


Figure 3-44. Rhombic antenna.

interesting to consider long antennas. The common types are the rhombic and the V antennas. Both have a marked directivity and a gain which may be 20 to 40 times the gain of an ordinary dipole.

The rhombic antenna of Fig. 3-44 is a nonresonant long antenna which is suitable when maximum gain is wanted in a given direction and when erecting space is available. The rhom-

bic is easily put up only for horizontal polarization, in which case the antenna lies in the horizontal plane.

The three parameters governing the performance are the opening angle α , the side length l , and the terminating resistance R .

As with all long antennas, the rhombic provides good results only if it is really long with respect to the wavelength. A side length of 3λ constitutes a minimum; the gain increases with l and may reach 15 db. It is consequently recommended to install the longest possible rhombic. If possible, l should be an odd number of quarter wavelengths to provide a maximum front-to-back ratio. This parameter is the ratio of sensitivity in the forward direction to the sensitivity in the backward direction. If the terminating resistance is correctly matched, the front-to-back ratio may theoretically reach infinity. In other words, the rhombic has zero sensitivity to waves arriving from the back.

Another factor which influences the length of the sides indirectly is the angle of arrival of the signal above the horizontal. In all practical cases this angle is very small and the opening angle α should be large, say between 80 and 90 degrees. Theoretically, the side length should then be large—up to 15λ for best results. Shorter lengths still ensure good performance but at reduced gain.

The rhombic preferably should be installed above flat ground, well in the clear, at a height of at least 15λ .

The terminating resistance R largely determines the front-to-back ratio as has been said, and also the aperiodicity of the antenna. Good adaptation may be secured by adjusting the resistance. An 800-ohm resistor is suitable if it has negligible inductance and capacitance at the working frequency.

The directivity is rather sharp, particularly for a long rhombic as shown in Fig. 3-45, and the diagram exhibits secondary lobes of small magnitude. Adjusting the value of the terminating resistance modifies the secondary lobes' distribution. This effect can be put to good use to reduce the side sensitivity in a given direction—for example, to eliminate a bothersome reflected wave.

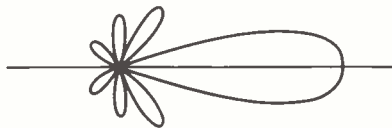


Figure 3-45. Rhombic antenna directivity diagram.

The rhombic impedance is on the order of 700 to 800 ohms. It has a very wide band, approximately one octave wide.

Two rhombics can be stacked one above the other in the same fashion as Yagis to improve the performance and eventually facilitate feeder matching.

3-24. V antenna

The V antenna is a nonresonant long antenna as shown in Fig. 3-46. The V lies in the horizontal plane for horizontal polarization. The parameters determining the performance are the side length l and the opening angle α . The gain increases with l .

The opening angle determines the directivity. It is chosen for maximum sensitivity along the axis of the V and should decrease when the length l increases.

When l varies from 4λ to 12λ , the angle α varies almost linearly from 50 to 30 degrees, and the gain with respect to a dipole varies almost linearly from 8 to 12 db.

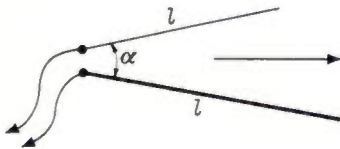


Figure 3-46. V antenna.

The V antenna can also be used as a resonant antenna by choosing l equal to a multiple of the half wavelength. It then becomes bidirectional, with equal sensitivity in both directions along the axis of the V. With an l of 8λ and an opening angle of 35 degrees, the gain reaches 10 db with

respect to the dipole. The resonant V can be made unidirectional by adding a second V antenna a quarter wavelength behind the first one. The second V fulfills the same function as a reflector in a Yagi. It also increases the gain.

V antennas can be stacked with resulting increased vertical directivity.

The directional diagram and the elevation of the main lobe can be adjusted between large limits by varying the opening angle and the height above ground.

3-25. Multiple antenna systems

In a location where only a limited number of stations can be received with good results, the best system installation is to use one antenna per channel. Each antenna can be tuned and oriented to provide optimum results.

If a number of stations can be received, economics dictates the use of a combination of wideband antennas, say one or sometimes two per band, mounted on a rotating mast so that the antenna can be oriented in any desired direction for best reception. A small electric motor rotates the antenna, and a remote control box is placed near the receiver.

A typical arrangement, generally of a Yagi type, includes one antenna for Channels 1 to 4 and another antenna for Channels 5 to 13. A common lead-in cable is used, as previously indicated. If some UHF station is available, the installation also includes an antenna tuned to the corresponding part of the UHF band; a fan or butterfly antenna is commonly used. A separate lead-in is generally provided, since in most cases UHF channels are received through separate r-f amplifier and mixing stages.

3-26. Antenna distribution

A single antenna may be used with a number of receivers. To avoid interaction, a sufficient amount of attenuation should be provided between any two receivers; 20 db constitutes a minimum.

If the received signal is strong enough, there is no objection to losing some amplitude, and simple passive distributors made of resistors can be used. Several arrangements are possible; Fig. 3-47 shows a common circuit. Each receiver is fed through a three-resistor combination. The values of the series R_s and parallel R_p resistors depend on the desired impedance and on the number of receivers. They are not highly critical, and the nearest standard values are used instead of the exact calculated values.

For the standard impedance of 300 ohms, Table 3-2 indicates common values for two to eight receivers connected to the same antenna system.

The resulting impedance match is satisfactory for practical purposes. The attenuation between any two receivers is not very high but in most cases it is sufficient to avoid troublesome local oscillator interference.

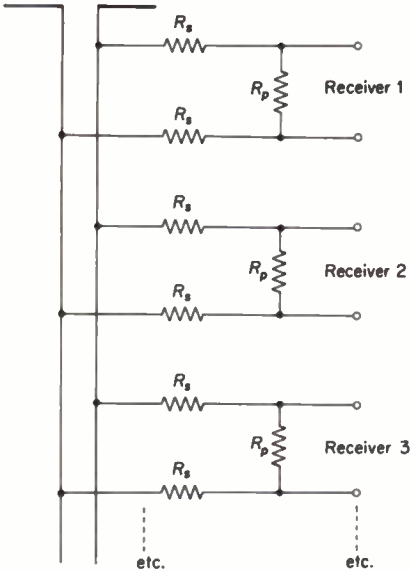


Figure 3-47. Passive antenna distributor.

If the received signal is weak and the attenuation of a passive distributor cannot be tolerated, it is possible to insert a preamplifier or booster before the distributor to bring the signal to the required level.

Table 3-2. Attenuator values

Number of receivers	Series R_s	Parallel R_p
2	220	400
3	400	400
4	480	330
5	600	330
6	720	330
7	960	330
8	1100	330

Another possibility is indicated schematically in Fig. 3-48. The antenna is connected to the grids of cathode-follower stages. Each cathode provides an output. The low cathode impedance is adjusted to provide a good match to the connecting cable. It is worth noting that a cathode-follower stage does not amplify but on the contrary introduces a certain amount of attenuation. The received signal should then be sufficiently strong to provide the necessary voltage level after the cathode follower.

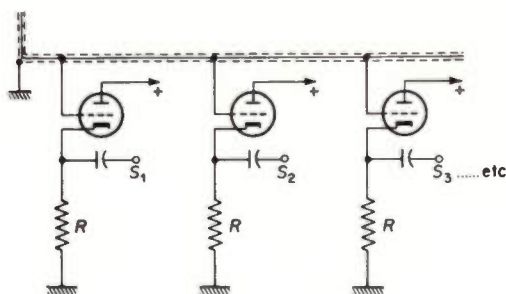


Figure 3-48. Antenna distributor using cathode followers.

3-27. Possibility of reception

The possibility of reception at any given point depends essentially on the ratio between the received signal level at the input of the receiver and the level of noise produced by the receiver. This signal-to-noise ratio S/N determines the quality of the picture.

A better way to express this is to use the noise factor F , because its value is independent of the conditions of measurement. Briefly, the noise factor expresses how much (in power) more noisy the receiver is than an ideal receiver which has the antenna resistance as the only source of noise. More precisely, the noise factor is given by

$$F = \frac{F_1}{F_2}$$

with

$$F_1 = \frac{\text{input signal power}}{\text{input noise power}} \quad \text{and} \quad F_2 = \frac{\text{output signal power}}{\text{output noise power}}$$

F is evidently equal to 1 for the ideal receiver with no internal noise so that the lower F , the better the actual receiver. High-quality receivers may have a noise factor on the order of 10. When the noise factor is known, the useful sensitivity of the receiver can be calculated. All that is necessary is to calculate the noise voltage V_n referred to the input, which is

$$V_n = (4 \times 10^{-21} \times R_a \times B \times F)^{1/2}$$

where R_a = antenna impedance,

B = bandwidth,

F = noise factor.

Using the standard impedance value of 300 ohms and a bandwidth of 4 mc, the noise voltage in microvolts can be established as

$$V_n = 2.2F^{1/2}$$

From this equation the minimum signal level at the input may be deduced. A signal-to-noise ratio of 20 db, or 10 times, is considered as a minimum for good quality. The signal level V_s in microvolts at the input should then be at least

$$V_s = 22F^{1/2}$$

Thus, for a signal-to-noise ratio of 20 db, a high-performance receiver having a noise factor of 9 would require a minimum signal of 66 microvolts at its input.

3-28. Attenuation

Excessive signal intensity may also be a drawback. It overloads the receiver and can lead to excessive contrast in the picture, synchronization instability, sound distortion, or cross modulation. The simplest way to reduce the intensity of a signal is to insert an attenuator between the antenna lead-in and the receiver input.

For the standard 300-ohm line, the attenuator is a symmetrical arrangement of five resistances as shown in Fig. 3-49. It can be wired conveniently on a small piece of insulating material. The required attenuation varies with the strength of the signal. Table 3-3 gives the values in ohms of the four series resistances R_s and of the parallel resistance R_p for the usual values of 10, 20, and 30 db.

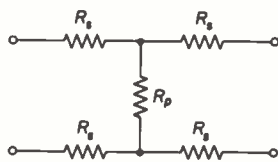


Figure 3-49. Symmetrical attenuator.

Table 3-3. Attenuator values

Decibels	R_s	R_p
10	82	220
20	120	62
30	150	22

Several attenuators can be cascaded to obtain higher attenuations.

Problems

1. Explain the difference between *radio horizon* and *visual horizon*.
2. A transmitting antenna is installed at the top of a building 300 m high. Calculate its optical and radioelectrical radius of coverage.
3. A transmitting antenna is at an altitude of 220 m. A receiving antenna is at an altitude of 18 m. Calculate the maximum allowable distance between them for direct radio propagation.

4. A station has an ERP of 10 kw. Calculate the theoretical electric field intensity at 10 and 35 miles away.
5. A transmitting antenna at a height of 120 m radiates a power of 30 kw on a frequency of 60 mc. A receiving antenna 65 miles away is at a height of 18 m. Calculate the field at the receiving antenna.
6. The straight line joining a transmitter and a receiver is 20 km long. On the perpendicular to this straight line at its center and two kilometers away from the straight line lies an obstruction which gives rise to a reflected wave at the receiver. The picture tube is 32 cm wide. Evaluate the distance between the main picture and the ghost due to reflection.
7. A station has an ERP of 32 kw on a frequency of 200 mc. A receiving dipole aerial is 42 miles away. Evaluate the voltage at the antenna terminals.
8. Compare ribbon lines and coaxial cables as lead-ins for television reception.
9. Calculate line losses for wet and dry conditions at 50 mc and 200 mc for 300-ohm ribbon and 300-ohm tubular line, and for RG59/U and RG11/U coaxial cables.
10. A lead-in line 35 ft long is mismatched. The picture tube is 32 cm wide. Calculate the distance between echos.
11. Explain why a coaxial cable should not be used as a symmetrical line.
12. Explain the advantages of grounding the center tap of the primary winding of a symmetrical antenna input.
13. Discuss the advantages of symmetrical lead-in line vs. coaxial nonsymmetrical cable.
14. Explain the necessity of matching the lead-in line to the receiver input.
15. Discuss the advantages and drawbacks of Yagi antennas.
16. Compare VHF and UHF antennas.
17. Compare a folded dipole and a single dipole.
18. Discuss the advantages and disadvantages of a rhombic antenna with respect to a Yagi antenna.
19. A receiver has a noise factor of 18, an input impedance of 300 ohms, and a bandwidth of 4.5 mc. Calculate the input noise voltage.
20. A receiver has a noise factor of 22, an input impedance of 300 ohms, and a bandwidth of 3.5 mc. Evaluate the signal voltage for signal-to-noise ratios of 10 db and 30 db.

Chapter 4

I-F AND R-F WIDEBAND CIRCUITS

Television, along with radar, is a typical example of wideband technique. Although the fundamentals of tuning and resonance are as good here as they are everywhere in electronics, it is nevertheless true that special circuits and special tubes are needed to extract usable gain from a stage having a bandwidth of several megacycles. Interstage coupling circuits are especially important in this respect.

4-1. Introduction

All the r-f or i-f tuned circuits of a picture receiver, and to a lesser extent the sound receiver, are wideband circuits and as such they rely on a specific technique. Whether in r-f or i-f stages, four main types of coupling circuits are commonly used. They are:

1. the antiresonant parallel circuit shown in Fig. 4-1;
2. the untuned primary transformer shown in Fig. 4-2;
3. the double-tuned transformer shown in Fig. 4-3;
4. the complex filter circuits shown in Fig. 4-4.

On the first three figures notice the parallel resistances shunting the tuned circuits. They are used to reduce the selectivity, that is, to increase the bandwidth.

A number of variations of these fundamental circuits can be and sometimes are used. Also, practical arrangements may differ in detail. Nevertheless, the four fundamental circuits indicated are by far the most common and justify individual theoretical study.

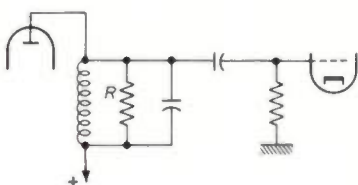


Figure 4-1. Antiresonant coupling circuit.

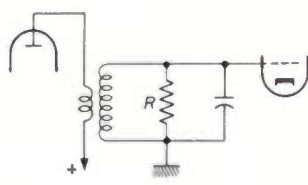


Figure 4-2. Untuned primary transformer coupling.

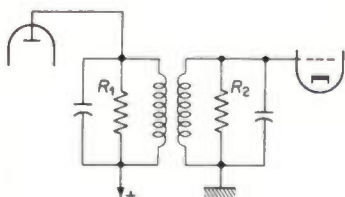


Figure 4-3. Double tuned transformer coupling.

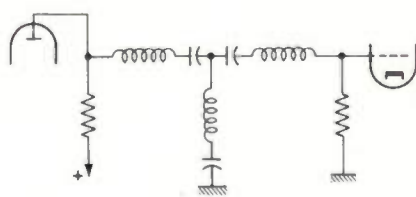


Figure 4-4. Filter type of coupling.

4-2. Antiresonant circuit

The equivalent circuit of Fig. 4-1 given in Fig. 4-5 will be used for calculations. The tube is represented by a current generator i and its plate resistance r_p .

The load impedance Z can be considered as a parallel combination of resistance R and the impedance X of the LC circuit. Hence,

$$Z = \frac{RX}{R + X} = \frac{R}{1 + R/X}$$

Let e_1 be the input voltage applied to the grid of the tube. Then

$$i = g_m e_1$$

where g_m is the mutual conductance of the tube.

Let e_2 be the output voltage across the anode load Z , so that

$$e_2 = iZ = g_m Z e_1$$

The gain of the stage is

$$K = \frac{e_2}{e_1} = g_m Z$$

Two implicit assumptions have been made to simplify calculations: the ohmic resistance of the inductance L is assumed to be negligible; the plate

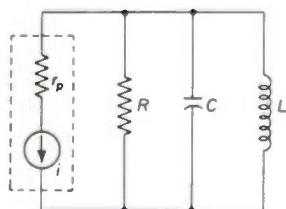


Figure 4-5. Equivalent circuit for the antiresonant circuit coupling.

resistance r_p of the tube is assumed to be much larger than R . These two assumptions are almost always valid in practice.

At resonance, that is, for

$$\omega_0 = (LC)^{-1/2}$$

where ω_0 is the resonant angular velocity or resonant pulsance, the impedance X of the LC combination becomes infinite so that

$$Z = R$$

and the gain at resonance is

$$K_0 = g_m R$$

Near resonance, X is no longer infinite, and the gain decreases to a value

$$K = g_m Z = g_m \frac{R}{1 + R/X}$$

K is smaller than K_0 by a factor A :

$$A = \frac{K_0}{K} = 1 + \frac{R}{X}$$

This expression is always larger than 1 and is called the *relative attenuation*. Its value can be calculated as follows, in terms of the circuit parameters.

The impedance X of the LC circuit results from the paralleling of the inductive reactance $j\omega L$ and the capacitive reactance $1/j\omega C$:

$$\frac{1}{X} = j\omega C + \frac{1}{j\omega L} = -j \left(\frac{1 - LC\omega^2}{L\omega} \right)$$

Consequently,

$$A = 1 + \frac{R}{X} = 1 - jR \left(\frac{1 - LC\omega^2}{L\omega} \right)$$

Squaring to eliminate the imaginary part, we obtain

$$A^2 = 1 + R^2 \left(\frac{1 - LC\omega^2}{L\omega} \right)^2$$

or

$$A = \left[1 + R^2 \left(\frac{1 - LC\omega^2}{L\omega} \right)^2 \right]^{1/2}$$

As a check, notice that at resonance $LC\omega_0^2 = 1$, and A becomes equal to 1 as it should. Near resonance, say at frequency $(\omega_0 + \Delta\omega)$, A becomes larger than 1. The angular velocity ω appears only in the term $(1 - LC\omega^2)/L\omega$ in the expression for A , so that this term will be studied separately.

At the angular velocity $(\omega_0 + \Delta\omega)$, this term becomes

$$\frac{1 - LC(\omega_0 + \Delta\omega)^2}{L(\omega_0 + \Delta\omega)}$$

Developing, neglecting a second-order term in $\Delta\omega^2$ and using $LC\omega_0^2 = 1$, provides

$$-\frac{2LC\omega_0\Delta\omega}{L(\omega_0 + \Delta\omega)}$$

If $\Delta\omega$ is small enough with respect to ω_0 , it can be neglected in the denominator, and the term simplifies to $-2C\Delta\omega$. The expression for A is then

$$A = [1 + (2RC\Delta\omega)^2]^{1/2}$$

or, also, $2RC\Delta\omega = (A^2 - 1)^{1/2}$

Replacing $\Delta\omega$ by $2\pi\Delta f$, we have

$$4\pi RC\Delta f = (A^2 - 1)^{1/2}$$

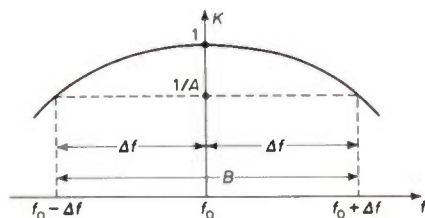


Figure 4-6. Bandwidth of the coupling circuit.

attenuation A is the same (Fig. 4-6). The bandwidth B between these two frequencies is

$$B = 2\Delta f$$

which can be used in the preceding expression to give

$$2\pi BCR = (A^2 - 1)^{1/2}$$

or

$$BCR = 0.159(A^2 - 1)^{1/2}$$

where B = bandwidth in cycles per second,

C = capacitance in farads,

R = resistance in ohms,

A = attenuation.

An easier expression for practical use is

$$BCR = 159(A^2 - 1)^{1/2}$$

with B in megacycles, C in micromicrofarads, and R in kilohms. For a 3 db bandwidth B_1 , the attenuation A is 1.41 so that

$$B_1CR = 159$$

The basic expression can be used to find any of the parameters if the others are known. It also provides the inductance value through the LC product if the resonant frequency is known.

4-3. Stage gain

The value of R is then

$$R = \frac{159}{BC}(A^2 - 1)^{1/2}$$

The gain at resonance is

$$K_0 = g_m R = 159 \frac{g_m}{BC} (A^2 - 1)^{1/2}$$

with B in megacycles, C in micromicrofarads, and g_m in milliamperes per volt. This expression shows that, for a given bandwidth at given attenuation, the gain is proportional to the mutual conductance g_m and inversely proportional to the total shunt capacitance C across the circuit—hence, our interest in a minimum shunt capacitance.

This capacitance includes:

1. any capacitor connected across the inductance;
2. the distributed capacitance of the coil;
3. the output or plate capacitance of the preceding tube;
4. the input or grid capacitance of the following tube;
5. the parasitic wiring capacitance.

Item 1 can be suppressed by eliminating all physical capacitors and tuning the circuit by adjusting the inductance value by means of an iron or brass core.

Item 2 can be reduced by special winding of the coil. At the r-f or i-f frequencies used, it is generally small.

Items 3 and 4 constitute the main part of the total capacitance; the only way to reduce them is to choose tubes which have the lowest input and output capacitances.

Item 5 can be somewhat reduced by careful wiring.

4-4. Special tubes

As far as tube selection is concerned, the best type would have maximum transconductance g_m and minimum input C_i and output C_o capacitances. This is summed up in what is called the *merit coefficient* M of the tube.

$$M = \frac{g_m}{C_i + C_o}$$

For wideband amplification use, the larger the M , the better the tube.

The merit coefficient should be used with some caution when comparing different tubes—for two reasons: First, it compares tubes in themselves, but not in circuit. When the other capacitances are taken into account, the results may be modified. Second, it disregards Miller effect in triodes, which increases the input capacitance because of the plate-to-grid capacitance. It is true that, except when special circuits are used, pentodes rather than triodes are usually found in i-f amplifiers, in which case Miller effect is negligible. Nonetheless, when triodes are used, as they frequently are in the r-f input stages, the results are misleading.

To indicate more precisely the necessity for special tubes having high transconductance and low capacitances, the gain will be evaluated for two tubes.

The first tube is a standard pentode used in radio receivers, the 6SK7. It has an input capacitance C_i of $4.6 \mu\mu\text{f}$ and an output capacitance C_o of $12 \mu\mu\text{f}$. Its transconductance g_m is 1.5 ma/v . Let the bandwidth B be 4.5 mc at 3 db attenuation, which corresponds to $A = 1.414$. Assume a wiring parasitic capacitance of $8.4 \mu\mu\text{f}$. The total shunt capacitance is

$$C = 4.6 + 12 + 8.4 = 25 \mu\mu\text{f}$$

Then

$$R = \frac{159}{4.5 \times 25} (2 - 1)^{1/2}$$

or

$$R = 1.43 \text{ kilohms}$$

The gain is

$$K_0 = g_m R = 1.5 \times 1.43 = 2.14$$

for the stage, which is evidently extremely low and of little practical interest. To obtain any sensible amount of over-all gain, say 8000, as many as 13 stages would be necessary, with drastic reduction in bandwidth.

The second example will use a standard television pentode, the 6CB6, which has input and output capacitances of 6.5 and $2 \mu\mu\text{f}$ respectively and a transconductance of 8 ma/v . Assume that careful wiring has reduced the parasitic capacitance to $4.5 \mu\mu\text{f}$. The total shunt capacitance is

$$C = 6.5 + 2 + 4.5 = 13 \mu\mu\text{f}$$

For the same bandwidth of 4.5 mc and the same attenuation of 1.41 , the resistance is

$$R = \frac{159}{4.5 \times 13} = 2.74 \text{ kilohms}$$

and the gain is

$$K_0 = g_m R = 8 \times 2.74 = 21$$

approximately.

Three stages would be sufficient to ensure a total gain of 8000.

Since the gain per stage is $K_0 = g_m R$ and the 3 db bandwidth $B_1 = 159/CR$ in practical units, the gain-bandwidth product is

$$K_0 B_1 = \frac{159 g_m}{C}$$

with B_1 in megacycles, g_m in milliamperes per volt, and C in micromicrofarads.

It can be seen that the gain-bandwidth product is a constant for given g_m and C . When the tube and the stage design have been chosen, g_m and C are fixed and gain can be traded for bandwidth and vice versa, as long as their product remains constant.

4-5. Input impedance of a tube

At high frequencies, the input impedance of a tube is composed of a resistance R_i in parallel with a capacitance C_i . This capacitance essentially consists of the grid-to-cathode capacitance C_{gk} plus the grid-to-anode capacitance C_{ga} reflected at the input by Miller effect, so that

$$C_i = C_{gk} + (K + 1)C_{ga}$$

where K is the stage gain.

For pentodes, C_{ga} is very small and Miller effect can be neglected. However, this is not the case with triodes.

The input resistance R_i can be safely ignored when the frequency is below a few megacycles. For the frequencies used in television it becomes an important factor. It is usually given by the tube manufacturer as a value R_a for a given frequency f_a . Since it varies as the reciprocal of the square of the frequency, its value R_b at a different frequency f_b is given by

$$R_b = R_a \left(\frac{f_a}{f_b} \right)^2$$

4-6. Transformer

The transformer is frequently used at the input of a receiver, as shown in Fig. 4-7, to couple the antenna to the first amplifying stage. The equivalent autotransformer shown in Fig. 4-8 is also common.

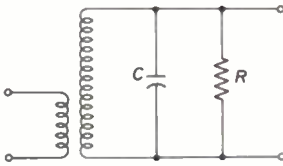


Figure 4-7. Untuned primary input transformer.

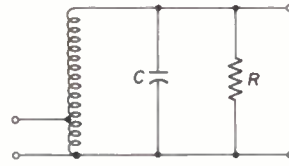


Figure 4-8. Input autotransformer.

A high coupling coefficient k between primary and secondary is desirable. With the transformer it is difficult to reach high values, even if primary and secondary are wound one above the other. High coupling is easier with an autotransformer, especially if the winding is long and has a magnetic core.

Let the mutual inductance between primary and secondary be M and assume for simplicity that the primary impedance is negligible with respect to the secondary impedance. Since Z_2 is the secondary impedance, it appears reflected in the primary as an impedance

$$Z_1 = \frac{M^2 \omega^2}{Z_2}$$

A voltage e_1 across the primary creates a primary current

$$i_1 = \frac{e_1}{Z_1} = \frac{e_1 Z_2}{M^2 \omega^2}$$

The induced secondary voltage e_2 is

$$e_2 = M \omega i_1 = \frac{e_1 Z_2}{M \omega}$$

The voltage gain K of the transformer is

$$K = \frac{e_2}{e_1} = \frac{Z_2}{M \omega} = \left(\frac{Z_2}{Z_1} \right)^{1/2}$$

4-7. Transformer bandwidth

The voltage across the secondary varies as a function of the frequency f or of the angular velocity $\omega = 2\pi f$. It is maximum at the resonant frequency f_0 , and decreases as the frequency moves away from the resonant frequency.

The classical selectivity formula

$$s = \left[1 + \left(\frac{2Q \Delta f}{f_0} \right)^2 \right]^{1/2}$$

gives the relative attenuation s as a function of the quality factor Q and of the distance Δf from the resonant frequency f_0 .

For a 3-db attenuation at distance Δf_1 ,

$$s = 1.414 = 2^{1/2}$$

and consequently

$$\left(\frac{2Q \Delta f_1}{f_0} \right)^2 = 1$$

or

$$f_0 = 2Q \Delta f_1$$

Since the 3-db bandwidth is $B_1 = 2 \Delta f_1$ as before, then

$$f_0 = QB_1$$

The secondary impedance is

$$Z_2 = \frac{Q}{\omega_0 C} \quad \text{or} \quad Z_2 = \frac{1}{2\pi B_1 C}$$

and since $K = (Z_2/Z_1)^{1/2}$, it follows that

$$K = \frac{1}{(Z_1 \cdot 2\pi B_1 C)^{1/2}}$$

Now, to avoid unwanted reflections at the input, the impedance Z_1 at the primary should match the lead-in characteristic impedance. Stated in another way, Z_1 is determined by factors external to the transformer. It can then be seen that the gain of the circuit for given bandwidth B_1 is limited by the shunt capacitance C as it was for the antiresonant circuit.

There is, however, an improvement. In the transformer or autotransformer the gain is inversely proportional to the square root of the shunt capacitance, whereas for the antiresonant circuit it is inversely proportional to the shunt capacitance.

Nevertheless, it still appears necessary for high gain to reduce C as far as possible, and the same factors will influence the choice of a suitable tube—namely low input and output capacitances and high transconductance.

4-8. Double-tuned transformer

When a double-tuned transformer is used as an interstage coupling circuit, the output capacitance of the preceding stage appears across the primary, whereas the input capacitance of the following stage appears across the secondary.

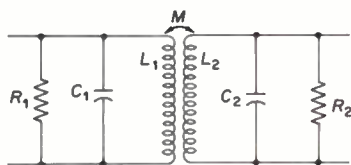


Figure 4-9. Components in a double-tuned transformer.

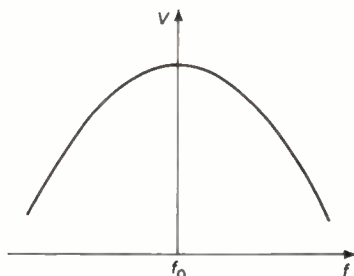


Figure 4-10. One-maximum response curve.

The total shunt capacitance is thus split into two parts, and it can be intuitively expected that the gain should be greater than in the case of a single-tuned circuit since we have seen that the gain is limited by the shunt capacitance.

The equivalent circuit is shown in Fig. 4-9. It is well known that the coupling coefficient k between primary and secondary determines the over-all shape of the response curve. At low values of k , the response curve, shown in Fig. 4-10, exhibits a single maximum and is reminiscent of the response of a single-tuned circuit. At high k , the response curve takes the distinctive shape of Fig. 4-11, with two maxima separated by a trough.

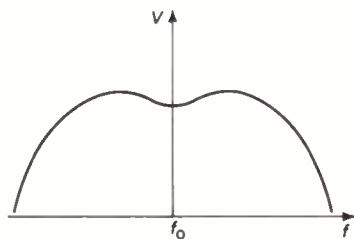


Figure 4-11. Two-maxima response curve.

The maximum value of k for which there exists only one maximum is the transitional coupling k_t . The value of k which provides maximum gain for given primary and secondary Q is the critical coupling k_c .

The transitional coupling k_t and the critical coupling k_c coincide only when the primary quality factor Q_1 and the secondary quality factor Q_2 are equal. In all other cases, k_t is larger than k_c . Therefore, it is evident that the shape of the response curve depends on the coupling coefficient and on the primary and secondary Q 's.

4-9. Attenuation curve

Instead of the *gain* of the transformer, we can speak of the *attenuation*, which is the reciprocal of the gain. This results in some simplification in the calculations. Moreover, the attenuation at resonance is taken equal to 1 so that all attenuation curves can be shown in normalized form for easy comparison. This can further be facilitated by using an auxiliary variable x which combines all the variables in the problem, that is, the frequency f , the coupling coefficient k , the primary Q_1 , and the secondary Q_2 . The intermediary steps in the derivation of x will be indicated to illustrate the method used.

First, define a relative frequency spacing d as the ratio of the frequency displacement Δf off resonance to the resonant frequency f_0 :

$$d = \frac{\Delta f}{f_0}$$

so that at resonance

$$d = 0$$

Now, define two parameters,

$$m = k^2 + \frac{1}{Q_1 Q_2} \quad \text{and} \quad n = \left(\frac{1}{Q_1} + \frac{1}{Q_2} \right)^2$$

The auxiliary variable is then

$$x = \frac{2d}{m^{1/2}}$$

and the classical expression for the attenuation of the transformer can be written as

$$A = \left[1 - 2 \left(1 - \frac{n}{2m} \right) x^2 + x^4 \right]^{1/2}$$

This expression is suited to the calculation of universal curves, the only parameter being $n/2m$. A family of such curves is shown in Fig. 4-12, which represents half the total response curve of the transformer. The other half is symmetrical with respect to the resonant frequency, for which $x = 0$. It gives the attenuation A as a function of x for various values of $n/2m$.

The gain can be calculated for any frequency for any curve, starting with the gain K_0 at resonance, which is

$$K_0 = g_m \frac{k}{m} \left(\frac{R_1 R_2}{Q_1 Q_2} \right)^{1/2}$$

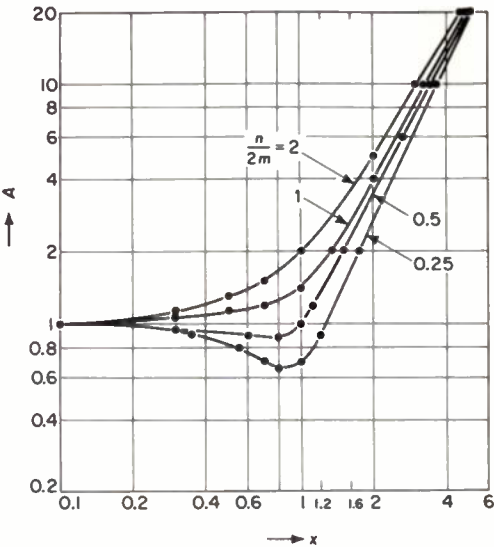


Figure 4-12. Universal attenuation curves.

where g_m = tube transconductance,
 k = coupling coefficient,
 $m = k^2 + 1/Q_1Q_2$,
 Q_1 = primary Q ,
 Q_2 = secondary Q ,
 R_1 = primary shunt resistance,
 R_2 = secondary shunt resistance.

4-10. Coupling coefficient

It is apparent, as shown in Fig. 4-12, that all curves having

$$\frac{n}{2m} < 1$$

exhibit a maximum attenuation smaller than 1 for some frequency besides resonant frequency. In other words, the corresponding response curves exhibit two maxima.

This can be checked by referring to the attenuation equation where x^4 , being an even power, is always positive. The only possibility of A 's being smaller than 1 occurs when the coefficient of x^2 , that is, $-2(1 - n/2m)$, is negative. This will occur only for

$$\frac{n}{2m} < 1$$

which checks with the given curves.

4-11. Transitional coupling

Transitional coupling k_t , then, corresponds to $n/2m = 1$, or, substituting for n and m ,

$$\left(\frac{1}{Q_1} + \frac{1}{Q_2}\right)^2 = 2\left(k_t^2 + \frac{1}{Q_1 Q_2}\right)$$

and solving for k_t ,

$$k_t = \left[\frac{1}{2}\left(\frac{1}{Q_1^2} + \frac{1}{Q_2^2}\right)\right]^{1/2}$$

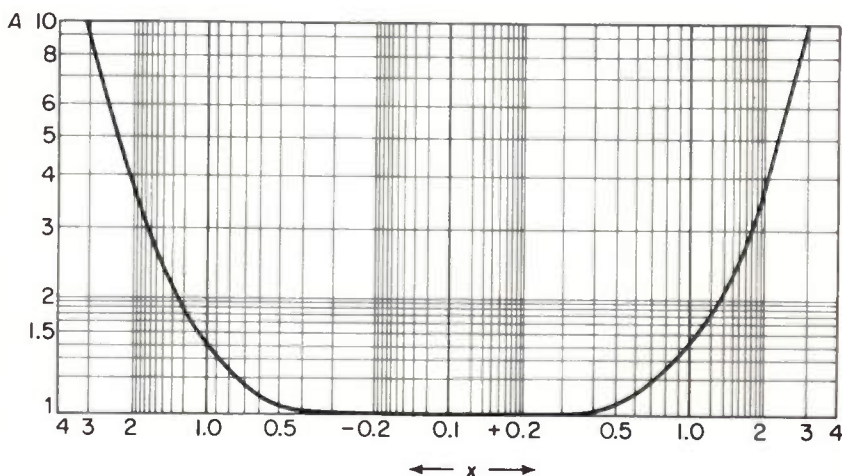


Figure 4-13. Attenuation curve for transitional coupling.

This value of k provides the flattest possible response curve. The attenuation equation reduces to

$$A = (1 + x^4)^{1/2}$$

and the corresponding response curve is given in Fig. 4-13.

Let x_1 be the value of x for which the attenuation is 3 db, that is, $A = 1.414$. The preceding expression gives $x_1 = \pm 1$. Since the corresponding relative frequency displacement is d_1 , it follows that

$$x_1 = \frac{2d_1}{m^{1/2}} = 1$$

or

$$d_1 = 0.5m^{1/2}$$

m can be calculated using the transitional value for k ; the result is

$$d_1 = \frac{1}{8^{1/2}} \left(\frac{1}{Q_1} + \frac{1}{Q_2} \right)$$

and since $d_1 = \Delta f_1/f_0$, we obtain

$$\Delta f_1 = 0.35f_0 \left(\frac{1}{Q_1} + \frac{1}{Q_2} \right)$$

The 3-db bandwidth B_1 is

$$B_1 = 2 \Delta f_1 = 0.71f_0 \left(\frac{1}{Q_1} + \frac{1}{Q_2} \right)$$

Since the primary and secondary Q 's are given, it is easy to evaluate the 3-db bandwidth of a transitionally coupled transformer. Referring to the curve, it is also easy to evaluate the bandwidth for any given attenuation or the converse.

4-12. One large Q

The bandwidth expression gives evidence of the effect of the primary and secondary Q 's. If Q_1 , say, is much larger than Q_2 , for example, because all the damping is concentrated on the secondary, the expression reduces to

$$B_1 = \frac{0.71f_0}{Q_2}$$

Under the same assumption, the transitional coupling becomes

$$k_t = \frac{0.71}{Q_2}$$

and the remarkable expression

$$B_1 = f_0 k_t$$

results, in which B_1 and f_0 should be expressed in the same units.

For example, an i-f transformer using transitional coupling, tuned on 30 mc and having a bandwidth of 4.5 mc at 3 db, in which only the secondary would be damped, should have a coupling coefficient

$$k_t = 4.5/30 = 0.15$$

to obtain a maximally flat response curve.

4-13. Practical design, one Q infinite

The practical design of a transformer having

$$Q_1 \gg Q_2$$

is straightforward.

The secondary quality factor is

$$Q_2 = 2\pi f_0 R_2 C_2$$

where R_2 is the secondary parallel total damping resistance, and C_2 the total secondary shunt capacitance. Since $k_t Q_2 = 0.71$, it follows that

$$2\pi f_0 R_2 C_2 k_t = 0.71$$

or

$$2\pi R_2 C_2 B_1 = 0.71$$

where R_2 is in ohms, C_2 in farads, B_1 in cycles per second. With practical units, that is, R_2 in kilohms, C_2 in micromicrofarads, and B_1 in megacycles, the expression becomes

$$B_1 C_2 R_2 = 113$$

which directly provides the bandwidth B_1 of a transformer whose secondary contains the shunt capacitance C_2 and the total damping in the form of a shunt resistance R_2 .

The gain per stage is easily found assuming a tube transconductance g_m . The expression previously given for K_0 becomes

$$K_0 = 1.41 g_m Q_2 \left(\frac{R_1 R_2}{Q_1 Q_2} \right)^{1/2}$$

and since $Q_1 = R_1 C_1 \omega_0$ and $Q_2 = R_2 C_2 \omega_0$, the final result is

$$K_0 = 1.41 g_m R_2 \left(\frac{C_2}{C_1} \right)^{1/2}$$

with g_m in amperes per volt, R_2 in ohms, C_1 and C_2 in farads. The same expression applies in practical units, that is, with g_m in milliamperes per volt, R_2 in kilohms, and C_1 and C_2 in micromicrofarads.

As we will show in the next paragraph, the case when one Q is very large provides, for the same bandwidth, a gain 1.41 times larger than if Q_1 and Q_2 were equal.

The drawback is that the first case corresponds to a transformer whose tuning is critical. Any mistuning, even slight, will result in a nonsymmetrical response curve. This is not true for the second case, which is therefore preferred for practical ease of tuning.

4-14. Equal Q 's

Simple expressions also result when the primary and secondary quality factors are equal. Let $Q_1 = Q_2 = Q$. Then

$$n = \frac{1}{Q^2}, \quad m = \frac{2}{Q^2}, \quad k_t = \frac{1}{Q}$$

Critical coupling k_c and transitional coupling k_t are equal. The 3-db bandwidth B_1 is

$$B_1 = \frac{1.41 f_0}{Q}$$

or

$$B_1 = 1.41 f_0 k_t$$

A comparison with the expression previously obtained when one Q is infinite shows that for identical bandwidth the equal Q 's case leads to a coupling coefficient 1.41 times smaller.

4-15. Practical design, equal Q 's

The expression for bandwidth, $B_1 = 1.41f_0/Q$, can be transformed by using $Q = R_1C_1\omega_0$ to provide

$$B_1 = \frac{1.41}{2\pi R_1 C_1}$$

or

$$B_1 C_1 R_1 = 0.225$$

with B_1 in cycles per second, C_1 in farads, and R_1 in ohms.

Using more practical units,

$$B_1 C_1 R_1 = B_1 C_2 R_2 = 225$$

with B_1 in megacycles, C_1 or C_2 in micromicrofarads, and R_1 or R_2 in kilohms.

The expression for the gain of the stage simplifies to

$$K_0 = \frac{1}{2}g_m(R_1 R_2)^{1/2}$$

4-16. Summing up

It may be helpful to sum up the results obtained for double-tuned transformers having transitional coupling, that is, maximally flat response.

Given:

- (a) the total shunt capacitances C_1 across the primary and C_2 across the secondary, in micromicrofarads;
- (b) the 3-db bandwidth B_1 in megacycles;
- (c) the tuning or center frequency f_0 in megacycles;
- (d) the tube transconductance g_m in milliamperes per volt.

Find:

- (a) the total damping resistances R_1 across primary and R_2 across secondary;
- (b) the gain K_0 at resonance;
- (c) the coupling coefficient k_t .

Two simple cases occur:

1. R_1 is infinite. All the damping is obtained by a resistor R_2 across the secondary. Then

$$R_2 = \frac{113}{B_1 C_2}$$

Notice that part of the secondary damping will be provided by the input resistance R_i of the following tube. In practical cases, other causes of damping are

negligible, so that the physical resistor R_p to connect across the secondary should have a value such that

$$\frac{1}{R_2} = \frac{1}{R_i} + \frac{1}{R_p} \quad \text{or} \quad R_p = \frac{R_2 R_i}{R_i - R_2}$$

The gain will be

$$K_0 = \frac{159 g_m}{B_1 (C_1 C_2)^{1/2}}$$

and the coupling coefficient

$$k_t = \frac{B_1}{f_0}$$

2. $Q_1 = Q_2$. This does not mean that $R_1 = R_2$, but that

$$R_1 C_1 = R_2 C_2$$

Then

$$R_1 = \frac{225}{B_1 C_1} \quad \text{and} \quad R_2 = \frac{225}{B_1 C_2}$$

As before, the physical resistor to connect across the secondary can be calculated by taking into account the input resistance of the following tube.

For the primary, a similar calculation can be done, taking into account the output resistance of the preceding tube. When this tube is a pentode as it generally is, its damping effect on the primary can be neglected.

The gain is

$$K_0 = \frac{1}{2} g_m (R_1 R_2)^{1/2}$$

and the coupling coefficient

$$k_t = \frac{0.71 B_1}{f_0}$$

Adjusting the coupling is facilitated by the fact that transitional coupling is equal to critical coupling, which corresponds to maximum gain at the resonant frequency. This provides an easy way of checking results or adjusting coupling.

4-17. Two-maxima curve

Overcoupling leads to a double-hump response curve. If the over-all response is considered, for the same bandwidth such a curve provides less attenuation at the edges than the maximally flat curve, at the price of a sag or trough in response around the resonant frequency (Fig. 4-14). Also, the overcoupled curve falls off more rapidly outside the passband, providing better adjacent selectivity or selectivity against adjacent channels.

Two means are available to obtain an overcoupled response—either increase the coupling coefficient k or increase the quality factor Q of the circuits. Increasing k makes for a wider frequency interval between the two maxima of

the curve. They move symmetrically away from the center frequency f_0 . Increasing Q does not actually modify the frequency distance between the two maxima, but the trough around center frequency becomes deeper. By referring to the curves of Fig. 4-12, a curve can be chosen to fit the wanted amount of sag on center frequency, and $n/2m$ is thus determined.

Because of phase distortion, a total sag of 3 db for the entire amplifier should not be exceeded. Some amount of compensation for the sag can be provided by single-tuned circuits in the r-f or i-f amplifiers. A typical value of

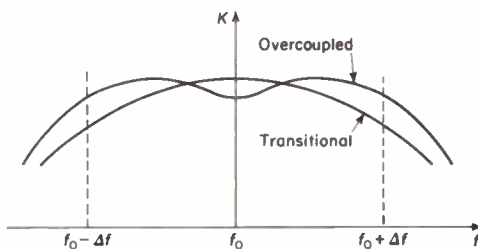


Figure 4-14. Transitional and overcoupled response curves.

sag for a single transformer would be 1 db, that is, 12 per cent. By solving the attenuation equation, it can be seen that this corresponds to the curve

$$\frac{n}{2m} = 0.5 \quad \text{or} \quad n = m$$

which goes through a minimum of

$$A = 0.88 \quad \text{at} \quad x = 0.8$$

The same curve goes through

$$A = 1 \quad \text{at} \quad x = 1$$

so that if the same 1-db attenuation is tolerated at the edges of the band as at the center of the band, the bandwidth is easily determined.

Assume identical primary and secondary Q 's for simplicity. Then

$$m = n = \frac{4}{Q^2}$$

and from $m = k^2 + 1/Q^2$, it follows that

$$k = \frac{3^{1/2}}{Q}$$

Also,

$$x = \frac{2d}{m^{1/2}} = dQ$$

and at the edge of the passband for which $x = 1$, we obtain

$$d = \frac{1}{Q}$$

Calling B_t the transformer bandwidth at 1 db,

$$B_t = \frac{2f_0}{Q}$$

and the transformer main characteristics are given by

$$Q = \frac{2f_0}{B_t} \quad \text{and} \quad k = \frac{0.866B_t}{f_0}$$

The resistances are then given by

$$R_1 = \frac{318}{B_t C_1} \quad \text{and} \quad R_2 = \frac{318}{B_t C_2}$$

with R_1 and R_2 in kilohms, B_t in megacycles, and C_1 and C_2 in micromicrofarads.

4-18. Practical design, overcoupled transformer

More generally, it is possible to design rapidly any overcoupled transformer using the preceding results.

The choice of a tolerated sag at the center frequency determines $n/2m$ with the help of the curves of Fig. 4-12, interpolating if necessary.

The tolerated attenuation at the edges of the desired bandwidth B determines x , which is

$$x = \frac{2d}{m^{1/2}}$$

From $B = 2df_0$ results

$$d = \frac{B}{2f_0}$$

so that m can be evaluated, and then n .

Two important practical cases occur.

1. *One large Q .* Let $Q_1 \gg Q_2$, which means that all the damping is on the secondary of the transformer. Then

$$n = \frac{1}{Q_2^2} \quad \text{or} \quad Q_2 = n^{-1/2}$$

and

$$m = k^2 \quad \text{or} \quad k = m^{1/2}$$

and since $Q_2 = 2\pi f_0 R_2 C_2$, it follows that

$$R_2 = \frac{1}{2\pi f_0 C_2 n^{1/2}}$$

2. *Equal Q's.* Let $Q_1 = Q_2 = Q$, which means $R_1 C_1 = R_2 C_2$. Now

$$n = \frac{4}{Q^2} \quad \text{or} \quad Q = \frac{2}{n^{1/2}}$$

Since $m = k^2 + 1/Q^2$, then

$$k = m - \frac{n}{4}$$

The Q is given by

$$Q = 2\pi f_0 R_1 C_1 = 2\pi f_0 R_2 C_2$$

so that

$$R_1 = \frac{1}{\pi f_0 C_1 n^{1/2}} \quad \text{and} \quad R_2 = \frac{1}{\pi f_0 C_2 n^{1/2}}$$

In practical units, these last two expressions become

$$R_1 = \frac{159Q}{f_0 C_1} \quad \text{and} \quad R_2 = \frac{159Q}{f_0 C_2}$$

with R_1 and R_2 in kilohms, f_0 in megacycles, and C_1 and C_2 in micromicrofarads.

4-19. Complex filters

Complex coupling circuits or filters are much less common than the single- or double-tuned coupling circuits. One reason is that tuning becomes more critical as the number of adjustable elements increases, and even a slight mistuning may have drastic effects on the over-all response. Also, coupling may become critical and manufacturing tolerances have to be tightened, with resulting increase in cost.

It can be said that complex filters are used only when it becomes almost impossible to obtain the desired response with single- or double-tuned circuits, or, in other words, when it must have a specific shape such as a steep falling off on one side, etc.

Such sophisticated techniques properly belong to wideband i-f amplifiers and will be touched upon in Chap. 5.

4-20. Practical graphs

In all practical cases, the results obtained by calculations are only approximate because of lack of accuracy in the given values or simplifying assumptions, or

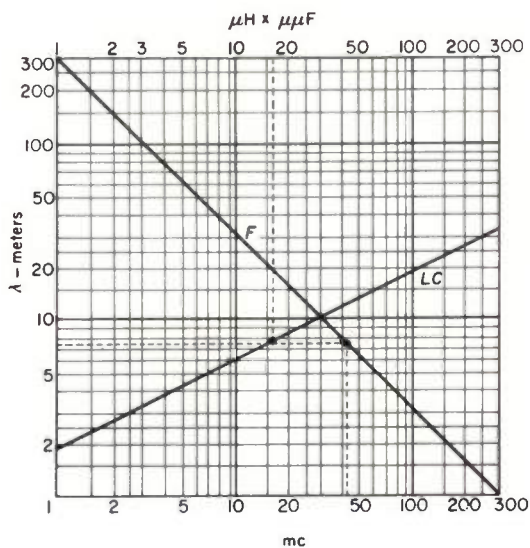
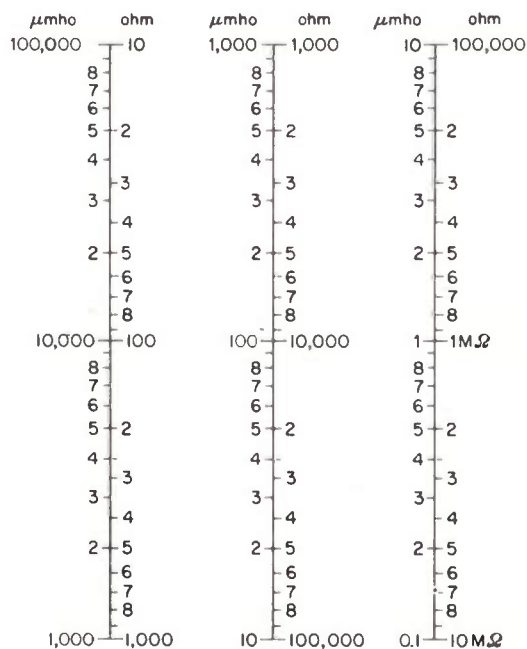
Figure 4-15. Graph relating frequency, wavelength, and LC product.

Figure 4-16. Nomogram to evaluate conductances.

neglect of secondary effects. Thus it is necessary to provide some easy way of proceeding to final adjustments; a common example is the use of adjustable magnetic cores to obtain the exact value of inductance. The range covered by such adjustments is generally quite wide, so that graphical methods can be resorted to for the calculations which will yield results of sufficient practical accuracy.

Figure 4-15 gives the frequency f corresponding to a wavelength λ in the range 1 to 300 mc or 300 to 1 m. It also provides the value of the LC product required to resonate on the given frequency in microhenrys $\times$ micromicrofarads. The dotted line example shows that a frequency of 42 mc corresponds to a wavelength of 7.14 m and requires an LC value of $16 \mu h \times \mu \mu f$.

Figure 4-16 is a conductance-resistance nomograph. A typical problem is the calculation of the value of the physical resistor R_p to connect across a tuned circuit already damped by the input resistance R_i of the following tube to obtain the correct total damping resistance R_t . The equation is

$$\frac{1}{R_p} = \frac{1}{R_t} - \frac{1}{R_i}$$

It is much easier to replace each resistance by its reciprocal, the conductance, putting

$$G_p = \frac{1}{R_p}, \quad G_t = \frac{1}{R_t}, \quad G_i = \frac{1}{R_i}$$

so that the equation becomes

$$G_p = G_t - G_i$$

The nomograph gives the transformation directly between R and G , and all problems of paralleling resistors reduce to simple additions or subtractions. The unit of conductance is the *mho*, which is simply the reciprocal of the ohm. A conductance of 1 micromho is equivalent to a resistance of 1 megohm.

4-21. Decibel

When comparing two powers P_1 and P_2 , the simplest way is to state their ratio P_1/P_2 . However, in television the simplest way is not the most convenient method because of the enormous range of power encountered. The input power to an antenna may be on the order of 1 micromicrowatt, and the output power to a speaker may be 10 watts. Their ratio would be 10^{13} . In such cases a logarithmic unit is convenient, and

$$G = \log_{10} \frac{P_1}{P_2}$$

expresses the relative values of P_1 and P_2 in bels when decimal logarithms are

used. The bel is a rather large unit; the decibel, which is one-tenth of a bel, is more commonly used so that

$$G \text{ db} = 10 \log_{10} \frac{P_1}{P_2}$$

where P_1 and P_2 must be expressed in identical units.

If the two powers P_1 and P_2 are developed across equal impedances, then the corresponding voltages across the impedances, E_1 and E_2 , and the currents through it, I_1 and I_2 , are related by

$$\frac{E_1}{E_2} = \frac{I_1}{I_2}$$

and

$$\frac{P_1}{P_2} = \frac{E_1 I_1}{E_2 I_2} = \left(\frac{E_1}{E_2} \right)^2 = \left(\frac{I_1}{I_2} \right)^2$$

so that

$$G \text{ db} = 10 \log_{10} \left(\frac{E_1}{E_2} \right)^2$$

or

$$G \text{ db} = 20 \log_{10} \frac{E_1}{E_2}$$

and similarly,

$$G \text{ db} = 20 \log_{10} \frac{I_1}{I_2}$$

The use of decibels is so convenient that it is common practice to use it to compare voltages or currents even when the impedances are different. Although it is incorrect on a strict basis, the practice is widely used and offers no serious drawback if the exact definition is kept in mind.

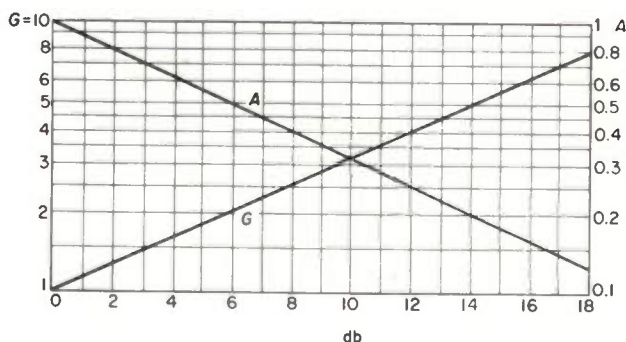


Figure 4-17. Graph relating gain and attenuation to decibels.

Figure 4-17 is a decibel graph. It gives the number of decibels corresponding to a given gain G or to a given attenuation A .

The graph is not too precise for small values; in such cases Table 4-1 will prove useful.

Table 4-1. Decibels

db	Gain	Attenuation	db	Gain	Attenuation
0.1	1.01	0.989	2.2	1.29	0.776
0.2	1.02	0.977	2.4	1.32	0.759
0.3	1.03	0.966	2.6	1.35	0.741
0.4	1.05	0.955	2.8	1.38	0.724
0.5	1.06	0.944	3.0	1.41	0.708
0.6	1.07	0.933	3.2	1.44	0.692
0.7	1.08	0.923	3.4	1.48	0.676
0.8	1.10	0.912	3.6	1.51	0.661
0.9	1.11	0.902	3.8	1.55	0.646
1.0	1.12	0.891	4.0	1.58	0.631
1.1	1.13	0.881	4.2	1.62	0.617
1.2	1.15	0.871	4.4	1.66	0.603
1.3	1.16	0.861	4.6	1.70	0.589
1.4	1.17	0.851	4.8	1.74	0.575
1.5	1.19	0.841	5.0	1.78	0.562
1.6	1.20	0.832	5.5	1.88	0.531
1.7	1.22	0.822	6.0	1.99	0.501
1.8	1.23	0.813	6.5	2.11	0.473
1.9	1.24	0.803	7.0	2.24	0.447
2.0	1.26	0.794	7.5	2.37	0.422

Problems

1. Starting from fundamentals, derive the relation between resonant frequency f_0 , the coefficient of quality Q , and 3-db bandwidth B of a parallel-tuned circuit RLC .

2. A parallel-tuned circuit has a capacitance C of 20 $\mu\mu\text{f}$ and must provide a bandwidth of 4 mc at a relative attenuation of 1.62. What should be the damping resistance?

3. A parallel-tuned circuit has a capacitance C of 16 $\mu\mu\text{f}$ and a damping resistance of 5000 ohms. Plot the attenuation as a function of the bandwidth.

4. An inductance has been connected between the grid of a stage and ground and shunted by a 10,000-ohm resistance. The measured 3-db bandwidth is 2.5 mc. What is the input capacitance of the stage? If the working frequency is 20 mc, what should be the value of the inductance?

5. Referring to a tube handbook, pick ten tubes suitable for wideband amplification and compare their relative performance. How are your conclusions modified if the tubes are used in a circuit which adds 12 $\mu\mu\text{f}$ parasitic capacitance?

6. The input resistance of a tube is 16,000 ohms at 50 mc. Calculate its input resistance at 20, 40, 80, and 160 mc.

7. Evaluate the gain of an input transformer matching a 300-ohm lead-in into a circuit tuned by a 18 $\mu\mu\text{f}$ capacitance and having a 3-db bandwidth of 4 mc.

8. A transitionally coupled, double-tuned transformer has a primary Q of 15 and a secondary Q of 18, and is tuned on 30 mc. What is the 3-db bandwidth? What is the 6-db bandwidth?

9. A transitionally coupled, double-tuned transformer has a primary Q of 100, a secondary Q of 5, and is tuned on 40 mc. Calculate the 3-db bandwidth and the coupling coefficient.

10. Evaluate the gain at resonance of a stage having as coupling circuit a transformer with primary Q of 120 and secondary Q of 6. The working frequency is 30 mc, and the primary and secondary are tuned with $16\ \mu\text{mf}$. Calculate the 3-db bandwidth. The tube transconductance is $6\ \text{ma/v}$.

11. A tube having $\mu = 800$ and $r_p = 1,000,000$ ohms is used in a stage having as coupling circuit a double-tuned transformer with identical primary and secondary. For each, the tuning frequency is 40 mc, the capacitance $14\ \mu\text{mf}$, and the damping resistance is 5000 ohms. Calculate the 3-db bandwidth and the stage gain.

12. Design an overcoupled, double-tuned transformer having a sag of 1.5 db at center, a 3-db bandwidth of 4 mc, and a center frequency of 40 mc for the case of a large secondary Q and for the case of equal Q 's. In both cases, shunt capacitances are $12\ \mu\text{mf}$ at primary and $16\ \mu\text{mf}$ at secondary.

13. Evaluate the number of decibels corresponding to: (a) a power gain of 4562; (b) a voltage gain of 1200; (c) a current gain of 1864.

14. A voltage amplifier stage has a gain of 26 times. How many stages are needed to obtain a total gain of 130 db?

Chapter 5

R-F AND I-F WIDEBAND AMPLIFICATION

In a television receiver, r-f and i-f amplifiers use several stages in cascade to produce enough gain over the necessary bandwidth. However, each stage is not designed separately, but rather the entire amplifier is considered as a complete system, and several design methods have been developed to provide satisfactory practical performance. The simplest case is that of the synchronous amplifier, where all stages have identical tuning frequencies. However, staggering of the tuning frequencies ensures better results and is frequently employed. Feedback between stages also may be employed for the same reason, and simple, practical design methods are available.

5-1. Cascading stages

A practical amplifier embodies a number of stages which are cascade-connected one behind the other. Three to five tubes may be used in a typical i-f amplifier, which means four to six coupling circuits.

On the other hand, r-f amplification preceding the mixer tube uses only one or two stages, that is, two or three tuned circuits. Because of the high center frequency at r-f, and because of the difficulty in obtaining a well-defined pass-band, the exact shape of the total response curve is mainly determined by the i-f amplifier. In this case, the r-f amplifier is designed to provide nearly constant amplification over the desired band, with the highest gain possible and with little regard for shaping the entire response curve.

Although the cascading of stages to obtain a given gain over a given bandwidth is a very general problem, it will be studied mainly with reference to the i-f amplifier.

The coupling circuits may be any of the types described in the preceding chapter.

The simplest way to design the amplifier is to design one stage around a type of tube and a type of coupling circuit and to use several identical stages in cascade. All circuits will be identical and tuned on the same frequency. For this reason, such an arrangement is called a *synchronous amplifier*.

Better over-all results are obtained if the successive coupling circuits are not tuned on the same frequency, thus creating a *stagger-tuned* amplifier.

Finally, a third method uses the properties of feedback around stages or groups of stages.

5-2. Synchronous amplifier

When n identical stages of gain at resonance K_0 are cascaded, the total gain at resonance K_t is given by

$$K_t = K_0^n$$

Let f_1 and f_2 be the two frequencies for which the gain of an individual stage falls down by 3 db, that is, reduced to $0.707 K_0$. The 3-db bandwidth per stage B_1 is

$$B_1 = f_2 - f_1$$

When several stages are cascaded, each one brings in 3-db attenuation at f_1 or f_2 so that the total attenuation for the amplifier is $3n$ db at f_1 or f_2 . This means that the 3-db bandwidth B_t of the amplifier is smaller than B_1 .

The greater the number of stages, the more B_t shrinks with comparison to B_1 . The reduction is expressed by the shrinking factor S , which can be evaluated as follows:

For a total amplifier 3-db bandwidth B_t , the total attenuation A_t should be

$$A_t = 1.414 = 2^{1/2}$$

so that each of the individual stages should contribute an attenuation A_i

$$A_i = (A_t)^{1/n}$$

Now the bandwidth B_i of one stage is linked to the attenuation per stage A_i through

$$B_i CR = 159(A_i^2 - 1)^{1/2}$$

and substituting for A_i ,

$$B_i CR = 159(2^{1/n} - 1)^{1/2}$$

The 3-db bandwidth B_1 for a single stage is given by

$$B_1 CR = 159$$

which shows that the ratio between 3-db bandwidths B_t for the entire amplifier and B_i for a single stage is

$$S = \frac{B_t}{B_i} = (2^{1/n} - 1)^{1/2}$$

Numerical values are given by Table 5-1.

Table 5-1. Shrinking Factor

<i>n</i>	1	2	3	4	5
<i>S</i>	1	0.64	0.51	0.44	0.39

Notice that the bandwidth shrinks drastically when the number of stages increases. This creates the necessity of designing each stage for a wider bandwidth with resulting loss of gain.

5-3. Stagger tuning

The double-hump response curve characteristic of an overcoupled, double-tuned transformer can be obtained with two stagger-tuned, single-tuned

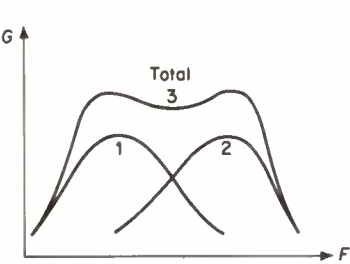


Figure 5-1. Combining response curves for two stages.

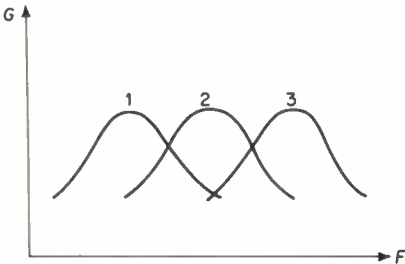


Figure 5-2. Three staggered circuits.

circuits. Figure 5-1 shows the individual response curves 1 and 2 for each stage. When the two stages are cascaded, the overall response curve 3 exhibits the shape shown, with two maxima and a sag or trough in the middle. To widen the bandwidth, it is necessary only to increase the difference in the resonant frequencies 1 to 2. Unfortunately, the trough in the center of the total response curve becomes deeper. We might try to compensate for this by adding another circuit tuned in between the two preceding circuits, which leads to the triplet arrangement shown in Fig. 5-2. With the curves indicated, however, there is overcompensation and the total response curve would have a hump in the middle corresponding to the resonant frequency of circuit 2.

A much better over-all curve results, as shown in Fig. 5-3, if the central circuit has heavier damping than the other two.

Figures 5-1 and 5-3 indicate the general method of designing stagger-tuned amplifiers. The circuits will be symmetrically tuned on each side of the desired center frequency. If their number is odd, one circuit will be left over and will

be tuned on the center frequency, with heavier damping, to compensate for the sag.

The over-all response curve evidently results from two sets of variables: the tuning frequencies and the shapes of the curves, that is the dampings of the circuits.

Mathematical analysis of the problem of obtaining the best over-all response curve leads to a set of design equations whose derivation will not be attempted here. Instead, a simplified version will be applied to practical design.

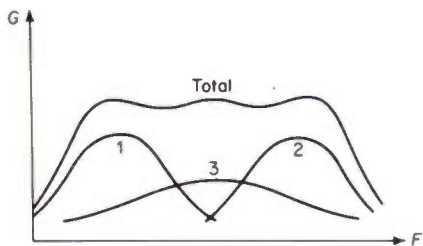


Figure 5-3. Three staggered circuits providing almost flat total response curve.

5-4. Staggered amplifiers

The amplifiers considered use single-tuned coupling circuits and have a variable number of stages. The over-all response curve will be defined by

its 6-db bandwidth, so that for single sideband reception one edge of the passband should coincide with the picture carrier frequency.

The total number of circuits will be split into two groups. For all practical cases having three to five tubes, that is, four to six circuits, each group will be a doublet (two circuits) or a triplet (three circuits):

- (a) four circuits—two doublets;
- (b) five circuits—one doublet, one triplet;
- (c) six circuits—two triplets.

For a doublet the two circuits are tuned on frequencies f_1 and f_2 , symmetrical with respect to the center frequency f_0 :

$$f_1 f_2 = f_0^2$$

For a triplet, the third circuit is tuned on f_0 . The attenuation reaches 6 db for the whole amplifier at frequencies f_a and f_b so that the 6-db bandwidth is

$$B = f_b - f_a$$

The center frequency is defined by

$$f_0^2 = f_a f_b$$

If the bandwidth is small compared with f_0 , the arithmetical mean can be taken instead of the geometrical mean with good approximation, giving

$$f_0 = \frac{1}{2}(f_a + f_b)$$

5-5. Design method

Given:

- (a) total gain K ;
- (b) 6-db bandwidth B in megacycles;
- (c) center frequency f_0 in megacycles;
- (d) total shunt capacitance C in micromicrofarads for each circuit.

Find:

- (a) number of stages;
- (b) tuning frequencies in megacycles;
- (c) shunt resistances in kilohms;
- (d) eventually, inductance values.

1. *Stages.* The average gain per stage is

$$K_i = \frac{159g_m}{BC}$$

with g_m in milliamperes per volt.

The number of stages n is given by solving

$$(K_i)^n = K$$

and rounding off n to the next integer. There will then be $n + 1$ circuits, divided into doublets and triplets.

2. *Doublet.* Calculate the relative bandwidth

$$M = \frac{f_0}{B}$$

and evaluate the stagger factor

$$a = 1 + \frac{0.35}{M}$$

The tuning frequencies are

$$f_1 = \frac{f_0}{a} \quad \text{and} \quad f_2 = af_0$$

The shunt resistances are

$$R_1 = \frac{225a}{BC} \quad \text{and} \quad R_2 = \frac{225}{aBC}$$

3. *Triplet.* Evaluate the stagger factor

$$a = 1 + \frac{0.43}{M}$$

The tuning frequencies are

$$f_0, \quad f_1 = \frac{f_0}{a}, \quad f_2 = af_0$$

The corresponding shunt resistances are

$$R_0 = \frac{159}{BC}, \quad R_1 = \frac{318a}{BC}, \quad R_2 = \frac{318}{aBC}$$

The physical resistances are evaluated as explained before, taking the input impedance of the following tube into account.

4. *Inductors.* Inductance values are calculated using Thomson's formula, since the resonant frequency and the total shunt capacitance are known for each circuit. Notice that there is no necessity for all capacitances to be identical in the expressions given above. Each value will determine the corresponding resistance. The only difficulty would concern the average gain, for which the average value of capacitance should be used.

If several stages have the same total shunt capacitances, it is usually expedient to manipulate Thomson's expression to obtain an easier formula. For example, for $C = 20 \mu\mu f$, the classical expression

$$\omega = \frac{1}{(LC)^{1/2}}$$

can be transformed into

$$L = \frac{1280}{f_2}$$

with L in microhenrys and f in megacycles.

5. *Order of staggering.* The over-all curve does not depend on the order of staggering of the circuits, which is determined by practical considerations.

Diode detectors, which are commonly used, introduce a heavy damping of the preceding circuit, where it may be necessary to use the most damped circuit of all, that is, the circuit tuned on f_0 . Furthermore, to minimize the risks of instability, tuned circuits separated by two stages of amplification should have resonant frequencies as widely separated as possible.

5-6. Overcoupled amplifiers

When the coupling circuits between stages are overcoupled, double-tuned transformers, they are generally all tuned to the same frequency; this will be assumed in the following discussion, where Dishal's method will be used. Figure 5-4 illustrates the symbols used for a transformer having a primary-to-secondary coupling through mutual inductance M , such that

$$M = k(L_1 L_2)^{1/2}$$

where k is the coupling coefficient. Figure 5-5 shows the symbols used with a transformer having combined coupling by mutual inductance M and an additional coupling capacitance C_3 .

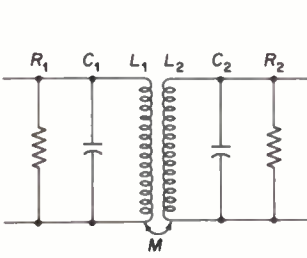


Figure 5-4. Components in a double tuned transformer.

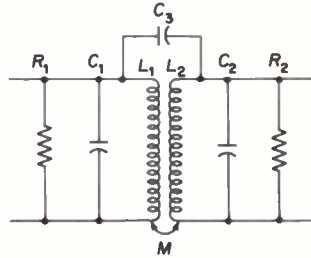


Figure 5-5. Additional capacitive coupling through C_3 .

The primary and secondary quality factors are equal:

$$Q = 2\pi f_0 R_1 C_1 = 2\pi f_0 R_2 C_2$$

where f_0 is the resonant frequency.

The response curve is illustrated in Fig. 5-6. B is the desired bandwidth for an attenuation

$$a = \frac{V_1}{V}$$

B_0 is the frequency interval between the two maxima. The attenuation at the center frequency with respect to maximum response is

$$b = \frac{V_1}{V_0}$$

which determines the center sag.

Once the shape of the response curve is decided upon, then a , b , and B are known.

Evaluate

$$\alpha = \frac{b}{(b^2 - 1)^{1/2}} \quad \text{and} \quad \beta = \left(\frac{\alpha - 1}{2} \right)^{1/2}$$

Then

$$\frac{B}{B_0} = [1 + 2\beta(1 + \beta^2)^{1/2}(a^2 - 1)^{1/2}]^{1/2}$$

determines B_0 , and Q follows from $Q = f_0/\beta B_0$, so that

$$R_1 = \frac{Q}{2\pi f_0 C_1} \quad \text{and} \quad R_2 = \frac{Q}{2\pi f_0 C_2}$$

The coupling coefficient is

$$k = \frac{\left(1 + \frac{1}{\beta^2}\right)^{1/2}}{Q}$$

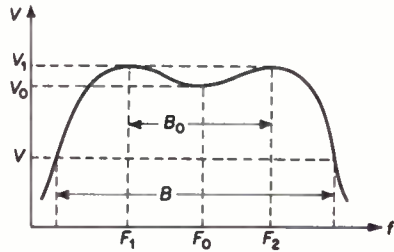


Figure 5-6. General response curve of double tuned transformer.

Let $C = (C_1 C_2)^{1/2}$ for pure magnetic coupling, and

$$C = [(C_1 + C_3)(C_2 + C_3)]^{1/2}$$

for combined coupling. The gain per stage is

$$K_0 = \frac{g_m}{4\pi B_0 C}$$

Calculations are notably simplified if the same attenuation is adopted at the edges and at the center of the pass-band, as shown in Fig. 5-7. In this case,

$$B_0 = 0.71 B_1$$

where B_0 is the frequency width between maxima and B_1 the desired bandwidth. B_0 can be evaluated directly from B_1 and the result used in the preceding expressions for Q , R , k , and K_0 .

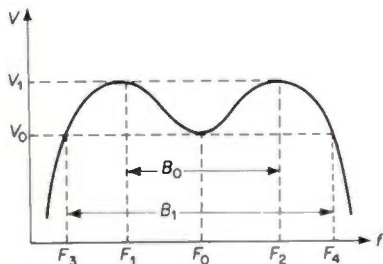


Figure 5-7. Response curve having same attenuation at the center and at the edges of the band.

5-7. Feedback

The principle of feedback is illustrated in Fig. 5-8, which shows two pentode amplifying stages using antiresonant coupling circuits. The tuning capacitance is entirely composed of the parasitic capacitances without physical capacitor, which is a common arrangement.

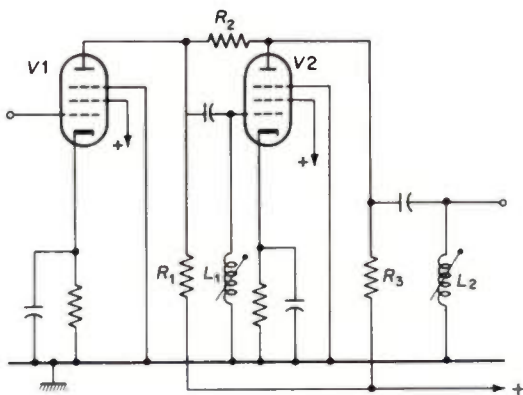


Figure 5-8. Anode to anode feedback through R_2 .

For r-f input frequencies, ground and $B+$ are actually equivalent, so that the anode resistances R_1 and R_3 appear in fact across L_1 and L_2 . Besides being

the anode load resistances, R_1 and R_3 are simultaneously used as damping resistances; this is also a commonly used arrangement.

Negative feedback is provided by the plate-to-plate coupling resistance R_2 . Notice that the circuit really provides plate-to-grid feedback around one stage through the interstage coupling capacitor. Only the second tube V_2 is subjected to feedback. This arrangement is called a *shunt feedback pair*.

The parasitic shunt capacitance across R_2 upsets the circuit somewhat and practical response curves are skewed for this reason. The alternative circuit

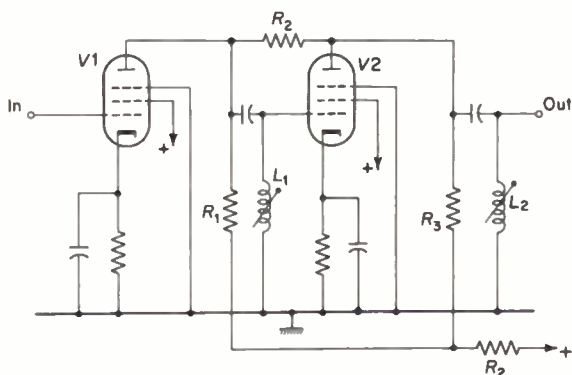


Figure 5-9. Anode to anode feedback of the series type.

shown in Fig. 5-9 is preferred because it provides more symmetrical curves. It is called a *series feedback pair*, and the feedback is due to the common anode resistance R_2 .

Both resonant circuits are tuned on the center frequency f_0 , and the amount of feedback determines the over-all response curve for the pair.

H. S. Jewitt's method will be used as a practical method to determine the response curve of the feedback pair.

5-8. Response curve

The amount of feedback depends on the values of R_1 , R_2 , and R_3 given in Figs. 5-8 and 5-9. It determines whether the over-all response curve will have one or two maxima through a parameter k called the *form factor*. Form factor k may vary between 2 and -2 . For $k = 2$ there is no feedback, and the circuits of Figs. 5-8 and 5-9 reduce to a synchronous two-stage amplifier.

Figure 5-10 shows several response curves corresponding to various values of k . The ordinate scale gives the attenuation y in decibels with respect to the

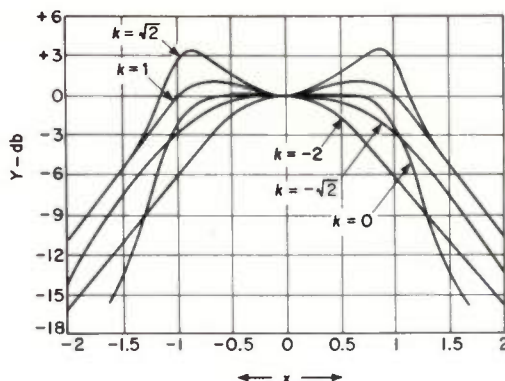


Figure 5-10. Response curves for the feedback pair.

gain at center frequency. The abscissa scale gives the relative frequency distance x from center frequency. Consequently, center frequency gives

$$x = 0 \quad \text{and} \quad y = 1$$

and the desired 3-db bandwidth obtains for

$$x = \pm 1$$

The combination of these curves between themselves and eventually with a standard single-tuned stage provides any shape of over-all response curve.

5-9. Feedback amplifiers

Whatever the number of stages, they can be separated into feedback pairs, eventually with one stage left over to be used as synchronous stage. In all cases, all circuits are individually tuned on the center frequency f_0 .

For two stages or a pair, Fig. 5-10 applies directly. The response is maximally flat for $k = 0$ with an attenuation of 3 db at the edges of the passband, that is, between

$$x = 1 \quad \text{and} \quad x = -1$$

For identical attenuation at the center and at the edges of the passband, the value should be

$$k = 1$$

and there are two maxima at about 1 db above center frequency response.

For three stages, a feedback pair with $k = 1$ may be combined with a synchronous stage. The principle is illustrated Fig. 5-11. The total response curve is maximally flat and provides 3-db attenuation for

$$x = 1 \quad \text{and} \quad x = -1$$

Four stages can be divided into two feedback pairs. If they are identical, the total response curve has the shape of the curves of Fig. 5-10, according to the value of k , but with attenuations which are twice as large. In other words,

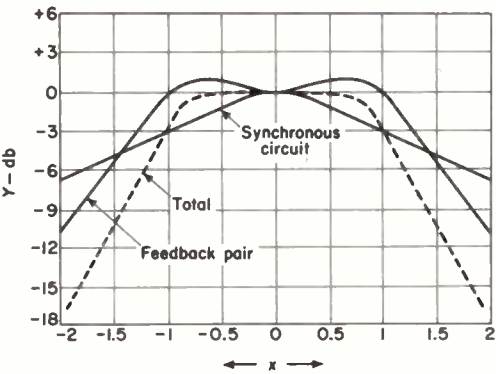


Figure 5-11. Combined response curves for three stages.

doubling the values of the ordinate scale of Fig. 5-10 gives the curves for two identical pairs.

For example, if each pair is maximally flat, the attenuation will be 6 db for

$$x = 1 \quad \text{and} \quad x = -1$$

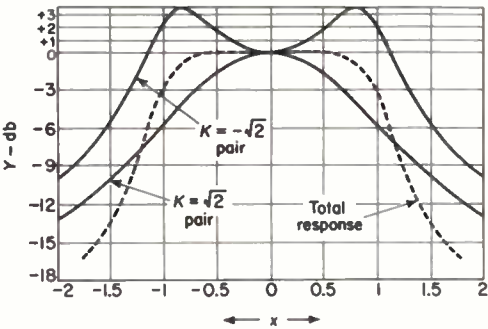


Figure 5-12. Combined response curves for four stages.

However, the combination of two maximally flat curves does not provide a maximally flat total response. Such a response is obtained with one pair having $k = 1.414$ (or $\sqrt{2}$) and another pair having $k = -1.414$. This combination is illustrated Fig. 5-12. The total curve has 3-db attenuation at

$$x = 1 \quad \text{and} \quad x = -1$$

Another possibility occurs with two pairs with $k = 1$. The total response curve has two 2-db humps in the passband. The response at the center of the band is equal to the response at

$$x = 1 \quad \text{and} \quad x = -1$$

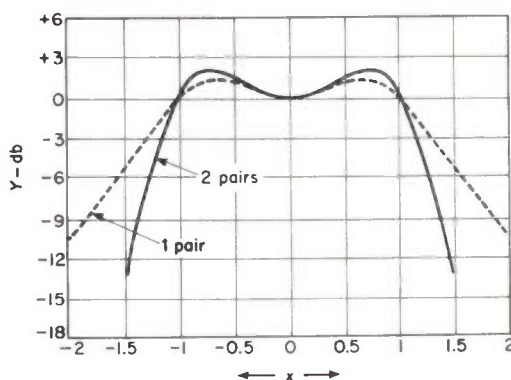


Figure 5-13. Response curves for one and two feedback pairs.

In this case, an attenuation of 6 db at the edges of the passband, suitable for single sideband reception, is obtained for a passband comprised between

$$x = 1.3 \quad \text{and} \quad x = -1.3$$

so that the ratio of the 0-db bandwidth ($x = \pm 1$) to the 6-db bandwidth ($x = \pm 1.3$) is equal to 0.78.

Figure 5-13 shows the response for one and two feedback pairs with $k = 1$.

5-10. Practical design

The elements which determine the shape of the response curve are the three resistances R_1 , R_2 , and R_3 for the shunt circuit of Fig. 5-8 and for the series circuit of Fig. 5-9.

Let B = bandwidth at 3 db,

C = total shunt capacitance across the inductance,

g_m = transconductance,

$R = \frac{1}{2}\pi BC$ = equivalent damping resistance,

$K_0 = g_m R$ = average gain per stage.

The values of the resistances are given by Table 5-2. As before, the values given for R_1 and R_3 are the total damping resistance. The physical resistor to use must be calculated by taking into account the input resistance of the following tube.

Table 5-2

<i>k</i>	<i>Shunt circuit</i>		<i>Series circuit</i>	
	<i>R</i> ₁ = <i>R</i> _s	<i>R</i> ₂	<i>R</i> ₁ = <i>R</i> _s	<i>R</i> ₂
$-\sqrt{2}$	$\frac{4K_0R}{1.52K_0 - 3.42}$	$1.17K_0R$	$\frac{4K_0R}{1.52K_0 + 3.42}$	$\frac{13.7K_0R}{2.32K_0^2 - 11.6}$
-1	$\frac{4K_0R}{2K_0 - 3}$	$\frac{4K_0R}{3}$	$\frac{4K_0R}{2K_0 + 3}$	$\frac{12K_0R}{4K_0^2 - 9}$
0	$\frac{2K_0R}{1.41K_0 - 1}$	$2K_0R$	$\frac{4K_0R}{3.46K_0 + 1}$	$\frac{4K_0R}{12K_0^2 - 1}$
1	$\frac{4K_0R}{3.46K_0 - 1}$	$4K_0R$	$\frac{2K_0R}{2K_0 + 1}$	$\frac{2K_0R}{2K_0^2 - 1}$
$\sqrt{2}$	$\frac{4K_0R}{3.7K_0 - 0.58}$	$6.9K_0R$	$\frac{4K_0R}{3.7K_0 + 0.58}$	$\frac{2.32K_0R}{13.7K_0^2 - 0.335}$

When only double-hump curves are used, that is, for positive values of *k*, and if the average gain per stage *K*₀ is at least 10, the expressions can be simplified to give results with sufficient practical accuracy according to Table 5-3.

Table 5-3

<i>k</i>	<i>Shunt circuit</i>		<i>Series circuit</i>	
	<i>R</i> ₁ = <i>R</i> _s	<i>R</i> ₂	<i>R</i> ₁ = <i>R</i> _s	<i>R</i> ₂
0	1.4 <i>R</i>	2 <i>K</i> ₀ <i>R</i>	1.16 <i>R</i>	0.33 <i>R</i> / <i>K</i> ₀
1	1.16 <i>R</i>	4 <i>K</i> ₀ <i>R</i>	<i>R</i>	<i>R</i> / <i>K</i> ₀
$\sqrt{2}$	1.1 <i>R</i>	6.9 <i>K</i> ₀ <i>R</i>	1.1 <i>R</i>	0.17 <i>R</i> / <i>K</i> ₀

5-11. Tube input impedance

In most amplifiers manual or automatic gain control is obtained by varying the bias of amplifying tubes, which should be of the remote cutoff type.

Changing the bias modifies the transconductance and, hence, the stage gain. Unfortunately, the input impedance has a resistive and a capacitive component, and both vary with the transconductance. The result is a detuning of the input circuit due to capacitance variation and a change in damping due to resistance variation.

Capacitance variation is due mainly to a modification of the space charge distribution between grid and cathode, with resulting change in the grid-to-cathode capacitance *C*_{gk}.

Capacitance variation due to Miller effect is almost negligible with pentode tubes because of the very small value of the plate-to-grid capacitance *C*_{gp}.

Since no physical capacitor is used across the tuned circuit when maximum gain is wanted, the input capacitance of the tube constitutes a substantial fraction of the total capacitance and its variation, however small, may cause mistuning.

When the transconductance is reduced to a low value, the input capacitance may decrease by a few micromicrofarads below its nominal value.

There are several ways to reduce the variation of input capacitance and, at the same time, the variation of input resistance. A small amount of negative

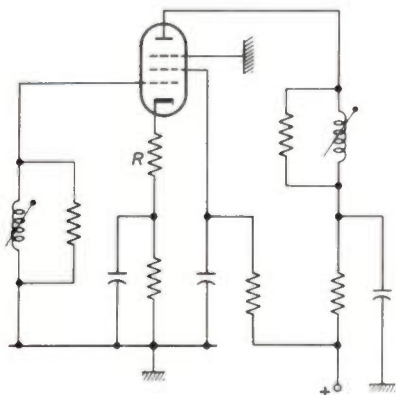


Figure 5-14. Cathode feedback through R .

feedback is a common remedy. It is easily obtained by providing an uncoupled resistance R in the cathode circuit, thus ensuring current feedback as shown in Fig. 5-14. Notice that in this case the cathode is no longer grounded and all decoupling capacitances should return to ground and not to the cathode. The feedback resistor has a relatively small value, usually 10 to 50 ohms. The drawback of the circuit is that the gain of the stage is reduced by the negative feedback.

Let R be the value of the feedback resistor. The feedback factor is

$$1 + g_m R$$

where g_m is the transconductance. The effective input capacitance C_f is smaller than the nominal input capacitance C_i :

$$C_f = \frac{C_i}{1 + g_m R}$$

We have previously stated that if g_m decreases, so does C_i . If the value of R is chosen with care, then it can be arranged for C_f to stay approximately constant, which is the desired result.

Let kC_i be the variation of C_i corresponding to a given variation of g_m . The new C_f value is

$$C_f = C_i - kC_i = C_i(1 - k)$$

so that if C_f is to be constant we must have

$$C_i(1 - k) = \frac{C_i}{1 + g_m R}$$

which can be solved for R :

$$R = \frac{k}{g_m(1 - k)}$$

This value of R will ensure exact compensation of the input capacitance only for the given transconductance variation used in the calculations. Compensation will be obtained for intermediary values of g_m only if the input capacitance variation is a linear function of transconductance. The results obtained are sufficiently accurate for practical purposes if the transconductance variation does not exceed 10 times.

As an example, a 6BX6 tube has an input capacitance of $9.4 \mu\mu\text{f}$ for a transconductance of 7 ma/v , and of $7.2 \mu\mu\text{f}$ for a transconductance of 0.7 ma/v . The capacitance variation is $2.2 \mu\mu\text{f}$ and $k = 2.2/9.4 = 0.23$ so that

$$R = \frac{0.23}{0.007(1 - 0.23)} = 42 \text{ ohms}$$

We have expressed g_m in amp/v to obtain R in ohms. The feedback factor is approximately

$$1 + g_m R = 1.3$$

and the gain of the stage is reduced by a factor of 1.3.

A compromise could be made by using a lower value for R , which would give a smaller reduction in gain at the price of a small input impedance variation.

Another entirely different method to reduce the variation in input impedance of pentodes is to apply bias to the third or suppressor grid. This bias should roughly be about 10 to 15 times larger than the first grid bias. Figure 5-15 illustrates a simple way of obtaining this result; resistance R_2 should be 10 to 15 times larger than R_1 . Since the control grid circuit returns to the junction of R_1 and R_2 , the first grid bias is only the voltage drop across R_1 . On the other hand, the suppressor grid is grounded, so that its bias is the full voltage drop across R_1 and R_2 .

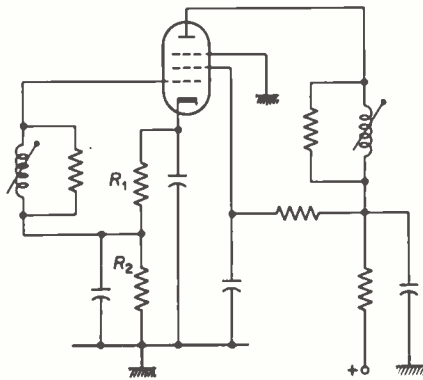


Figure 5-15. Biasing the suppressor grid.

The bias variation necessary for gain control is obtained by shunting a control potentiometer across R_1 and R_2 , adding a fixed resistor to set minimum bias. Figure 5-16 shows the resulting circuit in the case of a 6BX6 tube. The input capacitance variation is reduced by a factor of 20 times, and the input resistance variation is negligible.

When the gain control is obtained by varying a negative bias voltage, a resistance divider can be provided to maintain the right ratio between suppressor grid and control grid bias. Such is the role of R_1 and R_2 shown in Fig. 5-17.

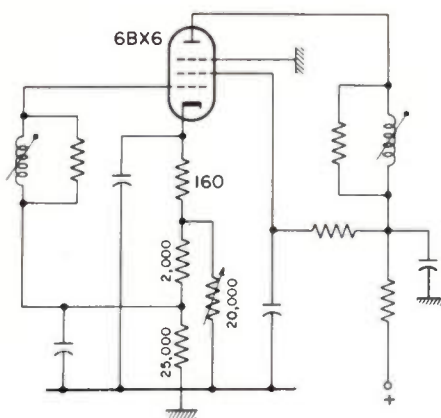


Figure 5-16. Practical circuit embodying gain control.

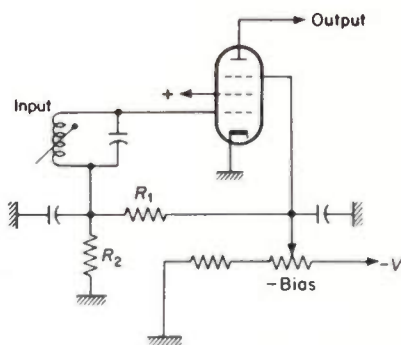


Figure 5-17. Simultaneous bias variation for the first and third grid.

Problems

1. Calculate the relation between the 3-db bandwidth B_1 of a single-stage and the 3-db bandwidth B_t of a synchronous amplifier using (a) 6 stages and (b) 8 stages.
2. Establish the relation between the 6-db bandwidth of a stage and the 6-db bandwidth of a synchronous amplifier using n such stages.
3. Compare synchronous and stagger tuning.
4. Design a synchronous amplifier having a total gain of 1000, a 6-db bandwidth of 4 mc, a center frequency of 40 mc, and using parallel-tuned coupling circuits. The total shunt capacitance is $10 \mu\text{mf}$. The tubes have a transconductance of 8 ma/v .
5. Design a synchronous amplifier having a gain of 750, a 3-db bandwidth of 4 mc, and a center frequency of 40 mc with a shunt capacitance of $16 \mu\text{mf}$. The tubes have a transconductance of 7.2 ma/v .
6. Design a synchronous amplifier having a gain of 100, a 6-db bandwidth of 4.5 mc, a center frequency of 180 mc, and using parallel-tuned coupling circuits with a shunt capacitance of $12 \mu\text{mf}$. The tubes have a transconductance of 9 ma/v .
7. Design a staggered amplifier having a gain of 1000, a 6-db bandwidth of 4 mc, a center frequency of 40 mc, and using parallel-tuned circuits with a shunt capacitance of $15 \mu\text{mf}$. The tubes have a transconductance of 8 ma/v .
8. Design a staggered amplifier having a gain of 750, a 3-db bandwidth of 4 mc, a center frequency of 32 mc, and using parallel-tuned circuits with a shunt capacitance of $16 \mu\text{mf}$. The tubes have a transconductance of 7.2 ma/v .
9. Design a staggered amplifier having a gain of 100, a 6-db bandwidth of 4.5 mc, a center frequency of 180 mc, and using parallel-tuned coupling circuits with a shunt capacitance of $12 \mu\text{mf}$. The tubes have a transconductance of 9 ma/v .

Calculate the physical damping resistance, the tubes having an input impedance of 16,000 ohms at 150 mc.

Evaluate the inductance values.

10. Establish a simple practical expression giving inductance directly in microhenrys as a function of frequency in megacycles to resonate with a capacitance of (a) 12 μmf , (b) 18 μmf , (c) 25 μmf .

11. Design an overcoupled, double-tuned transformer having equal primary and secondary Q 's, a 1-db bandwidth of 4 mc, a center frequency of 40 mc, and an attenuation at center frequency of 0.5 db.

Calculate the gain per stage for $g_m = 6 \text{ ma/v}$.

12. Design an overcoupled, double-tuned transformer having equal primary and secondary Q 's, a 3-db bandwidth of 4 mc, a center frequency of 24 mc, and a sag of 1 db at center frequency. Calculate the gain per stage for $g_m = 6 \text{ ma/v}$.

13. Design an overcoupled, double-tuned transformer having equal primary and secondary Q 's, a 1-db bandwidth of 4 mc, a center frequency of 44 mc, and a sag of 1 db at center frequency.

Calculate the gain per stage for $g_m = 7.2 \text{ ma/v}$.

14. Design a feedback pair for a 3-db bandwidth of 4 mc centered on 45 mc. The total shunt capacitance is 14 μmf , the transconductance is 7 ma/v. A 1-db irregularity of gain is tolerated in the passband.

15. Design a four-stage feedback amplifier having a 6-db bandwidth of 4 mc centered on 24 mc. The interstage total shunt capacitance is 18 μmf . The transconductance is 6 ma/v. The response curve should be as flat as possible in the passband.

16. Design a feedback pair for a 3-db bandwidth of 4.5 mc centered on 172 mc. Interstage total shunt capacitance is 12 μmf . Transconductance is 7.2 ma/v. A 3-db irregularity is tolerated in the passband.

Calculate the inductances and physical resistances. The input resistance of the tubes is 10,000 ohms at 120 mc.

17. Calculate the cathode feedback resistance to reduce the input impedance variation of a tube having an input capacitance of 12 μmf at 6.4 ma/v and of 9.2 μmf at 0.4 ma/v.

Evaluate the reduction in stage gain.

18. It is desired to design an amplifier having a gain of 1600 over a bandwidth of 4 mc at 6 db, with center frequency on 35 mc. The coupling circuits are parallel-tuned circuits with 16 μmf shunt capacitance. The tubes have a transconductance of 6 ma/v.

Synchronous or stagger tuning may be used. Compare the two methods.

19. An amplifier having a gain of 12,000 with a 6-db bandwidth of 4.5 mc centered on 40 mc may use stagger tuning or overcoupled, double-tuned

transformers. Input capacitance is $9\ \mu\text{mf}$, output capacitance $6\ \mu\text{mf}$. The tubes have a $7.2\ \text{ma/v}$ transconductance. Compare the two designs.

20. A two-tube r-f amplifier has a 3-db bandwidth of 4.5 mc centered on 180 mc. It uses tubes having a transconductance of $7\ \text{ma/v}$. Total input capacitance is $8\ \mu\text{mf}$, total output capacitance $7\ \mu\text{mf}$. Compare performances for: (a) synchronous tuning; (b) staggered tuning; (c) feedback pair; (d) overcoupled, double-tuned transformers.

Chapter 6

R-F AMPLIFICATION

It is usually customary to provide r-f amplification of the received signal before frequency changing. Besides providing extra gain, r-f amplification will secure some amount of selectivity on the signal frequency and will help to eliminate unwanted signals which may lead to interference if they are allowed to reach the frequency changer. In addition, r-f amplification will reduce the amount of local oscillation signal leaking to the antenna and radiating in the neighborhood.

However, the main reason for using r-f amplification is that it ensures a better over-all signal-to-noise ratio because r-f amplifier tubes produce less noise than frequency changers.

6-1. R-f stages

In most television receivers, some amount of amplification is provided between the antenna input and the frequency changer tube. One or two r-f stages, equipped with pentode or triode tubes, are used to accomplish this.

The main object of r-f amplification is to provide a better signal-to-noise ratio and some extra gain. R-f stages also reduce the possibility of the receiver's local oscillator radiating through coupling to the antenna and disturbing neighboring receivers.

Very few receivers have no r-f stage at all. Some have only one stage, but most use two stages, the tubes being triodes which are combined in a single glass envelope; special tubes have been developed for this purpose.

The almost universal custom is to have a separate subassembly, called a *tuner*, for the r-f and frequency changing stages with the associated coils and components.

In any case, the first r-f stage is a problem in itself because of its coupling to the antenna and because it determines the signal-to-noise ratio—essentially, in effect, the possibility of reception.

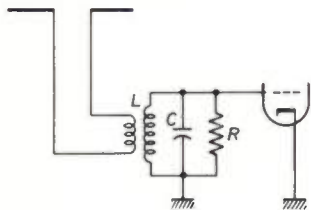


Figure 6-1. Antenna input using an untuned primary transformer.

6-2. First r-f stage

The antenna is coupled to the r-f stage through a tuned secondary transformer or through its equivalent, the autotransformer (this component has been previously analyzed). In the present case, the circuit is shown in Fig. 6-1 in simplified form up to the input to the first r-f tube.

The results of interest will be briefly recalled.

Let B = bandwidth for 3-db attenuation,

C = shunt capacitance,

Z_1 = primary input impedance,

Z_2 = secondary impedance,

k = transform ratio.

The secondary impedance is

$$Z_2 = \frac{1}{2\pi BC}$$

and the transform ratio

$$k = \left(\frac{Z_2}{Z_1} \right)^{1/2}$$

The secondary inductance is given by

$$L = \frac{1}{C\omega_0^2}$$

where ω_0 is the resonant angular velocity. Assuming that the transform ratio is equal to the turn ratio, the number of turns n_1 in the primary is deduced from the number of turns n_2 in the secondary through

$$n_1 = \frac{n_2}{k}$$

For example,

let $B = 4.5$ mc,

$C = 12 \mu\text{mf}$,

$Z_1 = 300$ ohms,

to match the standard lead-in impedance. The secondary impedance is

$$Z_2 = 10^6/\pi \times 4.5 \times 12 = 6000 \text{ ohms}$$

approximately, so that the transform ratio is

$$k = (6000/300)^{1/2} = 4.47$$

Thus it is seen that the input transformer brings in an interesting increase in signal level in addition to matching the impedances of the lead-in cable and the input circuit.

6-3. Double-tuned input

The input transformer might have its primary tuned by a suitable capacitance, as shown in Fig. 6-2, where the coupling element is the double-tuned trans-

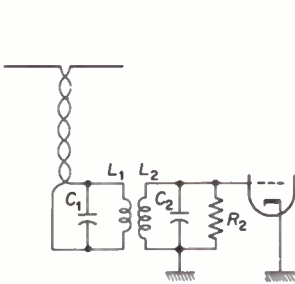


Figure 6-2. Double tuned input transformer.

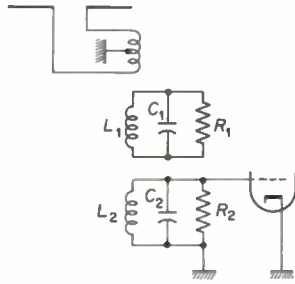


Figure 6-3. Intermediary tuned circuit input transformer.

former previously analyzed. For the same bandwidth this circuit should theoretically provide twice as much gain as the preceding autotransformer.

Another interesting possibility is illustrated in Fig. 6-3. An intermediary tuned circuit $L_1-C_1-R_1$ ensures the coupling between antenna input and grid-tuned circuit. Theoretically, the gain is higher than for a double-tuned transformer and the adjacent selectivity, that is, the slope of the sides of the response curve, is better. Tight coupling between the windings may be somewhat of a problem.

For comparison Fig. 6-4 shows the response of the three preceding circuits. Curve 1 applies to the single-tuned transformer or autotransformer input. Curve 2 refers to the double-tuned transformer. Curve 3 pertains to the three-winding circuit.

Generally, the primary winding has a center tap. This tap is grounded to balance the lead-in line. With this arrangement, parasitic voltages collected by

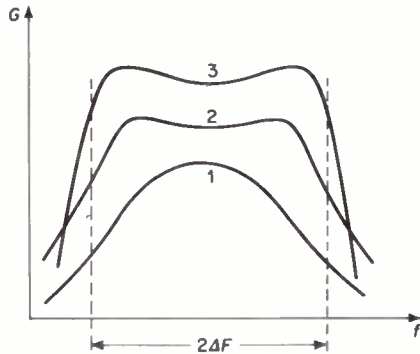


Figure 6-4. Response curves for three different types of input circuits.

the line, which are equal and in phase in the two conductors of the line, are applied in opposition between each end of the primary and ground and theoretically cancel each other in the secondary.

6-4. Noise

The simplest way to increase the sensitivity of a receiver is to increase the gain—by adding extra amplifying stages, for example.

Unfortunately, a limit is soon reached because of the noise. Notice that the use of the term “noise” might seem improper when it applies to a visual picture, although it is in common usage. Noise manifests itself on a picture by a large number of small black and white points, flickering all over the screen and variously described as “snow” or “salt-and-pepper.”

Noise has several distinct origins. The common sources are

- (a) industrial statics,
- (b) atmospheric statics,
- (c) cosmic statics,
- (d) thermal noise in the antenna,
- (e) thermal noise in the receiver.

Industrial statics are due to industrial electric or electronic machines or systems. Atmospheric statics are due to any form of atmospheric electricity, not necessarily in the form of lightning. Cosmic statics are electromagnetic radiations originating in distant parts of the universe. Thermal noise in the antenna is due to thermal agitation of electrons in the equivalent resistance of the antenna. Thermal noise in the receiver is due to electron thermal energy in the circuits and in the tubes.

It is immediately apparent that the first three causes of noise are exterior to the receiver and difficult to eliminate. The best that can be done at the receiver is to reduce their effect through the use of suitable circuits. In the case of industrial statics, parasitic signals can be very effectively suppressed at the source itself if it can be identified and if it can be reached.

Cosmic noise decreases as the cube of frequency as a first approximation. For practical purposes it can be neglected for frequencies on the order of 200 mc or more.

Thus, only the antenna and receiver noises remain to effectively limit the intrinsic performance of the receiving installation. Both antenna and receiver noises are of thermal origin. They are created by the random thermal movement of electrons in resistances. This agitation decreases with temperature and reaches zero for a temperature of -273°C , that is, at absolute zero; at this point thermal noise disappears. Since reducing the temperature of the antenna and receiver is hardly practicable, we must find other means of reducing thermal noise.

6-5. Thermal noise

Thermal noise, or hereafter more simply *noise*, then, is the result of the random movement of the electric charges, carried by the electrons, which produce voltage fluctuations across any resistance through which they move. Because of their very origin, these fluctuations have no definite frequency but appear equally distributed over the whole frequency spectrum.

A direct consequence of great practical importance is that the total noise increases with the bandwidth. Since television uses large bandwidths, the problem is acute.

The classical Nyquist expression

$$E = (4kTRB)^{1/2}$$

gives the rms noise voltage E in volts across a resistance R in ohms as a function of the bandwidth B in cycles per second and the absolute temperature T in degrees Kelvin.

k is Boltzman's constant, given by

$$k = 1.38 \times 10^{-23} \text{ joule/}^\circ\text{K}$$

The absolute temperature is given by the number of degrees centigrade plus 273. For example, the standard room temperature is 27°C or

$$27 + 273 = 300^\circ\text{K}$$

At standard room temperature, for a bandwidth of 4.5 mc, Nyquist's expression becomes

$$E = 27.3 \times 10^{-8} \times R^{1/2}$$

More simply, with E in microvolts and R in ohms,

$$E_{\mu v} = 0.273 \times R^{1/2}$$

For example, a resistance of 300 ohms at room temperature would give a noise voltage of

$$E = 0.273 \times 300^{1/2} = 4.7 \mu v$$

Nyquist's expression applies whether the resistance is real or only an equivalent resistance. Thus, a tuned circuit at resonance reduces to its equivalent shunt resistance and the noise can be calculated accordingly.

Figure 6-5 shows the noise voltage in microvolts as a function of the resistance in ohms for the standard bandwidth of 4.5 mc and the standard room temperature of 300°K or 27°C .

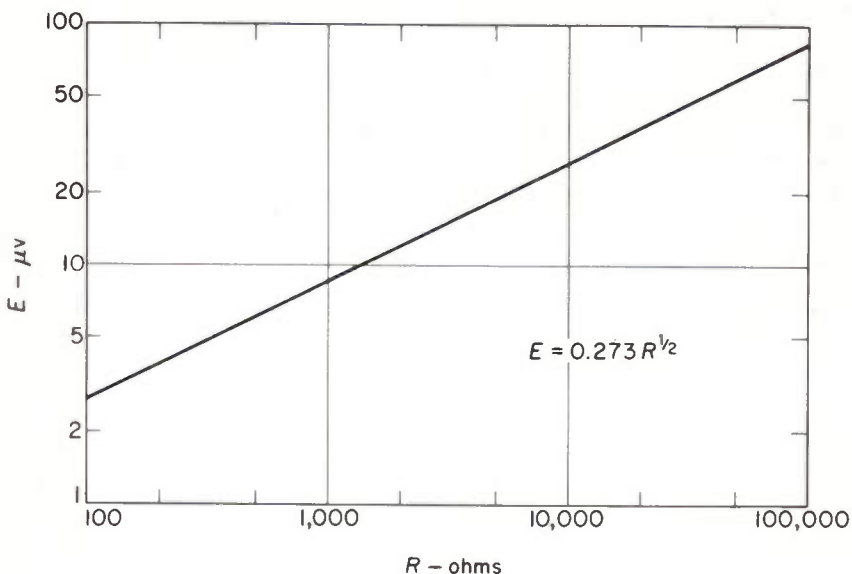


Figure 6-5. Graph for the evaluation of noise voltage.

6-6. Tube noise

The importance of tube noise depends on the tube's position in the amplifying chain. As soon as the gain of a stage exceeds, say 10, which is the general case, it is evident that the noise created before this stage is multiplied by the gain of the stage and becomes much more important than the noise created after the stage. In common receivers, this means that by far the most important noise is the noise created before the first stage. Consequently, all efforts to reduce noise will be applied to the input circuits and to the input tubes.

Sometimes, if the gain of the first stage is not too high, the noise contributed by the second stage must also be considered.

A tube produces noise principally for two reasons. First, electrons fall one at a time on the anode and produce the so-called *shot effect*, analogous to the noise produced by numerous droplets of rain falling on the ground. Second, if the electron beam is shared among several electrodes, the distribution of current will also take place with discontinuities on the electronic scale, giving rise to what is called the *partition noise*. This noise will appear every time there is a positive grid, as in pentodes, and it can generally be said that the more positive electrodes a tube has, the more noise it produces. Consequently, triodes, having no partition effect, will be less noisy than pentodes. Hence, we see the advantage of using triodes in the input stages, where noise must be reduced by all available means.

6-7. Noise resistance

The noise produced by a tube does not differ from the noise created by a resistance. It is convenient to imagine the tube noise as being produced by a fictitious resistance inserted in the grid circuit of the tube which is considered as noiseless. This resistance is the noise equivalent resistance R_s , which depends on the transconductance and on the eventual distribution of current between positive electrodes. Its value may be reasonably approximated by the expressions which follow.

Triode

Amplifier:

$$R_s = \frac{2.5}{g_m}$$

Mixer:

$$R_s = \frac{4}{g_c}$$

Pentode

Amplifier:

$$R_s = \frac{I_a}{I_a + I_s} \left(\frac{2.5}{g_m} + \frac{20I_s}{g_m^2} \right)$$

Mixer:

$$R_s = \frac{I_a}{I_a + I_s} \left(\frac{4}{g_c} + \frac{20I_s}{g_c} \right)$$

Multigrid mixer tubes

$$R_s = \frac{20I_a(I_k - I_a)}{g_c^2 I_k}$$

In all the above expressions,

g_m = transconductance in amperes per volt,

g_c = conversion transconductance in amperes per volt,

I_a = anode current in amperes,

I_s = screen current in amperes,

I_k = cathode current in amperes.

Common values would range between 200 and 1000 ohms for amplifying triodes, 800 and 4000 ohms for mixer triodes, 700 and 2000 ohms for amplifying pentodes, and 2000 and 6000 ohms for mixer pentodes.

It is obvious that the advantage is very marked in favor of the triodes. This more than balances the difficulties of using triodes as r-f amplifiers

because of the high plate-to-grid capacitance, and current designs use triodes almost exclusively in the amplifying stages before the mixer, resorting to special circuits as discussed below.

Roughly, the same tube is four times more noisy as a mixer than as an amplifier. This again points to the usefulness of having r-f stages before the mixer, even if the mixer is a triode. To give the reader an idea of the practical values of equivalent noise resistances, Table 6-1 lists the values in ohms for four triode and three pentode tubes.

Table 6-1. Equivalent Noise Resistance

<i>Tube</i>	<i>Amplifier</i>	<i>Mixer</i>
6C4	1140	4500
6J4	210	800
6J6	470	1800
12AT7	440	1700
6AK5	1880	8400
6BX6	1200	5500
6AC7	720	3200

6-8. Signal-to-noise ratio

Where reception is difficult because of a weak signal, the absolute magnitude of the signal is much less important than its relative magnitude with respect to noise.

For example, consider a 50-microvolt signal applied to two receivers. The first receiver has a noise of 50 microvolts at input, and the signal is drowned in the noise. The second receiver has a noise of 10 microvolts at input, and the signal is easily distinguished above the noise.

It is the signal-to-noise ratio, S/N , therefore, which determines the possibilities of reception. As we have said before, its value should be at least 10, which corresponds to 20 db, for good reception in television.

The best practical way to obtain a good S/N ratio is to use two low-noise triodes as r-f amplifiers before the mixer tube.

With good r-f amplification, if the signal is still drowned in noise, it is useless to increase the amplification because this would not change the (only) important factor S/N . It is even somewhat deceptive to attempt to reduce the bandwidth. Nyquist's fundamental equation shows that the noise voltage varies only as the square root of bandwidth. Consequently, to increase the S/N ratio by 6 db, the bandwidth should be reduced by a factor of 4. Furthermore, the grain of the noise on the screen (that is, the average size of the noise dots on the picture) increases when the bandwidth is reduced.

Given good r-f amplification, the best way to improve reception is to increase the signal level at input by improving the antenna system.

6-9. Load matching

Figure 6-6 shows a generator of internal resistance R delivering a voltage E and loaded by a resistance X . This diagram may represent the equivalent generator of an antenna feeding the input of a receiver.

What should be the value of X for the useful power, that is, the power dissipated on X , to be a maximum?

The current I through the circuit is

$$I = \frac{E}{R + X}$$

The power on X is

$$P = XI^2 = \frac{XE^2}{(R + X)^2} = E^2 \frac{X}{(R + X)^2}$$

To find the maximum value of P , take the derivative of P with respect to X :

$$\frac{dP}{dX} = E^2 \frac{R - X}{(R + X)^3}$$

This derivative goes through zero and P goes through a maximum for

$$X = R$$

This shows that a generator will deliver a maximum power to a load when this load is equal to the internal impedance of the generator. In this case, the voltage across the load is only $E/2$, which means a 50 per cent loss in signal voltage.

From a noise point of view, the signal voltage must be maximum, so that it might well be interesting to sacrifice some efficiency by deliberately mismatching the load to obtain a better S/N ratio.

Higher signal voltage across the load would be obtained for a load X higher than the optimum value R . If the efficiency does not decrease too rapidly when X/R increases beyond 1, then the scheme is worthwhile.

6-10. Load mismatch

For any value of X the power delivered to X is given by

$$P = XI^2 = X \left(\frac{E}{R + X} \right)^2$$

As shown before, this power is maximum for $X = R$ and is

$$P_{\max} = \frac{E^2}{4R}$$



Figure 6-6. Generator feeding a load X .

Let $k_1 = X/R$ and $y = P/P_{\max}$ to normalize the variables. Dividing P by $P_{\max}$ provides

$$y = \frac{4X/R}{(1 + X/R)^2} \quad \text{or} \quad y = \frac{4k_1}{(1 + k_1)^2}$$

Figure 6-7 shows how y varies as a function of k_1 . There is a maximum for $k_1 = 1$, but this maximum is not sharp and for values of k_1 larger than 1, the efficiency is still good. It drops to only 89 per cent, that is, 2 db for $k_1 = 2$ as shown in Fig. 6-7.

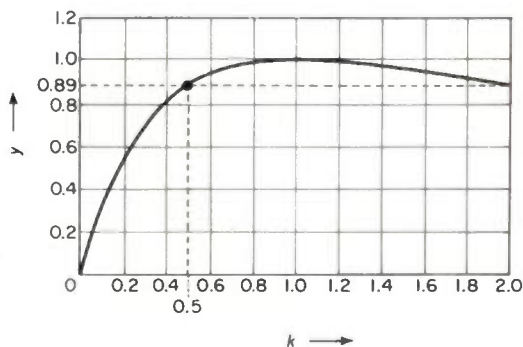


Figure 6-7. This curve shows how efficiency varies as a function of mismatch.

It is then feasible to use $k_1 > 1$ to obtain a better S/N ratio. The limit in this case is set by the possibility of reflections if the mismatch is too important.

6-11. Noise power

The noise produced by the generator is given by Nyquist's expression,

$$E_n^2 = 4kTRB$$

and the maximum noise power which will appear on the load will occur for matched load and will be

$$P_n = \frac{E_n^2}{4R} = kTB$$

This expression shows that in this case the fraction of generator noise delivered to the load is independent of resistance and will, of course, be the signal-to-noise ratio in power for matched case. This ratio, therefore, will not be modified by any impedance transformation (without losses) that will take place in the circuit.

Such impedance transformations are precisely what is used at the antenna input, whether the circuit uses a transformer (Fig. 6-8) or an autotransformer (Fig. 6-9). With a transform ratio

$$n = \left(\frac{R_2}{R_1} \right)^{1/2}$$

the secondary impedance R_2 is reflected in the primary as R_1 , the generator sees a load equal to its internal resistance, the impedances are matched, and the efficiency is maximum.

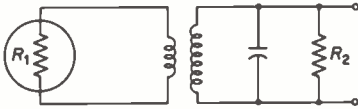


Figure 6-8. Equivalent input circuit for a transformer.

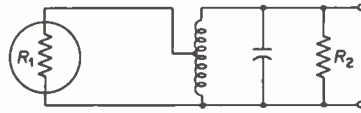


Figure 6-9. Equivalent circuit for an autotransformer.

However, the noise power is transferred to the secondary in the same manner. In fact, the antenna resistance reflected in the secondary is equal to the secondary resistance, and the antenna brings in just as much noise as the circuit.

As we have said before, an improvement may be effected in this respect by reducing the antenna's contribution to the total noise. Deliberate mismatching, by using a transformer ratio n different from optimum, will reduce the noise at secondary due to the antenna at the price of a loss in gain. The overall result is nevertheless a better S/N ratio.

Spectacular improvement by mismatching, however, should not be expected. The input tube adds a major noise contribution which is not modified by whatever is done in the circuit. All in all, the noise due to the antenna is only a fraction of the total noise of the stage, and the apparent effect of antenna noise reduction will be reduced accordingly.

The resulting total noise voltage V_t of several noise sources $V_1, V_2, V_3, \dots$ is not obtained by simple addition, but according to the rms law:

$$V_t^2 = V_1^2 + V_2^2 + V_3^2 + \dots$$

The simplest way to overcome this difficulty is to transform all noise voltages into the equivalent noise resistances. These resistances can be combined in the usual way, and the total resulting resistance gives the total noise voltage by application of Nyquist's formula.

6-12. Pentode r-f amplifier

Pentodes are easier to use than triodes and are preferred as r-f amplifiers if a very high signal-to-noise ratio is not required, that is, when the signal has sufficient amplitude.

The input circuit is generally a transformer or autotransformer as before. The coupling circuit to the following tube may be a single-tuned circuit, a double-tuned transformer, or sometimes a series resonant circuit. The anti-resonant circuit is used as indicated Fig. 6-10. The parasitic capacitances tune it to resonance, the exact tuning being obtained by adjusting the inductance value as indicated, generally with the help of an iron or brass core. The grid-leak resistance R is simultaneously used as damping resistance for the tuned circuit in combination with the input resistance of the tube. The coupling capacitance C_1 should have negligible reactance at the resonant frequency.

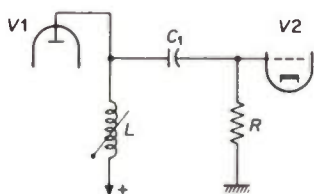


Figure 6-10. Tuned coupling circuit in the anode.

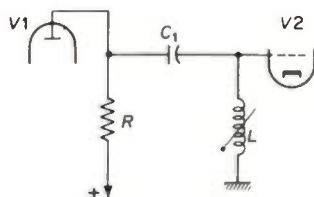


Figure 6-11. Tuned coupling circuit in the grid.

A variant of the circuit is shown in Fig. 6-11, which places the resistance in the anode circuit and the inductance in the grid circuit. Plate current now flows through the resistance, but the inductance has one end grounded, which may be convenient. Furthermore, the grid d-c time constant is much shorter.

In television the number of amplifying stages is large, and it is customary to use separate decoupling circuits in the plate supply to each stage to reduce the possibility of instability due to coupling between stages through a common lead. Some pentodes use the same voltage on the screen as on the plate, and a common decoupling circuit is used as shown in Fig. 6-12. Other pentodes need a lower screen voltage, and a dropping resistor R_2 with decoupling capacitor C_2 is used as shown in Fig. 6-13. Notice that R_2 is connected to the top of R_1 so that the screen supply is doubly decoupled.

The third or suppressor grid is grounded directly for better stability. When cathode bias is used as shown, the cathode resistance is decoupled by a capacitance large enough to offer negligible reactance at the working frequency.

The resistance across the inductance, in combination with the tube input resistance, provides the necessary bandwidth.

The double-tuned transformer shown in Fig. 6-14 is almost identical with the preceding transformer except for the coupling circuit. The same arrangements and the same variants are used.

The same remarks can be made for the series-tuned circuit shown in Fig. 6-15. The series inductance L is tuned to resonance by the parasitic capacitances

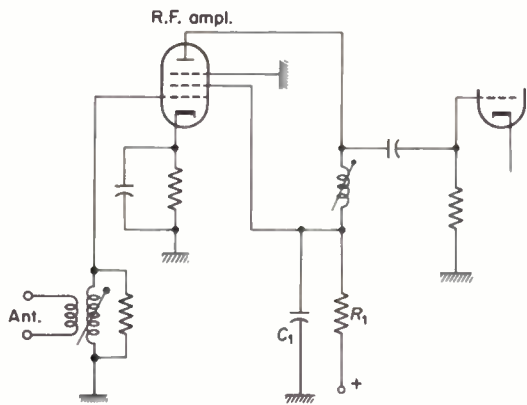


Figure 6-12. Screen directly connected to the anode voltage supply.

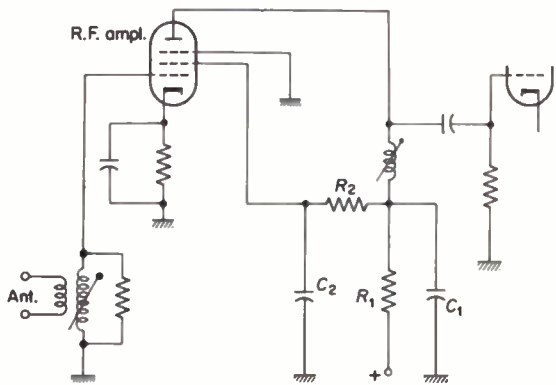


Figure 6-13. Separate decoupling circuit for the screen.

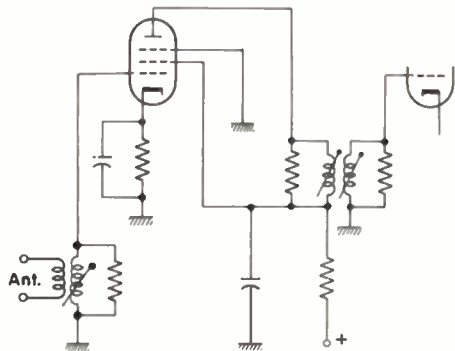


Figure 6-14. Double-tuned transformer coupling circuit.

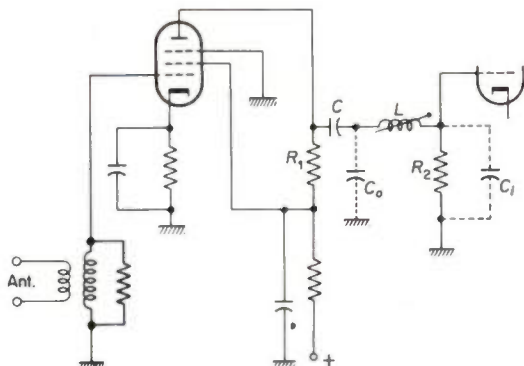


Figure 6-15. Series inductance coupling circuit.

indicated by dashed lines. C_o is the output capacitance of the r-f tube, and C_i is the input capacitance of the following tube.

C is a coupling capacitor of high value, having negligible reactance at the working frequency and insulating the grid from B+. The necessary damping is provided by R_1 and R_2 .

The advantage of the series circuit is that the effective capacitance across the inductance L is smaller than for the shunt resonant circuit shown in Fig. 6-12. The gain is therefore larger for same bandwidth. Furthermore, inductance L must have a larger value in the series case to resonate on any given frequency. This may be a practical advantage at high frequencies.

6-13. Triode r-f amplifier

The use of triodes as r-f amplifiers is very interesting because of the low noise they produce. Difficulties arise because of the relatively high parasitic plate-to-grid capacitance C_{gp} , which couples the output and input circuits and may provoke instability or even oscillation.

The geometric plate-to-grid (cold) capacitance is that of a capacitor having the areas of the grid and the plate for electrodes and the vacuum as dielectric. Its value is on the order of $1 \mu\mu\text{f}$, but the actual working capacitance is much larger because of the Miller effect.

If a voltage of $+1 \text{ v}$ is applied to the grid of the cold (not working) triode, then the total voltage across the grid-to-plate capacitance is 1 v . However, if the same voltage is applied to the grid of the working triode having a stage gain K , a voltage $-K \text{ v}$ will appear at the plate. The voltage across the grid-to-plate capacitance is now

$$1 - (-K) = (K + 1)\text{v}$$

Therefore, the simple fact of putting the tube to work has multiplied by $(K + 1)$ the apparent grid-to-plate capacitance. This is the *Miller effect*.

Through this capacitance part of the output voltage is fed back to the input. The classical way to take care of it is to feed back to the input an equal voltage, but of opposite phase, so that the two voltages reaching the input cancel out. The method is called *neutralizing*, and it can be applied in a number of ways. As shown in Fig. 6-16, the plate inductance constitutes in fact a transformer because of the B+ tap, and the signal at its low end has a phase opposite to the phase of the signal at the high end, which is the anode. The low-end signal is then suitable for neutralization and is applied to the grid through capacitance C , which is adjusted for exact neutralization.

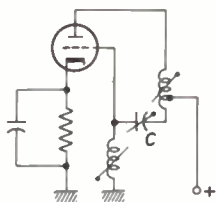


Figure 6-16. Neutralization of a triode through C .

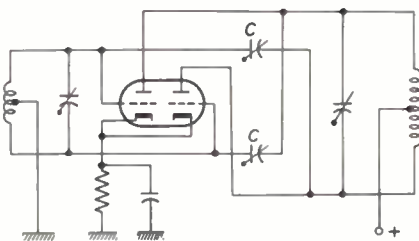


Figure 6-17. Symmetrical neutralization.

A convenient variation is shown in Fig. 6-17 for a symmetrical double-triode stage. In this circuit the signals at the plates are of opposite phase and suitable for the neutralization of the opposite triode. Plate-to-grid cross coupling by the capacitances C is all that is needed, and adjusting the capacitances provides exact neutralization.

Although neutralized triodes do work well, there is a general tendency to use other triode circuits which avoid the necessity for neutralizing and the resulting critical adjustment, or which provide better over-all performance.

6-14. Triode circuits

To eliminate the coupling between output and input due to the plate-to-grid capacitance, two methods can be used besides neutralizing: grounding the plate or grounding the grid—for the r-f signal.

The first method is represented by the circuit shown in Fig. 6-18. The anode is effectively grounded for r-f by the capacitance C . The output circuit is driven by the cathode, with R_1C_1 being the usual cathode biasing arrangement. This cathode-loaded circuit does not provide any gain; on the contrary, the signal voltage on the cathode output is lower than the signal voltage on the grid input. However, the output impedance is low and the cathode can be tapped down on the output circuit as shown. There is thus some gain due

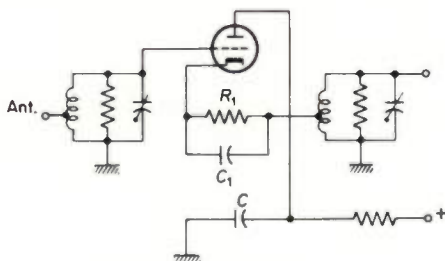


Figure 6-18. Grounded anode circuit.

to the transform ratio of both input and output circuits. A variation is illustrated in Fig. 6-19, in which the output circuit is untuned or aperiodic, an r-f choke (RFC) providing the r-f cathode load.

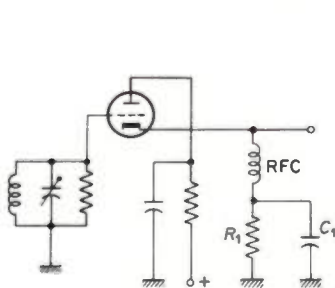


Figure 6-19. Untuned output cathode follower.

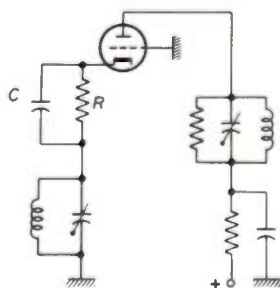


Figure 6-20. Grounded grid circuit.

The second method corresponds to the circuit shown in Fig. 6-20. The grid is directly grounded, and the input is at the cathode. In this fashion, the grounded grid acts as an electrostatic screen between the cathode input and the plate output. A peculiarity of this circuit is the low input impedance at the cathode, as a first approximation equal to $1/g_m$. For the high-transconductance type of triode used here, the cathode input impedance would be on the order of 100 to 300 ohms. A direct consequence is heavy damping of the input-tuned circuit, with resulting poor selectivity and low gain.

The input circuit, therefore, is not too useful and can be suppressed without much loss leading, to the diagram shown in Fig. 6-21 where the input is untuned or aperiodic across a resistance R . This resistance, in parallel with the tube cathode input

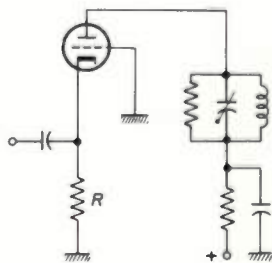


Figure 6-21. Untuned input grounded grid amplifier.

resistance, is chosen to match the impedance at input. Also, it can provide all or part of the bias. This circuit is sometimes used at the input of a receiver.

Another variation uses two grounded-grid triodes in a symmetrical arrangement, as shown in Fig. 6-22, which better suits a symmetrical lead-in line.

Figure 6-23 shows a circuit which is a hybrid between the grounded cathode and the grounded grid. The input circuit is grounded at a tap and drives the cathode and grid in a push-pull fashion. Since cathode and grid receive r-f signals of opposite phase, the total effect of the parasitic capacitances with

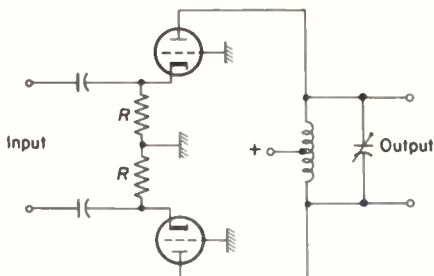


Figure 6-22. Symmetrical grounded grid amplifier.

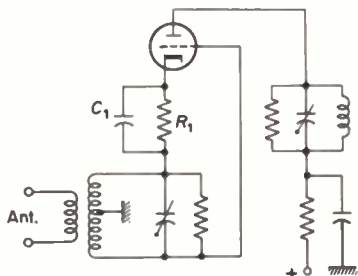


Figure 6-23. Grid and cathode input.

respect to the plate can be reduced by careful positioning of the grounded tap on the input circuit.

The preceding circuits can be used together, especially since a number of suitable tubes belong to the double-triode variety where two triodes are mounted in the same glass envelope.

For example, cascading the circuits of Figs. 6-19 and 6-21 leads to the diagram of Fig. 6-24, where there is a cathode-to-cathode coupling between the two tubes by the r-f choke (RFC). T_1 is the input circuit and T_2 the output circuit.

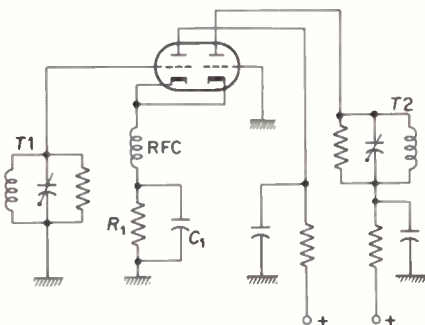


Figure 6-24. Cathode to cathode coupling.

A clever sophistication of the scheme appears in Fig. 6-25. The unused grid and plate of Fig. 6-24 have been put to work with input at T_3 and output at T_4 . The circuit now provides two separate amplification chains, one from T_1 to T_2 , and the other from T_3 to T_4 . These chains may be used for widely different frequencies, such as r-f and i-f. The adjustment of the circuit is somewhat critical.

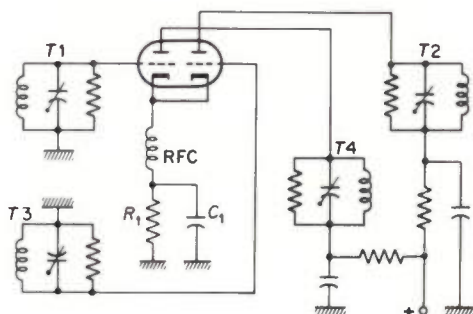


Figure 6-25. Double amplification in one stage.

6-15. Cascode

Miller effect is troublesome with the grounded-cathode triode because of the gain of the stage. If this gain is kept low, then the circuit can be used without

trouble. A practical application of the idea is illustrated by the cascode circuit shown in Fig. 6-26. The first triode is a classical grounded-cathode triode. It drives the cathode of the following stage, whose impedance is approximately $1/g_m$ and loads the first stage. Assuming both stages have identical transconductances, the gain of the first stage is roughly

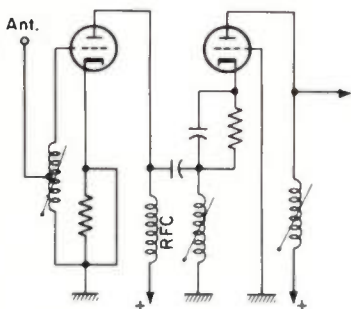


Figure 6-26. Theoretical diagram of a cascode circuit.

$$K = g_m \frac{1}{g_m} = 1$$

and Miller effect in this first stage can be ignored. A voltage gain of 1 may not seem too useful. However, the first stage has an output impedance much lower than its input impedance and provides a sizable power gain.

Assuming the load of the second stage to be R , the gain of the second stage, and also the total gain, is

$$K_t = g_m R$$

This gain can be high and the circuit is stable. There is some amount of negative feedback in the second stage because of the coupling impedance in its cathode circuit.

Because of the power gain of the first stage and the negative feedback in the second stage, the total noise factor is good and practically equal to the noise factor of the first stage.

The circuit can be improved if neutralizing of the first tube is used. Figure 6-27 shows a circuit which accomplishes this using a neutralizing inductance L . This neutralization is not critical and does not reduce the stability. Its

adjustment is very broad, and in many cases a fixed value inductance is used. A number of variations of the neutralizing circuit are used. Notice how the arrangement shown has taken care of the d-c voltages.

6-16. Direct-coupling cascode

Direct coupling between first anode and second cathode may be used with special triodes. As far as d-c supply is concerned, the two tubes are in series as shown in Fig. 6-28, and they share the supply voltage.

In the circuit shown the second triode has no cathode bias. The grid-leak resistance has a large value and the contact potential provides sufficient bias for the grid. The circuit is simple, and the lack of any tuned element in the coupling makes for a wide bandwidth.

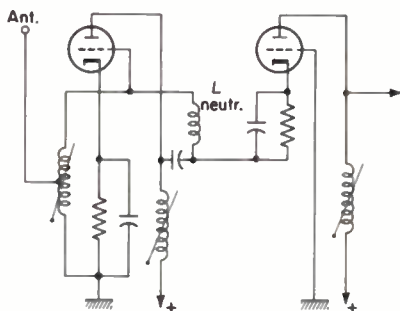


Figure 6-27. Neutralization of a cascode.

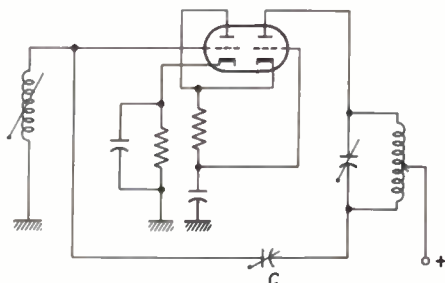


Figure 6-28. Direct-coupled cascode.

As explained before, neutralizing is used to obtain better performances. In this case, the tap on the output circuit provides at its low end a signal of suitable phase to be applied to the input grid through the neutralizing capacitance C .

The basic circuit of the cascode can be modified in numerous ways to obtain a number of variations. Some may offer practical advantages, but they all rely on the same fundamental principles.

6-17. Remote cutoff tubes

Remote cutoff or variable slope (variable μ) tubes are commonly used in r-f amplifying stages. It will be seen later that some kind of automatic gain

control, or AGC, is included in the television receiver. It is applied to the r-f as well as to the i-f amplifying tubes and for proper operation requires variable slope tubes.

In addition to contributing usefully to the over-all AGC efficiency, an AGC-controlled r-f stage is instrumental in reducing distortion with large input signals because it eliminates or reduces saturation of succeeding stages.

Problems

1. A low-loss lead-in line of 75-ohm characteristic impedance is matched to the input circuit through a transformer arrangement. The input-tuned circuit has a shunt capacitance of $12\ \mu\text{f}$ and a 3-db bandwidth of 4 mc. Calculate the gain of the input transformer.

2. A UHF lead-in line of 400-ohm characteristic impedance is coupled to a tuned input circuit having a shunt capacitance of $3.8\ \mu\text{f}$ and a bandwidth of 6 mc at 3 db. Evaluate the gain of the input circuit.

3. Calculate the thermal noise produced by a tuned circuit having a shunt capacitance of $18\ \mu\text{f}$ and a 3-db bandwidth of 4 mc at room temperature.

4. A tuned circuit has a shunt capacitance of $15\ \mu\text{f}$ and a 6-db bandwidth of 5 mc. Calculate its noise in its 3-db bandwidth at room temperature.

5. A tuned circuit has a shunt capacitance of $16\ \mu\text{f}$ and a 3-db bandwidth of 4.5 mc. Calculate its thermal noise when the receiver is switched on.

One hour after switching on, the temperature of the tuned circuit has increased by 18°C . Evaluate the increase in noise.

6. A triode tube has a transconductance of 6 ma/v. Calculate its noise at room temperature when used as a mixer and as an amplifier, with a 4.5-mc bandwidth.

7. A pentode has a cathode current of 12 ma and a plate current of 10 ma. Its transconductance is 9 ma/v as an amplifier. Calculate the noise it produces at a temperature of 45°C as an amplifier and as a mixer, with a bandwidth of 4.5 mc.

8. Compare the noise of a triode and a pentode amplifier having the same transconductance of 6 ma/v at room temperature and at 50°C . The pentode has a plate current of 9 ma and a screen current of 1.2 ma. The bandwidth is 6 mc.

9. Use a tube manual to find characteristics of three triodes and three pentodes for television. For each of them evaluate the equivalent noise resistance as amplifier and mixer. Calculate the noise they produce over a 4.5-mc bandwidth at room temperature.

10. An r-f pentode stage uses a tube having a transconductance of 6 ma/v, a plate current of 8 ma, and a cathode current of 10 ma. The bandwidth is 4.5 mc. The input circuit has a shunt capacitance of $14\ \mu\text{f}$. With antenna disconnected, calculate the noise due to the first stage, circuit and tube. Evaluate the improvement when the pentode is replaced by a triode with the same transconductance and parasitic capacitances.

An antenna having an equivalent noise resistance of 300 ohms is connected and injects a 45-microvolt signal. Evaluate the signal-to-noise ratio for the pentode and the triode stage. Assume room temperature.

11. An antenna having an equivalent resistance of 300 ohms is transformer-coupled, to match impedances, to a circuit having a 4-mc bandwidth and 12- μmf shunt capacitance. The signal in the antenna is 120 microvolts. Evaluate the signal-to-noise ratio when the antenna is directly connected to the tuned circuit and when it is connected through the transformer.

12. A generator of 6v emf and internal impedance of 5 ohms feeds a variable load R . Evaluate the power delivered to the load when R is equal to 1, 2, 4, 5, 6, 8, 10, and 20 ohms.

Plot the resulting curve.

Calculate the allowable range of R for a power in R less than 3 db below maximum.

13. Three noise sources are connected, directly or indirectly, to the grid of an input tube. One is due to the antenna and brings in 16 μv noise. The second is due to the tuned circuit and brings in 11 μv noise. The third is due to an additional resistance and brings in 26 μv noise. Calculate the total noise on the grid.

14. The third or suppressor grid of a pentode r-f amplifier can be connected to ground or to the cathode. Compare the two methods.

15. Discuss the advantages of a series-tuned coupling circuit with respect to a parallel-tuned coupling circuit.

16. A triode stage has a bandwidth of 4.5 mc, a total output capacitance of 10 μmf , an input capacitance of 6 μmf , and a grid-to-plate capacitance of 1.1 μmf . Neglecting the input resistance of the tube, calculate the damping resistance of the input circuit. The tube has 5.5 ma/v transconductance, and its plate resistance can be neglected.

17. Design a single-tube neutralizing circuit using a capacitance bridge and explain its operation.

18. Draw the equivalent bridge circuit for a symmetrical double-triode amplifier using cross neutralizing and derive the condition for neutralization.

19. Draw a grounded-grid triode r-f amplifier circuit, including all parasitic capacitances, and explain why neutralizing is unnecessary.

20. An r-f amplifier uses two triode tubes. Draw the circuits of all possible combinations, paying attention to the d-c voltage levels, and briefly explain operation.

21. In a cascode, the bandwidth is 4 mc and the first stage total input capacitance is 6 μmf . The transconductance is 5 ma/v. Assume the input impedance of the second stage is 200 ohms. Evaluate the power gain of the first stage.

22. The second stage of a cascode has a total output capacitance of $9 \mu\mu\text{f}$ and a bandwidth of 4 mc. The tube transconductance is 5 ma/v. Calculate the voltage gain of the stage.

23. A cascode uses two identical triodes of 8 ma/v transconductance, $5 \mu\mu\text{f}$ total input capacitance, $10 \mu\mu\text{f}$ total output capacitance. The bandwidth is 4.5 mc. Evaluate the voltage gain and power gain of the cascode.

24. A lead-in line of 300 ohms impedance is transformer-matched to the input circuit of a cascode. Calculate the gain between antenna input and output of the cascode, which uses two identical triodes of 6 ma/v transconductance, with input capacitance $7.5 \mu\mu\text{f}$ and output capacitance $11 \mu\mu\text{f}$. The bandwidth is 4 mc.

Chapter 7

FREQUENCY CHANGING

The total amplification necessary in a television receiver reaches high values and necessitates a number of amplifying stages. The gain of each stage is limited because of the large bandwidth over which amplification is accomplished. If all this amplification were to take place at the same frequency, the risk of instability would be great. For this reason we resort to frequency changing, so that part of the gain occurs at r-f frequency and part at i-f frequency.

In addition, frequency changing provides a fixed i-f independent of the channel to be received. All i-f circuits are consequently fixed-tuned, with a resulting simplification in i-f amplifier design.

The r-f, mixer, and local oscillator circuits use variable tuning to receive all channels. The corresponding tubes and circuits are generally combined into a single subassembly called a *tuner*.

7-1. Mixer and converter

A *mixer* is a tube which combines the signal coming from the antenna (and eventually through amplifiers) with the signal coming from a separate oscillator. A simple diode may be used as a mixer.

A *converter* is a tube which by itself provides the local oscillator signal to combine with the antenna signal. It is generally a multielectrode tube, i.e., pentode, hexode, or heptode. Converters are extensively used in radio receivers, but some modification is required to adapt conventional converters for television receivers.

When the frequency increases, both the output voltage and the stability of an oscillator decrease. At television frequencies in particular, the oscillating section of a standard converter tube is inadequate.

In the present state of the art the best solution to the problem of obtaining enough output voltage and sufficient stability on television frequencies seems to be the use of a separate oscillator tube (almost always a simple triode) and a separate mixer tube to ensure frequency changing. The mixer is generally a triode or pentode. Frequently, oscillator and mixer are combined in a single-envelope, special-purpose tube, either a double triode or triode-pentode.

The output from the mixer is a signal resulting from the beat of the antenna signal and the local oscillator signal. Its frequency is

$$f_i = f_o - f_r$$

where f_i is the beat, difference, or intermediary frequency, f_o is the local oscillator frequency, and f_r is the frequency received by the antenna. Generally, the oscillator frequency is higher than the received frequency, and the resulting intermediary frequency is 28.75 or 45.75 mc for the picture carrier. The sound carrier, of course, is 4.5 mc below, that is, on 24.25 or 41.25 mc.

7-2. Oscillator

High stability is a desirable quality for the oscillator; however, it is also a costly quality. The usual compromise is to design the oscillator for as good a stability as can be obtained with reasonable cost and to provide a manual fine tuning, or vernier, in order to be able to adjust the exact tuning and to compensate for frequency drifts.

Because of the wide bandwidth of the picture receiver, a small drift in oscillator frequency does not make a noticeable difference in the picture quality. Also, with the almost universal use of the intercarrier sound system, the sound i-f is independent of oscillator frequency. The worst effect of a drift in oscillator frequency is apparent only if the shift is large enough to set the trap circuits off-tune.

Vernier tuning is a cheap and easy remedy to obtain a fine tuning, but how can this tuning be maintained? Oscillator frequency drifts mainly for two reasons: voltage variation and/or temperature variation. Voltage variation is rarely large enough to create an objectionable shift. It could be remedied by precise voltage stabilization, but this method is costly.

Thermal drift may be due to the tube or to the oscillator components. Parasitic capacitances of the tube vary with temperature, but once a stable working temperature is reached, the variations are very small. In addition, their effect can be made negligible by using a large total capacitance across the oscillator circuit and by coupling the tube to the circuit through a small-value capacitance.

The main causes of oscillator frequency shift are the circuit components themselves. They should be located on the chassis in such a way that they

reach the working temperature rapidly to reduce the initial drift period and that they stay at the same temperature to reduce the long-term drift. The oscillator circuit should then be placed as far as possible from any heat-producing element and should be well ventilated. A good way is to build this circuit on a small separate unit, generally with the frequency changer tube included.

A common method to improve thermal stability is to compensate for components having a positive thermal coefficient, which is the usual case, by introducing a special component, usually a capacitor, having a compensating negative temperature coefficient. If the value of the compensating component and its temperature coefficient are properly chosen, good compensation will result over the temperature working range.

7-3. Oscillator circuits

A wide variety of oscillator circuits is known. Two basic arrangements with minor variations are commonly used in television—the Hartley and the Colpitts oscillators.

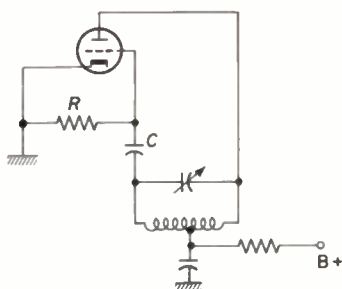


Figure 7-1. Series fed Hartley oscillator.

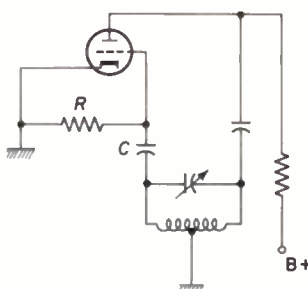


Figure 7-2. Parallel fed Hartley oscillator.

In the Hartley, the 180-degree phase shift between plate and grid necessary for oscillation is obtained through a tap on the oscillator inductance. Various commercial designs differ only in detail. Figure 7-1 shows the series Hartley, where the anode is fed B+ in series through part of the inductance. Figure 7-2 is the parallel or shunt Hartley, where B+ is shunt fed to the anode through a blocking resistance or r-f choke.

A variation of the basic Hartley is the cathode-coupled oscillator shown in Fig. 7-3, where the tuned circuit is in the grid and the cathode is connected to a tap. In all oscillator circuits, grid bias is obtained by grid current with the help of an RC circuit in the grid.

In the Colpitts, the phase shift is obtained through a tap on the oscillator capacitance, as shown in Fig. 7-4. The total tuning capacitance has been split

into two series capacitances C_1 and C_2 , and the common point grounded. Here again, practical arrangements differ in detail. Tuning may be accomplished by varying simultaneously C_1 and C_2 , by adding an extra variable capacitance across the circuit, by adjusting the inductance value, or, between narrower limits, by varying either C_1 or C_2 .

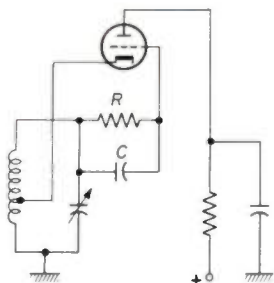


Figure 7-3. Cathode coupled oscillator.

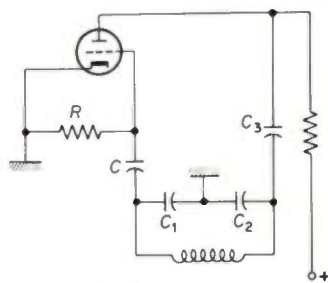


Figure 7-4. Colpitts oscillator.

At television frequencies C_1 and C_2 should have very small values. Since the plate-to-ground C_{pk} and grid-to-ground C_{gk} capacitances happen to be in shunt across C_2 and C_1 respectively, they are sufficient to maintain oscillations and there is no need for physical capacitances across the inductance, as shown in Fig. 7-5.

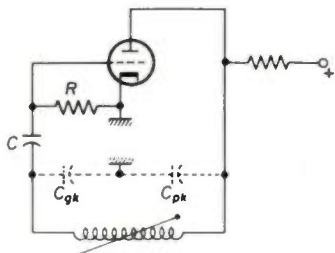


Figure 7-5. Colpitts oscillator for high frequencies, using the tube parasitic capacitances.

The grid coupling capacitance C must be large enough for oscillations to occur with sufficient amplitude and at the same time small enough to reduce the coupling between tube and circuit to increase stability. A practical value is chosen to provide enough oscillation amplitude with some margin of safety, usually 10 to 20 μf . The accompanying grid resistance R has the usual value of 10 to 20 kilohms.

Capacitance C_3 is necessary only to prevent plate voltage from being applied to the oscillator circuit. It is simply a d-c blocking capacitance of, say 100 μf .

The plate decoupling resistance is not critical and may have a value ranging from 3 to 10 kilohms.

7-4. Mixer

In general the mixer tube is a high-transconductance triode or pentode. The transconductance of a tube working as a mixer, or *conversion transconductance*,

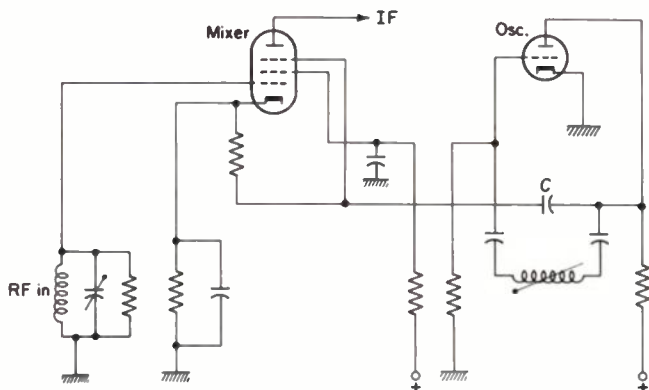


Figure 7-6. Capacitive coupling between oscillator and mixer.

is roughly one-third of the transconductance of the tube when used as a straight amplifier. Since the parasitic capacitances are identical in both cases, the gain of a mixer stage will be

$$K_0 = \frac{g_m R}{3}$$

for a tube of transconductance g_m used as a mixer with a load impedance R .

An optimum conversion transconductance results when the injected local oscillation voltage has a given value, and the coupling between mixer and oscillator has to be adjusted to provide maximum gain.

Oscillation voltage may be injected on any electrode of the mixer tube, but is always applied to the control grid along with the antenna signal in practical application. The advantage is that a low oscillator voltage may be used, 2 to 3 volts being a common value.

The coupling between oscillator and mixer may be capacitive or inductive. In the first case shown in Fig. 7-6 the local oscillator voltage is taken from one end of the oscillator tank circuit and applied to the grid of the mixer through capacitance C . The value of C chosen is as small as possible consistent with enough injected voltage on the grid of the mixer, and a practical value is about $1 \mu\mu\text{f}$.

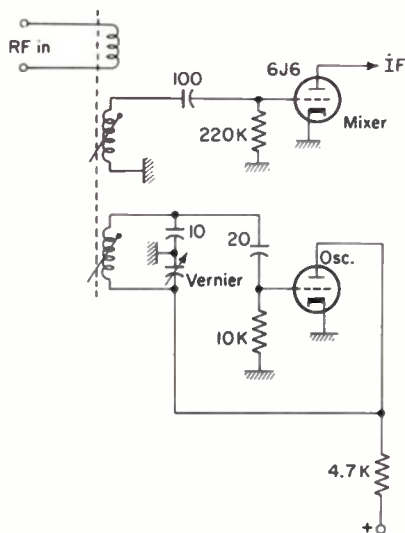


Figure 7-7. Inductive coupling between oscillator and mixer.

The second case is shown in Fig. 7-7 for a double triode 6J6, one-half of which is used as oscillator and the other half as mixer. The antenna signal is inductively coupled to the mixer grid circuit and so is the oscillator signal.

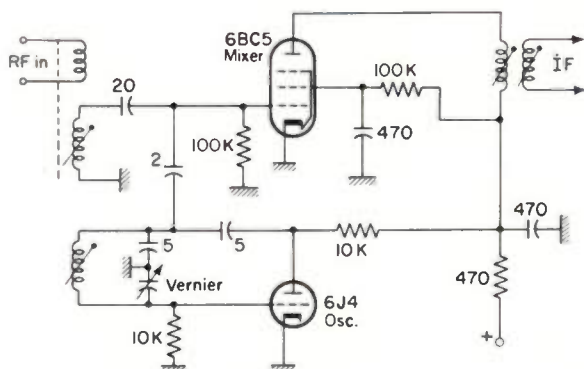


Figure 7-8. Pentode mixer and separate triode oscillator.

The necessary grid bias for the tube to work correctly as a mixer is obtained by using a large value of grid resistance—in this case, 220 kilohms—so that grid current creates a sufficient voltage drop. Alternatively, cathode bias could be

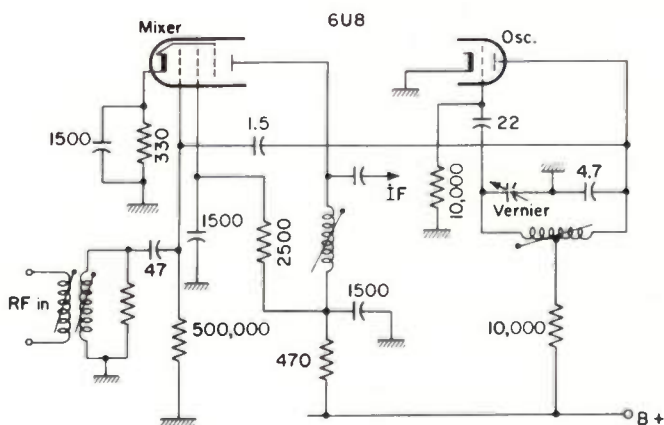


Figure 7-9. Triode-pentode frequency changer stage.

used, or both methods combined. The advantage of using a cathode bias is that in case of failure of injected oscillator voltage, there is still a safety bias left on the mixer.

Figure 7-8 shows a pentode mixer with a separate triode oscillator. The circuit uses a combination of inductive and capacitive coupling for oscillation signal injection.

A variation of this arrangement, shown in Fig. 7-9, uses a triode-pentode tube and capacitive coupling. Notice the combined bias in grid and cathode circuits for the pentode.

A peculiar circuit illustrated in Fig. 7-10 has a single pentode simultaneously used as both an oscillator and frequency changer. Oscillation is produced by a Colpitts circuit between first and second grid. Somewhere along the inductance is a neutral point where the oscillator signal is zero. At this point the r-f signal is injected. The usual vernier tuning and grid-biasing arrangements are provided as in previous circuits.

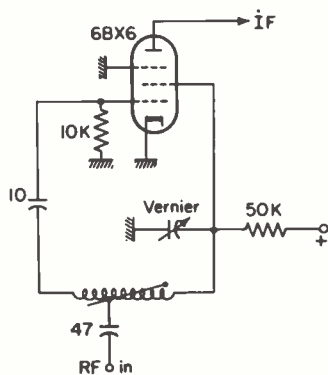


Figure 7-10. Single tube oscillator-mixer.

7-5. Tuning mechanisms

Of course, when receiving different channels, all r-f, mixer, and oscillator circuits must be tuned accordingly; there are several methods available. They can be divided broadly into continuous or discontinuous tuning arrangements.

Continuous tuning systems have either the capacitance or the inductance of each resonant circuit varying over a range of values. All variable elements are ganged together mechanically for ease of control, and the usual tracking components, trimmers or padders, are employed. There is not much difficulty in obtaining good tracking because of the relatively small range of frequency contained in a television band. Figure 7-11 shows a capacitance-tuned circuit. Notice the trimmer capacitances across the tuned inductances. The circuit of Fig. 7-11, as shown, covers only one band. Additional bands could be received by switching in different coils.

Continuous inductance tuning is very similar to capacitance tuning. The mechanical arrangement of the coils, however, must be carefully designed. Spiral coils have been used with success. The advantage of inductance tuning is that the shunt capacitance is not notably increased by the tuning system and is constant over the band.

Continuous tuning has been practically abandoned for the 13 channels of the VHF band, although it was used with some success in early type receivers. However, at UHF, where the number of channels is very large, continuous tuning becomes a decided advantage over switching.

Discontinuous tuning is done by switching over from one tuned-circuit channel to the next. The technique of switching inductances allows each one to be tuned at an optimum for the specific channel it covers and ensures best consistent performance over all channels. A standard multiposition switch can be used for the switching.

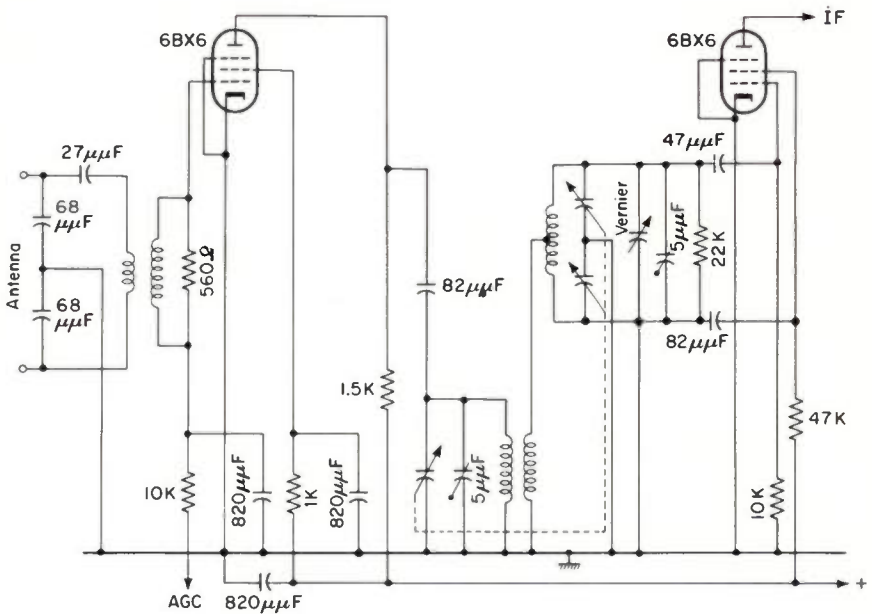


Figure 7-11. Continuous capacitance tuning.

7-6. Tuner circuits

Figure 7-12 illustrates a design using two double triodes, one as a cascode r-f amplifier and the other as an oscillator-mixer. Three channels are indicated for purposes of simplicity, but more can be accommodated depending on the type of switch used.

The input circuit is a double-tuned transformer. The secondary embodies at its ground end a capacitance bridge used for neutralizing purposes. A broad-band untuned inductance is inserted in series in the anode-to-cathode coupling of the cascode to improve performance at high frequencies.

A double-tuned transformer couples the r-f amplifier to the mixer.

The oscillator is a Colpitts with vernier adjustment, and the injection to the mixer is capacitive.

The most common tuning arrangement by far, however, is the turret-type tuner in which inductances are carried by a rotating drum and come into contact with fixed contacts connected to the r-f, mixer, and oscillator circuits. All tubes and components, from antenna input to i-f input, are grouped together in a small unit called the *tuner*. Similar techniques may be used with the other tuning arrangements, the switching mechanism being part of the tuner. In current designs, the set of inductances for one channel constitutes a sub-assembly which clips into the turret and can be removed or inserted in seconds.

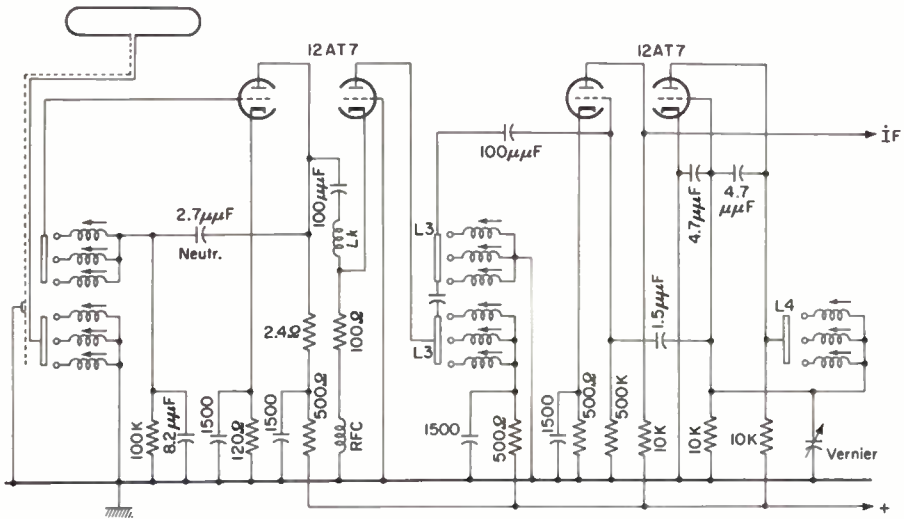


Figure 7-12. Switch type tuner.

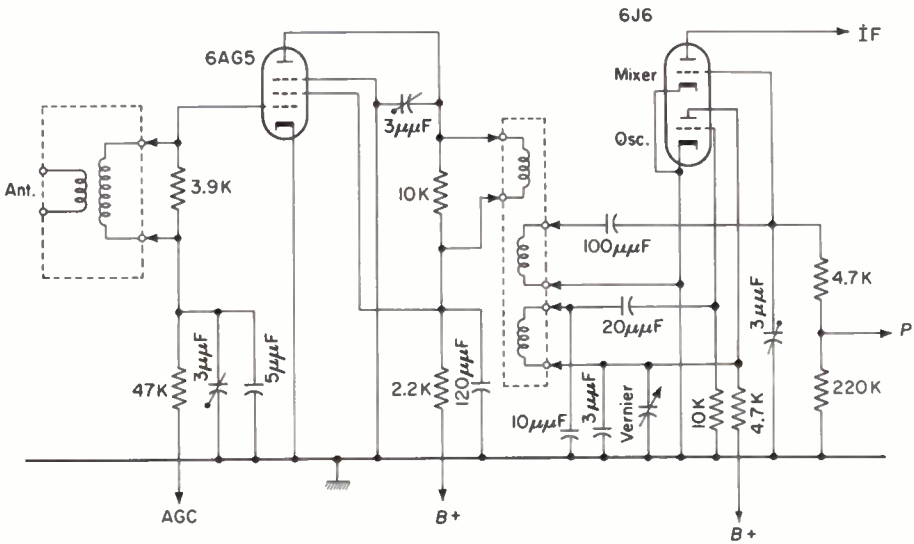


Figure 7-13. Turret type tuner.

Some inductances can be adjusted from the outside. A great variety of circuits can be used. Figure 7-13 shows clearly the ten contact points on the turret tuner. A pentode is used as an r-f amplifier and a double triode as an oscillator-mixer. Point *P* is a test point to check oscillator grid current. Notice the vernier tuning and 3- μf trimmer capacitors to adjust parasitic capacitances if the tubes are changed.

Figure 7-14 has a switching arrangement similar to the previous tuner, but uses a pentode r-f amplifier and a pentode converter.

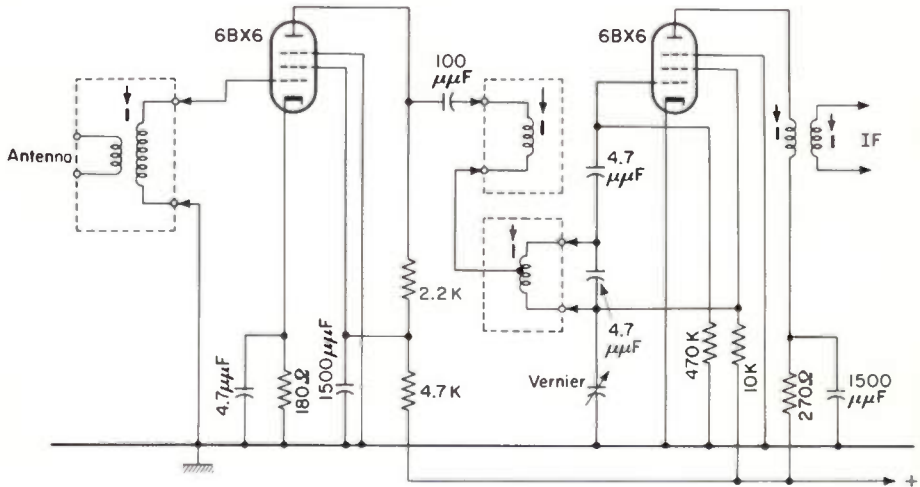


Figure 7-14. Tuner circuit using two pentode tubes.

Figure 7-15 uses a pentode r-f amplifier and a double triode oscillator-mixer. The input circuit is an autotransformer. The coupling circuit between r-f amplifier and mixer is a series type. Local oscillation injection is capacitive.

Figure 7-16 employs a double triode as a cascode r-f amplifier and another double triode as an oscillator-mixer. The input circuit is an autotransformer. The coupling between the two triodes of the cascode is tuned for each channel. The coupling between cascode and mixer belongs to the series type, and local oscillator injection is capacitive. Four coils are switched by the turret for each channel.

Figure 7-17 has a special double triode as an r-f amplifier in a direct-coupled cascode circuit. The first triode has a capacitance bridge neutralizing circuit. A double triode is used as oscillator-mixer, with inductive injection of the oscillator signal. Vernier adjustment is provided, as well as small trimmer capacitances to compensate for individual variations from tube to tube in case of replacement, and a test point *P* for oscillator injection control. Also,

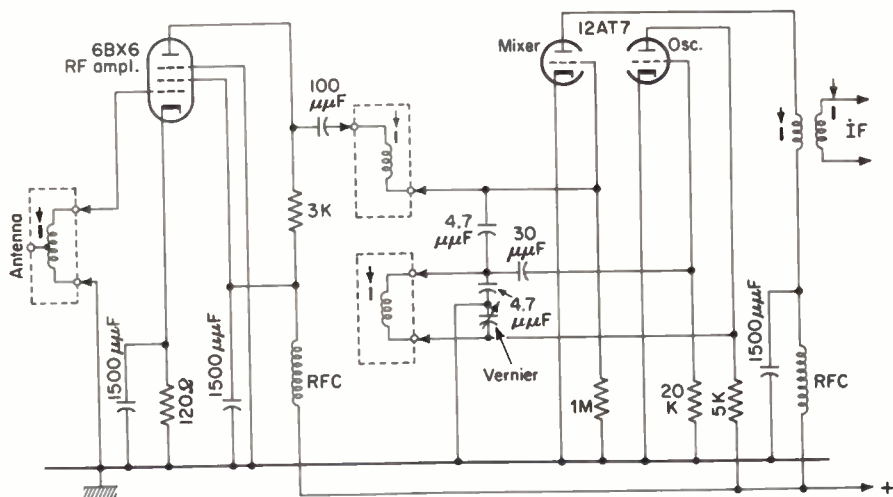


Figure 7-15. Tuner circuit using a pentode RF amplifier and a double triode mixer-oscillator.

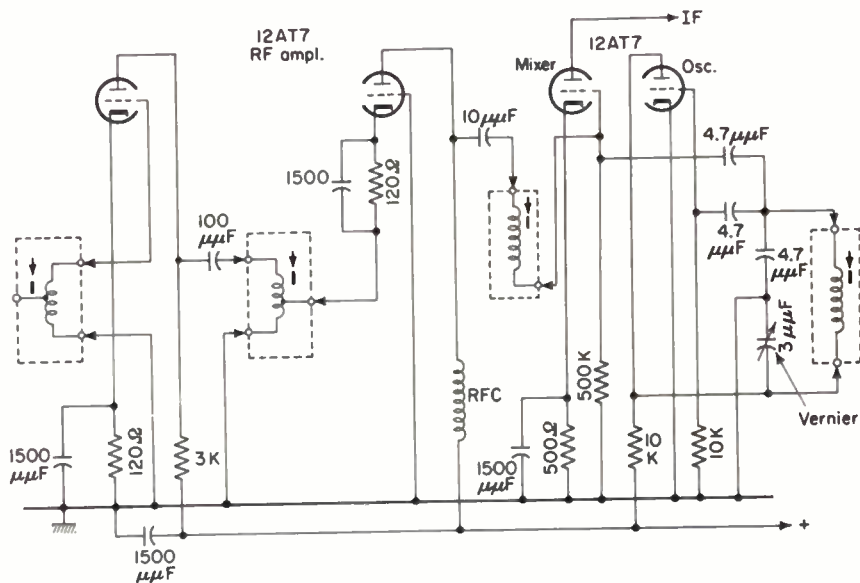


Figure 7-16. Tuner circuit using a double triode as cascode RF amplifier.

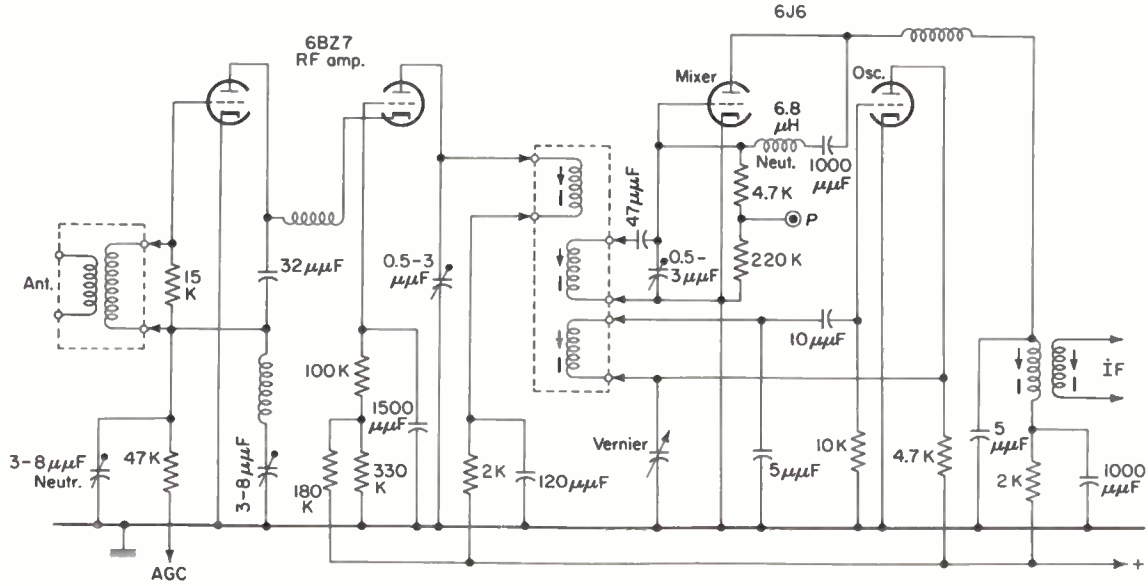


Figure 7-17. Tuner circuit using a dc-coupled cascode RF amplifier.

neutralizing of the mixer triode is obtained with the help of an inductance between plate and grid circuits.

7-7. UHF reception

UHF reception poses some problems of its own. It is generally accomplished with the help of a separate UHF section, existing either in the receiver or as an optional complement to the chassis, or as an external additional unit. UHF reception requires a separate antenna and lead-in line. We have previously stressed the importance of the antenna installation for the VHF bands. It becomes even more important at UHF, where the band to be covered is very wide and where the higher frequency generally means increased losses. Ghosts and echos may also be much more troublesome. Besides requiring a large bandwidth, a UHF antenna should have the highest possible gain. In regard to antenna size and gain, high frequencies (e.g., short wavelengths) have a decided advantage. The antenna may be large in terms of wavelength while being physically small, so that high gain and directivity may be secured without excessive mechanical difficulties.

A number of antennas may be used—multiple Yagi or helix for a narrow band or, for a wider band, conical, fan, corner, or other types. However, care must be exercised to avoid losses in the lead-in line. Sometimes a special low-loss type of cable is used, either tubular line or open wire line.

Because of the peculiarities of UHF propagation, which is very nearly optical in character, UHF reception is generally limited to a somewhat reduced coverage area around the transmitter and is strongly dependent on the local conditions.

Even more important than at VHF, the antenna should be installed as high in the clear as possible; it should have maximum gain and use a well-matched, low-loss, lead-in line for best UHF results.

7-8. Single and double frequency changing

The inclusion of UHF reception in a VHF receiver may be designed along two different principles.

In the first case, the UHF section, consisting of an antenna, r-f amplifier, mixer, and oscillator, feeds directly into the input of the i-f amplifier. This scheme is called *single* frequency changing and is particularly applicable if the i-f frequency is high.

In the second case, the UHF section delivers an i-f frequency which corresponds to some VHF channel unused in the area and feeds into the VHF tuner, which transforms it again into the receiver i-f. This is, of course, *double* frequency changing.

The UHF section is frequently made as a separate subchassis called a *UHF tuner*. In many VHF receivers where the UHF section is not included, a socket is provided to connect an additional UHF tuner and feed it from the existing receiver power supply.

In any case, an entirely separate UHF tuner with its own independent power supply may always be added to an existing VHF receiver.

7-9. UHF tuner

The UHF band extends from 470 to 890 mc, which is almost a 1-to-2 frequency range. Continuous tuning over this band, or sometimes over part of it, is universally used in UHF tuners if only because of the complexity of switching 60 channels if discontinuous tuning were adopted.

A number of tuning methods may be put to use. The popular arrangement is inductance variation of some sort because of the difficulty of using capacitance tuning at such high frequencies. Considering the irreducible parasitic capacitances, this last method leads to very small values of inductance.

Resonant lines are commonly used instead of coils. A shorted quarter-wave line behaves at its open end as an antiresonant circuit and is much more convenient to use at high frequencies. Tuning is accomplished by shorting out a variable length of line. Furthermore, such lines need not be straight; for example, they can be bent into circular shape for ease of tuning and compactness.

Various other tuning methods can be used—for example, some forms of cavity or re-entrant line tuning.

Also, capacitance tuning has been successfully used in association with resonant lines.

Tubes for UHF frequencies are somewhat of a problem. Triodes are used almost exclusively, frequently in some form of grounded-grid circuit. Types 6J6, 12AT7, and other double triodes can be used. Special tubes such as 6J4, 6F4, 6AF4, etc. have been developed.

Frequency changing uses a triode or a crystal diode which needs less oscillator input voltage and accordingly reduces the risk of oscillator frequency radiation through the antenna.

One or two stages of i-f amplification are sometimes included in the UHF tuner.

7-10. UHF tuner diagrams

A theoretical diagram of a simple UHF tuner using a crystal mixer and an oscillator is shown in Fig. 7-18. Quarter-wave lines are used. The shorting contacts slide along the line and are ganged together for single-knob tuning. The antenna is directly coupled to the input-tuned circuit by a hairpin loop.

The oscillator is a variation on the Colpitts design. The crystal mixer receives the input signal and the oscillator signal on both sides for frequency changing. The resulting i-f is applied to an i-f amplifier stage through a filter-tuned circuit. Tracking is ensured by the trimmer capacitances T across each line and by mechanical adjustment of line spacing or line width.

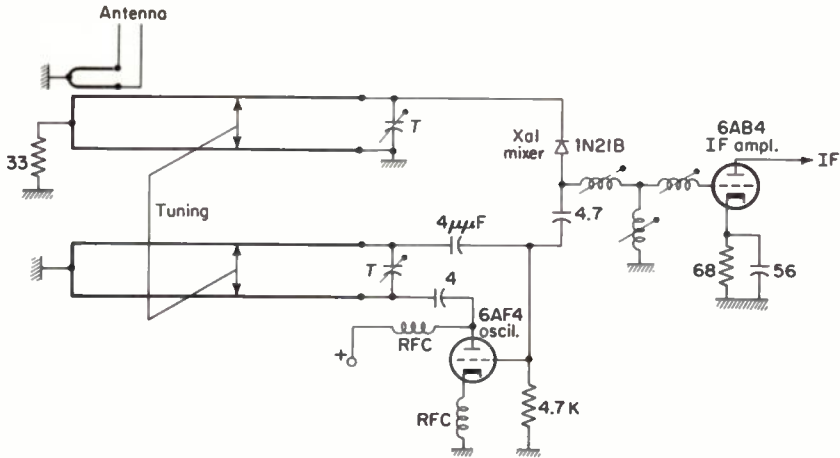


Figure 7-18. UHF tuner using lines as tuning elements.

A variation of this circuit has a third tuned line acting as a preselector between antenna and mixer circuit. Also, a high-pass filter is commonly used at the antenna input to eliminate unwanted image frequencies.

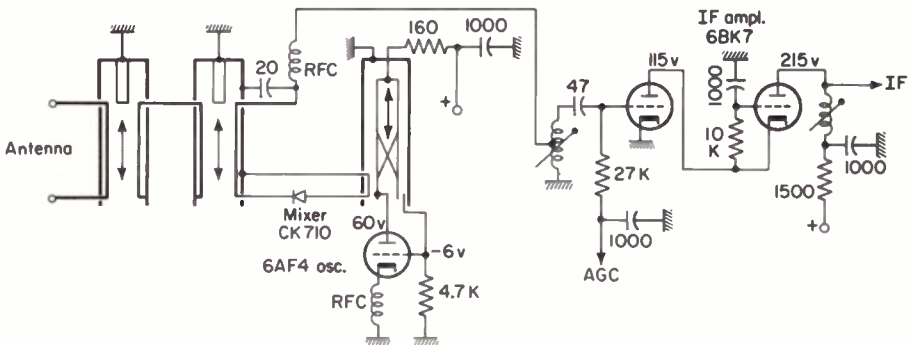


Figure 7-19. UHF tuner using a preselector circuit at the antenna input.

Figure 7-19 illustrates a practical design in which a preselector-tuned circuit is inserted between antenna and mixer input-tuned circuit. The mixer is a crystal diode. The oscillator uses a somewhat similar tuning arrangement.

Plate-to-grid feedback is provided by the grid probe shown, ensuring capacitive coupling of the oscillator circuit to the grid. The three tuned circuits are ganged together and carefully shielded—a general practice in UHF tuners. The resulting i-f is amplified by a low-noise cascode stage and then applied to the input of the i-f amplifier of the receiver.

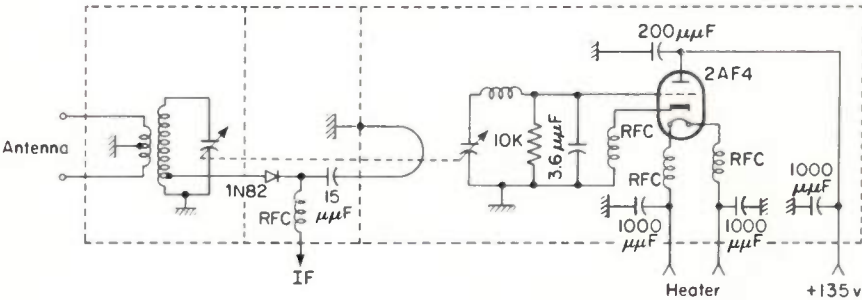


Figure 7-20. Theoretical diagram of an UHF tuner.

Frequently UHF tuner diagrams represent the theoretical circuit and do not try to indicate the physical appearance of the components. Such is the case with Fig. 7-20, which illustrates a tuned antenna input driving a crystal mixer and a triode oscillator. The coupling between oscillator and mixer is inductive through the loop protruding in the oscillator section. Supply lines

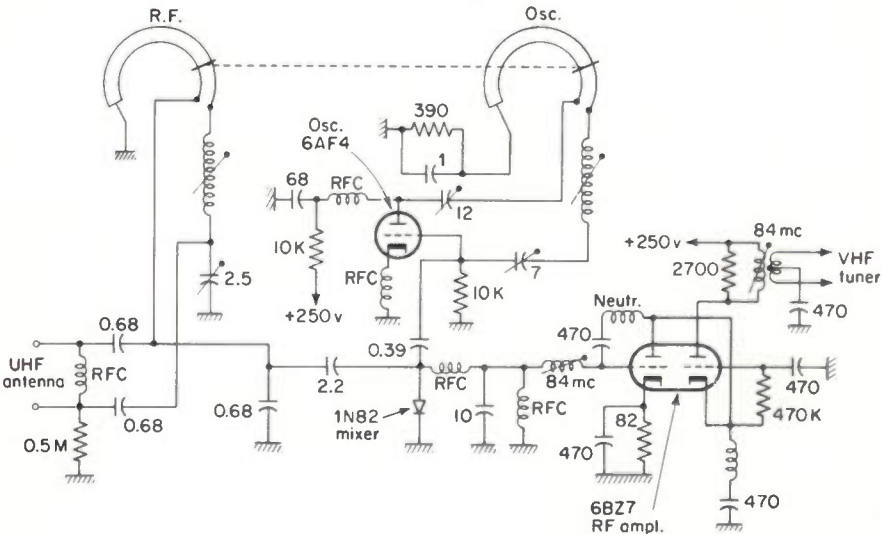


Figure 7-21. UHF tuner using circular lines.

to the oscillator are thoroughly decoupled, and the use of r-f chokes is common in the heater circuits. Shields separate the unit into three sections: r-f input, mixer, oscillator.

Figure 7-21 illustrates a UHF tuner in which the tuning elements are circular lines with movable short circuits. Small adjustable inductances and capacitances are connected in series with the lines for tracking and aligning. The oscillator is a UHF triode and the mixer is a crystal. The i-f provided by the crystal is on 84 mc and is amplified by a neutralized cascode stage before being applied to the VHF tuner set on the suitable channel.

Problems

1. Discuss the advantages and disadvantages of using an oscillator frequency above or below the received frequency.
2. Discuss the advantages and disadvantages of using a low or high intermediate frequency.
3. Because of the limited sound receiver bandwidth, i-f frequency drift has to be limited to ± 25 kc. Calculate the necessary stability percentage of the oscillator for Channel 2 and for Channel 13, assuming a picture carrier i-f of 41.25 mc.
4. Discuss the role of oscillator vernier tuning. What frequency range would you suggest the vernier adjustment should cover? How does this fit in with multi-channel reception?
5. Design an oscillator using a single-tuned circuit and having no tap, even indirectly, on the inductance or capacitance of the tuned circuit, and no additional winding.
6. Compare the series- and shunt-fed Hartley oscillators.
7. How do the values of the parasitic capacitances affect the operation of a Colpitts oscillator?
8. A mixer tube feeds a single-tuned i-f circuit having a center frequency of 44 mc, a 3-db bandwidth of 4 mc, and a shunt capacitance of $18 \mu\mu\text{f}$. The transconductance of the tube when used as an amplifier is 8 ma/v. Calculate the gain of the mixer stage.
9. The mixer stage is sometimes called the *first detector*. Why?
10. Compare the advantages of capacitive and inductive injection of the local oscillator signal into the mixer.
11. Calculate the tuning range ratio for the low VHF band, the high VHF band, and the UHF band.
12. Compare the advantages of capacitance and inductance tuning.

13. Calculate the oscillator frequency range for VHF and UHF bands, assuming a 45.75 mc i-f picture carrier. In each case calculate the tuning range ratio, e.g., the ratio of extreme frequencies in each range.
14. Compare the advantages of continuous and discontinuous tuning.
15. In discontinuous tuning, inductance or capacitance could be switched. Discuss the two possibilities.
16. Why are triode tubes largely used in tuners?
17. Explain the operation of the neutralizing inductance connected between plate and grid of a mixer triode.
18. Give a block diagram of a UHF tuner receiving Channel 74 and feeding the input of a VHF receiver tuned on Channel 13 and having a picture i-f carrier of 45.75 mc. In particular, indicate all relevant tuning frequencies.
19. Why are lines used as tuned circuits at UHF and not at VHF?
20. Assuming a total parasitic capacitance of $8 \mu\text{f}$, calculate the inductance of an antiresonant circuit tuned on Channels 16, 50, and 82.
21. Justify the use of a high-pass filter at the input of a UHF tuner.
22. Why is a preselector-tuned circuit used between the antenna and the mixer input-tuned circuit in some UHF tuners?
23. Suggest and discuss a method for ensuring tracking of the input circuit and oscillator circuit in a UHF tuner. Both circuits are straight lines of two $\frac{1}{8}$ -in. round wires, tuned by a movable short circuit.
24. A quarter-wavelength line is used to tune over the whole UHF band. Evaluate the displacement of the tuning movable short circuit.
25. A UHF tuner feeds into a VHF receiver tuned on Channel 12. Its local oscillator circuit is a quarter-wavelength line, tuned by a movable short circuit. Calculate the minimum and maximum length of the line.
26. Explain why a cascode i-f input stage is used on UHF tuners.
27. In a double frequency changer type of UHF reception, study the eventual effects of the harmonics of the two local oscillators.
28. Compare the merits of single and double frequency changing methods of UHF reception.
29. It has been proposed that all television channels be regrouped together. Discuss the advantages of this proposition.

Chapter 8

I-F AMPLIFICATION

By far the largest part of the total gain of a television receiver is obtained in the i-f amplifier section, which uses several stages in cascade. Furthermore, the shaping of the total response curve is mainly determined by the i-f section, so that tuning of the i-f circuits is critical as far as over-all bandwidth and selectivity are concerned. Interstage coupling circuits are insufficient to provide the desired results. Trap circuits are added, therefore, to improve performance by eliminating unwanted frequencies or bands of frequencies.

8-1. Gain requirement

For a good receiver the total noise referred to the input is on the order of 40 microvolts peak-to-peak. Maximum sensitivity would be required for a signal-to-noise ratio of only 1 and would thus apply to a 40-microvolt peak-to-peak input signal. Assuming sufficient sensitivity to fully modulate the picture tube, that is, roughly a 40-volt peak-to-peak output signal, a high sensitivity receiver would have an over-all voltage gain of

$$\frac{\text{output signal}}{\text{input signal}} = \frac{40}{40 \times 10^{-6}} = 10^6$$

that is, 120 db. For good performance of the detector, its input voltage should be on the order of 4 volts peak-to-peak, so that the gain of the detector-video amplifier section is 20 db. This leaves a 100-db gain to be provided by remaining sections.

Another 20-db gain, roughly, is provided by the r-f amplifier-mixer section, so that the remaining section of the picture receiver, that is, the i-f amplifier, should provide a gain of 80 db, or 10,000 times.

Four stages will easily provide the required amplification, and even three stages might suffice if high-transconductance tubes are used with high-gain coupling circuits, or if some of the bandwidth is sacrificed.

The standard i-f amplifier will thus have three or four tubes and a bandwidth of 3 to 4 mc. Standard i-f picture carrier frequencies are 25.75 and 45.75 mc. The latter value is recommended and actually is better suited to receivers having a UHF band. The higher frequency, therefore, will be adopted in the following treatment. However, it should be understood that what is said applies as well to the lower i-f value with suitable modifications.

8-2. Response curve

The r-f mixer section of the picture receiver provides a more or less constant amplification over the whole channel. One of the important functions of the i-f amplifier, in addition to gain, is to shape the total response curve. Practically, this means providing constant amplification over say 4 mc, ensuring enough gain at sound frequency for the intercarrier system to work and maintaining enough adjacent (i.e., off-channel) selectivity to discriminate against undesired signals.

If the receiver belongs to the split i-f category, in which a separate receiving chain is used for the sound, strong attenuation of the channel sound carrier is needed. However, most present receivers use the intercarrier system and the problem of filtering out the sound frequency is met within the video amplifier.

With the vestigial sideband system of transmission, the response curve of the receiver should have the peculiar shape shown in Fig. 8-1. The response

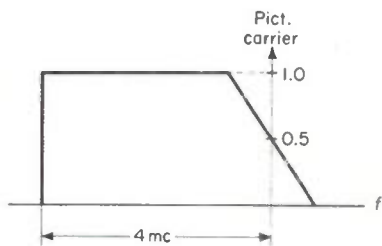


Figure 8-1. Theoretical receiver response curve.

on the picture carrier frequency should be 0.5 the response in the middle of the passband, which corresponds to a 6-db attenuation on the carrier frequency. The reason for this peculiarity is that, as we have seen before, the transmitted signal suppresses a sideband only partially. The result is that frequencies near the carrier, e.g., low frequencies, are really transmitted with both sidebands. On the other hand, frequencies away from the carrier, e.g., high frequencies,

are effectively suppressed in one sideband and are in fact transmitted in the single-sideband system.

At the same time, however, the efficiency of a detector is twice as good when detecting a double-sideband signal than when it detects a single-sideband signal. To compensate for this characteristic of the video detector, an attenuation of 6 db must be introduced somewhere, and this is done in the response

curve. There is a gradual transition from the double-sideband system to the single-sideband system along the side of the curve cutting the carrier frequency at -6 db.

Evidently a response curve with sharp corners is impossible to obtain, and the curve shown in Fig. 8-2 represents a much more reasonable goal without departing too much from the ideal. Nevertheless, this nonsymmetrical curve with rapid attenuation outside the passband is difficult to obtain with the simple wideband amplifier technique previously described. Moreover, adjacent channels may introduce bothersome signals.

8-3. Trap frequencies

With a picture carrier at 45.75 mc, the channel sound carrier will be located at 41.25 mc. The adjacent channels, above and below, will have carriers not too far away, as shown in Fig. 8-3. Below will be the adjacent picture carrier on 39.75 mc, and above will be the adjacent sound channel at 47.25 mc.

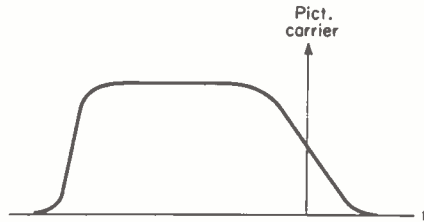


Figure 8-2. Practical receiver response curve.

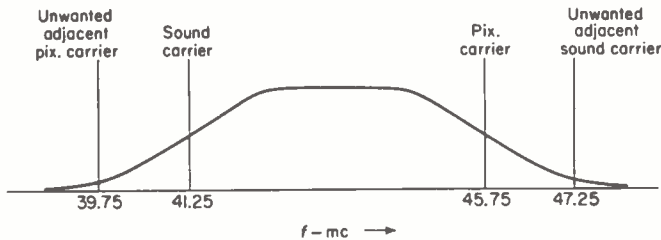


Figure 8-3. Distribution of the various carriers.

The sides of the i - f amplifier response curve fall away slowly as indicated, so that if the *adjacent* channel has a strong signal, it will be sufficiently amplified by the receiver to create undesirable interference. The best way to eliminate this difficulty is to use resonant circuits sharply tuned on the undesired frequencies, that is, 39.75 and 47.25 mc, to introduce additional attenuation. Such circuits are called *trap circuits*. They produce a sharp notch in the response curve as shown in Fig. 8-4.

In practice, an additional attenuation of some 40 db may be needed for the adjacent sound on 47.25 mc. This can be obtained with two trap circuits, each

providing about 20 db. Sometimes one trap circuit is enough, depending on the exact shape of the response curve.

The other frequency is the adjacent picture carrier on 39.75 mc. Here attenuation is somewhat easier because the frequency is quite far from the desired carrier. In fact, the resulting beat frequency is

$$45.75 - 39.75 = 6 \text{ mc}$$

which would be hardly visible on the screen. Furthermore, the adjacent picture carrier is on the channel sound carrier side, where the response curve may be well below the level shown in the above figures.

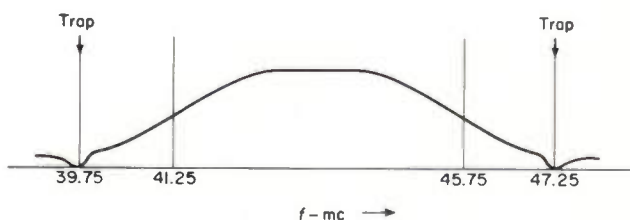


Figure 8-4. Location of the trap frequencies.

Generally, therefore, one trap at 39.75 mc is enough, and none may be provided depending on the shape of the response curve. The same conclusions apply to the other i-f standard, using frequencies of 27.25 mc and 19.75 mc respectively.

8-4. Trap circuits

Various arrangements can be used in the trap circuits and some possibilities that are commonly used will be described in this section. Figure 8-5 shows a coupling circuit LC which might be used between two i-f stages. A trap L_1-C_1 ,

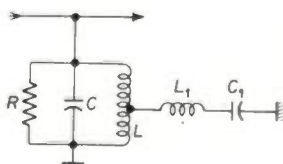


Figure 8-5. Connection of a series resonant trap circuit L_1C_1 .

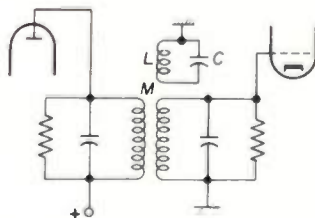


Figure 8-6. Coupled trap circuit LC .

tuned on the unwanted frequency, is connected to a tap on the inductance L and behaves as a short circuit for the unwanted frequency.

When double-tuned transformers are used as coupling components, a third winding LC may be added as shown in Fig. 8-6 and tuned on the unwanted

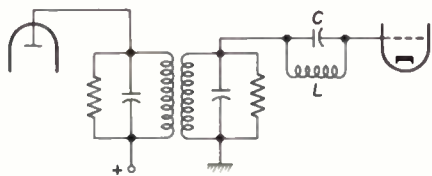


Figure 8-7. A parallel tuned trap circuit LC may be inserted in the coupling.

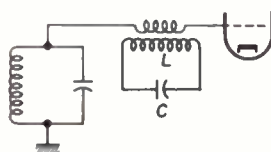


Figure 8-8. A trap circuit LC may be coupled in series into the coupling.

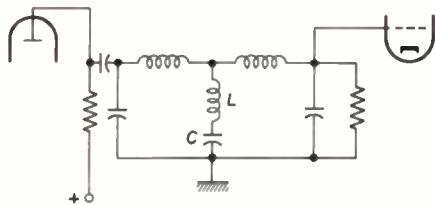


Figure 8-9. The trap circuit LC may be part of a coupling filter.

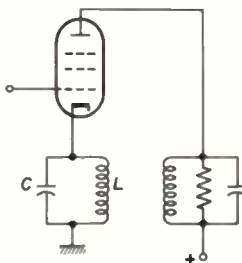


Figure 8-10. Cathode feedback trap circuit LC .

frequency. Also, as shown in Fig. 8-7, an antiresonant LC circuit may be inserted in the grid connection. From the point of view of impedance matching and sharp trapping, the arrangement shown in Fig. 8-8 is superior. In this diagram, the trap LC is coupled into the grid circuit through a small primary.

The trap is sometimes included in a complex coupling circuit, such as shown in Fig. 8-9. Here, trap LC virtually shorts out the coupling at the unwanted frequency.

Another principle is used in the feedback trap as shown Fig. 8-10. Trap LC has a large impedance at its resonant frequency and produces a large amount of cathode feedback, reducing the gain of the stage drastically. As shown in Fig. 8-10, the low cathode impedance would heavily damp the trap

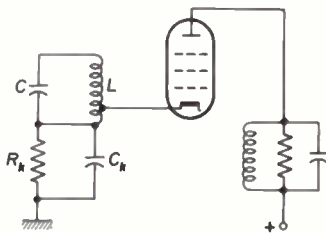


Figure 8-11. Better impedance match is obtained by connecting the cathode to the tap on the trap circuit LC .

circuit. A practical arrangement which matches impedances better and provides for cathode bias is represented in Fig. 8-11.

8-5. *I-f amplifier*

To illustrate the preceding points, Fig. 8-12 shows a slightly simplified diagram of a commercial i-f amplifier. It uses four tubes and has five single-tuned coupling circuits. Note the extensive use of trap circuits, six in all, five of them coupled to the interstage circuits and a sixth in the cathode circuit of the last tube.

The first three tubes receive a variable negative bias used as gain control on their control grid. A 47-ohm undecoupled resistance in each cathode reduces the variation of input impedance. The fourth tube is cathode-biased and works at constant gain. Two each of the traps coupled to the interstage circuits are tuned on the unwanted carrier frequencies of 47.25 and 39.75 mc. The tuning frequencies are indicated near the circuits on the diagram. The two remaining traps, one right at the input and the other coupled to the last cathode, are tuned on 41.6 mc. They create a rapid drop of the response curve at the low end and limit the passband to 4 mc approximately. There is enough response left on the desired sound frequency of 41.25 mc for the sound receiver to function properly.

Notice how carefully the grid and plate circuits have been decoupled. Also, it will be found that tuning frequencies indicated for the interstage circuits do not fit with the standard values calculated for a staggered four-stage amplifier, because the response curve is strongly shaped by the trap circuits. The effect of the traps and the staggering indicated leads to the proper response curve.

8-6. *I-f amplifier tubes*

The i-f section is a high-gain, large-bandwidth amplifier. As such, it must use wideband techniques—in particular, amplifier tubes having high transconductance and low input and output capacitances. This problem has been studied in preceding chapters. However, there is an additional requirement in the case of the i-f section. Its gain must be controlled, either manually or, more generally, through an automatic gain control or AGC circuit. This type of circuit will be analyzed later, but we should point out here that its proper operation requires variable- μ or remote cutoff tubes. The signal takes a larger amplitude as it progresses from the antenna through the stages, and the tubes should be able to handle it. Hence, it may be necessary to adjust AGC action according to the position of the tube in the amplifying chain. This point is especially important in r-f stages where the over-all signal-to-noise ratio is determined to a large extent and where gain should be at maximum for

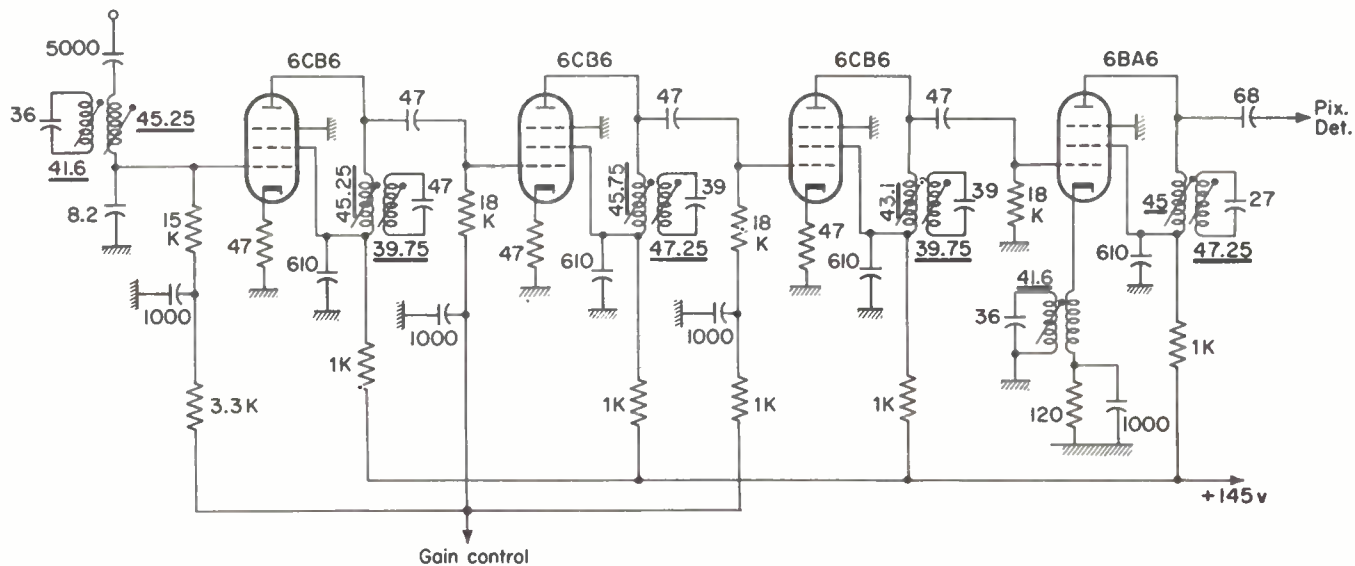


Figure 8-12. Theoretical diagram of a 4-stage i-f amplifier.

weak signals. It is customary to suppress AGC action on the r-f stages for weak signals, generally through a voltage delay circuit. Another related problem occurs in the last i-f stage, which should provide the detector with a signal whose amplitude may well reach several volts peak-to-peak. To handle such an amplitude, the last i-f stage is frequently not submitted to the AGC and works at full gain.

8-7. Response curve shaping

The usefulness of highly efficient trapping may be questioned, since adjacent channels are never used in the same geographical area. For example, Channels 3, 5, 7, 9, 11, and 13 will be used in a given area, and Channels 2, 4, 6, 8, 10, and 12 will not be assigned to the same area. There is, however, the case of fringe reception. Television receivers will be found in localities where adjacent channels may be received from distant transmitters in different geographical areas. Antenna directivity may not be sufficient to avoid interference, particularly since long-distance propagation is bound to be irregular.

There is also the possibility of freak reception over long distances from an adjacent channel transmitter which may be hundred of miles away.

Even if these cases do not apply, two reasons militate strongly in favor of a response curve that is shaped, through the use of trapping, in such a way that it just covers the desired bandwidth and cuts off sharply:

1. This shape will ensure minimum noise for maximum picture definition. Using a bandwidth wider than the standard channel will not add anything to picture definition, which in this case will be limited by the transmitter. It will, however, increase noise, which is proportional to bandwidth, and consequently will degrade the signal-to-noise ratio.

2. In each receiver, the local oscillator used for frequency changing is a small transmitter. Part of the r-f energy it produces finds its way to the receiving antenna through spurious coupling. The receiving antenna will radiate this energy, and it may well reach other receiving antennas. The amount of local oscillation leaking out of a receiver is restricted by manufacturing standards but is still sufficiently large to produce interference in the case of many antennas in close proximity. The risk is increased by oscillator harmonics, always present although generally weak.

Finally, there is in some cases the possibility of interference due to poorly installed equipment such as r-f heating or diathermy units in the neighborhood, to say nothing of legitimate r-f transmitters and the harmonics they may produce.

Evidently, the wider the bandwidth of a receiver, the more chance there is of receiving undesirable signals liable to produce interference. The visual result on the screen is an oblique or wavy spurious pattern superimposed on the

picture. It may be stationary or drifting and is variously described as a *venetian blind* or *herringbone* effect.

Trapping is used, therefore, not only to eliminate unwanted frequencies but also to help in obtaining better adjacent selectivity, that is, selectivity off-channel, by response curve shaping.

Problems

1. An economical receiver has no r-f stage, one frequency changer, two i-f stages, a detector, one v-f stage. Its picture tube is fully modulated by 25 v peak-to-peak. The bandwidth has been reduced to 2.8 mc. Making reasonable assumptions, calculate the sensitivity of the receiver.

2. Show why the efficiency of a detector is twice as good for a double-sideband signal than for a single-sideband signal.

3. What are the interesting trap frequencies for a receiver using a 25.75 mc i-f picture carrier?

4. Compare the merits of 25.75 and 45.75 mc i-f picture carriers.

5. Why should the r-f frequency changer part of a receiver have a flat response over the channel? Why is it not used to help shape the over-all response curve?

6. Draw a typical i-f response curve for a receiver using a 25.75 mc i-f picture carrier.

7. It is proposed to use selective feedback between two successive i-f stages to eliminate an undesired frequency. Investigate the merits and disadvantages of this trap system.

8. Draw a simplified diagram of a three-stage i-f amplifier using 25.75 mc i-f picture carrier. Indicate the position of all tuned circuits, including traps, as well as their tuning frequencies. Justify your arrangement.

9. Some receivers have no traps whatsoever. How is this possible?

10. What should be the quality factor Q of a trap for adjacent channel sound and for adjacent channel picture?

11. Why should frequency-modulated sound be troublesome in a picture receiver?

12. Explain why traps are not used in sound receivers.

13. Show graphically the effect of a cathode trap.

14. Evaluate the amplitude of the signal on the grid of the last i-f tube for a typical receiver.

Chapter 9

DETECTION

A detector stage is used to extract the video modulation, or v-f signal, from the i-f carrier. The familiar diode detector is commonly used for this purpose. Particular attention is paid to detection polarity because of the nonsymmetrical nature of the v-f signal.

The large v-f bandwidth imposes low detection load impedances, and this results in noticeable damping of the driving i-f circuit.

9-1. Detector

By far the most common type of detector used in television receivers is the diode, whether it is a tube or a crystal diode, germanium or silicon. The plate detector and the cathode detector are rarely used.

Although different in appearance and in some characteristics, tube and crystal diodes nevertheless obey the same fundamental laws, and a general theoretical study of the diode can be made without concern for the actual type of diode used. Except when specifically indicated, the standard diode symbol in the following diagrams will represent interchangeably a tube or a crystal diode.

Crystal diodes have a number of advantages and are widely used. They are small, economical, easy to use, need no socket, and do not consume heater power.

The usual detector circuit diagram is the series detector shown in Fig. 9-1. The tuned circuit applies the i-f signal to the diode, and the detected video frequency or v-f signal appears across load resistance R_L . Capacitance C_1 closes the circuit for i-f. The parallel detector of Fig. 9-2 could be used equally well. The components are connected differently but they play the same roles.

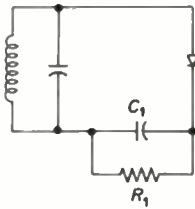


Figure 9-1. Series detection circuit.

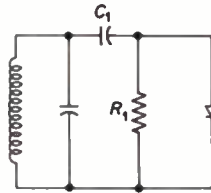


Figure 9-2. Parallel detection circuit.

9-2. Phase

In radio the polarity of detection is of no importance and either polarity is acceptable because the ear is incapable of distinguishing phase reversals. However, this is not so in television, which treats visual images. The modulated i-f carrier has the shape shown in Fig. 9-3. Positive detection using the

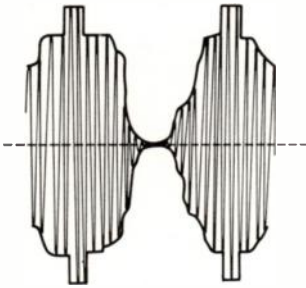


Figure 9-3. Modulated television carrier.

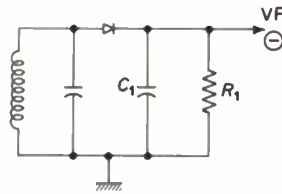


Figure 9-4. Negative v-f phase detection.

circuit given in Fig. 9-4 will allow only the positive part of the i-f signal to pass through the diode. The detected signal will be as shown in Fig. 9-5 and will be recognized as a negative-phase video signal. This is indicated by the minus sign in a small circle shown in both Figs. 9-4 and 9-5. The sign applies to the *phase* of the v-f signal and not to its actual voltage value. As can be seen, the negative-phase v-f signal is in fact entirely above the zero reference level, that is, in the positive voltage region. In a negative-phase v-f signal, the sync tips are up.

If the diode is reversed as shown in Fig. 9-6, providing negative detection, the resulting v-f voltage shown in Fig. 9-7 will have a positive phase and lie

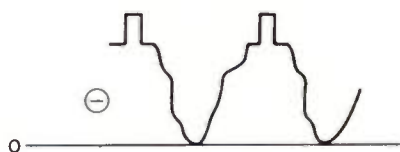


Figure 9-5. Negative phase v-f signal.

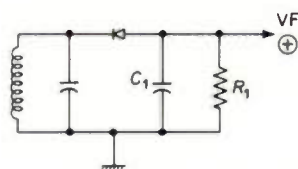


Figure 9-6. Positive v-f phase detection.

entirely below the zero reference line, in the negative voltage region. In a positive-phase v-f signal, the sync tips are down.

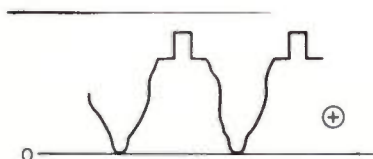


Figure 9-7. Positive phase v-f signal.

The reference line is at the same voltage level as the bottom of the detector load resistance. In the case of the figures, this is ground, or zero, potential.

Another way of obtaining the two v-f phases without reversing the diode connections is to connect the ground to another point of the circuit, as shown in Figs. 9-8 and 9-9. The first two circuits are preferred because the lower end of the input-tuned circuit is connected to ground, which is frequently of practical convenience.

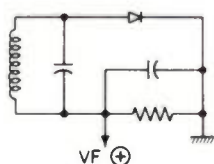


Figure 9-8. The ground can be connected directly to the diode to obtain positive v-f phase detection.

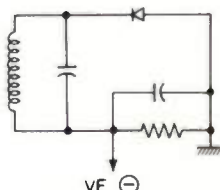


Figure 9-9. Reversing the diode provides negative v-f phase detection.

9-3. Detector polarity

A standard grounded-cathode amplifying stage reverses the phase of the signal between grid and plate. A positive-phase v-f signal applied to an amplifying tube will appear on its plate as negative-phase v-f. Furthermore, the picture tube can be modulated either by the grid with positive v-f or by the cathode with negative v-f.

Therefore, according to the number of amplifying stages between detector and picture tube and to the way that modulation is applied to the tube,

detection should be positive or negative. This is summed up in Table 9-1.

Table 9-1. Detection Polarity

<i>Tube modulation</i>	<i>Number of v-f stages</i>	<i>Detection polarity</i>

cathode	even	positive
cathode	odd	negative
grid	even	negative
grid	odd	positive

It must be remembered that a positive polarity detection provides a negative-phase video signal, and vice versa.

9-4. Resistance value

Besides the input circuit and the diode, the fundamental detection circuit contains only a resistance R_1 and a capacitance C_1 .

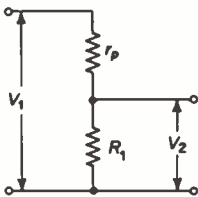


Figure 9-10. Resistance dividing effect of the diode resistance r_p .

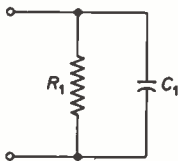


Figure 9-11. Equivalent output circuit.

For the detected or rectified v-f current flowing around the circuit, the load resistance is in series with the diode internal resistance r_p . The tuned circuit impedance can be neglected at v-f. Since V_1 is the total voltage, the output voltage V_2 is only that fraction of V_1 appearing across R_1 so that

$$\frac{V_2}{V_1} = \frac{R_1}{R_1 + r_p} = p$$

where p is the detection efficiency. For high p , the value of R_1 should be high and r_p should be small. The value of R_1 is unfortunately limited by video bandwidth considerations. The load circuit of the detector is as shown in Fig. 9-11. If all frequencies in the video frequency range are to be equally detected, the load impedance Z_1 should be nearly constant in the entire range. Since

$$Z_1 = \frac{R_1 X_1}{R_1 + X_1}$$

where X_1 is the reactance of C_1 , a detected current i will produce an output voltage

$$V = iR_1 \frac{X_1}{R_1 + X_1}$$

The term $X_1/(R_1 + X_1)$ is always smaller than 1 and varies with frequency because of X_1 :

$$X_1 = \frac{1}{j\omega C_1}$$

To minimize this variation, R_1 should be small with respect to X_1 .

The dilemma, then, is that on the one hand R_1 should be much larger than r_p for high detection efficiency, and on the other hand R_1 should be much smaller than X_1 for good frequency response.

Reducing C_1 as much as possible will help. Assuming a value of only 10 μf and a maximum video frequency of 4 mc, the reactance will have a minimum magnitude

$$X_1 = \frac{1}{2\pi \times 10 \times 10^{-12} \times 4 \times 10^6} = 4000 \text{ ohms}$$

approximately.

If a 3-db drop is tolerated at 4 mc, then

$$R_1 = X_1 = 4000 \text{ ohms}$$

In practice, various high-frequency correcting circuits are added to the load to obtain better performance at the high end of the v-f range. They are similar to the correction circuits used in v-f amplifiers and will be analyzed later.

9-5. Capacitance value

Capacitance C_1 also poses a problem of compromise. As we have just seen, it should be kept at minimum value for best v-f response. However, its role is to short circuit R_1 for i-f, so that its reactance X_1 at i-f should be much smaller than R_1 .

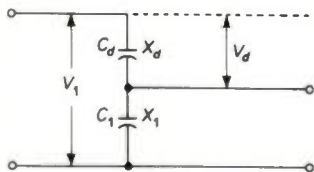


Figure 9-12. Capacitance dividing effect of the diode capacitance C_d .

Moreover, Fig. 9-12 shows that, for the i-f signal, C_1 constitutes a capacitive divider with the diode self-capacitance C_d . The voltage effectively detected is the voltage V_d across the diode, which is smaller than the applied voltage V_1 :

$$V_d = V_1 \frac{X_d}{X_d + X_1}$$

For V_d to be maximum, X_d should be much larger than X_1 . In other words, C_1 should be large with respect to C_d .

The conclusion is that a television detector should have an internal resistance r_p and a parasitic capacitance C_d both as low as possible. In both respects, crystal detectors outclass tube detectors and they are used extensively.

A good part of C_1 may be made up by parasitic capacitances, mainly the input capacitance of the first v-f tube and wiring capacitance. With a crystal diode, the value thus obtained is frequently insufficient and a physical capacitor of say $5 \mu\text{f}$ is connected at the output of the detector.

9-6. Detection damping

At each rectified alternation of the i-f signal, capacitance C_1 receives a small charge, and after a few cycles C_1 is almost charged up to the peak voltage of

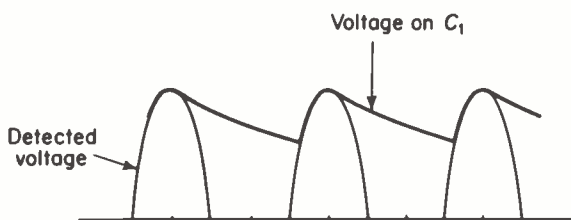


Figure 9-13. Charge of the load capacitance C_1 .

the rectified signal. Between peaks of the detected signal C_1 discharges exponentially through R_1 , as shown in Fig. 9-13, and the following alternation begins to charge it up again, and so on.

The voltage across C_1 must follow the amplitude variations of the i-f signal for faithful detection. To accomplish this, the time constant $R_1 C_1$ of the discharge must be short enough for C_1 to discharge sufficiently in one period to follow the fastest variation of i-f amplitude.

Since C_1 is constantly charged up almost to the peak value of the i-f signal, the diode conducts only for a brief moment at the peak of the i-f signal. If E is the rms value of i-f voltage, the detected voltage is the peak value $E\sqrt{2}$.

Let the current be i . Then

$$i = \frac{1.41E}{R_1 + r_p}$$

or, using the detection efficiency p ,

$$i = \frac{1.41pE}{R_1}$$

The power dissipated is

$$W = 1.41Ei = \frac{E^2}{R_1/2p}$$

Since $W = E^2/R$, the preceding expression shows that the effective damping

resistance value is $R_1/2p$. For example, a detector having an efficiency of 60 per cent and a load resistance of 3600 ohms would damp its driving i-f circuit as would a resistance of

$$\frac{3600}{2 \times 0.6} = 3000 \text{ ohms}$$

The calculation as indicated is grossly simplified and gives at best some idea of the effect of detection damping. The exact calculation is rather involved and does not readily lend itself to quick computation.

9-7. Detection filter

The output of the detector contains not only the desired v-f signal, but also a residual i-f signal, and even harmonics of the i-f frequency. Shunt capacitance C_1 prevents these unwanted components from having a large magnitude, but

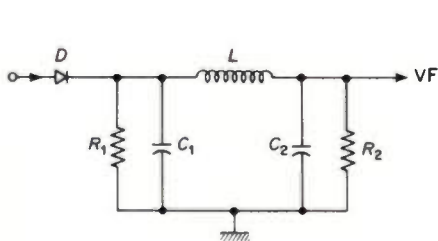


Figure 9-14. Low-pass output filter L — $C_1 - C_2$.

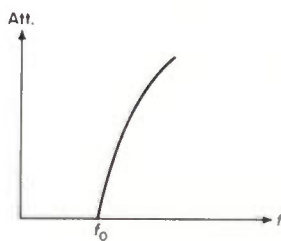


Figure 9-15. Attenuation curve of the low-pass filter.

since they could be amplified by the following v-f amplifier, it might be advisable to prevent them from reaching the grid.

A good solution is to use a low-pass filter. One section is generally sufficient, and Fig. 9-14 shows a pi section at the output of the detector. The theoretical attenuation curve of such a filter is given Fig. 9-15. There is no attenuation up to the cutoff frequency f_0 . Above f_0 the attenuation increases steeply. Cutoff f_0 may be taken at 1.5 or 2 times the highest video frequency, say

$$f_0 = 4 \times 1.5 = 6 \text{ mc}$$

The pi section is terminated by resistances R_1 and R_2 . As far as the diode load is concerned, R_1 and R_2 appear in parallel, and the efficiency drops. To avoid this, R_1 is simply omitted, and the circuit reduces to that shown in Fig. 9-16. The components should be

$$C_1 = C_2 = \frac{1}{2\pi f_0 R_2} \quad \text{and} \quad L = \frac{R_2}{\pi f_0}$$

For example, with $C_1 = C_2 = 10\ \mu\mu\text{f}$ and $f_0 = 6\ \text{mc}$, the values should be

$$R_2 = \frac{1}{2\pi f_0 C_1} = 2650\ \text{ohms}$$

and

$$L = 141\ \mu\text{henrys}$$

An easier way to eliminate or at least attenuate the unwanted components at the output of the detection is to insert an r-f choke in series with the connection to the grid of the v-f tube. The arrangement is shown in Fig. 9-17.

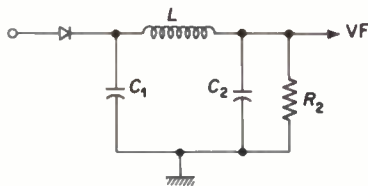


Figure 9-16. A single resistance R_2 may be used.

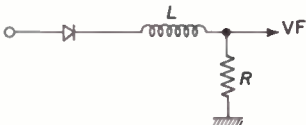


Figure 9-17. An r-f choke L may be used in the output.

When a high-frequency correction system embodying a series inductor is used, this inductor can double in as choke L .

9-8. Other detectors

Symmetrical detection as illustrated in Fig. 9-18 offers several advantages. Theoretically, there is no need for capacitance C_1 across the load resistance R_1 .

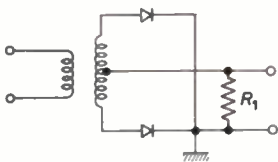


Figure 9-18. Symmetrical detection.

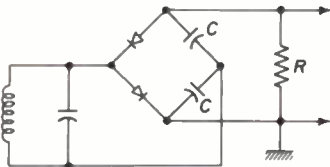


Figure 9-19. Voltage doubling detection.

The efficiency of a symmetrical detector is better than that of the single detector. Even harmonics of the i-f frequency cancel out at the output.

These advantages, however, have little practical value. C_1 cannot be lower than the parasitic capacitances. The efficiency is not greatly improved. The harmonics of i-f frequency are much less troublesome than the fundamental i-f.

On the other hand, the circuit has two very practical drawbacks. It needs two diodes, and the driving circuit should be center-tapped.

In this respect, the voltage doubling detector of Fig. 9-19 is superior. The output voltage is higher and the efficiency better than with a single detector. Capacitances C would be on the order of $10\ \mu\mu\text{f}$, and R is calculated as usual.

Plate detection is rarely used; a diagram is given in Fig. 9-20. Capacitances C are used for decoupling. They have a large value, say 30 to $100\ \mu\text{f}$, and are

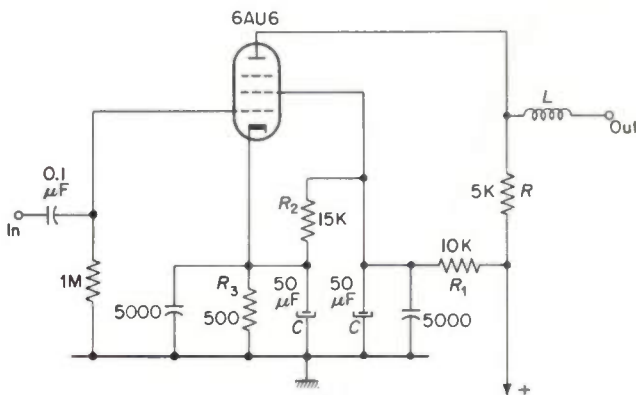


Figure 9-20. Plate detector.

shunted by ceramic or mica capacitors, ranging from 500 to $5000\ \mu\mu\text{f}$ as shown, to improve decoupling at high frequencies. The resistance bridge R_1 - R_2 across the supply voltage determines the screen and cathode voltages. The screen voltage should be the same as when the tube is used as an amplifier. The cathode bias should be such as to fix the working point at or near cutoff. R is the detector load, and a filter inductance L is provided in the output.

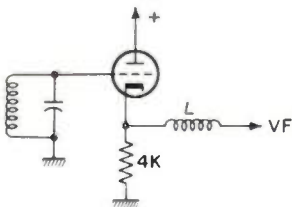


Figure 9-21. Cathode detector.

Cathode detection, shown in Fig. 9-21, is also called *infinite impedance* detection because it hardly loads the preceding circuit. It uses a high-transconductance triode or pentode which has the detection load in the cathode circuit. A filter inductance L is provided at the output.

9-9. Typical designs

A few representative detection circuits will be given to illustrate more recent practices. For the sake of completeness, high-frequency compensation circuits have been included, although they will not be studied until the next chapter. For each diagram they comprise the two inductances whose values are given.

Figure 9-22 shows half a double diode used in a typical circuit. Since parasitic capacitances are insufficient to provide the required value, a $5\text{-}\mu\mu\text{F}$ capacitor has been added in the circuit as previously described. Figure 9-23

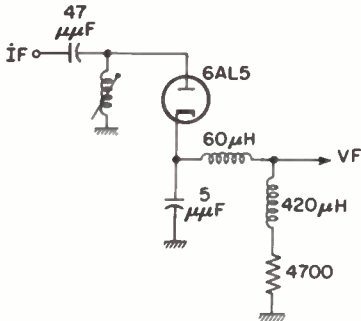


Figure 9-22. Typical diode detection circuit.

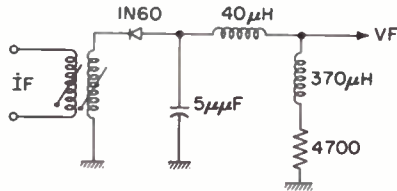


Figure 9-23. Common crystal detection circuit.

uses a crystal detector; except for the compensating inductance values, the circuit is almost identical to the preceding circuit. Figure 9-24 also uses a crystal detector. In its output, besides the compensating inductors and load

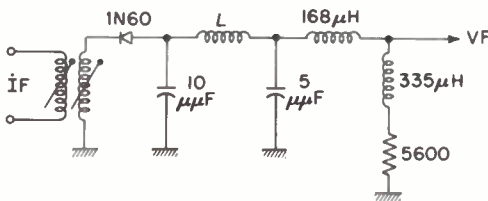


Figure 9-24. Crystal detection circuit embodying a pi filter.

resistor, there is a pi filter circuit comprising inductor L and the two 10- and $5\text{-}\mu\mu\text{F}$ capacitors in parallel with parasitic capacitances.

Problems

1. Compare a tube diode and a crystal diode when used as detectors.
2. During repairs a diode picture detector has inadvertently been connected the wrong way around. What will be the result?
3. Draw a diagram of a detector followed by two stages of v-f amplification driving the picture tube on its cathode. Indicate for each stage the shape of the input and output signal.

4. Assume a parasitic capacitance of $8\ \mu\text{f}$ at the output of the detector. What is the maximum allowable resistance load if the 6-db bandwidth must be 3 mc? No compensating circuit is used.
5. A diode detector has an efficiency of 72 per cent. Its output parasitic capacitance is $12\ \mu\text{f}$. Its load resistance must be such as to obtain a 3-db bandwidth of 3.6 mc without compensating circuit. What is the damping it reflects across its driving-tuned circuit?
6. A diode detector has an output parasitic capacitance of $10\ \mu\text{f}$ and must have a load resistance providing a 3-db bandwidth of 3.2 mc without compensation circuit. The diode internal resistance is 2400 ohms. Evaluate detection efficiency.
7. A diode detector has an internal resistance of 1800 ohms. Its output shunt capacitance is $9\ \mu\text{f}$, and the detector should ensure a 3.4 mc video bandwidth at 3 db without compensation circuit. Calculate the damping across the detection driving circuit.
8. A diode detector has an internal resistance of 2200 ohms and a load resistance of 4700 ohms. An i-f voltage of 1.3 v rms is applied to the diode. Calculate the power delivered by the driving i-f circuit.
9. Design a post-detection pi filter having a cutoff frequency of 5.5 mc. The diode resistance load should be 3800 ohms.
10. A diode detector has an output parasitic capacitance of $10\ \mu\text{f}$ and should provide a video bandwidth of 3 mc at 3 db without compensation circuit. Design a pi filter for the output of the detector.
11. Calculate the theoretical efficiency of symmetrical detection.
12. Calculate the theoretical efficiency of the voltage doubler detection.
13. Explain graphically the operation of a plate detector.
14. Explain graphically the operation of the cathode detector.
15. Using a tube diode and a crystal diode of your choice, design two detector stages having the same 3-db bandwidth. Making reasonable assumptions, compare the two circuits for performance.
16. Show why grid detection is not used in television.
17. Study the effect of a nonlinear characteristic curve for a picture detector. In particular, at low levels, most characteristics are approximated well enough by a parabola. Study the effects of this curvature with respect to signal level.
18. It is proposed to use the saturation knee of a pentode for detection. Discuss the merits of the proposition.
19. A double diode is available for picture detection. Propose three suitable diagrams and compare their performance.
20. A diode detector has developed a leakage resistance between its two electrodes. What will be the result of this defect?

Chapter 10

V-F AMPLIFICATION

With video-frequency signals extending from zero to several megacycles, constant amplification over the entire v-f bandwidth is a problem. Several means are usually combined to secure satisfactory practical performance: a low value of load impedance, high-frequency compensation, and low-frequency compensation. In addition, the d-c component and very low frequencies are taken care of by a separate circuit.

A number of compensation circuits are available for low as well as high frequencies. The final choice is, as always, a matter of judicious engineering compromise.

10-1. Wideband amplification

Theoretically, the range of frequencies that a video-frequency, or v-f amplifier should handle extends from zero to the highest modulation frequency, say 4 mc. In practice, the lowest frequency considered is the picture frequency, or 30 cps, and the d-c component of the signal is taken care of through either d-c coupling or d-c restoration.

It will be seen later that an amplifier which has a sharp cutoff above its high-frequency limit will introduce troublesome transient distortions. To reduce this unwanted effect, the response should fall off progressively above the high-frequency limit. A response curve of this type is shown in Fig. 10-1. The

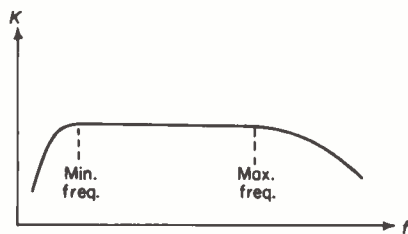


Figure 10-1. Typical v-f response curve.

gain is constant throughout the v-f band, say from 30 cps to 4 mc. It drops sharply at low frequencies and decreases progressively at high frequencies.

Constancy of gain in the v-f band, however, is not the only important consideration. Another desirable goal is a phase variation proportional to frequency, or, equivalently, a time-delay constant in the v-f band. Phase distortion is unimportant in audio work because the ear is insensitive to the relative

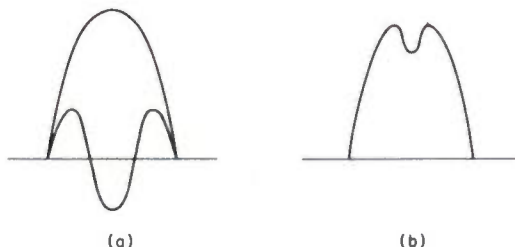


Figure 10-2. Composition of a fundamental and in-phase third harmonic.

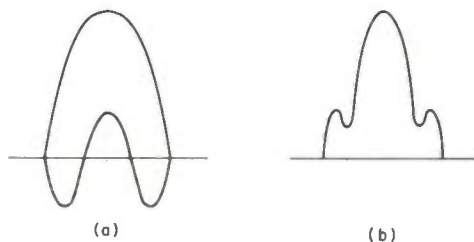


Figure 10-3. Composition of a fundamental and opposite phase third harmonic.

phase of the various frequency components that make up a complex wave. In television, however, phase distortion is very important because the relative phase of the components drastically modifies the shape of the resulting complex wave and the video image. Figure 10-2 illustrates this statement. In (a) are shown two components, the fundamental and the third harmonic, and they are in phase. Their combination gives the complex wave (b). Figure 10-3(a) shows the same fundamental and third harmonic components, but with different relative phases. The resulting complex waveform of Fig. 10-3(b) is very different from that of Fig. 10-2(b) and would result in quite a different pattern on the screen.

In order for an amplifier to have no phase distortion, the various frequency components of a complex wave should have at the output the same relative phase or position that they had at the input. This will, of course, occur if all components are delayed by an equal interval when going through the amplifier. As stated previously, no phase distortion is equivalent to constant time delay for all frequencies.

Now for a time delay equal to one period T , the phase delay will be 2π . Then, to a phase delay ϕ will correspond a time delay t such that

$$t = \frac{\phi T}{2\pi}$$

Since $T = 1/f$, where f is the frequency, then

$$t = \frac{\phi}{2\pi f}$$

For zero phase distortion, t should be constant. This will occur if

$$\phi = kf$$

that is, if the phase variation is proportional to frequency, k being the constant of proportionality.

The visible effect of phase distortion on the screen differs for low and high frequencies. At low frequencies, phase distortion modifies the general background shade, and large uniform shade areas exhibit a characteristic smearing. At high frequencies, phase distortion manifests itself by blurring of the details, the apparent effect being a lack of definition or a bad focusing. The blurring, however, will only affect the small-size details. Moreover, close inspection will show that the raster lines are properly focused.

To sum up the important characteristics of a good v-f amplifier, it may be said that the gain and time delay should be constant and independent of frequency in the entire v-f range.

10-2. Low-frequency distortion

A simple RC-coupled amplifier will satisfy the two fundamental requirements in a limited range of frequencies. It will fall short of ideal at low and high frequencies and must be modified to be an acceptable v-f amplifier.

Figure 10-4 shows a standard amplifying stage with cathode bias R_1C_1 , screen decoupling R_3C_3 , load R_4 , and coupling circuit R_5C_4 . The drop in gain

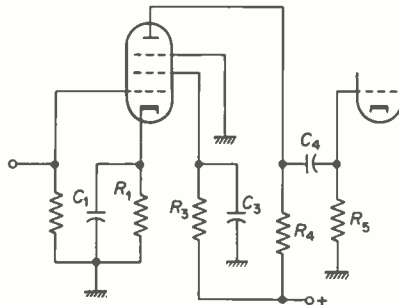


Figure 10-4. Resistance-capacitance coupling.

at low frequencies is essentially due to the effect of capacitances, whose reactance X increases when the frequency decreases:

$$X = \frac{1}{\omega C} = \frac{1}{2\pi f C}$$

Of the three capacitors used in the circuit, C_1 and C_3 are used as decoupling components. They should short circuit the resistances they shunt, R_1 and R_3 , for all frequencies in the v-f range. Generally speaking, the reactance X of the capacitor should be negligible in comparison with the resistance R it decouples.

Since R is a constant independent of frequency and X increases when frequency decreases, there is always a frequency low enough for this condition to be unsatisfied and the decoupling capacitor fails in its role.

For example, assume the typical values $R_1 = 150$ ohms and $C_1 = 50 \mu\text{f}$. At 30 cps

$$X_1 = \frac{1}{2\pi \times 30 \times 50 \times 10^{-6}} = 106 \text{ ohms}$$

approximately, which is by no means negligible across R_1 or 150 ohms. The cathode impedance at 30 cps will not be zero, there will be negative current feedback, and the gain will drop accordingly.

The same reasoning applies to R_3C_3 and screen-grid feedback also reduces the gain at low frequencies.

It might be thought that an easy solution would be to bias the grid negatively, and connect the cathode to ground and the screen directly to some suitable voltage source. This simply shifts the problem to the bias voltage and screen voltage filtering circuits.

The cathode feedback problem can be solved by discarding all capacitance across R_1 . There is then constant feedback all over the v-f range, and hence no variation of gain due to the cathode circuit. The price paid for this solution is the reduction in gain due to feedback.

Screen-grid feedback can be made negligible down to very low frequencies by using a large-value capacitance for C_3 , particularly since R_3 is usually large.

The same method can be used for the cathode circuit. Here the resistance is low, but the voltage is small and capacitances of several hundred microfarads become a practical proposition. All in all, cathode and screen feedback can be eliminated without too much difficulty down to frequencies well below 30 cps.

The real problem lies with the coupling circuit R_5C_4 . The signal V_1 at the plate of the tube is applied to C_4 and R_5 in series, whereas the driving signal V_2 for the following stage is taken across R_5 only. The equivalent circuit is given in Fig. 10-5, where X_4 is the reactance of C_4 . V_2 is given by

$$V_2 = V_1 \frac{R_5}{R_5 + X_4}$$

Using standard values in radio receivers, $C_4 = 0.01 \mu\text{f}$ and $R_5 = 0.5 \text{ megohm}$, reactance X_4 will be equal to R_5 when

$$\frac{1}{2\pi f C_4} = 500,000$$

or

$$f = 32 \text{ cps}$$

approximately. At this frequency, V_2 is only $0.71 V_1$ and, even worse, the phase shift reaches 45 degrees. The easy solution is to increase R_5 and C_4 .

A maximum limit for R_5 is set by the fact that it is connected in a grid circuit and that tube manufacturers set a limit to the permissible resistance, usually 1 megohm or less because of grid current.

As far as C_4 is concerned, its value may be increased many times, keeping in mind that the larger the capacitance, the bigger the capacitor—with consequent increase in parasitic capacitance to ground and the risk of higher leakage current. A large time constant $C_4 R_5$ may result in motorboating due to relaxation by grid blocking.

Common values used for $C_4 R_5$ are $0.1 \mu\text{f}$ and 0.5 megohm , which provide a 3-db loss and a 45-degree phase shift at 3.2 cps that is generally satisfactory.

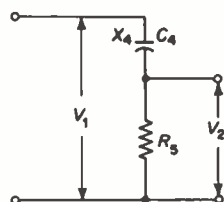


Figure 10-5. Equivalent coupling circuit.

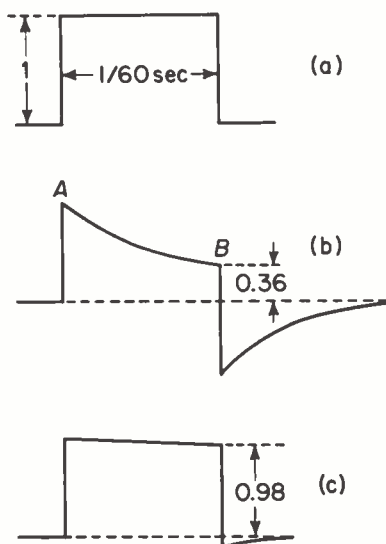


Figure 10-6. Distortion of rectangular signals.

10-3. Square-wave distortion

Low-frequency distortion takes on more immediate physical significance if square-wave signals having a long period are considered. For example, consider a long rectangular pulse of duration $t = \frac{1}{60}$ second or, approximately, $t = 0.017$ second, of unit amplitude. Such a waveform is shown in Fig. 10-6(a).

When this pulse is applied to an RC coupling circuit, the actual waveform at the output is the result of the capacitance charge and has an exponential shape that follows the expression

$$\epsilon^{-t/RC}$$

where $\epsilon = 2.718 \dots$. This expression shows that the distortion of the pulse depends on the ratio of its duration t

to the time constant RC of the circuit. For example, let $RC = t$. The output pulse has the shape illustrated Fig. 10-6(b). Its amplitude has fallen to

$$\epsilon^{-1} = \frac{1}{\epsilon} = \frac{1}{2.718} = 0.36$$

at the end of the pulse. As shown in Fig. 10-6(b), if point A at the beginning of the pulse corresponds to white, point B at the end of the pulse corresponds to a dark grey, whereas the original rectangular pulse of Fig. 10-6(a) was supposed to represent a uniformly white surface.

The drop at the end of the pulse, furthermore, will be transmitted as such by the capacitances with the result shown at point B of Fig. 10-6(b). The output voltage undershoots zero and the signal tends asymptotically to zero with a time constant RC after the pulse has ended.

Increasing the coupling circuit time constant to $RC = 5t$ results in an amplitude, at the end of the pulse, of

$$\epsilon^{-1/5} = 0.98$$

The 2-per cent droop during the duration of the pulse and the following 2-per cent undershoot are tolerable, and it may be said that a pulse of duration t is transmitted with tolerable distortion by a circuit having a time constant of $RC = 5t$ or longer. An all white picture, for example, is a rectangular pulse of duration $\frac{1}{60}$ second and is the longest such pulse in v-f waveforms. This, then, gives the desired time constant for almost zero distortion:

$$RC = \frac{1}{12} \text{ second}$$

With $R = 1$ megohm and $C = 0.1 \mu f$, the time constant is

$$RC = \frac{1}{60} \text{ second}$$

which is just suitable.

Frequency distortion, phase distortion, and rectangular signal distortion are three different aspects of the same basic phenomenon. Because television deals with visual displays, the tendency is to focus attention on distortion of waveforms, that is, rectangular signal or square-wave distortion. Let us consider some other aspects of distortion.

10-4. Low-frequency compensation

Low-frequency distortion can be reduced to a tolerable amount with judicious selection of the values of the components as previously described in Sec. 10-2. If for some reason this cannot be done, there is always the possibility of compensating for the distortion by deliberately introducing an opposite distortion by means of a low-frequency compensation circuit.

A common low-frequency compensation circuit is shown in Fig. 10-7. The added elements are R and C . At high and medium frequencies, C effectively

decouples R . At low frequencies, however, the impedance of C and R in parallel is no longer zero and adds to the actual load resistance R_4 . The gain of the stage increases, and this compensates for the decrease in output voltage at low frequencies due to $C_4 R_5$.

If R is several times larger than R_4 , good practical compensation can be obtained. The equivalent circuit shown in Fig. 10-8 may be represented as a bridge network, which will be balanced and independent of frequency for

$$CR_4 = C_4 R_5$$

In practice, the value of R_4 is determined by *high-frequency* considerations, to be taken up in Sec. 10-6. If C_4 and R_5 are known values, C is also known.

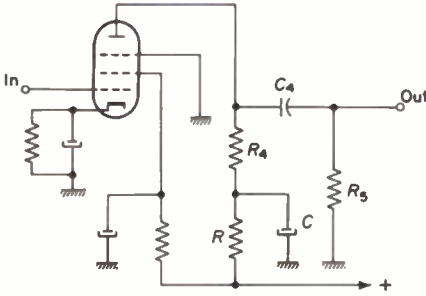


Figure 10-7. Low frequency compensation network RC .

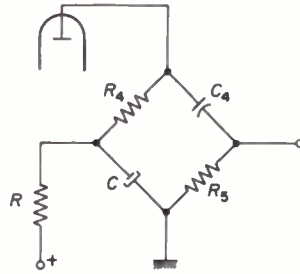


Figure 10-8. Equivalent bridge circuit.

The resistor R is then chosen as several times larger than load resistor R_4 , compatible with the voltage drop.

With this method of computing and providing compensation, the gain and time delay may be maintained at a constant value down to very low frequencies. Furthermore, RC acts as an extra filtering and decoupling circuit, reducing residual hum and increasing stability, and making low-frequency compensation a desirable technique.

10-5. Graphical solution

The graphical method, Fig. 10-9, permits a quick and sufficiently accurate solution of low-frequency compensation problems. C_1 and R_1 are the coupling capacitor and resistor respectively, R is the load resistor, and C the compensating capacitor.

The chart is used twice in the following manner:

1. Read R_1 on the ordinate R -scale and C_1 on the oblique lines. The intersection gives the time constant $\theta = R_1 C_1$ on the abscissa scale.

2. Read R on the ordinate scale, and θ on the abscissa scale. The intersection gives C , read on the obliques. C will be rounded off to the nearest lower standard commercial capacitance value. This lower value provides a small amount of overcompensation which will eventually compensate for a small amount of cathode or screen distortion.

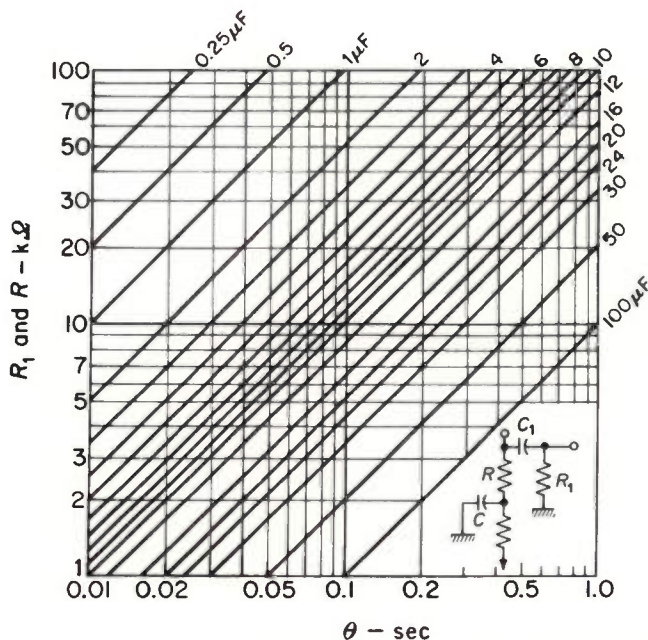


Figure 10-9. Graph to calculate low-frequency compensation network component values.

10-6. High-frequency compensation

High-frequency distortion is totally independent of low-frequency distortion and is due to parasitic shunt capacitance.

The total parasitic shunt capacitance C of a v-f amplifying stage is shown in Fig. 10-10. When the frequency increases, the reactance of C decreases and reduces the effective value of load, that is, the gain. Also, since the effective load consists of C and R in parallel, the phase angle varies with frequency.

To reduce the shunting effect of C across R , the first method is to reduce C to the barest possible minimum. When this has been done, the other recourse is to reduce R . Unfortunately, this also reduces the gain.

As in the case of low-frequency distortion, the only solution, then, is to use the compensation principle by introducing a network which will provide a

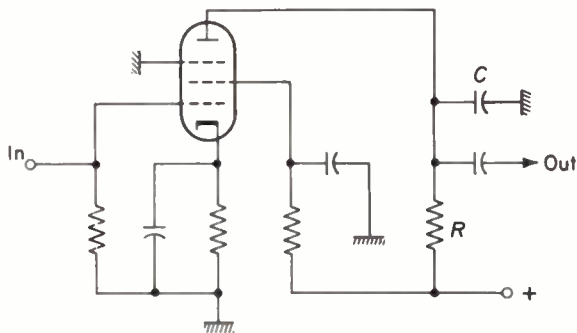


Figure 10-10. Parasitic capacitance C between anode and ground.

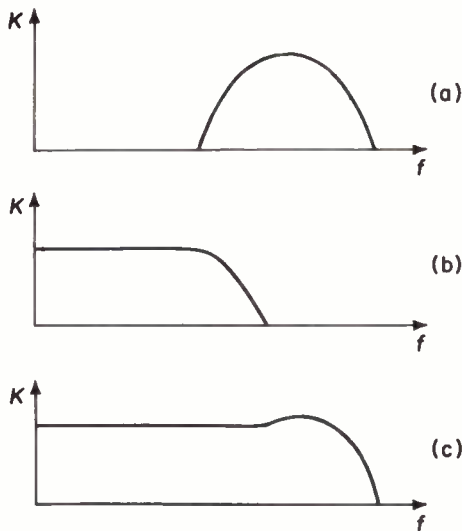


Figure 10-11. The addition of a damped tuned circuit (a) to an RC response curve (b) provides the extended response (c).

load which increases with frequency. Such a circuit could be a damped tuned circuit, resonating near the top of the v-f band and having the response curve as shown in Fig. 10-11(a). The v-f stage response on its load R is given by Fig. 10-11(b). Adding the tuned circuit in series with R as shown in Fig. 10-12 combines the two curves of Fig. 10-11(a) and (b) to provide the total resulting curve (c), which exhibits an increase in high-frequency response.

Numerous high-frequency correction circuits have been devised. They can be broadly classified in two categories—two-terminal and four-terminal circuits.

In two-terminal circuits input and output are common points as shown in Fig. 10-13, the circuit appearing in shunt across the terminals.

In four-terminal circuits as shown in Fig. 10-14, the input and the output are separated by the circuit, which then has at least one component in series between input and output and at least one shunt component. Four-terminal circuits separate the parasitic capacitance into two parts: output capacitance and input capacitance. Thus it is to be expected that they provide greater over-all gain than two-terminal circuits. These circuits will be considered in turn.

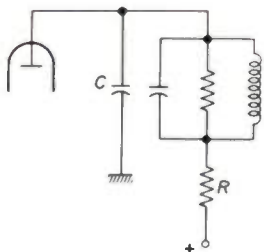


Figure 10-12. High frequency compensation circuit.

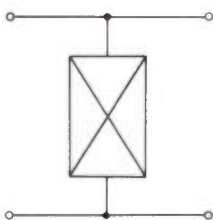


Figure 10-13. Generalized two-terminal circuit.

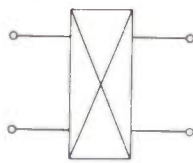


Figure 10-14. Generalized four-terminal circuit.

10-7. Two-terminal networks

All load circuits contain at least the load resistance R and the parasitic shunt capacitance C . For compensation purposes, various reactive elements are added to counterbalance the effect of C . The goal is to build, across the capacitance C terminals, the highest possible impedance having a constant value up to some maximum frequency f_0 , say 4 mc.

This problem can be solved very generally. It provides the useful following result: the highest constant impedance Z that can be built across C is given by

$$Z = 2X$$

where X is the reactance of C at frequency f_0 :

$$X = \frac{1}{2\pi f_0 C}$$

The tube transconductance being g_m , the maximum gain will be

$$K = g_m Z = 2g_m X$$

and is once again limited by the parasitic shunt capacitance.

The conclusions reached for high-frequency tubes then apply to v-f tubes. They should have maximum transconductance and minimum parasitic capacitances.

10-8. Shunt compensation

The simplest two-terminal compensating circuit is given in Fig. 10-15. The only additional component is an inductance L in the branch shunting the capacitance. A variation, shown in Fig. 10-16, is obtained by interposing L

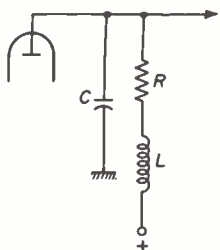


Figure 10-15. Shunt type compensation circuit.

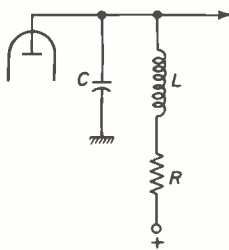


Figure 10-16. R and L can be permuted.

and R , the two circuits being equivalent. Let $X = 1/\omega_0 C$ be the reactance of C at the maximum angular velocity, $\omega_0 = 2\pi f_0$. Then the values of R and L are given by

$$R = X$$

and

$$L = \frac{0.5X}{\omega_0} \text{ or } L = 0.5CR^2$$

For $f_0 = 4$ mc, the following practical expressions can be used:

$$R = \frac{40}{C}, \quad L = \frac{800}{C}$$

with R in kilohms, C in $\mu\mu\text{f}$, L in μhenrys .

Shunt compensation is the classical circuit used in high-frequency compensation. Its gain will be taken as unity, therefore, to allow comparison with other compensating networks to be discussed in the following sections.

Different values can be adopted. For $R = 0.85X$ and $L = 0.3X/\omega_0$, the gain will be reduced to 0.85 times the gain of the standard shunt compensation, but the time delay is much more nearly constant. Corresponding practical values are

$$R = \frac{34}{C}, \quad L = \frac{480}{C}$$

in kilohms, $\mu\mu\text{f}$, and μhenrys respectively.

10-9. Other two-terminal networks

Figure 10-17 shows a two-terminal network that differs from the standard shunt compensation by the addition of a capacitor C_1 across L . Recommended practical values are

$$R = 1.2X, \quad L = \frac{0.54X}{\omega_0}, \quad C_1 = \frac{2C}{3}$$

These values provide a relative gain of 1.2 with respect to standard shunt compensation, and the time delay characteristic is intermediary between the two previous cases. Practical values are

$$R = \frac{48}{C}, \quad L = \frac{864}{C}, \quad C_1 = 0.67C$$

in kilohms, μf , and μhenrys respectively.

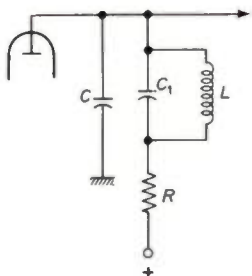


Figure 10-17. Variant of the shunt compensation circuit.

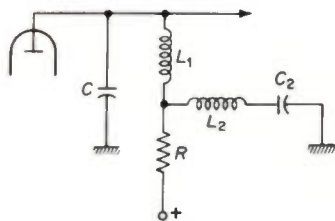


Figure 10-18. Complex two-terminal compensation circuit.

Figure 10-18 is a more complex two-terminal network for which

$$L_1 = \frac{1.8X}{\omega_0}, \quad L_2 = \frac{1.2X}{\omega_0}$$

$$R = 1.5X, \quad C_2 = 0.3C$$

The relative gain is 1.5, and the time delay characteristic is about as good as that of the preceding network. Again, for a maximum frequency of 4 mc, practical values are

$$L_1 = \frac{2880}{C}, \quad L_2 = \frac{1920}{C}$$

$$R = \frac{60}{C}, \quad C_2 = 0.3C$$

in kilohms, μf , and μhenrys .

10-10. Four-terminal networks

In a four-terminal network, as shown in Fig. 10-19, the total parasitic shunt capacitance C is split into two parts C' and C'' .

The general theory of four-terminal networks indicates that if X is the reactance at maximum frequency f_0 of the highest capacitance of C' or C'' (that is, the lowest reactance), the maximum possible gain is

$$K = 2g_m X$$

where g_m is the tube transconductance.

The expression is similar to that found for the two-terminal networks, but since $C = C' + C''$, in the most favorable case, where $C' = C''$, the gain of a four-terminal network may theoretically reach twice the gain of a two-terminal network.

Frequently, however, parasitic capacitances are such that $C'' = 2C'$, so that

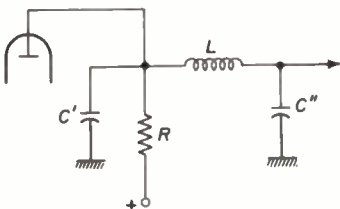
$$C = 1.5C'$$

and the relative gain of a four-terminal network is only 1.5 times the gain of the standard two-terminal network.

10-11. Series compensation

The simplest four-terminal compensating network is shown in Fig. 10-20. It uses only one extra component, the inductor L placed in series with the coupling circuit.

Let X' and X'' be the reactances of C' and C'' at the maximum frequency and let $C' = C''/a$. Then



$$R = \left(\frac{a}{2}\right)^{1/2} \times X'' = 0.71a^{1/2}X''$$

$$L = \frac{aX''}{2\omega_0}$$

In the usual case (Sec. 10-10) where $a = 2$, the expressions simplify to

$$R = X'', \quad L = \frac{X''}{\omega_0}$$

Figure 10-20. Series type compensation circuit.

and the relative gain is 1.5 times the gain of the standard two-terminal network.

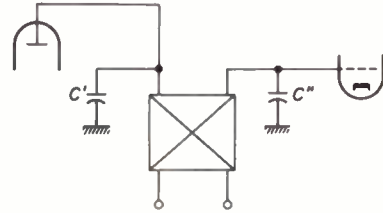


Figure 10-19. A four-terminal circuit separates the parasitic capacitance in two components C' and C'' .

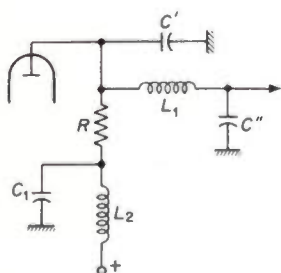
For the maximum frequency of 4 mc, practical expressions are

$$R = \frac{40}{C''}, \quad L = \frac{1600}{C''}$$

in kilohms, $\mu\mu\text{f}$, and μh .

10-12. Compound compensation

Combining series and shunt compensation networks gives the compound compensation shown in Fig. 10-21, with series inductance L_1 and shunt inductance L_2 . Values for this case are



$$C' = 0.8C'', \quad C_1 = 0.35C'', \quad R = 1.5X''$$

$$L_1 = \frac{2.25X''}{\omega_0}, \quad L_2 = \frac{0.675X''}{\omega_0}$$

and the relative gain reaches 2.25 times the gain of the standard two-terminal network.

Practical expressions for a maximum frequency of 4 mc are

$$R = \frac{60}{C''}, \quad L_1 = \frac{3600}{C''}, \quad L_2 = \frac{1080}{C''}$$

$$C' = 0.8C'', \quad C_1 = 0.35C''$$

in kilohms, $\mu\mu\text{f}$, and μh .

It should be noted, however, that four-terminal networks require a given ratio a between C' and C'' . The exact ratio could be obtained by adding extra capacitance in shunt with C' or C'' as required, but this is generally not recommended. The performances are not too critically dependent on the value of a .

10-13. Comparison

It is obvious from the above discussion that four-terminal networks provide a higher gain than two-terminal networks for comparable constancy of gain and time delay.

It may be generally stated that gain increases with the number of reactive components, that is, with the complexity of the network. Four-terminal networks, however, are more complex than two-terminal networks in that they are more difficult to adjust and generally more difficult to design for flat response over the v-f bandwidth. Two-terminal networks, correspondingly, are much more tolerant of variations in component value.

Four-terminal networks provide a response curve which falls off rather abruptly at high frequencies. The resulting transient response at high frequencies is not too good. This serious drawback is frequently corrected by establishing the four-terminal network for a higher maximum frequency than

necessitated by the v-f bandwidth. There is a loss of gain when this is done and it somewhat offsets the advantage of the network in this respect. Nevertheless, the commonest high-frequency compensation network is the compound type, followed by the shunt and series types.

10-14. Other types

Other compensation methods such as the voltage-divider or selective feedback can be devised. Both flatten the response curve by reducing its maximum, whereas inductance networks increase its minimum and are preferred for this reason.

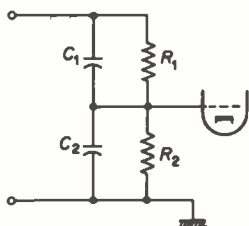


Figure 10-22. Capacitive voltage divider.

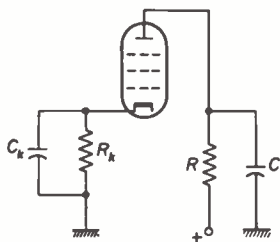


Figure 10-23. Cathode feedback.

Voltage-divider or time-constant compensation is shown in Fig. 10-22 in a grid circuit. C_2 is the tube input capacitance and C_1 is the compensating component. At low and medium frequencies, the dividing ratio is determined by the values of R_1 and R_2 .

At high frequencies, however, the dividing ratio is determined by C_1 and C_2 and this difference in the factor determining dividing ratio can be put to good advantage. The dividing ratio will be independent of frequency when

$$R_1 C_1 = R_2 C_2$$

Figure 10-23 shows the cathode feedback compensation. The cathode capacitance C_k is such that at low and medium frequencies its reactance is large. Current feedback is applied to the tube and the gain reduced.

When the frequency is high, C_k has a low reactance and effectively shunts R_k , ensuring the full gain of the tube.

When R is the anode load resistance and C the associated parasitic shunt capacitance, exact compensation is obtained for

$$R_k C_k = RC$$

10-15. Transient signals

An abrupt voltage change or transient such as shown in Fig. 10-24 is typical of v-f signals. For example, it corresponds to a sudden transition from black to

white in the picture. Such a signal is inevitably distorted and will appear at the output of the amplifier having the shape shown in Fig. 10-25. The amount of distortion depends on the amplifier characteristics.

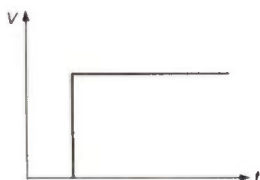


Figure 10-24. Ideal step signal.

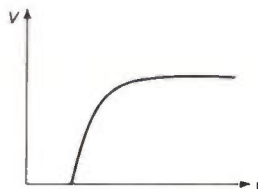


Figure 10-25. Step signal at the output of an amplifier.

For a given bandwidth, the more constant the gain and the time delay, the lower the distortion. For identical gain and time delay, distortion is lower for an amplifier having useful gain at the highest frequency.

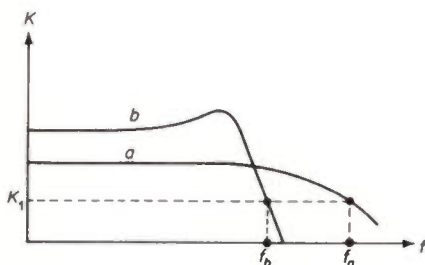


Figure 10-26. Of these two response curves, *a* provides the best transient response.

It follows that a response such as curve *a* in Fig. 10-26 will ensure better transient response than curve *b* because, for the same low gain K_1 , curve *a* has some response up to frequency f_a , whereas curve *b* provides response only up to lower frequency f_b .

Curve *a* is characteristic of two-terminal compensation, and curve *b* is characteristic of four-terminal compensation. On this basis, transient response is better for two-terminal compensation networks.

Actually, the shape of a rectangular or step signal after going through the amplifier depends on the component values, and this may be the basis of a method of designing the amplifier. A tolerable distortion is decided upon, and the component values are evaluated accordingly.

10-16. Transient response

The amplitude and phase characteristics as functions of frequency determine the performance of any compensation network. The final result (that is, the effect on the reproduced picture) is the ultimate criterion, however, and since

it is dependent directly on transient response, it seems that transient performance would be a more realistic way of appraising an amplifier in terms of visual results. This is customarily done by using, as a test signal, a rectangular or step signal, whose amplitude is normalized to unity for ease of comparison. The input voltage is then a unit step signal.

When such a signal is applied to the grid of an amplifier stage having a pure resistance load R , as shown in Fig. 10-27, the parasitic shunt capacitance C between plate and ground cannot charge up instantaneously but only with time constant $T = RC$ according to the exponential charge law. The anode signal, consequently, has the shape shown in Fig. 10-28.

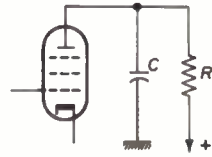


Figure 10-27. Anode voltage can not change more rapidly than allowed by time constant RC .

The abrupt input variation is replaced by a gradual transition, which can be considered as completed in a time equal to $5T$ or $5RC$. If the original signal corresponds to a jump from black to white, Fig. 10-28 shows that the jump will be replaced by a short but nevertheless progressive transition from black to white in time $5RC$.

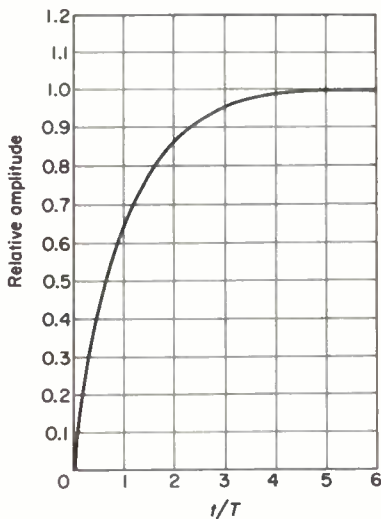


Figure 10-28. Transient response for an RC load.

For the typical values $R = 4000$ ohms and $C = 20\mu\text{f}$, the time constant is

$$T = RC = 0.08 \mu\text{sec}$$

and the transition has a duration

$$5T = 0.4 \mu\text{sec}$$

For a scanning velocity of $1 \text{ cm}/\mu\text{sec}$, typical of large-size screens, the transition occupies a length of 0.4 cm on the screen and can hardly pass unnoticed.

10-17. Compensation

High-frequency compensation is used, therefore, to improve transient response and eliminate undesirable "smearing" in the case of rapid transitions from black to white and vice versa. This point of view is complementary to the one previously adopted when dealing merely with compensation from a frequency response point of view.

Compensating inductances noticeably modify the shape of the transient response, which may appear as shown in Fig. 10-29. The initial rise is faster

than it is without compensation, but the response overshoots 1 and exhibits a damped oscillation asymptotic to 1.

The overshoot may be important and has to be considered, especially if two v-f amplifiers are cascaded, in which case the distortions are cumulative and add. Its amount depends on the R , L , and C values in the high-frequency compensation network. In particular, values can be found for which there is no overshoot and no oscillation. Unfortunately, these values do not coincide with values ensuring maximum gain, which in turn are different from values providing the flattest frequency response.

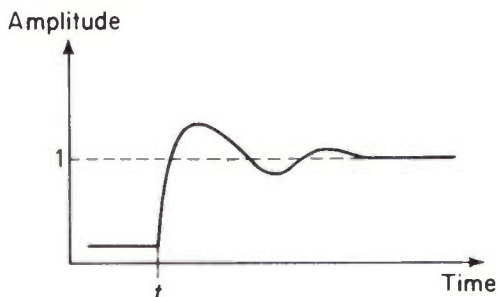


Figure 10-29. Overshoot in the transient response.

It is necessary, therefore, to reach a compromise, which is somewhat dependent on the specific performance expected from the amplifier.

10-18. Shunt compensation

Since the simplest compensation networks are of the shunt type, it is instructive and easy enough to work out the effect of component values on the transient response. At the same time, frequency and phase characteristics can be obtained along with a deeper insight into the relationship between the three response characteristics (frequency, phase, and transient).

To evaluate the components, let C be the total shunt capacitance and f_0 the frequency for which the gain has decreased by 3 db. Then

$$R = \frac{1}{2\pi f_0 C}, \quad L = k^2 C R^2$$

where k is the parameter whose value determines the shape of the gain and phase characteristics, as shown in Fig. 10-30. The diagram is made universal by using f/f_0 as the frequency variable.

For $k = 0$, there is no compensating inductance, and the standard RC load response curve is obtained, giving 3-db attenuation and a 45-degree phase shift at f_0 .

For $k = 0.71$, the inductance value is

$$L = 0.5CR^2$$

and there is a slight hump in the gain characteristic. Also, the transient response exhibits a marked overshoot. The bandwidth gain with respect to the RC load is important.

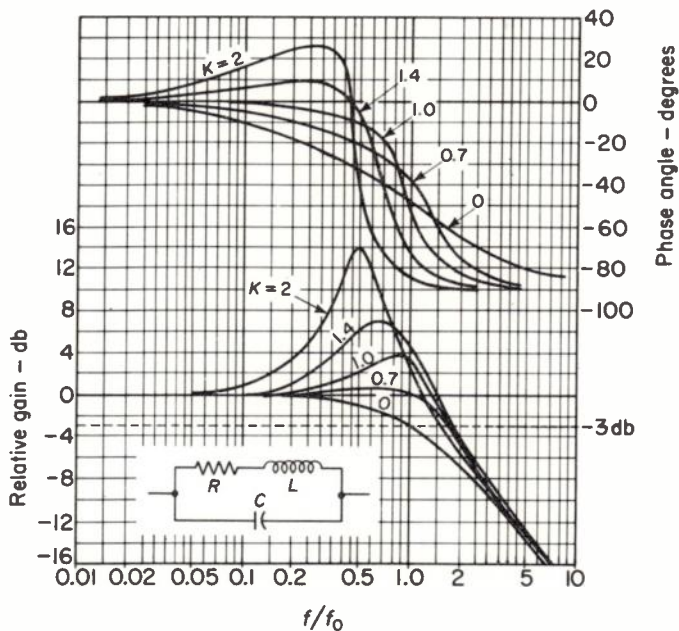


Figure 10-30. Graphical calculation of shunt compensation.

For $k = 0.64$, the gain curve has a minimum of irregularity in the passband and is said to be maximally flat. This provides the best frequency response, and the inductance value is

$$L = 0.41CR^2$$

The corresponding transient response is shown in Fig. 10-31; it has a maximum overshoot of 7 per cent. The gain is lower than in the preceding case, the gain and phase characteristics are good, and the transient response is fair.

For $k = 0.55$, the inductance value is

$$L = 0.3CR^2$$

and we have already pointed out that this case provides a nearly constant time delay, that is, the best phase characteristic. The corresponding transient response is shown in Fig. 10-32 and has the fastest rise for no overshoot. The gain is lower than for the preceding case. As may be seen from Fig. 10-30, the gain decreases with k .

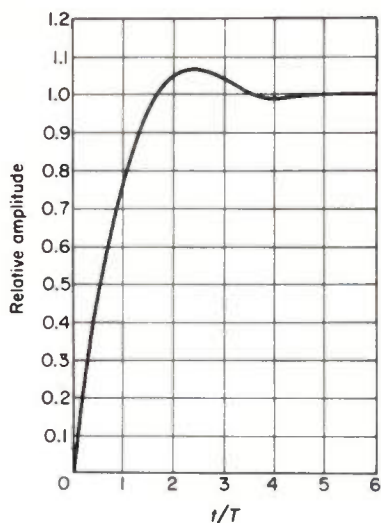


Figure 10-31. Transient response of the shunt compensation network providing best frequency response.

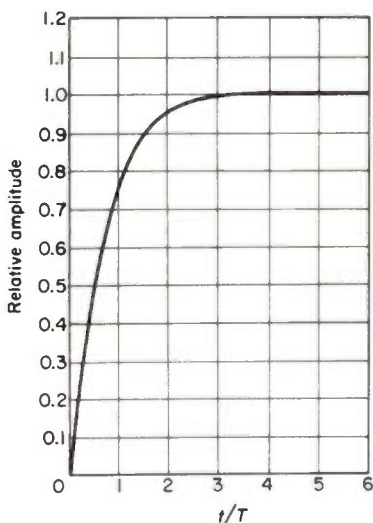


Figure 10-32. Transient response of the shunt compensation network providing best phase response (flat response).

10-19. Choice of k

No hard and fast rule can be set for the choice of k , but the following will serve to indicate the value selected:

1. $k = 0.71$ gives the maximum gain and a poor transient response, the latter tolerable only if one v-f stage is used and if some quality can be sacrificed for higher gain.
2. $k = 0.55$ provides a relatively small gain and a good transient response, which is interesting if quality is a prime consideration or if several stages are cascaded.
3. $k = 0.64$ constitutes somewhat of a compromise between the two preceding values that is especially applicable to the case of two v-f amplifying stages.

Actually, one should not attach too much importance to the overshoot obtained, for the following reasons: first, the resulting shade variation on the picture as a result of overshoot is not too important and the r-f and i-f circuits introduce a similar but unavoidable distortion of the same type; second, the ideal step voltage does not exist because of the limited pass-band of the transmitter and of the r-f and i-f amplifiers. The signal never goes up vertically, but progressively, and this reduces the actual overshoot.

In actual practice, shunt compensation networks are designed to give either maximally flat frequency response or maximum gain, that is, with $k = 0.64$ or $k = 0.71$, respectively.

10-20. Flat series and compound networks

For the series compensation network, the total parasitic shunt capacitance C is usually split into C' and C'' such that $C'' = 2C'$, and the component values

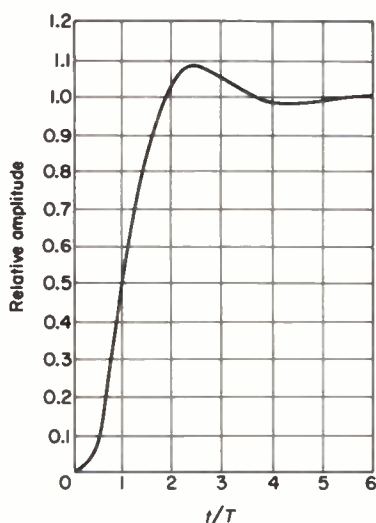


Figure 10-33. Transient response for flat series compensation network.

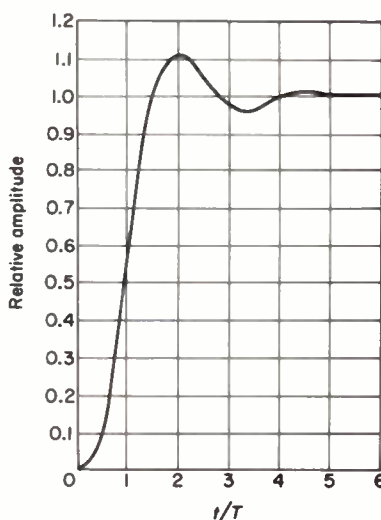


Figure 10-34. Transient response for flat compound compensation.

are $R = 1.5X$ and $L = 1.5X/\omega_0$, where X is the reactance of C at the maximum frequency f_0 .

The corresponding transient response is shown in Fig. 10-33, and practical values for 4 mc are:

$$R = \frac{60}{C}, \quad L = \frac{2400}{C}$$

in kilohms, $\mu\mu f$, and μh .

For compound compensation network, again with $C'' = 2C'$, and L_1 and L_2 being respectively the series and shunt inductances, the values are

$$R = 1.8X, \quad L_1 = \frac{1.68X}{\omega_0}, \quad L_2 = \frac{0.39X}{\omega_0}$$

The transient response is shown in Fig. 10-34. Practical values for 4 mc are

$$R = \frac{72}{C}, \quad L_1 = \frac{2688}{C}, \quad L_2 = \frac{624}{C}$$

For both series and compound networks, the component values given ensure a maximally flat frequency response.

10-21. Tube choice

Tubes used as v-f amplifiers should provide a maximum gain for a given bandwidth. As we have seen, the possible gain increases with transconductance and decreases when the parasitic shunt capacitance increases.

The conclusions are identical to those reached for r-f or i-f amplifiers, and the same remarks apply, especially with regard to the merit coefficient M .

If there is only one v-f stage, the total input capacitance comprises:

1. detection output capacitance,
2. v-f stage input capacitance,
3. wiring capacitance.

The total output capacitance comprises:

1. v-f stage output capacitance,
2. picture tube input capacitance,
3. wiring capacitance.

The last v-f stage, which drives the picture tube, must use a tube able to provide a large current swing in the linear portion of its characteristic because most picture tubes require 50 to 100 volts for full modulation and the load impedance is small.

Assuming a safe voltage swing to be 100 volts and a load of 4000 ohms, the linear current swing should be at least

$$100:4000 = 25 \text{ ma}$$

and it is evident that the required output power may well reach several watts. As a result, the v-f output tube will be a power tube. As a matter of fact, audio power tubes are sometimes used for this purpose.

10-22. Coupling detector—v-f

The coupling circuit between detector and v-f should fulfill the same requirements as the anode load of the v-f tube, and the same considerations apply. In particular, high-frequency compensation of some kind is used. Also, if a capacitance coupling is used, as shown in Fig. 10-35, low-frequency compensation by RC could be included. In this figure, high-frequency compensation is of the shunt type by L .

However, the d-c level at detector output is low, which makes it practical to use d-c coupling between the detector and v-f amplifier. The direct-coupled circuit is simpler, as shown in Fig. 10-36, and all video frequency components

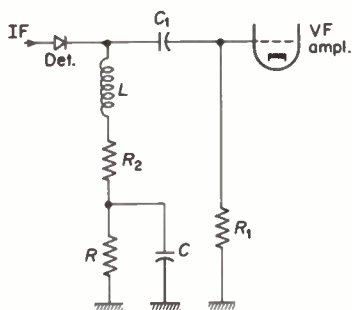


Figure 10-35. Resistance-capacitance coupling between detector and v-f amplifier.

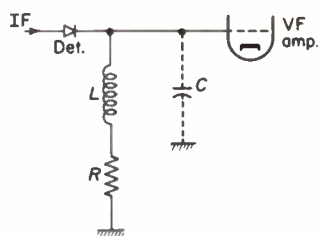


Figure 10-36. Direct coupling between detector and v-f tube.

are applied to the v-f tube, down to and including zero frequency or the d-c component. The diagram of Fig. 10-36 shows shunt compensation but series compensation (Fig. 10-37) or compound compensation (Fig. 10-38) could be used just as well.

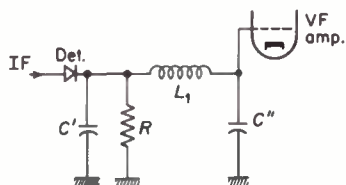


Figure 10-37. Series compound coil L_1 .

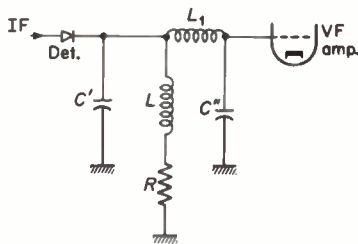


Figure 10-38. Compound compensation circuit $L - L_1$.

The last two methods shown each have the same practical advantage. The series inductance in both Figs. 10-37 and 10-38 may double in as a detector output filter to eliminate unwanted detection by-product frequencies.

With vacuum diodes especially designed for television, or crystal diodes, the wiring can be very compact and the parasitic capacitances reduced to minimum value.

10-23. V-f tube bias

When direct coupling is used between the detector and v-f amplifier, the d-c component of the video signal is applied to the grid of the tube along with the a-c components. According to the type of detection, the v-f signal will be entirely above (Fig. 10-39) or entirely below (Fig. 10-40) zero level.

In the first case, shown in Fig. 10-39, the signal can only drive the grid more positive than the quiescent level. In the second case, shown in Fig. 10-40, it can only make it more negative. The net effect is that the bias will be different for both cases so as to set the quiescent point at one end or the other of the linear part of the characteristic.

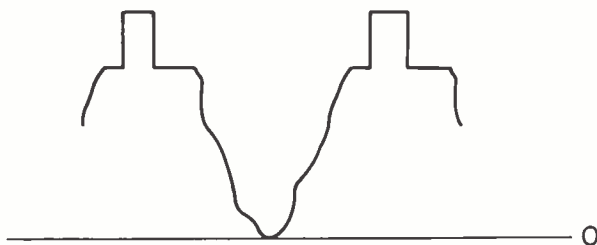


Figure 10-39. V-f signal above zero level.

For example, with the video signal in negative phase (as shown in Fig. 10-39) entirely contained in the positive region above zero reference level, the quiescent point P will be placed at the end of the linear part of the characteristic near cutoff. This is obtained with a large bias voltage V_B . The quiescent plate current of the tube is low and increases when a signal is applied.

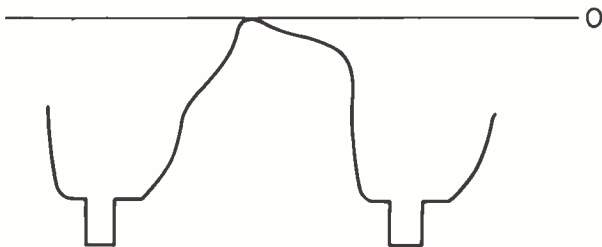


Figure 10-40. V-f signal below zero level.

For the video signal in positive phase (as shown in Fig. 10-40) entirely contained in the negative region below zero reference level, the quiescent point P will be placed at the top of the linear part of the characteristic near saturation. This is obtained with a low bias voltage V_B . The quiescent plate current is high and decreases when a signal is applied.

The circuit design shown in Fig. 10-41 provides a positive-phase v-f signal on the plate. If this signal directly drives a picture tube, it has to be applied to its grid.

In the other design, as shown in Fig. 10-42, the plate v-f signal would have negative phase and would be suitable for modulating a picture tube by the cathode.

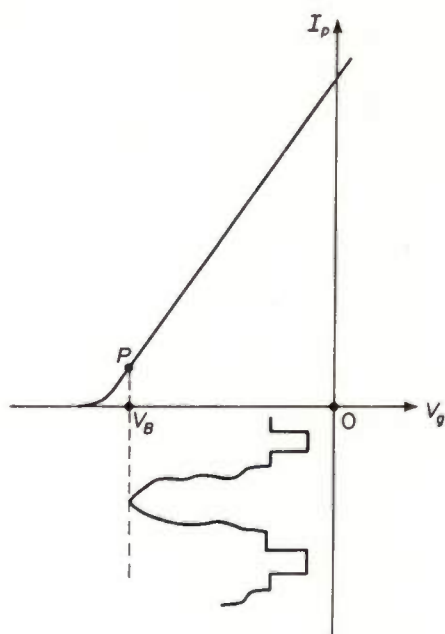


Figure 10-41. Choice of quiescent point P near cutoff.

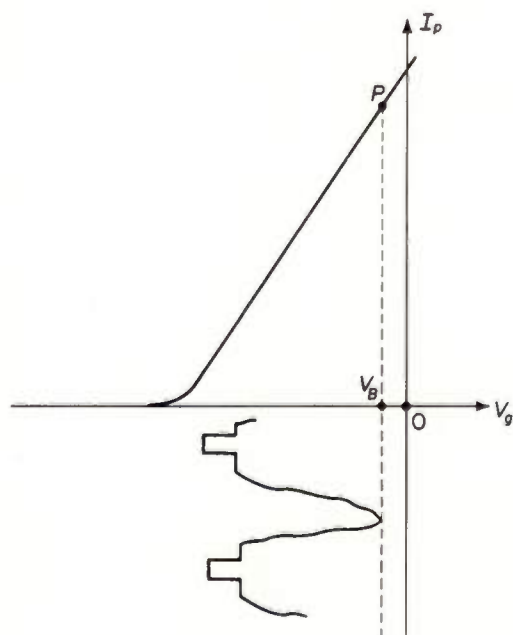


Figure 10-42. Choice of quiescent point P near saturation.

10-24. Picture tube coupling

In most receivers where there is only one v-f stage, and when it is d-c coupled to the detector, the plate signal will contain the video and d-c components of the modulation.

For the d-c signal to be faithfully reproduced, all decoupling capacitors should have an infinite capacitance value and the voltage supply should be stabilized. Neither of these conditions is satisfied in practice, but nevertheless the d-c component on the plate is close enough to its ideal value to be of practical usefulness for average brilliance setting purposes.

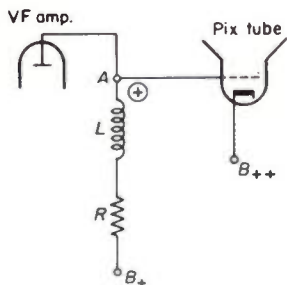


Figure 10-43. Direct coupling between v-f amplifier and picture tube.

The easy way to apply the d-c component to the picture tube, along with the v-f components, is to resort to d-c coupling between the v-f tube and the modulating electrode, cathode or grid.

Grid modulation requires a positive-phase v-f signal, and the simplest d-c coupling is direct coupling as shown in Fig. 10-43. This places the grid at a d-c voltage, equal to the plate voltage at point *A*, that is strongly positive. To obtain correct bias on the picture tube, its cathode is connected at *B* to a positive voltage higher than the voltage at *A*. In this way, the grid of the picture tube is negative with respect to its cathode, as it should be. For example, assume the supply voltage to be 250 v, the plate current 10 ma, and the load resistance *R* 4000 ohms.

The voltage drop across *R* is

$$0.01 \times 4000 = 40 \text{ v}$$

The plate voltage at *A* is

$$250 - 40 = 210 \text{ v}$$

and this is also the picture tube grid potential. Assume the normal picture tube bias is -30 v . Then its cathode is tied to a voltage of 240 v, so that the grid-to-cathode bias of the picture tube is

$$210 - 240 = -30 \text{ v}$$

as required.

The disadvantage of the circuit shown in Fig. 10-43 is apparent immediately. If, for any reason, the preceding v-f tube does not draw plate current, its anode voltage jumps up to 250 v, and so does the picture tube grid voltage. This makes the grid 10 v positive with respect to the cathode at 240 v, an unhealthy state of affairs at best.

10-25. Compromise coupling

A safer arrangement is provided by the circuit shown in Fig. 10-44. In this case, coupling between v-f tube and picture tube is done by a voltage divider R_1 - R_2 . Assume for simplicity that $R_1 = R_2$. Using the same values as before, the potential on the anode of the v-f tube would be 210 v, so that the potential on the grid of the picture tube would be

$$210:2 = 105 \text{ v}$$

For the same bias of -30 v , the picture tube cathode would be set at a voltage of 135 v.

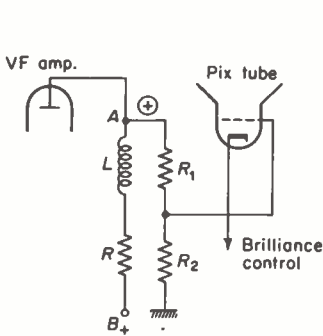


Figure 10-44. Reduction of d-c voltage.

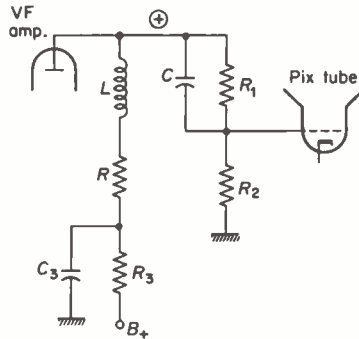


Figure 10-45. Practical circuit.

Now, if plate current of the v-f tube falls to zero, the plate voltage jumps up to 250 v as before, but the grid of the picture tube reaches only half of this value, that is, 125 v.

This still leaves a safety bias of

$$125 - 135 = -10 \text{ v}$$

on the grid of the picture tube.

The disadvantage, of course, is that if the grid of the picture tube receives only half the d-c voltage, it also receives only half of the v-f signal. This may be easily solved by the circuit shown in Fig. 10-45. A capacitance C is shunted across R_1 and all the high- and medium-frequency components of the signal are directly applied to the grid. Low-frequency components are progressively attenuated when frequency decreases, down to the d-c component for which the total gain is reduced to 50 per cent.

The problem here is similar to low-frequency attenuation by a coupling circuit and the same low-frequency compensation circuit R_3C_3 (see Sec. 10-4) is used. Evidently, compensation will be obtained for

$$R_3 = R$$

because in this case the d-c load becomes $(R + R_3)$ and has twice the value of the high-frequency load R , thus compensating for the 50-per cent loss due to $R_1 R_2$. The capacitance value is given by

$$R_3 C_3 = R_1 C$$

or

$$C_3 = \frac{C R_1}{R_3}$$

Notice that the diagram is identical with that of the previously low-frequency compensation circuit but that the components are evaluated differently. Sometimes the compensating circuit $R_3 C_3$ is omitted, and the somewhat reduced d-c component is considered satisfactory as such.

10-26. Cathode modulation

When the picture tube is modulated by its cathode, a negative-phase v-f signal is needed. Direct coupling to the v-f stage may again be used advantageously.

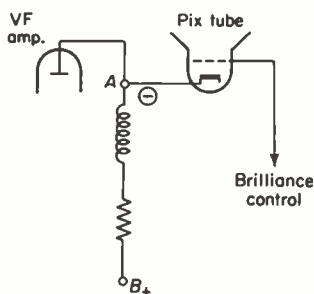


Figure 10-46. Direct coupling to the cathode of the picture tube.

A simple circuit using shunt high-frequency correction is shown in Fig. 10-46.

In this case the cathode of the picture tube is at the same potential as the anode of the v-f tube. To ensure correct bias, the grid of the picture tube is connected to a voltage slightly less positive. Assume the same values as before. The tube would have its cathode at 210 v, and for -30 v bias its grid would be at 180 v.

In the event of v-f tube failure, the cathode of the picture tube now becomes more positive, bias increases, and the picture tube is cut off so that the arrangement is safe. Moreover, parasitic capacitance is smaller for the cathode than for the grid of a picture tube. It would appear that cathode modulation would be preferable to grid modulation.

The circuit shown in Fig. 10-46 has two drawbacks. First, the heater-to-cathode insulation of the picture tube must be able to withstand the voltage applied to the cathode. Second, the first anode must be a few hundred volts positive with respect to the cathode of the tube. However, if the cathode is already at a high voltage, it might be not easy to feed the first anode. Nevertheless, in actual practice, neither drawback is a serious one. The heater-to-cathode insulation is usually sufficient to withstand the voltage applied to it, and anode voltage is provided by the boosted voltage in the horizontal sweep output stage, as will be seen later.

Regardless of the type of coupling used between v-f tube and picture tube, high-frequency compensation is always included. The compensation technique

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follows the pattern of the circuits already analyzed, and component values are established in the same way, using the picture tube cathode or grid capacitance but keeping in mind the fact that the connection between the v-f stage and the picture tube socket may well be long and a few extra micromicrofarads are added.

Both cathode and grid drive of the picture tube are used, with cathode modulation preferred for the reasons described above.

10-27. No v-f tube

Picture tubes have been designed so that they can be fully modulated with a signal amplitude of a few volts, say 10 volts peak-to-peak. It then becomes a

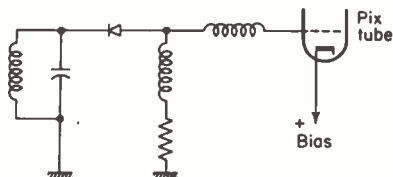


Figure 10-47. Direct coupling between detector and picture tube.

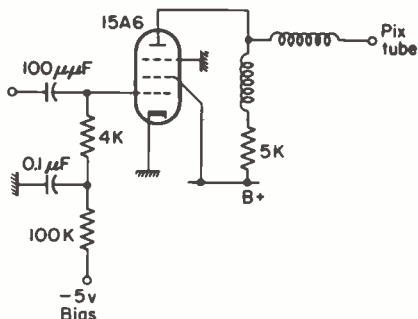


Figure 10-48. Direct coupling between plate detector circuit and picture tube.

practical proposition to eliminate v-f amplification, the picture tube being driven directly off the detector. In this case, d-c coupling is a logical choice, and a simplified circuit is shown in Fig. 10-47.

The detector is designed to work at high level, and this may create a problem relating to the last i-f stage. Assuming a detection efficiency of 50 per cent, the i-f tube driving the detector must provide a 40-v peak-to-peak signal and must be designed accordingly.

Another uncommon case is provided by the plate detector circuit shown in Fig. 10-48. In this diagram the tube is biased near cutoff by a negative voltage applied to the grid. The operation here is very similar to that illustrated in Fig. 10-41 except that the grid receives the full modulated r-f signal instead of a v-f signal and that detection is produced in the tube itself. Plate detection is rarely used, except perhaps in receivers where economy is the major consideration. The circuit shown may provide 40 to 50 volts v-f signal and uses a compound compensation for high frequencies.

10-28. Several v-f stages

On the other hand, some receivers may use two or three v-f stages. Here it is necessary to have *RC* coupling between the stages unless a high supply voltage is available, permitting the two tubes to be fed in series for d-c.

Wherever *RC* coupling is used, low-frequency compensation should (theoretically) be provided for the reason explained earlier in this chapter. In practice, however, it is frequently omitted between v-f stages.

Also, since the capacitance coupling suppresses the d-c component of the v-f signal, d-c restoration should be included. This is not always the case, and a number of receivers entirely neglect the d-c component.

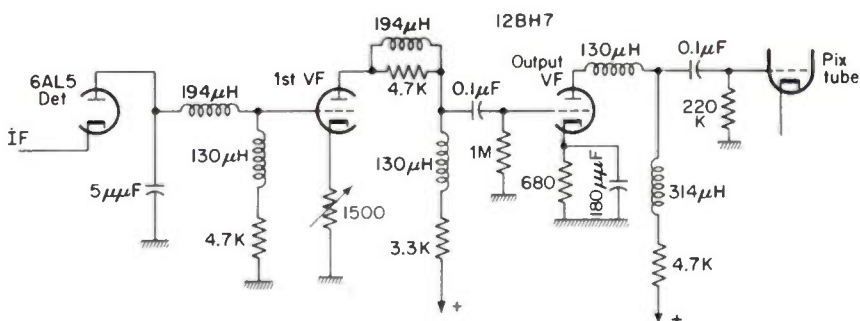


Figure 10-49. Two-stage v-f amplifier using a double triode.

A simplified diagram of a two-stage v-f amplifier using a double triode is shown in Fig. 10-49. Detection is provided by half a double diode, and the v-f signal is d-c coupled to the grid of the first triode with a compound high-frequency compensation circuit. The first v-f tube has a variable cathode resistor acting as a gain, or contrast, control. It is capacitance-coupled to the second v-f stage and compound high-frequency compensation is provided. The 4700-ohm resistor across the series inductor damps out eventual oscillations of this inductor when shock-excited by fast transients. This is a common arrangement. Also, notice that the series inductor is placed directly in the anode circuit—before the coupling circuit. This circuit design is equivalent to the previous design as far as high-frequency compensation is concerned but it leads to somewhat different values of capacitances and may be advantageous from this point of view.

The second v-f stage uses the other half of the tube. It has a cathode resistor decoupled by a low-value capacitor to provide increased gain at high frequencies. Note that compound compensation is used in the anode circuit. Coupling to the picture tube is also of the *RC* type.

Figure 10-50 shows a somewhat simplified diagram of a two-stage v-f amplifier with d-c coupling between stages. It uses a double triode tube. The

first stage is biased by the grid voltage developed across the large grid resistance. Its plate voltage is provided by the drop across the cathode resistor of the second stage and is on the order of 75 volts. Compound high-frequency compensation is used and the series inductance is damped as before. The second stage is fed a high plate voltage and is *RC*-coupled to the tube with compound high-frequency compensation.

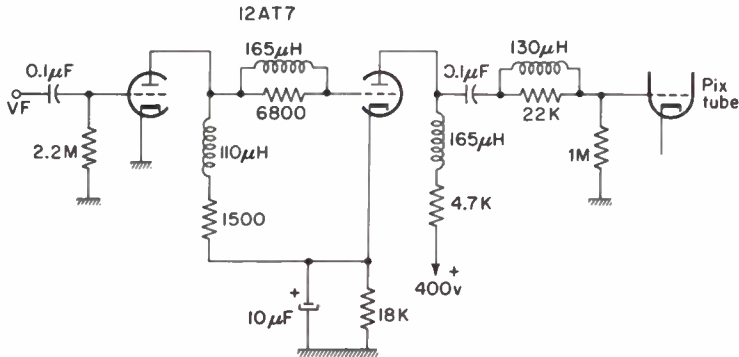


Figure 10-50. Two-stage v-f amplifier with d-c coupling.

10-29. Feedback pair

When two v-f stages are used, high-frequency compensation may be obtained simply by selective negative feedback around the pair of stages. The principle is illustrated in Fig. 10-51. A feedback resistance *R* is connected between the two load resistors *R*₁ and *R*₂. Since the two plate voltages are opposite in phase, there is negative feedback the value of which increases when *R* decreases. Moreover, parasitic capacitances *C*₁ and *C*₂ have reactances which decrease when frequency increases so that the amount of feedback decreases

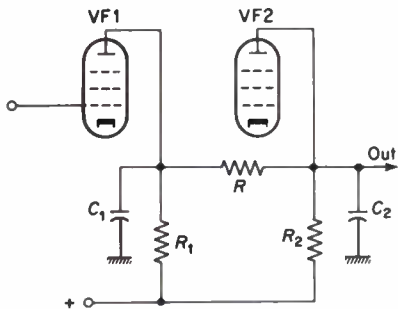


Figure 10-51. Principle of v-f feedback.

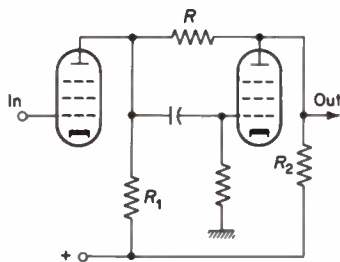


Figure 10-52. Practical feedback arrangement.

when frequency increases, thus providing relatively higher gain at high frequencies.

High-frequency compensation is thus obtained simply by using the existing parasitic capacitances at the price of the usual loss of gain due to negative feedback. When the interstage coupling circuit is added, the diagram appears as shown in Fig. 10-52.

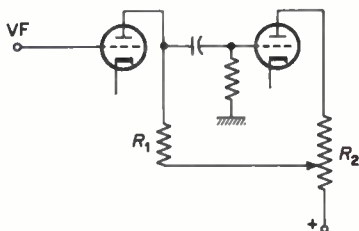


Figure 10-53. Adjustable feedback on R_2 .

Instead of using a separate feedback resistance R , a variation of the circuit uses a tap on R_2 to which R_1 is connected as shown in Fig. 10-53. The position of the tap on R_2 determines the amount of feedback. If R_2 is a potentiometer and R_1 is connected to the moving arm as shown, a simultaneous control of gain and bandwidth results.

Values of R_1 and R_2 may be adjusted in order for conditions to be exactly right, as far as bandwidth is concerned, when feedback is at maximum. The resulting simple circuit is shown in Fig. 10-54.

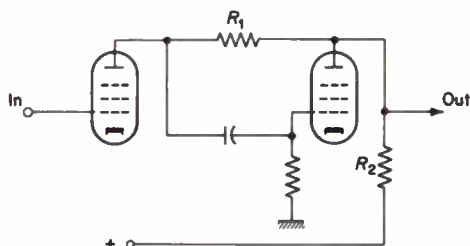


Figure 10-54. Direct plate-to-plate feedback.

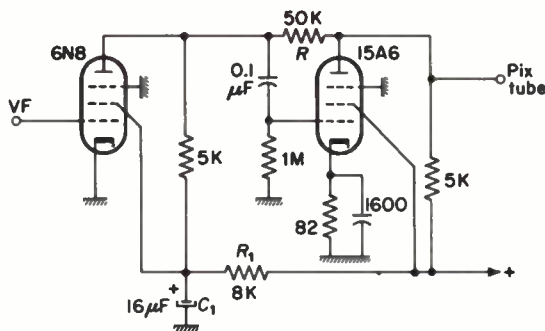


Figure 10-55. Practical v-f feedback amplifier.

A practical design is shown in Fig. 10-55. The first tube is a double diode-pentode whose diodes are used for detection. Its plate and screen voltages are supplied through a filtering and decoupling circuit R_1C_1 . The second stage has some amount of extra high-frequency compensation due to the low value of the cathode decoupling capacitance. Feedback around the second stage is provided by resistance R .

10-30. Typical designs

A few typical diagrams of v-f amplifiers will be given to illustrate the application of design principles. Since examples of two-tube circuits have already been shown, the illustrative circuit diagrams will be restricted to single-stage v-f amplifiers. Auxiliary circuits have been omitted for simplicity.

Figure 10-56 shows compound high-frequency compensation and a resistance bridge R_1R_2 at the output to reduce the d-c voltage applied to the cathode of the picture tube. Capacitance C_1 applies the full v-f signal to the picture tube. Similar circuits are used in the other diagrams.

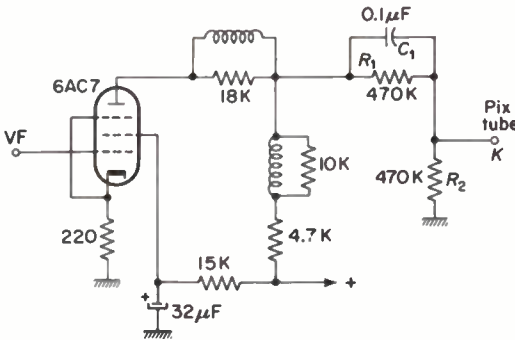


Figure 10-56. Single stage v-f amplifier.

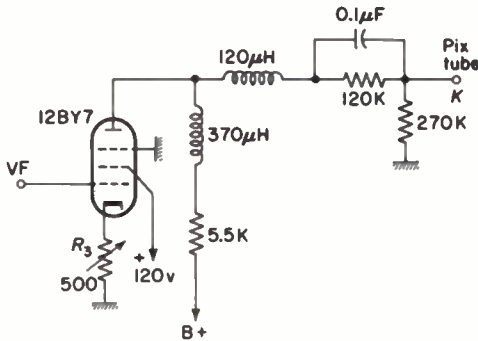


Figure 10-57. Gain control R_3 in a v-f amplifier.

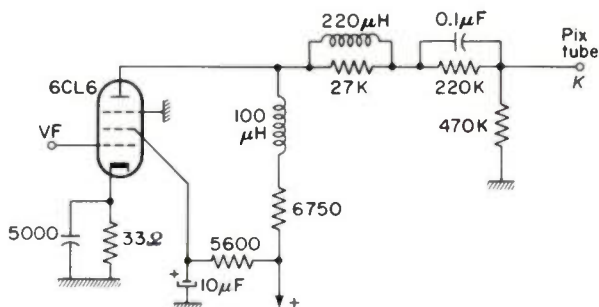


Figure 10-58. Complex high frequency compensation in a single stage v-f amplifier.

Figure 10-57 includes gain control by the variable cathode resistance R_3 , compound high-frequency compensation in its plate, and a voltage reducing circuit at the output.

Figure 10-58 has some amount of high-frequency compensation due to the low value of the cathode-decoupling capacitor and compound high-frequency compensation in its plate circuit. It also incorporates a voltage reducing circuit at the output.

Problems

1. The cathode of a tube contains a bias circuit using 100 ohms and $10\ \mu\text{f}$ in parallel. Plot the impedance of this circuit as a function of frequency.
2. The screen grid of a tube contains a decoupling circuit using 20,000 ohms and $0.1\ \mu\text{f}$. Plot the impedance of this circuit as a function of frequency. Evaluate the value of capacitance necessary for this circuit to have negligible impedance down to 5 cps.
3. An interstage coupling circuit has a capacitance of $0.01\ \mu\text{f}$ and a resistance of 100,000 ohms. Plot the output voltage as a function of frequency. The input voltage has a constant amplitude of 10 v.
4. A square wave with a 10-v amplitude and 60 cps recurrent frequency is applied to a coupling circuit using a $0.02\text{-}\mu\text{f}$ capacitor and a 330,000-ohm resistor. Plot the voltage across the resistor.
5. A tube has a 4700-ohm load resistance and is followed by a coupling circuit $0.1\ \mu\text{f}$ -100 kilohms. Design a low-frequency compensation network.
6. A tube has a 6200-ohm load resistor shunted by a parasitic $26\text{-}\mu\text{f}$ capacitor. The transconductance is 5 ma/v and the internal resistance of the tube can be neglected. Plot the gain as a function of frequency.
7. The first a-f stage of an audio amplifier is a pentode with a load of 22,000 ohms and a transconductance of 2.5 ma/v. It receives a 1-v grid signal. It is followed

by a coupling circuit $0.01\ \mu\text{f}$ -100 kilohms, and the parasitic shunt capacitance is $24\ \mu\text{f}$. Plot the voltage at the grid of the following stage as a function of a frequency.

8. Design a shunt compensation circuit for a bandwidth of 4.2 mc and a parasitic capacitance of $16\ \mu\text{f}$. The time delay should be nearly constant.

9. Design a complex two-terminal compensation circuit for a bandwidth of 4.5 mc and a parasitic capacitance of $18\ \mu\text{f}$. Maximum gain is sought.

10. Design a series compensation circuit for a bandwidth of 4.5 mc. The parasitic capacitance before the series inductor is $9\ \mu\text{f}$. The parasitic capacitance after the series inductor is $6\ \mu\text{f}$.

11. Design a series-shunt compensation circuit for a bandwidth of 4.3 mc. The parasitic capacitances are 8 and $10\ \mu\text{f}$.

12. A video stage has a transconductance of 6 ma/v. Its output total parasitic capacitance is $6.4\ \mu\text{f}$. The input total parasitic capacitance of the following stage is $8\ \mu\text{f}$. The bandwidth is 4.5 mc. Calculate the gain of the stage for no compensation, shunt, series, and series-shunt compensation. In each case, design the circuit.

13. Use a tube manual to pick a good v-f amplifier pentode. Assume $10\ \mu\text{f}$ extra parasitic capacitance for wiring and input of the following tube. Bandwidth is 4 mc. Design a cathode feedback compensation circuit. Evaluate the loss in gain.

14. A v-f tube load is 4700 ohms shunted by $18\ \mu\text{f}$. The v-f signal on its grid jumps abruptly from black to white. Calculate the width of the visible transition on the screen of a 17-in. picture tube. Use the United States standard.

15. Design a maximally flat shunt compensation circuit for a bandwidth of 4.5 mc and a parasitic capacitance of $16\ \mu\text{f}$.

16. Design a maximally flat series compensation circuit for a bandwidth of 4.5 mc and parasitic capacitances of $5\ \mu\text{f}$ and $10\ \mu\text{f}$.

17. Design a maximally flat series shunt compensation circuit for a bandwidth of 4.5 mc and parasitic capacitances of 5 and $10\ \mu\text{f}$.

18. Discuss the bias necessary for a v-f tube when no d-c component is present. Compare with the case when d-c component is retained, especially with regard to signal handling ability.

19. Draw a diagram of a detector, two-stage v-f amplifier, and picture tube. Indicate the shape of the signal at all points of interest.

20. It is suggested that a picture tube could be modulated push-pull fashion on its grid and cathode. Discuss this suggestion. Design and discuss a suitable circuit, paying attention to the d-c component.

21. Use a tube handbook to pick out a good v-f tube and design around it a complete stage of v-f amplification including low- and high-frequency compensation. Bandwidth is 4.2 mc. Make reasonable assumptions.

Chapter 11

D-C COMPONENT AND SYNC SEPARATION

The usual v-f amplifier uses capacitance coupling and does not transmit the picture d-c component. Fortunately, simple circuits are available to restore this d-c component and ensure correct relative levels in the v-f signal.

Restoration of d-c is especially necessary at the input of the sync separator, which works on an amplitude basis and functions properly only if the d-c component is present.

The sync separator separates the sync pulses from the v-f signal so that only sync pulses can be applied to the sweep sections and result in proper synchronization. Numerous sync separator circuits are known. They are always followed by some kind of sorting circuit so that vertical and horizontal sync pulses can be sorted out on a duration basis.

11-1. D-c component

We have previously shown that it is (theoretically) necessary to have the d-c component of the v-f signal present at the modulating electrode of the picture tube. The average brilliance level is thus fixed and the correct relative levels of brightness are obtained.

It is also necessary to have the d-c component present when separating the sync signals from the modulation proper. This separation is done on an amplitude basis and will be carried out properly only if all sync tips reach the same level, which occurs, in turn, only if the d-c component is present. This is illustrated in Fig. 11-1.

Figure 11-1(a) shows two signals of similar shape but different amplitude as they appear when the d-c component is present. They might represent a white line and a black line on the picture tube. With d-c component, sync tips fall at the same level in both cases and a clipper circuit working at level C will provide the output signal shown in Fig. 11-1(c), which contains sync pulses no matter what the picture modulation is.

In Fig. 11-1(b) the d-c component has been removed from the two signals, say by a coupling capacitance. An average zero level has established itself in

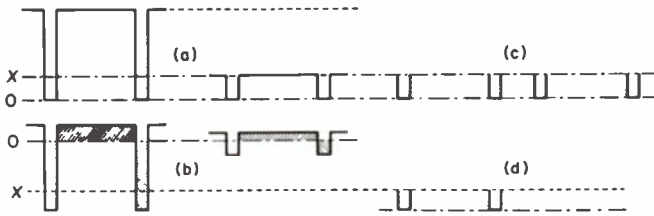


Figure 11-1. White line simplified signal with d-c component in a and without d-c component in b . The results of clipping are shown in c and d .

such a way that the shaded area above zero level and the shaded area below zero level are equal, which is another way of saying that there is zero d-c component.

Since this area equality holds good whatever the shape of the signal, the two waveforms of Fig. 11-1(a) will now look as represented. With the same clipping level set at C , the output of the clipper will be as shown in Fig. 11-1(d), where no sync pulses are obtained for low levels of picture modulation. It is obvious that some method must be devised to bring the a-c video signals to the same relative level that existed before the d-c component was removed by the previous v-f amplifiers using RC coupling. This method is called *d-c reinsertion* or *restoration*.

11-2. Diode restorer

A simple way to align all the sync tips on the same level, that is, to reinsert a d-c component in the v-f signal, is shown in Fig. 11-2. It uses a diode, either tube or crystal.

Capacitor C removes the d-c component, if any, from the v-f signal and applies this d-c free (a-c) signal to the diode. During negative alternations, shown as A in Fig. 11-2, the diode conducts. During positive alternations, B , the diode is reverse biased and is cut off.

Let the applied voltage be V_e as shown in Fig. 11-3 and the voltage across the capacitor be V_c . During a negative half-cycle, V_e has the polarity indicated

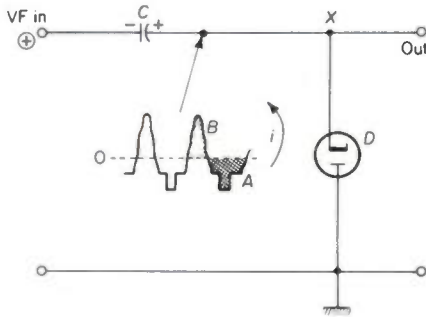


Figure 11-2. D-c component restoration.



Figure 11-3. Voltage distribution in the restoration circuit.

and the diode allows a current i to flow as shown. This current charges up capacitor C with the polarity indicated. As long as the diode provides current, it keeps on charging up C , but it will be noticed that the voltage V_c across C is opposite in polarity to the input voltage V_e . Therefore, the voltage V_d across the diode is

$$V_d = V_e - V_c$$

and decreases when C charges up. Equilibrium is attained when no current flows and

$$V_d = 0$$

that is, when

$$V_e = V_c$$

so that the capacitor has charged up to the negative peak voltage of the input signal.

This charging-up process occurs very quickly because of the low resistance of the diode when conducting. The result is such that, after the short charging period, the voltage at point X never goes negative. It goes down to zero only for negative peaks of the input signal. This means, precisely, that the negative peaks of the output signal will be aligned on zero, or that d-c component has been reinserted.

During the positive half-cycles B of Fig. 11-2, the diode is cut off, and a resistor R must be provided to close the circuit as shown in Fig. 11-4.

But now C can discharge through R with time constant RC , upsetting the operation of the d-c restorer. To avoid this, or rather to reduce it to tolerable proportions, time constant RC should be made large in comparison with the period of the input signal.

If CR is a coupling circuit, this condition is already satisfied. The mere presence of the shunting resistor, however, corresponds to a small discharge of

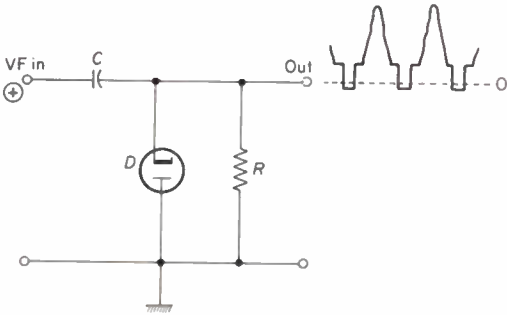


Figure 11-4. A resistance R closes the d-c path.

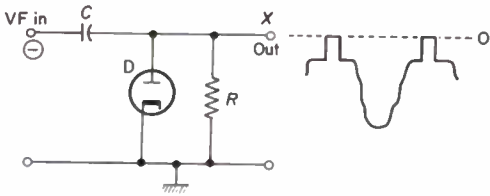


Figure 11-5. D-c restoration for positive phase v-f signal.

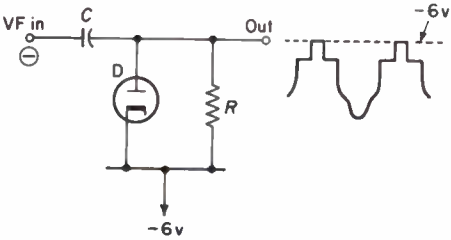


Figure 11-6. Setting the d-c restoring level.

the capacitor between negative peaks, so that the output signal is not exactly aligned on zero, but slightly below. The difference is practically negligible. This type of circuit, acting on a negative-phase v-f signal, is called a *negative d-c restorer*.

If the v-f signal happens to be in positive phase, all that is necessary is to reverse the diode, as shown in Fig. 11-5, and a positive d-c restorer is obtained which will align all sync tips on zero level and keep the v-f signal below zero.

In practice, alignment can be obtained at any voltage level by the simple expedient of setting the entire d-c restoring circuit at that level. For example, as shown in Fig. 11-6, the signal will be aligned on -6 volts with respect to ground.

11-3. Grid restoration

Normally, the output of the d-c restorer would be applied to the grid of the following stage.

If the signal is in negative phase, the diode restorer can be omitted as shown in Fig. 11-7. The cathode and control grid of the tube may replace it, exactly as in a grid detector.

Since there is no bias provided, the sync tips drive the grid positive and grid current charges up the capacitor C as before. The voltage developed

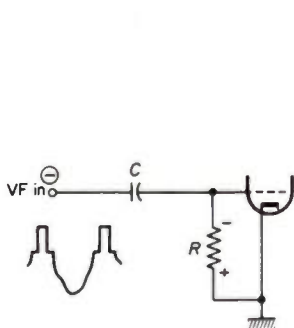


Figure 11-7. Grid d-c restoration.

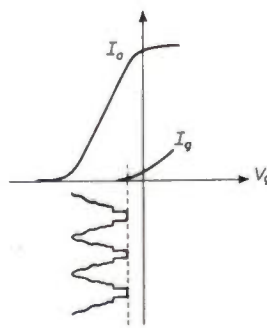


Figure 11-8. Operation of grid d-c restoration.

across R creates a negative grid bias just sufficient to maintain the positive peaks of the signal at the onset of grid current. The corresponding characteristic curve is shown in Fig. 11-8. Grid current begins for a slightly negative voltage, say -0.5 v and sync tips are aligned at this value. Although not quite

correct, for simplicity it is often said that grid detection in this circuit aligns the grid signal on zero level.

The output signal on the plate of the tube is then also aligned on some positive voltage value.

If the tube is driven by a positive-phase v-f signal, it becomes necessary to add a d-c restorer across its grid leak resistance, as shown in Fig. 11-9. The bias of the

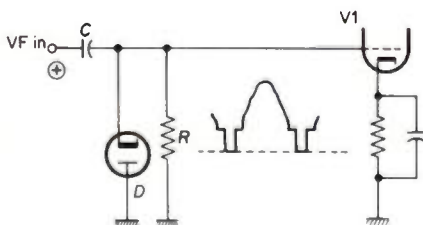


Figure 11-9. Diode d-c restorer in grid circuit.

tube must be high enough so that the grid can never be driven positive by the signal. If it were, the grid would detect and the circuit would contain *two* diodes with opposite polarity at the input.

Usually the quiescent point is set at the low end of the linear part of the characteristic, near cutoff.

11-4. Cathode restoration

Actually, any detector circuit may replace the diode of the d-c restorer with identical results. It has just been done for the grid detector, and it may just as well be done for the cathode detector.

As previously shown, cathode detection is obtained by inserting a large-value resistance in the cathode circuit of a tube. When applied to d-c restoration, the diagram appears as shown in Fig. 11-10. A peculiarity of this circuit makes it very different from the other two. For no signal, the high cathode

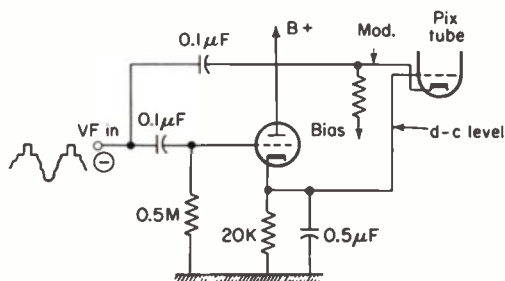


Figure 11-10. Cathode restoration.

resistance maintains the tube virtually at cutoff, and only the sync tips drive it into conduction. The output voltage on the cathode then contains only the aligned sync pulses, but the modulation part of the v-f input signal has been clipped off. It falls beyond cutoff on the tube characteristic.

The cathode restorer only provides sync pulses and d-c level. Modulation must be applied through another channel. A convenient method is to use both grid and cathode of the picture tube as shown.

11-5. Noise sensitivity

When noise pulses are superimposed on the v-f signal, they appear as sharp spikes of large amplitude and the d-c restorer, following the maximum instantaneous amplitude of the signal, will exhibit a high sensitivity to noise. When the coupling capacitor has been charged up to a large value by a noise pulse, it discharges slowly because of the long time constant, so that the noise pulse is artificially lengthened and appears on the screen as a streak instead of a point.

Clipping at maximum signal level might be used to reduce the effect of noise pulses. However, a simpler way is to reduce the effectiveness of d-c restoration, either by inserting a resistance of a few kilohms in series with the diode, or by reducing the shunt resistance to a value of, say, 100 kilohms to reduce the time constant. Both methods obviously are detrimental to good d-c restoration.

A variation is to feed the diode d-c restorer through a low-pass filter which will almost eliminate the noise pulses and (at the same time) the high-frequency components of the v-f signal. The d-c restorer provides only the d-c or low-frequency components and high-frequency components must be applied through a separate channel, as in the case of a cathode d-c restorer.

A suitable circuit is shown in Fig. 11-11, where both methods are used. The d-c restorer is fed by a shorter than usual time-constant circuit. Its output is smoothed out by the filter RC before being applied to the grid of the picture tube.

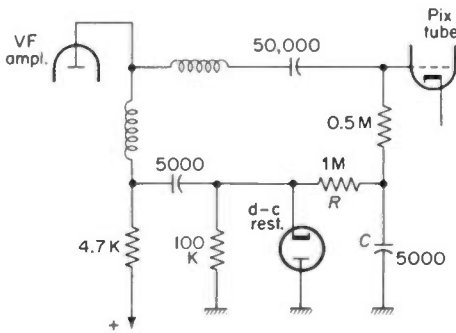


Figure 11-11. Reduction of noise sensitivity.

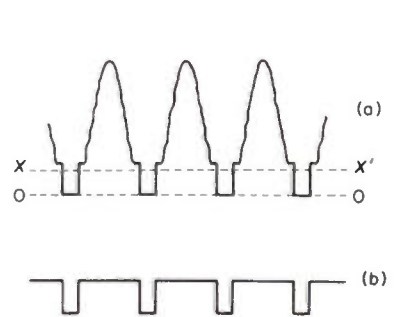


Figure 11-12. Separation of synchronization pulses.

11-6. Synchronization separation

Synchronizing pulses must be separated from the modulation part of the v-f signal before being applied to the horizontal and vertical sweep oscillators.

Sync separation is accomplished on an amplitude basis by means of clipping. The v-f signal with its d-c component is shown in Fig. 11-12(a). A simple clipping at level XX' or lower will eliminate the upper portion of the v-f signal. The resulting signal is shown at (b) and contains only the sync pulses or, more precisely, the lower part of the sync pulses.

Any type of clipper circuit may be used for sync separation purposes. As a matter of fact, the cathode d-c restorer previously described constitutes a sync separator since its output contains only the sync pulses and no v-f modulation.

A simple diode might just as well be used with its clipping level adjusted at the proper value. In actual practice, the only circuits used are those that ensure sync separation over a wide range of v-f signal amplitude without adjustment.

11-7. Shunt diode

Figure 11-13 shows a shunt diode with the clipping level determined by the cathode bias V . Assume that the clipping level XX' shown in Fig. 11-12

corresponds to 3 volts. The diode cathode is biased at +3 v. As long as the input signal is below the threshold level of 3 v, the diode plate is less positive than its cathode, the diode is nonconducting, and the input signal appears across R . When the instantaneous level of input signal exceeds 3 v, the diode anode becomes positive with respect to its cathode, and the diode conducts and shorts the output. Resistance R is on the order of 10,000 ohms and is introduced to prevent the diode from shorting the input at the same time.

Thus, only that part of the signal at a level below 3 v, that is, the lowest part of the sync pulses, appears at the output. There is a difficulty, however,

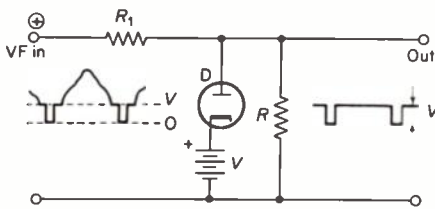


Figure 11-13. Parallel diode separator.

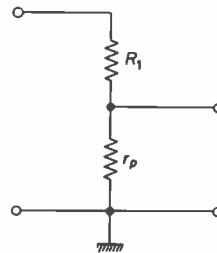


Figure 11-14. Equivalent resistance divider.

due to the diode internal resistance r_p . Assume that R is comparatively large and can be neglected. When the diode conducts, the equivalent circuit appears as shown in Fig. 11-14. A fraction $r_p/(R_1 + r_p)$ of the input voltage appears at the output. Assuming $R_1 = 10,000$ ohms and $r_p = 2000$ ohms, it can be seen that 17 per cent of the input signal appears at the output. It is not possible to increase much the value of R_1 because of the resulting distortion of the signal. In practice, the residual percentage of v-f modulation at output is reduced to some 10 per cent, and the shunt diode, therefore, is not too good a sync separator.

For a v-f signal of opposite phase, the diode can be reversed and used in a similar circuit.

11-8. Series diode

The diode sync separator may be series-connected, as shown in Fig. 11-15, where V defines the clipping level. As long as the instantaneous level of input signal keeps below V , the diode conducts and the signal appears at output. When the instantaneous level at the input exceeds V , the diode cathode becomes more positive than its anode and the diode cuts off, thereby disconnecting output from input. The output signal contains only that part of the input signal below V volts, that is, the sync pulses as indicated.

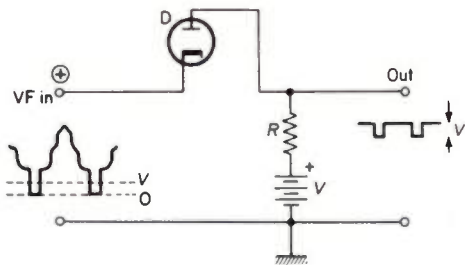


Figure 11-15. Series diode separator.

If the v-f input signal has negative phase, the diode and battery are reversed as shown in Fig. 11-16. In actual practice, of course, the clipping level V is obtained not by a battery, but by using d-c voltages existing in the receiver.

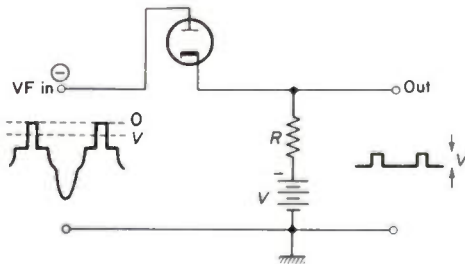


Figure 11-16. Negative phase series diode separator.

Figure 11-17 shows how the plate supply voltage can be put to work to accomplish this purpose.

This time the difficulty is due to the parasitic capacitances or, more precisely, to the diode interelectrode capacitance C and to the total parasitic shunt capacitance C' across the output.

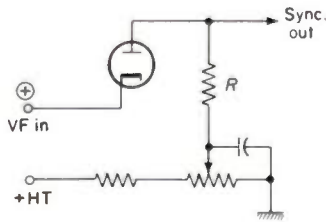


Figure 11-17. Adjustable separation level.

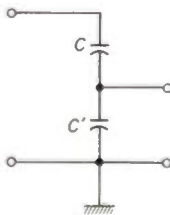


Figure 11-18. Equivalent capacitive divider.

When the diode is nonconductive, these two capacitances constitute a voltage divider as shown in Fig. 11-18, so that a fraction $C/(C + C')$ of the input voltage appears at the output. Assume a diode capacitance C of $1\ \mu\text{mf}$ and an output capacitance C' of $10\ \mu\text{mf}$. Then $1/(1 + 10) = 0.09$, so that 9 per cent of the input signal appears at the output. To reduce this proportion, the diode should have a very low interelectrode capacitance, which is the case for a crystal diode. Also, an extra capacitor could be added across the output. It should have the largest value compatible with the tolerated signal distortion it may introduce.

11-9. Triode separator

Triode or pentode tubes are commonly used as sync separators because they provide good separation without the need to adjust a clipping level and because they ensure at the same time amplification of the separated sync pulses.

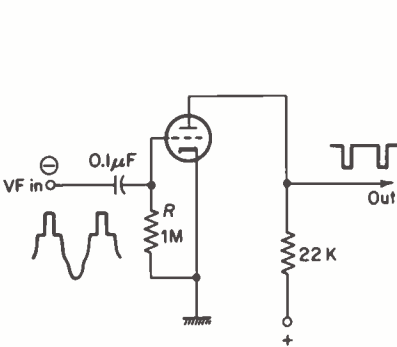


Figure 11-19. Grid detection triode separator.

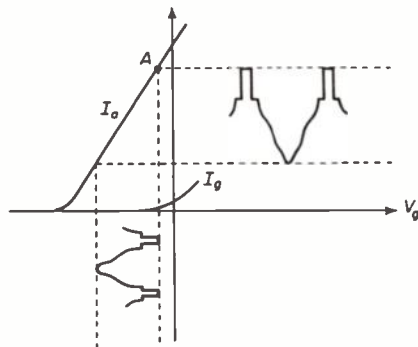


Figure 11-20. Operation of an amplifier tube.

Figure 11-19 shows a triode sync separator. The v-f signal in negative phase is applied to the input and the tube works as a grid detector. Separated sync pulses appear across the plate load resistance. Since the diagram looks so very much like a grid d-c restorer, one may wonder why it should work as a separator. The difference lies in the grid cutoff voltage.

In the case of the d-c restorer, shown in Fig. 11-20, the quiescent point is at A , at the onset of grid current. The signal amplitude and the tube working conditions are such that the whole input signal falls inside the grid base and appears amplified at the plate.

In the case of the separator, shown in Fig. 11-21, the quiescent point A is identically placed. This time, however, input signal amplitude and tube working conditions are such that the largest part of the v-f signal, including all

video modulation and part of the sync pulses, falls beyond cutoff and is not amplified by the tube. The plate output signal then contains only the separated sync pulses.

Under suitable conditions, separator tubes may have a cutoff voltage as low as -1 v. Assuming 25 per cent sync in the v-f signal, no modulation will appear at the output as long as the input v-f signal has a minimum amplitude of 4 v. Since the v-f signal on the picture tube is on the order of 50 to 100 v, there is a wide safety margin.

Despite the short grid base, and because of tube amplification, the separated pulses at the plate may easily have an amplitude of from 10 to 40 v.

Notice that the tube provides its own d-c restoration, and that the input v-f signal may or may not have its d-c component. It must, however, be in negative phase.

Separation is actually obtained by plate current cutoff. The same principle can be used with other tubes, especially pentodes, with some attending advantages.

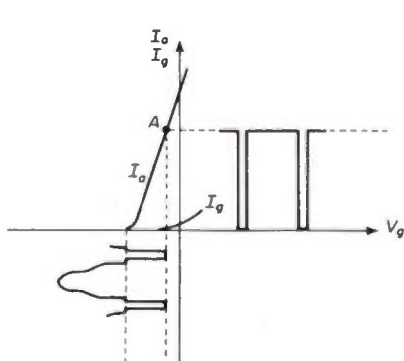


Figure 11-21. Operation of a separator tube.

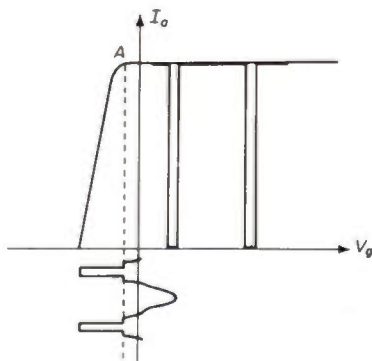
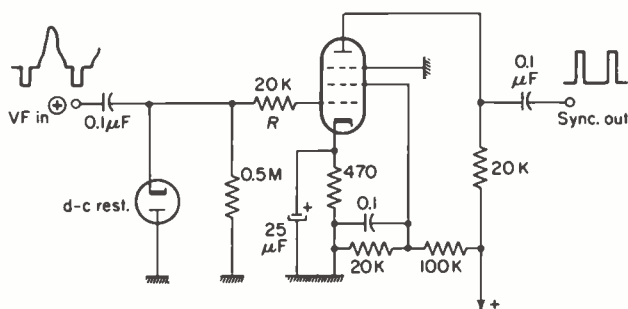


Figure 11-22. Separation at saturation.

11-10. Pentode separator, positive phase

A pentode separator can be used exactly as the triode in the preceding circuit, a pentode replacing the triode. It offers other possibilities, in addition. By reducing both screen and plate voltages, a characteristic curve may be obtained which has a short grid base and which also saturates for low values of plate current.

Such a curve, as shown in Fig. 11-22, may be used to obtain separation by saturation of plate current. A positive v-f signal of sufficient amplitude is applied to the grid, and the d-c component must be present for all sync tips to be aligned. Working conditions are chosen so that saturation at A occurs

Figure 11-23. Grid limiting resistance R .

below modulation level of the v-f input signal. The plate output signal then contains only the separated and amplified sync pulses.

If the input signal is large enough, sync tips may fall beyond cutoff and the scheme provides double clipping of the sync pulses.

Notice that because the input signal is in positive phase, a diode d-c restorer must be used on the grid. The corresponding diagram is shown in Fig. 11-24. The grid is positive during the modulation part of the v-f signal, and the low grid-cathode impedance under these conditions (on the order of 1000 ohms) loads the input circuit. To reduce this effect, a resistor R is added in series in the grid circuit. It should have the highest value which will not distort the sync pulses excessively. This arrangement, incidentally, has two beneficial side effects:

1. It isolates the input capacitance of the separator tube from the input circuit.
2. It reduces the amplitude of picture modulation with respect to sync pulse amplitude in the following manner. During sync pulses, the grid is negative and receives all the input signal. During modulation, the grid is positive and receives only a fraction of the input signal because its low grid-to-cathode resistance constitutes a voltage-divider with R , the added series grid resistor. The value of R cannot be made very large. With the tube input capacitance, it constitutes a low-pass filter which distorts the pulses. If R is too large, the distortion becomes intolerable.

11-11. Pentode separator, negative phase

With a v-f input signal in negative phase, a pentode can ensure separation by cutoff as did the triode, with the grid detection taking care of d-c restoration. Moreover, with low screen and plate voltage, saturation of plate current will occur at low currents and double clipping of the signal may be obtained as shown in Fig. 11-24. The separated sync pulses in the anode have constant

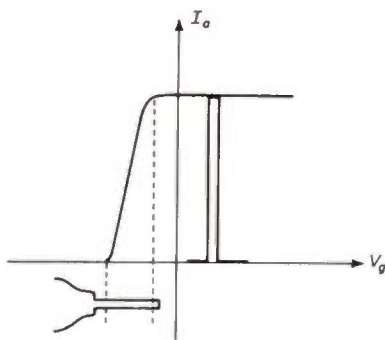


Figure 11-24. Double clipping.

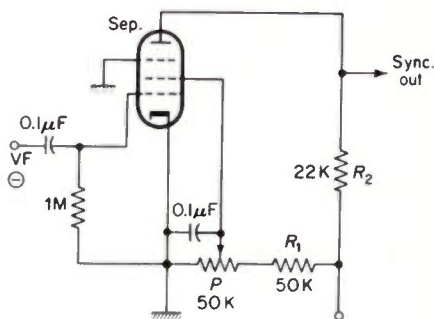


Figure 11-25. Adjustable screen grid voltage.

amplitude. In addition, at the input the top of the sync pulses may contain noise, which is eliminated at the output by the saturation clipping.

This arrangement, in fact, cuts a slice (double limiting) in the sync pulses, the height of the slice being equal to the separation between cutoff and saturation. As in all sync separators, the v-f input signal amplitude should be as large as possible to ensure wide safety margins. It should be pointed out that, because of nonlinearity in the amplifier stages, it may well be that the proportion of sync pulses is below the nominal 25 per cent of the total v-f signal.

Generally, the v-f signal for the separator is taken in the picture receiver where it has a maximum amplitude, that is, after the last v-f amplifier.

The screen voltage could be adjusted to set the working conditions for the tube, as shown in Fig. 11-25. However, this is almost never done and a fixed screen voltage of fairly low value, say on the order of 20 volts, is obtained by a resistance divider or sometimes with the help of a large series dropping resistance.

11-12. Tube operation

The separator should evidently use a sharp cutoff tube. Furthermore, it is interesting to have a high-transconductance tube for reasons somewhat similar to those put forward for amplifiers.

The separator load resistance should be large in order to provide large-amplitude sync pulses but its value is limited by pulse distortion, especially for line pulses.

Line pulses are rectangular pulses with a repetition or recurrence frequency of 15,750 cps, having a short rise time. Such a rectangular signal can be decomposed into a fundamental sine wave of 15,750 cps and a large number of

harmonic sine waves. It may be assumed that distortion can be tolerated if harmonics up to the 20th are correctly reproduced, higher harmonics being attenuated. The corresponding frequency limit is

$$15,750 \times 20 = 315,000 \text{ cps}$$

or roughly 0.3 mc, and this should be the bandwidth of the separator stage. Assuming a parasitic total shunt capacitance of $20 \mu\text{mf}$, a bandwidth of 0.3 mc at 3 db will be obtained for a load resistor

$$R = \frac{1}{C\omega} = \frac{1}{20 \times 10^{-12} \times 3 \times 10^5 \times 2\pi} = 27,000 \text{ ohms}$$

approximately.

Notice that when the separator tube conducts, that is, during sync pulses, its plate resistance shunts the external load resistance and reduces its effective value.

The relatively small bandwidth needed for correct sync pulse handling makes it easy to obtain extra amplification if needed. Sometimes extra amplification is provided before the separator. More frequently, it occurs after separation and is combined with a second clipping of the pulses.

For the same reason, the v-f signal to drive the separator is usually taken after the high-frequency compensation circuits so as not to increase the parasitic shunt capacitance on the v-f amplifying tube unnecessarily. The signal is then taken across the v-f tube load resistance R only. Moreover, a series resistance R_1 of a few thousand ohms is frequently connected in series at the take-off point, as shown in Fig. 11-26. Its role is to isolate the separator input capacitance from the v-f tube. Its value is chosen sufficiently small so as not to distort the sync pulses.

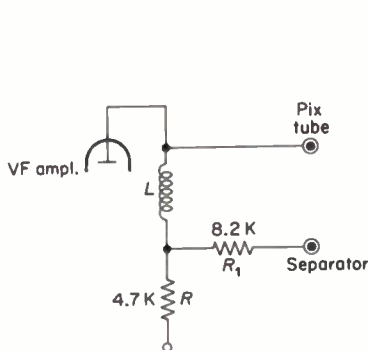


Figure 11-26. Resistance R_1 isolates the separator input.

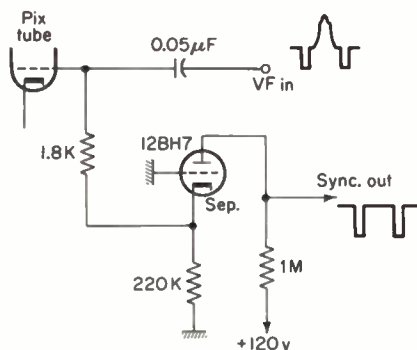


Figure 11-27. D-c restoration by the grid-cathode diode, with sync separation in the plate circuit.

11-13. Separation at d-c restorer

A separate clipper tube is not absolutely essential for sync separation if the receiver contains a d-c restorer. As we have seen, a d-c restorer conducts only during the tips of the sync pulses, so that its anode current contains the separated sync pulses and the d-c restorer can double as a sync separator. The only necessary modification is the addition in the circuit of a load resistance across which are obtained the separated sync pulses. Their amplitude is generally small, but the sync pulses can be increased by suitable bias of the d-c restorer, with detrimental effects on d-c restoration.

A better way of accomplishing combined d-c restoration and sync separation is to use the grid and cathode of a triode as d-c restorer diode, and to use the anode circuit to obtain the separated sync pulses. A diagram illustrating this technique is shown in Fig. 11-27, where d-c restoration is obtained in the picture tube grid circuit and sync pulses are obtained in the plate circuit of the triode.

11-14. Other circuits

The separator tube may carry load resistors in both anode and cathode circuits, as shown in Fig. 11-28. This provides output sync pulses of both polarities.

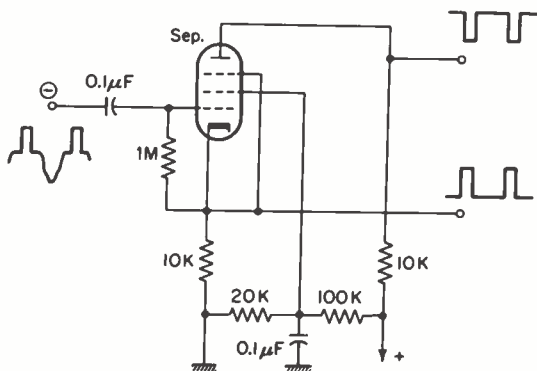


Figure 11-28. Phase splitting at the separator.

When a positive-phase separator is used, there is a d-c restorer diode in its grid circuit and there is no need to connect another one in the grid of the picture tube. The same diode performs both functions by connecting the picture tube grid to that point on the diode cathode, as shown in Fig. 11-29, where the signal is present with its d-c component.

Two diode separators are shown in Figs. 11-30 and 11-31. Both are d-c coupled to the v-f amplifier. In the first figure, a series diode receives the v-f

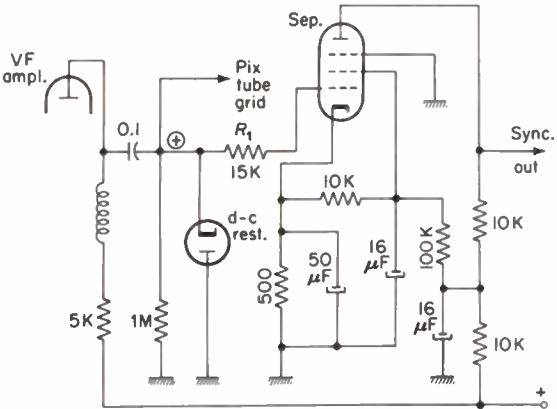


Figure 11-29. The separator d-c restoration diode can also be used for the picture tube.

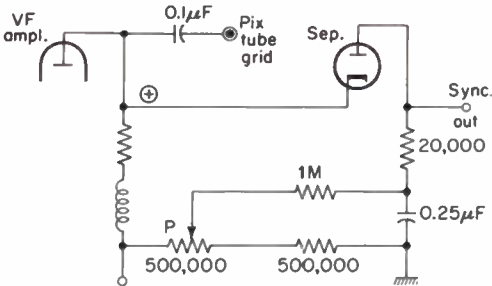


Figure 11-30. D-c coupling to a series diode separator.

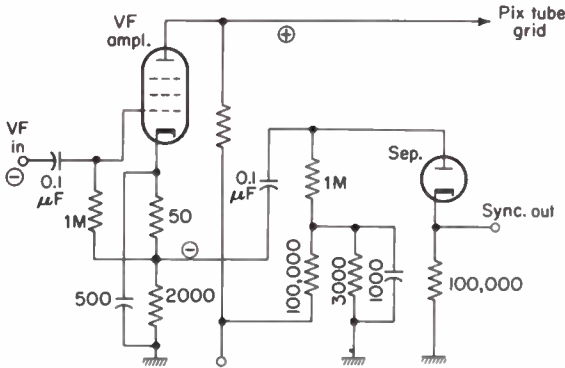


Figure 11-31. D-c coupling to a series diode separator for negative phase.

signal in positive phase from the anode of the v-f tube and has a clipping level adjusted by P . In the second, a series diode is driven by the cathode of the v-f tube and has a fixed clipping level determined by a resistance divider across the plate supply voltage.

If the available v-f signal happens to be in positive phase and a grid detection type of separator is to be used, a good solution is to amplify the v-f signal before separation and at the same time reverse its phase. A suitable circuit, which uses a double triode tube, is shown in Fig. 11-32. The first triode receives the v-f signal in positive phase from the anode of the v-f amplifier. Resistor R at the input isolates the tube input capacitance and, in combination with it, constitutes a low-pass filter to reduce bandwidth and noise sensitivity. The v-f signal appears in negative phase on the anode and is applied to the second triode, working as a sync separator.

Figure 11-33 shows a commonly used circuit of pentode separator for positive-phase v-f signal in which a d-c restorer diode is used in the grid circuit. The separated sync pulses are positive-going. Figure 11-34 represents a commonly used circuit of pentode separator for negative-phase v-f signal. The screen voltage is fixed by a resistance divider as shown. In a variation shown in Fig. 11-35, screen voltage is fixed by a large-value series dropping resistance. Both circuits provide negative-going separated sync pulses.

11-15. Typical circuits

A wide variety of sync separation circuits are available. A few examples will be given to show how the principles previously explained have been applied to practical designs.

A double triode is used in the circuit shown in Fig. 11-36. The first tube works as a separator. Notice the provision of grid detection and the low plate voltage. The second tube is a d-c restorer and clipper similar to the circuit previously described but working on sync pulses only. The arrangement provides sync pulses of constant 20-v amplitude. Pertinent voltages have been indicated on the diagram.

A practical example of a triode used simultaneously as a d-c restorer and a sync separator is shown in Fig. 11-37.

Figure 11-38 shows a version of the pentode separator for a v-f signal of negative phase. Notice the low values of screen and plate voltage.

11-16. Pulse discrimination

The output of the sync separator provides sync pulses of constant amplitude but the problem of discriminating between the horizontal and vertical pulses and applying them to the corresponding sweep oscillators still remains.

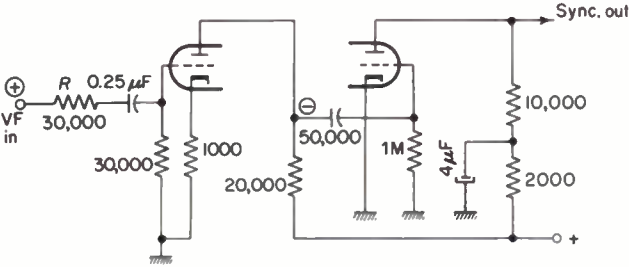


Figure 11-32. Combined sync amplifier and separator.

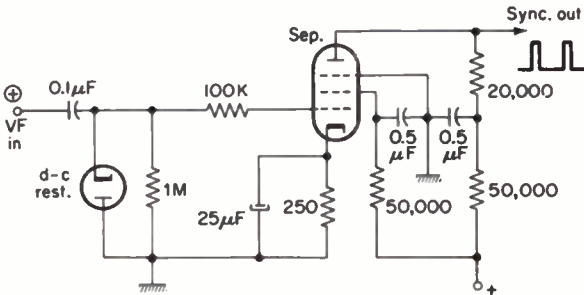


Figure 11-33. Positive phase pentode sync separator with diode d-c restorer.

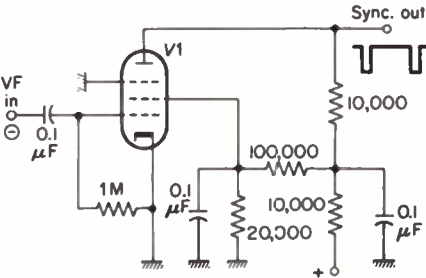


Figure 11-34. Resistance bridge for screen grid voltage.

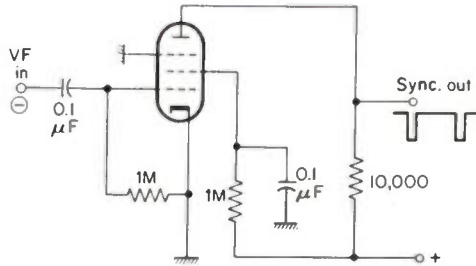


Figure 11-35. High series dropping resistance in screen circuit.

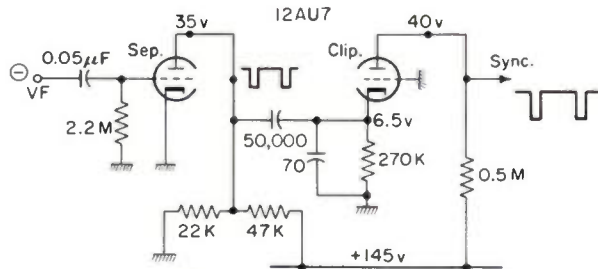


Figure 11-36. Sync separator and clipper.

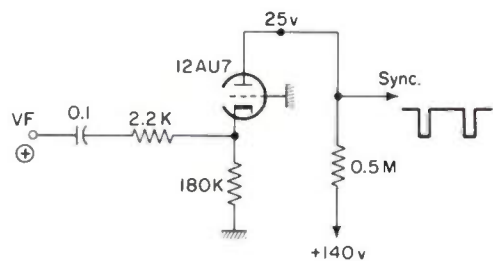


Figure 11-37. Combined d-c restorer and sync separator.

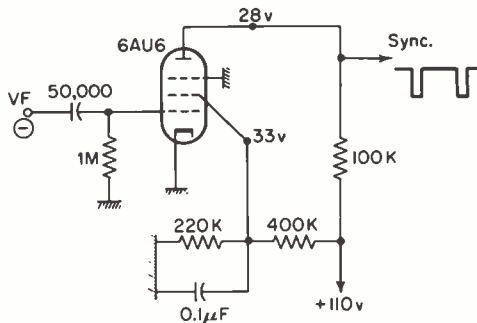


Figure 11-38. Typical pentode sync separator.

Pulse separation is based on the duration difference between the horizontal and vertical pulses. The horizontal pulse is short and has a duration of 4.5 microseconds. The vertical pulse is long and has a duration of 190 microseconds. It is interrupted by line pulses, as described in Chap. 1 when the standards were studied.

A common way to sort the pulses is to use a circuit which transforms duration variations into amplitude variations. Simple examples of such circuits are the differentiator and the integrator. Essentially, a *differentiator* may be considered as a circuit sensitive to short pulses and an *integrator* as a circuit sensitive to long pulses.

Another important problem is premature synchronizing of the vertical oscillator. Premature synchronizing may be due to the pulse produced by the horizontal oscillator when it switches over. If this switching pulse is allowed to reach the vertical oscillator, it will trip it before the vertical pulse does. Since the normal synchronizing vertical pulse occurs between two line pulses for every odd field to ensure interlace, the error due to the line pulse fed back from the horizontal oscillator upsets the interlace.

Precautions must therefore be taken to make certain that the back pulse from the horizontal line oscillator has negligible amplitude when it reaches the vertical field oscillator. A good method is to insert a unidirectional device (for example, a tube) that will let the sync pulse pass through from grid to plate but will not allow the back pulse to work its way back from plate to grid.

Widely different schemes are used, as we will see shortly.

11-17. Differentiating circuit

An elementary differentiating circuit may be made of a capacitor C and a resistor R as shown in Fig. 11-39. V_i is the input voltage and V_o the output voltage. Let i be the current through the circuit due to the applied voltage V_i .

If the RC product is small, that is, if both R and C are small, the reactance of C is

$$X = \frac{1}{\omega C}$$

and is large compared to R . The current is then essentially determined by the capacitance and has a value

$$i = C \frac{dV_i}{dt}$$

The voltage across R is the output voltage V_o :

$$V_o = iR$$

so that

$$V_o = RC \frac{dV_i}{dt}$$

The output voltage is proportional to the derivative of the input voltage, hence the name of the circuit.

The smaller the RC product, the better the approximation made to calculate i . Unfortunately, RC appears in the expression for V_o , so that the better the differentiation, the smaller the output signal.

In practice the differentiation is considered as acceptable if

$$RC \leq \frac{t_1}{4}$$

where t_1 is the pulse duration. The units are ohm, farad, and second, but the same equation applies for more practical units, namely megohm, $\mu\mu f$, and μsec .

For horizontal pulses the differentiator time constant RC should then be on the order of 1 microsecond or less.

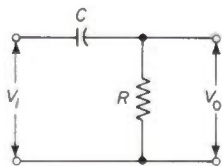


Figure 11-39. Elementary differentiating circuit.

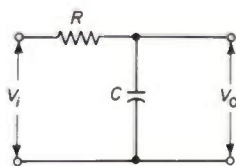


Figure 11-40. Elementary integrator circuit.

11-18. Integrating circuit

An elementary integrating circuit consisting of a resistor R and capacitor C is shown in Fig. 11-40. Let i be the current through the circuit due to applied voltage V_i .

If the RC product is large, that is, if both R and C are large, the reactance of C will be small and can be neglected in comparison to R . The current is

essentially determined by the resistance and has a value

$$i = \frac{V_i}{R}$$

The output voltage V_o is the voltage across the capacitance and is

$$V_o = \frac{1}{C} \int i \, dt$$

so that

$$V_o = \frac{1}{RC} \int V_i \, dt$$

The output voltage is proportional to the integral of the input voltage, hence the name of the circuit.

The larger the RC product, the better the approximation made for i , the better the integration, and unfortunately, the smaller the output voltage.

In practice, acceptable integration is obtained for

$$RC \geq 4t_1$$

where t_1 is the pulse duration. The expression holds good for ohms, farads, and seconds, or just as well for megohms, $\mu\mu\text{f}$, and μsec .

For vertical pulses, the integrator time constant should then be on the order of 760 microseconds or more.

11-19. Other differentiators

The RC circuit is not the only differentiator. An inductive RL circuit may produce differentiation also. When the input signal is a current i , it is sufficient to have it flowing through an inductance L for the voltage V_L across the inductance to be

$$V_L = L \frac{di}{dt}$$

that is, proportional to the derivative of the input signal. The corresponding circuit is shown in Fig. 11-41. In all circuits using an inductance, the time constant is L/R .

The inductance can be replaced by a mutual inductance M as shown in Fig. 11-42 with identical results. This circuit may be used with blocking oscillators, where the transformer is already provided.

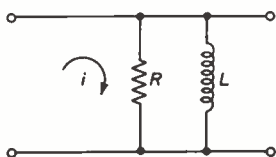


Figure 11-41. Inductance current differentiator.

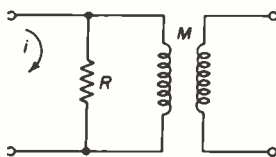


Figure 11-42. Mutual inductance current differentiator.

When the input signal is a voltage, the inductance differentiator takes the form shown in Fig. 11-43.

Except where inductance is already provided in the circuit for other reasons, capacitance differentiators are much preferred. They are more economical and any inductance has parasitic resistance and capacitance which can upset the differentiation. The usual differentiator for a voltage signal is the circuit of Fig. 11-39.

Several differentiators can be cascaded. Since the attenuation of the signal will be important, a large input signal is necessary.

The effect of a differentiator is a strong distortion of the signal. Theoretically, a perfect differentiator that is fed a parabolic signal would provide a linearly increasing output or ramp signal. If fed a ramp signal, it would provide a constant output.

If the input signal were a rectangular pulse, the differentiator would provide a pair of opposite sharp pulses, corresponding to the front and back side of the rectangular pulse. Also, a sawtooth wave applied to a differentiator would come out as a rectangular wave.

It is frequently said that a differentiator selectively passes fast variations of a signal with respect to slow ones, or that the shorter the duration of a signal, the better it is transmitted by a differentiator.

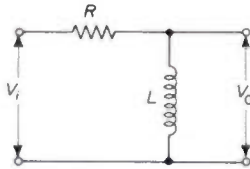


Figure 11-43. Inductance voltage differentiator.

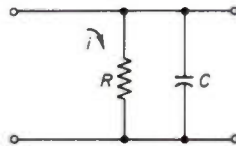


Figure 11-44. Capacitance current integrator.

11-20. Other integrators

The voltage-driven RC integrator of Fig. 11-40 has an equivalent when the input signal is a current as shown in Fig. 11-44. The operation of the circuit is due to the fact that the voltage across a capacitance is proportional to the integral of the current i flowing through it.

An inductor L can also be used as an integrator as shown in Fig. 11-45. This circuit is sometimes used in connection with blocking oscillators where the inductance L is already provided in the blocking transformer. Because of the parasitic resistance and capacitance unavoidably associated with inductance, and because of economic considerations, the RC integrator is by far the most common circuit.

Frequently several integrators are cascaded. Because of the high total attenuation, a large input signal is generally required.

Generally speaking, an integrator is a circuit which (similar to a differentiator) distorts the signal markedly. If it is fed a constant voltage, it provides ideally a linear continuously increasing output or ramp voltage. If fed a ramp signal, it transforms it into a parabolic signal. Also, if fed a rectangular wave, its output will be a sawtooth wave. It is worth noting in passing that in this case a practical integrator will provide a sawtooth which is not quite linear, but has exponential (instead of linear) rising and falling slopes. The larger the integrator time constant, the smaller this exponential distortion. It can be seen, therefore, that the integrator has an action opposite to that of the differentiator.

Summing up, it may be said that an integrator selectively passes slow variations of the signal with respect to fast variations or, again, that longer pulses are better transmitted than short pulses.

This last property and the opposite property of the differentiator are used in sorting vertical and horizontal sync pulses.

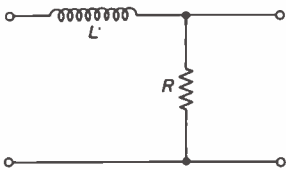


Figure 11-45. Inductance integrator.

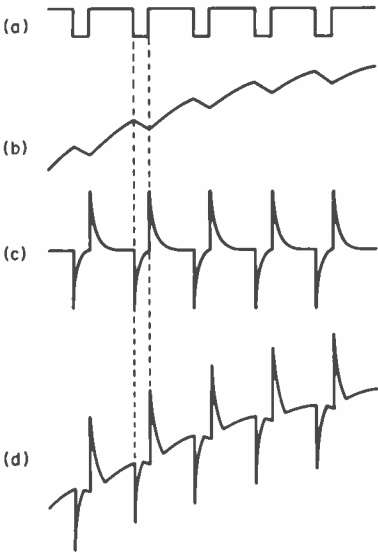


Figure 11-46. Combined integration and differentiation.

11-21. Combined circuits

Figure 11-46 shows at *a* a series of rectangular pulses representative of a recurrent sync signal. The result of integrating this series of pulses is shown at *b*, whereas the effect of differentiation of the waveshape *a* is illustrated at *c*.

A sharp triggering of the sweep oscillators, that is, a precise synchronization, is easier to obtain with fast-rising sharp pulses similar to those obtained at C through differentiation.

On the other hand, an integrator facilitates a discrimination based on *amplitude* because of the rising output signal it produces. Integrator and differentiator action can be combined to obtain both advantages at the same time. The basis of the scheme is that the voltage across the resistance of an RC integrator differs very little from the input signal and can be used to drive a differentiator.

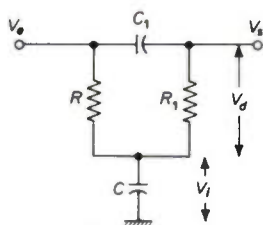


Figure 11-47. Voltage driven integrator-differentiator circuit.

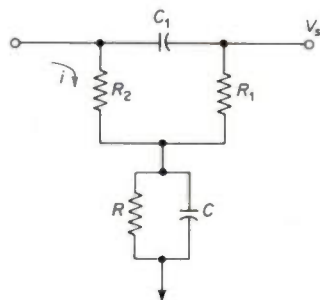


Figure 11-48. Current driven integrator-differentiator circuit.

A circuit to accomplish this purpose is shown in Fig. 11-47. The input voltage V_e is applied to the integrator RC and the integrated voltage V_i appears across C . The voltage across R is applied to a differentiator C_1R_1 and the differentiated voltage V_d appears across R_1 .

Since R_1 is connected to the top electrode of C , the addition of V_i and V_d across R_1C provides the total output signal V_s , which is the sum of the differentiated and integrated signals and has the form illustrated in Fig. 11-46(d). The component values are chosen so that there is a minimum disturbance of the integrator by the differentiator, and vice versa.

A similar circuit may be devised when the input signal is a current instead of a voltage as shown in Fig. 11-48. RC is the integrator. An additional resistance R_2 , which is small enough not to disturb the integrator, has been inserted in series, and the voltage appearing across it, proportional to the signal input current, is used to drive the differentiator C_1R_1 . As before, the low end of R_1 returns to the top electrode of C , so that the output voltage, taken across R_1C , is the sum of the differentiated and integrated signals.

In this circuit integrator and differentiator are better separated than in the preceding arrangement, and consequently element values are more independent.

11-22. Sync pulse sorting

A simple circuit to sort vertical and horizontal sync pulses and to apply them to the corresponding sweep oscillator is shown in Fig. 11-49. It simply consists of a differentiator C_2R_2 for line sync and an integrator R_1C_1 for field sync. Representative values would be

$$C_2 = 100\ \mu\mu\text{f}, \qquad R_2 = 10\ \text{kilohms}$$
$$C_1 = 10,000\ \mu\mu\text{f}, \qquad R_1 = 100\ \text{kilohms}$$

Since the circuit is placed between the anode of the separator and the grids of the horizontal and vertical sweep oscillators, care has to be taken of the d-c

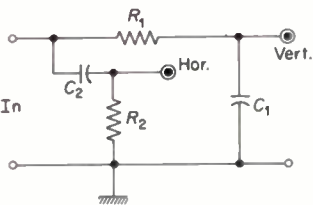


Figure 11-49. Sorting horizontal and vertical sync pulses.

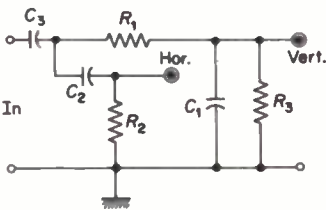


Figure 11-50. Practical sorting circuit.

voltage levels. In actual practice, the circuit would appear as shown in Fig. 11-50, where C_3 is a coupling capacitor to isolate the separator plate voltage and may have a value of $0.1\ \mu\text{f}$. The other added component is R_3 , which is simply a grid return for the vertical oscillator.

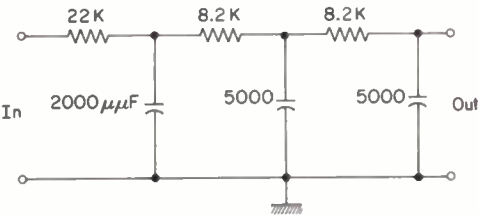


Figure 11-51. Triple integrator.

In practice, a single integrator is rarely used. The separated sync pulses have an amplitude of from 10 to 40 volts and only a few volts (or even a fraction of a volt) are sufficient to synchronize the vertical oscillator. Several integrators may then be cascaded with better results, as far as vertical sync sorting is concerned, and the output signal may still be large enough to ensure good synchronization. Figure 11-51 shows such a typical triple integrator

circuit. Notice that the time constants chosen are rather short for each integrator and ensure only partial integration. Cascading three integrators, however, provides satisfactory results.

11-23. AFC systems

Direct synchronization of the horizontal oscillator by the sync pulses is rarely used because the system is extremely sensitive to noise pulses, which appear as short spikes riding on the v-f modulation signal and reaching into the sync signal region, as shown in Fig. 11-52. These noise pulses pass through the sync separator and trigger the oscillator before the legitimate sync pulse does. The usual result is horizontal tearing of the picture.

An entirely different system of synchronization is based on the fact that all oscillators have a frequency that depends more or less on the tube voltages and especially on the grid voltage. If the sync section of the receiver provides a suitable d-c voltage to control the horizontal oscillator and keep its frequency at the correct value, noise problems are eliminated. The system is called *AFC—Automatic Frequency Control*. A variation of the scheme provides a d-c voltage to control a reactance tube, shunt-connected across the frequency-determining circuit of the oscillator.

The AFC control voltage should be zero when the oscillator is working at the correct frequency, that is, at the repetition frequency of the horizontal sync pulses. It should become positive or negative according to whether the oscillator is working on a frequency lower or higher than the sync pulses' recurrence frequency.

A good way to determine whether the horizontal oscillator is in step with the horizontal pulses is to compare their frequencies or phases. The circuit which performs this function is called a *phase comparator*.

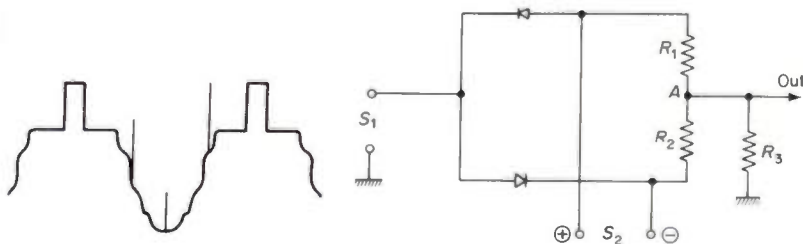


Figure 11-52. Noise superposed on the v-f signal.

Figure 11-53. Double diode phase comparator.

11-24. Phase comparator

Numerous phase comparator circuits can be designed. Figure 11-53 illustrates the basic circuit of a double diode comparator. One of the signals, S_1 , is

applied simultaneously to the two diodes, connected back to back. The other signal S_2 is applied symmetrically to the other side of the diodes. Equal load resistors R_1 and R_2 are used. Assume S_1 is zero. The equal and opposite voltages S_{2+} and S_{2-} due to S_2 create equal and opposite currents through R_1 and R_2 and point A is at zero potential.

Assume now that S_1 has some value and has the same frequency as S_2 . On one alternation S_1 will add to S_{2+} . On the other alternation S_1 will add to S_{2-} . The total result will be symmetrical again and the voltage at A will still be zero.

Assume now that S_1 and S_2 have different frequencies. The voltages applied to the two diodes, each equal to $S_1 + S_2$, will not have the same value because of the changing phase relation between signals at different frequencies. One of the diodes will pass a larger current than the other, the two currents will not cancel at A , and the *difference* current will flow through R_3 creating across it a voltage which is a function of the phase difference between S_1 and S_2 .

This basic circuit easily lends itself to horizontal synchronization when S_1 is the line oscillator voltage or a signal derived from it, and S_2 is the horizontal sync pulse voltage.

In the same way, the oscillator signal could be applied symmetrically as S_2 and the sync signal nonsymmetrically as S_1 .

11-25. Phase inverter

A separate winding could be provided on the line output transformer, and a grounded center tap would ensure symmetrical voltages for the phase comparator.

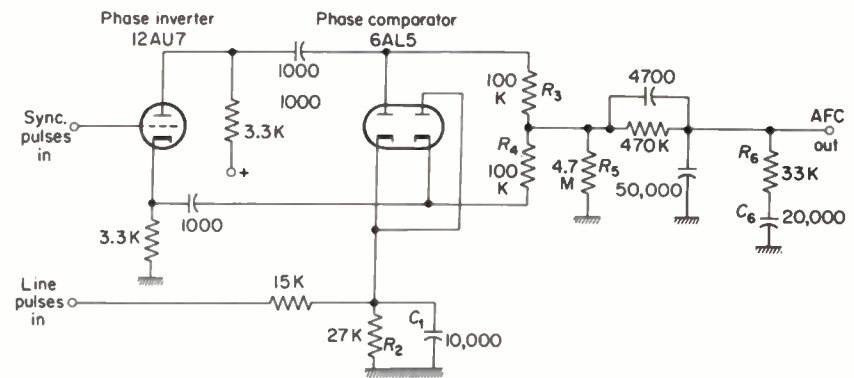


Figure 11-54. Practical circuit of a phase comparator.

More commonly, a phase inverter is used on the sync pulse to apply it symmetrically to the phase comparator, and the horizontal line pulse is taken somewhere in the line time base. Figure 11-54 shows a commonly used

practical circuit. A triode is used as a phase inverter. Its grid is directly connected to the plate of the sync separator and it carries equal loads in the cathode and anode circuits. The symmetrical signals across these loads are applied to the comparator, which uses a double diode.

The line pulse comes from the line output transformer. A simple shaping circuit $R_1-R_2-C_1$ transforms it into a sawtooth wave for phase comparison. Equal load resistors R_3 and R_4 are used for the diodes and the AFC voltage appears across R_5 . It is filtered before being used for frequency control.

Auxiliary circuit R_6-C_6 is called the *anti-hunt* circuit. Its function is to eliminate an eventual instability of the entire AFC system which would send it into low-frequency oscillation.

11-26. Triode comparator

Another fairly common comparator circuit is shown in Fig. 11-55. The basis of the operation of the triode comparator is that the grid current of a triode

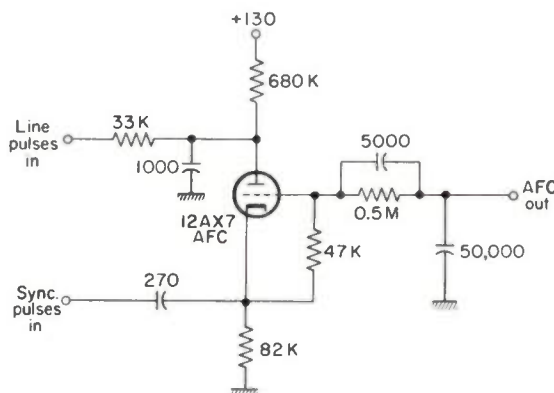


Figure 11-55. Triode comparator.

depends on the grid *and* plate voltages. Consequently, if these voltages happen to be periodic, grid current depends on the phase relationship between them. In the circuit shown, sync pulses are applied to the cathode, thus supplying the grid-cathode signal. Line pulses are applied to the anode through an integrator circuit which transforms them into a sawtooth wave.

Grid current flowing through the grid resistance provides the AFC voltage, which is filtered for frequency control use.

A variation of the triode comparator appears in Fig. 11-56. Sync pulses provided by a phase inverter are applied symmetrically between grid and cathode. Line pulses are integrated into a sawtooth wave applied to the anode. Plate current flows only during that fraction of the time when the sync pulses

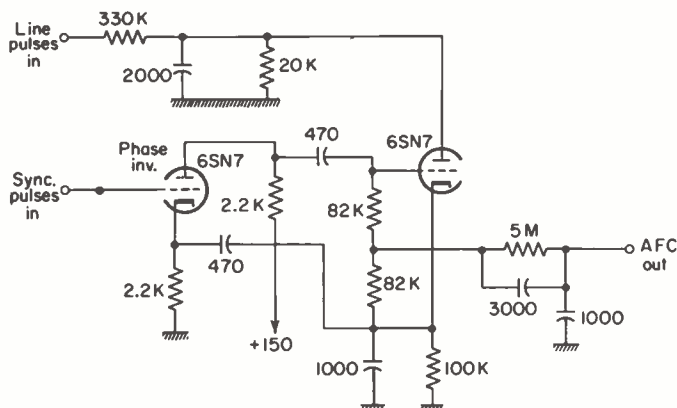


Figure 11-56. Symmetrically driven triode comparator.

drive the grid positive and the cathode negative. The short burst of plate current, moreover, has a value which depends on the instantaneous value of the sawtooth wave on the anode. Therefore, plate current is a function of the relative phase of sawtooth and sync pulses, that is, of line pulses and sync pulses.

The voltage drop the plate current produces across the cathode resistor is used as the AFC voltage. Actually, the AFC voltage is picked off as shown halfway between grid and cathode. This eliminates a strong component at sync pulse frequency.

Problems

1. Show that it is necessary for the d-c component to be present if correct modulation levels are to be respected.

2. In view of clipping the sync pulses, is it better to align the tops of the sync pulses or the black level?

3. Show graphically what happens in a diode restorer circuit when the time constant RC is equal to the period of recurrence of the pulses.

4. It is suggested that d-c restoration be produced at the grid itself of the picture tube. Discuss the feasibility of the idea.

5. Illustrate graphically how strong noise pulses affect the d-c restorer circuit, and show how simple additional circuits can greatly reduce the effects of noise.

6. A diode has an internal resistance of 1200 ohms and a parasitic capacitance of $1.4 \mu\text{f}$. It is to be used as simple sync separator. Study the performance of the possible circuits, assuming sensible values.

7. Show that a grid-detection separator works satisfactorily whether or not the d-c component is present in the signal it receives.

8. Choose a pentode tube which you think suitable for separation work and obtain its characteristics in a tube handbook. Design a separator circuit so that double limitation is obtained. Calculate the output voltage.

9. Using a tube handbook, find the characteristics of a triode to be used as sync separator. Design the circuit. Calculate the amplitude of the output pulses. Evaluate the minimum value of input voltage necessary for good separation.

10. Two sync separators are used in the same receiver, one for horizontal pulses and one for vertical pulses. Calculate the bandwidth necessary for each. Compare designs.

11. An additional stage of amplification is used in the sync section between v-f separator pick-off and sync separator. Design the stage, using a tube of your choice.

12. Show graphically the relation between input and output signals for a differentiator when the input signal is (a) a sine wave and (b) a square wave.

13. Show graphically the relation between input and output signals for an integrator when the input signal is (a) a sine wave and (b) a square wave.

14. Design an integrator-differentiator circuit using inductances.

15. A triple integrator has for each section a time constant equal to 5 times the vertical pulse duration. The input pulse has an amplitude of 100 v. Calculate the amplitude of the output pulse.

16. A differentiator circuit for horizontal pulses has a time constant of 0.4 microsecond. The input pulses have an amplitude of 40 v. Evaluate the amplitude and shape of the output signal.

17. A television system using 405 lines and 50 fields per second has horizontal pulses of 10 microseconds duration and vertical pulses of 400 microseconds. Design a sorting circuit for this system.

18. Explain how noise pulses can produce tearing of the picture.

19. A shaping circuit is used to modify the form of the pulses coming from the horizontal sweep circuit before applying them to the phase comparator. Justify its use.

20. Explain the role of the filter used in the output of an AFC comparator.

21. In a double diode comparator, the sync pulses are assumed to be applied symmetrically to the diodes. Study what happens if symmetry is destroyed, including the case when one diode receives no sync pulses.

22. Compare a double-diode and a single-triode phase comparator.

23. It is suggested that some form of comparator could use a tube having two control grids, e.g., hexode, with one grid receiving the line pulses and the other the sync pulses. If the tube is cut off except when both grids receive pulses, the plate current is a function of the overlap of the two pulses, that is, of their phase relation. Investigate this suggestion.

Chapter 12

SOUND RECEPTION

The sound accompanying a picture is transmitted on a separate frequency-modulated carrier. The sound carrier goes through the frequency changer and results in a sound i-f, different from the picture i-f, which can be amplified separately. This is the basis of split i-f sound receivers.

The more common method, however, is to allow the sound i-f to go through the picture i-f amplifier. The picture detector will then produce a mixing of picture and sound i-f carriers. The resulting beat frequency is called the *inter-carrier* frequency. It contains the useful sound frequency modulation and the undesired picture amplitude modulation. This last modulation is removed by limiting, so that an intercarrier sound receiver is similar to an FM receiver.

12-1. FM receiver

The sound signal uses frequency modulation with a maximum deviation of ± 25 kc and a pre-emphasis of 75 microseconds. The sound carrier is 4.5 mc above the picture carrier.

A standard FM receiver incorporates the following stages: r-f amplification, frequency changing, i-f amplification, FM detection, and audio amplification.

The first two sections, r-f amplification and frequency changing, are usually with the picture receiver so that in television the sound receiver circuitry comprises only i-f amplification, detection, and a-f amplification.

Auxiliary functions provided in the sound circuitry are amplitude limitation and de-emphasis. Amplitude limitation in some form is always used to remove noise and other unwanted amplitude modulation. It also provides the FM detector with a constant amplitude signal so that there is no need for

automatic gain control. Amplitude limitation frequently takes place in the last i-f stage before detection or in the detector itself.

De-emphasis is required to attenuate the higher audio frequencies which were purposely overamplified as at the transmitter.

12-2. Split i-f reception

Two different systems are used for the sound receiver: the split i-f and the intercarrier.

In the split i-f system, the sound i-f is taken off the picture receiver, generally after the frequency changer stage or after the first i-f stage, and usually across a sound trap circuit. The sound i-f thus obtained is 4.5 mc below the picture i-f and is amplified by two or three sound i-f stages before detection.

The disadvantage of this system lies in the narrow bandpass of the sound i-f amplifier, which is 50 kc as a minimum and in practice closer to 100 kc. If the local oscillator varies in frequency by 0.1 mc or 100 kc, which is not very much considering its frequency of operation, the resulting sound i-f shifts by the same amount and falls outside the sound i-f amplifier passband: the sound disappears or is strongly attenuated.

For this and other reasons, split i-f sound reception is rarely used nowadays in comparison to intercarrier reception. Also, except for sound take-off and i-f frequency, a split i-f sound receiver differs very little from an intercarrier receiver.

The intercarrier system will therefore be analyzed in detail with the understanding that most of the analysis applies just as well to a split i-f receiver.

12-3. Intercarrier reception

Intercarrier reception is based on the fact that sound and picture carriers are separated by a constant 4.5 mc. Consequently, the picture carrier can be used to beat against the sound carrier and provide a resulting sound i-f of 4.5 mc. It is true that this signal will contain picture amplitude modulation as well as sound frequency modulation, but an efficient limiter will remove all amplitude modulation and let only the sound signal pass through.

The advantage of the system is that the resulting 4.5-mc beat frequency is independent of the local oscillator frequency or instability. Moreover, it leads to a more economical design than the split carrier system.

The beat frequency will be produced only if sound and picture i-f carriers are applied to a nonlinear stage. Such a stage already exists in the picture receiver; it is the video detector.

Therefore, no extra stage is needed in this respect. The only requirement is that the response curve of the picture receiver be such that enough sound i-f signal is available at the video detector. This is easily ensured by suitable tuning of the picture i-f circuits.

12-4. Sound pick-off

The sound i-f at 4.5 mc may then be picked off at any point in the picture receiver after video detection.

Sometimes a point between detector and v-f amplifier is used, but more frequently the sound is picked off at the output of the v-f amplifier, just before the picture tube.

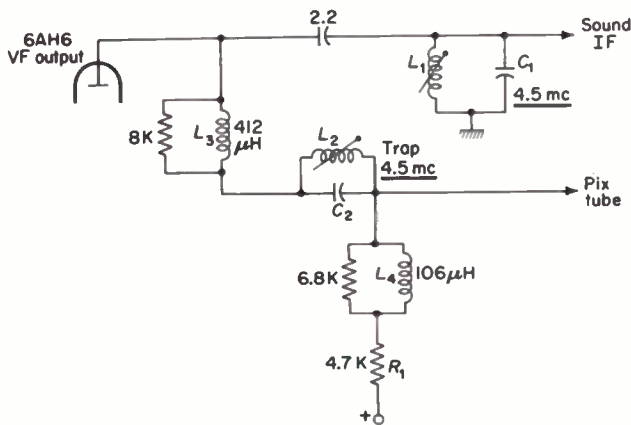


Figure 12-1. Capacitive coupling to the sound i-f.

A variety of circuits can be used, employing either capacitive coupling or some form of inductive coupling.

Figure 12-1 shows one form of capacitive coupling through a small capacitance linking the v-f output and the first tuned circuit L_1C_1 of the sound i-f.

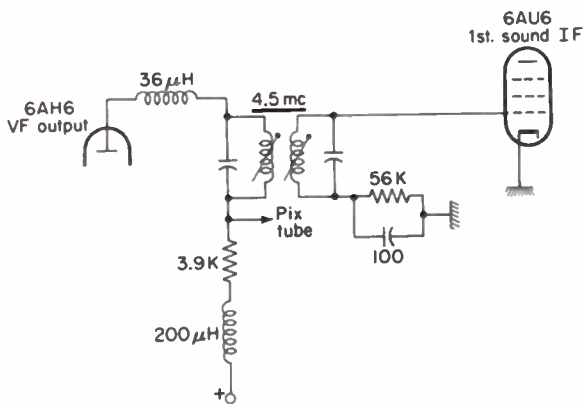


Figure 12-2. Double tuned transformer sound pick-off.

The anode circuit of the v-f tube contains the usual compensation inductances L_3 and L_4 and the load resistance R_1 , as well as a circuit L_2C_2 tuned on 4.5 mc.

The beat at 4.5 mc, if allowed to reach the screen of the tube, would be apparent on the picture as a parasitic modulation. It is reduced to unobservable proportions by a circuit tuned to 4.5 mc, which is the trap circuit L_2C_2 .

Figure 12-2 illustrates another sound pick-off circuit which uses a transformer. In this case, no extra trap circuit is necessary in the connection to the picture tube.

Other coupling circuits are used for pick-off, but they are all based on the same principles.

12-5. I-f amplifier

The sound i-f amplifier is tuned on 4.5 mc and generally contains one or two tube stages, the last stage frequently being used as a limiter at the same time.

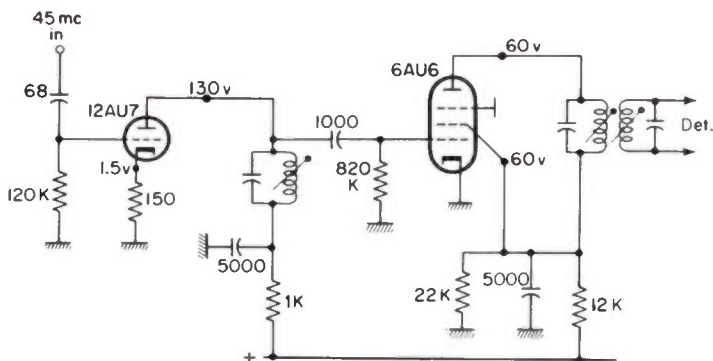


Figure 12-3. I-f amplifier and i-f limiter stages.

Figure 12-3 shows a two-stage sound i-f amplifier using a triode as the first stage and a pentode as the output stage. The second stage is also used as a double limiter. It has no cathode bias and a large grid resistance so that grid detection occurs and limits the positive excursion of input signal. In addition, plate and screen are fed low voltages and another limitation occurs on the negative excursion of the input voltage. The mode of operation is analogous to that of a double clipping separator stage.

A one-stage sound i-f amplifier is shown in Fig. 12-4. It uses the same double limitation principle as before, employing grid detection and low screen voltage. Still another one-stage sound i-f amplifier is shown in Fig. 12-5. It uses only grid detection, limiting only the positive peaks of the input signal and thus affording some noise improvement.

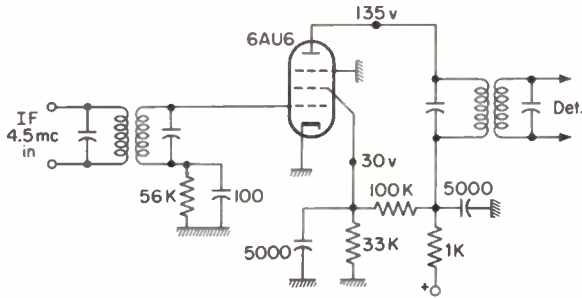


Figure 12-4. I-f limiter stage.

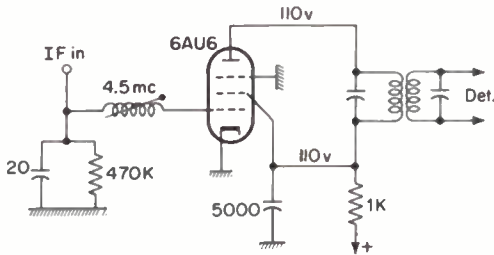


Figure 12-5. I-f amplifier-limiter stage.

12-6. FM detection

Several FM detectors, or discriminators, have been devised. The simple circuit shown in Fig. 12-6 resembles an amplitude detector. The difference is that the

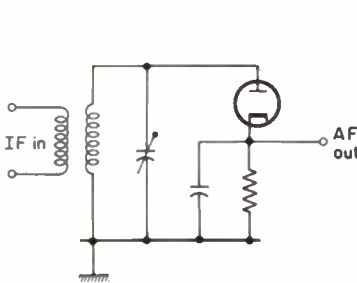


Figure 12-6. FM slope detection circuit.

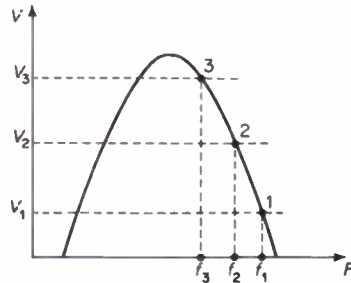


Figure 12-7. Operation of FM slope detection.

discriminator is off-tuned deliberately so that the carrier frequency falls on the side of the response curve. The farther away the received frequency is displaced from the resonant frequency, the smaller the output signal. This is shown in Fig. 12-7, where frequency f_1 corresponds to amplitude V_1 , frequency

f_2 to amplitude V_2 , frequency f_3 to amplitude V_3 , etc. It is apparent that if the received carrier is set at f_2 , frequency variations will be transformed into amplitude variations.

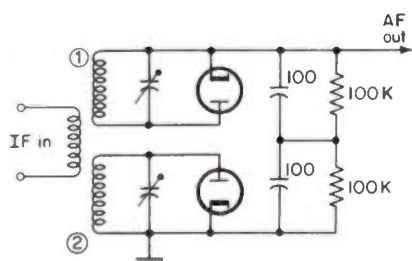


Figure 12-8. Symmetrical discriminator.

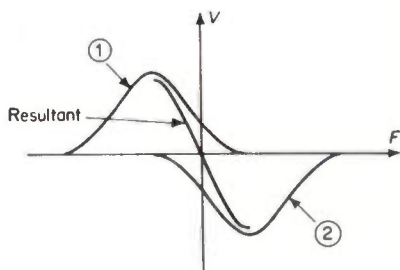


Figure 12-9. Response curve of the discriminator.

The system is called *slope detection*. It has several disadvantages, the most evident being that the detection is not linear. This can be improved by using two identical circuits, each with resonant frequencies above and below the

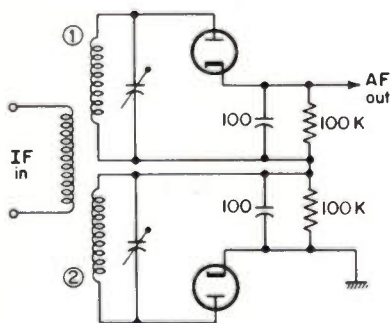


Figure 12-10. Travis discriminator.

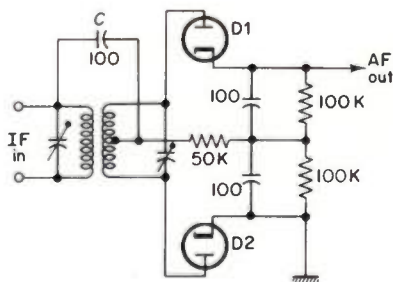


Figure 12-11. Foster-Seeley discriminator.

received frequency respectively, as shown in Fig. 12-8. A symmetrical arrangement is employed, as shown in Fig. 12-9, so that response curves 1 and 2 of the two circuits combine to give an almost linear total S-curve.

A variation, called the *Travis discriminator*, is shown in Fig. 12-10.

12-7. Foster-Seeley detector

The diagram shown in Fig. 12-11 represents the Foster-Seeley FM discriminator or detector. Both primary and secondary of the input transformer are tuned on the sound frequency.

Each diode receives equal and opposite signals from the center-tapped secondary. The primary signal is 90 degrees out of phase with the secondary signal and is applied in series to the two diodes through a capacitor connecting primary to secondary center tap. Each diode then receives equal signals and the output voltage is zero.

However, the symmetrical phase relations above hold true only at the exact resonant frequency. If the received frequency is above resonant frequency, then symmetry is destroyed, the angle of phase increases, and D_1 receives a larger signal than D_2 .

If the received frequency is below resonant frequency, the angle of phase decreases and D_2 receives a larger signal than D_1 .

The voltages detected by the two diodes are opposed to each other, so that the resulting difference signal is a function of the frequency of the input signal and FM detection has been accomplished. A variation of the Foster-Seeley diagram is shown in Fig. 12-12. The additional primary voltage in the secondary is obtained by means of an untuned tertiary winding L .

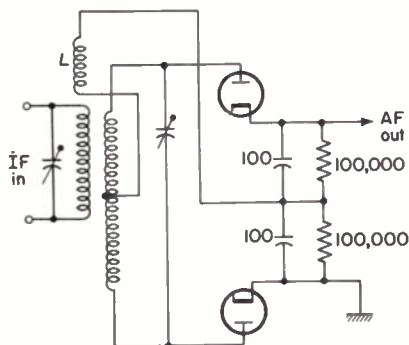


Figure 12-12. Inductive link between primary and secondary center tap.

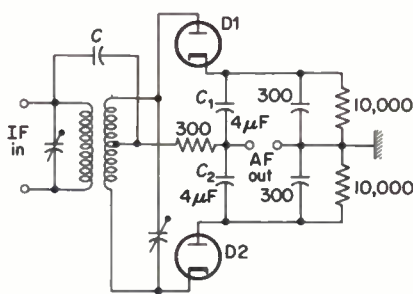


Figure 12-13. Ratio detector.

12-8. Ratio detector

The most common FM detector is probably the ratio detector of Fig. 12-13. It is derived from the preceding two-diode circuits by the simple reversal of a diode. A number of variations have been designed. In all cases, however, the two diodes are connected in opposite polarities and the a-f output signal is the difference between the two detected voltages.

Since C_1 and C_2 are large-value capacitors, they behave practically as a short circuit at audio frequencies. The ratio detector is insensitive, therefore, to amplitude modulation and hence has a very useful immunity to noise pulses. It can use a double secondary as shown in Fig. 12-14 or, equally as well, an inductive coupling between primary and secondary, as shown in the

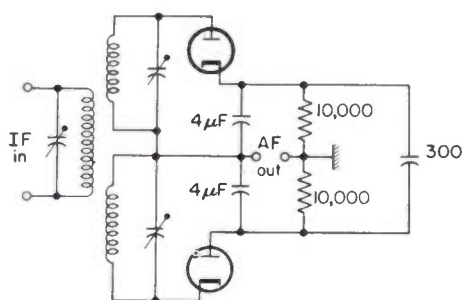


Figure 12-14. Variant of the ratio detector.

commonly used circuit of Fig. 12-15. Notice that a single large-value capacitor C_1 is used across the diode loads and that capacitor C_2 in shunt across the output attenuates the high frequencies for de-emphasis.

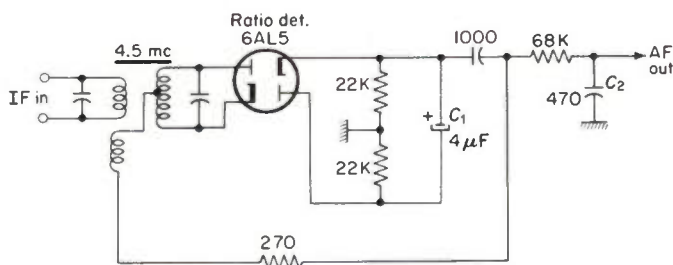


Figure 12-15. Common diagram of ratio detector.

12-9. Quadrature detector

A different type of FM detector is represented by special tubes of the hexode variety, developed expressly for the purpose of FM detection.

These tubes have two control grids, the second one called the *quadrature grid*. It receives a signal 90 degrees out of phase with the signal applied to the first control grid. Plate current flows only when the two control grids are simultaneously positive. If the phase difference between the two grids is smaller or larger than 90 degrees, plate current will flow for a longer or shorter time, thus producing a detected output signal.

The nominal 90-degree phase difference at carrier frequency may be obtained by using a double-tuned transformer, as done for the double diode detectors. A simpler way of achieving the same result is shown in Fig. 12-16. A resonant circuit LC , tuned on 4.5 mc, has been inserted between quadrature grid and ground. It plays the same role as the secondary of a double-tuned transformer, the coupling in this case being electronic and due to the effect of

space charge on the quadrature grid inside the tube. Notice the shunt capacitance across the plate load resistance to attenuate high audio frequencies for de-emphasis. This circuit is also called a *gated-beam discriminator*.

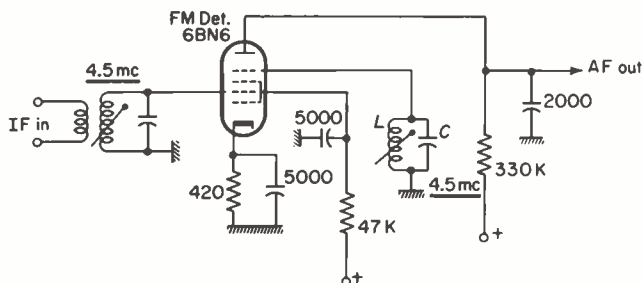


Figure 12-16. Quadrature grid tube FM detector.

12-10. De-emphasis

In comparison to the low audio frequencies, the high audio frequencies are overamplified at the transmitter and this must be compensated for by suitable high audio-frequency attenuation in the receiver. A de-emphasis circuit is used for this purpose, which might take the form shown in Fig. 12-17. The load of an a-f tube contains not only the customary resistance R , but also a shunt capacitance C , chosen of such value that $RC = 75 \mu\text{sec}$. High audio frequencies are thus progressively attenuated with the same time constant as when they were overamplified, and the over-all result is a more faithful a-f reproduction.

The idea behind the system, of course, is to improve the signal-to-noise ratio, a result especially desirable at higher audio frequencies.

In actual practice, a simple version of the de-emphasis circuit is commonly used. It consists of a capacitance of suitable value connected across the output from the FM detector.

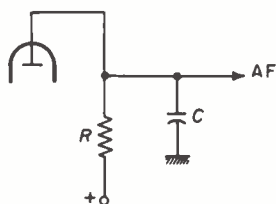


Figure 12-17. De-emphasis RC circuit.

12-11. A-f amplification

Little need be said about a-f amplification. It is exactly similar to the corresponding section of a radio receiver and the same circuits are used, eventually with the same refinements or improvements such as tone control, push-pull output, etc.

Very briefly and generally, two tubes are employed; the first is a voltage amplifier, the second a power output amplifier.

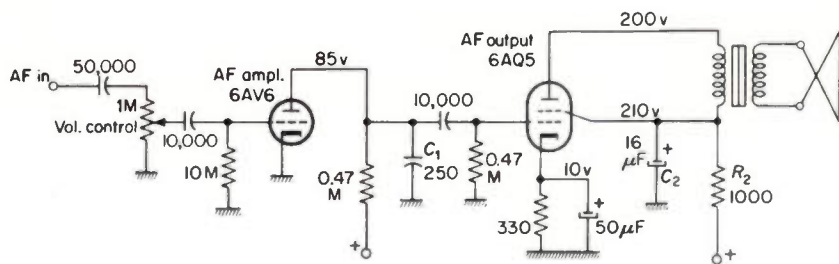


Figure 12-18. Typical AF section of a sound receiver.

A typically representative practical diagram is shown in Fig. 12-18. The voltage amplifier uses a triode, and grid bias is obtained by a high-value resistance in the grid circuit, the cathode being grounded. This arrangement is fairly common in actual practice. The plate circuit contains a capacitor C_1 across the load resistor to reduce amplification at the high audio frequencies.

The output amplifier uses a beam power tube in a conventional circuit. Notice the decoupling arrangement R_2C_2 in the plate and screen voltage supply. This is customarily included to eliminate any interaction with the other sections of the television receiver due to current variations in the sound output stage.

Problems

1. Explain the usefulness of pre-emphasis at the transmitter.
2. The standard FM radio broadcast maximum deviation is 75 kc. Why wasn't this value adopted for the television sound?
3. With illustrative diagrams, explain the role of amplitude limitation.
4. Draw a diagram of a pentode i-f amplifier providing double limitation. Show graphically how the tube operates, using a characteristic curve.
5. The picture transmitter of a station has a breakdown and goes off the air. The sound transmitter is still working and is used to make an announcement saying that the picture transmitter will be back on the air in twenty minutes. Compare the usefulness of this announcement for a split i-f receiver and an intercarrier receiver.
6. Intercarrier sound uses the picture carrier. However, if video modulation reaches 100 per cent, the picture carrier disappears. What happens in this case? What remedy would you suggest?
7. The intercarrier beat can be taken off the output of the detector or after v-f amplification. Compare the two possibilities.
8. A double-tuned transformer is used as coupling between two stages of 4.5-mc sound i-f amplification. Making reasonable assumptions, design this transformer.
9. What are the disadvantages of slope detection?

10. Using phase relationships between the various voltages, show graphically how the Foster-Seeley discriminator operates.
11. Using phase diagrams, show graphically how the ratio detector operates. Also, give an explanation of its insensitivity to amplitude modulation.
12. What happens in a ratio detector if the large-value capacitance across the diode loads opens up?
13. Using phase diagrams, show graphically how the quadrature grid detector operates. How does it behave in the presence of amplitude modulation and strong noise pulses?
14. It is proposed to use pre-emphasis and de-emphasis in an amplitude-modulated radio system. Evaluate the proposal.
15. Assuming sensible numerical values, show that the pre-emphasis and de-emphasis system materially improves television sound reception.
16. Design a two-tube sound audio amplifier containing negative feedback and tone control.
17. Design a high-quality sound audio amplifier using a push-pull output stage.
18. One of the reasons put forward to justify the use of FM sound in television was that FM radio broadcasting stations could be received at the small cost of adding the necessary frequency band to the television tuner. Discuss the merits of this proposal in view of the present state of the art.
19. Evaluate the total gain of the sound section of a typical intercarrier receiver.
20. A large-size receiver has a screen 50 cm wide. Evaluate the shape and size of the visible pattern which would be produced if the intercarrier 4.5-mc frequency were allowed to reach the modulating electrode.
21. When a television receiver is misadjusted, so-called *sound bars* appear on the screen. These are grey or black bars moving up and down at the rhythm of the sound modulation. Explain how this might happen, especially in view of the fact that the sound is frequency-modulated.
22. A typical defect of sound reception is a characteristic buzz superimposed over the desired sound modulation. How does this happen?
23. What would be the effect of mistuning the sound i-f transformer driving the FM detector?

Chapter 13

PICTURE TUBE

The picture tube presents on its screen the over-all result of the complex operations taking place in the receiver, that is, the received picture.

To a large extent, the type of picture tube used influences the mechanical and electrical design of the receiver. Public taste dictates the use of larger screens and shallower cabinets, which in turn have led to the necessity for wider deflection angles in picture tubes, pending some new development leading to a flat type of picture display.

For the time being, electromagnetic deflection is the only practical method of scanning wide-angle picture tubes. Focusing may be electromagnetic or electrostatic.

Auxiliary functions such as ion trapping and adjustable centering are also associated with picture tubes.

13-1. Picture tube

The picture tube receives from various sections of the television receiver all of the supply voltages necessary for its proper operation. The modulating v-f signal is provided by the v-f amplifier. Scanning is accomplished with the help of two sweep generators, vertical and horizontal. Auxiliary circuits are added for focusing, centering, brilliance control, blanking, and other secondary functions.

The picture tube presents on its screen the ultimate result of the operation of the entire receiver and is an important element. It determines to a large extent the design of the circuits and also the mechanical design of the receiver.

Almost without exception, currently used picture tubes employ electromagnetic deflection. Electrostatic deflection will be briefly described,

however, for the sake of completeness and also because it is still used in monitors, electronic viewfinders, and oscilloscopes.

A picture tube has the shape shown in Fig. 13-1. It is a glass envelope, evacuated to a high vacuum, with a cylindrical section containing the electrodes and a conical section whose dimensions depend on the size of the tube. The bottom E of the conical part represents the screen of the tube. It carries on its internal face a layer of a special material, called a *phosphor*, which fluoresces and emits visible light upon impact by electrons. At the other end

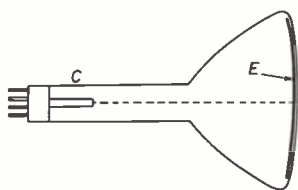


Figure 13-1. Basic picture tube.

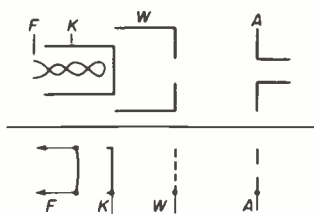


Figure 13-2. Triode type electron gun.

of the tube, in the cylindrical part, an electron gun C projects an electron beam on the phosphor layer, or *screen*, which becomes luminous at the point of impact.

The *electron gun*, as shown in Fig. 13-2, consists of a cathode source of electrons K heated by a heater F , a control electrode or grid or Wehnelt cylinder W , and a first or accelerating anode A . The electrons emitted by the cathode are attracted to the anode, which is at a high positive voltage. The electrons are thus accelerated and shot as a narrow beam through the hole in the center of the anode. Such a structure as represented is called a *triode gun* and its usual symbolic representation is given in Fig. 13-2.

Grid W has a role similar to the role of the grid in an ordinary vacuum tube. It controls the intensity of the electron beam.

The luminous area, or *spot*, produced on the screen by such a simple system would be rather large because the beam has a large cross section and moreover is divergent. To obtain a small and luminous spot, it becomes necessary to concentrate or focus the electron beam. Focusing is obtained with electromagnetic or electrostatic fields.

13-2. Electrostatic focusing

To produce electrostatic focusing, a fourth electrode is added to the electron gun, as shown in Fig. 13-3. It is called the *second anode*. The usual representation is indicated.

The cathode being taken as the zero reference voltage, the grid W is negative by, say, 30 v; the first anode A_1 is positive by, say 300 v, and the second anode A_2 is strongly positive, say 1000 v. The geometry of the anode electrodes is such that the electrostatic field between them has the distribution shown in Fig. 13-4. The equipotential lines have been drawn to show the analogy with an optical lens. When the diverging electron beam is shot through this field, it undergoes focusing just as a light beam is focused by an optical lens. Because of this similarity, the electron focusing device is known as an *electron lens*.

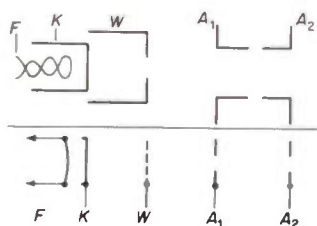


Figure 13-3. Tetrode type electron gun.

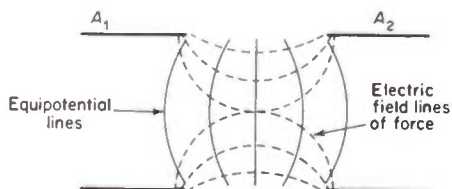


Figure 13-4. Electronic focus lens.

The electron beam coming out of the lens is convergent, so that its cross section decreases, goes through a minimum, and then increases again. By modifying the voltage on the first anode, the convergence of the lens can be adjusted and the minimum cross-section of the beam can be made to coincide with the screen. In this way, a small spot is obtained whose diameter may be a fraction of a millimeter.

13-3. Electromagnetic focusing

Focusing can also be obtained by means of a magnetic field parallel to the axis of the tube. In this case, electrons moving parallel to the axis do not undergo any modification due to the field. However, electrons diverging from the axis move along a direction at an angle with the magnetic field and are acted upon by a force simultaneously perpendicular to their direction of motion and to the field. This can be shown by considering an electron of charge e and mass m , accelerated by a voltage E to a velocity v .

The acceleration energy is eE . The kinetic energy acquired by the electron is $\frac{1}{2}mv^2$. By the law of conservation of energy,

$$eE = \frac{1}{2}mv^2$$

hence

$$v = \left(\frac{2eE}{m} \right)^{1/2}$$

Substituting numerical values for e and m :

$$e = 16 \times 10^{-20} \text{ coulomb}$$

$$m = 9 \times 10^{-28} \text{ gram}$$

and using convenient units,

$$v = 595E^{1/2}$$

where v is in km/sec and E is in volts.

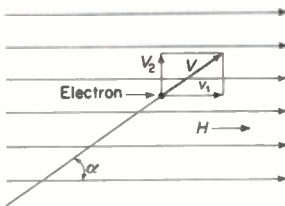


Figure 13-5. Effect of a magnetic field on an electron.

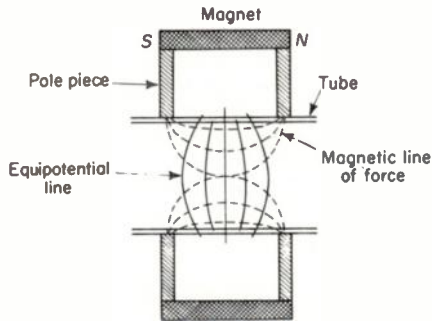


Figure 13-6. Magnetic field focusing.

When the electron enters a magnetic field H at an angle α , as shown in Fig. 13-5, its velocity vector v may be resolved into two perpendicular components. The first is parallel to the field and is the axial velocity v_1 . The second is perpendicular to the field and is the radial velocity v_2 . Since

$$v_2 = v \sin \alpha$$

the force acting on the electron is

$$F = Hev_2$$

or

$$F = Hev \sin \alpha$$

and is perpendicular to both H and v . When v is parallel to H , angle α is zero, F is zero, and no force acts on the electron, which moves undisturbed through the field.

When the electron motion is at an angle α with the magnetic field, it is submitted to the force F which modifies its trajectory in such a way that the electron has a tendency to wrap itself around the magnetic lines of force and travel in a circular path. Since the electron has been subjected to the electrostatic force of the anode driving it forward through the tube, the resultant path of the electron will be helical as it moves from cathode to the screen.

In practical cases, the situation is somewhat more complicated because the magnetic field is not homogeneous. Figure 13-6 might help the reader to visualize what happens. An annular magnet with two pole pieces has been

placed around the neck of the tube, and the field it produces has the distribution indicated. If equipotential lines are drawn, they can be seen to outline a lens shape similar to the one found for an electrostatic field and reminiscent of an optical lens. The effect is similar; the electron beam is made to converge and has a very small cross-section at the point of convergence. Adjusting the magnetic field intensity allows the point of convergence to coincide with the screen surface.

13-4. Focusing control

In current practice, an annular permanent magnet or an electromagnet *C* is slipped over the neck of the tube as shown in Fig. 13-7, a little behind the anode. In the case of a permanent magnet, focusing may be adjusted with

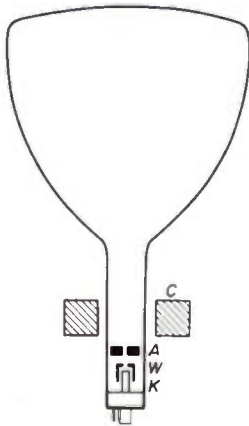


Figure 13-7. Focus electromagnet or magnet *C*.

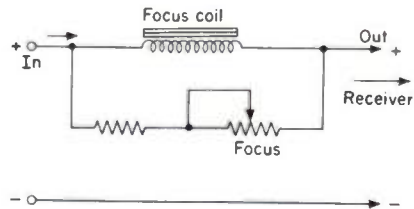


Figure 13-8. Focusing control for a series type coil.

variable magnetic shunts or perhaps with an auxiliary electromagnet. Also, two ring magnets may be used and their spacing varied.

An electromagnet is commonly called a *focus* or *focusing coil*. Its field is easily adjusted by controlling the current flowing through the coil. Roughly, a few hundred ampere-turns are necessary.

Two types of focus coils are used:

1. The low-resistance type has a small number of turns, say 3000, and uses a rather large current, say 0.2 amp. It then provides 600 ampere-turns. This type is frequently called a *series coil* because of its usual connection, as shown in Fig. 13-8. The coil is inserted in series with the high-voltage supply, like a filter inductance which it can replace. Current is adjusted by a variable shunt resistance.

2. The high-resistance type of focus coil has a large number of turns, say 30,000, and uses a small current, say 20 ma, again providing 600 ampere-turns. Such a coil is called a *shunt coil* because of its common connection across the high-voltage supply. A variable series resistance adjusts the current as shown in Fig. 13-9.

To ensure a more uniform field and better efficiency, the focus coil is generally encased in a shell of magnetic material, and an annular internal gap shown in Fig. 13-10 produces and localizes the field inside the tube.

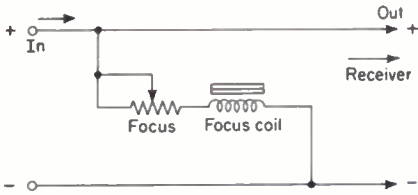


Figure 13-9. Focusing control for a shunt coil.

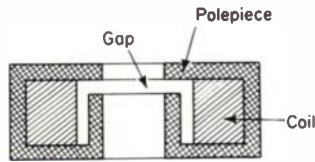


Figure 13-10. Shielded focus coil.

In the expression giving the force acting on the electron we find the velocity, which varies with the accelerating voltage. Therefore, the magnetic field intensity needed for focusing increases with the accelerating voltage.

Now that a small spot has been obtained on the screen, it is necessary to sweep it over the screen area, and this also can be done by electrostatic or electromagnetic fields.

13-5. Electrostatic deflection

Electrostatic deflection of the electronic beam is obtained by adding two electrodes, the deflection plates P_1 and P_2 , shown in Fig. 13-11, on each side of the beam after it emerges from the final accelerating anode A_2 . The average potential of P_1 and P_2 is equal to the potential of A_2 , so that the deflection plates do not modify the velocity of the electrons.

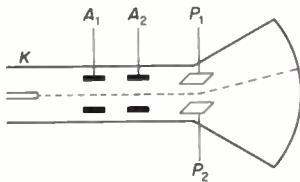


Figure 13-11. Electrostatic deflection plates.

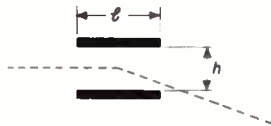


Figure 13-12. Effect of an electrostatic field on an electron.

When P_1 and P_2 are at the same potential, they attract the electron beam equally and their effects balance out. When P_1 is more positive than P_2 , there is a net attraction force toward P_1 and the electron beam is deflected upward. Similarly, the beam is deflected downward if P_2 happens to be the more positive plate.

This qualitative reasoning may be replaced by a more precise quantitative calculation, which will illustrate how electron ballistics operate.

Let l and h be the length and spacing of the deflection plates (Fig. 13-13), e and m the electron charge and mass, E the acceleration or high voltage, and U the deflecting voltage or voltage between the deflection plates.

The axial velocity of the electrons has already been given as $v_1 = (2eE/m)^{1/2}$ and is not modified by the deflection plates.

When electrons enter the field created by the deflecting plates, they have no transverse velocity.

Assume a deflecting voltage U exists between the two deflection plates. The corresponding voltage gradient F is

$$F = \frac{U}{h}$$

Now, an electron starting with zero velocity, which is the case for transverse velocity, and acted upon by a voltage gradient F , reaches a velocity

$$v_2 = \frac{F t_1 e}{m}$$

where t_1 is the time during which the voltage gradient acts upon the electron.

It is easy to evaluate t_1 . Since the axial velocity v_1 of the electron is not modified, and since the deflection plates have a length l , the electron will go through the plates in time

$$t_1 = \frac{l}{v_1}$$

Therefore, it will acquire a transverse velocity

$$v_2 = \frac{Uel}{mhv_1}$$

When it emerges from the deflecting plates, the electron has an axial velocity v_1 , a transverse velocity v_2 , and will move according to the resultant γ of these two velocities.

Figure 13-13 shows a method of evaluating deflection sensitivity. Let L be the distance from the center of the deflection plates to the screen and d be the deflection of the spot on the screen. There are two right triangles in which

$$\frac{d}{L} = \frac{v_2}{v_1}$$

Substituting for v_2 ,

$$d = \frac{UelL}{m\hbar v_1^2}$$

Substituting for v_1 ,

$$d = \frac{LU}{2\hbar E}$$

which is the fundamental equation giving the apparent deflection d as a function of the applied voltages and of the tube parameters. The units need only be homogeneous—for example, U and E in volts, and h , l , L , and d in centimeters.

For a given tube, l , L , and h are fixed. Let $k = lL/2h$. Then

$$d = k \frac{U}{E}$$

so that the amplitude of deflection is linearly proportional to the deflecting voltage and inversely proportional to the accelerating voltage.

The deflection sensitivity of a tube may be defined in a number of ways. The easiest definition is to state that deflection sensitivity, for a given accelerating voltage, is the deflection in millimeters observed on the screen when a deflection voltage of one volt is applied to the deflection plates.

Deflection sensitivity will increase for

1. larger l —that is, longer deflection plates;
2. smaller h —that is, smaller spacing between deflection plates;
3. larger L —that is, greater distance between deflection plates and screen.

In addition, it can be seen that deflection will increase when the accelerating voltage is reduced. Unfortunately, brilliance decreases simultaneously.

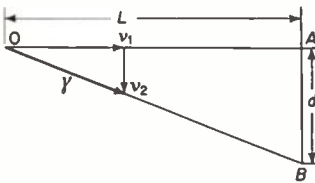


Figure 13-13. Evaluation of deflection sensitivity.

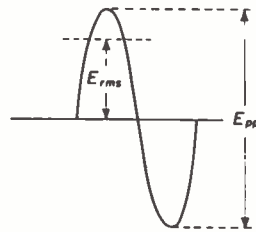


Figure 13-14. Peak to peak and rms values of a sine voltage.

13-6. Deflection sensitivity

Since tubes may be used with various accelerating voltages, a somewhat more precise way of defining deflection sensitivity is to state that the tube has a

sensitivity of, say, $600/E$ mm/v, where E is the accelerating voltage, or, just as well, 0.6 mm/v/kv. This means that for an accelerating voltage of 1000 volts, or 1 kv, the deflection reaches 0.6 mm per applied deflection volt. Incidentally, deflection volts so stated are always d-c volts.

The sensitivity for the usual picture tube is on the order of a fraction of a millimeter per volt. Its value is given in the tube data and can be used to evaluate the voltage necessary for a given deflection. For example, a tube having a sensitivity of 0.4 mm/v will give a deflection of 100 mm for

$$100 : 0.4 = 250 \text{ volts d-c}$$

If the applied deflection voltage is an a-c sine wave, the above result applies to the peak-to-peak E_{pp} value, from which the rms value E_{rms} can be obtained as shown in Fig. 13.14 through

$$E_{rms} = 0.35E_{pp}$$

In the above case,

$$E_{rms} = 250 \times 0.35 = 87.5 \text{ v}$$

The deflection voltage necessary to cover, say, one screen diameter does not depend so much on the tube dimensions as on the accelerating voltage. The real criterion is not the screen size, but the maximum deflection angle. For example, a 3-in. tube and a similar 4-in. tube will have the same electrode arrangement but the 4-in. tube will be longer than the 3-in. tube. For the same accelerating voltage, the same deflection voltage will be necessary to cover one diameter in both cases. This is because the 4-in. tube has a larger diameter on the one hand, but it also has a larger distance between deflection plates and screen with resulting improved deflection sensitivity.

However, the same accelerating voltage will give less luminosity or brightness on the larger screen. This may well compel the use of a higher accelerating voltage, with resulting loss of deflection sensitivity.

In other words, scaling up a small tube may require larger deflection voltages, not because of the larger size but because of the necessity to maintain sufficient brightness on the screen.

13-7. Electrostatic tube

One pair of deflecting plates moves the beam in two opposite directions only. To sweep the entire screen surface, a second pair, perpendicular to the first, is necessary. It is placed beyond the first pair and will consequently provide reduced sensitivity.

A complete electrostatic cathode-ray tube, as shown in Fig. 13-15, will contain a heated cathode K , a grid W , two or three anodes A_1, A_2, A_3 , a pair of horizontal deflection plates P_1 , and a pair of vertical deflection plates P_2 .

The deflection plates modify the electron trajectory by acting only on the transverse velocity; they should not modify the axial velocity. To accomplish this, it is necessary that the average potential of the deflection plates be equal

to the electron beam potential at the location of the plates. This has a bearing on the nature of the deflection circuit, as will be seen in the following paragraph.

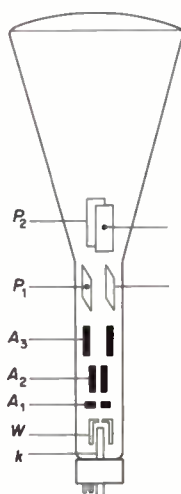


Figure 13-15. Electrostatic picture tube with tetrode gun.

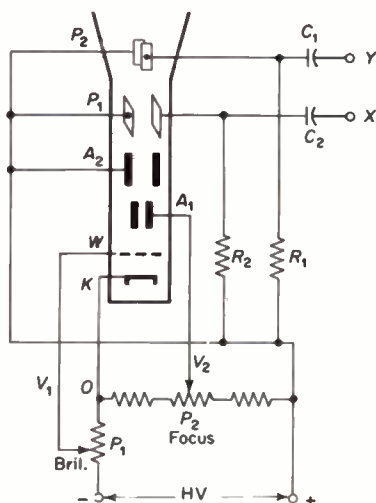


Figure 13-16. Supply voltages for an electrostatic tube.

13-8. Symmetrical and nonsymmetrical deflection

The simplest deflection system is illustrated in Fig. 13-16, along with the tube electrode connections. The cathode K potential is taken as reference zero; the grid W is set at a small negative voltage V_1 , say -20 v, by the action of potentiometer P_1 which is the brightness or brilliance control. The first anode A_1 is at a positive voltage of, say, 200 v, adjustable through P_2 which is the focus control. The second and final accelerating anode A_2 is connected to the accelerating voltage, say 1000 v, and so is one deflection plate of each pair. The other two plates receive the deflection signals through capacitances C_1 and C_2 and return to the accelerating voltage through large-value resistances R_1 and R_2 . The deflection plates do not take any current and there is no d-c voltage drop across R_1 or R_2 .

This simple system is used only for small screen sizes because it introduces trapezoidal distortion. This occurs because one plate of the first pair is at constant voltage, that is, $B+$, whereas the other plate is at variable voltage, $B+$ plus deflection signal, hence nonsymmetrical deflection. Consequently, the average voltage between the two plates varies above and below $B+$ and accelerates or decelerates the electrons, modifying their axial velocity. The deflection sensitivity of the second pair of plates varies accordingly, producing

on the screen a raster which is smaller on one side than on the other, as shown in Fig. 13-17.

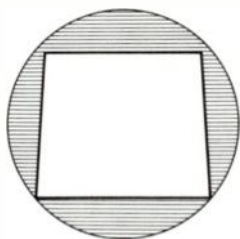


Figure 13-17. Trapezoidal distortion.

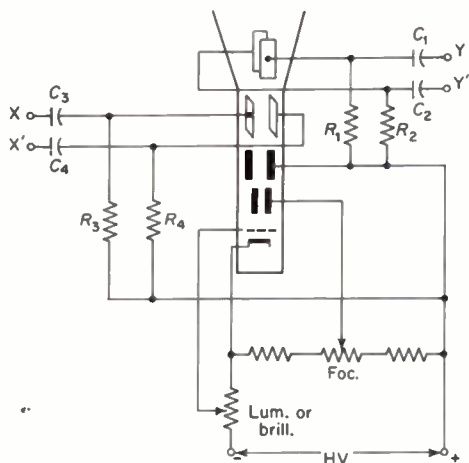


Figure 13-18. Symmetrical deflection.

Special construction of the deflection plates can in some measure correct for this defect, but all large-screen tubes use symmetrical deflection instead. In this deflection technique, as shown in Fig. 13-18, the four deflection plates are connected to the accelerating voltage through large resistances R_1 , R_2 , R_3 , and R_4 and each pair is driven push-pull fashion by the deflection signal through capacitances C_1 , C_2 , C_3 , and C_4 .

In this way, *both* plates of the same pair vary symmetrically in voltage above and below the accelerating voltage. The average potential is constant and equal to the accelerating voltage and there is no trapezoidal distortion.

13-9. Centering

In the absence of deflecting voltages, theoretically the spot should be at the center of the screen. This is not true in practice because of manufacturing tolerances. For this reason, and also for convenience in use, centering controls are provided. The customary centering circuit is shown in Fig. 13-19 applied to a symmetrical tube. Auxiliary tube control circuits have not been represented for clarity. Observe the difference in comparison to the preceding diagram in that one deflection plate of each pair returns to point A . The other plate is connected to the slider of a potentiometer shunted across points B and C .

Now B is more positive and C less positive than A . By adjusting the slider, the plate connected to it may be made more or less positive than the opposite plate and a permanent displacement of the picture will be produced.

One potentiometer may control vertical centering, and the other horizontal centering in this manner.

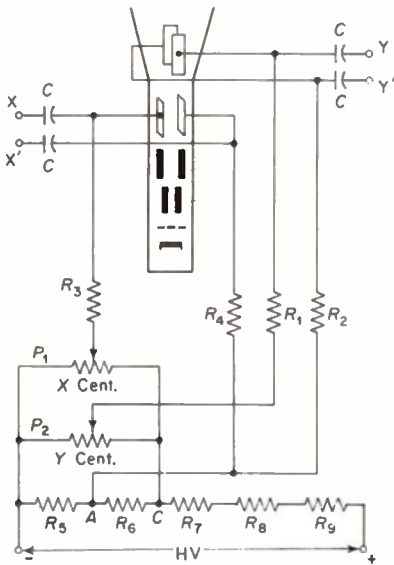


Figure 13-19. Centering adjustments.

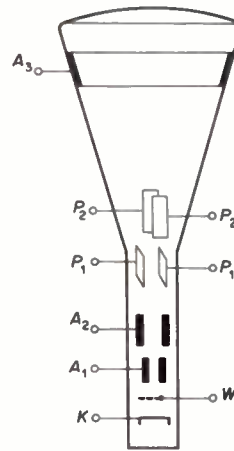


Figure 13-20. Post-acceleration electrode A_3 .

13-10. Post-acceleration tube

As stated previously, deflection sensitivity increases when the acceleration or high voltage decreases, but the brightness decreases at the same time.

It would appear that electrons should be slow for deflection and fast for brightness. These two seemingly contradictory requirements may be satisfied by accelerating the electrons *after* they have been deflected. The principle employed is called *post-acceleration* and is shown in Fig. 13-20 applied to a cathode-ray tube in the form of an annular extra (post-accelerating) anode A_3 near the screen. The high voltage applied to A_2 is relatively small and the electron beam is easily deflected by the deflecting plates, ensuring good sensitivity. After deflection, the beam is accelerated again by the extra anode A_3 connected to a voltage several times higher than the voltage applied to A_2 . The resulting brilliance is high.

The principle of post-acceleration is simple, but its practical realization is not too easily achieved. The post-acceleration anode is commonly made of a ring or spiral of colloidal graphite, or aquadag, deposited on the inside surface of the tube.

Even when there is no post-acceleration, the inside of the tube is generally coated with aquadag, from the electron gun up to a short distance from the

screen. This coating is connected to the final anode or may even constitute the final anode, especially in magnetic deflection tubes.

The aquadag accomplishes several useful purposes. It prevents light coming from the rear to backlight the screen. It reduces the amount of light originating in the screen itself and reflected back on the screen by the tube's glass envelope. It collects wandering electrons, especially secondary electrons produced when the electron beam hits the screen, and returns them to the positive terminal of the supply.

13-11. Magnetic deflection

We have already seen that for magnetic focusing an electron traveling with velocity v and entering a magnetic field H at a right angle to the velocity is submitted to a force F simultaneously perpendicular to H and v and given by

$$F = Hev$$

The force F modifies the direction of electron trajectory, but not the velocity. The electron moves in a circular path of radius R such that the centripetal force F due to the field balances the centrifugal force due to curvature of the trajectory.

For a better idea of the process, assume an electron of charge e and mass m moving at velocity v and acted upon by a magnetic field H . At any point on the trajectory, the centrifugal force mv^2/R and the centripetal force Hev are equal, R being the radius of curvature of the trajectory at this point. Hence

$$\frac{mv^2}{R} = Hev$$

or solving for R ,

$$R = \frac{mv}{He}$$

In constant magnetic field H , and since m , v , and e are constant, this shows that the radius of curvature of the trajectory is constant. In other words, the trajectory is an arc of circle.

13-12. Deflection sensitivity

Referring to Fig. 13-21, a is the length occupied by magnetic field H , R is the trajectory radius, and α is the resulting final angle of deflection. At point A , the electron exits from the magnetic field and moves along the tangent to the curved trajectory, hitting the screen at a distance x from the center. The deflection is x , and D is the distance from the screen to the center of deflection.

The two angles α , shown in Fig. 13-21, are equal because their sides are perpendicular geometrically as radius and tangent. Hence

$$\sin \alpha = \frac{a}{R}$$

Now $x = D \sin \alpha = Da/R$, and using the value previously found for R ,

$$x = \frac{DaeH}{mv}$$

Velocity v is a function of the accelerating voltage E , given by $v = (2eE/m)^{1/2}$, as already seen, and substituting for v gives

$$x = \frac{HaD}{(2mE/e)^{1/2}}$$

which is the fundamental equation for magnetic deflection. For a given tube, D is constant, and since e and m are constant, the expression can be written

$$x = k \frac{aH}{E^{1/2}}$$

where k is a constant. This equation shows that deflection is directly proportional to field intensity and to the length of the field. Also, it is inversely proportional to the square root of the high voltage, and in this respect magnetic deflection scores an important advantage over electrostatic deflection. With large-size tubes, sufficient brilliance is obtained only by a large increase in high voltage. If the high voltage were quadrupled, electrostatic deflection sensitivity would decrease to one quarter of its original value, whereas magnetic deflection sensitivity would decrease only to one half.

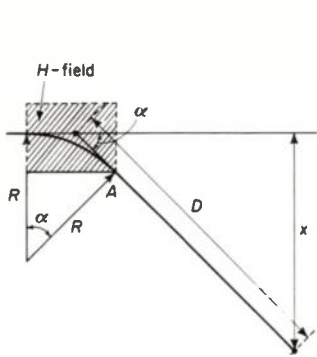


Figure 13-21. Electromagnetic deflection sensitivity calculation.

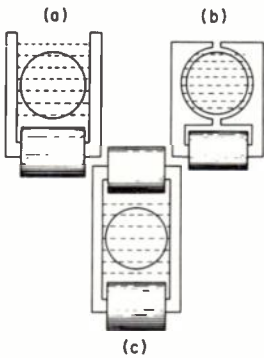


Figure 13-22. Various designs for vertical deflection coils.

13-13. Deflection coils

The magnetic deflection field is obtained with the help of external coils slipped over the neck of the tube. Practical designs differ. In most cases, a magnetic core of high-permeability material is used to increase efficiency. This core could be designed, as shown in Fig. 13-22, to channel into the tube a field

produced by either one or two coils. However, the most common type of coil takes a shape such as is shown in Fig. 13-23, designed to fit tightly around the neck of the tube. The only active part of such a coil is the straight part parallel to the axis of the tube. Seen end on, these coils produce a magnetic field, as



Figure 13-23. Coaxial pair of coils.

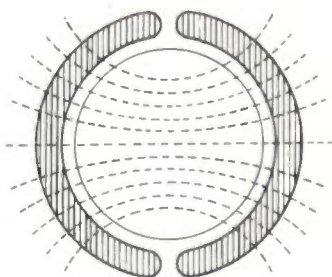


Figure 13-24. Magnetic deflection field inside the coils.

shown in Fig. 13-24, which is not quite linear. However, only the central part of the field is used for deflection. A pair of coils, as shown in Fig. 13-23, is capable of producing deflection in two opposite directions only, and a second

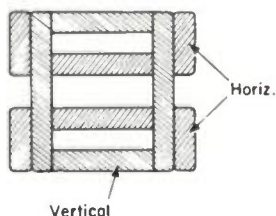


Figure 13-25. Side view of a coaxial deflection yoke.

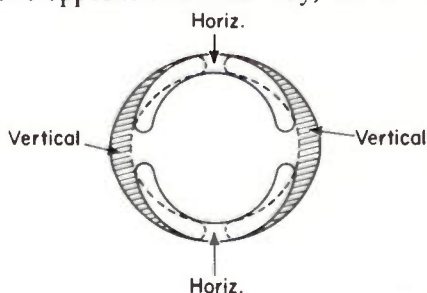


Figure 13-26. End view of a coaxial deflection yoke.

pair is necessary for complete scanning of the face of the tube. The second set is made shorter than the first and fits inside it as shown in Fig. 13-25. The longest pair of coils is used for line or horizontal deflection. The shortest pair is used for field or vertical deflection.

Figure 13-26 represents the same set of coils viewed endwise. It shows how the two pairs are combined 90 degrees apart to produce horizontal and vertical scanning. Such an assembly is called a *deflection yoke* and the coils are usually surrounded by a ring of high-permeability material to concentrate the field inside the tube and increase efficiency. Also, the coils frequently flare outward and fit snugly against the conical part of the tube.

13-14. Which deflection?

Seen from the cathode-ray tube manufacturer's point of view, it is much easier to make a tube for electromagnetic deflection than for electrostatic deflection.

The reverse might seem to be true for the consumer. It is relatively easy to obtain voltage sawtooth waves and feed them directly to deflection plates. However, when the screen size increases and the high voltage must also be increased to preserve sufficient brightness, large-amplitude sawtooth voltages have to be produced with good linearity, and this sets voltage supply problems for the sweep generators. Moreover, in view of field distortions between deflecting plates, deflection angles much larger than 10 degrees are a rarity. A direct consequence is that a large-screen tube becomes rather cumbersome in length.

The limit of practicability for electrostatic deflection type tubes seems to be about a 7-in. screen diameter. Under 7 inches, electrostatic deflection is extensively used. Above, electromagnetic deflection is virtually unchallenged.

Electromagnetic deflection requires a current sawtooth through the coils, and the sweep generators therefore can use a normal plate supply voltage. Increasing the high voltage does not reduce deflection efficiency too rapidly and brightness can be high. The magnetic field for deflection is homogeneous over a fairly large volume, so that large deflection angles can be tolerated, up to 130 degrees or more in present tubes. By this method, a large-size screen can be obtained for a relatively small depth.

Since public taste prefers screens of 17 inches or more, it is easy to understand why all present television receivers use electromagnetic deflection.

13-15. Fluorescent screen

A large number of known compounds emit light when they are hit by an electron beam. Television tubes use mixtures of complex compounds, called *phosphors*, based on various chemical salts such as sulphates of zinc or cadmium. A good phosphor should have high light efficiency, a very nearly white color, and a persistence, remanence, or memory compatible with television use, that is, on the order of the picture period.

The layer of phosphor inside the tube should not be too thick, because the picture produced on its back surface must be observed through the thickness of the layer. At the same time, it cannot be too thin, because electrons could shoot through it and reach the glass surface, where their remaining energy would be dissipated as heat and lost for the light-producing process.

Commercial screens at present do not produce a pure white light, but rather a bluish-white color. Persistency is kept below $\frac{1}{60}$ second. Luminous efficiency, that is, the amount of light produced per watt of electric energy

furnished to the tube, reaches 10 per cent, which is as good as that of most incandescent lamps.

Progress in the direction of improved efficiency is highly desirable because it provides increased light output without modification of the electrical or mechanical characteristics of the tube. Active research is going on in the field of phosphors, and theoretical considerations show that an increase of several times the present light efficiency is not an impossibility.

13-16. Flat screen tube

The curvature of the screen of a picture tube produces a geometrical distortion as shown in Fig. 13-27. The picture appears cramped at the periphery of the curved screen because of the angle under which it is seen. The spherical shape

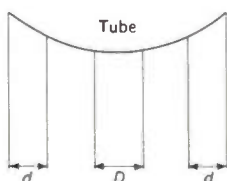


Figure 13-27. Apparent geometrical distortion.

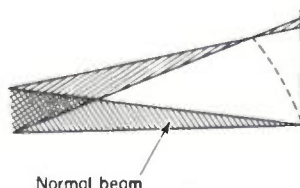


Figure 13-28. Defocusing for an ordinary electron beam.

of the screen had been adopted for two reasons. First, the optimum focus surface is a sphere having for its center the deflection center. Second, the screen convexity greatly increases a tube's mechanical performance. Since the tube is evacuated, atmospheric pressure is applied to its outside and especially to the screen, which is structurally the weakest part. The face of a 21-in. tube must resist a total atmospheric pressure of some 1500 kg and a curved surface can resist pressure to a greater extent.

However, since a plane picture is superior from the viewer's side, glass techniques have been developed to produce slightly convex or cylindrical screens that are almost flat except for a small curved part joining the screen to the cone. The main difficulty that faced the manufacturer was to obtain this result without an undue increase in glass thickness, which would have resulted in increased weight, price, and light loss.

The structural glass problems solved, there still remains the focusing problem when the tube face is flattened. The solution adopted is to produce a narrow and slowly converging electron beam. With a normal beam, as shown in Fig. 13-28, focusing is optimum only near the center of the flat screen, and deteriorates rapidly outward where the beam hits the screen after having gone through its minimum section. With a narrow beam of low divergence, as shown in Fig. 13-29, the spot on the screen is still small enough at the

periphery of the picture, and the lack of sharpness in focusing is not apparent at normal viewing distances.

Commercially, however, there is a limit to the size of tube which can be economically manufactured using present techniques. It seems to be on the order of 30 inches and this may be explained in part by the resulting size of the cabinet in the viewer's home. Apparently, there is no pressure by the public for larger screens and little effort has gone into their development.

Nevertheless, a large-size picture is sometimes desirable, and one way to surmount the difficulties associated with mammoth tubes is to use projection-type tubes.



Figure 13-29. Defocusing for a thin beam.

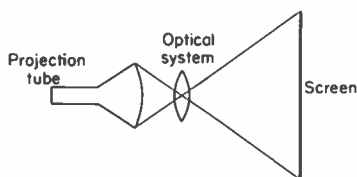


Figure 13-30. Simple optical projection system.

13-17. Projection tubes

The first projection idea to come to mind, if one is faced with enlarging the tube image, is shown in Fig. 13-30. An optical lens projects an enlarged image of the tube face on a large screen. The drawback lies in the poor efficiency of the lens, which is on the order of 5 per cent. To obtain on the screen a sufficiently bright picture, the brilliance of the tube should be increased by a minimum factor of 20. Special tubes, nevertheless, have been made for the purpose of projection by this method. These use very high voltages, between 25,000 and 100,000 volts, and the phosphor screen is designed accordingly. They generally have a small diameter—5 in. is typical—and produce a high-brightness, high-contrast picture, which may undergo enlargement up to several feet or even several yards.

If space is limited, as it would be if the whole system were installed in a single cabinet, the optical path can be folded to reduce bulk. Figure 13-31 shows the use of a mirror and a vertically set projection tube. The screen can be ground glass or some plastic material. It receives the picture on its back face, and viewers in front see it through the screen material.

A liquid magnifying lens has also been used, as shown in Fig. 13-32. This is made of a transparent plastic casing filled with a suitable refractive liquid such as certain oils. It may be large in size and can be placed directly on the tube face with a resulting apparent increase in tube size.

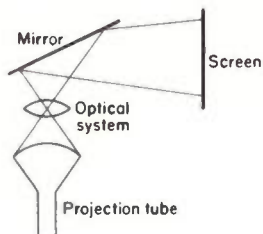


Figure 13-31. Folding the optical path with a plane mirror.

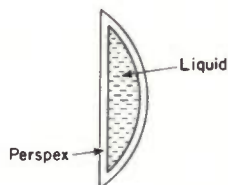


Figure 13-32. Liquid filled lens.

13-18. Schmidt optical system

The poor efficiency of standard lenses, however, has led to the use of better optical systems. A successful type, originally developed for astronomy, is the Schmidt arrangement, one form of which is illustrated in Fig. 13-33. Essentially, the optical components are a spherical concave mirror and a special lens

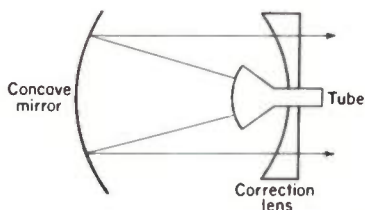


Figure 13-33. Schmidt type of optical system.

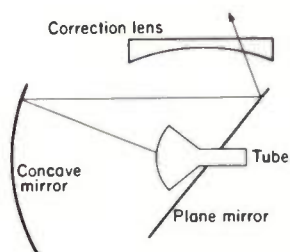


Figure 13-34. Schmidt optical system and plane mirror.

to correct distortions. Since the tube protrudes through the correcting lens, one might wonder why its shadow does not appear on the screen. However, the picture on the tube face is not projected point by point, but each point is the origin of an infinity of light rays reflected by the spherical mirror and focused at one point on the screen. When some of these rays are suppressed by an obstruction, such as the tube, the result is a slight decrease in over-all brightness without a localized shadow.

The optical path of the rays may be folded to reduce the dimensions of the system, as shown in Fig. 13-34. A plane mirror is used and the tube is located in an opening in the center of this mirror. The four elements—tube and associated components, spherical mirror, plane mirror, and correcting lens—are assembled into a small-sized single unit in one commercial version of the idea.

The increase in brilliance of the projected picture when a Schmidt system is used may reach 8 times with respect to an ordinary lens system. Moreover, a spherical mirror has only spherical aberration in comparison to lenses, so that the Schmidt system provides magnification with less distortion than a lens system.

13-19. Large-screen projection

Home-type projection systems have been developed using a 3- to 5-in. diameter tube and a 20,000 to 30,000 volts accelerating high-voltage supply. A picture 3 to 4 ft wide with sufficient brilliance and contrast for home use can be obtained. Apparent luminosity or brightness can be increased with a special type of screen using the light directive properties of certain materials.

For public use in theaters, however, larger sizes are necessary and pictures several yards wide have been obtained. Such projection systems call for special tubes working under very high voltages, on the order of 100,000 volts. Some tube manufacturing problems arise, such as obtaining sufficiently high insulation strength and providing a phosphor screen capable of withstanding the bombardment of the high-energy electron beam. The phosphor layer thickness is increased, therefore, and a sizable amount of brilliance and contrast would be lost if the picture were observed from the side of the screen opposite the electron gun. Instead, the picture appearing on the inside face of the phosphor is used for projection. This necessitates offsetting the electron gun at an angle, as shown in Fig. 13-35, so that it does not obstruct the path of the optical rays. With this type of construction, normal scanning would give a trapezoidal picture; the top of the picture would be larger than the bottom because it is farther from the deflection center. This defect can be remedied by using a keystoneing correction arrangement, which consists essentially of modulating the amplitude of the horizontal sweep with the vertical sawtooth. A simplified circuit to accomplish this is shown in Fig. 13-36. A variable-gain

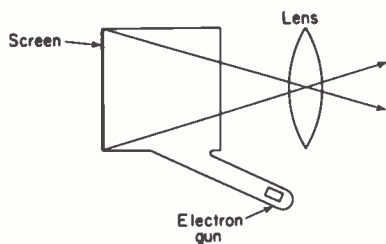


Figure 13-35. Special oblique beam tube for projection.

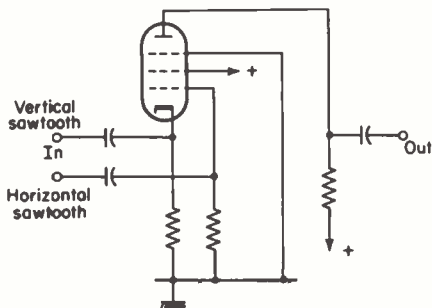


Figure 13-36. Keystoneing correction amplifier.

tube is used, and its slope is controlled by the vertical sawtooth. The resulting horizontal signal has the shape illustrated in Fig. 13-37. The number of horizontal cycles has been greatly reduced for clarity of the diagram.

Another distortion due to the same cause is that the horizontal lines in the picture do not have a constant spacing, but a spacing which increases from bottom to top of the picture. This distortion is corrected by using a vertical sawtooth which is not linear but has the shape shown in Fig. 13-38. This waveshape is easier to obtain and uses simpler circuits than the linear vertical sawtooth represented by the broken lines in Fig. 13-38.

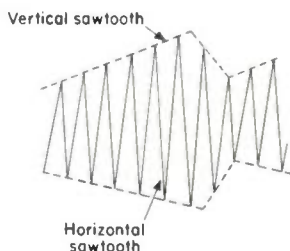


Figure 13-37. Keystoning correction waveform.

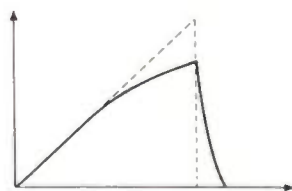


Figure 13-38. Vertical sawtooth linearity compensation.

13-20. Metal-glass tubes

A temporary shortage of the special glass used to make picture tubes originally resulted in the development of metal-glass tubes. In these tubes, shown in Fig. 13-39, glass is used only where necessary and unavoidable, e.g., for the neck, which must be transparent to magnetic fields, and for the screen, which must be transparent to light. The remainder of the tube, that is, the greatest part of the cone, is made of stainless steel. Two large metal-to-glass seals are unavoidable, and satisfactory techniques have been established to obtain sufficient bonding strength. The largest seal especially, between the front face plate and cone, is under heavy strain due to atmospheric pressure on the face of the tube, and the type of seal used is shown in Fig. 13-40. The external air pressure pushes the glass against the metal, whose elasticity moreover absorbs any additional strain due to temperature variations and improves the seal.

The metal cone constitutes the final anode. It is connected to a high voltage of 10,000 or 20,000 volts and serious precautions must be taken for its insulation, especially at the periphery of the screen.

Manufacture of metal-glass tubes is more difficult than the manufacture of standard all-glass tubes, and there is little point in using metal-glass construction when glass is available.

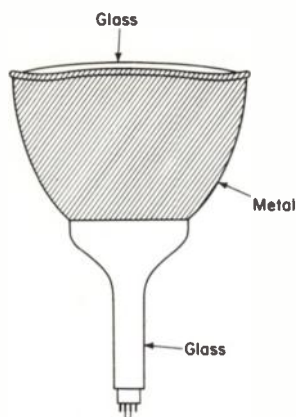


Figure 13-39. Metal-glass picture tube.

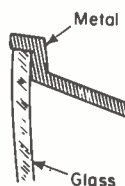


Figure 13-40. Metal-glass junction around the screen.

13-21. Rectangular tubes

Early picture tubes had symmetry of revolution and a circular face, mainly to ease glass-blowing fabrication problems. A circular screen, however, is illogical for television use, where the rectangular picture, shown in Fig. 13-41,

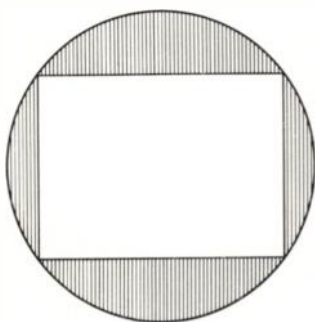


Figure 13-41. A rectangular picture does not fit well a round face tube.

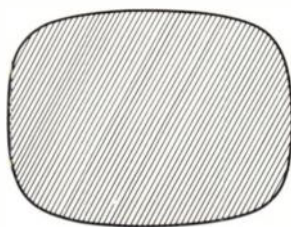


Figure 13-42. Shape of the screen of a rectangular tube.

leaves a good part of the screen unused. Corresponding increases in cabinet size and glass weight, as well as in total pressure exerted on the tube, accompany this useless area.

It was only logical for tube manufacturers to drop circular faces and replace them by rectangular faces adapted to television format, roughly in the ratio of

3 to 4, as soon as manufacturing techniques permitted this step. The angles are slightly rounded off in the present rectangular tubes to ease manufacturing difficulties, and the screen is plane, slightly convex, or cylindrical. The general shape of a so-called "rectangular tube" looks like the shaded portion of Fig. 13-42.

A convex screen may produce troublesome reflections, as shown in Fig. 13-43. It is recommended that television be viewed with some amount of ambient light to reduce eye fatigue, and a lamp will be reflected by a convex

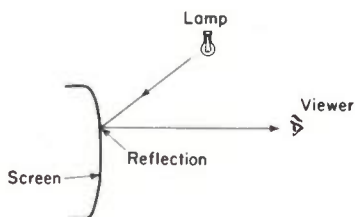


Figure 13-43. Light reflection on the screen.

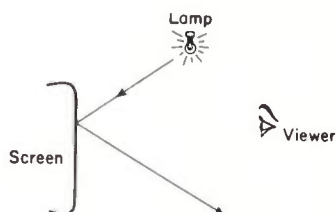


Figure 13-44. Advantage of a flat screen.

screen as an annoying luminous patch on the picture. Two remedies are available. The external face of the screen may be ground, etched, or frosted, thus producing diffuse reflection and reducing the defect. Also, the screen may be made convex only along the horizontal direction and straight only along the vertical direction. This is called a *cylindrical screen*, and any light source above level will be reflected downwards and will not bother the viewer, as shown in Fig. 13-44. An additional precaution to avoid reflections is to tilt the screen slightly forward, as shown in Fig. 13-45.

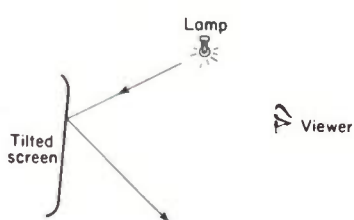


Figure 13-45. Tilting the screen slightly forward improves matters.

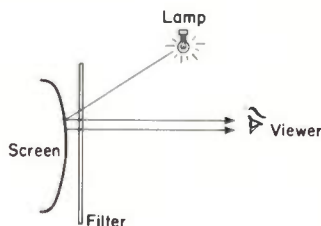


Figure 13-46. Neutral filter in front of the screen.

13-22. Gray filter

Ambient light reduces eye fatigue but at the same time it reduces apparent contrast by lighting the dark areas of the picture. An improvement can be effected by placing a neutral gray filter in front of the screen. Light rays coming from the screen go through the filter only once, and assuming the

filter transmission coefficient to be 50 per cent, these direct rays are reduced to 50 per cent. (Fig. 13-46.)

Light rays coming from some source external to the receiver and reflected by the screen have to go through the filter twice and are reduced to 25 per cent. It is thus evident that the filter improves the ratio of emitted light to reflected light and hence the apparent contrast of the picture.

The apparent brightness, however, is also reduced but this is easily corrected by readjusting the brilliance control.

Some tubes have a built-in gray filter—the glass used for the face is not colorless but tinted gray. For other tubes, a glass or plastic filter may be placed in front of the screen, replacing the usual safety plate.

On a sunny summer day, the contrast in nature may reach the enormous value of 10,000 outdoors. The contrast of the same scene televised is only on the order of 100 on a picture tube screen, and could be doubled by using a gray filter. It is somewhat surprising that such a reduced ratio, compared with nature's range, still provides a satisfactory picture.

13-23. *Ion blemish*

Regardless of how high the vacuum has been developed inside a picture tube, there is always some amount of residual gases left in the tube, and the electrodes themselves may also give off some gases when heated. Electrons shooting away from the cathode will collide unavoidably from time to time with gas molecules and, when trapped by the molecules, will give them a negative charge and transform them into negative ions.

Such an ion will have the same charge as an electron, but its mass will be several thousand or even several hundred thousand times larger. It will be accelerated and attracted by the high-voltage positive field and projected onto the screen through the deflection field.

Now a fundamental difference between electrostatic and electromagnetic deflection becomes apparent. Reference to the deflection equations will show that the mass of the particle does not appear in the expression for electrostatic deflection, so that in this case the heavy ions are deflected and fall uniformly all over the screen.

For electromagnetic deflection, however, the particle mass appears in the expression for deflection, with the result that the heavy ions will be considerably less deflected than the very light electrons. These heavy negative ions will fall on a narrow limited zone around the center of the screen.

Continuous bombardment of this central area by heavy ions will progressively impair the light-emitting property of the phosphor in this area, and a brown or darkened blot, called an *ion burn* or *blemish*, will become apparent after some extended use of the tube.

This defect is peculiar to electromagnetic deflection tubes and will become

noticeable after a time which depends on how good a vacuum prevails inside the tube. Two methods are used to eliminate it: aluminization and ion trapping.

13-24. Aluminization

An extremely thin layer of aluminum is coated over the phosphor layer by vacuum deposition, as shown in Fig. 13-47. Its thickness is 50 millimicrons or less, so that the light and small electrons, accelerated to high velocity, shoot straight through it and reach the phosphor layer to excite luminescence.

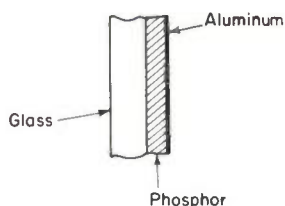


Figure 13-47. Aluminum layer at the back of the phosphor.

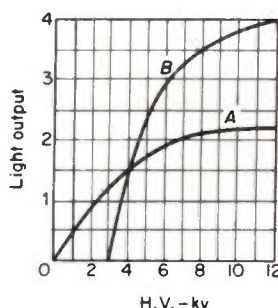


Figure 13-48. Comparison of luminosity with and without aluminum backing.

Ions are big and heavy—and what is more, they are accelerated to only a relatively slow speed because their mass appears in the expression for velocity. They are stopped by the aluminum layer and thus effectively prevented from reaching the phosphor and ruining it eventually.

Aluminization brings other valuable advantages as well. In an ordinary tube, as previously explained, the high-voltage circuit is indirectly closed through secondary electrons that are knocked off the screen and fall on the final anode, e.g., the graphite internal aquadag coating. As long as the secondary emission coefficient is equal to 1, the screen has zero total charge and the current flow mechanism functions correctly.

However, if the high voltage is increased to obtain greater brightness, the secondary emission phenomenon is partially inhibited and the screen takes on an over-all negative charge. One result is that this negative charge slows down the electrons hitting the screen and thus reduces the amount of light they produce. The phenomenon is called “sticking,” because the electrons seem to stick to the screen.

The visual result produced by sticking is a kind of saturation of light. Above a given value, increasing the acceleration voltage produces practically no increase in light output, as shown by curve A of Fig. 13-48.

When the tube is aluminized, however, the metallic film offers an easy path to the graphite coating and directly completes the high-voltage circuit. No charge appears on the screen, and evidently there is no light saturation due to sticking. The curve of light output vs. acceleration voltage is shown in *B* of Fig. 13-48. The comparison clearly indicates the advantage of aluminized screens above a certain value of high voltage in increased light output with no saturation.

Another advantage is that aluminization increases the apparent brilliance and contrast. In an ordinary tube, shown in Fig. 13-49, half of the total

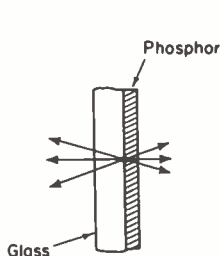


Figure 13-49. Loss of emitted light.

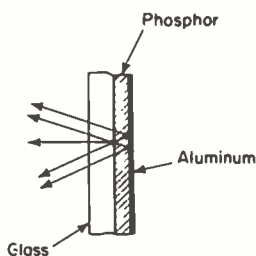


Figure 13-50. Mirror effect of the aluminum backing.

emitted light of luminescence is directed back inside the tube and is lost. Moreover, this light can be reflected by the tube walls, and fall back on the screen, thus reducing the apparent contrast.

In an aluminized tube shown in Fig. 13-50, light rays starting inward are reflected on the metallic layer and proceed outward, thus increasing apparent brilliance and contrast.

13-25. Ion trap

The different behavior of electrons and ions can be put to good use to sort out the desired electrons and eliminate the undesired ions. Figure 13-51 represents the oblique electron gun used for this purpose. By far the greatest part of the dangerous ions is produced in the vicinity of the cathode, so the oblique arrangement is designed to eliminate them.

Both electron and ion beams are accelerated by the anode positive voltage along the axis of the gun. A magnetic field provided by an external magnet bends the electron beam to bring it along the axis of the tube, where it proceeds forward to the screen. However, the ions remain practically undeflected by the magnetic field; they hit the anode wall and are eliminated.

Another common method (Fig. 13-52) uses an oblique slit electron gun as shown. The oblique electric field deflects both electrons and ions and projects

them onto the anode wall. However, an external magnetic field brings the electrons back along the axis, but it has almost no action on the ions and lets them reach the anode wall and be eliminated as before.

The magnetic field is usually provided by a small magnet, called an *ion trap*, which is slipped over the neck of the tube. The field may sometimes be adjusted through a magnetic shunt. In all cases the magnet assembly may be

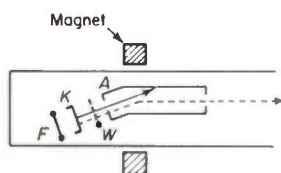


Figure 13-51. Off-axis electron gun.

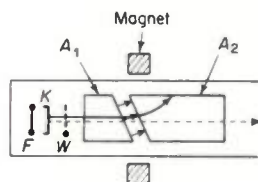


Figure 13-52. Oblique slot electron gun.

moved forward or backward and rotated around the tube until the correct position is found. This adjustment is easy: optimum position of the magnet corresponds to a maximum of screen brightness.

13-26. External coating

As we will see later, most high-voltage supplies work at a relatively high frequency and good filtering can be obtained with capacitances of a few hundred micromicrofarads. Nonetheless, high-voltage insulation makes such

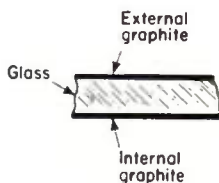


Figure 13-53. Double graphite coating.

components costly, and a clever method of eliminating them is to coat the outside of the bulb of the picture tube with a conducting layer of graphite. The tube then carries two coatings (Fig. 13-53), the internal one being the anode coating. With the glass between them as a dielectric, the two conducting layers constitute a good capacitor of 500 to 2500 $\mu\mu f$, which is quite suitable as a filter capacitance by the simple expedient of grounding the external layer.

13-27. Permanent magnet focusing

The focusing magnetic field can be obtained by the use of a focus coil as previously indicated, with good results. However, a coil is an expensive component, and the necessary current means that some heat will be produced in the resistance of the coil, with a resulting resistance variation. The focus will then vary, and it will be necessary to readjust the focusing adjustment perhaps several times until a stable working temperature is attained. Also, if the supply voltage varies, the focus may exhibit a corresponding change.

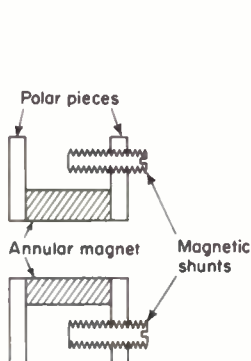


Figure 13-54. Magnetic shunt adjustment of a focus magnet.

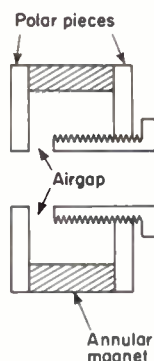


Figure 13-55. Variable airgap adjustment of a focus magnet.

For all these reasons, permanent magnet focusing is frequently used in one form or another. Focusing adjustment can still be obtained by using magnetic shunts, as shown in Fig. 13-54, or an adjustable airgap, as shown in Fig. 13-55.

The external magnetic field may be intense and may interfere with the magnetic field deflection system if deflection coils are near the focus assembly. The result is defocusing at the periphery of the picture, for which a simple remedy is given in Fig. 13-56: two magnets are used back to back, and focusing is adjusted by modifying their spacing.

Frequently, a permanent magnet focus adjustment is not one of the controls provided by the manufacturer as immediately accessible to the purchaser. It is located on the neck of the tube and reached at the back of the receiver. Permanent magnet focusing is rarely adjusted once the initial adjustment is made, for the reason that permanent magnet focusing is essentially independent of supply voltages and that some amount of variation is tolerable. The result is that the viewer has one less control to take care of, with the consequent simplification of adjustments.

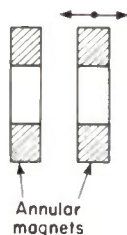


Figure 13-56. Variable spacing adjustment of a double focus magnet.

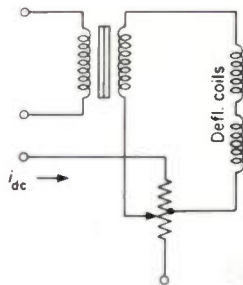


Figure 13-57. Electrical centering circuit.

13-28. Electrostatic focusing

Focusing coils or even magnets can be dispensed with if electrostatic focusing is employed.

Early types of electromagnetic deflection-electrostatic focusing tubes simply used an electron gun adapted from the all-electrostatic tube described in Secs. 13-1 and 13-2. The disadvantage is that in the design of electrostatic tubes, the focus anode should be connected to an adjustable voltage roughly equal to one-fifth of the final accelerating voltage. With large-size tubes having a high voltage on the order of 10,000 volts or more, the first anode necessitated a *variable* 2000-volt supply. Providing such a control was in no way a convenient proposition and carried with it a certain danger to the user. A first improvement was to change the geometry and disposition of the electrodes in the gun in such a way that the first anode current was practically zero. Nevertheless, there remained the insulation problem and the difficulty of obtaining a 2000-volt supply.

The breakthrough came with the development of a new electron gun, mechanically more complex than the standard type, but presenting the big advantage of a 200- to 400-volt optimum voltage for the focusing electrode. Such a voltage is readily available in a television receiver and poses no particular insulation or safety problems.

Moreover, this type of gun has been improved to the point that acceptable focusing is obtained for zero voltage on the focus electrode, which can then be directly tied to the cathode, thus eliminating the focusing adjustment entirely.

13-29. Electromagnetic centering

Centering controls are easily installed for use with electrostatic deflection tubes, as we have seen, since it is necessary merely to vary the d-c potentials of

the deflection plates. However, centering is slightly more complex with electromagnetic deflection tubes.

One method would be to superpose a d-c current over the sawtooth deflection current through the coils. The circuit of Fig. 13-57 shows how this current could be adjusted in magnitude and direction to move the picture bodily in two opposite directions. A practical circuit might include one or two capacitances to short the potentiometer for alternating current.

Another method is to mount the focus coil or magnet in such a way that it can be mechanically tilted in any direction by a few degrees and locked in position. The whole picture moves in the direction of the tilt. The drawback is that focusing may suffer somewhat in the process, especially at the periphery of the screen. Best over-all focusing is obtained when the focus assembly is coaxial with the tube. Nevertheless, the system is simple and economical and widely used to correct for the common, usually small decentering.

A third means of centering the picture is provided by one or two small adjustable centering magnets, which are slipped over the neck of the tube. Figure 13-38 illustrates a simple design: the magnet is a piece of steel wire producing a magnetic field and spring-clamped on the tube neck. Rotating the magnet and moving it along the tube adjusts centering.

Usually two such magnets are used, and their combined use allows a centering adjustment in all directions on the order of 0.1 tube diameter, which is sufficient for practical purposes.



Figure 13-58. Magnetic centering with an auxiliary magnet.

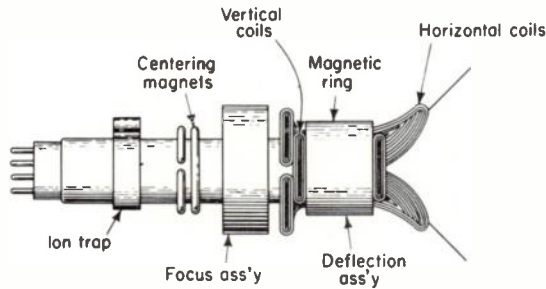


Figure 13-59. Components on the picture tube neck.

13-30. External components

The number of auxiliary components slipped over the tube neck may be large. For example, as shown in Fig. 13-59, starting from the socket end of the tube neck, the first component would be the ion trap, then the centering magnets, then the focus coil or magnet assembly, and then the deflection coil assembly.

Some of these components, of course, may be missing in any particular case. Not all tubes need an ion trap, for example. Also, an arrangement other than ring magnets can be used for centering. Similarly, an electrostatic focus tube would not have any focusing assembly, etc. The only component certain to be present (since electrostatic deflection is never used in large tubes) is the deflection yoke assembly with its two sets of coils for horizontal and vertical deflection and, generally, a ring of high-permeability material around the coils.

Problems

1. Compare electrostatic and electromagnetic deflection from the point of view of the circuit designer.
2. Compare electrostatic and electromagnetic focusing from the standpoint of circuit simplicity.
3. Calculate the velocity of an electron accelerated by (a) 2000 v, (b) 10,000 v, (c) 20,000 v, (d) 100,000 v.
4. A gas molecule has been ionized by trapping an electron. Its weight is 10^{-23} g. Calculate the velocity with which it hits the screen of a picture tube which has a high voltage of 15,000 v. Calculate the kinetic energy of the molecule.
5. An electron accelerated under 12,000 v enters a magnetic field of 600 ampere turns at an angle of 26 degrees. Evaluate the force acting on the electron.
6. A gas molecule weighing 6.2×10^{-23} gram has been ionized by trapping an electron and has been accelerated by 14,000 v when it enters at an angle of 15 degrees a magnetic field of 21 gauss. Calculate the force acting on the molecule.
7. Show in detail how an electromagnetic lens behaves for electrons as an optical lens for light.
8. Compare series- and shunt-type focusing coils. Discuss their advantages and disadvantages.
9. An electrostatic tube has deflection plates 3 cm long and 1 cm apart, 25 cm from the screen. The high voltage is 2000 v. Calculate the length of the trace on the screen when a sine voltage of 15 v rms is applied between the deflection plates.
10. Calculate the deflection sensitivity of a tube that has deflection plates 2 cm long, 0.6 cm apart, and 16 cm from the screen.
11. Calculate the transverse velocity of an electron accelerated under 1000 v and shooting between deflection plates 2.5 cm long, 0.8 cm apart, between which a d-c voltage of 65 v has been applied.
The deflection plates are 19 cm away from the screen. Calculate the electron deflection.
12. A gas molecule weighing 4×10^{-23} g has been ionized by trapping an electron and has been accelerated by 1000 v. It enters the space between deflection

plates 2.5 cm long, 0.8 cm apart. There is a d-c voltage of 65 v between plates. Calculate the transverse velocity of the ion.

The deflection plates are 19 cm away from the screen. Calculate the ion deflection.

13. A tube has a deflection sensitivity of 0.4 mm/v/kv and a high voltage of 2400 v. A 24-v rms sine voltage is applied to the deflecting plates. Calculate the deflection.

14. A 12-cm diameter tube has a deflection sensitivity of 0.3 mm/v/kv and is operated under 1800 v high voltage. Calculate the rms sine voltage necessary to sweep the beam across the tube face.

15. Design a circuit for electrostatic centering where the average d-c potential between plates is kept constant.

16. Compare symmetrical and nonsymmetrical electrostatic deflection.

17. Discuss the advantages and disadvantages of post-acceleration.

18. An electron accelerated under 12,000 v enters a deflection field of 240 ampere turns. Evaluate the radius of curvature of its trajectory.

19. An electron enters a magnetic deflection field of 7500 gauss extending over 5 cm. The high voltage is 16,000 v, and the distance from the screen to the deflection center is 30 cm. Calculate the deflection.

20. A gas molecule weighing 10^{-22} g has been ionized by trapping an electron and accelerated under 11,000 v. It enters a magnetic field created by deflection coils 5.5 cm long, having 600 turns, and carrying a peak current of 1.2 amp. The screen is 25 cm away from the end of the coils. Calculate the ion deflection and the deflection of an electron under the same conditions.

21. Suggest three arrangements for vertical and horizontal deflection coils other than the coaxial type.

22. What would be the best value of screen persistency? Justify your answer.

23. A tube face has a radius of curvature of 60 cm and a width of 40 cm. It displays vertical bars spaced 4 cm apart. For a viewer sitting several yards away in the axis of the screen, calculate the apparent spacing of the bars and evaluate the resulting geometrical distortion.

24. Calculate the total force due to atmospheric pressure on the face of a tube that has a rectangular screen 17 in. in diagonal.

25. Explain why the phosphor layer should be thick in a projection tube.

26. Compare aluminization and ion trapping.

27. Explain the appearance of ion burn or blemish.

28. The metallic layer on an aluminized screen fulfills several functions. Explain how in detail.

29. A round tube has a conical part coated on both sides with graphite, starting 2 cm from the face and ending 5 cm from the neck. The face diameter is 41 cm; the neck diameter is 2.5 cm. The cone total angle is 90 degrees. The glass is 0.3 cm thick, and it has a dielectric constant of 5.3. Calculate the capacitance between the two graphite layers.

30. Design a complete centering system using d-c current through vertical and horizontal deflection coils.

Chapter 14

ELECTROMAGNETIC DEFLECTION

Electromagnetic deflection is universally used in television receivers. It requires sawtooth currents to produce variable magnetic fields that deflect the electron beam.

Two sets of deflection coils, combined into a deflection yoke, are required for vertical and horizontal deflection. Inductance then comes into play, with widely different effects for vertical and horizontal sweeps because of the large difference in frequencies, respectively 60 and 15,750 cps. The fact that electromagnetic deflection is used bears heavily on the circuit design of the vertical and horizontal sweep sections.

14-1. Deflection current

At the very outset, a distinction must be made between sweep and deflection in order to fully comprehend electromagnetic sweep and deflection circuit requirements. A linear sweep produces a displacement of the spot on the tube screen which is exactly proportional to time. A linear deflection produces an angular deflection of the electron beam exactly proportional to time.

These two linearity requirements coincide only when the tube face is a perfectly spherical surface having its center at the deflection center. Since present commercial picture tubes have almost flat screens and do not fulfill this condition, a linear deflection does not produce a linear sweep. This can be easily compensated for by circuitry, however, as will be seen, and linear deflection will be considered as the desirable goal.

Since deflection is proportional to the current through the deflection coils, the problem of producing linear deflection reduces to the problem of producing a linear deflection current. In typical commercial circuits, deflection current has a peak-to-peak value of a few tens of milliamperes for vertical sweep and of a few hundred milliamperes for horizontal sweep. Such intensities can be obtained only with power tubes, and only with special power tubes for horizontal sweep.

14-2. Deflection voltage

A deflection coil or pair of coils not only has inductance L , but also some ohmic resistance R and distributed capacitance C . In addition, unavoidable

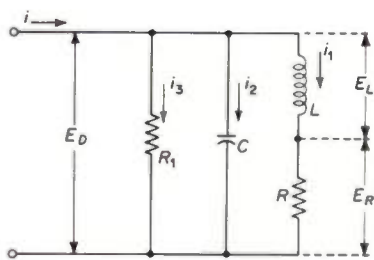


Figure 14-1. Equivalent deflection circuit.

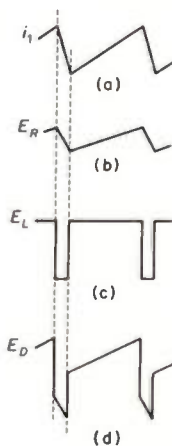


Figure 14-2. Waveforms in a deflection circuit.

losses in insulating or magnetic material or in shields are equivalent to a parallel resistance R_1 , in which may also be included any additional damping. In comparison to a simple inductance, a more precise representation of a deflecting coil looks as shown in Fig. 14-1.

The fundamental requirement is that current i_1 through L be linear. This linear waveform i_1 is represented in (a) of Fig. 14-2. This current flowing through R will create across it a voltage drop E_R represented in (b) of the same figure.

The voltage E_L across the inductance is

$$E_L = L \frac{di_1}{dt}$$

that is, proportional to the derivative of current. It is then a rectangular wave as shown in (c) of Fig. 14-2.

The deflection voltage E_D , therefore, is equal to the graphical or vector sum of the voltages across inductance and resistance,

$$E_D = E_L + E_R$$

and is shown in (d) of the figure.

But this voltage E_D will also be applied to capacitor C , creating a current i_2 , and to resistor R_1 , creating a current i_3 . This last current is simply

$$i_3 = \frac{E_D}{R_1}$$

and therefore will have the shape of Fig. 14-2(d).

The capacitance current i_2 is

$$i_2 = C \frac{dE_D}{dt}$$

that is, proportional to the derivative of the deflection voltage. Since the deflection voltage E_D contains two vertical transitions, its derivative would contain two large-amplitude narrow pulses at these points. The transition in the negative direction precedes the retrace and can be neglected, retrace linearity being of no importance. The positive transition at the beginning of the sweep should correspond to a large narrow pulse charging up the capacitance at the beginning of the sweep. Since such a pulse is difficult to obtain by simple means, it is purely omitted. The importance of this distortion depends on the value of C and is generally small. The result is a slowing down of the beginning of the sweep, as shown in Fig. 14-3.

14-3. Nonlinearity

If the ohmic resistance of the deflection coil could be omitted, it would be necessary only to keep the voltage across the coil constant in order to obtain a linear current. This results directly from

$$E = L \frac{di}{dt}$$

because if E is constant, so must be di/dt and the current, having a constant slope, is linear.

In reality, there is always some resistance, however small, and the current will vary exponentially, its derivative di/dt decreasing when i increases.

Nonlinearity can be expressed as the variation x of di/dt during a sweep period. R being the coil resistance, x is as a first approximation given by

$$x = \frac{TR}{L}$$

where T is the sweep duration in seconds, R is in ohms, and L in henrys.

For horizontal sweep

$$T = 53 \text{ microseconds}$$

approximately. Assuming a maximum nonlinearity of 10 per cent is tolerated, then

$$x = 0.1 \geq \frac{53 \times 10^{-6} R}{L}$$

or

$$\frac{R}{L} \leq 1880,$$

There is not much difficulty in obtaining a ratio of 1880 ohms/henry for horizontal deflection coils.

For vertical sweep

$$T = 15 \text{ milliseconds}$$

and for 10-per cent nonlinearity,

$$x = 0.1 \geq \frac{15 \times 10^{-3} R}{L}$$

or

$$\frac{R}{L} \leq 6.7$$

A common value for the R/L ratio of deflection coils would be 1000 to 2000 ohms per henry. This would suit requirements for horizontal scanning, but it is evident that the problem of vertical linearity is serious and that a rectangular wave drive cannot be used for vertical coils.

14-4. Total current

If the capacitance current i_2 is neglected, total current i is

$$i = i_1 + i_3$$

and is a combination of sawtooth and rectangular waves, in a proportion which depends on component values.

The shape during retrace is of no interest. The only important point is that retrace should be fast enough to be accomplished within the duration limits set by the standard. In fact, retrace is never linear.

During the sweep period, E_L is constant and E_R increases linearly if i_1 is linear. The relative amplitudes are easily found. Let I be the amplitude of deflection current i and T be the sweep duration. Then

$$E_L = \frac{LI}{T}$$

is the amplitude of the rectangular component, and

$$E_R = RI$$

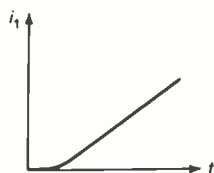


Figure 14-3. Slow start of a sawtooth current.

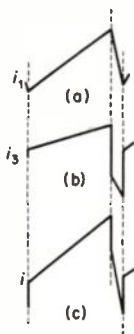


Figure 14-4. Combination of rectangular and sawtooth signals.

is the amplitude of the sawtooth component. Their ratio is

$$\frac{E_R}{E_L} = \frac{TR}{L} = x$$

Figure 14-4 shows in (a) deflection current i_1 , in (b) parallel resistance current i_3 , and in (c) total current i , which is the sum of the two preceding currents.

14-5. Horizontal and vertical sweep

Since the sweep duration T appears in the ratio of sawtooth to rectangular components, for the same R/L ratio, the waveshapes will look very different for horizontal and vertical scanning. T is approximately 53 microseconds for horizontal and 15 milliseconds for vertical scanning, which corresponds to a duration ratio of 283. Assuming an identical R/L ratio for the two sets of deflection coils, the proportion of sawtooth component is 283 times larger for vertical scanning than for horizontal scanning.

Therefore, for vertical scanning the coil inductance is negligible compared with its resistance, so that the sawtooth component largely predominates in the deflection voltage. At field frequency, shunt capacitance can be totally neglected, and total current i should be a sawtooth wave essentially.

For horizontal scanning, the coil resistance is negligible compared with its inductance, and the deflection voltage is essentially a rectangular wave. Deflection current, therefore, is a sawtooth wave, parallel resistance current is a rectangular wave, and total current i is a combination of both. Moreover, at line frequency the parasitic shunt capacitance can no longer be neglected and must be taken into account.

Another interesting conclusion concerns the amplitude of voltage across the coils.

For vertical sweep, deflection voltage during sweep is

$$E_D = RI$$

very nearly, and reaches a few volts.

For horizontal sweep, deflection voltage during sweep is

$$E_D = \frac{LI}{T}$$

very nearly, and easily reaches 100 volts or more.

During retrace, the same current variations will occur in a shorter time and will result in a much larger voltage across the coils. If horizontal retrace is 12.5 per cent of the period, then for linear retrace the voltage across the horizontal coils would be 8 times higher than during sweep and could easily exceed 1000 volts. Moreover, retrace is never linear, and consequently instantaneous current variation will have a value much larger than average. During retrace, voltages of several thousand volts are commonly developed in horizontal scanning systems.

14-6. Power tube

The usual scanning arrangement contains a relaxation oscillator tube and a power amplifier. It is worth pointing out that the grid of the power tube is always fed a positive-going sawtooth, i.e., a sawtooth having a positive slope during sweep, so that anode current may reach high values with the usual supply voltages.

The power stage may have different modes of operation. When it operates as Class A, amplification is approximately linear and the grid of the tube receives a voltage such that the desired plate current is obtained.

The power stage may also work as a switch, in which case the tube is alternately switched on and off, and the anode load is essentially inductive to obtain a current sawtooth waveform.

Finally, as a third method, feedback circuits may be employed to compel plate current to vary linearly or in conformity with the voltage signal applied to the grid.

The first and third systems are mainly used for vertical scanning and the second for horizontal scanning.

Because of the difference in frequency, horizontal scanning will need much more power than vertical frequency. A few watts are sufficient for vertical scanning, whereas several tens of watts are necessary for horizontal scanning. Also, because of the large voltages developed during horizontal retrace, the anode of a line power tube will have to be well insulated. Customarily, a

separate anode connection is made at the top of the tube. Finally, instantaneous values of horizontal deflection current are large, and in some cases the tube must provide them at rather low anode voltage.

14-7. Deflection coil coupling

The deflection coil could be driven directly by the power tube, as shown in Fig. 14-5, and in this case it is said to be a *high-impedance coil*. It could also be a low-impedance coil if driven through a transformer as shown in Fig. 14-6.

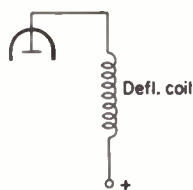


Figure 14-5. Direct drive of the deflection coil.

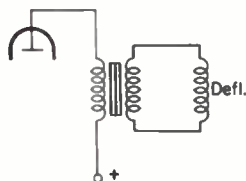


Figure 14-6. Transformer coupling of the deflection coil.

With a high-impedance coil, the simple circuit of Fig. 14-5 is unsuitable because the d-c plate current flowing through the coil would produce a large permanent deflection. In common practice, the arrangement used is shown in Fig. 14-7. A capacitance coupling is used to block direct current, and a choke

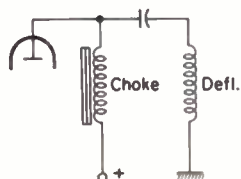


Figure 14-7. Capacitance coupling of the deflection coil.

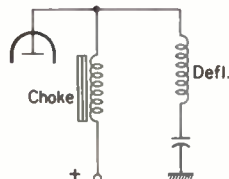


Figure 14-8. Practical coupling for high impedance coil.

is inserted in the anode circuit. A resistor is rarely used in place of the choke because of the high current and resulting high-voltage drop. The circuit of Fig. 14-8 may sometimes be more convenient in practice because one side of the capacitor is grounded. If such a circuit is employed, the capacitor must have a large value to avoid distortion of the signal and it is generally of the electrolytic variety.

With a high-impedance vertical deflection coil, the number of turns is large, say several thousand, and the wire has a small diameter to reduce bulk

so that the coil L also has high resistance R . The R/L ratio then introduces nonlinearity, and an easy way to correct for distortion is shown in Fig. 14-9. The anode load is made of L_1 and R_1 in series. Compensation is obtained for

$$\frac{R}{L} = \frac{R_1}{L_1}$$

and R_1 is made adjustable, usually as a high-dissipation, wire-wound potentiometer of a few thousand ohms. This control is called the *vertical linearity control*.

The above problem does not arise for line scanning, where the coil resistance is almost negligible compared to the coil inductance.

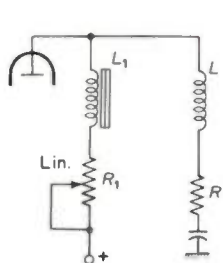


Figure 14-9. Linearization adjustment by R_1 .

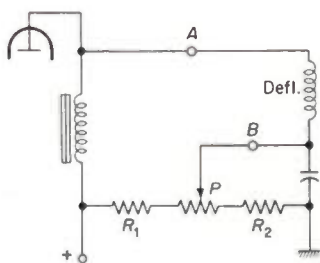


Figure 14-10. Electrical centering circuit.

14-8. Centering

Electrical centering may easily be included in high- or low-impedance deflection systems. The idea is to push a d-c centering current through the coil while providing a separate path for the sawtooth current.

A circuit suitable for a high-impedance coil appears in Fig. 14-10. A resistance bridge R_1 - P - R_2 across the supply voltage applies at point B a voltage which may be adjusted to a potential above or below that at point A . In this way, a controllable current can be made to flow in either direction through the deflection coil. Such a control circuit may be applied to horizontal and vertical centering, respectively.

Figure 14-11 illustrates an adaptation of the idea for a transformer-driven low-impedance coil. A center-tapped centering potentiometer P is used. Notice that the d-c centering current flows through the transformer secondary and could partially saturate its core. This can be avoided by variations of the circuit, if necessary. However, no permanent d-c current due to the tube flows through the coil in the two preceding diagrams and the centering adjustment need correct only for small displacements.

14-9. Transformer drive

Introducing a transformer between tube and deflection coil undoubtedly offers many advantages. The transformer can be of the step-down type and deliver to the coil a current several times larger than the tube current. This facilitates the design of the tube circuit. In addition, the coil may have a small number of turns of large-size wire, thus reducing its inductance and also its resistance. The pulse voltage developed during flyback is smaller, the insulation problem is easier to solve, and the coils are mechanically stronger and easy to produce.

However, all is not on the advantage side for transformer-driven deflection coils, because the insulation difficulties are now transferred to the primary of the transformer, where peak voltages across the coils are multiplied by the transform ratio. Also, transformers are in no way perfect components—they introduce parasitic elements with resulting distortion or undesirable effects that have to be eliminated or in some way corrected.

The circuits differ widely for horizontal and vertical deflection and will be investigated in separate chapters. However, common basic principles are identical and will be introduced presently.

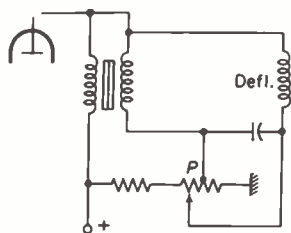


Figure 14-11. Electrical centering circuit with transformer drive.

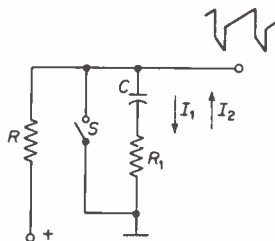


Figure 14-12. Resistance R_1 provides the rectangular component.

14-10. Rectangular and sawtooth waves

An important linearity requirement, as we have seen, is that the complex sweep signal should contain the proper proportion of sawtooth and rectangular components.

Fortunately, this is easy to obtain, and the circuit is indicated in Fig. 14-12. It shows a charge circuit RC and a relaxation oscillator symbolically represented by a switch S . The only extra element is resistance R_1 . When the capacitor C charges up, a current I_1 flows and creates a positive voltage drop $R_1 I_1$. During discharge, an opposite and larger current I_2 flows and creates a larger negative voltage drop $R_1 I_2$. These two opposite voltage drops add

to the sawtooth voltage across C to produce the desired waveform. Figure 14-13 illustrates the current waveform I , the voltage V_C across the capacitance, the voltage V_R across the resistance, and the output voltage V_T which is simply the graphical sum of V_C and V_R . The proportion of rectangular wave in the final signal is evidently controlled by R_1 . Accordingly, R_1 is a potentiometer of a few thousand ohms and as a control has been variously labeled *linearity*, *peaking*, *form*, or *drive adjustment*. With R and L as the

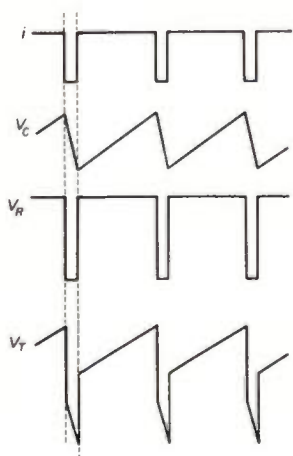


Figure 14-13. Waveforms obtained in the peaking circuit.

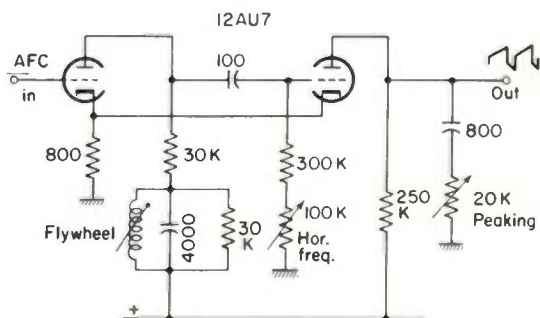


Figure 14-14. Practical circuit to provide the horizontal driving signal.

resistance and self-inductance of the deflection coil, respectively, correct adjustment results when the two time constants are identical, or

$$R_1 C = \frac{L}{R}$$

For reasons previously given, this circuit is used only in horizontal sweep generators. Figure 14-14 illustrates a practical arrangement applied to a multivibrator.

14-11. Deflection coils

Two sets of deflection coils are necessary—one for horizontal or line deflection and one for vertical or field deflection. They are combined in a deflection yoke slipped over the neck of the tube. Some magnetic material is commonly employed to increase efficiency. Since horizontal sweep current has a recurrent frequency of 15,750 cps and a sawtooth waveform, it will

contain harmonics of the fundamental frequency up to perhaps the 20th, or roughly 300 kc, and the magnetic material has to be suitable for such high frequencies. Magnetic ceramics of the ferrite variety are commonly used for this purpose.

The axial distribution of the magnetic field intensity created by a deflection coil has the form shown in Fig. 14-15. Field intensity depends on the distance from the center of the coil and decreases progressively, to become negligible at points *A* and *B*. It has a maximum value H_a at the center of the coil. The total magnetic efficiency of the coil may be expressed by the area under the curve. Let *S* be this area.

The same area would be enclosed by the rectangle shown in dashed lines, having a height H_a and a width

$$l = \frac{S}{H_a}$$

This rectangle is said to represent the field of an equivalent perfect coil of length *l* having constant field intensity H_a inside the coil and zero field intensity outside the coil. Such an equivalent coil would produce the same deflection as the real coil with the same current flowing and can be used instead of the real coil in deflection problems, with resulting simplification of the calculations; *l* is also called the *equivalent length* of the coil.

14-12. Deflection angle

The problem of electromagnetic deflection has already been approached in Chap. 13, devoted to picture tubes. It will now be considered from a deflection coil point of view, and the result will have some bearing on the practical design of the deflection coil.

The following two expressions have been established previously:

$$v = \left(\frac{2eE}{m} \right)^{1/2}, \quad r = \frac{mv}{He}$$

where *v* = electron velocity,

m = electron mass,

e = electron charge,

E = accelerating voltage,

H = magnetic field intensity,

r = radius of curvature of electron trajectory.

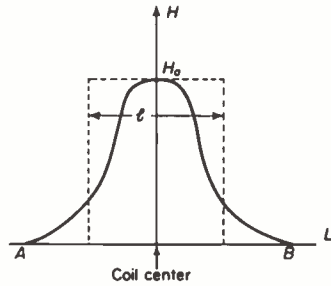


Figure 14-15. Practical magnetic deflection field and theoretical equivalent.

Actually, the expression for velocity neglects the fact that the electron mass increases with velocity due to relativity. The relativity correction can be neglected if the accelerating voltage is lower than 20,000 v, which is generally true in television.

Referring to Fig. 14-16, l is the equivalent length of the coil over which magnetic field H acts upon the electron. When the electron escapes the field, its flight follows the tangent to the circular deflection trajectory. This tangent cuts the axis of the tube at point O which is the center of deflection and as a

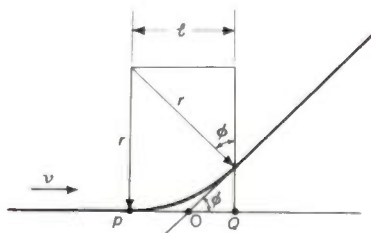


Figure 14-16. Evaluation of magnetic deflection.

first approximation is fixed at the center of the coil. The two angles ϕ indicated are equal because their sides are perpendicular two by two. Then

$$l = r \sin \phi$$

and substituting for r ,

$$l = \frac{\sin \phi m v}{e H}$$

Substituting for v ,

$$\sin \phi = 0.7 l H \left(\frac{e}{E m} \right)^{1/2}$$

or in practical units,

$$\sin \phi = \frac{0.3 l H}{E^{1/2}}$$

with l in centimeters, E in volts, and H in gauss.

This expression shows that a large deflection angle will be obtained with coils having a large length l and producing a large field H , that is, having a large number of turns and passing an intense current. Also, it shows again that the deflection angle varies inversely with the square root of the acceleration voltage E . In addition, the preceding expression, embodying e/m , shows that the deflection angle depends on the charge-to-mass ratio of the particle and will be much smaller for ions than for electrons.

A high deflection current implies large power and power tubes and is costly. It would seem, then, that the best way to increase deflection sensitivity would be to increase the length and the number of turns of the coil. However, both possibilities have limitations.

14-13. Length of the coil

Figure 14-17 shows the tube cut through its axis in the deflection region. The equivalent length of the coil is l , and the deflection angle is ϕ . A is the center of curvature of the circular trajectory, and as shown before, the angle

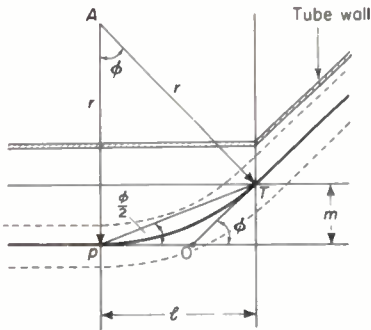


Figure 14-17. Calculation of deflection coil length.



Figure 14-18. Typical shape of a coaxial coil.

indicated at A has a value of ϕ . The angle indicated at P between the tangent and the chord then has a value $\phi/2$ as shown. Consequently,

$$m = l \tan \frac{\phi}{2}$$

Assume the tube neck has an inside radius of 18 mm. The electron beam in the deflection region has a radius of 1.5 mm typically. Therefore m is limited to

$$18 - 1.5 = 16.5 \text{ mm}$$

because if it were larger, the electron beam would hit the neck of the tube.

For a so-called "90-degree tube," the horizontal deflection angle is the largest and would be roughly 80 degrees, which corresponds to $\phi = 40$ degrees. Substituting for m and $\phi/2$ provides

$$16.5 = l \tan 20$$

$$16.5 = 0.36l$$

$$l = 46 \text{ mm or } 1.8 \text{ in.}$$

approximately, so that the length of the coil is severely limited. An improvement can be effected by using coils of the shape shown in Fig. 14-18. The

flared-up front fits over the conical part of the tube. This improvement is especially useful with tubes that have a wide deflection angle.

14-14. Deflection coefficient

The only part of the magnetic field which is useful for deflection is the volume inside the tube. For the ideal coil producing a constant field intensity H in a cylinder of length l and radius a equal to the internal radius of the tube neck, the useful volume V_u is

$$V_u = \pi a^2 l$$

The energy W accumulated in the field is

$$W = \frac{H^2 V_u}{8\pi}$$

Using the practical expression for $\sin \phi$ to solve for H provides

$$H = \frac{\sin \phi \cdot E^{1/2}}{0.3l}$$

Substituting for H and V_u in the expression for W ,

$$W = \frac{E a^2 \sin^2 \phi}{0.72l}$$

with E in volts, a and l in cm, and W in ergs.

Energy is provided to the coil, of self-inductance L , as a current I . This energy has a value $LI^2/2$, and a fraction $kLI^2/2$ is useful for deflection, k being the deflection coil efficiency, which is always smaller than 1. It follows that

$$\frac{kLI^2}{2} = \frac{E a^2 \sin^2 \phi}{0.72l}$$

and solving for I ,

$$I = a \sin \phi \left(\frac{E}{0.36kLl} \right)^{1/2}$$

This expression shows the effect of the various parameters determining the deflection current. Since economy dictates that deflection current should be kept to a *minimum*, the following conclusions result:

1. a , the inside radius of the tube neck, should be small, and if full deflection is difficult because of other parameters, a smaller tube neck could be a solution. This would be the case, for example, for wide deflection angles.

2. ϕ , the maximum deflection angle, depends on the tube. The current tendency is to increase ϕ to make the tubes shorter. Standard values have been 72, 90, and 110 degrees for the diagonal total angle.

3. E , acceleration voltage, has its known effect and has been discussed previously.

4. k , coil efficiency, should evidently be as large as possible, and a ferrite magnetic circuit helps in this respect.

5. l , the length of the coil, should also be as large as possible, but is limited as has been seen in Sec. 14-13. Incidentally, reducing a would also compel reduction of l , but the over-all result is still in the right direction.

6. L , the self-inductance of the coil, should be large. In particular, if the shape of the coil as defined by a and l is fixed, L is proportional to n^2 , where n is the number of turns. Therefore, l is inversely proportional to n .

For a given tube, then, a , ϕ , and E are fixed, and the deflection coefficient depends on the shape and inductance of the coil. The deflection coefficient increases with the length and number of turns of the coil.

It has already been seen, however, that length is limited, because of the risk that the electron beam would hit the tube neck and produce a corner shadow on the screen. The number of turns is also limited, as we will see presently.

14-15. Number of turns

During the rise of the deflection current sawtooth waveshape, deflection current starts from zero and increases linearly, building up a strong magnetic field (around the deflection coil) in which energy is stored. During retrace, all this energy must be quickly dissipated so that the next sweep period can start again from zero. If all the energy is not dissipated when the following period starts, the beginning of the sweep moves in the direction of scanning and part of the picture may be lost.

The time taken for the energy to be dissipated determines the retrace or flyback time of the coil, and it must be smaller than the blanking time allowed by the standard if no portion of the picture is to be lost.

It is remarkable that all modifications of the coil shape that improve the coil efficiency simultaneously reduce its retrace time. When the coil shape has been optimized, another possibility is offered by the fact that return time is minimum when the coil is at critical damping, for which

$$R = \left(\frac{L}{4C} \right)^{1/2}$$

where R is the damping resistance, L the coil inductance, and C the coil capacitance. Adding a suitable resistance across the coil will then minimize retrace time.

In addition, since the coil time constant is $(LC)^{1/2}$, retrace time may be reduced by minimizing C —for example, by a special type of winding. The improvement is not large and limited in any case.

14-16. Defocusing

The ideal homogeneous deflection field, needless to say, is never obtained. It could be graphically represented as illustrated in Fig. 14-19, where the magnetic lines of force are parallel and equidistant. The circle represents the cross section of the electron beam. All electrons in the beam are acted upon by the same field value and deflection is said to be homogeneous. On the screen, the spot will have the same circular size wherever it is observed.

In actuality, the above ideal conditions are rarely the case. It can easily be checked on a given receiver by suppressing vertical sweep, thus reducing

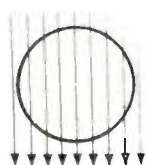


Figure 14-19. Uniform deflection field.

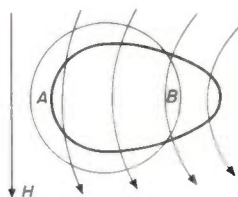


Figure 14-20. Beam cross section distortion by a concave field.

the raster to a single horizontal line in the middle of the screen. With focus correctly adjusted, the line will be thicker near the sides of the screen. Barring poor focusing, this phenomenon is due to deflection defocusing, which in turn is due to a nonhomogeneous deflection field. This field can be concave, in which case the lines of force curve in as shown in Fig. 14-20. Electrons at *A* in the beam are almost normally deflected, while electrons at *B* are in a more intense field and undergo a larger deflection. The circular cross section of the beam is distorted and drawn out along the direction of deflection.

The deflection field can also be convex, as shown in Fig. 14-21. The distortion is opposite to that of the preceding field. Electrons at *B* are deflected less than electrons at *A*, and the circular cross section of the beam is distorted and squeezed along the direction of deflection.

In both of the above cases, since the lines of force are nonlinear, any fraction F of the field, as shown in Fig. 14-22, can be split into a component H along the desired direction of field and a component N along the perpendicular direction. According to whether the field is concave or convex, this distortion component N will be positive or negative and will distort the beam cross section in opposite ways along the direction perpendicular to deflection.

It is important, then, to obtain as homogeneous a deflection field as possible. Several means can help to obtain this result. Common techniques are special profile polar pieces, multisection coils, and variable turn density coils.

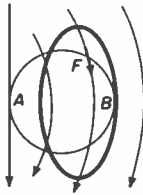


Figure 14-21. Beam cross section distortion by a convex field.

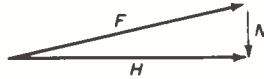


Figure 14-22. Components of a curved field.

14-17. Coaxial coils

The only useful deflection field is inside the tube itself, and the ratio of this useful field to the total field is the coil efficiency. It can be maximized by a coil shape such as shown in Fig. 14-23. The two coils are connected in such

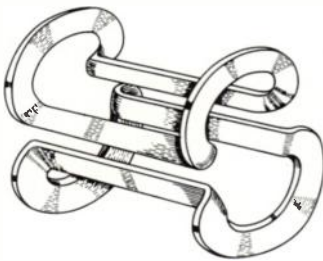


Figure 14-23. Pair of coaxial deflection coils.

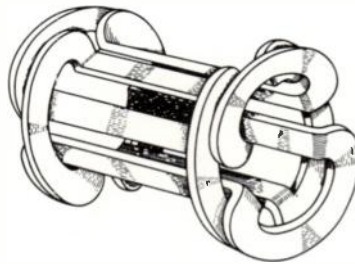


Figure 14-24. Coaxial deflection yoke.

a way that their fields add by either parallel connection or, more commonly, series connection.

To obtain full scanning of the tube, two such pairs are combined at right angles to each other as shown in Fig. 14-24, the longer pair being used for horizontal scanning where better deflection sensitivity is useful.

Considering only one pair of coils, say for horizontal deflection, the straight parts are the only useful flux-producing portions. When cut at half length, the cross section of the straight parts is distributed around the tube neck as shown in Fig. 14-25. The spacing e between two coils of a pair controls the homogeneity of the field. If e is too small, as shown in Fig. 14-26, a concave field results. If e is too large, as shown in Fig. 14-27, a convex field is obtained. The exact spacing depends on the method of winding the coils. It is generally on the order of a few millimeters.

Another method is to use cosinusoidal or multisection coils. In these coils, the winding is made in such a way that the number of turns distributed

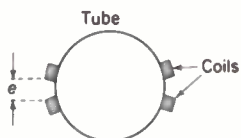


Figure 14-25. Active sections of the coils around the neck.



Figure 14-26. Production of a concave field.



Figure 14-27. Production of a convex field.

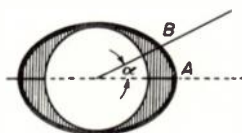


Figure 14-28. Cosine distribution of turns.



Figure 14-29. Cosine wound deflection coil.

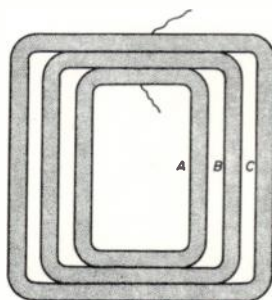


Figure 14-30. Section wound deflection coil.



Figure 14-31. Section wound deflection coil after forming.

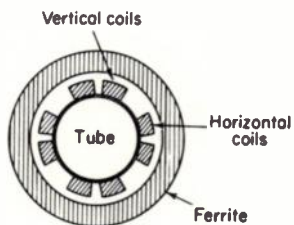


Figure 14-32. Addition of a magnetic ring circuit.

between A and B is proportional to $\cos \alpha$, as shown in Fig. 14-28. Thus the straight parts of the coils are very wide and of variable thickness, and almost wrap around the entire surface of the tube neck. Such a pair of coils is shown in Fig. 14-29. Winding such coils is not too easy and an approximation is commonly used to obtain the desired flux distribution. Each coil is wound in several sections, as shown in Fig. 14-30. Starting from center, each section has a longer active straight part, more turns and greater thickness than the preceding section, and a good approximation to continuous turn distribution can be obtained. Once bent to final shape, a pair of multisection coils would appear as shown in Fig. 14-31.

Most deflection yokes have a ring of high-permeability magnetic material around the coils, as shown in Fig. 14-32, to increase coil efficiency by concentrating the magnetic field inside the tube. Ferrites are commonly used for the rings enclosing the coil as shown.

Still better efficiency can be obtained with the special shape shown in Fig. 14-33. The ferrite ring just fits around the neck of the tube, and the coils are located within the inside notches.

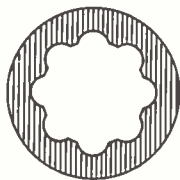


Figure 14-33. Notched magnetic ring.

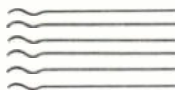


Figure 14-34. Parasitic vertical deflection.

14-18. Parasitic oscillations

The two pairs of coils, vertical and horizontal, are distributed around the neck and are never rigorously orthogonal to each other. There is a small but unavoidable magnetic coupling between them as a result. In addition, a capacitive coupling cannot be avoided.

During horizontal flyback, a high-amplitude transient appears in the line coils, which will be coupled to the vertical coils. If the induced voltage is large enough, it will shock-excite the vertical coils and start them ringing on their resonant frequency, thus creating a spurious vertical deflection in the form of a damped sinusoidal oscillation. Since the transient appears during horizontal retrace, ringing will occur immediately after, that is, at the left side of the picture where the resulting spurious vertical deflection will cause the raster lines to appear wiggly, as shown in Fig. 14-34. This defect can be eliminated by suitable damping of the vertical coils, for example, by shunting them, as in Fig. 14-35, with low value resistances. Sometimes damping of the

vertical coils is achieved with a capacitance of such a value that it behaves practically as a short circuit at line frequency, as shown in Fig. 14-36.

Another defect of the same type occurs because the two coils making up the horizontal deflection coil do not have balanced capacitances. In this case, the flyback transient shock-excites a differential oscillation between the two coils, thus superposing a damped sine wave velocity modulation to the

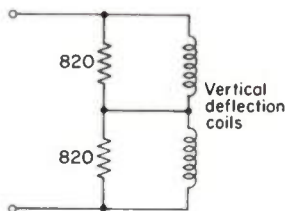


Figure 14-35. Resistance damping of the vertical coils.

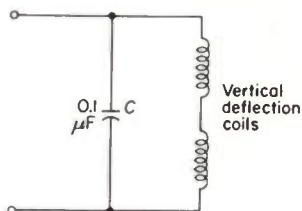


Figure 14-36. Shunting the vertical coils by a large value capacitance C .

normal horizontal sweep. The result is several dark bands at the left of the screen but the lines do not wiggle. The remedy consists in balancing of the capacitances and is generally obtained by connecting a small capacitance, 40 to 100 $\mu\mu\text{f}$, across the top horizontal coil as shown in Fig. 14-37. Because of the retrace transient, this capacitance should be well insulated.

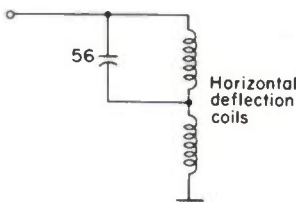


Figure 14-37. Balancing the parasitic capacitances across the horizontal coils.

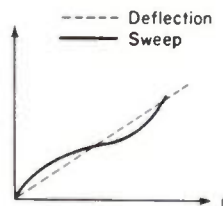


Figure 14-38. Characteristic S distortion.

14-19. Distortions

A deflection angle linear with time, as previously shown, would produce a spot displacement on the screen, or sweep, which would be linear with time only if the screen were a spherical surface having as its center the deflection center. However, this is never the case. Since the screens are flat or almost flat, the sweep speed increases when the spot moves away from the center of the screen. For linear deflection, then, the sweep would take the characteristic S form shown in Fig. 14-38, and produce what is called S distortion. This

distortion occurs in both vertical and horizontal deflection and a theoretically rectangular raster would appear on the screen in the pin-cushion form shown in Fig. 14-39.

The defect could be corrected if the deflection coils produced a field more intense at the center of the tube. In this case, deflection would not be linear, but would follow the shape shown in Fig. 14-40(a). Actually, such a shape is almost always obtained, at least to some extent. Its curvature could be adjusted so that it is exactly opposite to that of the S distortion of Fig. 14-40(b). Total results would then be the linear sweep of Fig. 14-40(c). In fact, S distortion compensation is by no means critical and the same deflection yoke can suit a number of tubes of similar types of screen.



Figure 14-39. Pin-cushion picture distortion.

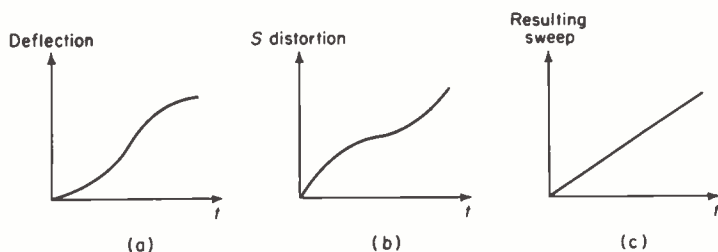


Figure 14-40. Compensation of S distortion.

Since the deflection field is nonlinear, there will be some amount of focus distortion which will add to the focus distortion due to a flat or almost flat screen. For both distortions, the remedy consists in using a thin electron beam to minimize the defect. Moreover, in practice, focus is adjusted for best results halfway between the center and periphery of the screen. In this way, focus at the center is slightly below and at the sides slightly above optimum, and the apparent over-all focus is improved.

Finally, because S distortion is more troublesome for horizontal deflection, it might be pointed out that it is easily corrected electrically without additional components for line scan. This will be shown in Chap. 17, which is devoted to horizontal deflection.

Problems

1. Explain the desirability of a linear deflection current.
2. A deflection coil is made up of an inductance L and a resistance R in series. It has a shunt capacitor C and a shunt resistor R_1 . Draw the waveforms of voltage across and current through each component when the current through L is a linear sawtooth.

3. A circuit is made up of inductance L , capacitance C , and resistance R in parallel. A rectangular voltage wave is applied to it. Draw the waveforms of current through the three components.

4. A circuit is made up of capacitance C and resistance R in parallel. A sawtooth current flows through R . Draw the waveforms of current and voltage for the whole circuit.

5. A circuit is made up of capacitance C , inductance L , and resistance R in parallel. A sawtooth of current flows through L . Draw the waveforms of the applied voltage and total current.

6. A deflection coil has a self-inductance of 0.3 henry and a resistance of 450 ohms. Plot nonlinearity for sweep durations from 1 to 200 μsec .

7. A deflection coil has a self-inductance of 1.2 henrys and a resistance of 1500 ohms. A 60-cps sawtooth current of 50 ma peak-to-peak flows through it. Draw the waveform to scale of the voltage across the coil and indicate voltage values.

8. Explain why there is so much difference in the problems of vertical and horizontal deflection.

9. Investigate the effect of a large capacitor shunted across the vertical deflection coils.

10. A deflection coil has a self-inductance of 0.1 henry and a resistance of 65 ohms. A 15,750-cps sawtooth current of 2 amp peak-to-peak flows through it. Retrace is linear and has a duration equal to 14 per cent of the period. Evaluate the peak voltage across the coil during sweep and during retrace.

11. Investigate the requirements for a tube to be used as power amplifier for horizontal scanning.

12. Investigate the requirements for a tube to be used as power amplifier for vertical scanning.

13. A high-impedance deflection coil has inductance L and resistance R . It is coupled through a large-value capacitance C to the plate of a tube having for load and inductance L_1 in series with resistance R_1 . The R/L ratio of the deflection coil introduces nonlinearity, and this can be corrected by proper choice of R_1 and L_1 . Establish the suitable relation.

14. Design a complete vertical and horizontal electrical centering system for high-impedance coils and suggest suitable component values.

15. Design a complete vertical and horizontal electrical centering system for low-impedance coils and suggest suitable component values.

16. Explain the role of the peaking adjustment.

17. Could peaking be used for vertical deflection?

18. Design a blocking relaxation oscillator including peaking adjustment. Give component values for horizontal frequency.

19. It is suggested that a deflection yoke could be designed along the lines of an electric machine multipolar stator. Discuss this suggestion.

20. Evaluate the field intensity necessary to scan a 110-degree tube having an acceleration voltage of 15,000 v and equivalent deflection coil length of 5 cm.

21. A magnetic field of 35 gauss is necessary to scan a 72-degree picture tube having a high voltage of 12,000 v. Evaluate the equivalent length of the deflection coils.

22. A 90-degree tube has a high voltage of 14,000 v and deflection coils of 45-mm equivalent length. The deflection current has a value of 1.5 amp peak-to-peak. Calculate the number of turns in the deflection coils.

23. A 110-degree tube has a neck with an inside diameter of 22 mm. The electron beam radius will be assumed to be 1 mm. Evaluate the allowable equivalent length of the deflection coils. The high voltage is 16,000 v. (a) Calculate the deflection field. (b) Calculate the energy accumulated in the field.

24. A 90-degree tube has a high voltage of 12,000 v, and coils of 45-mm equivalent length with an efficiency of 58 per cent, and an inductance of 80 millihenrys. The neck of the tube has an internal radius of 18 mm. Calculate the deflection current.

25. Establish a relation giving the deflection current as a function of the deflection coil characteristics and accelerating voltage only.

26. Find (in the technical and periodical literature) inductance and resistance values for the vertical and horizontal coils of a deflection yoke. Calculate in both cases the time constant and the critical damping resistance, assuming or measuring capacitances.

27. Discuss the advantages and disadvantages of coaxial deflection coils.

28. Justify the use of a ferrite ring around a deflection yoke.

29. A flat-face, 21-in. tube has a distance of 12 in. between the center of the screen and the deflection center. Plot S distortion.

Chapter 15

SAWTOOTH GENERATORS

Electronic sawtooth generators commonly rely on the charge and discharge of a capacitor to produce the required waveform, although other means are available. Such sawtooth generators belong to the category of relaxation oscillators, common examples of which are the multivibrator and the blocking oscillator.

Sine wave oscillators are also used for sawtooth generation. The sinusoidal signal they produce is distorted to obtain a suitable waveform.

If the image reproduced on the screen of the picture tube is to be recognizable, its elements should occur in correct order, that is, in the same order that they were scanned at the transmitter. This means that both vertical and horizontal scannings should be in synchronism at both the receiver and transmitter. Synchronization is obtained by applying sync pulses to the sawtooth generators. The pulses trigger the generators and ensure that the scanning periods start at the same time at both receiver and transmitter.

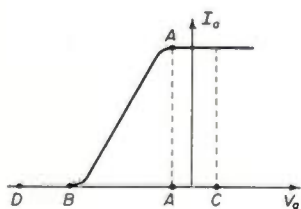


Figure 15-1. Plate current limits at cutoff B and saturation A .

15-1. Square wave

A tube has two natural limits beyond which there is no current variation. They are, as shown in Fig. 15-1, the saturation at A and the cutoff at B .

At saturation point A , current reaches a maximum value and cannot increase any further, with plate voltage at a minimum. At cutoff point B , current reaches zero and cannot decrease

further, with plate voltage at a maximum and equal to the applied voltage.

When a large-amplitude signal is applied to the grid, and the grid voltage swings rapidly between points *D* and *C* indicated on the diagram, the tube is swept past both the above limits. As long as the grid voltage stays at *C* or, more precisely, above *A*, the tube saturates at maximum current. As long as the grid voltage stays at *D* or, more precisely, below *B*, the tube bottoms at zero current.

This indicates that the exact *shape* of the grid drive voltage has no importance below *B* or above *A*. If grid voltage V_g varies rapidly between *A* and *B*, as indicated in Fig. 15-2(a), the plate current I_a will switch rapidly between zero and saturation as shown in (b). At the same time, plate voltage V_p will swing rapidly between maximum and minimum as shown in (c). Notice that the waveforms of plate current and plate voltage have identical shapes but are opposite in phase.

Such a waveform is generally called a *rectangular wave*. In the case of Fig. 15-2, positive and negative alternations are equal and this special case is properly called a *square wave*, or sometimes a 50/50 (per cent on-off) rectangular wave.

The same principle may be used to produce rectangular waves having, for example, a 90/10 ratio as illustrated in Fig. 15-3. These waveforms are

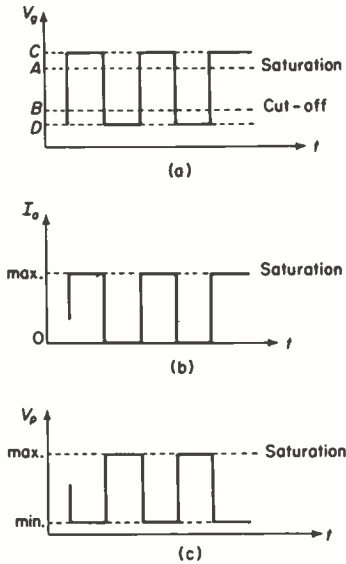


Figure 15-2. Double clipping of a large amplitude signal.

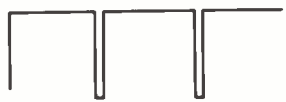


Figure 15-3. Nonsymmetrical rectangular wave.



Figure 15-4. Integration of the rectangular wave gives a sawtooth.

extensively used in television, because a simple integration transforms the rectangular wave into a fast-return sawtooth, as shown in Fig. 15-4, which is typical of sweep waveforms.

15-2. Tube switch

Notice that in the above type of operation, no use is made of the actual characteristic of the tube. In fact, the tube is used as a switch—closed at *A*, open at *B*.

At saturation A , the current through the tube is maximum, the voltage across the tube (between anode and cathode) is minimum, so that the tube appears practically as a zero resistance or at least as a very low resistance, and the switch is closed.

At cutoff B , the current through the tube is zero, the voltage across the tube is maximum, so that the tube appears practically as an infinite resistance, and the switch is open.

This peculiar type of operation may be represented symbolically, as shown

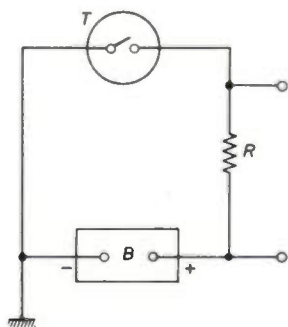


Figure 15-5. Using a tube as a switch.

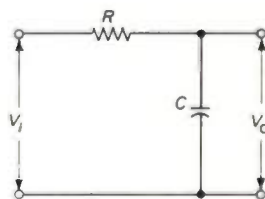


Figure 15-6. RC type integrator circuit.

in Fig. 15-5, where the tube has been replaced by an equivalent switch, controlled by grid voltage.

The equivalent circuit contains the tube switch T , the plate voltage supply B , and the load resistor R across which is taken the output waveform. As such, the circuit will provide a rectangular signal. To obtain a sawtooth, all that is needed is an integration circuit, the simpler version of which is the RC type shown in Fig. 15-6. For a suitable RC time constant it has been shown previously that this circuit provides an output signal V_o which is the integral of the input signal V_i .

Since the integral of a rectangular wave is a sawtooth wave, the combination of the two circuits of Figs. 15-5 and 15-6 will constitute a sawtooth generator. Moreover, a resistor may be eliminated in the process, because the same resistor R may be used simultaneously as the load resistor and as the integrating resistor. The combined equivalent is represented in Fig. 15-7.

For practical purposes, then, the addition of a simple capacitor to a rectangular wave generator transforms it into a sawtooth generator.

Since this circuit is so fundamental to the concept of sawtooth generators, a closer investigation is in order.

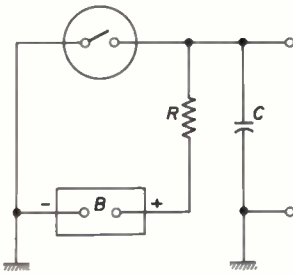


Figure 15-7. Combining the tube switch and the integrator circuits.

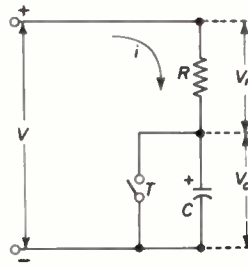


Figure 15-8. Equivalent circuit.

15-3. Capacitance charge

The equivalent diagram of Fig. 15-8 shows a voltage V applied to an RC circuit. V_R is the voltage across resistor R , and V_C the voltage across capacitor C . For the time being, switch T is open as shown. V_C is zero.

When V is applied, the capacitor has no charge and all the voltage appears across R . A current $i = V/R$ flows through the circuit as shown and begins charging up the capacitor. As soon as some charge has been accumulated, a small voltage V_C appears across C . Now evidently, as shown on the diagram,

$$V = V_C + V_R$$

or

$$V_R = V - V_C$$

Consequently, as soon as V_C appears across C , V_R decreases and the charge current becomes

$$i = \frac{V - V_C}{R}$$

that is, smaller than before, and the rate of charge decreases. Nevertheless, C keeps on charging up, V_C increases, V_R decreases, i decreases, and the rate of charge decreases even more.

To sum up, the rate of charge decreases when the charge increases, so that the voltage V_C across C will not vary linearly. Actually, it varies exponentially, as will be shown shortly.

Evidently, more time will be needed to reach a given voltage V_C if C is larger, because it will take longer to charge up. Also, if R is made larger, the charge current will decrease and it will take longer to reach a given charge voltage V_C .

Both R and C , then, determine the time taken to reach a given voltage V_C , whence of course the name *time constant* given to the product RC . The equation

$$T = RC$$

gives the time constant T in seconds when C is in farads and R in ohms.

15-4. Exponential charge

Let i be the charge current of Fig. 15-8. The voltage across the resistance is

$$V_R = iR$$

and the voltage across the capacitance is

$$V_C = \frac{1}{C} \int i \, dt$$

In the circuit,

$$V = V_R + V_C$$

or

$$V = iR + \frac{1}{C} \int i \, dt$$

Differentiate and divide by R :

$$\frac{di}{dt} + \frac{i}{RC} = 0$$

or

$$\frac{di}{dt} + \frac{1}{T} i = 0$$

Separate the variables:

$$\frac{di}{i} = -\frac{dt}{T}$$

Take the integral:

$$\int \frac{1}{i} \, di = -\int \frac{1}{T} \, dt + K$$

K being the integration constant, to be determined later. Integrate:

$$\log_e i = -\frac{t}{T} + K$$

Take the antilogarithm:

$$i = e^{-t/T} \cdot e^K$$

e is the base of neperian logarithms and is equal to 2.718. . . .

Let $e^K = k$. Then

$$i = ke^{-t/T}$$

To determine the integration constant k , it is only necessary to know one boundary condition. This is easily provided by the fact that at the time of

applying the voltage, at $t = 0$, the capacitance carries no charge and the instantaneous current i_0 flowing is $i_0 = V/R$. But at $t = 0$, the equation for i above becomes

$$i_0 = ke^0 = k$$

so that $k = V/R$. Therefore,

$$i = \frac{V}{R} e^{-t/T}$$

The voltage across the resistance is

$$V_R = iR = Ve^{-t/T}$$

and the voltage across the capacitance is

$$V_C = V - V_R = V(1 - e^{-t/T})$$

15-5. Charging curve

The equation $V_C = V(1 - e^{-t/T})$ then gives the voltage V_C across a capacitor C after a time t , the voltage V being applied through a resistor R such that $T = RC$; e , of course, is equal to 2.718. . .

The charge equation can be normalized as

$$\frac{V_C}{V} = 1 - e^{-t/T}$$

and a universal charging curve can be drawn, as shown in Fig. 15-9, using V_C/V for ordinates and t/T for abscissas. This curve is the exponential rise of charge in capacitor C .

It may be seen that the charge is strongly nonlinear. However, if it is kept well below the maximum, that is, if only the beginning of the curve is used, then a fair approximation of a straight line results. This point is more clearly seen in Fig. 15-10, which is merely an enlarged version of the beginning of the curve of Fig. 15-9. In practice, the charge is considered as linear if it is limited to $V_C/V = 0.1$, or approximately $t/T = 0.1$, so that

$$RC = 10t$$

This can be shown more precisely. In $V_C/V = 1 - e^{-t/T}$, the exponential second term can be expanded into a series, giving

$$\frac{V_C}{V} = \frac{t}{T} - \frac{1}{2}\left(\frac{t}{T}\right)^2 + \frac{1}{6}\left(\frac{t}{T}\right)^3 - \dots$$

If t/T is small compared to 1, all terms containing powers of t/T can be neglected compared with t/T , so that

$$\frac{V_C}{V} = \frac{t}{T}$$

which evidently indicates a linear variation of V_C as a function of time:

$$V_C = kt$$

where $k = V/T$ is a constant.

Referring back to Fig. 15-9, the curve tends asymptotically toward 1 but theoretically never reaches it, so that the charge is never completed. For

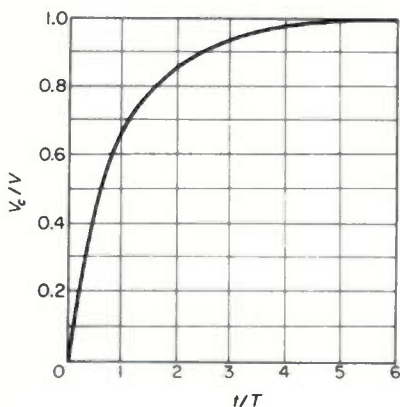


Figure 15-9. Exponential charge of a capacitance.

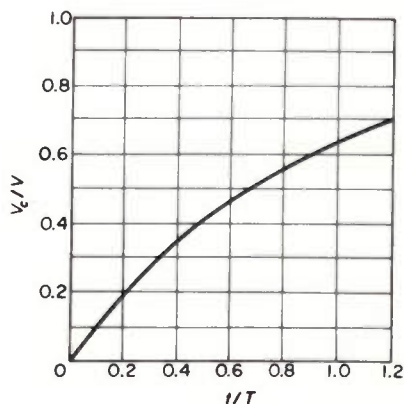


Figure 15-10. Expanded start of the exponential charge.

$t = T$, the charge voltage reaches 63 per cent of the applied voltage. Similarly, for

$$t = 2T, \quad 86\%$$

$$t = 3T, \quad 95\%$$

$$t = 4T, \quad 98\%$$

For most practical purposes, the charge is considered as complete after a time $t = 5T$.

15-6. Amplitude control

In practical cases, the rectangular wave to be integrated has a given frequency. Stated another way, the resulting sawtooth has a fixed duration. To adjust the amplitude of the sawtooth, it is then sufficient to change the time constant of the integrator circuit.

As shown above, when the charge voltage V_C is limited to 10 per cent of the applied voltage V , the charge may be considered as linear. In this case, the charge voltage reached after a given time is proportional to the time constant. If the time constant is large, the capacitor has accumulated relatively little charge in time t , as shown in curve *A*, Fig. 15-11. The resulting charge

voltage, that is, the amplitude of the sawtooth, is V_1 . If the time constant is short, the charge voltage reaches a larger value V_2 after time t , as shown by curve B , and the resulting sawtooth has a larger amplitude.

Controlling the time constant, therefore, provides an easy way of controlling the output sawtooth amplitude.

In actual practice, the integrator time constant will be varied by changing the resistance value. It is convenient in this case to replace the resistor by a potentiometer, called the *amplitude control*.

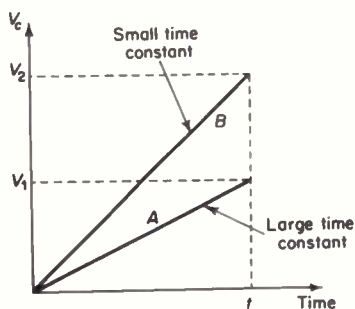


Figure 15-11. Effect of time constant on amplitude.

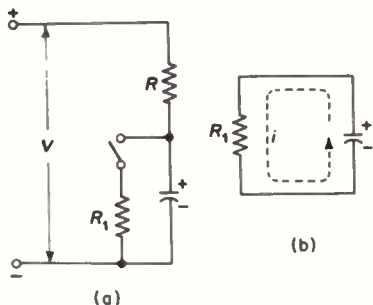


Figure 15-12. Equivalent circuits for charge and discharge.

15-7. Discharge curve

When the switch T is closed in the simplified diagram shown in Fig. 15-8, the capacitor is short-circuited and instantaneously discharged.

Actually, the tube T does not have zero resistance when conducting, but a small resistance R_1 . A more exact equivalent circuit, therefore, would be represented by Fig. 15-12(a). When the switch is closed at $t = 0$, the circuit reduces to Fig. 15-12(b), so that the capacitor simply discharges its charge through R_1 , with a time constant

$$T_1 = R_1 C$$

A calculation analogous to the one performed for the charge gives

$$V_C = V e^{-t/T_1}$$

with V_C = voltage across capacitance at time t ,

V = voltage across capacitance at time $t = 0$,

$e = 2.718$,

t = time since switch was closed at $t = 0$,

$T_1 = R_1 C$ = time constant.

The discharge is also exponential. Since R_1 is much smaller than R , the discharge time will be much shorter than the charging time. In fact, for

almost linear output, charge and discharge durations are in the ratio of R/R_1 , approximately.

It may then be seen that integration of a rectangular wave by an RC circuit will not provide a linear sawtooth, but rather the exponential fall and exponential rise sawtooth of Fig. 15-13. For all practical purposes, this sawtooth will be a close enough approximation of linear waveform when

$$RC = 10 \theta$$

where θ is the sawtooth period.

The amplitude of the resulting sawtooth will then be 10 per cent of the applied voltage. Voltages of 100 to 300 v are usually available in a receiver to produce the rectangular wave, so that it is apparent that good sawtooth waves of from 10 to 30 v amplitude will be obtained easily.



Figure 15-13. An RC integrator produces an exponential sawtooth.

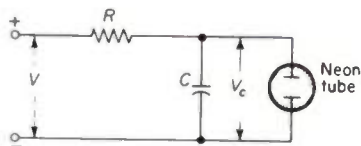


Figure 15-14. Gas diode elementary relaxator.

15-8. Neon tube relaxator

A relaxator or relaxation oscillator is an oscillator that produces a non-sinusoidal waveform such as a rectangular or sawtooth wave.

An elementary sawtooth generator using a neon tube is illustrated in Fig. 15-14. A neon tube simply consists of two electrodes in a glass envelope filled with low-pressure neon gas. Gas in a tube envelope is symbolically indicated by crosshatching or by a black dot.

Neon tubes have a peculiar behavior, which is the basis for their use in a number of circuits. When a small voltage is applied across a neon tube, no current flows. The internal resistance of the tube is very high—practically infinite, and the tube behaves as an open circuit.

Then the voltage across the neon tube is increased. At first, nothing happens. When the voltage reaches a critical value, say 100 v, called the *ionization or firing voltage*, the tube fires. That is, the neon suddenly becomes ionized and produces a bright orange glow, the tube internal resistance falls to a very low value, almost zero, and the tube behaves as a short circuit. In fact, the current flowing through the tube would destroy it, and some current-limiting arrangement must be provided.

Now if the voltage is reduced, a remarkable thing happens: the gas stays ionized and the tube conducts down to a voltage, called the *extinction*

voltage, of say 70 v. Below 70 v, the glow disappears and the tube again behaves as an open circuit.

There is then a range of voltage from 100 to 70 v where the tube conducts if it has been ionized previously. The phenomenon is nonsymmetrical—the tube fires at 100 v, but turns off at 70 v.

It is now an easy matter to understand the operation of the circuit shown in Fig. 15-14. Capacitance C being discharged, assume a voltage V of 200 v is applied. C charges up with time constant RC . As long as the voltage V_C across C is below the firing voltage, say 100 v, of the neon tube, nothing happens and the capacitance continues to charge up. When its charge reaches

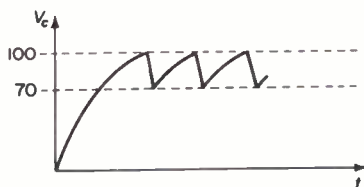


Figure 15-15. Operation of the gas diode relaxator.

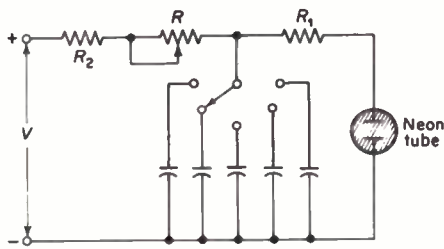


Figure 15-16. Adjustment of relaxation frequency.

100 v, the neon tube fires and behaves as a short across C , discharging it rapidly until the voltage V_C drops below the extinction voltage, say 70 v. Then the tube turns off and its internal resistance becomes infinite.

Capacitor C charges up again through R , this time, however, starting from 70 v, until it reaches 100 v, and the cycle repeats itself. The resulting waveform is illustrated in Fig. 15-15 and is a sawtooth wave of

$$100 - 70 = 30 \text{ v}$$

amplitude, and period essentially determined by the time constant $T = RC$.

A convenient way to control the recurring frequency of the sawtooth is to change C or R . To obtain a wide variation, the capacitance is generally switched, and the resistance is continuously variable. The resulting variable-frequency neon tube relaxator appears in Fig. 15-16. Resistor R_1 , in series with the tube, limits the discharge current to a safe maximum value. Resistor R_2 similarly limits the current flowing through R when the potentiometer is set at a low resistance value. This type of protection circuit is frequently used and will appear in many relaxator diagrams.

Since neon tubes are inexpensive, a neon relaxator is an economical sawtooth generator. However, it has several drawbacks, chiefly frequency instability, especially as a function of temperature, and difficulty of synchronization. For these reasons it is never used in television.

15-9. Thyatron

A thyatron is also a gas tube, that is, its glass envelope is filled with a low-pressure gas and it has three electrodes, including a heated cathode. Actually, then, it is a gas triode, with cathode, grid, and anode.

Considering only cathode and anode, the tube would behave as the preceding gas diode, except that the heated cathode makes for a much lower firing voltage. It would be used as a relaxator with a diagram similar to the diagram given for the neon tube.

The grid, however, greatly modifies the behavior of the thyatron. As in an ordinary triode, it has better control of the electron flow than the anode, and a voltage variation of 1 v on the grid is just as efficient as a voltage variation of K v on the anode. K , which would be the tube gain for a triode, is called the *control ratio* for a thyatron. A value of 30 for K is typical.

Assume now that a thyatron with a control ratio of 30 has a firing voltage of 100 v and that its plate is at 80 v, so that the tube is switched off. If the grid voltage is now increased by 1 v, this is equivalent to increasing the anode voltage by 30 v, so that the equivalent plate voltage jumps to

$$80 + 30 = 110 \text{ v}$$

which is above ionization voltage, and the tube fires instantaneously.

The remarkable point is that this is also a one-way phenomenon. Once the ionization is established and the tube conducts, the grid loses control.

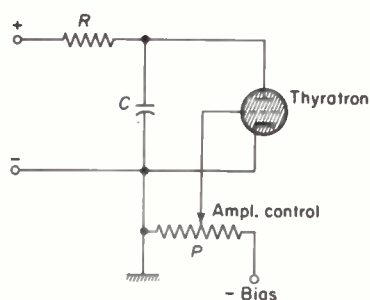


Figure 15-17. Gas triode relaxator.

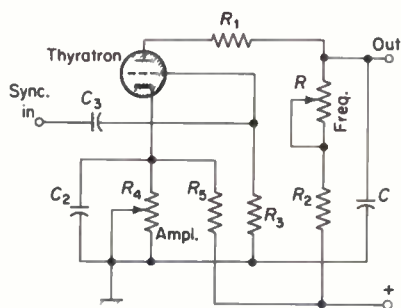


Figure 15-18. Practical thyatron relaxator.

No matter how negative it is made, the tube will not switch off. The only way to turn it off is to reduce the anode voltage below the extinction level.

Since plate voltage corresponding to extinction is fixed and since plate voltage corresponding to firing depends on the grid voltage, the voltage difference between the two levels is a function of grid voltage. If the thyatron is used as a sawtooth generator, as shown in Fig. 15-17, instead of a neon tube, adjusting the grid bias will control the amplitude of the sawtooth wave produced across the capacitance.

As for an ordinary triode again, bias may be obtained by making the cathode positive through a cathode bias RC circuit. A more elaborate version of the thyatron oscillator then would be as represented in Fig. 15-18. Cathode bias is obtained through R_5 - R_4 - C_2 and is made adjustable in order to control amplitude. The charge circuit is in the anode, C being the capacitor, and potentiometer R plus safety resistor R_2 being the total charging resistance. In the same manner as for the neon tube, R_1 limits the thyatron discharge current to a safe value.

Since bias is obtained in the cathode, the grid normally should be grounded. However, it offers a very convenient point to apply synchronization signals because of its good controlling characteristics. A coupling circuit R_3 - C_3 is used to inject sync pulses on the grid.

The cathode bias components, R_5 and C_2 , are sometimes omitted.

15-10. Practical circuits

Figure 15-19 represents a typical horizontal thyatron sawtooth generator for operation at 15,750 cps. The cathode potentiometer adjusts amplitude and the anode potentiometer controls frequency; this is true at least when the

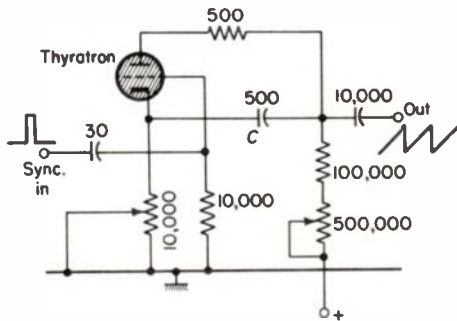


Figure 15-19. Typical horizontal thyatron relaxator.

thyatron is synchronized. When it oscillates freely, both adjustments are somewhat interdependent.

Horizontal sync is applied through the grid coupling circuit, whose components' values have been chosen to provide differentiation and to apply sharp pulses to the grid so as to obtain precise synchronization. Input sync pulses should be positive-going as indicated.

A similar circuit, as shown in Fig. 15-20, can be used for the vertical relaxator. Note that the component values are different from Fig. 15-19 and suited to 60 cps. In addition synchronization pulses are applied to the grid through an integrator circuit as is fit for the long vertical sync pulse.

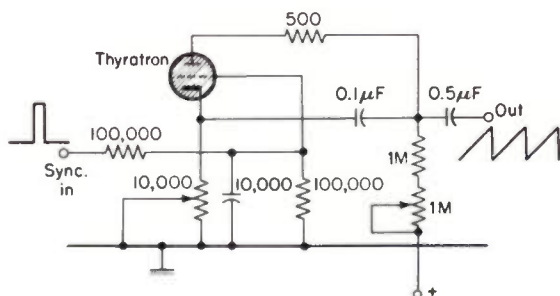


Figure 15-20. Typical vertical thyatron relaxator.

15-11. Synchronization

Since synchronizing processes are identical for the thyatron as well as for the other relaxators, it is interesting to investigate synchronization more closely. Referring to the horizontal sawtooth oscillator shown in Fig. 15-19, assume the thyatron has a control ratio of 30. The bias has been adjusted so that the grid is at -3 v with respect to the cathode, and the firing anode voltage is 100 v.

Anode supply voltage is applied and capacitor C charges up, so that anode voltage V_a rises toward 100 v, as shown in Fig. 15-21. Before it reaches 100 v, say at time t_1 when it is at 70 v, a 2-v positive sync pulse is applied to the grid. For this value, the firing voltage of the thyatron is reduced to

$$100 - (2 \times 30) = 40 \text{ v}$$

because of the grid control ratio of 30. Since the anode at that instant is at 70 v, that is, above 40 v, the thyatron fires instantaneously and discharges the capacitor C . The process then repeats itself.

It can be seen that the sync pulse triggers the thyatron and determines the exact time at which the sawtooth ends. Notice that for synchronization to be effective, the free-running relaxation period should be longer than the synchronized relaxation period. In other words, the relaxator should oscillate on a free-running frequency lower than the recurrent frequency of the sync pulses. Also, since sync pulses trigger the discharge before normal charge voltage is attained, the amplitude of the synchronized sawtooth will be smaller than the amplitude of the free-running sawtooth. Finally, when sync pulses occur they should have sufficient amplitude to trigger the thyatron.

One might think that there is no maximum limit to sync pulse amplitude. However, sync pulses too large would drive the grid positive, with resulting grid current and various side effects, and they could produce excessive synchronization with consequent waveform distortion.

In fact, a thyratron is easily synchronized with a small pulse amplitude, say a fraction of a volt, and it is always possible to reduce sync signal amplitude—for example, by adjusting the values of the components in the grid circuit to obtain better differentiation or integration.

The thyratron will be triggered, then, only if the sync pulse occurs near the end of the sawtooth rise, or inversely, if the free-running frequency is not too far below pulse recurrence frequency. Therefore, the relaxator will lock on the pulses only if its free-running frequency is within pulling range of the pulse frequency.

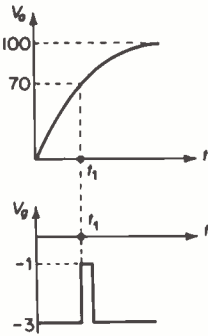


Figure 15-21. Applying the synchronization pulse.

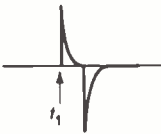


Figure 15-22. Differentiated sync pulse.

15-12. Pulse waveform

It is the front side of the pulse, the positive rise, which trips the relaxator. Once ionization is established, the grid loses control so that the shape of the sync pulse from then on is without importance.

Precision of synchronization then depends on the front rise of the pulse, which should be as steep and abrupt as possible. A differentiated pulse as shown in Fig. 15-22 is very suitable in this respect. The front rise is very sharp and ensures accurate triggering at time t_1 . The remaining shape of the pulse is unimportant to synchronization.

Figure 15-23 illustrates what happens with a distorted sync pulse, for example, when parasitic shunt capacitances limit the rise time. The time scale is expanded to make the time differences evident. Assume triggering level to be V_1 . The top pulse will trip the relaxator at time t_1 . The next pulse below, although arriving exactly at the correct time, has a reduced amplitude. Now V_1 corresponds to time t_2 , and a synchronizing error

$$\tau = t_2 - t_1$$

will result. It may be concluded that if the rise time of the sync pulse is not short, all sync pulses should have the same shape and the same amplitude.

Thyratrons are never used in television nowadays, although they are still employed in the less expensive oscilloscopes. In television, the common relaxation oscillators are the multivibrator and the blocking oscillator.

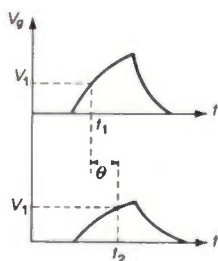


Figure 15-23. Synchronization error with a distorted sync pulse.

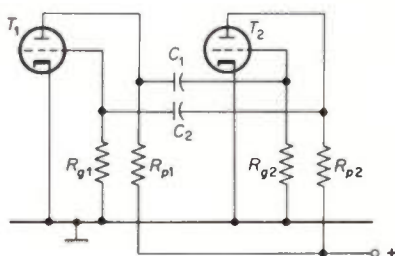


Figure 15-24. Basic circuit of a multivibrator.

15-13. Multivibrator

A multivibrator consists of two *RC*-coupled amplifying stages, in cascade, as shown in Fig. 15-24, the output of the amplifier being connected back to the input. Each tube introduces a phase shift of 180 degrees, so that the total phase shift for two tubes is 360 degrees. The amplifier output signal is consequently in phase with the input signal and the feedback is positive.

Any accidental variation causing, say, a decrease in plate current of tube 1 will result in an increase in plate voltage of tube 1. This positive voltage variation will be applied to the grid of tube 2 through the *RC* coupling, thus increasing the plate current and decreasing the plate voltage of tube 2. This negative voltage variation will be applied to the grid of tube 1 through the *RC* coupling, thus further reducing the plate current of tube 1, and so on. The process is cumulative, and because of the high gain of the two tubes in cascade, the very small original disturbance is quickly amplified to the point where tube 1 has a minimum of plate current and a maximum of plate voltage, whereas tube 2 has a maximum of plate current and a minimum of plate voltage.

In this way, it is possible to reach cutoff for tube 1 and saturation for tube 2 very rapidly. Saturation being a rather ill-defined condition, since actually the current does not reach maximum past saturation but continues to increase slowly, it is the cutoff of tube 1 which sets a limit to the process.

When tube 1 reaches cutoff, there is no further variation of plate current of tube 1, which is zero, and little variation of plate current of tube 2, which has

some maximum value. A stable state has been reached in a very short time after the original disturbance, because of the very fast change in grid voltage.

However, the multivibrator is not going to stay indefinitely in this stable condition. The reason for this is that tube 1 is held to cutoff by the charge accumulated on capacitor C_2 , which keeps the grid negative. But C_2 can discharge through R_{g1} , the voltage supply source, and R_{p2} , with time constant

$$\theta_2 = C_2(R_{g1} + R_{p2})$$

The charge on C_2 thus leaks away, and after some time the decreasing negative voltage on grid 1 will cross cutoff. Tube 1 begins to pass a small current, and its plate voltage drops slightly. This negative voltage variation is applied to the grid of tube 2 through the RC coupling, reducing its plate current and increasing the plate voltage. This positive voltage variation is applied to the grid of tube 1 through the RC coupling, thus further increasing grid 1 voltage, and so on. As before, the process is cumulative and the large amplification results in a very fast reversal of conditions. After a short time, tube 2 reaches cutoff and tube 1 passes a maximum current. The situation is exactly symmetrical to the preceding one, and is again stable. However, this time tube 2 is held to cutoff by the charge accumulated on C_1 , which will discharge with time constant

$$\theta_1 = C_1(R_{g2} + R_{p1})$$

so that after some time the multivibrator will switch over again to its first state, tube 1 at cutoff and tube 2 conducting, and so on.

In conclusion, the multivibrator will swing periodically between two symmetrical states. In each state, one tube is cutoff and the other is heavily conducting. The switching or transition between the two equilibrium states is very fast.

The original accidental variation assumed previously is not necessary. At switching on, as soon as one tube begins to draw current, the process will start.

15-14. Relaxation period

If the two time constants θ_1 and θ_2 are equal and the two tubes are identical, the multivibrator is symmetrical and its two stable states have equal durations. The voltage variation on the anodes in this case is a square wave, as shown in Fig. 15-25, with opposite phases on the two anodes.

When the two time constants θ_1 and θ_2 are different, the multivibrator is nonsymmetrical and will produce a rectangular square wave as illustrated in Fig. 15-26, with a ratio of positive alternation to negative alternation durations approximately equal to θ_1/θ_2 . This value is also called the *mark-to-space* or *black-to-white* ratio.

In television, strongly nonsymmetrical waves are used, the mark-to-space ratio being at least on the order of 10.

Generally, grid resistances are much greater than plate resistances, and as a first approximation

$$\theta_1 = C_1 R_{g2} \quad \text{and} \quad \theta_2 = C_2 R_{g1}$$

The period of oscillation T of the multivibrator is then roughly given by

$$T = C_2 R_{g1} + C_1 R_{g2}$$

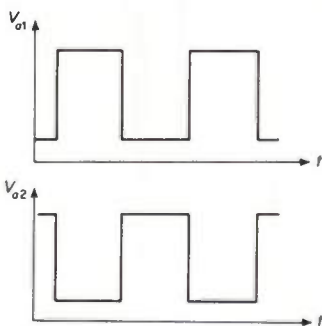


Figure 15-25. A symmetrical multivibrator produces a square wave.



Figure 15-26. A non-symmetrical multivibrator produces a rectangular wave.

In television, where one time constant, say θ_1 , must be at least 10 times larger than θ_2 , a crude approximation is

$$T = C_1 R_{g2}$$

so that the relaxation frequency is roughly

$$f = \frac{1}{C_1 R_{g2}}$$

This approximation is close enough for practical work, because in any case a frequency adjustment must be provided and covers a range which largely includes the calculation error.

15-15. Cathode-coupled multivibrator

In the multivibrator just described, two plate-to-grid couplings are used to ensure positive feedback and oscillation.

One of these couplings may be replaced, as shown in Fig. 15-27, by a cathode-to-cathode coupling. It may be shown that positive feedback is

obtained as follows: Assume a small positive voltage variation on grid 2. This will create a positive voltage variation on cathode 2, applied to cathode 1 through the RC coupling. An increase in voltage on cathode 1 will reduce current in anode 1, so that anode 1 voltage increases. This positive variation is applied through the RC coupling to grid 2, where it reinforces the original variation.

Thus, phase relations are correct for the effect to be cumulative and oscillations to start. If the two tubes are identical, the two cathode biases are equal, and the cathode-to-cathode capacitance can be discarded. At the same

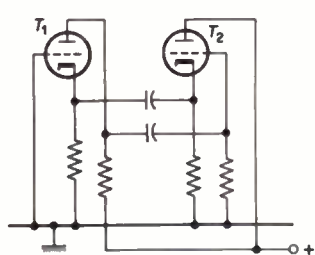


Figure 15-27. Capacitive cathode-to-cathode coupling.

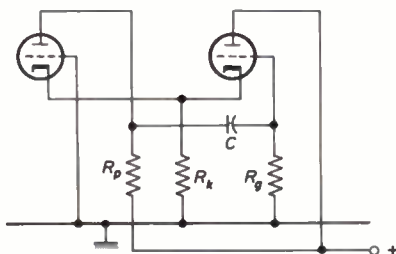


Figure 15-28. Direct cathode-to-cathode coupling.

time, a single common resistance may be used for bias. The resulting simplified circuit, which is more economical than the previous type, appears in Fig. 15-28.

The waveform is rectangular, strongly nonsymmetrical, and quite suitable for television work. The relaxation frequency is roughly given by

$$f = \frac{1}{CR_g}$$

The short alternation corresponds to the return or retrace time of the sweep. It can be made shorter by reducing cathode resistance R_k . However, this reduction, if carried to far, will cause coupling between the two tubes to become insufficient and oscillations will cease.

A double triode tube is frequently used in the circuit. A triode-pentode tube can be employed in a similar arrangement.

Variable relaxation frequency can be obtained by modifying the time constant. A convenient way, as shown in Fig. 15-29, is to use in the grid circuit a potentiometer P with limiting resistor R .

15-16. Waveforms

The simplified analysis of the operation of the multivibrator neglects several secondary effects such as grid current or parasitic capacitances, which

somewhat modify the shape of the output wave. Actually, grid voltage V_g and anode voltage V_a , respectively, look more as shown in Fig. 15-30. Observe the lower portion of the grid voltage curve, which represents the exponential discharge of the coupling capacitance.

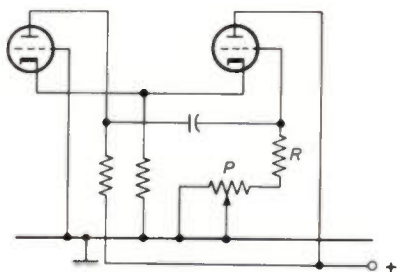


Figure 15-29. Frequency control by potentiometer P .

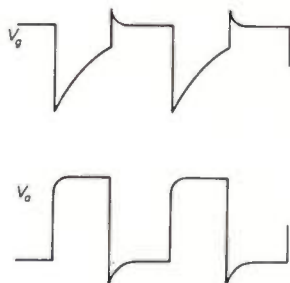


Figure 15-30. Waveforms observed on a multivibrator.

The two waveforms shown in Fig. 15-30 apply to a square-wave multivibrator. The same general type of distortion appears in the case of a non-symmetrical rectangular wave, shown in Fig. 15-31.

When a better square or rectangular wave is required, a good method is to clip off the upper and lower parts of the waveforms using double-clipping

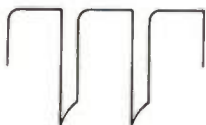


Figure 15-31. Non-symmetrical practical waveform.

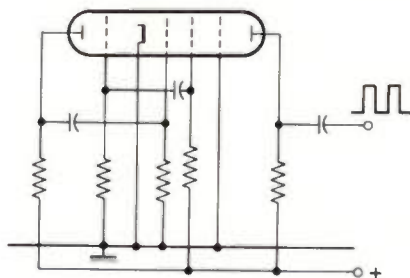


Figure 15-32. Triode-pentode multivibrator.

or slicing. Some amount of clipping may be obtained without additional tubes if a triode-pentode is used for the multivibrator, as shown in Fig. 15-32. In this case, the cathode, control grid, and screen grid of the pentode constitute the second triode of the relaxator, but the output signal is taken off the pentode anode. The same principle can be applied twice, as shown in Fig. 15-33, if the multivibrator is equipped with two pentodes.

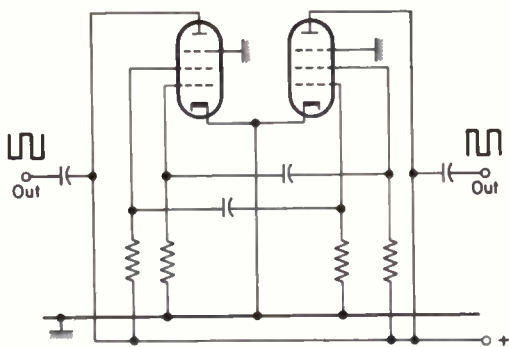


Figure 15-33. Screen cross coupling with pentodes.

15-17. Sawtooth multivibrator

To be used in television, the multivibrator output must be transformed into a sawtooth generator and provisions must be made for synchronization. The simpler cathode-coupled multivibrator is well suited to this purpose. It has one grid that is not used for oscillation, which provides a convenient point of injection for sync pulses. It also has a plate directly connected to the supply voltage and available to introduce an integrating circuit.

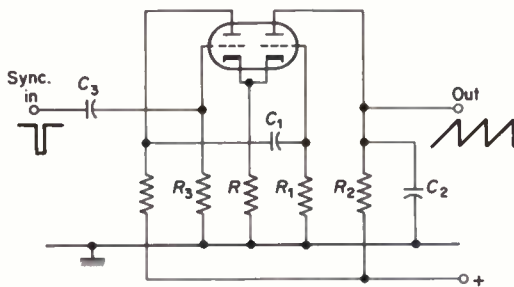


Figure 15-34. Synchronization of a multivibrator.

The resulting arrangement is shown in Fig. 15-34, where R is the cathode coupling resistance, C_1R_1 the coupling circuit, R_2C_2 the integrator, and R_3C_3 the sync injection circuit.

Sync pulses should be negative-going as shown, and a small pulse amplitude, say a fraction of a volt, insures good synchronization. The integrator circuit delivers a sawtooth voltage with quick retrace. Frequency and amplitude of the output signal can easily be controlled by making R_1 and R_2 variable.

A practical circuit for the vertical oscillator at 60 cps is shown in Fig. 15-35. It includes the frequency and amplitude controls and an integrating

circuit at the sync input. A similar circuit, except for component values, is given in Fig. 15-36 for the horizontal frequency oscillator.

The cross-coupled multivibrator is not used too frequently. Figure 15-37 illustrates an application with a triode-pentode tube for line frequency. No amplitude control is provided, the horizontal amplitude being controlled by circuits external to the multivibrator.

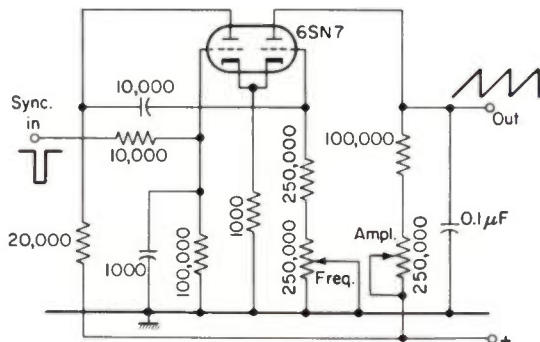


Figure 15-35. Practical diagram of a vertical multivibrator.

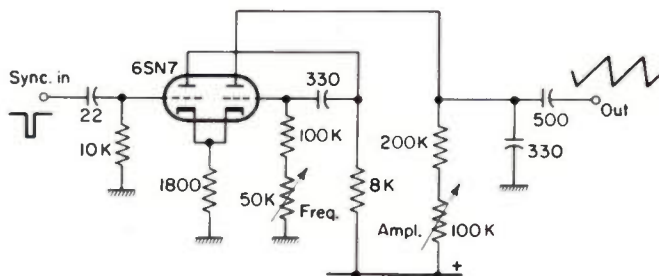


Figure 15-36. Practical diagram of a horizontal multivibrator.

15-18. Blocking oscillator

In addition to the multivibrator, the blocking oscillator is a very common relaxator in television. Its diagram is given in Fig. 15-38. At first sight, it looks just like an ordinary oscillator, but different values for the components make for a particular mode of operation.

The grid time constant RC is long, and the coupling between plate and grid by transformer T is very tight. When the anode voltage is applied, the tube oscillates strongly and a large grid current i flows through R as indicated. A large voltage drop appears across R and charges C negatively, cutting off the tube and killing the oscillation. Grid current disappears, and the charge accumulated on C leaks away through R and the transformer. When the

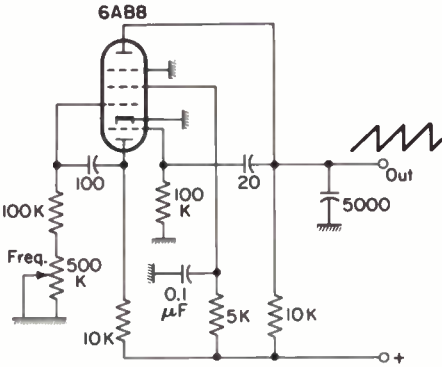


Figure 15-37. Cross coupled horizontal multi-vibrator.

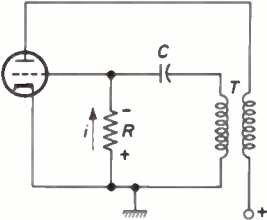


Figure 15-38. Basic circuit of a blocking oscillator.

decreasing negative voltage on C reaches cutoff, oscillations start again and the cycle repeats itself.
The result is a periodic quick charge and slow discharge of the capacitance, that is the production of a sawtooth wave.

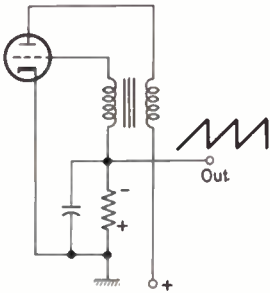


Figure 15-39. Production of a sawtooth.

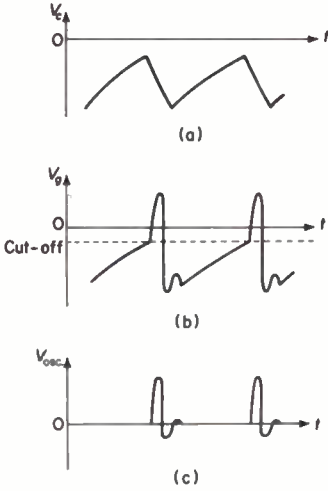


Figure 15-40. Waveforms observed in a blocking oscillator.

The oscillation frequency may have any value. Usually, it is on the order of 10 times the sawtooth frequency, that is, in the sonic or ultrasonic range, and T would have an iron core. Also, since the sawtooth is obtained across C , it might be convenient to have one end of C grounded. The circuit then looks like that shown in Fig. 15-39. The waveforms for this circuit are shown in Fig. 15-40. At (a) is the voltage V_C across the capacitor. At (b) is the grid

voltage V_g . At (c) is the damped oscillation across the transformer T . Curve (b) is evidently the sum of (a) and (c). It shows that the oscillation should be sufficiently damped so that the second positive alternation does not reach cutoff and does not trigger the tube. If necessary, a resistor is shunted across the transformer to ensure this result.

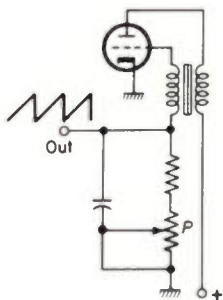


Figure 15-41. Frequency control by potentiometer P .

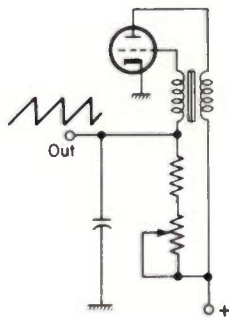


Figure 15-42. The grid can return to plate supply voltage.

Relaxation frequency is conveniently controlled by modifying the grid resistance as shown in Fig. 15-41.

The resulting sawtooth is strongly exponential because the capacitance charges up almost to the full applied voltage due to grid current. The wave-shape can be improved in this respect by connecting the grid to the plate supply, thus increasing the applied voltage as shown in Fig. 15-42.

15-19. Anode load

Similar to the multivibrator, the blocking oscillator switches between cutoff and heavy conduction and its plate current is a rectangular wave. The addition of an integrator R_1C_1 in the anode circuit, shown in Fig. 15-43, will then provide a sawtooth wave. Plate current is zero while the tube is held below cutoff and jumps suddenly to a large value when the tube oscillates. This is represented by the I_a curve shown in Fig. 15-44. The voltage V_a at the top of R_1 has the same form but opposite phase. Adding C_1 produces integration and the resulting voltage V_{C1} is the sawtooth shown in the figure.

In this diagram, the output signal is taken across C_1 and the sawtooth produced in the grid circuit is not used.

Amplitude may be controlled by adjusting R_1 and frequency by adjusting R , as shown in Fig. 15-45.

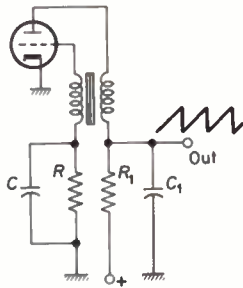


Figure 15-43. Production of a sawtooth in the anode circuit.

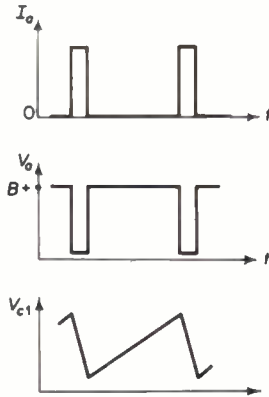


Figure 15-44. Idealized waveforms in the output circuit.

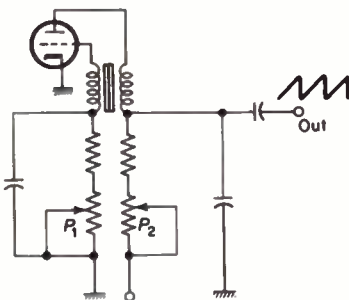


Figure 15-45. Frequency control by P_1 and amplitude control by P_2 .

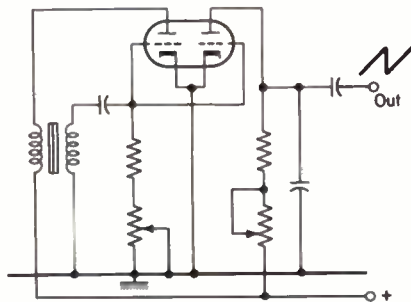


Figure 15-46. Addition of a discharge tube.

15-20. Discharge tube

In free oscillation, the amplitude and frequency controls are somewhat interdependent. This can be remedied by using one tube for oscillation and another for integration.

A diagram based on this principle appears in Fig. 15-46. One half of a double triode is a standard blocking oscillator. It controls the second triode, cutting it off periodically and producing in its plate circuit a rectangular wave. An integrator in the anode will then provide a sawtooth output.

The two functions—relaxation and integration—are thus entirely distinct and the adjustments are independent.

This circuit offers another advantage. In all single-tube blocking oscillators, the discharge path of the capacitance is through the grid-to-cathode space in the tube and through one of the transformer windings. This slows

down the discharge. When a separate tube is used as shown, the integrating capacitance discharges through the tube and no winding is in the path.

A similar circuit can be obtained without resorting to a double triode. A pentode can be used, as shown in Fig. 15-47. Relaxation is produced by the triode constituted by cathode, control grid, and screen grid. The integrating circuit is in the anode and independent.

In fact, a pentode used in this way is equivalent to a double triode. The second triode consists of a fictitious cathode between screen grid and suppressor; its control grid is the suppressor, and its anode is the pentode plate.

The coupling between the two triodes is not a grid-to-grid connection, but a purely electronic coupling by the electron beam between real cathode and fictitious cathode.

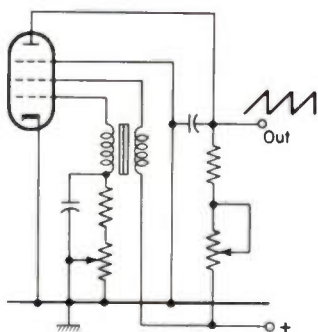


Figure 15-47. Screen grid relaxation.

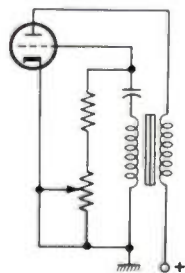


Figure 15-48. Variant of the blocking oscillator.

15-21. Variations

A number of variations of the fundamental blocking oscillator circuit can be designed; many are the result of a simple change in components arrangement as, for example, that shown in Fig. 15-48. Others are more varied, such as the cathode-coupled relaxator of Fig. 15-49, the anode relaxator of Fig. 15-50, the cathode-loaded relaxator of Fig. 15-51, or the symmetrical output relaxator of Fig. 15-52.

15-22. Synchronization

Like all relaxators, the blocking oscillator can be synchronized by a signal applied to almost any tube electrode, preferably in the form of sharp pulses for precise triggering.

The grid is the preferred synchronizing electrode for practical reasons. It offers a high-impedance input and requires a relatively small sync pulse

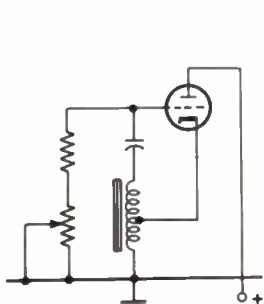


Figure 15-49. Cathode coupled blocking oscillator.

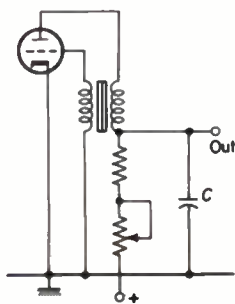


Figure 15-50. Anode relaxation.

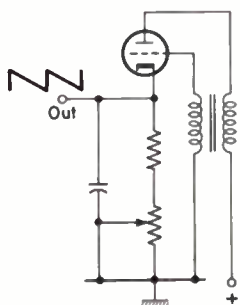


Figure 15-51. Cathode relaxation.

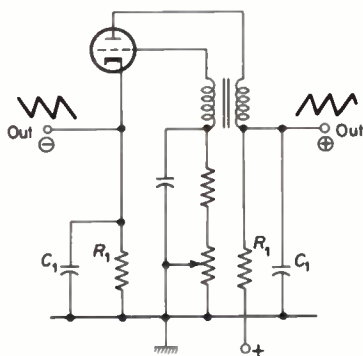


Figure 15-52. Symmetrical sawtooth outputs.

amplitude. The blocking oscillator in this respect is less sensitive than a thyatron or multivibrator, since good synchronization requires several volts.

Since retrace is initiated when the increasing grid voltage crosses cutoff, a positive-going pulse on the grid is necessary to trigger the tube. The coupling circuit may be a differentiator or integrator, and a number of variations are used. Figure 15-53 shows sync injection at the cold end of the grid winding through a coupling circuit CR which could be a differentiator. More simply, as shown in Fig. 15-54, the positive-going sync pulse is applied directly to the grid.

If a negative-going sync pulse is available, it can be applied to the anode. The transformer will transmit it to the grid with phase reversal, so that the grid will again receive a suitable positive pulse, as shown in Fig. 15-55.

Another idea is to add a third winding to the transformer and drive it by the sync pulses. Whether the pulses are positive- or negative-going is easily

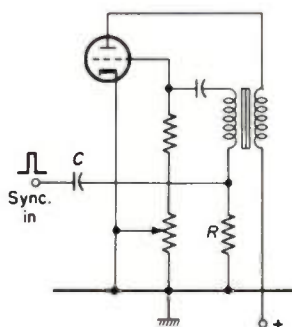


Figure 15-53. Synchronization of a blocking oscillator.

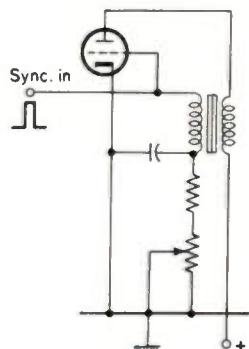


Figure 15-54. Direct synchronization on the grid.

taken care of by crossing the connections to the tertiary winding as shown in Fig. 15-56.

When a blocking oscillator is triggered, a strong pulse is generated which might find its way back up the synchronization circuits. This can be avoided

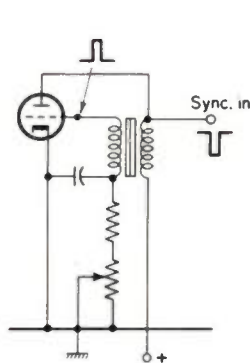


Figure 15-55. Synchronization on the anode.

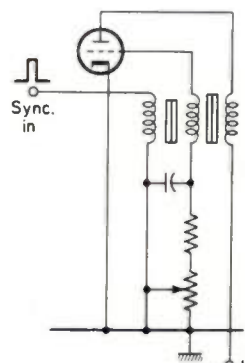
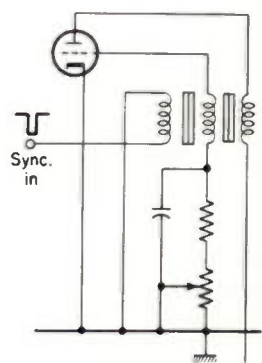


Figure 15-56. Separate winding for synchronization.



by inserting a sync pulse amplifier before the relaxator. Besides amplifying the sync signals, it effectively stops the back parasitic pulse and prevents it from reaching the synchronization section.

15-23. Practical circuits

The only special component in the blocking oscillator is the transformer. It generally uses a small-size core of iron laminations, and has a 1-to-1 ratio.

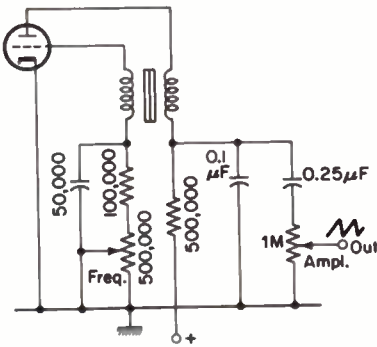


Figure 15-57. Typical vertical blocking oscillator.

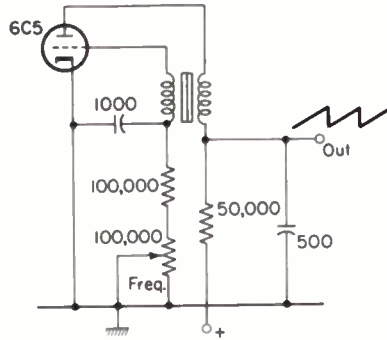


Figure 15-58. Typical horizontal blocking oscillator.

For horizontal frequency, each winding could typically have 300 turns, whereas for vertical frequency 3000 turns could be used. Fine wire is used to keep the size down and at the same time to provide some amount of damping. Losses in the iron core also contribute to the damping so that a low-quality core can be used with advantage.

Component values are not critical. Figure 15-57 shows a vertical frequency relaxator, including the amplitude control at the output. This simple arrangement can be employed because of the low frequency. Figure 15-58 applies to horizontal frequency, and Fig. 15-59 gives a variant with a pentode tube. Figure 15-60 illustrates the case of a double triode as relaxator-discharge tube for vertical frequency. Sync pulses are injected through a double integrator.

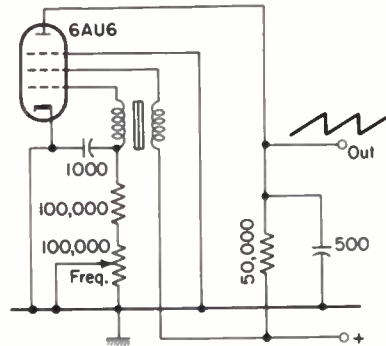


Figure 15-59. Pentode horizontal blocking oscillator.

The same circuit is shown in Fig. 15-61 for horizontal frequency. This time sync pulses are injected through a differentiator.

15-24. Other relaxators

A great number of relaxators are known; they are either derived from the preceding types or use other principles. The most common are the multi-vibrator and the blocking oscillator.

Whatever the circuit used, relaxators can easily deliver a sufficiently good approximation of a sawtooth with an amplitude of up to 50 v, which is

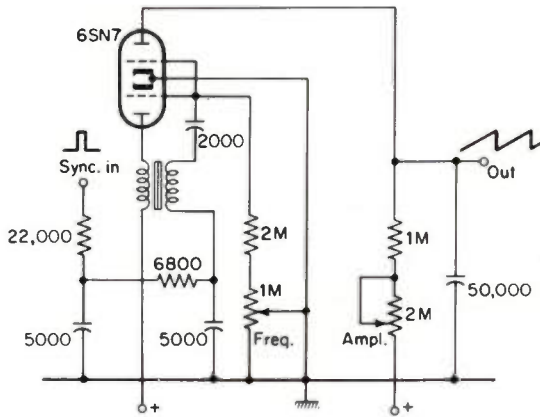


Figure 15-60. Practical circuit for a vertical discharge tube.

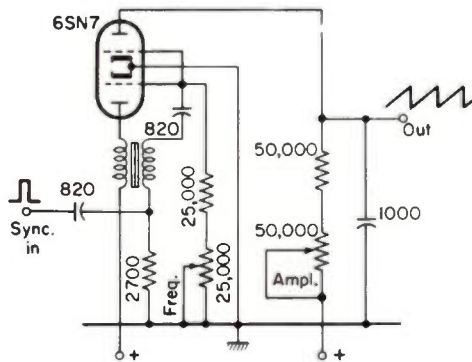


Figure 15-61. Practical circuit for a horizontal discharge tube.

more than sufficient to drive the following stage. If good linearity were necessary, various linearization circuits could be employed. These are rarely, if ever, used because the following stages introduce their own distortions and employ linearizing circuits besides.

15-25. Oscillators

Ordinary sinusoidal oscillators can be used with their output distorted to provide a reasonable approximation of the required waveform. These are only employed for horizontal frequency and one of their advantages is a stable oscillating frequency which can be controlled automatically by a reactance tube.

Figure 15-62 illustrates a cathode-coupled oscillator. Tuned circuit LC has adjustable inductance and resonates on the horizontal frequency of 15,750 cps. An additional circuit, not represented, controls the frequency.

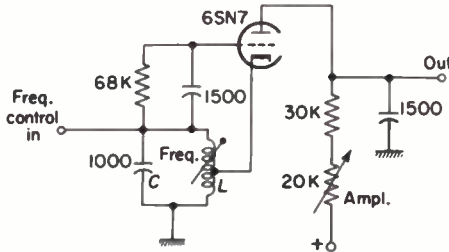


Figure 15-62. Cathode-coupled oscillator.

The tube working conditions and the anode circuit are established in such a way that the desired waveform is obtained at the output. For reasons which will be explained later, this waveform is not a pure sawtooth. Its amplitude reaches 150 v. On the grid the signal is almost sinusoidal and has an amplitude of over 100 v, which shows that the tube works Class C and that plate

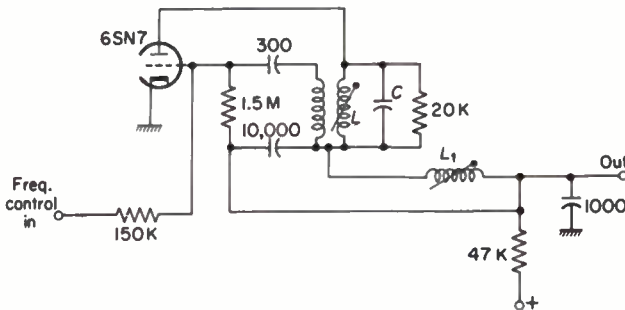


Figure 15-63. Grid-to-plate coupling oscillator.

current is made of short pulses, somewhat comparable to the current pulses produced by a relaxator and integrated in the plate circuit by a standard integrator with amplitude control.

Figure 15-63 shows another circuit in which grid-to-plate coupling is used. Tuned circuit LC has its inductance adjusted to resonate on horizontal frequency. The tube works Class C again, and the effect of the distorting inductance L_1 , with the other circuits components, is to produce suitable current pulses that an integrator circuit transforms into a sawtooth wave of 100 v amplitude.

A frequency control circuit, not represented, is connected to the grid.

15-26. Automatic frequency control

The importance of automatic frequency control, or AFC, has been pointed out with respect to horizontal frequency in Chap. 11, devoted to synchronization (see Sec. 11-23). Circuits have been indicated that provide the required d-c control voltage for the line oscillator.

All relaxators are frequency-sensitive to the voltage on their electrodes, and especially on the grid. A typical circuit is given in Fig. 15-64 for a multivibrator. The AFC control voltage is simply applied to the available grid of

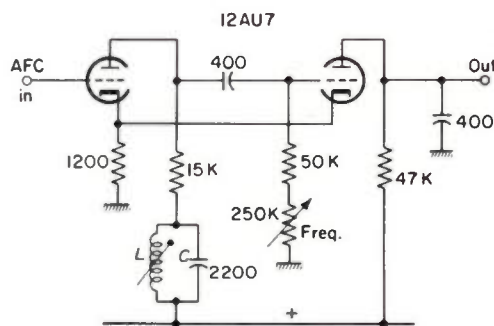


Figure 15-64. Flywheel circuit LC .

the relaxator. The only peculiarity is the circuit LC , tuned on horizontal frequency, in the plate circuit. Its function is to stabilize the multivibrator relaxation frequency. It has the same effect as a mechanical flywheel and is in fact, commonly called a *flywheel circuit*. It may be placed elsewhere in the multivibrator circuit with the same results.

For a sine-wave oscillator, similar arrangements could be used as previously indicated by the frequency control connection to the grid. A complete system, shown in Fig. 15-65, uses a reactance tube to shift the oscillating frequency. The oscillator is a triode with cathode coupling and the L_1C_1 circuit in the grid is tuned on the horizontal frequency. A transformer arrangement is used to drive the frequency comparator or discriminator, which also receives the sync pulses from the synchronization section. The oscillator load is a tuned circuit LC , and the output voltage has a pulse form and switches a discharge tube on and off. An integrator in the anode circuit of the discharge tube provides the sawtooth wave. The d-c voltage resulting from the comparator operation is filtered and applied to the control grid of a pentode used as a reactance tube. This pentode is connected across tank circuit L_1C_1 and controls the oscillating frequency.

A similar arrangement is illustrated in simplified form in Fig. 15-66. The oscillator is cathode-coupled. The reactance tube is connected across the

frequency. The plate pulse it produces at each period is integrated in the plate circuit by an R_1C_1 integrator and transformed into a sawtooth wave. The effect is that, with coil L_3 short-circuited, the waveform shown in Fig. 15-68(a) appears at the output.

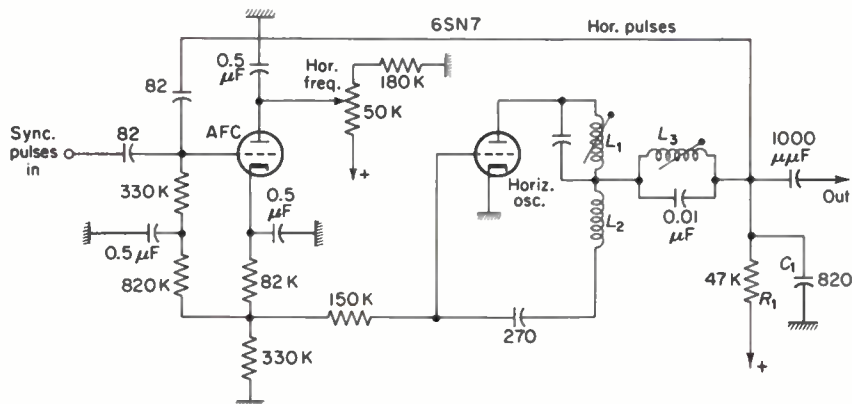


Figure 15-67. Synchroguide horizontal oscillator.

When in circuit, coil L_3 resonates with a shunt capacitor and is shock-excited by the plate voltage pulses. It produces the sinusoidal waveform shown in Fig. 15-68(b), which adds to the integrated sawtooth to produce the total output signal illustrated in Fig. 15-68(c). This waveform is suitable to drive the horizontal output tube.

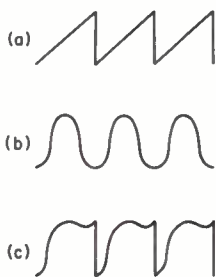


Figure 15-68. Waveforms in synchroguide circuit.

The left-hand triode works as a triode AFC tube (see Chap. 11) or comparator. It receives the sync pulses coming from the sync separator and the output pulse coming from the right-hand triode oscillator simultaneously. Its plate current has a value which depends on the instantaneous value of the total signal on the grid, and hence on the phase relation between sync pulses and oscillator pulses. This anode current flows through a large resistance in the cathode circuit, producing across it the voltage used for AFC purposes. The AFC voltage is applied to the grid of the right-hand triode oscillator in the usual way.

15-28. Practical circuits

Three common practical circuits will be given to illustrate application of the principles previously explained.

Figure 15-69 represents a double diode AFC tube controlling a double triode horizontal multivibrator. The AFC comparator receives on the one hand pulses coming from the line output transformer and, on the other hand,

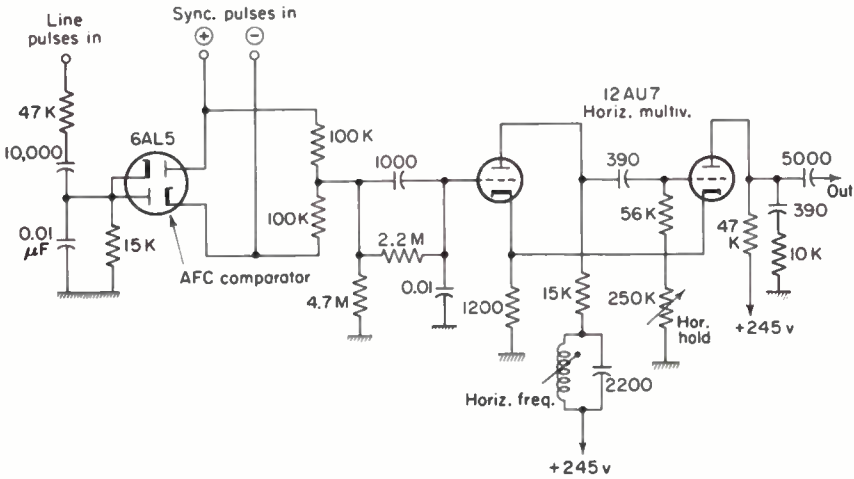


Figure 15-69. Double diode AFC circuit.

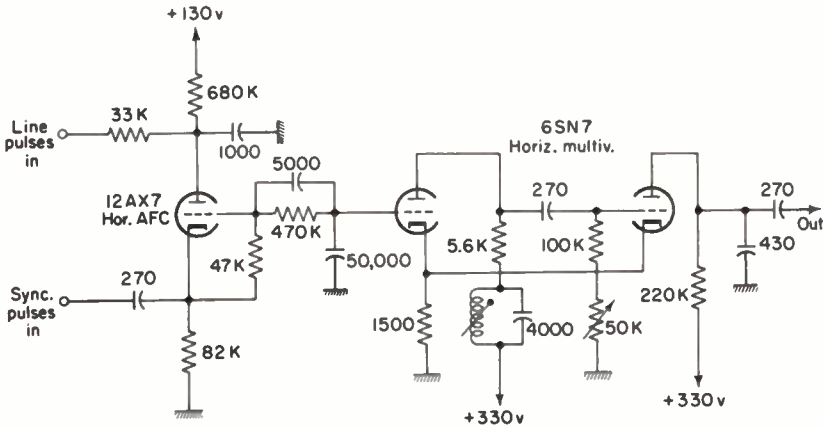


Figure 15-70. Triode AFC circuit.

sync pulses of opposite polarities, provided by a sync phase inverter. The AFC voltage it produces is applied to the control grid of the cathode-coupled multivibrator. A frequency stabilizing circuit is included in the multivibrator.

Figure 15-70 shows a triode AFC tube controlling a double triode

horizontal multivibrator. The triode comparator receives on its cathode the separated sync pulses and on its anode a line pulse coming from the line output transformer. The AFC voltage it produces is filtered and applied to the synchronizing grid of the multivibrator. This multivibrator is cathode-coupled and stabilized by a flywheel circuit in its anode.

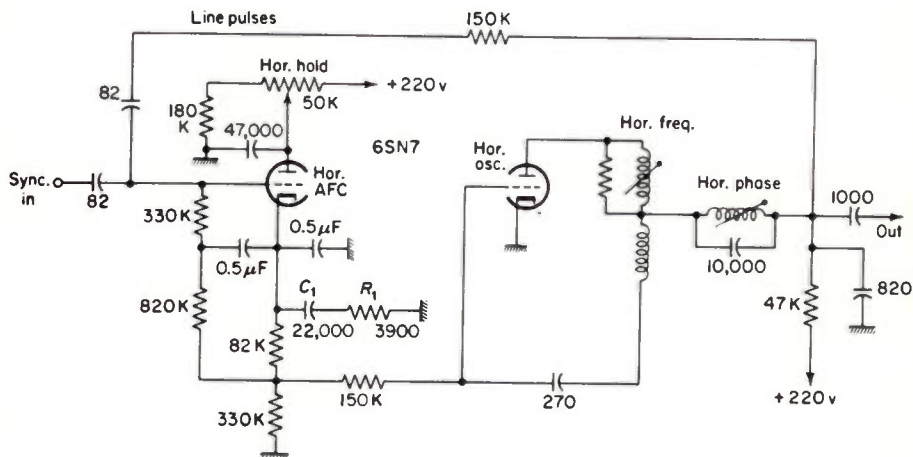


Figure 51-71. Practical synchroguide circuit.

Figure 15-71 illustrates a practical synchroguide circuit. It is almost identical with Fig. 15-67, except for details. The AFC function is assured by one of the triodes of a 6SN7. The other half is the horizontal oscillator. Circuits of this type sometimes exhibit a tendency to enter into low-frequency oscillation due to the fact that there is a feedback loop around the two tubes. When this oscillation occurs, the visual result is that a horizontal sinusoidal displacement is superimposed on the picture, which appears wavy. This oscillation is eliminated by an anti-hunt circuit, identified as C_1 - R_1 on Fig. 15-71.

Other types of AFC circuits are used in connection with horizontal oscillators. They all work either by modifying the d-c voltage on the grid of a tube in the case of relaxators or by changing the value of some reactance connected across a resonant circuit in the case of sinusoidal oscillators. The examples given illustrate the principles, and no difficulty should be experienced in analyzing other commercial circuits.

Problems

1. Explain why a large-amplitude signal is necessary to produce a square wave by clipping.

2. A sine wave is applied to the grid of a tube, and its amplitude is increased progressively. Sketch the plate current waveform for several representative values of amplitude.
3. Show graphically how an integrator RC circuit will transform a rectangular wave into a sawtooth.
4. Explain the fact that the charge of a capacitance through a resistance is not linear.
5. A capacitance of $1\ \mu\text{f}$ charges up from a voltage source of $100\ \text{v}$ through a resistance of $1\ \text{megohm}$. Plot the voltage across the capacitance as a function of time, calculating at least ten values for times between 0 and 10 seconds.
6. A capacitance of $1\ \mu\text{f}$ charged up to $100\ \text{v}$ is connected to a resistance of $1000\ \text{ohms}$. Plot the voltage across the capacitance as a function of time, calculating at least ten points.
7. The charge equation of a capacitance C through a resistance R from a voltage source V can be represented by a series expansion. If the charge is assumed linear, only the first term of the series is kept. A better approximation would be to use the first two terms in the series. Calculate the relative correction obtained when the second term is taken into consideration for the following values of t/RC : $0.01, 0.1, 0.3, 0.5, 0.8, 1, 2, 5$, and 10 . Plot the resulting curve.
8. Plot on a large scale the first 20 per cent of an exponential curve and the linear approximation.
9. Show mathematically how varying the time constant of an integrator will control the amplitude of the integrated signal.
10. Design a neon tube relaxator producing a rectangular wave.
11. Suggest a way of synchronizing a neon relaxator.
12. Describe a method used to calculate the value of the current-limiting resistance in series with a neon tube.
13. Compare the neon tube and the thyatron as relaxators.
14. Considering the thyatron as a triode, how does the addition of gas in the tube modify the operation?
15. Explain how bias can be obtained in the cathode of a thyatron, since the tube does not pass any current most of the time.
16. The charge capacitance in a thyatron relaxator may be connected between plate and ground or between plate and cathode. Compare the two methods.
17. Draw a circuit in which a thyatron relaxator is synchronized by the anode.
18. Draw a circuit in which a thyatron relaxator is synchronized by the cathode.
19. Design a thyatron relaxator producing rectangular waves.

20. Design a multivibrator circuit producing a rectangular wave with a 3/1 mark-to-space ratio and a recurrence frequency of 6000 cps.
21. It is desired to produce a series of very sharp pulses, equally spaced, alternatively positive and negative, and having a recurrence frequency of 800 cps. Suggest a suitable circuit.
22. Design a multivibrator circuit producing a rectangular wave with a minimum mark-to-space ratio of 90/10 and a recurrence frequency variable between 1000 and 2500 cps.
23. Design a multivibrator circuit producing a square wave with a mark-to-space ratio of 50/50 to 1 per cent precision, and a recurrence frequency of 400 to 1600 cps.
24. Using a double triode multivibrator and diode clippers, design a square-wave generator providing good waveform, adjustable amplitude, and frequency variable between 500 and 1000 cps.
25. Design a multivibrator sawtooth generator for a fixed frequency of 60 cps and a large mark-to-space ratio.
26. Design a multivibrator sawtooth generator for a fixed frequency of 15750 cps and a large mark-to-space ratio.
27. Explain how the frequency of sinusoidal oscillation in a blocking relaxator depends on the desired waveform.
28. Design a blocking oscillator synchronized in the cathode circuit.
29. Show graphically how a Class C sinusoidal oscillator can be used as a sawtooth generator.
30. Design an AFC circuit for a sinusoidal oscillator. It should use a reactance tube behaving as a variable inductance.
31. Design an AFC circuit for a blocking relaxator.
32. To obtain better AFC, it is suggested that the d-c control voltage provided by the phase comparator should be amplified before being applied to the relaxator. Suggest a suitable circuit.
33. Explain why AFC is not used for vertical relaxators.
34. Explain in detail the role of the flywheel-tuned circuit in a relaxator.

Chapter 16

VERTICAL SWEEP SECTION

The vertical sweep section produces the 60-cps vertical scanning of the picture. It feeds the field deflecting coils with a current waveform suitable to obtain linear scanning. These coils are commonly coupled to the vertical output tube through a transformer or autotransformer.

Good vertical linearity is not easy to obtain using economical components. A number of correction circuits have been designed to improve linearity; these will be discussed in this chapter.

16-1. High impedance

Because of the low frequency of vertical scanning and because voltages across field deflection coils are low, parasitic shunt capacitance can be neglected and resistance predominates largely over inductance.

High-impedance vertical coils can be wound up to several henrys. A convenient way to apply a sawtooth current is illustrated in Fig. 16-1. It is a simple RC coupling with a long time constant. A resistance load can be used in the anode circuit because plate current is not unusually high, and a power dissipation of a few watts is suitable for the resistance.

A deflection current of from 10 to 40 milliamperes peak-to-peak is usually necessary and is frequently provided by tubes of the audio power category or similar triodes, tetrodes, or pentodes.

At 60 cps, resistance is the dominant component for the vertical coils and an equivalent diagram is shown in Fig. 16-2. The sawtooth current, therefore, will be divided between R and R_1 , so that R should be large with respect to R_1 . Since R_1 is on the order of several kilohms, R should be, say, 20 kilohms for good efficiency. Unfortunately, assuming a 15-ma d-c plate current, R would

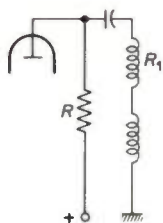


Figure 16-1. Resistance-capacitance coupling to the deflection coil.

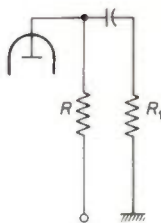


Figure 16-2. Equivalent circuit.

dissipate 4.5 watts, which by itself is not too serious except that it results from 300-v voltage drop, which can be tolerated only if plate supply voltage is high. The alternative, reducing the value of R , leads to poor efficiency and can be used only if there is a good supply of vertical deflection power to spare.

16-2. Coupling capacitance

The coupling capacitor also poses a problem. Its value should be large enough for the sawtooth signal to be transmitted without distortion, but this is not an easy requirement because of the relatively low resistance of the vertical coil. Assume $R_1 = 2000$ ohms. For negligible distortion, time constant CR_1 should be at least 10 times larger than the sawtooth period. The minimum value of CR_1 is then

$$CR_1 = 0.17 \text{ sec}$$

so that C should have a minimum value

$$C = 85 \mu\text{f}$$

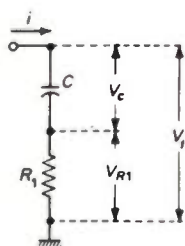


Figure 16-3. Voltage distribution in the deflection circuit.

and since it must be isolated for plate supply voltage, this component is bulky and costly.

A smaller capacitor, say $16 \mu\text{f}$, may be used and some distortion inevitably results. As shown in Fig. 16-3, let i be the linear deflection current through R_1 . Voltage V_{R1} is then linear. Voltage V_C across C is

$$V_C = \frac{1}{C} \int i \, dt$$

and is proportional to the integral of i . Since the integral of a sawtooth has a parabolic shape, V_C is parabolic. Total voltage V_T across the coil is a combination of linear and parabolic waveforms. This is shown in Fig. 16-4, which shows in (a) the linear deflection current, in (b) the capacitance voltage, in (c) the resistance voltage, and in (d) the total voltage across the deflection coil.

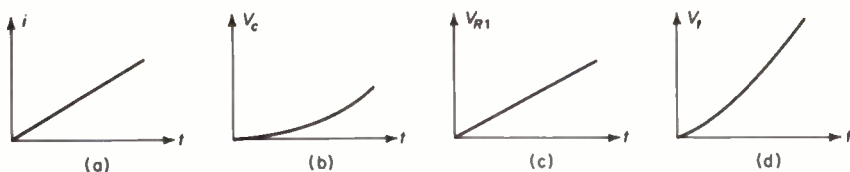


Figure 16-4. Waveforms in the high-impedance deflection circuit.

The parabolic voltage component necessary for linear deflection current must be present at the plate of the tube. Since the load is a resistance, this means that plate current must contain a parabolic component. Two fairly common methods are used to secure this component.

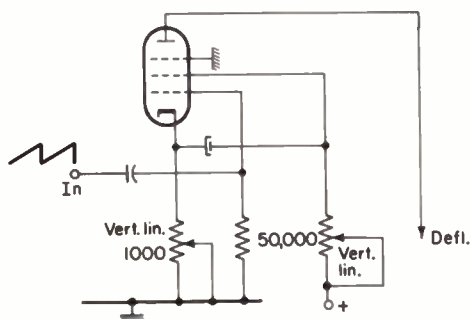


Figure 16-5. Linearity adjustments.

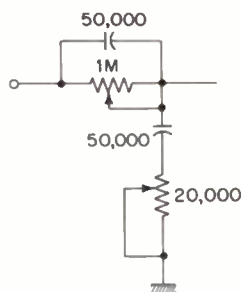


Figure 16-6. Pre-distortion circuits.

The first method, shown in Fig. 16-5, is to put to good use the curvature of the tube characteristic. By selecting the proper operating point, a linear sawtooth grid drive will produce a nonlinear plate current, and the curvature, if properly chosen, will provide a close enough approximation of the parabolic component. The corresponding cathode bias adjustment is labeled *vertical linearity*. Sometimes screen voltage is adjustable for the same reason. Both controls modify vertical amplitude, so that vertical amplitude and linearity adjustments of this type are interdependent.

The second method, shown in Fig. 16-6, keeps the tube working on a linear part of its characteristic, i.e., class A, but applies to the grid a driving signal containing a suitable parabolic component. Starting with a sawtooth, this is easily obtained through integrating circuits introducing partial integration. Generally, such pre-distortion circuits also contain a differentiator to take care of the inductance of the coil, so that good linearity can be easily adjusted. Such a circuit, combining integrator and differentiator, is shown in Fig. 16-6.

Since the integral of a sawtooth is a parabolic wave and since its derivative is a rectangular wave, this circuit will provide a waveform containing the

original sawtooth plus controllable amounts of parabolic and rectangular components. Playing on the two adjustments will allow good over-all vertical linearity to be obtained within a large margin of compensation.

The above techniques, however, do not overcome the problem of high voltage drop and its consequent power loss.

16-3. Choke inductance coupling

The main disadvantages associated with resistance load—high power dissipation and high voltage drop—can be eliminated if an inductance load is used as shown in Fig. 16-7, because its ohmic resistance can be low.

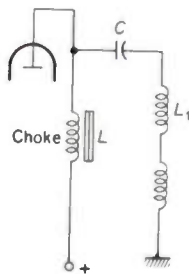


Figure 16-7. Choke inductance coupling.

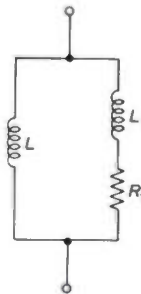


Figure 16-8. Equivalent circuit.

The choke coil should have a high inductance to eliminate undesirable distortion. The equivalent circuit shown in Fig. 16-8 has an inductance L which shunts the deflection coils L_1 - R_1 . Evidently, such a circuit is frequency-dependent (because the impedance of an inductance is ωL , which varies with frequency, whereas that of a resistance has a constant value and does not depend on frequency). Since a sawtooth contains a large number of harmonics, frequency-dependence means unequal harmonic gain and distortion.

Nonlinearity in this circuit can be measured by $TR/(L + L_1)$, where T is the signal period. For low distortion, L may have to be too high for practical convenience. The problem can be solved as shown in Fig. 16-9. A resistor R is added in series with L . When $R/L = R_1/L_1$, the circuit is balanced and frequency-independent, i.e., without distortion. A practical diagram would look as shown in Fig. 16-10, where resistance R is the linearity adjustment. It should have sufficient power dissipation because of the d-c plate current flowing through it.

If vertical flyback is very short, it may create high transient voltages across the choke coil. If such is the case, transient voltages can be limited by the simple expedient of shunting the inductor by a resistor ranging from 10,000 to 50,000 ohms.

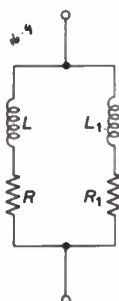


Figure 16-9.
Balanced circuit.

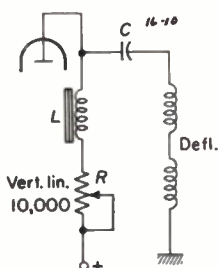


Figure 16-10. Practical
linearization circuit.

16-4. Practical circuit

The complete circuit for a vertical time base appears in Fig. 16-11. A triode is used as a blocking oscillator, and this stage contains the vertical

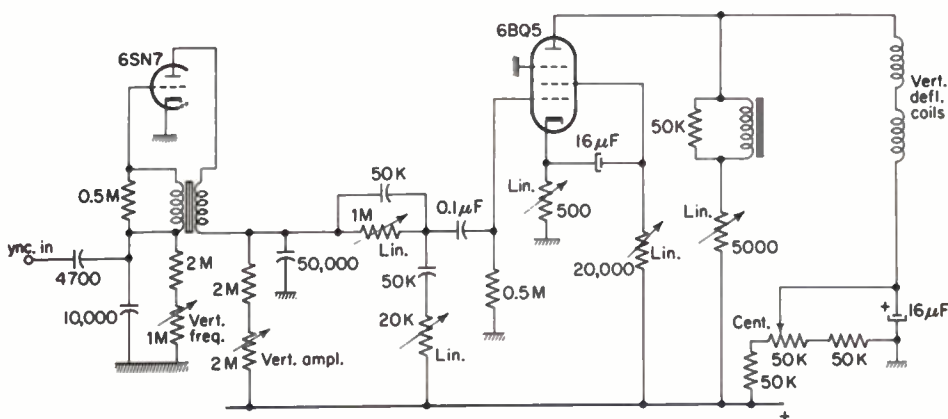


Figure 16-11. Complete diagram of a high-impedance vertical deflection system showing the possible linearity adjustments.

frequency and amplitude or height control. The output stage uses a power tube and high-impedance deflection coils. At the input, an integrator-differentiator type of linearity control has been provided. Also, characteristic curvature controls have been indicated in the cathode and screen connections. In actual practice, however, either one or the other (cathode or screen) type only, not both, would be employed. The anode circuit contains a choke inductance shunted by a transient limiting resistor and in series with it is a linearity adjustment by time-constant control.

A standard centering circuit is provided in the return lead of the deflection coils.

16-5. Transformer coupling

Low-impedance deflection coils are by far more common nowadays. They are coupled to the power tube by a transformer or alternatively by an autotransformer, as shown in Fig. 16-12.

At 60 cps field frequency, the 20th harmonic of the sawtooth is at 1200 cps, so that the frequencies of interest lie in the low a-f range. Standard iron laminations may be used, and the diagram looks, in fact, very much like the corresponding part of a sound output stage of an audio amplifier. The necessary transform ratio could be evaluated as in a-f by considering the total resistance at secondary R_2 and the optimum load resistance R_1 for the tube:

$$n = \left(\frac{R_1}{R_2} \right)^{1/2}$$

It is not always possible to use this optimum transform ratio because of the limited current-handling ability of the output tube and also because the present tendency is to use lower and lower plate supply voltages. In this last case particularly, the important point is that the tube working conditions and the transform ratio should be such that a sufficient peak-to-peak value of deflection current is obtained through the deflection coils.

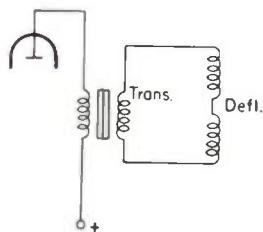


Figure 16-12. Transformer coupling to the deflection coils.

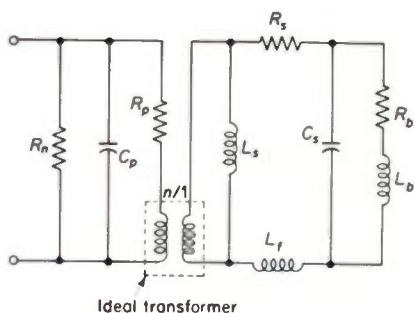


Figure 16-13. Equivalent circuit.

16-6. Equivalent diagram

The transformer or autotransformer constitutes an extra component and also introduces additional distortions. A relatively complex equivalent circuit is represented in Fig. 16-13. In this figure an ideal transformer has been

completed by several additional components whose role is to represent the imperfections of an actual transformer. They are

- R_p = primary resistance,
- R_n = equivalent loss resistance (laminations, etc.),
- C_p = primary parasitic capacitance,
- L_s = secondary inductance,
- R_s = secondary resistance,
- L_f = leakage inductance,
- C_s = secondary parasitic capacitance,
- R_b = deflection coil resistance,
- L_b = deflection coil inductance.

This complex diagram may be simplified if capacitances are neglected because of the low frequency. Then, let

$$R_t = R_s + R_b$$

and

$$L_t = L_f + L_b$$

and the circuit shown in Fig. 16-14 results.

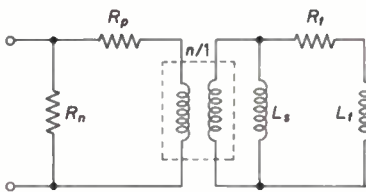


Figure 16-14. Simplified equivalent circuit.

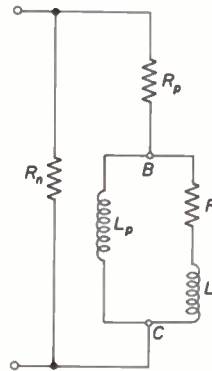


Figure 16-15. All elements have been transferred to the primary circuit.

Now all elements connected to the secondary may be multiplied by n^2 and transferred to the primary, finally resulting in the equivalent diagram shown in Fig. 16-15. L_p is the primary inductance through which flows magnetizing current, whereas R and L are the values of total secondary resistance and inductance transferred to the primary. The transformer-type equivalent circuit shows clearly that R_p and L_f should be minimum. This means that the

secondary should have low resistance and that the primary-to-secondary coupling should be maximum.

Similarly, in the last equivalent diagram, it is apparent that primary resistance R_p should be minimum and that primary inductance L_p should be maximum. These two conditions are somewhat contradictory, because a large inductance means a large number of turns. If bulk is to be of reasonable size, this requires a small-diameter wire and, consequently, high resistance.

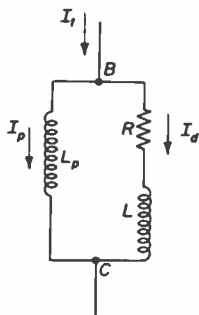


Figure 16-16. Evaluation of non-linearity distortion.

16-7. Nonlinearity

Neglecting R_p and R_n for the time being, the equivalent circuit shown in Fig. 16-15 reduces to the primary inductance L_p across the deflection coils $R-L$, as shown in Fig. 16-16. Total current I_t is split into a magnetizing current I_p and a deflection current I_d .

If the value of I_p is small, then linear I_d would be approximately produced by linear I_t . It is interesting to calculate what the conditions would be for I_d to exhibit reasonable distortion, starting with a linear I_t .

A convenient way to measure the nonlinearity of I_d is to evaluate the variation x of its derivative, that is, of its slope, as pointed out previously. The method of calculation has also been given and in this case provides

$$x = \frac{TR}{L + L_p}$$

where, in this case, $T = 15$ milliseconds. If 10 per cent nonlinearity is tolerated,

$$0.1 = \frac{0.015R}{L + L_p}$$

R is the power tube load; for a typical value, $R = 5000$ ohms, the above expression leads to

$$L + L_p = 750 \text{ h}$$

L is the inductance of the deflection coils multiplied by n^2 and barely reaches a few henrys. Therefore, the primary inductance L_p should have a value of several hundred henrys for good linearity, which requires a transformer that is heavy, bulky, and costly. Engineering economics, however, dictates the use of a relatively inexpensive transformer with the attendant necessity to compensate for its shortcomings by means of correction circuits.

16-8. Waveform

The easy solution in this case is to work the problem backwards, assuming a linear deflection current I_d . The voltage across points B and C is a composite

of sawtooth and rectangular waves. The retrace shape is of no interest as long as retrace takes place within the time limits imposed by the standard and, therefore, only the forward sweep will be considered.

The deflection current may be written as

$$I_d = kt$$

since it is a linear function of time. The total voltage across R and L is

$$E_d = RI_d + L \frac{dI_d}{dt} = Rkt + Lk$$

This voltage appears across L_p . Therefore, the current I_p through L_p is such that

$$E_d = L_p \frac{dI_p}{dt} = Rkt + Lk$$

Integrating,

$$I_p = \frac{1}{2} \frac{Rkt^2}{L_p} + \frac{L}{L_p} kt$$

Total current provided by the tube is

$$I_t = I_d + I_p$$

or

$$I_t = \left(1 + \frac{L}{L_p}\right) kt + \frac{Rkt^2}{2L_p}$$

The first term is linear. The second term constitutes a parabolic component, whose value decreases when R decreases or when L_p increases.

16-9. Parabolic component

The appearance of the parabolic component can also be explained by a method using simplified physical reasoning. Assume the deflection coil in-

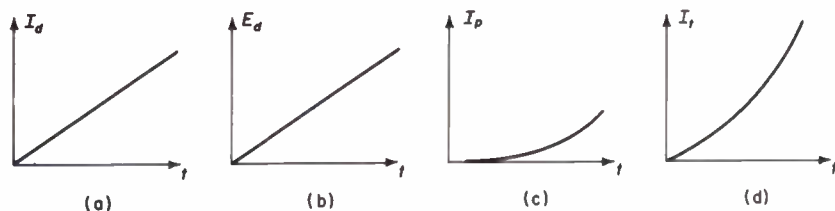


Figure 16-17. Components of the total deflection current.

ductance L is negligible compared to R . Since deflection current I_d is linear, as shown in Fig. 16-17(a), deflection voltage V_d will be

$$V_d = RI_d$$

as shown in (b), and is applied to L_p . The current I_p flowing through L_p will be proportional to the integral of voltage V_d . Since V_d is linear, its integral is parabolic, so that I_p will have the shape shown in (c). Total current I_t shown in (d) is the sum of I_p and I_d and contains linear and parabolic components. Moreover, if the deflection coil inductance L cannot be neglected, it also contains some amount of rectangular wave.

Similarly, if additional components such as R_p and R_n , shown in Fig. 16-15, are taken into account, the relative proportions of sawtooth, parabolic, and rectangular waveforms are modified, but the general shape of the current required from the tube is unchanged.

16-10. Current consumption

Using economical and practical components, the plate current will be strongly parabolic. It is worth noting that such a waveform leads to a marked reduction in current consumption.

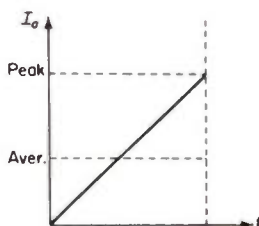


Figure 16-18. Parabolic plate current.

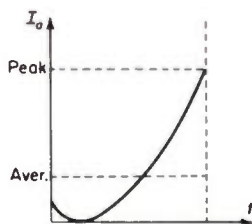


Figure 16-19. Linear plate current.

Figure 16-18 illustrates a linear current having a given peak-to-peak value. The average current is also shown.

Figure 16-19 shows a parabolic current having the same peak-to-peak value. The average current, however, is much lower than that of the preceding case. It can be shown that in this case the average current goes through a minimum for

$$\frac{TR}{L_p} = 3.45$$

where T = sweep duration,

L_p = primary inductance,

R = resistance of the coils transferred to the primary.

For $T = 15$ milliseconds, this leads to

$$L_p = \frac{R}{230}$$

16-11. Linearization

If the parabolic component of waveform is absent or insufficient, the deflection current becomes exponential in shape and the lower part of the picture is compressed.

The problem of a parabolic component for high-impedance coils has already been considered, and the same methods can be applied for low-impedance

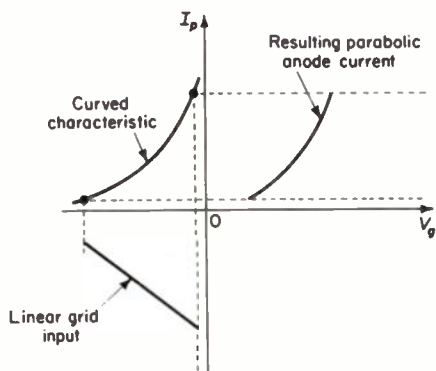


Figure 16-20. Distortion by curvature of characteristics.

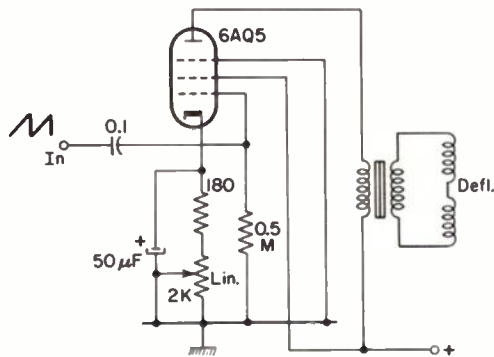


Figure 16-21. Linearization through bias adjustment.

coils. In particular, curvature of tube characteristics may be used by selecting the best working point as previously described. The principle is illustrated in Fig. 16-20. A linear voltage sawtooth applied to the grid results in a good approximation of parabolic current sawtooth in the plate. Notice that the grid sawtooth must be positive-going. A practical circuit diagram is shown in Fig. 16-21.

Similarly, pre-distortion circuits may be used. A simplified circuit of this type is shown in Fig. 16-22. C should be much larger than C_1 . If, as shown in

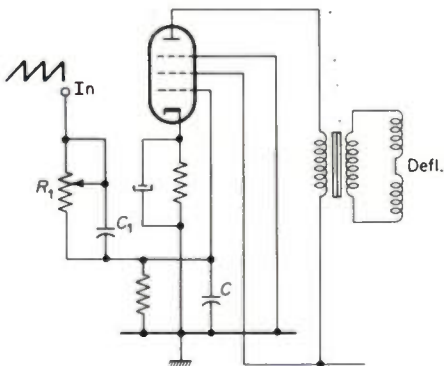


Figure 16-22. Pre-distortion circuit R_1C_1 .

Fig. 16-16, L_p is the primary inductance, and R and L the resistance and inductance of the deflection coils transferred to the primary, optimum values are

$$R_1C_1 = \frac{L + L_p}{R}$$

Adjusting R_1 controls linearity.

16-12. Feedback

Both the preceding linearization methods depend somewhat on tube characteristics, and linearity can change with time or with supply voltages. Feedback systems are frequently preferred because of their inherent stability. They are easily adapted to vertical scanning circuits.

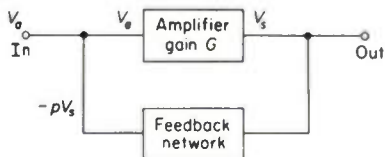


Figure 16-23. Principle of feedback.

The degeneration or negative feedback principle is simple. Consider (Fig. 16-23) an amplifier of intrinsic gain K_0 ,* so that

$$V_s = K_0 V_e$$

where V_e and V_s are respectively the input and output voltage of the amplifier itself.

Now a fraction pV_s of the output voltage V_s is applied through the feedback network to the input, but *in opposite phase to the original input voltage V_a* . Amplifier input voltage is now

$$V_e = V_a - pV_s$$

* In some of the literature, intrinsic gain K_0 is represented by the letter A_0 or A , and the feedback fraction p represented by the letter β .

Substituting for V_s and solving for V_e

$$V_e = \frac{V_a}{1 + pK_0}$$

The output voltage is

$$V_s = \frac{K_0 V_a}{1 + pK_0}$$

so that the gain between input and output of the whole system has become

$$\frac{V_s}{V_a} = \frac{K_0}{1 + pK_0}$$

and is smaller than the intrinsic amplifier gain K_0 . If pK_0 is large enough, 1 can be neglected in the denominator, and total gain becomes

$$\frac{V_s}{V_a} = \frac{1}{p}$$

The gain of the arrangement is the reciprocal of the negative phase feedback fraction p , and is independent of the characteristics of the amplifier.

For example, assume $K_0 = 10,000$ and $p = 0.01$. Then

$$\frac{K_0}{1 + pK_0} = 99$$

which is $1/p$ to 1 per cent. Moreover, for the amplification to change by 1 per cent, K_0 should vary by 100 per cent. A large improvement in gain stability has been obtained then at the price of a large decrease in gain. This last point is rarely important, since it is generally easy to obtain large gain.

Since amplitude or frequency distortions are produced by amplifier non-linearity, they may be practically eliminated by a large amount of feedback. This will be true only if pK_0 is much larger than 1 and, as implied above, if the reinjected signal has a phase exactly opposite to that of the input signal.

The reinjected signal may be proportional to output voltage or output current, depending on whether there is voltage or current feedback. In the case of current feedback, output current flows through a resistor and the voltage developed across it is used as the feedback voltage.

16-13. Feedback utilization

It is relatively easy to obtain a voltage sawtooth, but it is more difficult to obtain a current sawtooth in practical circuits.

In a current feedback system, a voltage proportional to output current is fed back to the input. Such a feedback tends to compel the *output* current to vary as the *input* voltage and the tube works as a linear current amplifier. For

example, deflection current can be made to flow through a resistance, and the resulting voltage drop is fed back to the input in proper phase relationship.

Two different ways of using feedback, therefore, should be clearly distinguished. In one case, a linear voltage sawtooth is applied to the grid and

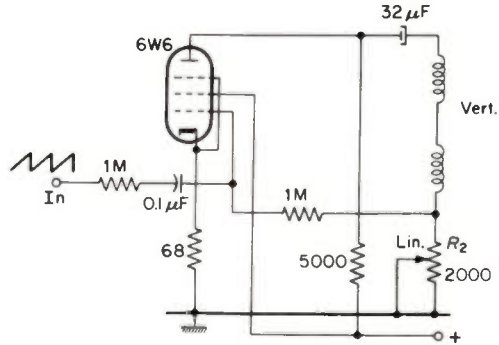


Figure 16-24. Feedback circuit for high-impedance coils.

deflection current is forced by feedback to vary as grid voltage, that is, linearly. In the other case, linear deflection current is obtained for a given shape of plate voltage. A voltage signal of similar shape is applied to the grid, and feedback is used to compel the plate voltage to vary as grid voltage.

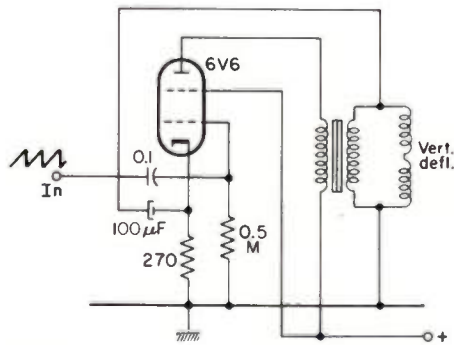


Figure 16-25. Feedback circuit with an output transformer.

The voltage feedback principle shown in Fig. 16-23 takes the practical form shown in Fig. 16-24 for high-impedance deflection coils. R_2 can be used as an amplitude or height control.

A variation for low-impedance deflection coils is shown in Fig. 16-25. The transformer has been used to obtain, across the secondary, a signal opposite in phase to the one existing at the anode. Feedback is then applied to the cathode

through a large value capacitor ($100\ \mu\text{f}$) to avoid shorting cathode bias. This arrangement is used in conjunction with pre-distortion circuits. Because the transformer is included in the feedback loop, feedback corrects for the transformer deficiencies, at least partially, as well. Notice that in this circuit, since the deflection coils are essentially resistive, feedback voltage is the voltage drop across the resistance of the coils due to deflection current.

16-14. Balanced circuit

Inductance and resistance are distributed in a deflection coil and it is impossible to separate them so as to have access to the voltage developed across the resistance only, which would be proportional to deflection current and exactly

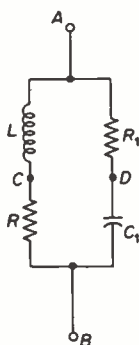


Figure 16-26. Deflection coil and auxiliary circuit R_1C_1 .

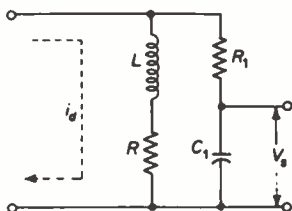


Figure 16-27. Output voltage across capacitance.

what is needed for current feedback. The introduction of an additional resistor, as was just indicated, in series with the coil, solves the problem but reduces efficiency.

A circuit which artificially provides access only to the resistance of the coils is shown in Fig. 16-26. L and R are the coils' inductance and resistance, either directly or transferred to the primary of the output transformer. R_1C_1 is a shunt auxiliary circuit. The arrangement is, in fact, a bridge circuit. It will be balanced when

$$\frac{L}{R} = C_1 R_1$$

and in this case the voltages at points C and D will be equal. C is not accessible in a given coil as explained, but D may be used just as well. In other words, the voltage across R_1 is equal to the voltage across L , and the voltage across C_1 is equal to the voltage across R . Two useful circuits result from these considerations. The first, shown in Fig. 16-27, applies to current

feedback. The feedback voltage V_s is taken across C_1 and is equal to the voltage across R :

$$V_s = Ri_d$$

where i_d is the deflection current. If a linear variation is imposed on V_s , then i_d is linear.

The second diagram is shown in Fig. 16-28 and applies to voltage feedback. The feedback voltage V_s is taken across R_1 and is equal to the voltage across L :

$$V_s = L \frac{di_d}{dt},$$

which is proportional to the derivative of the deflection current, that is, to its

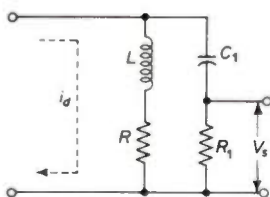


Figure 16-28. Output voltage across resistance.

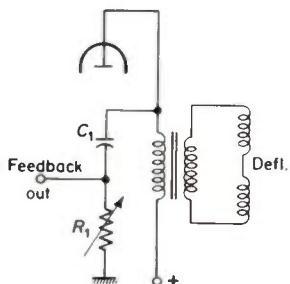


Figure 16-29. Application to a transformer coupled output.

slope. If V_s is contrived by the above technique to be held constant, the current will be linear.

Conversely, as V_s varies in proportion to the nonlinearity of the deflection current, the voltage feedback can be used to correct for it.

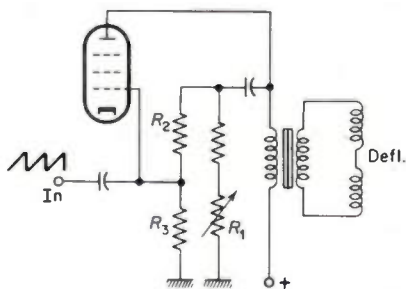


Figure 16-30. Resistance voltage divider.

The last arrangement is commonly used in the practical diagram shown in Fig. 16-29, in which R_1 would normally be a potentiometer to adjust bridge balance and would be called the *linearity adjustment*.

The feedback voltage so obtained is generally too high and would be reduced by a resistance bridge R_2 – R_3 before being applied to the grid as shown in Fig. 16-30. The reduction factor is commonly $\frac{1}{5}$ to $\frac{1}{10}$. A corresponding practical circuit is given, as shown in Fig. 16-31, and could be applied to any power output stage. Notice

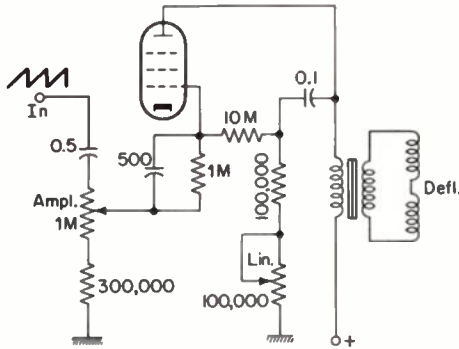


Figure 16-31. Practical circuit.

that the amplitude control is included in the dividing resistance bridge to the grid. As a result, when amplitude is increased, feedback is increased as well, and this makes for good over-all linearity.

The $500\text{-}\mu\text{f}$ capacitor C_1 is optional. Its role is to reduce the retrace time. It modifies the linearity slightly.

16-15. Blumlein circuit

A variation of the preceding circuit, shown in Fig. 16-32, is known as *Blumlein's circuit*. The feedback voltage is, as before, proportional to the anode current variation, but instead of its being used directly, it is integrated by a

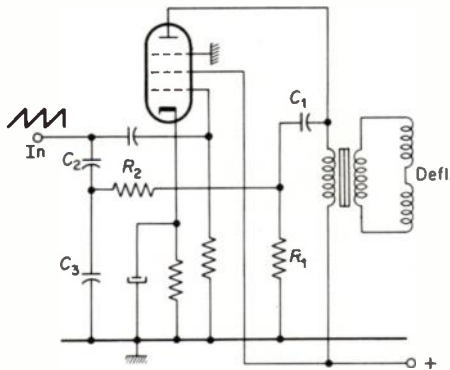


Figure 16-32. Blumlein circuit.

circuit $R_2\text{--}C_3$ and provides across C_3 a voltage proportional to anode current. This voltage is fed back to the grid in a peculiar way. The integrator capacitance of the relaxator has been split into two series capacitances C_2 and C_3 , and C_3 is at the same time the feedback circuit integrating capacitance. Thus,

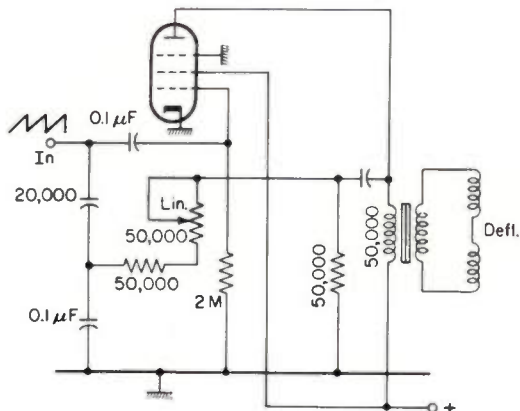


Figure 16-33. Practical Blumlein circuit.

the input sawtooth and the feedback voltages are simultaneously applied to the grid. In actual practice, the circuit has the form shown in Fig. 16-33, which is applicable to any output tube.

16-16. Miller integrator

Miller effect has already been analyzed. By way of review, its effect is to multiply by $(K + 1)$ the apparent value of the total capacitance C between grid and plate of a tube. K is the stage gain, and C includes the parasitic grid-to-plate tube capacitance plus any external capacitance connected between grid and plate as shown in Fig. 16-34. Multiplying the capacitance by a factor of $(K + 1)$ multiplies the time constant by the same factor. For a given charge voltage, this is equivalent to improving the linearity or keeping the charge current I more nearly constant.

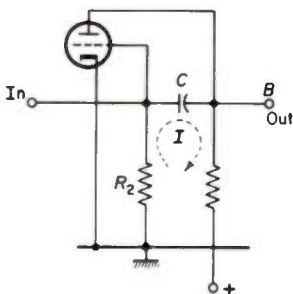


Figure 16-34. Miller integrator.

The circuit shown in Fig. 16-34 evidently applies to a voltage sawtooth but it can easily be adapted to a current sawtooth. An advantage can be taken from the fact that the Miller circuit compels the voltage variation at B to be very nearly linear or from the fact that current I is nearly constant as well. Such circuits are said to belong to the *magnetic* type of Miller integrator.

16-17. Miller circuits

Using the previously developed principle of the balanced bridge with $R_1 C_1$ across the deflection circuit, the voltage for the Miller integrator, as shown in Fig. 16-35, can be taken across C_1 and is equal to the voltage across R .

The Miller integrator compels the voltage at B in Fig. 16-35 to vary linearly. Consequently, the equal voltage across R varies linearly, and so must the deflection current through R .

In Fig. 16-36, the Miller integrator maintains current I through R_2 constant. Hence the voltage drop across R_1 is constant, and so is the equal voltage drop across L . This means that the deflection current varies linearly. With this circuit, the amount of feedback may be too large, and it can be reduced by splitting C into two series capacitances. If this is done, the arrangement becomes the Blumlein circuit previously given (Sec. 16-15).

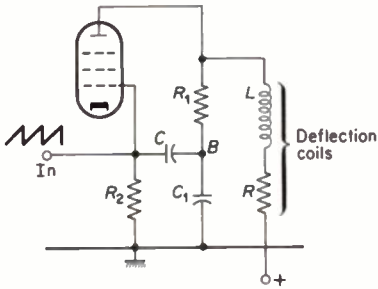


Figure 16-35. Miller integrator in deflection circuit.

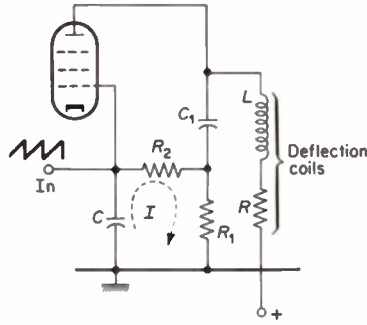


Figure 16-36. Variant of the integrator.

The two Miller circuits can be combined by taking the voltage for the integrator across L plus R , that is, at the anode. The coupling to the grid should then contain resistance and capacitance in series. A practical and commonly used circuit is shown in Fig. 16-37. Feedback is fixed and linearity is adjusted by the cathode bias.

16-18. Autotransformer

The vertical output transformer is frequently replaced by an autotransformer, as shown in Fig. 16-38. Capacitance C prevents a fraction of the d-c plate current from flowing through the deflection coils and producing a permanent decentering. It should have a large value, and an electrolytic capacitor C is used as shown.

An electrical centering circuit may easily be added, as shown in Fig. 16-39. This arrangement ensures that the capacitance will never receive a voltage of the wrong polarity because in no case can the voltage at the base of the deflection coils exceed the plate supply voltage.

Vertical decentering is rarely important and few receivers incorporate a vertical centering circuit. To eliminate the d-c component through the

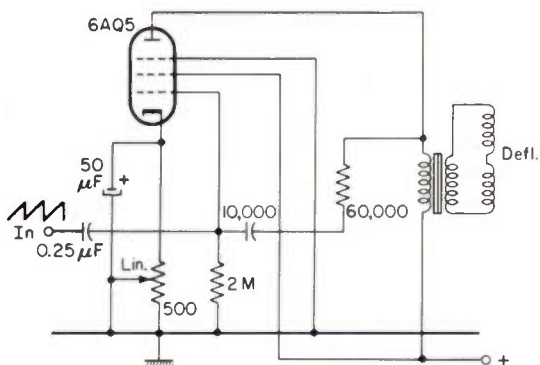


Figure 16-37. Practical application.

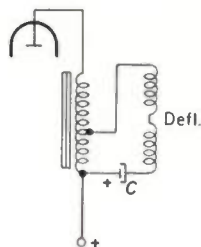


Figure 16-38. Auto-transformer output coupling.

deflecting coils, plate voltage can be supplied at their center point as shown in Fig. 16-40. In this way, plate current is divided into two equal parts flowing in opposite directions through the coils, and no decentering results.

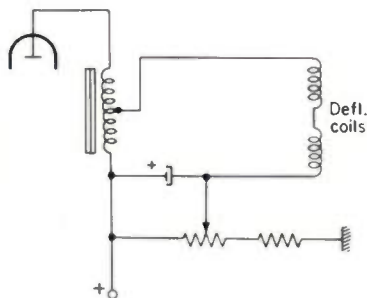


Figure 16-39. Centering circuit with an autotransformer.

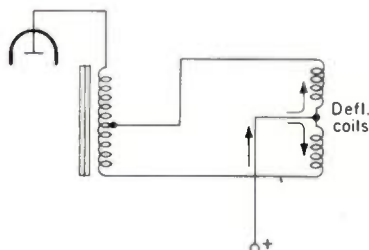


Figure 16-40. Feeding the plate current through the center tap of the deflection coils.

This problem of d-c component decentering does not occur with transformer coupling. If, however, centering is desired in this case, a number of circuits are available. Figure 16-41 shows the use of a center-tapped centering potentiometer, which is fed a centering current through a resistance bridge across the plate supply. The centering current could just as well be the current flowing in any circuit, such as a plate supply or, as shown in Fig. 16-42, the cathode current of a power tube. In some circuits, the secondary has its low end connected to the supply voltage, as shown in Fig. 16-43, and a simple centering arrangement may be provided.

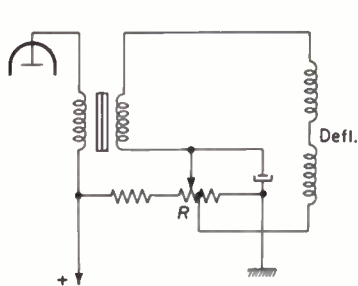
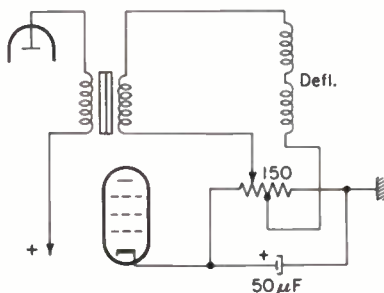
Figure 16-41. Electrical centering by potentiometer R .

Figure 16-42. Centering by cathode current.

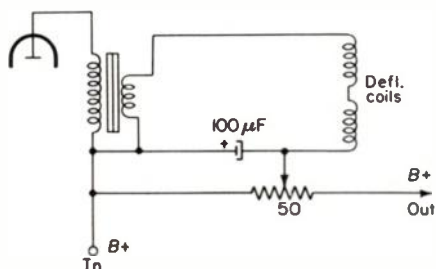


Figure 16-43. Centering by plate current supply.

16-19. Practical diagrams

Figure 16-44 represents the complete diagram of a commonly used vertical time base circuit employing a single triode-pentode tube. This illustration embodies a number of the circuits analyzed in this chapter.

The triode portion is used as a blocking oscillator, with frequency control by the grid circuit time constant. The anode current pulses are integrated by a fixed time constant circuit, and a fraction of the resulting sawtooth is taken off the amplitude or height control potentiometer, point d .

The pentode is the power amplifier. It drives the low-impedance deflection coils through a step-down matching transformer. A standard feedback circuit is connected between plate and grid and controls linearity. The plate supply voltage is rather high and comes from the boosted voltage in the horizontal time base, as we will see in the next chapter.

Working conditions have been measured as follows:

1. cathode voltage—11 v,
2. cathode current—11.8 ma,
3. pentode anode current—9.6 ma,
4. pentode screen current—1.5 ma,
5. total current for the whole circuit—15 ma.

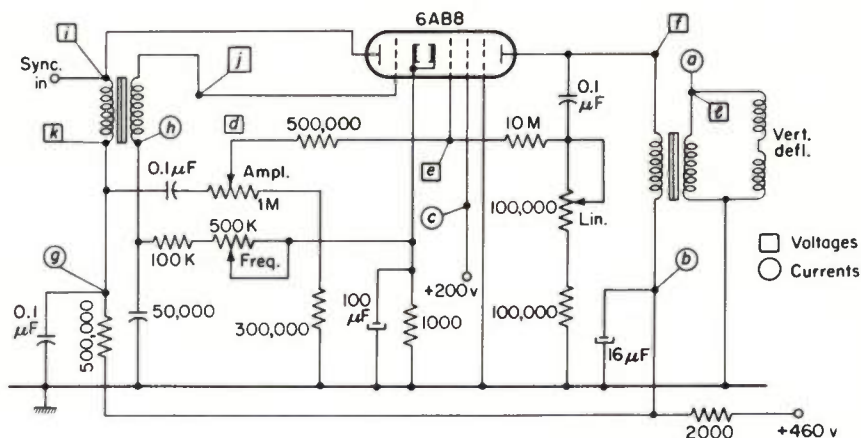


Figure 16-44. Complete diagram of a single tube vertical deflection system.

In addition, oscillograms of the shapes of various currents and voltages in the circuit have been reproduced in Fig. 16-45, with the same letters of reference as used on the diagram. Notice particularly the deflection current at *a*, very linear during sweep and at *l* the deflection voltage with sawtooth, parabolic, and pulse (rectangular) components; in *b* the strongly parabolic anode current; in *d* the sawtooth input to the grid of the pentode; in *e* the grid drive after feedback has added its contribution.

Figure 16-46 shows a vertical sweep section using a double triode. The first half of the tube is a cathode blocking oscillator, directly coupled to the second half which is the power amplifier. Plate voltage of the blocking oscillator is on the order of 100 v, the grid of the second tube is at +100 v, and correct bias is obtained by using a large-value cathode resistance, part of which is an adjustable linearity control. The output uses an autotransformer, and the plate supply voltage is applied at the center of the deflection coils to avoid decentering. Vertical coils are damped by a low shunt resistance.

Figure 16-47 also uses a double triode, of which the first half is a standard blocking oscillator and the second half is the power amplifier. A certain amount of peaking is provided in the output integrator of the relaxator by resistance *R*. The output circuit employs a transformer and provision has been made for a centering arrangement using an undecoupled low-value potentiometer.

16-20. Power relaxator

The two functions of relaxation oscillator and power amplifier can be combined in a single tube called a *power relaxator*. For example, Fig. 16-48 shows

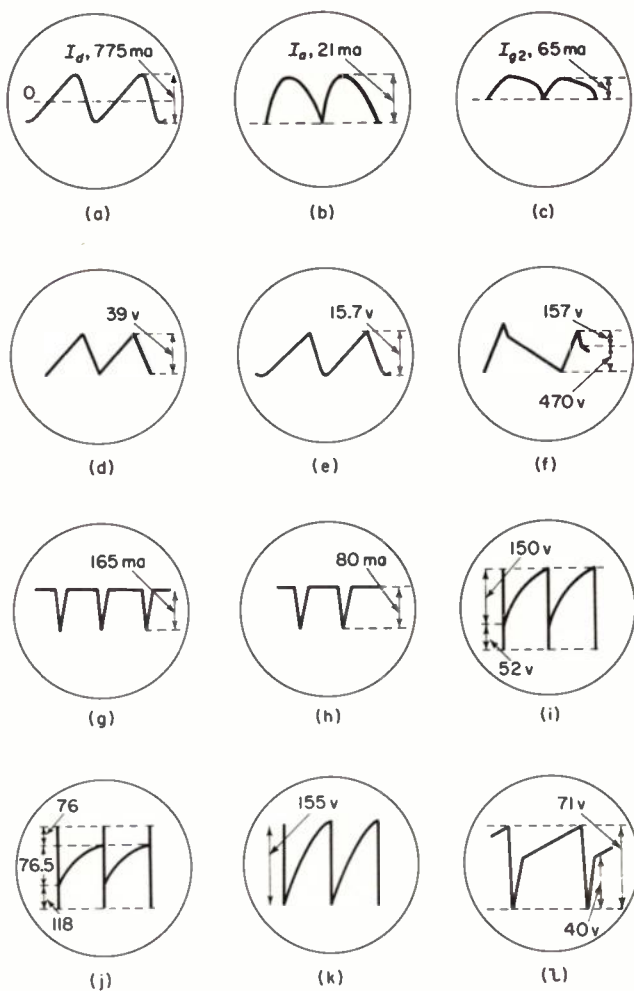


Figure 16-45. Waveforms observed in the preceding diagram.

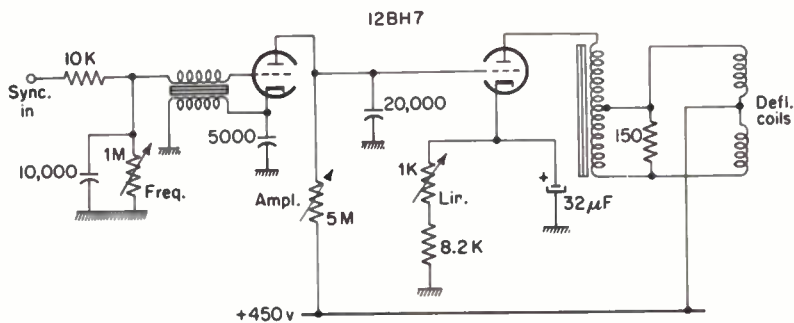


Figure 16-46. Double triode vertical sweep section.

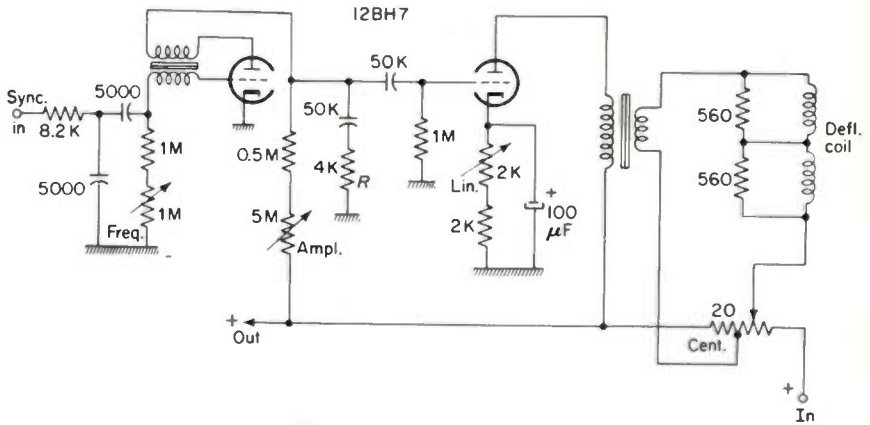


Figure 16-47. Vertical sweep section with electrical centering circuit.

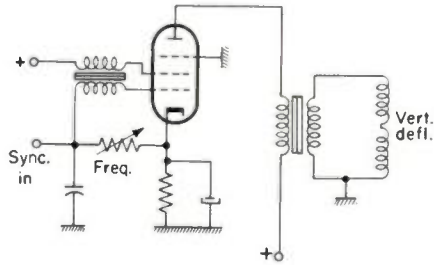


Figure 16-48. Single pentode vertical sweep section.

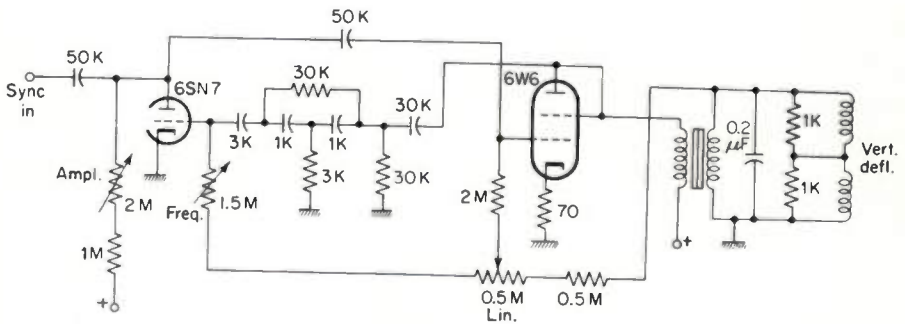


Figure 16-49. Multivibrator type of combined sweep section.

Vertical blanking pulses are taken off the screen grid and are applied to the grid of the picture tube through a simple shaping circuit.

Problems

1. Justify the use of audio output power tubes in vertical deflection output stages.
2. Calculate the minimum value of coupling capacitance to high-impedance coils having an ohmic resistance of 3300 ohms.
3. Show, with diagrams, what distortion results from using a coupling capacitance of too low a value with high-impedance coils.
4. Explain why a parabolic component of current is necessary in vertical deflection.
5. Show, with diagrams, the effect of an integrator pre-distortion circuit on a sawtooth.
6. Show, with diagrams, the effect of a differentiator pre-distortion circuit on a sawtooth.
7. Compare resistance and choke inductance coupling to high-impedance coils.
8. A high-impedance deflection coil measures 1.8 henry and 1800 ohms. Evaluate the value of the choke inductance coupling for a maximum distortion of 10 per cent.
9. Compare high-impedance and low-impedance deflection coils.
10. Compare choke inductance and transformer coupling to the deflection coils.
11. A low-impedance deflection coil measures 0.4 h and 300 ohms. It is matched to the optimum load value of the power tube by a transformer having a ratio of 5. Evaluate the transformer primary inductance for a maximum distortion of 10 per cent.
12. An output transformer has a primary impedance of 60 h and drives a deflection coil of 0.5 h. The equivalent primary resistance is 3000 ohms. For linear deflection, calculate the ratio of parabolic to linear current component of plate current.
13. Suggest three different ways of producing the parabolic current component.
14. Explain the difference between voltage and current feedback.
15. Compare the Blumlein circuit and the Miller integrator.
16. Discuss the various methods of linearization using feedback from a practical point of view.
17. A deflection coil measures 0.35 h and 520 ohms. Design a balanced shunt circuit for feedback purposes.

18. Compare plate-to-grid feedback in the output tube circuit with feedback between cathode and secondary of the output transformer.
19. Design an amplitude control circuit which will provide maximum feedback for the desired amplitude.
20. Compare transformer and autotransformer driving of the deflection coils.
21. Design a vertical centering circuit for a transformer-coupled coil, using the vertical output tube cathode current.
22. Compare a vertical deflection stage and an audio output stage.

Chapter 17

HORIZONTAL SWEEP SECTION

The horizontal deflection coils producing the horizontal lines on the picture tube receive horizontal deflection current from the horizontal sweep section. Because of the high value of line sweep frequency (15,750 cps), two important problems occur in horizontal deflection. The first relates to the large amount of power needed. The second pertains to transients in deflection circuits.

The problem of power is handled through the use of high-efficiency circuits and by applying the principle of recuperation of power.

The problem of transients requires a satisfactory practical solution through the use of some form of damping.

A horizontal line output transformer is generally used to match the deflection coils to the power stage. Its design imposes some serious problems because of the frequencies involved and because of transients.

17-1. Direct coupling

As in the case of vertical scanning, horizontal scanning may use two different types of coupling between the power stage and the deflection coils. In the high-impedance type of circuit, as shown in Fig. 17-1, the horizontal deflection coils are driven directly by the anode of the tube through a choke-capacitance coupling circuit. The purpose of this arrangement is, as explained for vertical deflection, to prevent the d-c plate current from flowing through the coils and producing a permanent decentering.

The same variations used for vertical deflection may be used for horizontal deflection. In particular, the capacitance may be connected between the low end of the coils and ground, as shown in Fig. 17-2, and a centering circuit can be provided. Because of the high sweep frequency, the capacitance may

have a much smaller value than was the case for vertical deflection and still offer negligible reactance.

For good efficiency, the choke inductance L should be much larger than the horizontal deflection coils' inductance, and this again offers no difficulty.

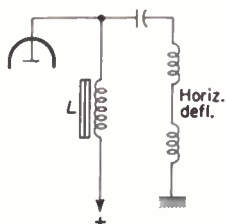


Figure 17-1. Choke inductance coupling to the deflection coil.

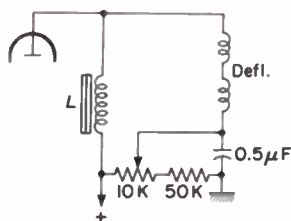


Figure 17-2. Centering circuit.

The equivalent circuit then reduces to that shown in Fig. 17-3, where L is now the equivalent total inductance, which is practically equal to the deflection coil inductance, C is the total parasitic capacitance, and R is the total shunt resistance (physical, tube, losses, etc.).

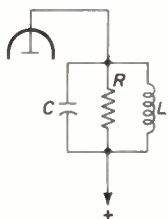


Figure 17-3. Simplified equivalent circuit.

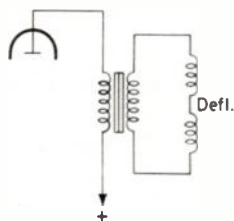


Figure 17-4. Transformer output coupling.

17-2. Transformer coupling

With low-impedance deflection coils, a transformer is necessary, as shown in Fig. 17-4, to couple the deflection circuit to the anode of the power tube.

This transformer is by no means a perfect element. It has resistance and capacitance in both primary and secondary and also a leakage inductance. At line frequency, the resistances may be neglected compared to the inductances for both primary and secondary and also for the deflection coils. Capacitances, however, become important at this frequency.

Figure 17-5, therefore, shows an equivalent diagram in which the actual transformer has been replaced by an ideal transformer to which extra elements

have been added to represent the various imperfections of the actual transformer. Resistances have been neglected, except for a total equivalent shunt resistance R representing all ohmic losses, parallel damping, tube shunting effect, etc. The components are

- R = total shunt resistance,
- C_p = primary shunt capacitance,
- L_s = secondary inductance,
- L_f = leakage inductance,
- C_s = secondary shunt capacitance,
- L_b = deflection coils inductance.

Neglecting the leakage inductance, all secondary components can be transferred to the primary. Then all inductances and all capacitances can

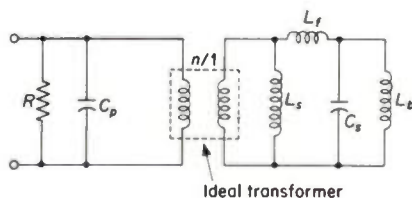


Figure 17-5. Equivalent circuit.

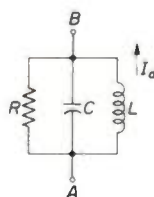


Figure 17-6. Resonant circuit.

be grouped together in L and C and the simple equivalent circuit shown in Fig. 17-6 is obtained. It is the familiar antiresonant LC circuit with shunt resistance R . It is well known that this circuit can be oscillatory or nonoscillatory, depending on the relative component values.

If damping is small, that is, if R is large, the circuit is resonant and oscillating. If damping is large, that is, if R is small, the circuit is nonresonant or nonoscillating.

The largest or critical value R_c of R for which the circuit is just nonoscillating is called the *critical damping*. It is given by

$$R_c = \left(\frac{L}{4C} \right)^{1/2}$$

In all practical cases for a horizontal line output transformer, R is much larger than critical value and the circuit is oscillating. This means that if these oscillations are to be suppressed, an extra damping resistance must be connected across the circuit.

17-3. Waveforms

The desired horizontal sweep waveform is a linear sawtooth deflection current I_d , shown in Fig. 17-7 (at least for the up-going or sweep part of the sawtooth).

The exact shape of the descending part (or retrace, return, or flyback) is immaterial as long as retrace is accomplished inside the time limits allowed by the standard.

Since horizontal line deflection coils may be considered (to a good approximation) as purely inductive, the deflection voltage V_d across them is

$$V_d = L \frac{dI_d}{dt}$$

that is, proportional to the derivative of current and therefore has the rectangular waveform shown in Fig. 17-8.

V_d is also the voltage appearing across the whole circuit given in Fig. 17-6. To the sawtooth current I_d through L will be added a current through



Figure 17-7. Linear sawtooth deflection current.



Figure 17-8. Rectangular deflection voltage.

C and a current through R , to make up the total current I_t that the tube should deliver for linear deflection.

During the period of rising linear sweep, V_d is constant, so that the current I_r through R is constant and the current I_c through C is zero. Total voltage is then simply the rectangular waveform of Fig. 17-8 plus a d-c component.

During retrace, however, the process becomes more complicated. At the end of the duration of the linear sweep, maximum current flows through L and a rather high constant voltage exists across L . When the flyback is triggered, the grid of the tube suddenly becomes negative, and the voltage V_d across L should instantaneously change in value. This is evidently impossible because of the electrical charge stored on the charged capacitance C and because of the magnetic energy stored in L in the form of the existing maximum magnetic field.

Actually, since the circuit is resonant, the steep descending transient shock-excites it into oscillation. The oscillation will last as long as the electric and magnetic stored energies have not been dissipated on R .

Current starts from the maximum value I_a reached at the end of the sweep and executes a sine-wave oscillation as indicated in Fig. 17-9(a). At the same time, voltage starts from value V_a at the end of the sweep and oscillates as shown in (b). Notice on Fig. 17-6 that the anode is connected to B . The current during sweep is taken as positive and flows from A to B through L , so that B is negative with respect to A .

Since A has been chosen as zero reference level for the voltages across the circuit, B is negative during sweep. In other words, the voltage across the circuit is negative during sweep. This explains why curve b starts from a negative V_a value whereas curve a starts from a positive I_a value. Also, notice the first high positive peak of the voltage curve, which will be used to produce the high voltage necessary for the picture tube as will be shown later.

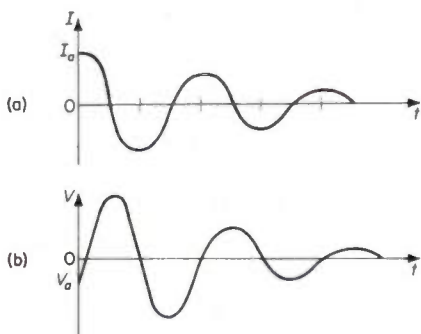


Figure 17-9. Voltage and current damped oscillations.

The exact shape of the damped oscillations depends on the damping obtained. For little damping, i.e., a high value of shunt R , the oscillation amplitude decreases slowly.

For a large amount of damping, i.e., shunt resistance below critical value, no oscillation is produced and the transient waveform has the shape illustrated in Fig. 17-10. It will have minimum duration at critical damping.

Whether the flyback transient is oscillating or not, however, the main problem is that of duration. If its duration exceeds the allowed retrace time, the last part of the transient will be superposed on the sawtooth current

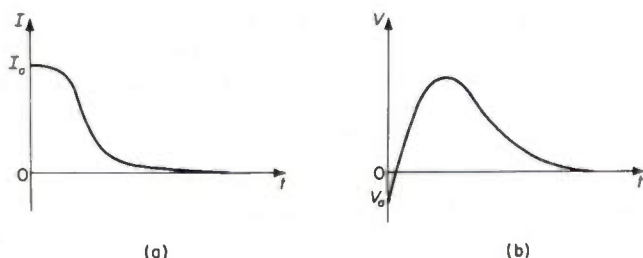


Figure 17-10. Waveforms at critical damping.

provided by the tube and will produce the waveform of Fig. 17-11, resulting in an intolerable distortion of the horizontal sweep. Specifically, Fig. 17-11 applies to an oscillating transient, but even a nonoscillating transient will produce a marked sweep distortion.

17-4. Damping

A simple way to eliminate this difficulty would be to shunt the circuit with a resistor whose resistance value is chosen to bring it just to critical damping.

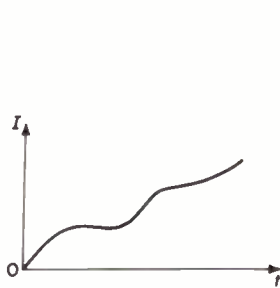


Figure 17-11. Deflection current distortion.

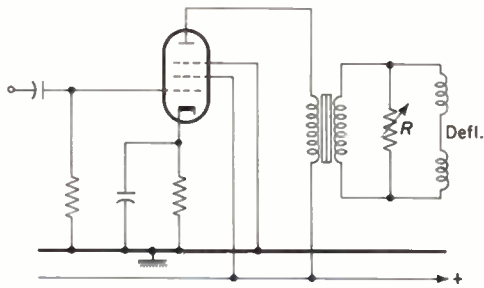


Figure 17-12. Damping by resistance R .

This minimizes the transient duration. If the duration is still too long, the inductance can be reduced, but this reduces the deflection efficiency at the same time.

Such a circuit is shown in Fig. 17-12. The resistance R has a value of a few thousand ohms and is adjustable in order to obtain critical damping. Since it dissipates considerable power, it requires a fairly high power rating.

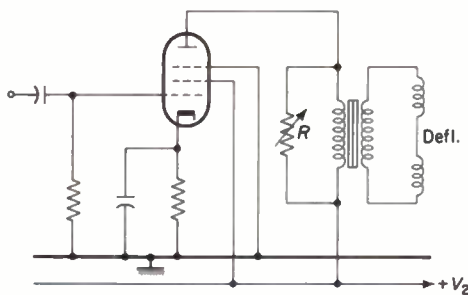


Figure 17-13. Damping across the primary of the transformer.

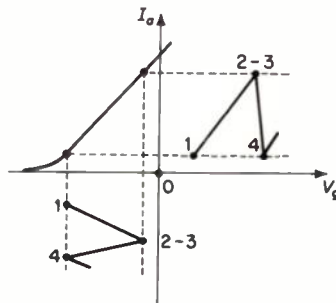


Figure 17-14. Class A tube operation.

The variable damper could equally well be shunted across the primary of the transformer, as shown in Fig. 17-13, with the same results. In this case it should have a higher ohmic value. Also, since voltage transients are larger in the primary than in the secondary, insulation problems may arise.

In this mode of operation, the tube conducts and provides current during the whole scanning cycle. This is Class A operation, as shown in Fig. 17-14.

At the end of the linear sweep, let I be the plate current through the transformer inductance L . The energy accumulated in L is $\frac{1}{2}LI^2$ and will be dissipated in the variable resistance, R .

Assume typical values $L = 0.003$ h and $I = 1$ amp. Then

$$\frac{1}{2}LI^2 = 0.0015 \text{ per cycle}$$

For 15,750 cps, the power dissipated per second is

$$15,750 \times 0.0015 = 23.6 \text{ watts}$$

approximately. Such a value is considerable and the damping system as shown in Fig. 17-13 is almost never used for this reason.

17-5. Tube operation

It is, nevertheless, enlightening and important to investigate deeper into the mode of tube operation if linear horizontal sweep is to be obtained. The anode

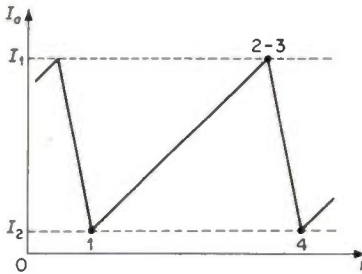


Figure 17-15. Plate current variation.

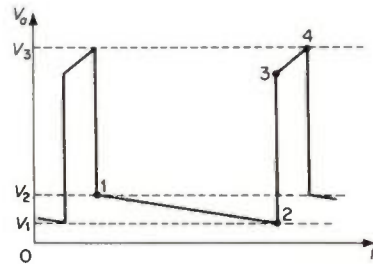


Figure 17-16. Plate voltage variation.

current must be the linear sawtooth shown in Fig. 17-15, which will be obtained for a linear sawtooth grid input to the tube. The corresponding plate voltage would be a rectangular wave for reasons described in Sec. 17-3. When the small series resistance in the deflection circuit is taken into account, a small proportion of sawtooth is added to the rectangular wave, and the anode voltage looks as shown in Fig. 17-16.

The linear relation between grid drive and plate current resulting from Class A operation has already been shown in Fig. 17-14. Points 1, 2, 3, and 4 in Fig. 17-14 identify the same points in the cycle for all the subsequent figures.

At the beginning of the sweep at point 1, the anode voltage is equal to supply voltage V_2 , shown in Fig. 17-16. As anode current I_a increases as shown in Fig. 17-15 during sweep, anode voltage decreases down to value V_1 at the end of sweep, at point 2.

At this instant retrace is triggered. At point 3, anode voltage jumps to a high value, where it stays approximately during retrace, reaching a maximum value V_3 at point 4. The corresponding operation diagram shown in Fig. 17-17, represents the cycle on the anode characteristic.

The last three figures indicate several interesting points. The tube is subjected to high positive peak voltages, capable of reaching several times the supply voltage. Insulation must be established accordingly in the anode circuit. A maximum current I_1 flows precisely when the anode voltage is at minimum V_1 , and the tube may have to be especially designed to meet this requirement. Finally, during retrace, the internal resistance of the tube reaches high values and the tube is practically incapable of damping the circuit.

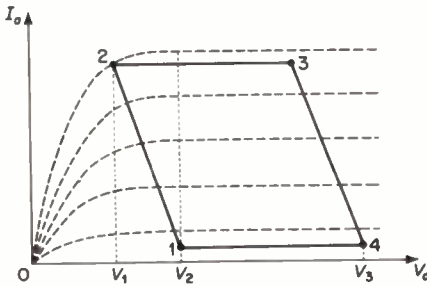


Figure 17-17. Operating cycle of the tube.

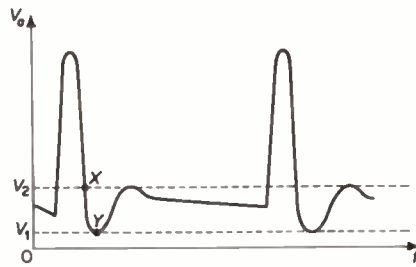


Figure 17-18. Sinusoidal retrace.

17-6. Parasitic oscillations

We may now turn to the problem of parasitic oscillations. The simplified explanation previously illustrated totally neglected the effect of the parasitic shunt capacitance which is particularly important during retrace. When retrace is triggered, the deflection circuit is shocked into oscillation, so that the anode voltage is not rectangular as assumed and shown in Fig. 17-16, but sinusoidal, as shown in Fig. 17-18. The first positive alternation is, for all practical purposes, not damped, because the tube with high plate voltage has high plate resistance. The first negative alternation is damped, however, because at a low value of plate voltage the tube has low plate resistance. After that, the oscillation is damped quickly enough by the tube resistance but might last long enough, nevertheless, to exceed the allowable retrace and superpose undesirable distortion on the deflection current sawtooth as shown in Fig. 17-19. In fact, this is usually the case with high-efficiency deflection and once again parasitic oscillations must be eliminated only during the horizontal sweep period and not necessarily during retrace.

17-7. Diode switch

This last remark gives a clue to a possible solution. Parasitic oscillations actually need to be damped out only during sweep, the shape of the retrace being of no interest. Theoretically, this could be done as shown in Fig.

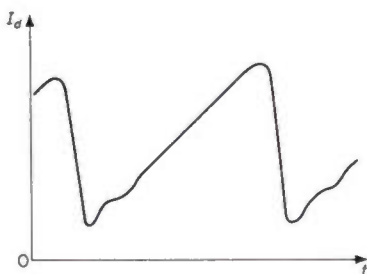


Figure 17-19. Oscillatory distortion of the sweep current.

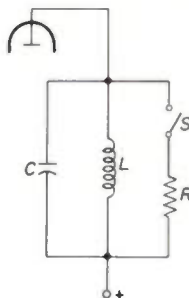


Figure 17-20. Switching of the damping resistance R .

17-20, with the help of a switch S which would connect damping resistor R across the deflection circuit at the right time only. The switch should operate automatically every scan period, and a simple version uses a diode as a switch as shown in Fig. 17-21.

The diode anode is at supply potential V_2 through damping resistance R . The diode cathode is at the same potential as the power tube anode.

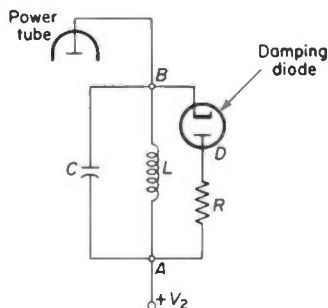


Figure 17-21. Automatic diode switch.

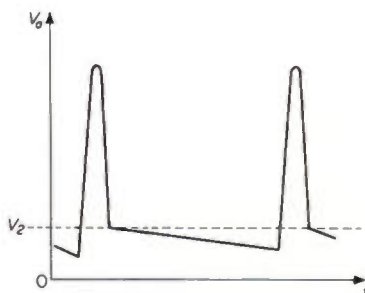


Figure 17-22. Deflection voltage.

Referring back to Fig. 17-18, the diode cathode is at a high positive potential during retrace. The diode is cutoff, and the circuit oscillates freely.

At point X , the diode cathode goes below V_2 . The diode is properly biased and conducts. R is switched across the circuit at point X and kills the oscillation.

The resulting current variation is given in Fig. 17-22. Parasitic oscillations have disappeared during the sweep. Retrace has a sinusoidal waveform, which is as good as any, and lasts approximately a half-cycle of the LC oscillation period.

There still is considerable power stored in L during each cycle and dissipated in damping resistor R . Could this power be recuperated to do useful work instead of being wasted? Obviously, in order to be useful it must be put to work rather than dissipated.

17-8. Recuperation

Power recuperation is possible at the price of a minute modification of the preceding circuit. The addition of capacitance C_1 across R as shown in Fig. 17-23 changes the operation notably. The circuit becomes similar to

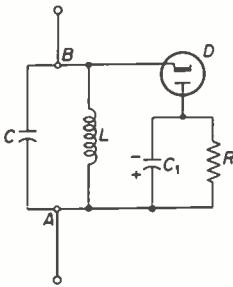


Figure 17-23. Addition of capacitance C_1 .

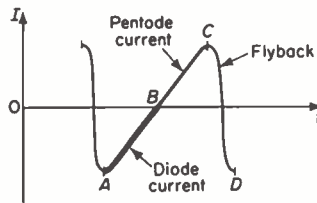


Figure 17-24. Tandem operation of diode and pentode.

that of the peak voltmeter for negative peaks, or of the restoration diode as well.

Capacitance C_1 charges up to the peak negative voltage value across the circuit shown in Fig. 17-23. Since point A is at fixed supply voltage V_2 and since point B has a peak negative value V_1 (shown in Fig. 17-18), capacitance C_1 charges to a voltage V_L such that

$$V_L = V_B - V_A = V_1 - V_2$$

which is negative, so that the charge on C_1 has the polarities indicated on the diagram.

The operation of the circuit is as follows: At the instant that V_a goes through a minimum V_1 , that is, at point Y of Fig. 17-18, the cathode of the diode reaches the same potential as the anode. The diode opens up and conducts, connecting C_1 across L . Thus, the charge voltage V_L is applied across L . If the values are correct, V_L remains constant, and since there is a constant voltage across L , the current I_a through L must vary linearly.

The magnetic energy stored in L at the end of the sweep cannot produce oscillations and is dissipated in the form of a current which decreases linearly to zero through the diode as shown in Fig. 17-24.

The slope of this current is determined by the voltage V_L across the deflection coils, but then this same voltage V_L is obtained across the coils when the pentode provides current. If working conditions are adjusted in such a way that the pentode begins to pass current when the diode ceases conduction, the mode of operation shown in Fig. 17-24 results.

From *A* to *B*, the diode provides current and cuts off at *B*. From *B* to *C*, the pentode provides current and cuts off at *C*. From *C* to *D*, neither diode nor pentode conducts. The circuit oscillates freely along a half-sinusoid, down to point *D* where the diode conducts and the cycle repeats itself.

The capacitance is used as a reservoir, restoring during the first part of the cycle the energy it accumulates during the second part. If there were no losses, all the energy provided by the pentode to the coil and stored in the magnetic field would be recovered. The *LC* circuit oscillation would not be damped, and its negative peak *D* would have the same amplitude as its positive peak *C*. The circuit would operate without consuming power.

Needless to say, the above is an idealized case, but it nevertheless shows that this type of operation may be obtained with little expenditure of power. In other words, recuperation circuits lead to a considerable economy in power consumption.

17-9. Shunt diode

Figure 17-25 illustrates a recuperation circuit in which diode and pentode are directly shunted across each other, with opposite polarities. The control voltage on the grid of the pentode is such that the tube provides current I_p during the first part of the scanning cycle. Then the pentode is cut off, and *LC* executes a half-sinusoid free oscillation. After that, current I_R flows through the diode in the opposite direction, providing the second half of the sweep and restoring that energy to the power supply (represented by a battery) which was borrowed from it by the pentode.

If there were no losses, the efficiency would be 100 per cent, and the circuit would work with zero over-all power consumption but this is hardly possible.

Moreover, during both parts of the sweep, deflecting coil *L* is connected across the power supply, either through the pentode or through the diode. A constant voltage *V* is thus maintained across the coil and, neglecting resistances, this means that current varies linearly.

Figure 17-26 illustrates how a complete cycle results. In (a), a positive current I_p is provided by the pentode. In (b) the circuit executes a free half-oscillation, providing current I_{LC} . In (c) a negative current I_R flows through the diode.

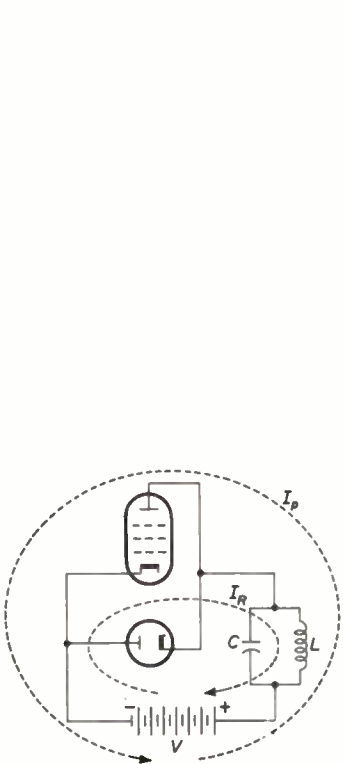


Figure 17-25. Shunt diode operation.

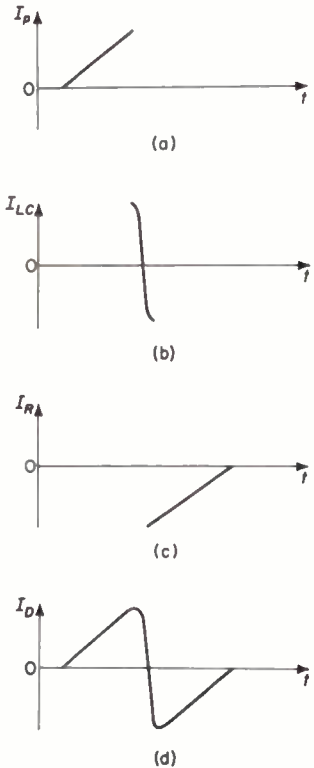


Figure 17-26. Components of total deflection current.

In (d) the sum of the three preceding components shows a complete cycle of the deflection current I_D .

17-10. Power tube control

The diode automatically switches on and off during the above cycle at the proper time. The power tube, usually a pentode, must be controlled by a suitable waveform applied to its grid. Since it works only during part of the cycle, it does not operate in Class A any longer, and its efficiency is improved.

A common arrangement is to obtain bias by grid rectification. A resistance may then be provided in the cathode as a safety precaution in case of grid drive failure. The circuit shown in Fig. 17-27 evidently works as a d-c grid restorer, with the result that the positive peaks of the grid voltage are aligned on zero volt, as

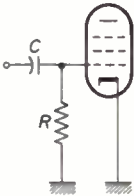


Figure 17-27. Grid control of the output tube.

shown in Fig. 17-28. Then, adjusting the amplitude of the grid drive will control the conduction period T of the pentode as shown.

A sawtooth grid drive as indicated on Fig. 17-28 may be unsuitable for the following reason: During flyback, a very high positive voltage appears

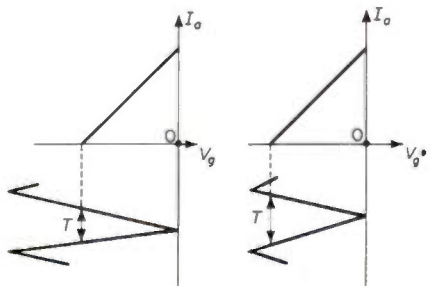


Figure 17-28. Adjusting the working cycle of the pentode.

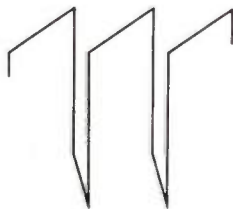


Figure 17-29. Complex grid drive waveform.

at the anode. If the tube is to be held at cutoff, a large negative grid voltage must be applied during flyback.

Now in the best case, the pentode in theory operates for half the sweep. Consequently, the grid sawtooth amplitude is twice cut off and the resulting negative peak value is insufficient to hold the pentode at cutoff.

To remedy this, the sawtooth drive is replaced by the sawtooth-plus-rectangular waveform shown in Fig. 17-29, which reaches high negative values during retrace. The circuit to obtain this waveform is recalled in Fig. 17-30. It is the familiar peaking (or shape, form, or drive) circuit.

By adjusting the value of R , the proportion between sawtooth and rectangular components is modified, which is a simple way of controlling the tube working period T of Fig. 17-28.

17-11. Type of operation

As a matter of fact, there is no need for the pentode to take over at the precise instant when the diode cuts off. This is really a limiting case of operation. At the other extreme, both might conduct for the whole duration of the sweep, as shown in Fig. 17-31. If the pentode current I_p is linear and if the diode current I_R is linear, their combination provides a resulting linear deflection current I_D . Moreover, if I_p and I_R are not quite linear, a fairly linear sum current I_D may still be obtained by a compensating adjustment of the working conditions.

Between the two extreme cases of sequential working (shown in Fig. 17-24) and simultaneous working (shown in Fig. 17-31), all intermediary cases of overlapping are possible.

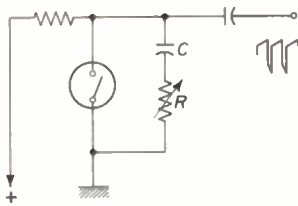


Figure 17-30. Peaking resistance R .

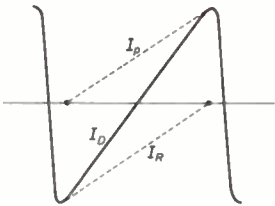


Figure 17-31. Simultaneous operation of diode and pentode.

17-12. Efficiency

Losses are unavoidable in actual practical circuits, so that the deflection circuit oscillation is damped somewhat. Its negative alternation will be smaller than its positive alternation. All the energy cannot be recovered, with the result that the efficiency is lower than 100 per cent and the circuit requires power. It is of interest, however, to reduce all losses as far as possible in good designs.

Figure 17-32 illustrates the power balance for the sequential type of operation. The top shaded area is equivalent to energy provided by the

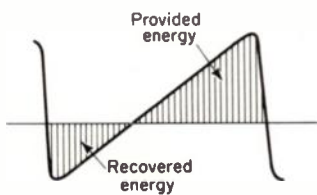


Figure 17-32. Sweep system efficiency.

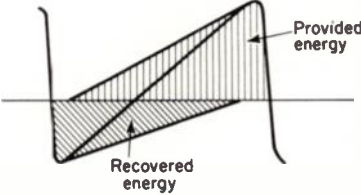


Figure 17-33. Energy recovery.

pentode. The bottom shaded area is equivalent to energy recovered through the diode. The difference between the two areas is the power absorbed by the losses.

Figure 17-33 gives the same diagram for simultaneous operation. It shows clearly that the most economical type of operation is the sequential mode of Fig. 17-32 since less energy is being lost.

We should point out that the amplitude ratio of the negative to the positive alternation of the flyback is equal to the square root of the damping factor α of the circuit, which is

$$\alpha = \frac{R}{2L}$$

where R is the resistance of the deflection coils and L their inductance.

For good efficiency, the negative alternation must be as nearly equal to the positive alternation as possible. That is, α should be small, which means that the resistance R should be small and the inductance L large.

17-13. Tube as a switch

The power tube could also be used as a switch. That is, the grid voltage would only turn the tube on and off, without controlling the shape of the anode current. In this case, anode current would vary in a way determined only by the anode load. Since this load is essentially inductive, the current would vary in a *nearly* linear fashion.

More precisely, current would increase exponentially with a rate determined by the time constant L/R of the anode load.

The exponential curve would be more nearly linear for a long time constant, so that L should be large and R small.

As previously indicated, nonlinearity of the pentode current could be corrected by suitable nonlinearity of the diode current.

17-14. Linearity

This entire question of linearity is worth a closer investigation. Because of losses, recuperation of power is never complete, and the junction point of pentode and diode currents in the high-efficiency sequential operation occurs somewhat to the left of the middle of the sweep, as shown in Fig. 17-32. If the two currents do not join correctly, there is a short variation of horizontal

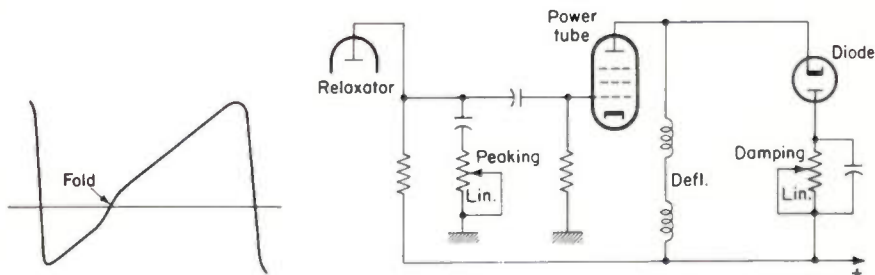


Figure 17-34. Distortion at current switching point.

Figure 17-35. Two possible linearity adjustments.

sweep speed, as shown in Fig. 17-34, which appears on the receiver screen as a vertical white bar or fold, slightly to the left of the center.

The position of the fold gives an estimate of efficiency. The nearer it is to the center, the higher the efficiency.

A smooth junction between diode and pentode currents may be insured by adjusting one of the currents or both, as shown in the circuit of Fig. 17-35.

Pentode current may be adjusted, for example, by the peaking control. Diode current may be adjusted by the damping resistance control. The first adjustment is preferred because it is more economical both in component requirements and efficiency.

Actually, the operation of the diode-pentode team has been somewhat idealized in Fig. 17-34 for clarity. The pentode current does not begin suddenly, but progressively, and is not linear. Similarly, diode current is not linear and does not cut off abruptly. In actuality, sequential operation would look more like that shown in Fig. 17-36. Fortunately, the curvatures

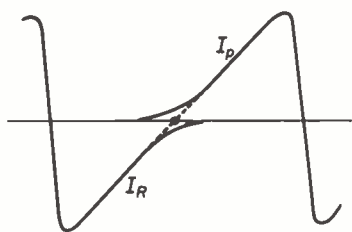


Figure 17-36. Nonlinearity of diode and pentode currents.

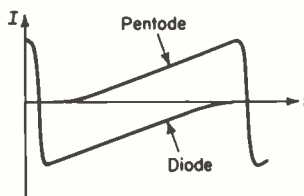


Figure 17-37. Linearity adjustment is less critical for simultaneous diode and pentode operation.

are opposite, and a good amount of compensation can be obtained by a judicious choice of working conditions. In particular, the figure shows that the pentode should switch on slightly before the diode switches off.

This horizontal linearity adjustment is somewhat critical in the above-illustrated case of sequential operation.

For simultaneous operation, shown in Fig. 17-37, adjustment is much less critical for two reasons: Curvature occurs at the extremities of the sweep and when one of the currents has maximum nonlinearity, the other one is almost linear and vice versa. Moreover, there is no necessity for both currents to have the same slope. All that is needed is that they both be approximately linear. This is true only for 100 per cent simultaneous operation. For all other modes up to sequential working, the *slopes* of the two currents must be as nearly equal as possible.

17-15. Damping resistance

How is it possible that the diode current can be made to have the same slope as the pentode current?

To find out, the operation of the diode circuit will be examined in detail with the help of Fig. 17-38. Sequential operation is assumed, so that the

pentode may be neglected during the working period of the diode. Also, the circuit is grounded for ease of understanding.

As previously explained, C_1 charges up to the value V_L of the peak of the first negative alternation of the flyback oscillation as shown in Fig. 17-39. During the pentode working period, however, C_1 discharges somewhat through R , so that at the next flyback the diode anode is not quite at V_L , but at a slightly smaller value of negative voltage. In this way, when the cathode is brought to V_L , it becomes negative with respect to the anode and the diode

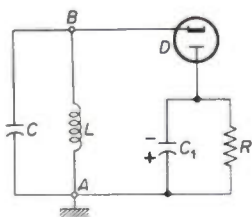


Figure 17-38. Resonant circuit and damping diode.

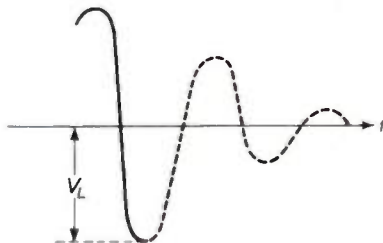


Figure 17-39. Retrace oscillation.

conducts. C_1 charges up again to voltage V_L , so that then the diode should cut off. But it does not!

What keeps it from reaching cutoff?

As shown in Fig. 17-38, it is the presence of R across C_1 . The capacitor discharges through the resistor with time constant RC_1 , and with the anode voltage falling, the diode remains conductive.

An equilibrium is reached automatically when the diode passes a charge current such that it exactly compensates for the discharge current through R .

Stated another way, the current flowing through the diode, which is the deflection current, varies exponentially in a fashion governed by time constant RC_1 . Adjusting R then controls the linearity.

If a nearly linear diode current is desired, therefore, time constant RC_1 should be made much longer than the sweep period, say by a factor of 10. With a capacitance of 0.1 microfarad, this would lead to a resistance value of 5000 ohms, roughly. By reducing this value, the curvature of the diode current might be adjusted for better over-all linearity as previously explained.

17-16. Diode connection

For a high-impedance circuit, the diode connection is as shown in Fig. 17-40. The diode is directly shunted across the deflection circuit. Its cathode receives high-amplitude positive pulses during flyback and must be insulated accordingly. A special heater transformer could be used, with high-voltage insulation between primary and secondary. The diode heater would have

one end connected to the cathode, so that there would be no cathode-heater insulation problem. The difficulty is that the heater transformer should have a very low secondary parasitic total capacitance, since this capacitance in effect appears across the deflection circuit where it is most unwelcome.

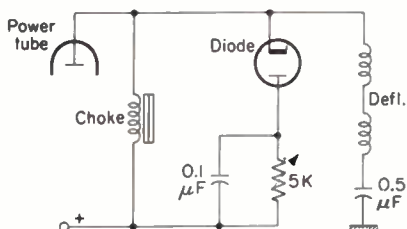


Figure 17-40. High-impedance deflection circuit.

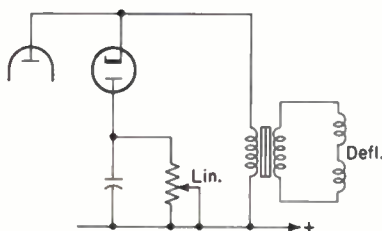


Figure 17-41. Damping diode across the transformer primary.

A better solution lies in the use of especially developed diodes that have sufficient insulation between cathode and heater. To improve matters, the cathode is frequently connected to a well-insulated cap at the top of the tube. The diode should also have a very low internal resistance.

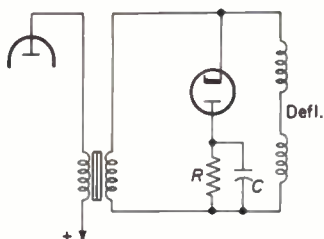


Figure 17-42. Damping diode across the transformer secondary.

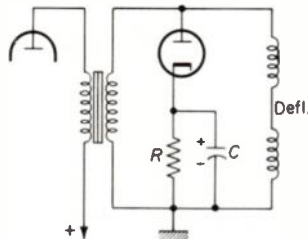


Figure 17-43. Grounding one end of the secondary.

The same diagram could be used for low-impedance deflection coils, with the diode across the primary of the transformer as shown in Fig. 17-41. It is much more convenient, however, to connect it across the secondary as shown in Fig. 17-42 because transient voltages are much lower and also because d-c levels are no problem. The whole secondary circuit can be set at ground level or at any other convenient voltage. The diode should again have a very low internal resistance, even more than when it is connected across the primary because it is shunting a lower impedance. Several tubes have been developed for this purpose.

The positions of the diode and the RC circuit may be interchanged as shown in Fig. 17-43. In this case, a positive voltage accumulates across C , which may be of some value.

17-17. Boosted voltage

This positive voltage developed across C in Fig. 17-43 can be put to good use and added to the existing $B+$ to provide a higher voltage supply. The low end of the secondary is connected to $B+$, say 200 volts. If a voltage of 100 v exists across C as shown in Fig. 17-44, the voltage between ground and point

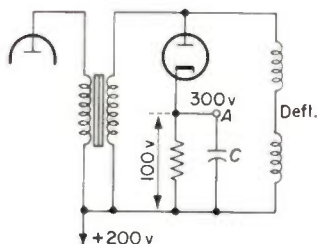


Figure 17-44. Voltage developed by the damping diode.

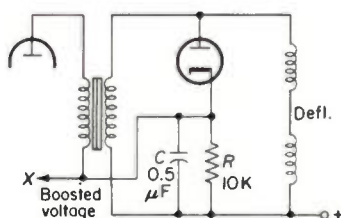


Figure 17-45. Boosted voltage production.

A reaches 300 v. This boosted voltage can be used advantageously for power tubes requiring a high voltage supply and for the horizontal line output tube as well. The diagram then takes the form shown in Fig. 17-45. Other circuits may be connected at point X , such as picture tube first anode,

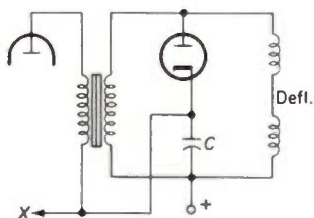


Figure 17-46. Auxiliary circuits fed on boosted voltage.

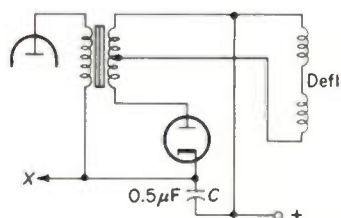


Figure 17-47. Special transformer ratio for the diode.

vertical power tube, relaxators, etc. All these circuits absorb power and behave as resistances connected across capacitor C . If their total equivalent resistance reaches the optimum value for R , say 10 kilohms, there is no need for a physical resistance across C and the circuit takes the form shown in Fig. 17-46. In fact, the total equivalent resistance of the circuits connected to point X need only be such that the total resistance is below the critical value of R . In this case, linearity is still good—perhaps even better—but efficiency decreases.

To bring efficiency up to a maximum possible value, the diode may be driven by a voltage different from the voltage across the deflection coils.

This simply means a different tap on the secondary winding, as shown in Fig. 17-47, and results in a higher boosted voltage. It is not uncommon for boosted voltage to reach twice the $B+$ voltage or even more.

17-18. Diode at primary

The diode could equally as well be connected to a tap on the primary, as shown in Fig. 17-48, by inverting its polarity. Cathode-to-heater insulation should be sufficient in this case to withstand pulse voltages at the tap. The

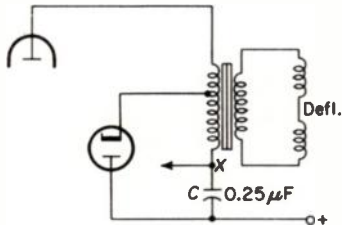


Figure 17-48. Primary connection of the damping diode.

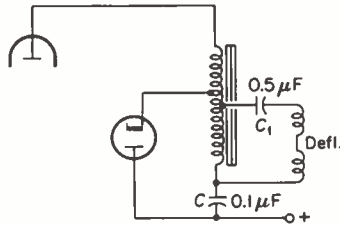


Figure 17-49. Autotransformer horizontal output.

boosted voltage appears at the top of C at point X indicated and is used, as before, to feed the line output tube and eventually other circuits.

In fact, this arrangement is mandatory if an autotransformer is used instead of a transformer in order to couple the deflection coils to the power tube, as shown in Fig. 17-49, as is frequently done. Capacitor C_1 eliminates d-c current through the coils.

Adjustment of optimum working conditions in these diagrams reduces to determining experimentally the best tap for the diode.

If the capacitor C has a value slightly too low to smooth out the boosted voltage entirely, there remains a small ripple at line frequency which modulates the line output tube plate voltage and consequently the horizontal deflection. By proper choice of capacitor C , this can be made to compensate satisfactorily for S distortion without further complication.

17-19. Output transformer

The horizontal output transformer is undoubtedly the most critical single element in a television receiver. It must provide correct matching of the various impedances; that is, the turn ratios of the various windings must be correct. It must be extremely well insulated in view of the high pulsed voltages existing during flyback. This insulation requirement is all the more important when these pulse voltages are used to generate the tube high voltage, which is common practice as will be seen later. The transformer, in addition, must

not distort waveforms, which is equivalent to saying that it should handle equally well all frequencies up to the 20th harmonic of line frequency, that is, roughly up to 300 kc.

The transformer windings, furthermore, should have minimum ohmic resistance and minimum parasitic capacitance.

Iron and copper losses should be at a minimum and coupling coefficient at a maximum.

This impressive list of requirements would make the design of a good horizontal line output transformer an arduous task, if it were not for ferrite cores. Their good frequency performance and high permeability make them ideally suited to the purpose.

Using ferrite cores, the coupling coefficient should reach at least 0.98. A lower figure leads to prohibitive leakage inductance values. The decrease in efficiency is secondary. The real drawback in this case is that leakage inductance resonates with various parasitic capacitances and creates troublesome oscillatory transients that are practically impossible to eliminate.

17-20. Horizontal centering

Horizontal centering may be obtained by means external to the deflection coils such as focus-assembly tilting, centering magnets, etc.

Electrical centering circuits are all based on providing a variable d-c current through line coils. For example, the simple arrangement shown in Fig. 17-50 uses the line output tube plate current. This technique is commonly found in several variations.

Centering circuits analogous to those given for vertical deflection could also be used but are rarely required in view of the simplicity of the basic circuit shown in Fig. 17-50.

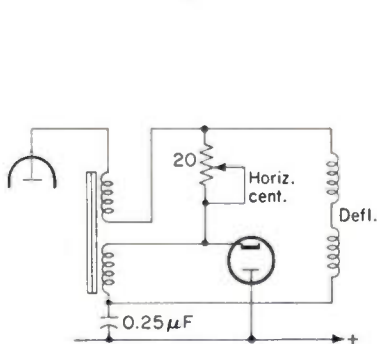


Figure 17-50. Electrical centering adjustment.

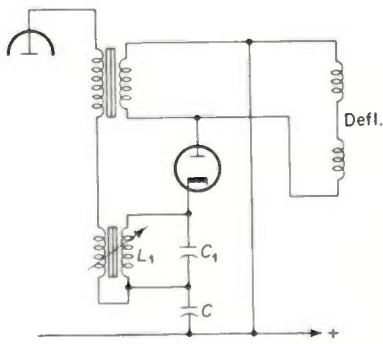


Figure 17-51. Linearization by tuned circuit L_1C_1 .

17-21. Linearity adjustment

Intrinsically, the linearity of horizontal sweep circuits using a pentode-diode combination is reasonably good if proper design care has been exercised. One of the causes of nonlinearity is attributed to the diode itself. When the diode is conducting, the voltage across deflection coils L in the basic diagram shown in Fig. 17-38 is equal to the voltage across C_1 plus the voltage drop across the diode. Assuming the voltage across C_1 to be constant, linear variation of current will be obtained if the voltage across the diode is constant. However, precisely because the current through the diode varies from a maximum value to zero, constant voltage across it could be obtained only if its resistance varied inversely to the current. Several circuits may be used to obtain a reasonable approximation of the wanted variation. In Fig. 17-51, a resonant circuit L_1C_1 approximately tuned on horizontal frequency is inserted in the diode cathode circuit. The voltage produced across L_1C_1 is adjusted in phase and amplitude for best compensation of diode nonlinearity.

Another common circuit is reproduced in Fig. 17-52. In this diagram, the capacitor in the diode cathode circuit has been split into two capacitors C_1 and C_2 , used at the same time to tune the inductance L_1 to 15,750 cps.

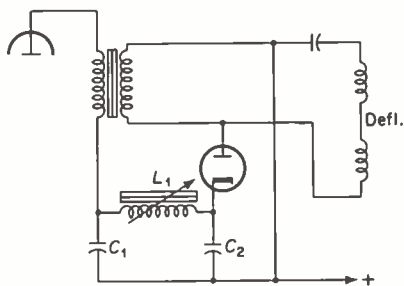


Figure 17-52. Split capacitance C_1C_2 .

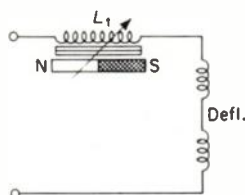


Figure 17-53. Linearity adjustment by saturated inductance.

The operation of this circuit is identical to that of the preceding circuit. The necessary phase shift is obtained by the split capacitance now instead of being produced by the phase relations between primary and secondary of a transformer.

Another method, basically different yet accomplishing the same purpose, is illustrated in Fig. 17-53. Here the nonlinearity due to the diode is accepted as such but compensation is introduced between diode and deflection coils. The operation is based on a saturated inductor L_1 , whose inductance value varies with the current flowing through it. Saturation is obtained by a

permanent magnet, represented as shown. Since L_1 is in series with the deflection coils, the magnitude of deflection current depends on L_1 , whose value in turn depends on the current.

For the right value of L_1 , the current amplitude variation due to saturation compensates for the variation due to nonlinearity.

To avoid the necessity of a mechanical adjustment of the permanent magnet, an electromagnet may be used as shown in Fig. 17-54. Adjusting its excitation current controls horizontal linearity.

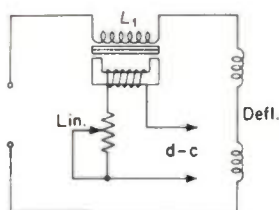


Figure 17-54. Adjustment of saturation current.

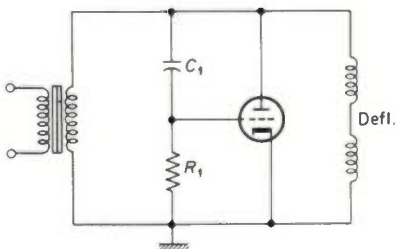


Figure 17-55. Triode damping circuit.

17-22. Damping triode

Instead of a damping diode, a triode could be used. Since its internal resistance may be controlled by the grid voltage, linearity could be adjusted without difficulty. The principle of the triode circuit is illustrated in Fig. 17-55. The triode is connected across the deflection coils as a damping diode. Its grid is driven by the voltage across the coils, partially differentiated by coupling circuit C_1R_1 . The anode voltage V_a is shown in Fig. 17-56(a). After partial differentiation, this results in the grid voltage V_g illustrated in (b). Figure 17-56 shows that at the beginning of the sweep, when current is maximum, the grid is driven positive so that the internal resistance of the tube is low. Then, as scanning progresses and current decreases, the grid becomes less positive and internal resistance goes up.

In this way, the voltage drop across anode and cathode of the tube is maintained more or less constant. Notice that the damping diode works in this circuit only during the first part of the cycle, so that grid voltage of Fig. 17-56(b) is made to vary only during the corresponding part of the cycle in the necessary fashion.

In actual practice, the circuit appears as shown in Fig. 17-57. Bias through grid current is provided by R_2C_2 , and the usual damping circuit R_3C_3 is connected in the cathode. Adjusting R_1 controls the grid drive and adjusts the horizontal linearity.

The working conditions for the triode are quite stringent. Special tubes have been and may have to be designed and economic considerations make their use rather rare in present receivers, which are usually designed around commercially available components.

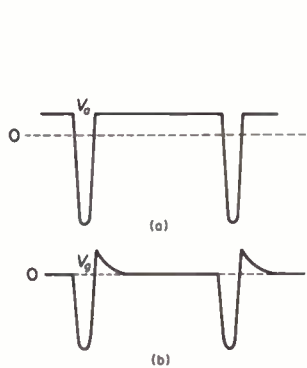


Figure 17-56. Waveforms in the triode damping circuit.

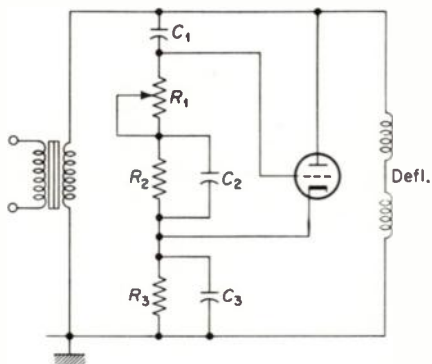


Figure 17-57. Practical triode damping circuit.

17-23. Amplitude control

Horizontal amplitude or width control is almost always applied after the output tube. Besides the reasons previously given, there is the fact that the picture tube high voltage is obtained from horizontal flyback and would be modified by amplitude controls acting on the power tube.

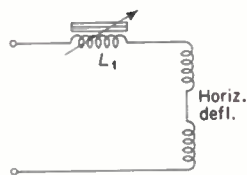


Figure 17-58. Width control by series inductance.

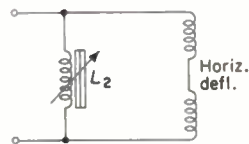


Figure 17-59. Width control by parallel inductance.

Furthermore, because of power dissipation, resistive amplitude controls are abandoned with preference for reactive controls. A simple example is given in Fig. 17-58. Inductor L_1 is placed in series with the deflection coils, and adjusting its value adjusts horizontal amplitude.

The same principle may be applied with a shunt inductor L_2 as shown in Fig. 17-59.

These two simple circuits have the same drawback in that they both modify (although in opposite directions) the total inductance—that is, the deflection circuit load.

By using both systems simultaneously, as shown in Fig. 17-60, the load variation can be minimized by suitable choice of L_1 and L_2 .

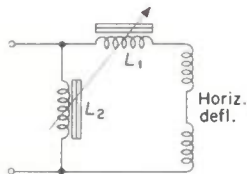


Figure 17-60. Width control by combined parallel and series inductances.

17-24. Grid drive

In many cases, no adjustment of the operating conditions of the pentode-diode pair is available in the diode part of the circuit. Hence, working conditions may have to be adjusted by the pentode grid drive. We have already seen that the peaking resistance, as shown in Fig. 17-61 can be used for this purpose.

With some forms of relaxation oscillator circuits, the time constant of the charging circuit can be adjusted. For example, as shown in Fig. 17-62, in the case of a pentode the anode circuit is virtually independent of the relaxator between control and screen grids.

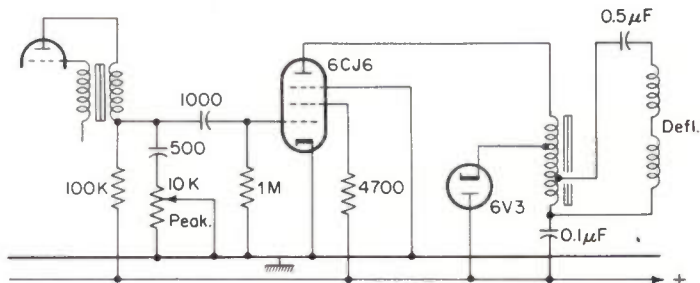


Figure 17-61. Grid drive adjustment by a peaking control.

A fairly common arrangement uses a capacitance voltage divider C_1 - C_2 to control the amount of grid drive, as shown in Fig. 17-63.

Other circuits use, in some cases, controls which unavoidably modify the picture tube high voltage on the basis that they are adjusted once and for all. Such would be an adjustment of the supply voltage to the relaxator or a variable screen grid voltage in the output tube.

17-25. EHV

The picture tube high voltage will be called *extra-high voltage* or, for simplicity, *EHV*. It is on the order of 10 to 20 kv for most tubes, and the current consumption is low, on the order of 100 microamperes. Although

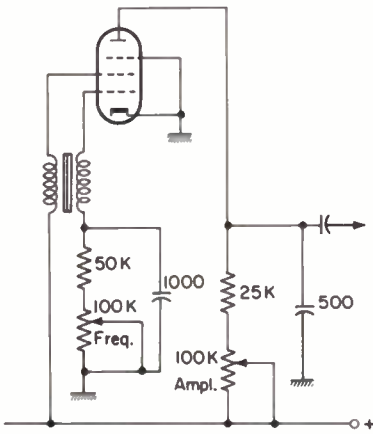


Figure 17-62. Amplitude control by potentiometer.

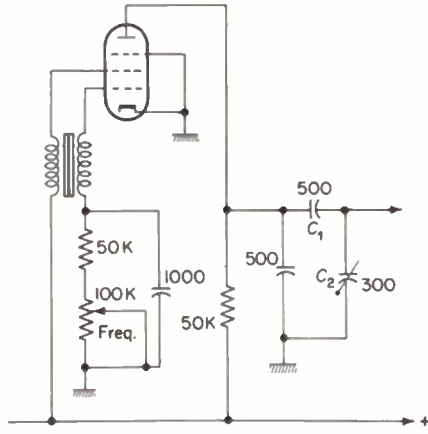


Figure 17-63. Amplitude control by capacitance C_2 .

a separate chapter is devoted to EHV generation, a few words will be said here about its almost universal production from the line output stage, because it bears on the design of the horizontal output circuits.

Voltage pulses developed across the line output transformer during flyback easily reach several kilovolts across the primary. By using an extra winding on the transformer as a step-up on the primary, the available pulse voltage is suitable for EHV use after rectification.

Because of the low EHV current requirements, a small-power diode is used as a rectifier, and its heater is fed by another small winding on the horizontal output transformer, as shown in Fig. 17-64, to simplify insulation problems. This simple and economical circuit is used virtually as such in all present-day receivers. It slightly increases, say by 2 watts, the power burden imposed on the line output stage.

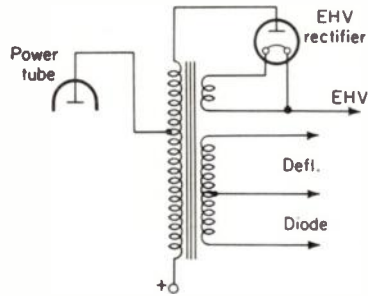


Figure 17-64. Production of the EHV output.

17-26. Self-relaxator

The usual team of relaxator-power tube may be replaced by a single tube for purposes of economy. This may pose a few problems related to synchronization, horizontal amplitude, or interdependence of adjustments, and the

arrangement, therefore, is not too popular except perhaps in certain small or inexpensive receivers where cost is a major factor.

Figure 17-65 shows a power pentode in which the cathode, control grid, and screen grid constitute the triode of a classical blocking oscillator. The

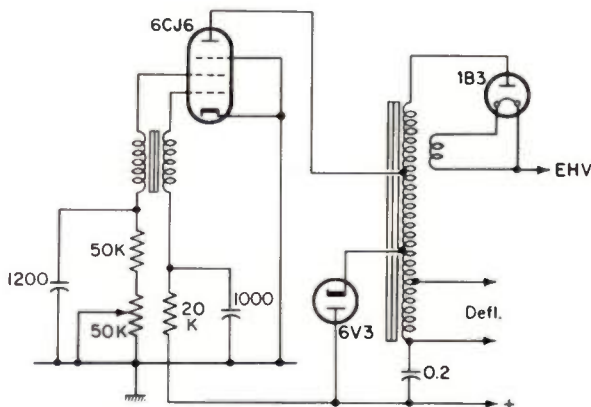


Figure 17-65. Autorelaxator line output tube.

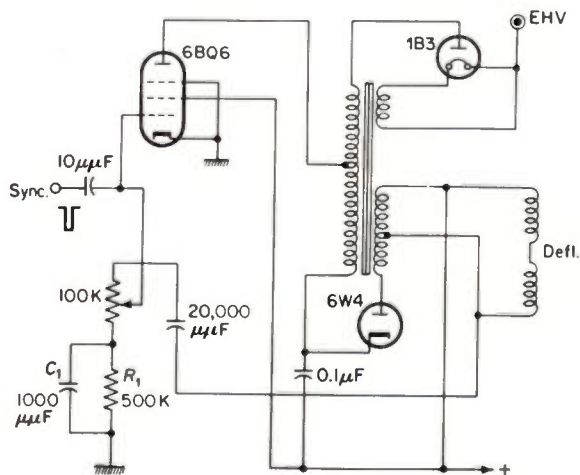


Figure 17-66. Plate-to-grid coupling of the autorelaxator.

anode circuit contains an output autotransformer feeding the damping diode, the deflection coils, and the EHV rectifier.

Figure 17-66 is a variation of the preceding diagram. Blocking oscillation is obtained with the help of the output transformer feeding back to the control grid. R_1C_1 is a circuit that improves linearity by shaping the grid voltage. The output uses a transformer with damping diode in the secondary.

A more sophisticated use of the output transformer is illustrated in Fig. 17-67 in a somewhat simplified version. The relaxator used here belongs to the purely inductive category. The necessary energy storage and release is not provided by a capacitance, but by the magnetic field in the transformer itself. In fact, the basic circuit uses no capacitance whatever. Frequency is adjusted by the grid potentiometer and synchronization signals are applied to the screen grid with sufficient amplitude.

A diagram used in actual practice appears in Fig. 17-68. Despite the screen-to-suppressor grid coupling, which recalls a transitron oscillator, oscillation is obtained by inductive coupling from the output transformer to the control grid.

Synchronization is applied to the screen grid in the form of negative-going pulses. Damping is provided simply in the case illustrated by a resistor and capacitor across the deflection circuit.

The voltage developed on the screen is applied to the suppressor grid. The coupling circuit has a variable time constant, which permits modifying the shape of the suppressor grid drive and adjusts linearity.

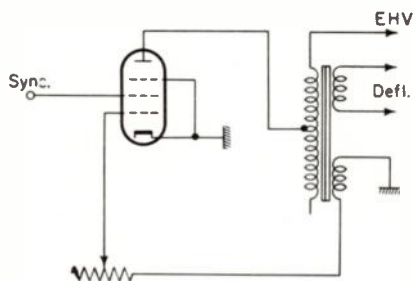


Figure 17-67. Inductive plate-to-grid coupling of the autorelaxator.

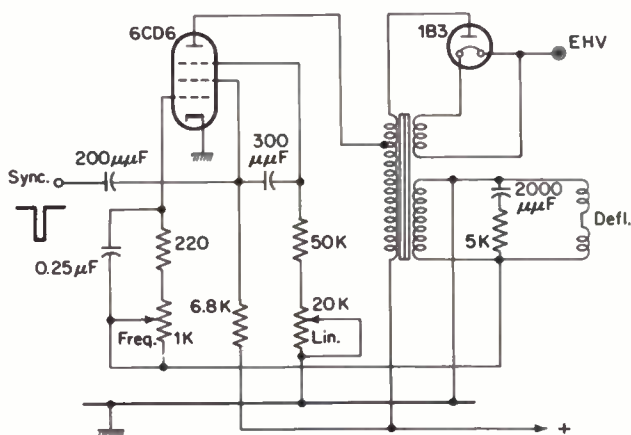


Figure 17-68. Practical circuit for an inductive autorelaxator.

17-27. Combined circuit

In a combined circuit, there is a relaxator stage and an output power tube, but the output tube is part of the relaxation circuits.

Several combinations may be used. For example, feedback may be arranged between a winding on the output transformer and a tube driving the power tube.

Another combination is shown in Fig. 17-69. The cathode, control grid, and screen grid of the output tube are used as second triode for a multi-vibrator. Crossed grid-to-plate couplings are provided. Negative-going sync

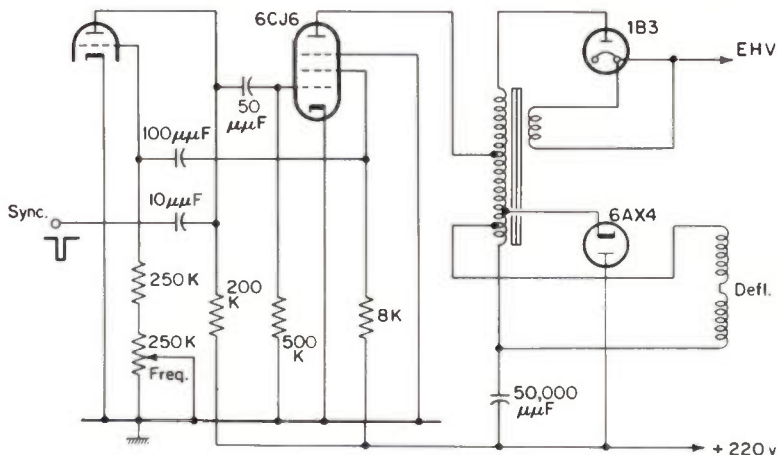


Figure 17-69. Horizontal sweep section using a combined multi-vibrator.

pulses are applied to the grid of the power tube. The output part of the circuit is conventional, including booster diode and EHV rectifier.

It should be pointed out that these circuits are not too popular. For one thing, in the example shown, the first triode could be used equally as well as a separate blocking oscillator with complete function separation and independence. For another thing, efficiency might be lower than if the output tube were used in the conventional fashion, and efficiency is extremely important in the horizontal sweep section.

17-28. Direct drive

The transformer or autotransformer associated with the line output stage is a critical and costly component. A clever circuit that eliminates this component is called the *direct drive* system; it can also be identified as a *reactive sweep* circuit. Figure 17-70 shows the basic diagram. A minimum number of components are used. Besides the power tube V_1 , damping diode V_2 , and deflection coils L_1 , it uses an inductor L and a capacitor C .

To simplify analysis, assume for the time being that L is short-circuited as shown in Fig. 17-71 and consider what happens during the increasing first

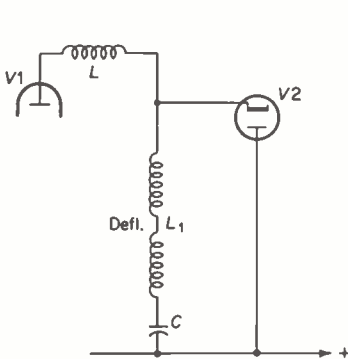


Figure 17-70. Simplified direct drive sweep circuit.

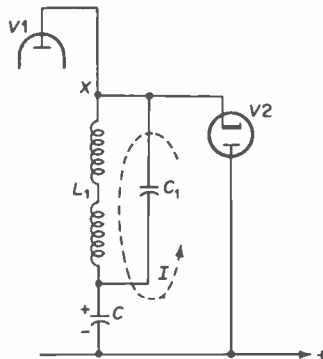


Figure 17-71. Operation of the direct drive circuit.

part of the cycle OA of Fig. 17-72. As shown, current I increases linearly from O to C . At C , grid drive voltage cuts off V_1 and the power tube becomes nonconductive; L_1 starts to oscillate on a frequency determined by its inductance value and by the parasitic capacitor C_1 . The voltage V across L_1 has the shape shown and reaches a high positive peak as usual. After a half-period, the voltage across L_1 reverses, and since point X is now negative, the diode is switched on. It short-circuits point X to $B+$ and kills the oscillation, which stops abruptly at point B shown on Fig. 17-72. The voltage at point X cannot go below $B+$ because of the damping effect of the diode.

During this time, current I has also completed a half-cycle of oscillation but with a phase difference of 90 degrees with respect to voltage. Starting from a positive value at C (shown in 17-72), it has reached negative value D .

Flowing around the circuit through the switched-on diode in the direction shown, current I charges up capacitance C with the polarities indicated and provides the second half of the sweep cycle.

Seen from the anode of V_1 , the voltage across C adds to the supply voltage to provide boosted voltage.

As described above, however, the circuit cannot work. The reason is that, owing to unavoidable losses, the oscillation of L_1C_1 is damped, so that the amplitude at point D of Fig. 17-72 is smaller than the amplitude at point C , and the sweep waveform decays and dies away.

In order for the circuit to work, the amplitude of the negative peak of current should be larger than the amplitude of the positive peak. In this way, the sweep amplitude increases until some equilibrium is reached between losses and the provided extra energy.

To accomplish this, extra energy is given to L_1 (each time it starts to oscillate) to be added to the energy stored in the magnetic field of L_1 . This additional energy is supplied by an additional inductance L . With its

parasitic capacitance C_2 , this inductor L resonates approximately at the same frequency as L_1C_1 . During retrace and neglecting V_1 which is cutoff, the equivalent circuit has the configuration shown in Fig. 17-73. C_3 is the parasitic capacitance between ground and plate of V_1 . The circuit is easily identified as two distinct capacitance-coupled tuned circuits.

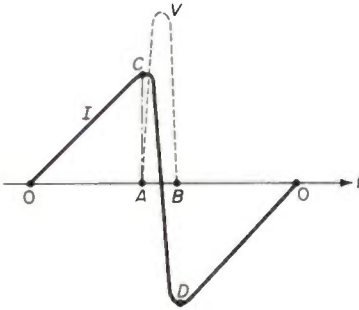


Figure 17-72. Waveforms in the circuit.

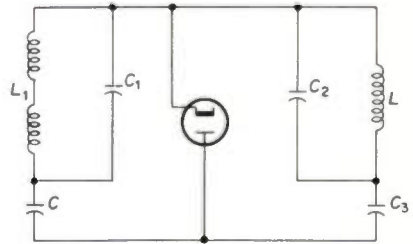


Figure 17-73. Equivalent circuit.

If the resonant frequencies are slightly different and if the currents reverse with a small phase difference, energy from L is effectively transferred to L_1 , with the result that the amplitude of the negative current peak is larger than the amplitude of the positive peak.

C has a large value, ranging typically from 0.1 to 0.5 μf . The coupling between the two tuned circuits and also in some measure the resonant frequency of LC_2 depend on the value of C_3 . Adjusting C_3 will then control efficiency by controlling the amplitude.

Notice that since the diode conducts during the corresponding part of the sweep cycle, auxiliary circuit LC_2 could just as well be connected in series with the diode. Frequently an inductance type of linearity control is placed in series with the diode and is sufficient to maintain operation without an additional inductor L .

17-29. Complete circuit

The pulse voltage on the anode of V_1 may reach as high as 4 or 5 kv, part of which appears across L and part across L_1 . The deflection coils should be well insulated, as is customary with high-impedance coils.

The circuit can be made to produce EHV for the picture tube's final anode by stepping up the 2- or 3-kv voltage across L as shown in Fig. 17-74 and rectifying it with a small rectifier. Heating power for the rectifier can also be provided by the EHV autotransformer as shown.

The EHV autotransformer can be an air-cored type, or a magnetic ferrite core can be used. A typical component, which is small and economical, is shown in Fig. 17-75.

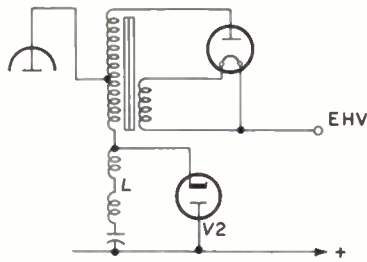


Figure 17-74. Autotransformer to provide EHV output.

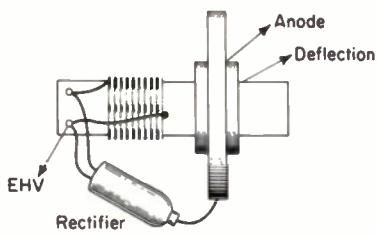


Figure 17-75. Practical construction of the EHV autotransformer.

A complete circuit is illustrated in Fig. 17-76. The damping diode is heated by a separate insulating transformer whose secondary has a parasitic capacitance to ground of $65\ \mu\text{f}$ as indicated by the dotted lines. Amplitude

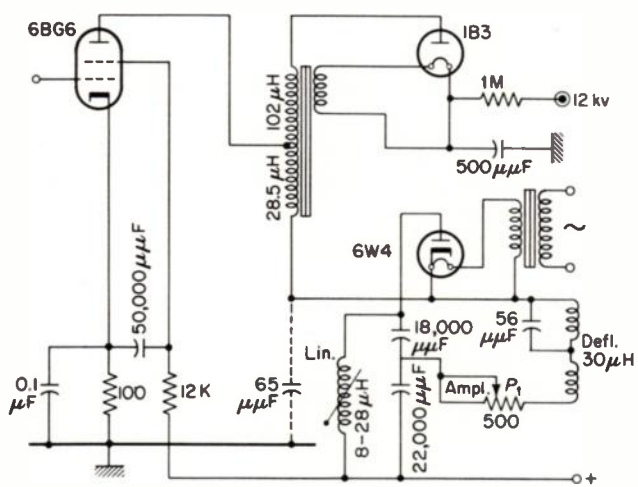


Figure 17-76. Complete circuit for direct drive sweep.

is adjusted by modifying the d-c anode voltage with P_1 . The linearity control circuit in the damping diode is a conventional practice.

Another version is given in Fig. 17-77. A high insulation voltage capacitor of $20\ \mu\text{f}$ is connected between ground and anode of the power tube; its value controls amplitude. No linearity circuit is provided. As such, linearity distortion is below 10 per cent except at the extreme left of the picture.

17-30. Practical circuits

A few horizontal sweep circuits that are typical of current practice will be given, sometimes in slightly simplified version, to illustrate application of the basic principles explained in this chapter.

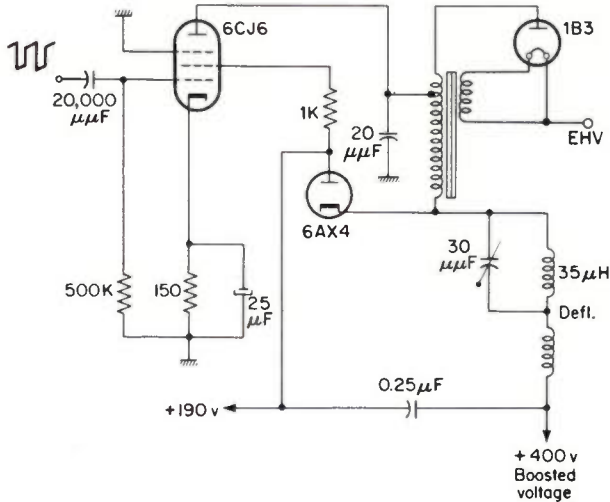


Figure 17-77. Practical direct drive circuit.

Figure 17-78 shows a straightforward and classical circuit. The low-value resistor in the control grid of the power tube is intended to kill eventual oscillations at very high frequency that are common when large-transconductance tubes are used in pulse circuits.

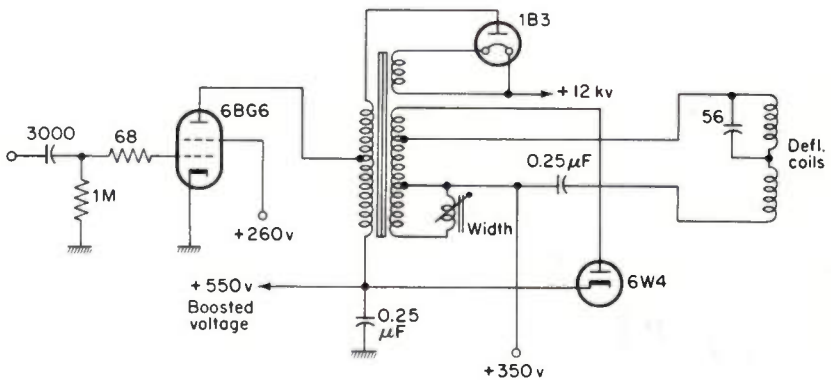


Figure 17-78. Transformer type of output circuit.

The output circuit uses a transformer, providing EHV through a step-up primary winding. The secondary combines three separate windings. One drives the deflection coils, another feeds the damping diode, and a third winding is used solely for amplitude or width control. No linearity adjustment is provided.

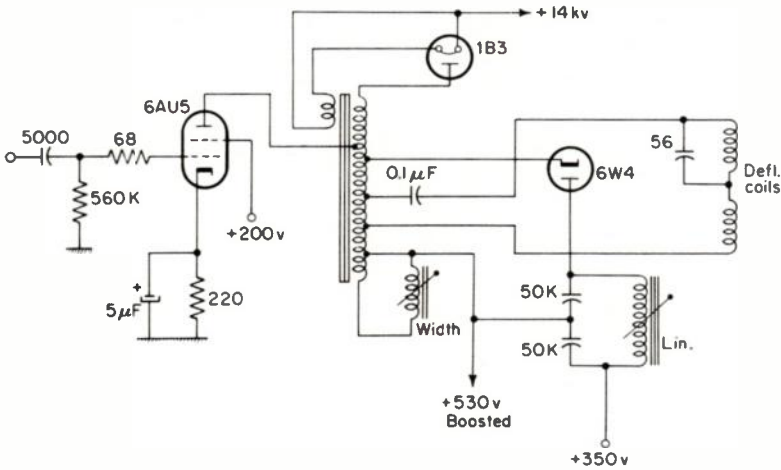


Figure 17-79. Autotransformer type of output circuit.

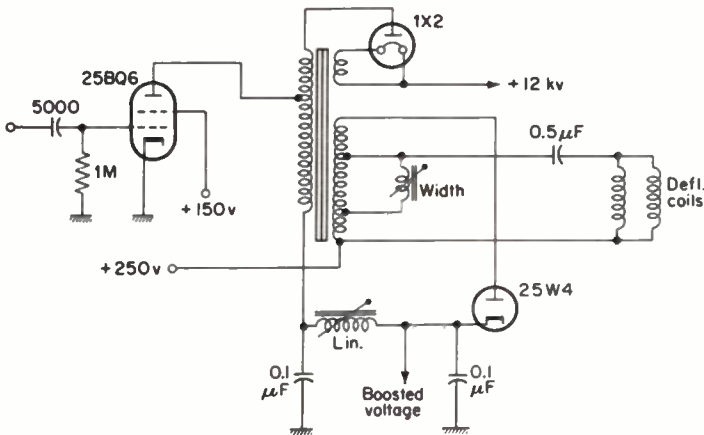


Figure 17-80. Transformer output circuit with parallel deflection coils and linearity inductance adjustment.

Figure 17-79 employs an output autotransformer, so that the damping diode connections are reversed with respect to the preceding diagram. EHV and width control circuits are identical but now the damping diode circuit contains a linearity control.

The circuit shown in Fig. 17-80 uses a-c—d-c tubes and a transformer. EHV is provided by the autotransformer winding on the primary. The secondary drives the damping diode with standard linearity control in the cathode. Two taps are provided for the width control and the deflection coils, which are connected in parallel this time.

Figure 17-81 shows a simple circuit with an autotransformer at the output. Its peculiarity is the provision of an amplitude or width control by adjusting the screen voltage on the output tube.

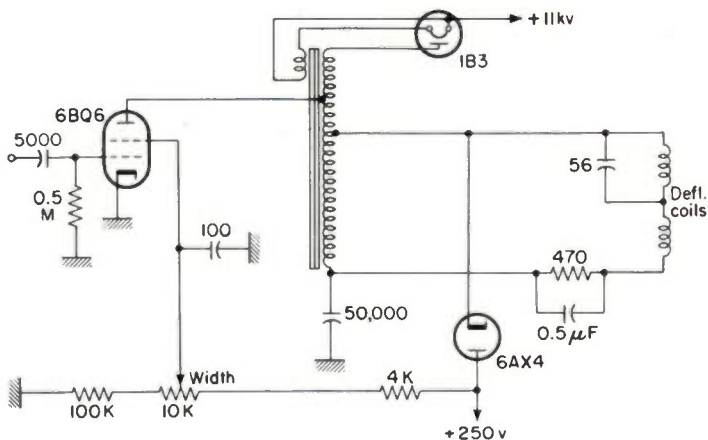


Figure 17-81. Autotransformer output with variable screen voltage.

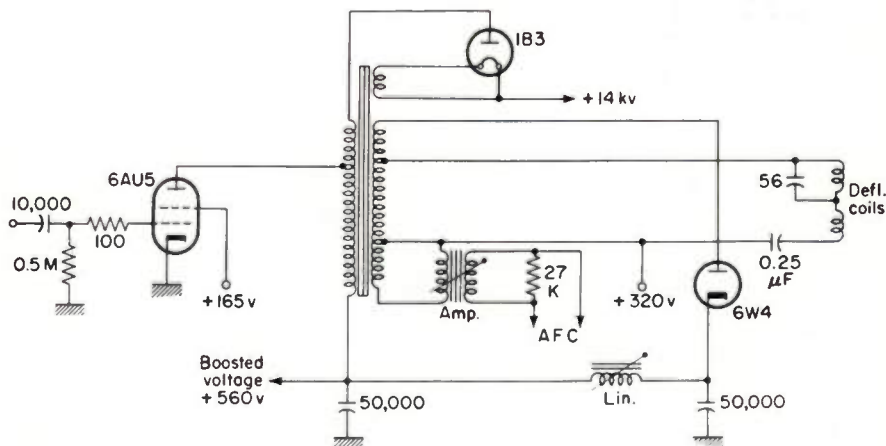


Figure 17-82. Transformer output with inductive amplitude control.

The circuit of Fig. 17-82 uses a transformer with a conventional EHV circuit. The secondary feeds the damping diode, whose cathode contains the linearity control circuit. Part of the secondary drives the deflection coils. Another part is connected to the amplitude control coil, which in this case is a transformer. Its secondary voltage is used for AFC purposes.

Problems

1. Compare high- and low-impedance deflection coil systems for horizontal sweep.
2. Explain the fundamental differences in the problem of horizontal and vertical scanning.
3. Design a linearity circuit for a high-impedance type of horizontal sweep.
4. A parallel-tuned circuit consists of inductor L , resistor R , and capacitor C in parallel. Calculate the value of R which provides critical damping.
5. An inductor L has an ohmic resistance R and is tuned by a capacitor C in parallel. Calculate the value of R providing critical damping.
6. Explain why it is not necessary for flyback to be linear.
7. A deflection current of 0.4 amp peak-to-peak flows through a 30-millihenry deflection coil. Calculate the voltage developed during retrace, assuming a flyback time of 15 per cent of the horizontal scanning period.
8. A sawtooth current of 0.6-amp peak value flows through horizontal deflection coils having an inductance of 16 millihenrys and a resistance of 80 ohms. Plot the voltage across the coils.
9. A deflection circuit consists of inductor L , resistor R , and capacitor C in parallel, and is at critical damping. Calculate, in terms of component values, the time for the flyback transient to decay to 10 per cent of its peak value.
10. A nonoscillatory deflection circuit has a transient which exceeds retrace time. Study the resulting sweep distortion and plot the sweep current waveform.
11. A deflection current of 0.09-amp peak value flows through deflection coils having a 22-millihenry inductance. Calculate the energy stored in the coils. If all this energy is dissipated in a damping resistor, evaluate the power rating of the resistor.
12. Compare Class A and Class C operation of the horizontal output tube.
13. The power tube is in effect connected across the deflection circuit. Investigate its damping effect on the circuit.
14. Deflection coils have an inductance value of 3.6 millihenrys. Calculate the maximum allowable parasitic parallel capacitance if retrace is to be accomplished in 16 per cent of the horizontal period.
15. Explain the operation of the damping diode.
16. Explain the operation of the booster voltage circuit.
17. The damping diode can be connected to the primary or to the secondary of the line output transformer. Compare the circuits.
18. Explain the function of the capacitor across the damping diode load resistor.

19. Investigate in detail the operation of the shunt circuit for a damping diode.
20. Explain why a simple sawtooth grid drive of the horizontal output tube is unsatisfactory.
21. Deflection coils have an inductance of 4 millihenrys and a resistance of 110 ohms. Assuming a maximum positive value of deflection current of 0.4 amp, calculate the maximum negative value.
22. State the desirable requirements for a damping diode and justify them.
23. Design a horizontal centering circuit for a transformer type of output circuit.
24. Show diagrammatically how a saturated inductance may be employed for linearity adjustment.
25. State and justify the desirable requirements for a triode used as a damping tube.
26. It is proposed to use variation of the EHV voltage as width control. Investigate the idea.
27. A damping diode and power recovery circuit could be used with vertical deflection. Study the advisability of such a circuit.
28. Compare a standard horizontal sweep section with its direct drive counterpart.
29. Suggest three practical ways of controlling amplitude in a direct drive circuit.
30. Design a centering adjustment for a direct drive circuit.
31. Compare the efficiency of a direct drive and of a standard circuit.
32. Explain why horizontal output tubes are always tetrodes or pentodes and never triodes.

Chapter 18

POWER SUPPLY

Three types of power supply are used to satisfy the voltage requirements of a television receiver: heater voltage, plate voltage, and picture tube high voltage (EHV). The EHV supply relies on special techniques and will be taken up in Chap. 19. The first two supplies use classical circuits more or less and take their power only from the main line voltage available, usually 115 volts a-c.

Heaters of various tubes may be connected in series or in parallel, and in most cases may use either a-c or d-c power.

Plate voltage supplies are broadly divided into two types: a-c only or a-c—d-c. The former can be subdivided further, according to whether or not it uses a transformer; the latter is not very common.

Classical filtering circuits are used with plate supplies. In some cases voltage or current stabilization is necessary.

18-1. Power supplies

A television receiver needs a plate voltage power supply, a high-voltage or EHV supply for the final anode of the picture tube, and heater power.

An EHV supply poses peculiar problems and will be analyzed in the next chapter. Heater and plate power are usually always obtained from the available mains supply. A few receivers, however, have been designed to work off batteries or accumulators.

Generally speaking, power supplies may be divided into two broad categories. The first is the a-c—d-c type, so called because it works either off a-c or d-c mains. The second is the a-c type, which may be subdivided further according to whether or not it uses a transformer or an autotransformer.

The transformerless a-c type should not be mistaken for an a-c—d-c type

of supply. Although it uses no transformer, it does not work off d-c mains. A typical example would be a voltage doubler.

18-2. Parallel heater chain

Tubes can use either alternating or direct current for heating purposes in television receivers. In some cases, they use audio or radio frequency or even pulse voltages, as is the case with the EHV rectifier.

For this reason, heater supply arrangements are not classified as a-c or a-c—d-c, but rather according to whether the heaters are connected in parallel or in series.

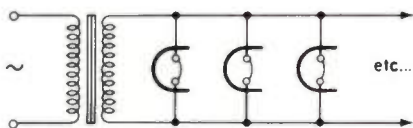


Figure 18-1. Parallel connection of heaters.

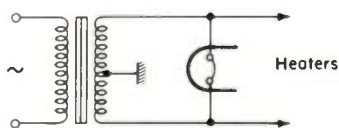


Figure 18-2. Center tapped winding.

For tubes of the so-called *alternative* series, all heaters are fed the same voltage and therefore are connected in *parallel*. Different heater powers are obtained by different heater currents.

The total heater power reaches from 10 to 50 watts, and the only economical way to obtain this power practically is to use a transformer or autotransformer, as shown in Fig. 18-1. Since this component works only on alternating current, this type of heater circuit is sometimes called “alternative,” although “parallel heater circuit” is a better description.

Modern tubes for parallel heater chains have been standardized at 6.3 v and take currents of 0.3 to 2 amp.

The heater-to-cathode insulation of some tubes is not made to withstand high potential differences. Moreover, there is a risk of induced hum. It is also customary not to allow the heater chain to float but to connect it to ground somehow. The circuit of Fig. 18-2 grounds a center tap on the heater winding of the transformer. Heaters are then fed by two connection wires, which may advantageously be twisted together to reduce risks of hum induction.

The circuit of Fig. 18-3 grounds one end of the heater winding. As before, two wires could be used for heaters wiring but the common practice is to use only one wire and rely on the general ground for the return circuit.

Voltage tolerances for heaters are such that they largely cover any variations in main supply voltage. If these variations become too important, they manifest themselves by apparent symptoms on the picture long before the heaters are really in danger or harmed.

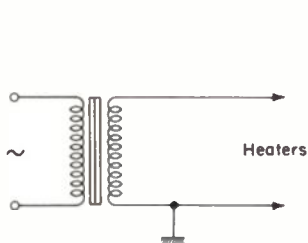


Figure 18-3. Grounding one end of the winding.

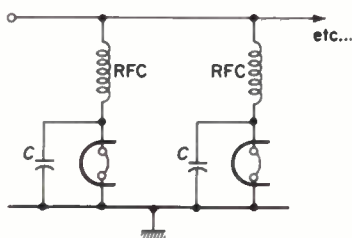


Figure 18-4. Parallel decoupling of heaters.

18-3. Decoupling

It may happen that instabilities or oscillations are created by interstage coupling due to r-f signals propagating along the heater connections. The remedy is the decoupling of the heater chain. A typical decoupling circuit consists of a capacitor and an r-f choke (RFC). The capacitor usually has a value of 500 to 10,000 $\mu\mu\text{f}$, and the r-f choke should be made of wire of sufficient diameter for the voltage drop to be negligible, as shown in Fig. 18-4.

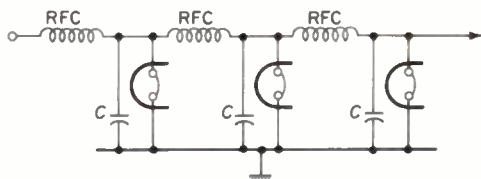


Figure 18-5. Series decoupling of heaters.

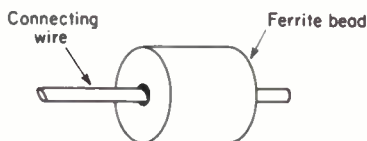


Figure 18-6. Use of a ferrite bead.

Instead of decoupling each heater separately, the arrangement shown in Fig. 18-5 may be used. The decoupling circuits are cascaded, which is preferable from a decoupling point of view. However, the current through each r-f choke increases with the number of heaters that follow, and the choke current ratings must be chosen accordingly. Sometimes the heater connecting lead itself is coiled in pigtail fashion to make an r-f choke. Another method is to use short tube lengths or beads of high-permeability material such as a ferrite around the heater wire, thus greatly increasing its inductance locally and producing the equivalent of an r-f choke as shown in Fig. 18-6.

Total heater current in a television receiver reaches a high value. If a single connecting wire is used, it should have sufficient carrying capacity or diameter. In addition, it would be a good idea to first feed the power tubes that have a large heater current to reduce voltage drop in the remaining heater connections.

A preferable, more simple method that is commonly used is to wire several separate heater chains, thus reducing current in each wire as shown in Fig. 18-7. For example, one wire would feed the sweep section, another the video receiver, and a third the sound receiver.

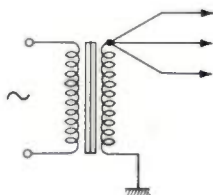


Figure 18-7. Multiwire heater circuit wiring.

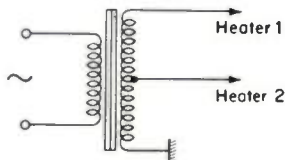


Figure 18-8. Connections for different heater voltages.

18-4. Different heater voltages

It may happen that it is necessary to use a special tube whose heater voltage is different from the others. The simplest solution is to use a separate winding

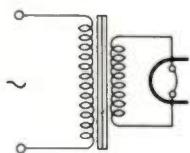


Figure 18-9. Separate heater transformer.

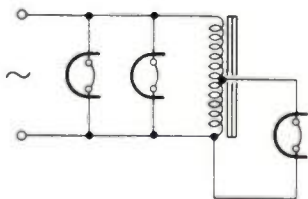


Figure 18-10. Heater autotransformer.

on the transformer or, equally satisfactory, a tapped winding as shown in Fig. 18-8. Another method is to add a separate small heater transformer as shown in Fig. 18-9. Its primary may be fed from the line voltage supply or from the general heater voltage. Such a circuit is also used when a heater has to be strongly insulated—for example, if the cathode is at high voltage.

In ordinary cases, the autotransformer shown in Fig. 18-10 may be used just as well, whether it is for step-up or step-down purposes as shown.

Also, if the required voltage V_1 is lower than the available heater voltage V , a simple dropping resistor may be used as shown in Fig. 18-11. Where I is the tube current, the resistance value should be

$$R = \frac{V - V_1}{I}$$

and it dissipates a power

$$W = I \times (V - V_1)$$

Since parallel heated tubes are voltage-standardized, heater circuits should be checked by measuring the voltage across filaments. Variations of 10 per cent are tolerated. If one heater is burnt out or otherwise open, the other heaters in the chain are not affected.

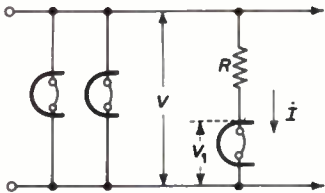


Figure 18-11. Dropping resistance.

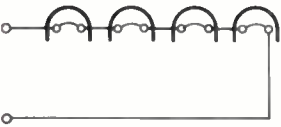


Figure 18-12. Series string of heaters.

18-5. Series heater chain

With tubes of the a-c—d-c series, heaters all take the same current and therefore must be connected in series, as shown in Fig. 18-12. Since various tubes require different heater powers, heater voltage varies between tubes in a series string.

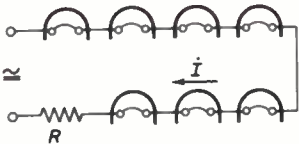


Figure 18-13. Dropping series resistance.

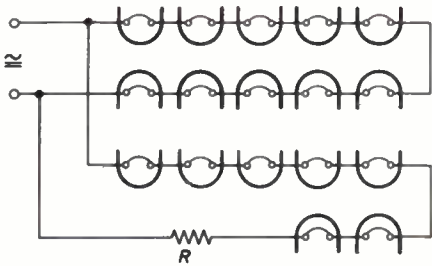


Figure 18-14. Two series strings of heaters in parallel.

Total heater voltage for a series chain is equal to the sum of the individual heater voltages. It can be provided either by a heater winding on the power transformer or directly from the main supply. In the last case, it is immaterial whether the main voltage supply is a-c or d-c, since no transformer is required (hence the name “a-c—d-c” sometimes given to this type of heater circuit).

If the total heater voltage V_1 is smaller than the main supply voltage V , a dropping resistor R may be used as shown in Fig. 18-13. I being the heater current, the resistance has a value

$$R = \frac{V - V_1}{I}$$

and dissipates a power

$$W = I \times (V - V_1)$$

If the total heater voltage exceeds the line voltage, heaters may be distributed between two chains, as shown in Fig. 18-14. Each chain may have its dropping resistor or, for economy, one chain is arranged to add up exactly to line voltage and therefore does not require a dropping resistance.

In actual practice, one end of the heater chain is grounded, as shown in Fig. 18-15, and this poses the problem of heater insulation.

The voltage between any heater and ground increases when going up the chain, starting from ground. The risks of defective insulation and 60-cycle induction of hum between heater and cathode increase simultaneously. For this reason the dropping resistor is at the top of the chain, since it poses no insulation problem.

As for the tubes, insulation between heater and cathode has been designed to take the full main supply voltage easily in a-c—d-c series applications. Nevertheless, in some types of power supply such as voltage doublers, the main voltage supply is not connected to ground and is at high d-c potential with respect to ground. In this case, heater-to-cathode insulation has to take this d-c voltage plus the peak value of the a-c main supply voltage.

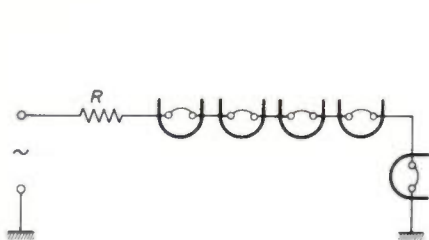


Figure 18-15. One pole of the mains supply is grounded.

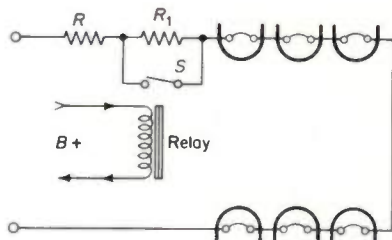


Figure 18-16. Heaters protection relay.

18-6. Switching-on transient

Tube heaters have a cold resistance that is much lower than their working resistance when hot. Immediately upon switching on, a high heater current flows for a short time. It may reach several times the normal current and decreases while the heaters heat up to a working temperature and reach their normal resistance.

This current surge is dangerous to heater life and should be avoided or at least reduced to acceptable proportions. A circuit which accomplishes this is shown in Fig. 18-16. It includes an additional protective resistor R_1 in addition to the normal dropping resistor R . Immediately upon switching on, both R

and R_1 are in the circuit and limit the current surge. After some time, the heaters are hot and R_1 is short-circuited. Short-circuit switch S is automatically controlled, either by a time-delay circuit or by plate supply current as shown, flowing through an electromagnetic relay. In one version, the electromagnetic relay is combined with the filter choke.

Another solution is to use a material that has resistance characteristics opposite to that of the heaters—that is, it should have high resistance when cold and low resistance when hot. Such materials, sometimes called *thermistors* are said to have a *negative temperature coefficient* or *NTC*. The corresponding circuit is shown in Fig. 18-17. An NTC element designed for the nominal current is inserted in series with the heater chain and virtually eliminates the current transient at the instant of switching on.

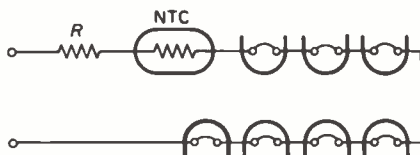


Figure 18-17. Use of a NTC resistance.

18-7. Series wiring

With series connection of the heaters, the only way to connect r-f chokes and capacitances for decoupling purposes is given in Fig. 18-18. The current is the same through the entire chain, generally on the order of 0.1 to 0.3 amp, and it is not very difficult to make r-f chokes that have negligible voltage drop.

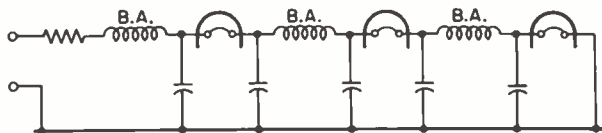


Figure 18-18. Decoupling circuits in a series heaters string.

For the same reason, the connection wire does not pose any problem in series wiring.

In the same way as for a parallel chain, ferrite beads may be used instead of r-f chokes, or the connecting wire may be wound to play the same role in decoupling. The tendency, of course, is to use a minimum of heater decoupling circuits, both for economy of components and for ease of wiring.

18-8. Different heater currents

It may be necessary to use tubes that have different heater currents in the receiver.

A possible solution is to group all tubes having the same current requirements and to have each group constitute one chain, as shown in Fig. 18-14.

Another solution is to design the chain for the tube or tubes that have the largest current, and shunt those remaining by a resistor R designed to accept the excess current, as shown in Fig. 18-19.

Let I = the chain current,

I_1 = the current required by the low-current group,

V = the voltage required by the low-current group,

R = the equalizing resistance.

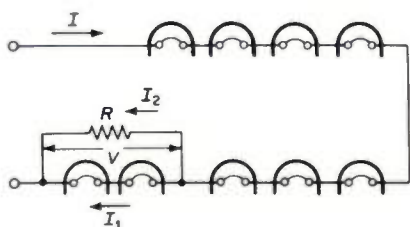


Figure 18-19. Current shunting resistance R .

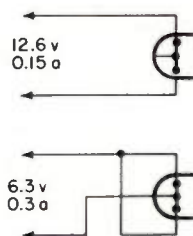


Figure 18-20. Double voltage standard heater.

The current I_2 flowing through R is

$$I_2 = I - I_1$$

Since the voltage across R must be V ,

$$R = \frac{V}{I_2} = \frac{V}{I - I_1}$$

and the power dissipated by R is

$$W = VI_2 = V(I - I_1)$$

Series heater tubes are calibrated for current and heater circuit check should be made with an ammeter inserted in series with the chain. Variations of up to 10 per cent are considered as normal.

Most tubes use either 150-ma or 300-ma heater current. Some are calibrated for both current and voltage and can be connected either in series or in parallel. Some have a center-tapped heater. According to whether both halves of the heater are used in series or in parallel, the tube will require 12.6 v and 0.15 amp or 6.3 v and 0.3 amp, as shown in Fig. 18-20.

In series heater circuits, an open heater in any portion of the string prevents the whole chain from heating up.

18-9. Order of connection

Frequently the total voltage required by the series-connected heaters of a television receiver exceeds the main supply or house voltage, and two chains are used.

The order of the heaters in the chain is somewhat important. Near the ground end of the chain are placed the sensitive heaters such as those of hum-sensitive tubes (a-f voltage amplifier, detector, sync separator) or those of tubes that have a high cathode voltage (such as the picture tube).

For noncritical tubes, the order has less importance and is frequently dictated by practical wiring considerations.

Finally, the rugged rectifiers and output tubes, whose heaters take several watts and whose large cathodes are well insulated, are placed at the end of the chain near the ungrounded main voltage input side, just after the dropping resistances and NTC resistances if any.

To illustrate these principles, Fig. 18-21 shows the heater circuit of a transformerless receiver. Most of the heaters have been arranged in two chains connected in parallel. The remaining heaters are those of the picture tube, the vertical power output stage, and the tuner section. These have been connected in such a way that the total current they need is equal to the sum of the currents in the two chains.

Notice the decoupling circuits and the order of connection of the heaters.

18-10. Plate power supply

The plate power supply or B+ voltage may essentially be obtained in three ways. The first method uses a transformer or autotransformer and works only on alternating current. The second is transformerless but works only on alternating current. The third is transformerless and works on alternating or direct current. Needless to say, the a-c types require rectification and filtering.

Transformers are heavy, costly, and cumbersome. In addition, they radiate a magnetic field which may cause undesirable inductions. The same remarks apply to autotransformers, although to a lesser degree. On the other hand, with autotransformers the chassis is live—that is, it is connected to one pole of the main supply voltage.

The a-c transformerless power supplies include all voltage-doubling or multiplying circuits. They also have the disadvantage of a live chassis.

The a-c—d-c type of power supply is rarely used because of the low B+ voltage it provides; it also has a live chassis.

The above classifications clarify matters, but they are not absolute. For example, voltage doublers are commonly used with autotransformers.

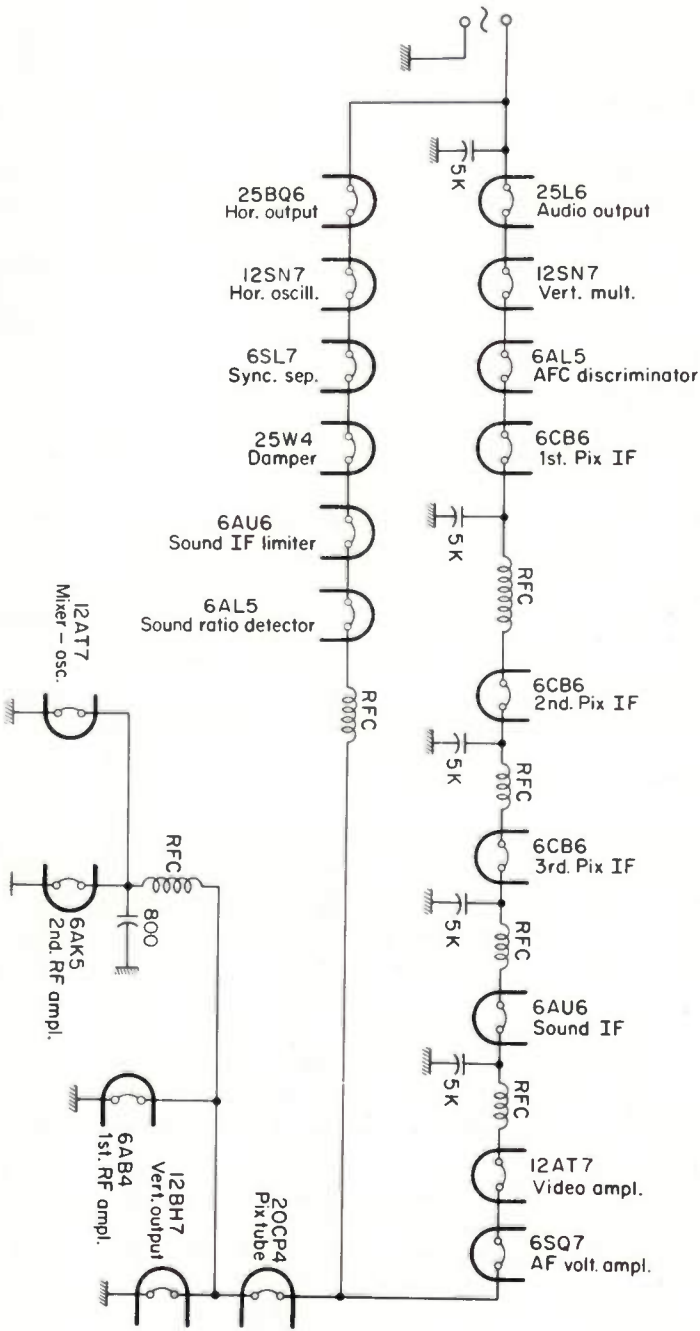


Figure 18-21. Practical series circuit for heaters.

18-11. Transformer

A classical transformer type of power supply is shown in Fig. 18-22. The primary circuit contains a fuse and has several taps so that the transformer can be matched to the available voltage.

Three secondaries are provided. The bottom one, shown in Fig. 18-22, provides heater voltage for all the tubes except the rectifier. The middle one is the high-voltage winding, which is center-tapped to provide full-wave rectification with a double-diode rectifier. The top secondary is used to feed the rectifier heater, which must be insulated since it is the B+ take-off point.

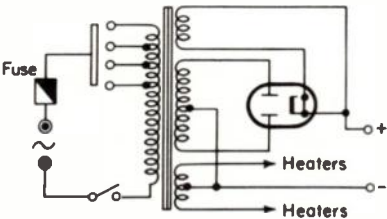


Figure 18-22. Plate supply using a transformer.

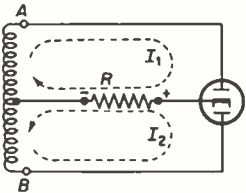


Figure 18-23. Operation of the rectifier.

The equivalent circuit for rectifier operation is shown in Fig. 18-23, where R is the rectifier load. During one alternation of the a-c voltage across the winding, point B is negative, and no current flows through the diode connected to B . At the same time A is positive. The diode connected to A conducts and delivers a current I_1 to the load.

During the following alternation, conditions are reversed. A is negative, and the top diode is cut off. B is positive and the bottom diode conducts and provides a current I_2 to the load. Whatever the alternation, current I_1 and I_2 both flow in the same direction, and rectification is said to be full-wave.

Figure 18-24(a) shows the a-c current I flowing through the transformer winding. In (b) is shown the way in which rectified currents I_1 and I_2 are combined. Although always positive, the total rectified current $I_1 + I_2$ contains a strong a-c component, shown by the dotted lines. This a-c component has a frequency that is double the supply frequency and must be eliminated. Filter circuits are used to accomplish this.

18-12. Filter circuit

The rectified current of Fig. 18-24(b), when analyzed for its components, is found to contain a desirable d-c component and undesirable a-c or hum components. Hum components consist of a fundamental component at twice the

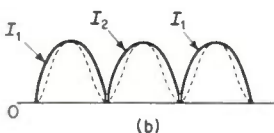
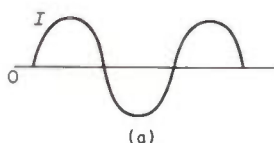


Figure 18-24. Waveforms in the rectifier circuit.

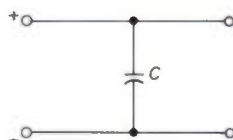


Figure 18-25. Capacitance filtering.

mains supply frequency and of all its even harmonics, whose amplitudes decrease with increasing frequency.

A method used to reduce hum components, as shown in Fig. 18-25, is to shunt a large capacitor C across the output of the rectifier. It has a reactance

$$X = \frac{1}{2\pi fC}$$

which decreases when frequency f increases, so that the capacitor more or less shorts out a-c components if its value is large enough.

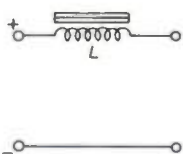


Figure 18-26. Inductance filtering.

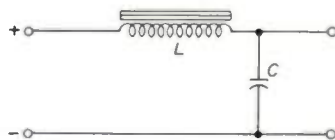


Figure 18-27. Inductance input filter.

Additional filtering, as shown in Fig. 18-26, is obtained by inserting an inductor L in series with rectified B+ supply. It has a reactance

$$X = 2\pi fL$$

which increases when frequency f increases. If L is large enough, it presents a high reactance to the flow of a-c components.

Both L and C are usually combined, either in the form shown in Fig. 18-27, with choke input, or that of Fig. 18-28, with capacitor input. Adding an output capacitor C_2 as shown gives the frequently used *pi filter*.

The inductor should have a large value for good filtering. Also, since it is in series with the B+, it should have low ohmic resistance to reduce voltage

drop, particularly in the case of high current supplies. These two requirements are somewhat contradictory if the component is to be made economically. Usually a few henrys are considered as sufficient, and the ohmic resistance may be from ten to a few hundred ohms.

Filter capacitors have high values, usually between say $8\ \mu\text{f}$ and $200\ \mu\text{f}$. This is economically feasible only with electrolytic types of capacitors, which

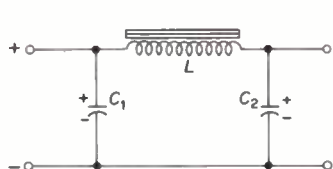


Figure 18-28. Pi filter.

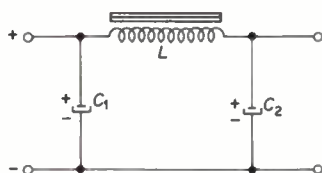


Figure 18-29. Use of electrolytic capacitors.

are polarized and must be connected to the respective polarities as shown in Fig. 18-29.

When the current is not too high or when a large voltage drop can be tolerated, a sizable economy is possible by replacing the inductor with a simple resistor as shown in Fig. 18-30.

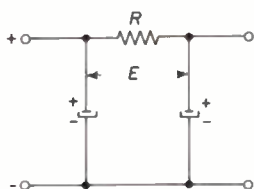


Figure 18-30. Resistance-capacitance filter.

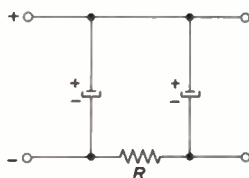


Figure 18-31. Filter in the negative conductor.

The larger R is, the better the filtering, the larger the voltage drop E across R , and also the larger the power P dissipated by R :

$$P = \frac{E^2}{R}$$

There is no absolute requirement for the filter inductor or resistor to be in series with the ungrounded B+ connection. It could equally well be in the B—, as shown in Fig. 18-31. Several filter circuits may be cascaded to improve filtering, as shown in Fig. 18-32.

When different circuits must be fed from a common supply, individual filtering circuits may be required to avoid interaction. This can be accomplished according to the parallel arrangement shown in Fig. 18-33 or according

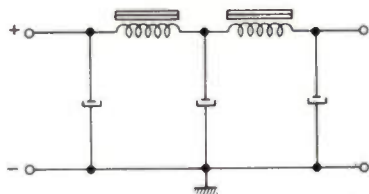


Figure 18-32. Double pi filter.

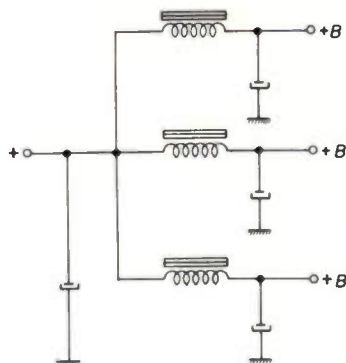


Figure 18-33. Parallel filter circuits.

to the series arrangement of Fig. 18-34. The last circuit has the advantage that filter circuits are cascaded, and the disadvantage that current through each inductance increases as shown when going back to the power supply.

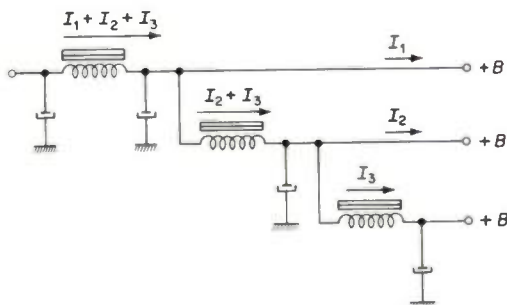


Figure 18-34. Cascaded filter circuits.

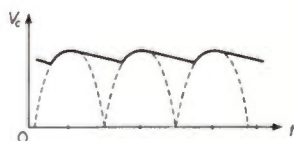


Figure 18-35. Voltage across the filter input capacitance.

18-13. Rectified voltage

With the capacitor input type of filter shown in Fig. 18-28, the input capacitor charges up to the peak value of the applied rectified voltage. Then the capacitor provides B+ current and partially discharges, to be charged up again by the next alternation, and so on.

The voltage V_c across the capacitor then has the shape shown in Fig. 18-35, and is seen to be much nearer to a real d-c voltage than the raw rectified waveform shown by the dotted lines.

Let V be the rms value of the a-c voltage applied to the rectifier. The corresponding peak value is

$$V_p = 1.41V$$

For raw rectified voltage such as is shown in Fig. 18-24(b), the d-c component is approximately equal to 0.64 times the peak value, that is, $0.9V$.

When there is an input capacitance, it charges up to $1.41V$, which is the peak voltage.

With the choke input type of filter shown in Fig. 18-27, the inductance provides the first filtering action. The capacitor after the inductor charges up to the average d-c voltage, that is, $0.9V$.

The addition of an input capacitor to the filter, therefore, brings a marked increase in the available B+ voltage, in the ratio of $1.41/0.9 = 1.57$, approximately.

In actual practice, assuming a 110-v rms a-c supply, a choke input filter would provide a B+ voltage of 100 v, whereas a capacitor input type of filter would deliver 150 v.

The mode of operation of the rectifier, however, is harder on the rectifier tube for a capacitor input filter. With a choke input filter, the output capacitor charges almost continuously from the rectifier through the inductor. With a capacitor input filter, the input capacitor receives a surge of charging current only during a short time at the maximum of each alternation. For equal power delivered by the rectifier, this means that the shorter charge current has to be much more intense with capacitor input than with choke input.

18-14. Rectifier

A rectifier tube is almost always used with transformer-type power supplies. It is a double-anode tube for full-wave rectification, and it must be able to satisfy the B+ power requirements of the receiver. Representative values would be 350 v and 200 ma.

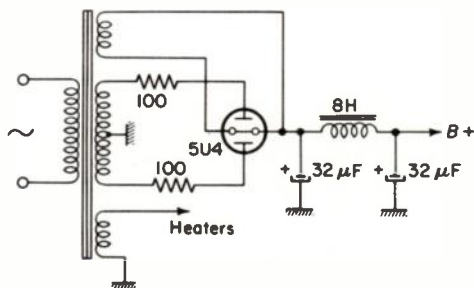


Figure 18-36. Practical plate power supply.

In addition, the rectifier must stand the surge of current during the short period when the input capacitor of the filter charges up.

The surge current may be limited by inserting a resistor in series with each anode, as shown in Fig. 18-36. If the secondary winding has enough ohmic

resistance, the resistors may be omitted. The result in any case is a decrease in $B+$ voltage.

The voltage drop across the rectifier and the power it dissipates increase with $B+$ current. If one rectifier is insufficient, two rectifiers simply can be connected in parallel, which doubles the power handling ability.

Sometimes a negative voltage is needed—for bias purposes, for example. It can be obtained easily from the standard power supply through the circuit shown in Fig. 18-37. A resistor R has been inserted between center tap of the

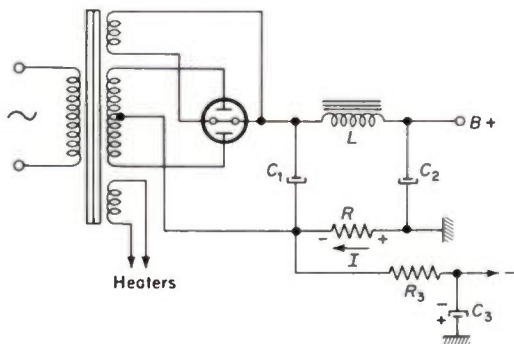


Figure 18-37. Negative bias can be obtained from the plate power supply.

secondary and ground. $B+$ current I flows through R and creates a voltage drop as shown. The resulting voltage RI is negative with respect to ground and can be adjusted by varying R . Notice that the filter input capacitor C_1 is connected to the ungrounded end of R , which insures better filtering.

The resulting negative voltage can be filtered, generally by an R_3C_3 filter as shown.

18-15. Practical circuits

A few representative circuits of transformer-type power supplies are given to illustrate typical applications of the principles explained previously.

Figure 18-38 is a straightforward diagram. The high-voltage secondary winding provides 2×285 volts rms. After rectification, 270 volts d-c are obtained. A pi filter is used that provides a filtered $B+$ of 260 v. A separate 6.3-v winding feeds the heaters of the receiver tubes.

Figure 18-39 has a similar rectifying section, except that the use of a 2×325 v winding insures a rectified d-c voltage of 335 v. It is used as such, with only the filtering due to the input capacitor of the pi filter. Also, a well-filtered 310-v output is available after the pi filter, and another output at 150 v is obtained by using a high-value resistor in a resistance-capacitance filter.

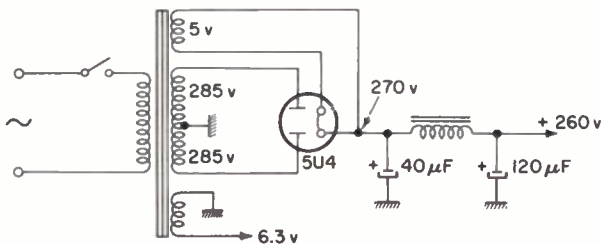


Figure 18-38. Typical plate power supply.

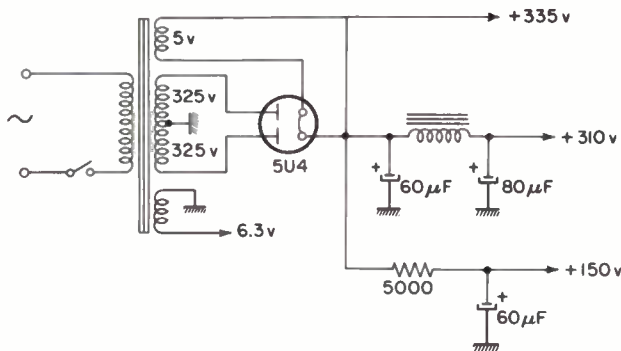


Figure 18-39. Inductance and resistance filters.

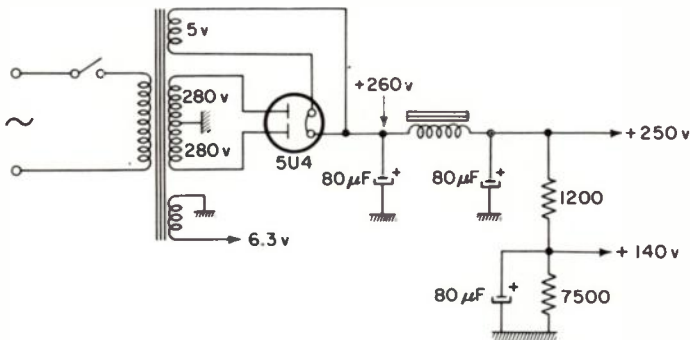


Figure 18-40. Resistance voltage divider.

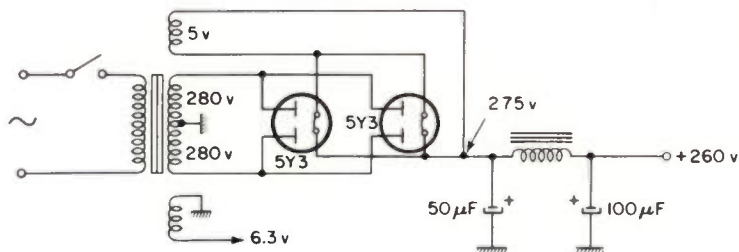


Figure 18-41. Use of two rectifier tubes in parallel.

Figure 18-40 is almost identical with Fig. 18-38, except for the provision at the output of a resistance divider to obtain 140 v, with an extra filtering circuit.

Figure 18-41 is also a classical circuit. Note that it uses two rectifiers in parallel to satisfy B+ current requirements.

18-16. Autotransformer

The basic autotransformer circuit of Fig. 18-42(a) shows how a step-up autotransformer provides a-c voltage for the rectifier. Two secondary windings are used in addition, one for the rectifier heater and one for the other heaters in the receiver.

The autotransformer, for the same power, is usually smaller and less costly than a transformer and may lead to some economy. It is rarely used, however, since a more economical way of obtaining comparable results is to use voltage doublers.

The autotransformer circuit shown rectifies only one alternation every period, when the rectifier plate goes positive. It is a half-wave rectifier. Figure 18-42(b) shows the voltage applied to the rectifier and Fig. 18-42(c) shows the resulting raw rectified voltage. The fundamental hum component is shown by the dotted line. It has the same frequency as the supply main and will be more difficult to filter out than the fundamental hum component of full-wave rectification, which was twice the main frequency. In this respect, filtering may be made easier by the fact that B+ voltage in this circuit is not high, so that high-value electrolytic capacitors are a practical proposition.

18-17. Transformerless circuit

Transformerless circuits working directly on a-c mains only are the voltage-doubler or voltage-multiplier circuits, using tubes or semiconductor rectifiers.

The basic principle of voltage multiplication is simple. Large-value capacitances are charged from rectified alternating current. Then they are commuted in series, so that total voltage is equal to the sum of the individual voltages.

This total voltage is the desired $B+$. Of course, capacitance commutation is obtained automatically at the frequency of the mains.

The simplest voltage multiplier is evidently the voltage doubler. Two circuits are in common use, the *Latour doubler* and the *Schenkel doubler*.

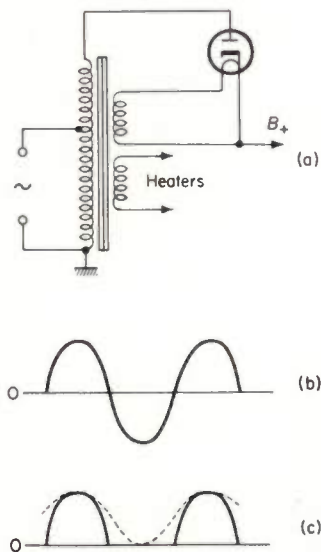


Figure 18-42. Single phase rectifier and waveforms.

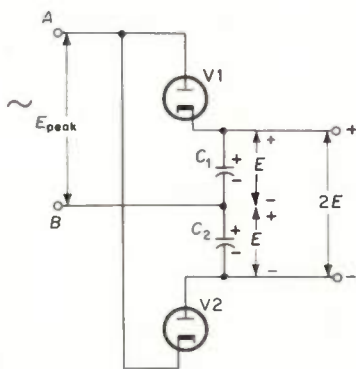


Figure 18-43. Latour voltage-doubling circuit.

18-18. Latour doubler

The basic Latour voltage-doubling circuit is shown in Fig. 18-43. It uses two rectifiers and two capacitors.

During the first alternation, point A is positive. Diode V_2 , whose cathode is positive, is cut off, and the diagram reduces to Fig. 18-44. Diode V_1 , whose anode is positive, delivers a current I which charges capacitance C_1 with the polarities indicated, up to the peak a-c voltage E_1 .

During the following alternation, point A becomes negative and diode V_1 , whose anode is negative, cuts off. The diagram reduces to that shown in Fig. 18-45. Diode V_2 , whose cathode is negative, delivers a current I which charges capacitance C_2 as indicated up to the peak a-c voltage E_2 .

For the $B+$ circuit, C_1 and C_2 are series-connected as shown in Fig. 18-46, so that the total $B+$ voltage is $E_1 + E_2$, that is, twice the peak a-c voltage. For example, with a main supply of 110 volts rms

$$E_1 = E_2 = 110 \times 1.41 = 154 \text{ v}$$

and the output voltage is 308 v.

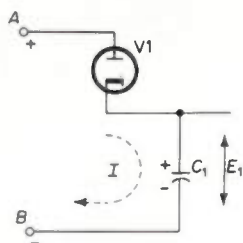


Figure 18-44. First alternation current flow.

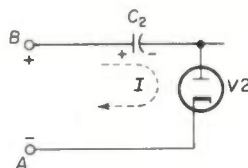


Figure 18-45. Second alternation current flow.

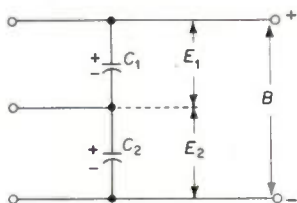


Figure 18-46. Voltage doubling across the capacitances.

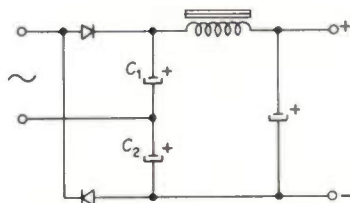


Figure 18-47. Complete Latour voltage-doubler circuit.

Actually, the capacitors do not quite charge up to the peak a-c voltage, but to a lower value. The higher the current required, the lower the voltage obtained. This can be corrected by using large-value capacitors. Since they carry charges of constant polarity, they can be of the electrolytic variety. Moreover, each capacitor is submitted to only half the output voltage, say 150 v working for 110-v mains. All in all, then, there is no difficulty in using large values, say 200 μ f.

The voltage-doubling capacitors simultaneously can play the role of filter input capacitance as shown in Fig. 18-47. Since both alternations are rectified, the fundamental hum component is at twice the frequency of the main.

Notice that in the Latour circuit the main supply has no direct common point with the output d-c voltage. In particular, the lower end of the main source is separated from B— (that is, ground) by C_2 , which is charged up to 150 v. On the positive peak of a-c input voltage, at 150 v peak, the instantaneous voltage between mains and B— or ground reaches 300 v. This may be a drawback with series heater circuits as already pointed out, since the heater-to-cathode insulation must withstand 300 v.

18-19. Schenkel doubler

The basic diagram of the Schenkel doubler circuit is illustrated in Fig. 18-48. It also uses two rectifiers and two capacitors, but the mode of operation is different.

During one alternation, point A is negative. Diode V_2 is cut off, and the circuit reduces to Fig. 18-49. Diode V_1 conducts and passes a current I which charges C_1 up to the peak value E_1 of the a-c supply, with the polarities shown.

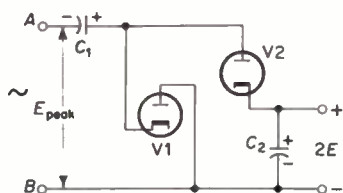


Figure 18-48. Schenkel voltage doubler.

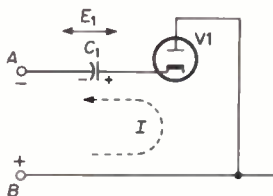


Figure 18-49. First alternation current flow.

During the following alternation, point B is negative. Diode V_1 is cut off, and the circuit reduces to Fig. 18-50. The main supply voltage, positive at A , is in series with the voltage E_1 accumulated on C_1 , and the total peak value $2E_1$ is applied to diode V_2 which conducts and charges C_2 up to peak value $2E_1$.

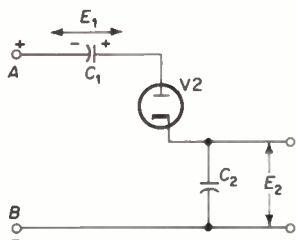


Figure 18-50. Second alternation current flow.

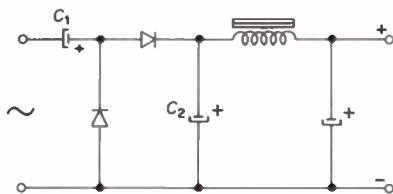


Figure 18-51. Complete Schenkel voltage doubler circuit.

The output d-c voltage taken across C_2 is then twice the peak a-c input voltage. As before, the 110-volt a-c main would theoretically produce 300 v d-c, and practically less, depending on current requirements and capacitance values.

Once again, capacitors always charge up with the same polarities and can be of the electrolytic type. However, C_2 must now withstand the full doubled d-c voltage, say 300 v. C_1 must pass a large a-c component but needs to be insulated only for 150 v working. Its value varies with B+ current and is on the order of 100 μf for 200 ma, 150 μf for 300 ma, and 300 μf for 400 ma.

C_2 doubles also as filter input capacitor as is apparent from Fig. 18-51.

In the Schenkel doubler circuit, the a-c supply main has a pole directly connected to B— or ground, an advantage for use with series heater circuits.

Figure 18-54 represents a Schenkel voltage doubler. A fuse is again provided for diode protection as well as a resistor to limit the current surge. Heaters are connected in a series chain across the main voltage supply.

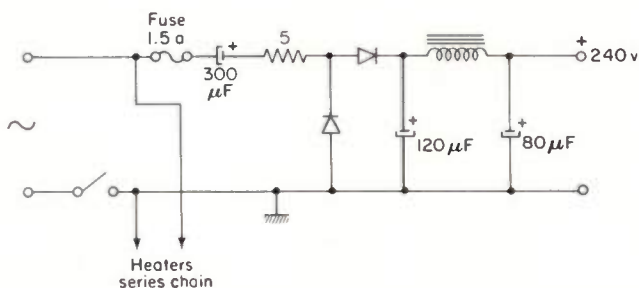


Figure 18-54. Practical Schenkel voltage - doubler circuit.

18-21. Voltage multipliers

The principle used in voltage doubling can be extended to higher orders of voltage multiplication.

For example, adding the voltages provided by a simple rectifier and a voltage doubler gives the voltage-tripling circuit shown in Fig. 18-55. The rectified d-c voltage easily reaches 350 v for the standard 110-v main supply.

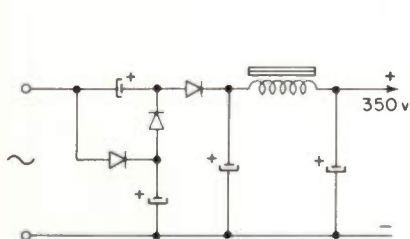


Figure 18-55. Voltage-tripler circuit.

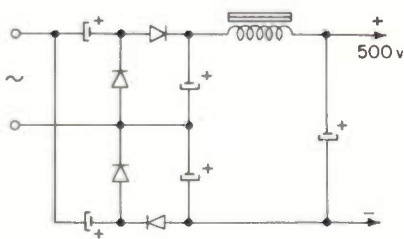


Figure 18-56. Voltage-quadrupler circuit.

Similarly, two voltage doublers can be connected in series to provide the voltage quadrupler shown in Fig. 18-56. Notice in this circuit that once again the main is not connected to the B— output and that insulation problems may arise with series heater chains.

Simple rectifiers and voltage multipliers may be combined in still another way—to provide both positive and negative voltages with respect to ground. Figure 18-57 illustrates the idea with a Schenkel doubler providing +250 v and a single rectifier providing -120 v. With two doublers as in Fig. 18-58, +250 v and -250 v are available, and in this arrangement the mains have

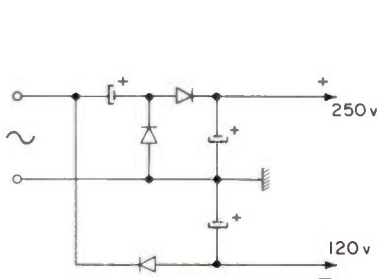


Figure 18-57. Positive and negative voltage supplies.

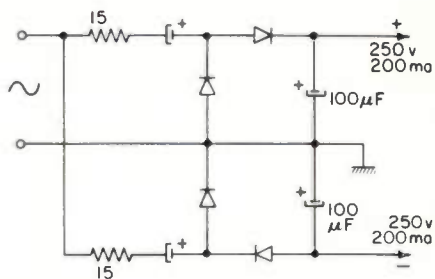


Figure 18-58. Double voltage-doubler circuit.

one pole grounded. Nevertheless, the full quadrupled voltage can be applied to tubes which require it by connecting the cathode and grid returns to -250 v instead of ground, as shown in Fig. 18-59. Notice in this circuit the grounded screen grid, which in this way is now at $+250$ v with respect to the cathode.

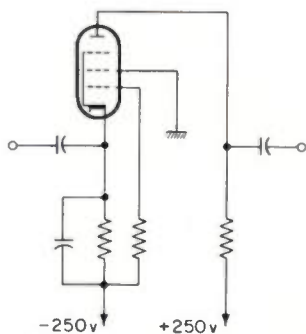


Figure 18-59. Negative voltage cathode supply.

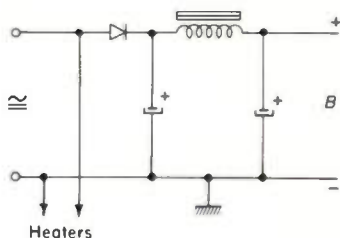


Figure 18-60. Direct mains voltage rectification.

18-22. A-c—d-c power supply

An a-c—d-c power supply must work as well on direct current as on alternating current and therefore cannot use transformers, autotransformers, or voltage multipliers. This means, of course, that on d-c the available $B+$ voltage is at most the d-c mains supply voltage, say 115 v. This is none too high for standard circuits. On a-c, however, the available $B+$ could reach the peak a-c value, say 150 v, which is somewhat of an improvement.

The basic diagram of Fig. 18-60 simply rectifies the main voltage. The rectifier is actually not necessary for direct current, but in this latter case the right polarity must be respected: the positive d-c pole of the mains should be connected to the rectifier.

Safety precautions must be taken in all circuits where the supply mains have one pole connected to the chassis. The supply main has a pole which is at ground potential and harmless as far as the risk of electrocution is concerned. This pole should be connected to the chassis. A convenient way of insuring this highly desirable result is to use a three-prong plug, which is polarized and fits the supply main receptacle only one way. Failing this, or as an additional precaution, receivers should be designed so that the user cannot come into contact with any conducting material which may be connected to

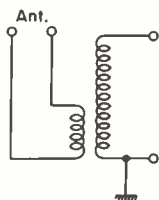


Figure 18-61. Input transformer to isolate the aerial.

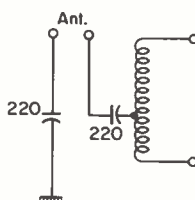


Figure 18-62. Aerial isolating capacitances.

the chassis, such as metallic knobs, grille, antenna, etc. The antenna can be insulated from the chassis through the use of a separate input primary winding, as shown in Fig. 18-61, or, if an autotransformer input is used, with two small-value capacitors, as shown in Fig. 18-62.

18-23. Voltage stabilization

In some cases it may be necessary to stabilize a voltage or a current; for example, voltage stabilization may be useful for the plate supply of an oscillator or for the accelerating voltage of a picture tube. Current stabilization may be provided in association with focusing current. Stabilization of the EHV is a special case and will be examined in the next chapter.

Stabilizing the B+ supply is rarely done on domestic receivers but is occasionally used, at least for critical stages, in professional installations. Two voltage stabilizers are in common use: the gas regulator diode and the vacuum-tube series circuit.

The common way to stabilize a current is to rely on the properties of the pentode tube or of nonlinear devices.

18-24. Gas-tube voltage regulator

A gas diode consists of two cold electrodes in a glass envelope filled with gas at low pressure. Its useful property is that the voltage across it is independent of a wide range of the current through it, called the *regulation* or *stabilization*

range. The circuit is shown in Fig. 18-63. The device to be fed a stabilized voltage is connected at X . Resistor R is adjusted so that for normal working conditions the current through the gas diode is in the middle of the stabilization range. In this way, the voltage at X is constant even if current consumption varies above or below the rated value.

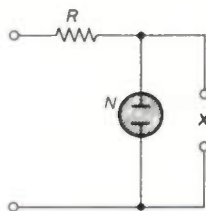


Figure 18-63. Gas diode voltage stabilization.

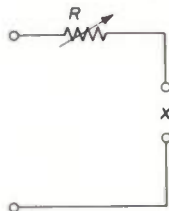


Figure 18-64. Adjustable dropping resistance R .

18-25. Tube regulator

The basic principle of a tube voltage regulator is illustrated in Fig. 18-64. At X is connected the device which should be fed constant voltage despite current variations. This can be done by modifying the value of the resistance R . In a practical arrangement, R will be the internal resistance of a tube, controlled by

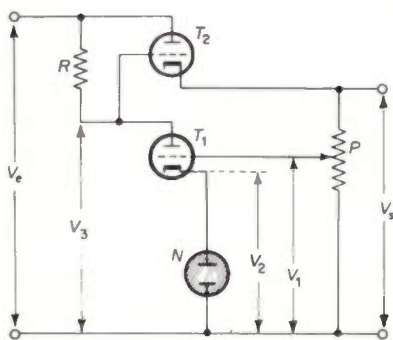


Figure 18-65. Tube voltage stabilizer.

the grid voltage. A simplified circuit diagram appears in Fig. 18-65. A fraction V_1 of the output voltage V_s is taken on potentiometer P and applied to the grid of a d-c amplifier T_1 whose cathode is fixed at constant voltage V_2 by the gas diode N .

The bias on T_1 is $V_2 - V_1$ and determines a plate voltage V_3 , which is used as control voltage for the grid of tube T_2 . For this value, T_2 has an internal resistance such that the voltage drop across T_2 is $V_c - V_s$.

Assume, for example, that the current consumption at the output V_s increases. The voltage V_s tends to decrease. V_1 decreases proportionately and V_3 increases. The control grid of T_2 becomes more positive, the internal resistance of T_2 decreases, and so does the voltage drop across T_2 . The output voltage V_s tends to increase, thus compensating for the decrease due to the variation in current.

The output voltage can easily be stabilized to a fraction of 1 per cent.

There are numerous variations of this circuit. For example, T_2 may be

replaced by several power tubes in parallel to pass the required current. Also, larger d-c amplification may be obtained by replacing T_1 by two stages.

Figure 18-66 shows a practical form of the circuit which provides 150 v regulated to 1 per cent and can deliver 200 ma. Resistance R is adjusted so that the current through the gas diode is 2.5 ma. Notice that a negative voltage supply is necessary in this circuit to provide the voltage across the gas diode.

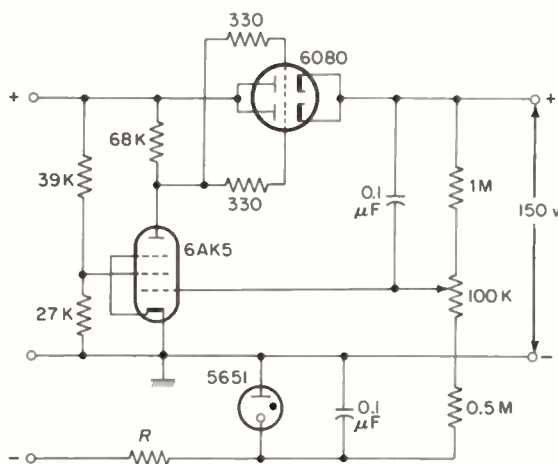


Figure 18-66. Practical diagram of a voltage stabilizer.

The screen voltage for the d-c amplifier tube is obtained through a resistance bridge. A low-value resistance is inserted in each grid connection to eliminate a possible high-frequency oscillation of the high-transconductance power tube.

The $0.1 \mu\text{f}$ capacitor at the output feeds back a-c components to the d-c amplifier and greatly reduces hum.

18-26. Current stabilization

In a pentode, plate current is practically independent of plate voltage above a minimum value of 50 v or so. This is apparent on a plate characteristic curve, as is shown in Fig. 18-67, for constant control grid and screen grid voltages. Adjusting the grid bias V_g modifies the plate current as shown.

This peculiarity may be used to provide a constant current to a load X , as shown in Fig. 18-68, whatever the variation in resistance of X within wide limits. Screen voltage is maintained at a constant value by a gas diode N . Potentiometer P adjusts the bias, hence the current through X .

This circuit is useful to stabilize current to a focus coil. In the usual arrangement, when the focus coil temperature rises due to power dissipation,

its resistance increases and the focusing current decreases, so that the focus control has to be readjusted.

If the focus coil is connected at X of Fig. 18-68, focusing current will remain constant despite resistance changes in the coil.

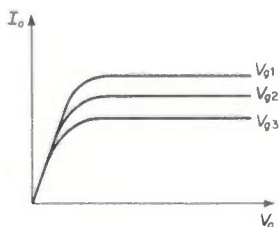


Figure 18-67. Family of anode characteristics.

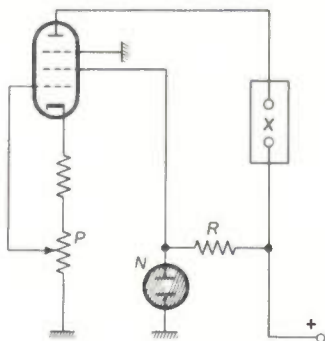


Figure 18-68. Current stabilizer.

18-27. Mains stabilization

The a-c voltage at the supply mains may vary and upset the proper working of the circuits. To avoid this, a-c voltage regulators may be inserted between the supply mains and the receiver.

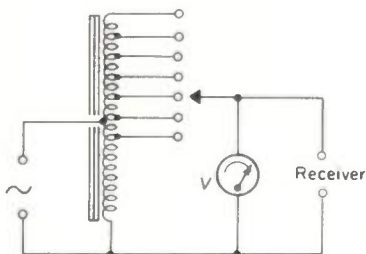


Figure 18-69. Manual adjustment of mains voltage.

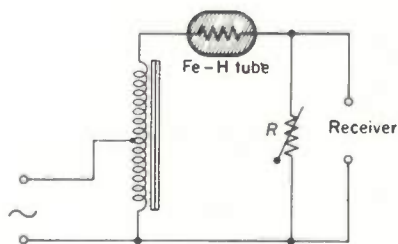


Figure 18-70. Voltage regulator using an iron-hydrogen tube.

Manual regulators may be used to correct for known regular variations such as daytime and nighttime differences. A simple autotransformer of sufficient rating with taps connected to a selector switch is convenient. A voltmeter monitors the output, as shown in Fig. 18-69, to obtain the desired receiver input.

Evidently, an automatic regulator is much more convenient. It can be based, as shown in Fig. 18-70, on the properties of an iron-hydrogen tube.

This tube consists of a filament of iron in a hydrogen atmosphere and its internal resistance varies strongly with current. The resistance variations oppose the current variations, so that the device is, in fact, a current stabilizer. In the diagram, the autotransformer is used to compensate for the voltage drop across the tube. Resistance R is adjusted to set the tube in the middle of its regulating range for normal working conditions.

Another circuit uses a saturating inductance, according to the diagram shown in Fig. 18-71. The output current flows through part of the inductance L , whose value is a function of current, so that its impedance varies as current changes.

With proper dimensioning of the components, a-c voltage stabilization is accomplished. The details of the operation of this so-called *ferroresonant* circuit are much more complicated, and the design of the regulator is somewhat empirical. Nevertheless, it does work well and offers the advantages that it is fully automatic, uses no tubes, and is extremely rugged. A number of variations of the basic circuit are commonly used for input voltage stabilization.

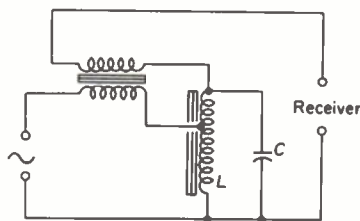


Figure 18-71. Saturating inductance voltage stabilizer.

Problems

1. Using the theoretical diagram of a typical television receiver, evaluate the total plate supply requirements in volts, amperes, and watts.
2. Using the theoretical diagram of a typical television receiver, calculate the heater power requirements.
3. Using the theoretical diagram of a typical television receiver and assuming a transformer efficiency of 95 per cent, calculate the a-c current it requires from the supply mains.
4. Calculate the heater power for typical tubes used in the following sections: frequency changer, i-f amplifier, v-f amplifier, sound output, horizontal oscillator, vertical output, horizontal output, damper, rectifier, EHV rectifier.
5. Compare parallel heater wiring using two wires and using one wire and ground.
6. Explain how hum could be induced from heater to cathode and what its effects would be.
7. Total heater current in a receiver reaches 7.2 amp. Calculate the size of wire required to carry it if the voltage drop is not to exceed 0.1 v over a 1 m length of wire.
8. A 5Y3 tube is to be fed from a 6.3-v line. Calculate the value and power dissipation of the series dropping resistor.

9. Compare parallel and series connections of heaters.
10. A series chain contains five 6CB6 tubes and is fed from a 116-v main. Determine the value and power rating of the dropping resistance. Calculate the power efficiency of the arrangement in terms of useful heater power.
11. A 6AB4 and two 6AK5 tubes must be fed heater power by a 0.3-amp series heater chain. Design the necessary circuit.
12. Plot the reactance value at 60 cps and 120 cps of a capacitor between 30 and 300 μf .
13. Plot the reactance value at 60 cps and 120 cps of an inductor between 1 and 20 henrys.
14. An a-c voltage of 250 volts rms is applied to a rectifier. Calculate the raw rectified d-c value when a 200- μf capacitor is connected at the output of the rectifier and when it is disconnected.
15. Design a single transformer power supply to provide separate 200- and 300-v d-c supply using separate rectifiers.
16. Design a single transformer power supply to provide +250 and -250 v with respect to ground.
17. On a standard transformer power supply explain why the winding feeding the plates of the rectifier must have a grounded center tap.
18. Design an autotransformer power supply providing +150 and -60 v with respect to ground.
19. Compare Latour and Schenkel voltage doublers.
20. Design a voltage-multiplying circuit for a multiplying factor of 5.
21. Design a voltage-multiplying circuit for a multiplying factor of 6.
22. Design a transformerless a-c power supply providing +130 and -250 v with respect to ground.
23. Explain with figures why purely a-c—d-c power supplies are rarely used in television receivers.
24. Refer to a tube handbook and design a gas-tube voltage regulator for 105 v and 30 ma. The available voltage is 150 v.
25. Refer to a tube handbook and design a gas-tube stabilizer for 255 v and 18 ma. The available voltage is 375 v.
26. Refer to a tube handbook and design a pentode current stabilizer for 30 ma. The available voltage is 300 v. The stabilized circuit has a resistance of 3000 ohms.
27. It is suggested that a-c voltage stabilization could be effected by two clipping diodes suitably biased. Design a circuit and discuss its performance.
28. Discuss the possibilities of using a battery or accumulator supply for a television receiver.

Chapter 19

PICTURE TUBE HIGH-VOLTAGE SUPPLY

Picture tubes require extra-high accelerating voltages, or EHV, to produce sufficient brightness—20,000 volts is a typical value. Current consumption is low and reaches only a fraction of a milliampere. Consequently, EHV power is on the order of a few watts, so that it becomes an entirely practical matter to produce EHV with the help of special circuits. Such an arrangement is by far more economical than a type of supply converted from the supply mains as in the previous chapter.

A-f, r-f, and pulse oscillators may be used as EHV power sources. However, the most common arrangement is the flyback type, which takes advantage of the high voltage transients produced in the horizontal deflection circuits.

19-1. EHV supplies

The final accelerating voltage for the picture tube, or extra-high voltage EHV, increases with the size of the tube. Values of 10 to 20 kv are common in domestic television receivers. Higher values, up to 100 kv, are used in professional or projection-type receivers.

Domestic television receiver EHV current would be on the order of 100 microamperes. Assuming a typical EHV output of 15 kv, the power needed by the EHV circuit is

$$15,000 \times 0.0001 = 1.5 \text{ w}$$

which is a small amount and explains why special types of circuit can be designed to supply EHV.

With such low current and power consumption, it becomes entirely practical to use a resistance-capacitance filter. The voltage drop across a 1-megohm resistor would be only

$$1,000,000 \times 0.0001 = 100 \text{ v}$$

which is negligible at EHV. This makes possible the use of a high resistance and a small-value capacitance. Since the capacitor must be insulated for the full EHV, it is economically very interesting to be able to use a small value. As we pointed out in Chap. 13, such a capacitor may even be an integral part of the tube.

Because also of the low current consumption, single-alternation rectification is employed, so that a single diode or rectifier can be used. It can be a low-power component, and the only problem will be that of sufficient insulation.

Four types of EHV supplies are used: mains, oscillator, pulse, and flyback.

19-2. Mains supply

The mains EHV supply is a classical half-wave, single-alternation rectifying circuit as shown in Fig. 19-1. The rectifier may be a diode as shown or a semiconductor element, in which case the heater winding is not necessary and the circuit reduces to that of Fig. 19-2.

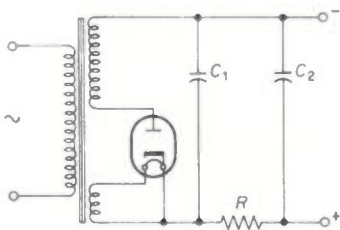


Figure 19-1. Half-wave EHV rectifier.

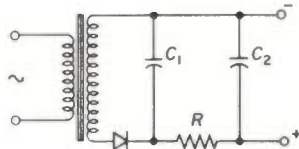


Figure 19-2. Use of a semiconductor rectifier.

Filtering is accomplished by the pi filter R - C_1 - C_2 . When R has a high value, the voltage drop across it may reach several hundred volts and the resistance must be chosen accordingly. Alternatively, several resistors may be connected in series to reduce the voltage across each component. A higher resistance is sought because, as we have just indicated, smaller capacitors can be used. This is especially interesting in the present case because the fundamental of the hum component is at 60 cps. On the other hand, a high resistance reduces the regulation of the EHV supply—that is, it changes for the worse the variation of voltage with variations of current.

Capacitor values of $0.1 \mu\text{f}$ constitute a minimum. To economize one capacitor, the diagram of Fig. 19-3 is useful. The output filter capacitor C_2 is omitted. R is now a protective resistor to limit current in case the power supply is short-circuited. R_1 is a good safety measure which should be present on all a-c mains types of EHV supply. Its function is to discharge C_1 when the power is turned off. Its value is large enough for its shunting effect to be negligible as far as EHV output voltage is concerned.

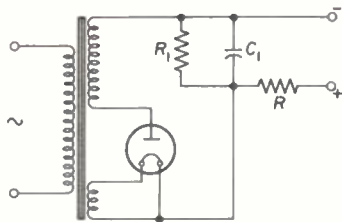


Figure 19-3. Simplified filter circuit and safety resistance.

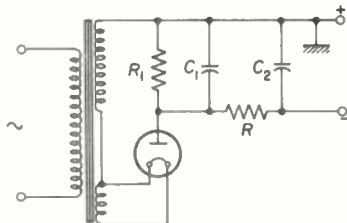


Figure 19-4. Negative voltage supply.

The insulation of the diode must withstand approximately three times the rms a-c voltage. Assume this voltage is 10,000 volts rms, that is,

$$10,000 \times 1.4 = 14,000 \text{ v peak}$$

C_1 charges up to 14,000 volts approximately on a positive alternation. On the following negative alternation, the diode anode reaches a negative peak voltage of 14,000 v. Assuming negligible discharge of the capacitance, its cathode is still at +14,000 v, so that the peak inverse voltage, or PIV, across the diode is

$$14,000 + 14,000 = 28,000 \text{ v}$$

In other words, the PIV on the diode is

$$2 \times 1.41 = 2.82$$

which, as we said before, may be rounded off roughly to three times the rms a-c voltage.

The same problem arises with the transformer, because the same PIV appears between heater and EHV windings. The transformer must be well insulated, and becomes more costly as the voltage increases.

Regarding the diode, its anode is generally connected to an anode cap at the top of the tube to improve insulation.

This type of EHV supply is mainly used when the voltage does not exceed a few kilovolts (as in the case of electrostatic tubes) because of its simplicity and its excellent intrinsic regulation. It is commonly found in oscilloscope diagrams. In these circuits, generally, the positive pole of the EHV is grounded

and the negative pole is insulated. A minor modification of the circuit, as shown in Fig. 19-4, accomplishes this.

19-3. Voltage multipliers

To facilitate transformer design, EHV outputs can be obtained from a moderate a-c supply voltage by the use of voltage multipliers. However, this method results in poorer EHV regulation.

The diagrams of voltage multipliers in Chap. 18 may be used as such for EHV. It is generally economical to use semiconductor rectifiers, because tubes require heater windings with attendant insulation problems.

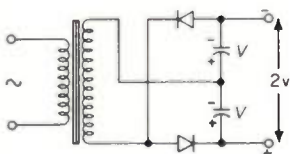


Figure 19-5. Voltage-doubler circuit.

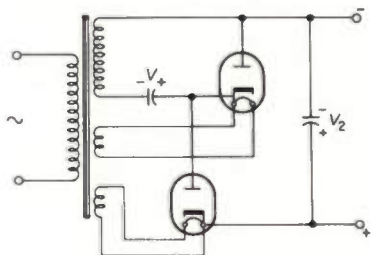


Figure 19-6. Tubes in a voltage-doubler supply.

Both types of voltage-doubler circuit previously discussed may be employed. The Latour doubler is illustrated in Fig. 19-5 for semiconductor rectifiers, and the Schenkel doubler is shown in Fig. 19-6 for tube rectifiers.

In this type of supply, the Latour doubler is more interesting because each capacitor must withstand only the a-c peak voltage. Moreover, the fundamental hum component is at twice the frequency of the supply mains. These two considerations justify the use of capacitors of lower value and lower working voltage than is the case with the Schenkel doubler, with resulting economy.

The usual disadvantage of the Latour doubler—no pole of the a-c supply is grounded—is of no importance in the present case of EHV. In fact, either the negative or positive side of EHV can be grounded according to the polarity desired for EHV.

As for plate power supplies, the principle of voltage multiplication can be extended to higher orders. An example is shown in Fig. 19-7, where a cascade multiplier uses seven semiconductor rectifiers fed by the plate supply winding of the mains transformer. The first two rectifiers R_1 and R_2 , with capacitor C_1 , constitute a Schenkel doubler which charges C_4 up to twice the peak a-c voltage. In addition, the negative side of R_1 is not grounded, but connected to the B+ of 350 volts, thus added to the voltage on C_4 . The charging process of the capacitors thus proceeds at each alternation, each rectifier charging its

capacitor up to the peak a-c value. For 350 volts rms a-c input, the peak value is

$$350 \times 1.41 = 500 \text{ v}$$

approximately. C_1 charges up to 500 v, and all the other capacitances up to 1000 v. The total voltage will be

$$500 + (5 \times 1000) = 5500 \text{ v}$$

to which are added 350 v of B+ as seen, bringing the available EHV to 5850 v across C_7 .

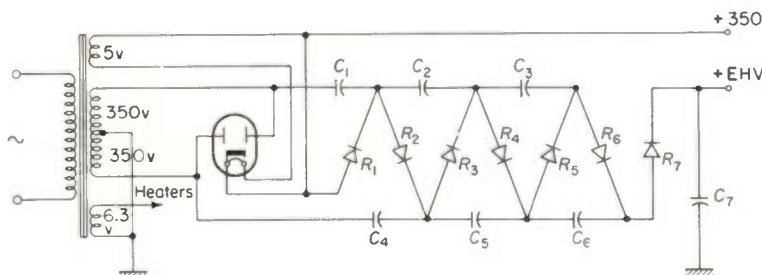


Figure 19-7. Cascade voltage multiplier.

Notice that, except for C_7 , no single capacitor has to withstand more than 1000 volts.

19-4. Advantages and disadvantages

An EHV mains supply is simple and rugged. It enjoys better regulation than the other systems, even if voltage multiplication is also employed, due to the fact that the input power is practically unlimited and that the supply internal resistance is low. Consequently, an increase in current reduces the voltage very little. This advantage is also a drawback. From a safety point of view, a mains EHV supply constitutes a dangerous hazard that is all the more serious if the EHV is high. But this is not its only disadvantage.

The supply is generally bulky, heavy, and costly, mostly because of insulation difficulties which increase rapidly with EHV. It is rarely used if the voltage exceeds a few kilovolts and is never found in domestic receivers.

19-5. A-f oscillator

Since the necessary EHV power is on the order of 1 or 2 watts, it is entirely sensible to think of producing the corresponding a-c power separately. The major advantage is that a frequency well above the frequency of the mains can

be used, with attendant reduction in cost. For example, a frequency in the audio range would simplify filtering problems greatly as well as transformer design.

This principle has been used in the circuit shown in Fig. 19-8. The few watts of a-c power are produced by an audio output tube in an a-f oscillator circuit. The fact that the available power is limited by the tube is an advantage. If the EHV current increases beyond reasonable limits, the a-c source cannot satisfy the power requirements, and the voltage immediately drops. Accidental

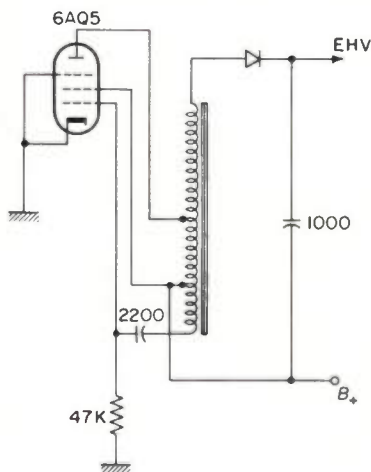


Figure 19-8. AF oscillator type of EHV supply.

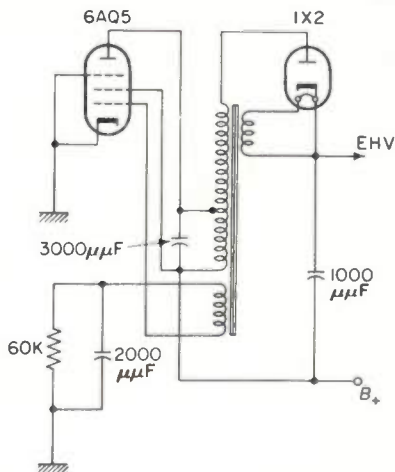


Figure 19-9. Heating the rectifier by an AF current.

contact with such EHV, then, is not generally deadly, the voltage almost disappearing on contact. This type of safety is in some measure common to all EHV supplies using a limited power source—that is, to all types except the mains supply discussed in the previous paragraph.

In Fig. 19-8 oscillation is due to the plate-to-grid coupling produced by the output transformer. As is usual with oscillators, grid bias is obtained by grid current. A step-up winding is added to the primary to bring the a-f voltage to the required value, and a semiconductor rectifier provides the d-c EHV. The filter is reduced to a single capacitor, which may have a relatively small value because of the high frequency. Notice that the low side of the filter capacitor is connected to B+, which in effect adds B+ voltage to the available EHV with respect to ground.

A somewhat similar circuit, shown in Fig. 19-9, uses a transformer at the output. The plate winding is tuned by a capacitor to the oscillating frequency, 15,000 cps approximately. The rectifier is a tube, and its heater power is provided by a small winding on the output transformer.

A-f power supplies can easily provide EHV outputs of 10,000 v or more, if necessary, with the help of voltage doublers. Their main disadvantage is due to the use of an iron core transformer.

19-6. R-f oscillator

By using a still higher frequency—that is, in the r-f range—the transformer may be air-cored or use a simple powdered iron core. Figure 19-10 illustrates

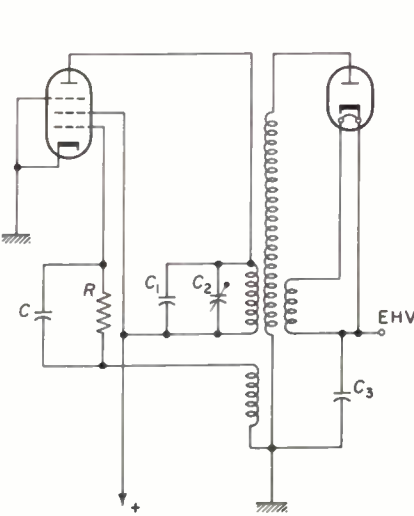


Figure 19-10. RF oscillator type of EHV supply.

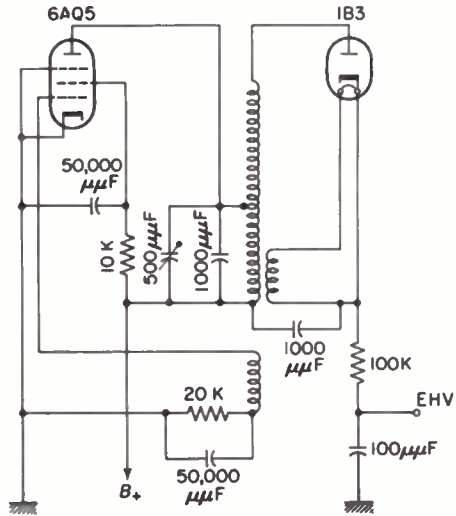


Figure 19-11. Typical RF supply circuit.

the basic design. A power tube is connected as an r-f oscillator with plate-to-grid coupling. Grid bias is obtained by grid current with the usual RC arrangement in the grid circuit. A high-voltage winding is coupled to the oscillator and drives a rectifier whose heater is also fed an r-f voltage. A filter capacitor C_3 is provided, which may have a relatively low value because of the high frequency.

The high-voltage winding, with its total parasitic shunt capacitance, resonates at a given frequency. The oscillation frequency is adjusted by tuning the plate winding with a fixed and a variable capacitor.

EHV output reaches a maximum when the oscillating frequency coincides with the high-voltage secondary resonant frequency, so that adjustment of the oscillating frequency constitutes a convenient way of controlling EHV output.

A practical diagram is given in Fig. 19-11. It differs from Fig. 19-10 by the use of an autotransformer at the output. Also, the input filter capacitance is

connected to $B+$, to add plate voltage to the EHV output. A full pi filter is included.

19-7. Winding

The r-f transformer or autotransformer has a special winding pattern. In particular, the EHV secondary has a high voltage across it and is made of several separate coils so that insulation difficulties are avoided. At the same

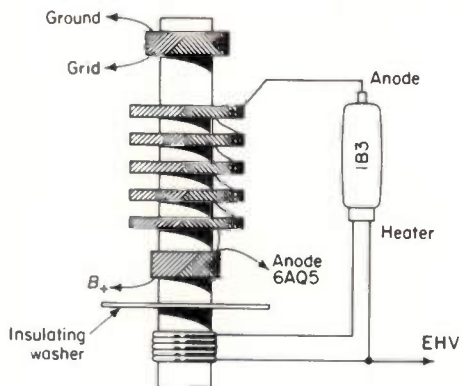


Figure 19-12. Practical disposition for an RF power supply.

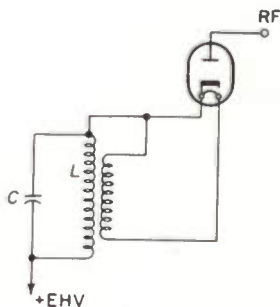


Figure 19-13. Heating the rectifier by a tuned transformer.

time this reduces parasitic capacitance and increases the quality factor of the coil winding.

Grid and plate windings are placed on each side of the EHV secondary.

The heater winding is made of a few turns of wire, strongly insulated, and placed near the anode winding. The general arrangement is illustrated in Fig. 19-12.

Such transformers provide EHV outputs up to 10,000 v without much difficulty. Beyond this value, it may be advisable to use voltage multipliers. When the required EHV power is high, a large-power tube may be used. Alternatively, several tubes may be connected in parallel. The circuit can deliver EHV currents of several milliamperes up to 40 kilovolts.

The major problem, of course, is insulation, especially with respect to the heater winding. A neat way of solving the difficulty is illustrated in Fig. 19-13. A fraction of the available r-f energy goes through the diode because of anode-to-cathode capacitance. It is collected by a tuned circuit LC in the cathode circuit, and fed to the heater through a step-down winding. The small tuned-primary r-f heater transformer is totally independent and can easily be strongly insulated. Heater power is adjusted by tuning the primary.

19-8. Pulsed tube

When a pulsed current is made to flow through an inductor, high values of voltage are developed across the inductor, as occurred in the case of the horizontal output transformer.

These high-voltage peaks can be put to work to provide EHV. The basic diagram appears in Fig. 19-14. A power tube is turned on and off by a pulse voltage applied to its control grid. The anode load is inductive and high

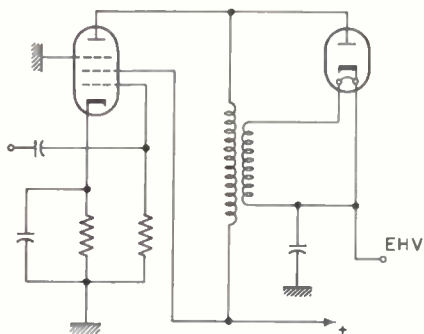


Figure 19-14. Pulse type EHV supply.

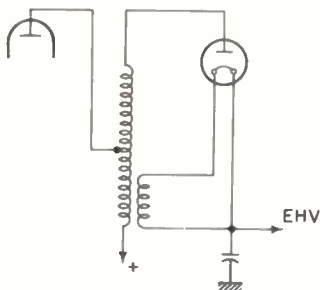


Figure 19-15. Autotransformer for pulse voltages.

transient voltages appear on the anode. They are rectified by a diode to provide EHV. The rectifier tube can be heated by a small winding coupled to the anode inductance as shown.

When the pulse voltage at the anode of the power tube is insufficient, the autotransformer principle can be used, shown in Fig. 19-15, or voltage multipliers can be employed.

Grid pulse voltage can also be provided by a relaxator. Furthermore, the tube itself can be connected as a relaxator. The waveform on the grid has little importance as long as it produces a steep variation of current every cycle.

The anode of the tube must withstand high pulse voltages and must be insulated accordingly—for example, with a top anode cap. The requirements are similar to those that apply to a horizontal line output tube, with the difference that much less power is necessary; a few watts are sufficient for EHV purposes.

The inductor also has to withstand high pulse voltages, and, as for the r-f supply, it is frequently made of several cascaded windings. Because of the relatively low frequency, it is advantageous to provide a magnetic core which allows a large reduction in the number of turns. Ferrite cores have also been used for this purpose.

19-9. Pulse-type supply

An EHV power supply of the pulsed variety appears in Fig. 19-16. It is designed to feed 25,000 v to a projection tube whose current may reach several hundred microamperes. A triode-connected pentode is used as a blocking oscillator on a frequency of the order of 1000 cps. It drives a power tube whose anode load is inductive. On top of the anode inductor, a step-up winding brings the peak voltage up to 9000 v.

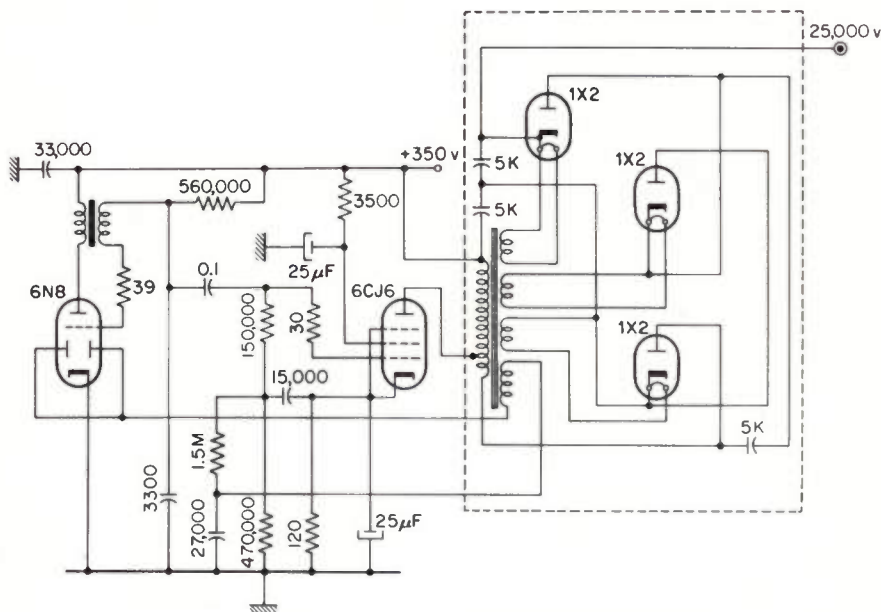


Figure 19-16. Voltage tripler 25,000 volts supply.

A voltage tripler using three small EHV rectifiers provides 25,000 volts d-c. The heaters are fed by small windings on the output autotransformer. The entire EHV circuit inside the dotted line is contained in an oil-filled casing to obtain sufficient insulation.

Since intrinsic regulation is poor, an auxiliary circuit has been added to improve it. A separate winding on the output autotransformer samples the EHV pulse voltage, which is rectified by the diodes of the relaxator tube. A negative d-c voltage is obtained which is proportional to pulse voltage and used as grid bias for the power tube.

When the EHV has a tendency to decrease, say because of an increase in EHV required current, the voltage rectified by the diodes decreases. The bias on the output tube decreases, and the tube delivers more power to its plate load.

The operation of this regulation circuit is strongly reminiscent of that of an AVC circuit.

An oscillator or pulse-type EHV supply is independent of the other circuits of the receiver. Its power is limited, and it does not generally constitute a lethal hazard. EHV can easily be adjusted if necessary.

Despite the fact that it requires a power tube, special transformer, rectifiers, and filter, a pulse-type supply is more economical than a mains type of supply, less bulky, and lighter.

Regulation is none too good unless special circuits are provided. In the case of an r-f oscillator, the oscillating frequency must be chosen so that it does not interfere with radio reception. Sometimes, the entire EHV section must be shielded.

19-10. Flyback-type supply

From the pulsed type of power supply to the flyback type is but a small step.

In a pulsed supply, the circuit essentially contains a relaxator driving a power tube with an inductive anode load. This is precisely what is found in the horizontal sweep section of a receiver.

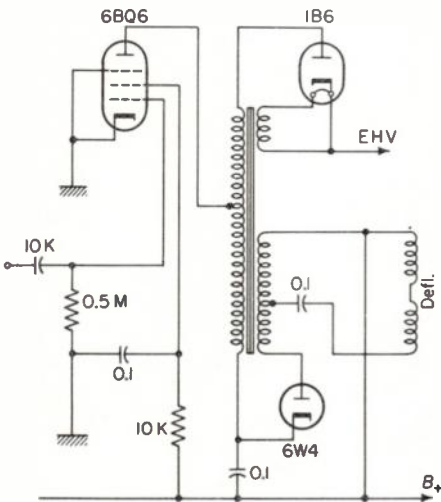


Figure 19-17. Horizontal flyback EHV supply.

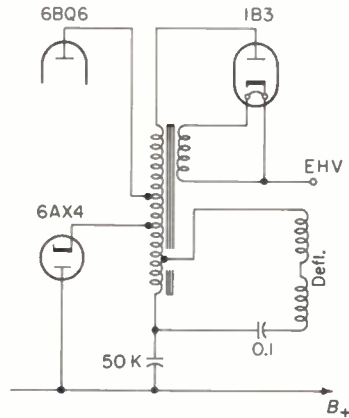


Figure 19-18. Autotransformer output and EHV circuit.

Very little need be done to obtain EHV from the horizontal line output tube as shown in Fig. 19-17. The anode winding has a step-up section to bring the pulse voltage to a suitable value, and a rectifier transforms it into direct current. The rectifier is heated by a small extra winding on the line transformer.

The economy is evident. It is not much more difficult to wind the output transformer with the EHV winding, and the rectifier and eventual filtering circuit must be provided anyway whatever the type of EHV supply. The output voltage may reach 20,000 volts without much difficulty.

Because of its simplicity and economy and despite some shortcomings, this circuit is in almost universal use in television receivers.

EHV power is evidently borrowed from the line output tube, which is already put to hard work. However, recent tubes and high-efficiency ferrite transformers make for an adequate power reserve, and EHV regulation is no worse than that of the other systems. It is not easy to adjust EHV output, and it is set at a fixed value by the transformer characteristics.

Figure 19-18 illustrates the application of this idea to the case of an autotransformer. A number of practical diagrams have been given in Chap. 17 which deals with the horizontal sweep section.

19-11. Voltage multiplication

When high EHV voltages are needed, insulation becomes a problem above, say, 20 kilovolts. The usual solution in such cases is to use voltage multipliers, but owing to the shape of the pulses, a problem arises which we have not investigated.

A classical tripler circuit is given in Fig. 19-19. Point *B* is grounded or connected to *B+*, so that the diodes are fed the voltage appearing between

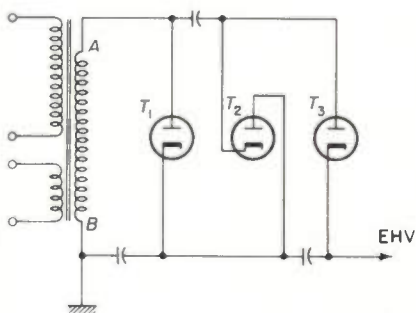


Figure 19-19. Triple diode circuit.

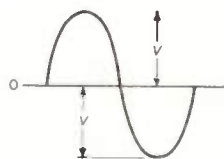


Figure 19-20. Peak voltages for a sine wave.

point *A* and ground. Then T_1 and T_3 will work on the positive alternation and T_2 on the negative alternation.

When the applied voltage is sinusoidal, the waveform is symmetrical with respect to the zero axis, and the positive and negative peak voltages have identical values V , as shown in Fig. 19-20. The three diodes work on the same a-c voltage, and the rectified d-c voltage is effectively tripled.

This is no longer true, however, if the applied voltage is a pulse waveform. In this case, positive peak voltage has a value V_1 and negative peak voltage a different value V_2 , as shown in Fig. 19-21. Diode T_2 rectifies V_2 and contributes very little to the output d-c voltage, which is scarcely higher than twice the

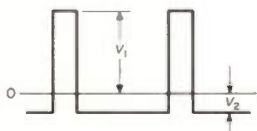


Figure 19-21. Peak voltages for a pulse waveform.

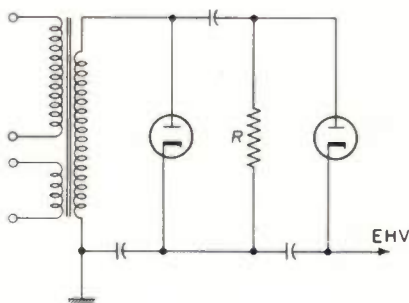


Figure 19-22. EHV supply circuit using a resistance instead of a diode.

peak a-c voltage. It is then sensible to omit T_2 and replace it by a resistor R , shown in Fig. 19-22, in which case the circuit is practically a voltage doubler, whereas the circuit containing T_1 and T_2 would not double the voltage. Resistor R is on the order of 1 or 2 megohms.

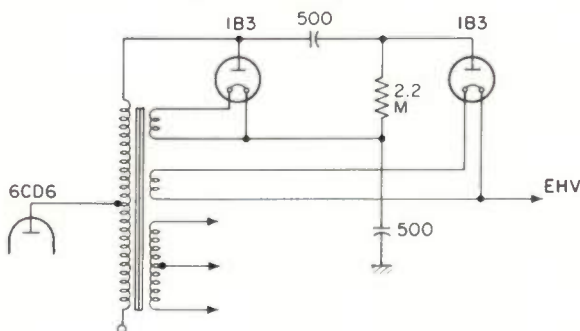


Figure 19-23. Voltage doubler with two diodes.

A practical circuit is given in Fig. 19-23. It uses two diode tubes heated by two well-insulated heater windings on the output transformer.

Notice that if a voltage multiplier uses five diodes in a circuit similar to that shown in Fig. 19-19, the d-c output voltage will be approximately three times the peak a-c input voltage, because two of the diodes will contribute very little to EHV output. They can then be replaced by resistors, and the circuit of Fig. 19-24 results. As can be seen, all diodes are connected so as to conduct on the positive alternations.

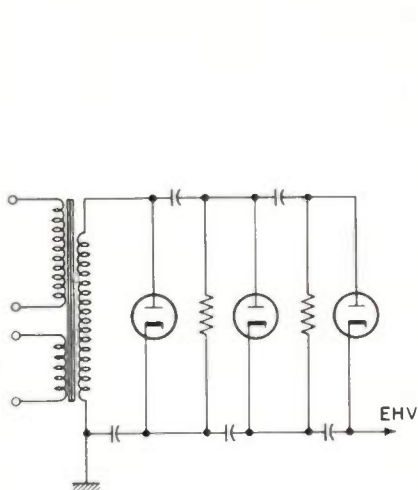


Figure 19-24. Voltage tripler with three diodes.

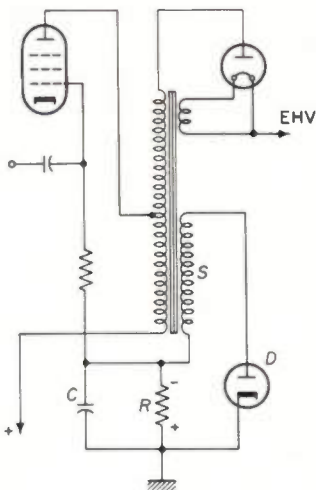


Figure 19-25. Diode EHV regulation.

19-12. Automatic regulation

The intrinsic regulation of the EHV supply is not too good; it may be improved by a number of automatic regulation circuits. In Fig. 19-25, a separate winding S provides a voltage proportional to the pulse voltage. It is rectified by diode D , and a negative d-c voltage appears across RC as shown and is used as a bias for the output tube. The power provided by the output tube is then controlled by the bias in such a way that the peak pulse voltage is kept constant. This type of circuit is used in the supply of Fig. 19-16.

Another arrangement is illustrated in Fig. 19-26. A fraction of the EHV output is taken across the resistance divider and applied to the grid of tube V_1 . The cathode of this tube is held at constant potential by a gas regulator N , fed from $B+$ through resistance R_1 . The variations of EHV output, reduced by the resistance divider, control the grid of V_1 and appear amplified at the anode. Since the screen grid of the power tube V_2 is connected to the anode of V_1 , the variations of EHV control the screen voltage of V_2 and therefore the power it delivers to the EHV circuit, thus maintaining the EHV output approximately constant. The principle of operation is similar to that used in series regulation of $B+$ supplies.

These regulation circuits are especially applicable to the case where the EHV is provided by a separate power tube.

When a flyback type of EHV supply is used, the same tube provides horizontal scanning, and it is necessary to use special circuits to prevent the EHV regulation from influencing the operation of the sweep system. An example of

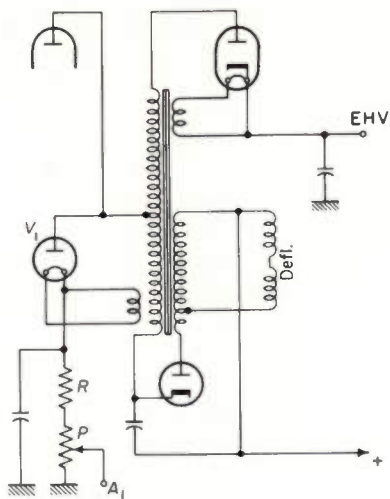


Figure 19-28. Two diodes produce two different EHV voltages.

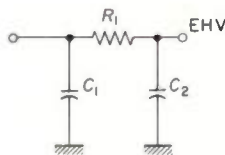


Figure 19-29. Resistance-capacitance filter.

19-14. Filter circuit

EHV filters are always of the resistance-capacitance type. In flyback-type supplies, the relatively high line frequency of 15,750 cps allows the use of capacitors of reduced values, usually 500 or 1000 μmf . This makes for economically acceptable capacitors despite the high working voltage.

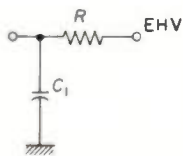


Figure 19-30. Simplified filter.

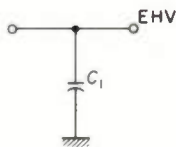


Figure 19-31. Single capacitance filter.

The standard filter circuit is the pi circuit shown in Fig. 19-29. For economy, the output capacitor is frequently omitted, and the filter reduces to the circuit shown in Fig. 19-30. R provides some protection if the EHV is accidentally shorted. It can be omitted, in which case the filtering is due only to a capacitor C_1 as shown in Fig. 19-31.

As we explained in Chap. 13, the glass envelope of the tube carries two conductive coatings, one on the inside and one on the outside; they constitute

a high-insulation capacitor, which has a value of several hundred micro-microfarads and is used as C_1 of Fig. 19-31. In this case, *no* filtering component is required in the EHV circuit. The filter capacitor is integral with the tube.

The outside conductive coating of the picture tube should be well grounded, generally through some spring contact arrangement.

Problems

1. Compare the advantages and disadvantages of half-wave, single-alternation and full-wave double-alternation rectification to provide EHV.
2. Compare a tube diode and a semiconductor diode as EHV rectifiers.
3. An a-c voltage of 1200 volts rms is rectified by a diode. Evaluate the peak inverse voltage across the diode.
4. A high peak inverse voltage appears between the heater winding and the high-voltage winding of an EHV transformer for the classical rectifier circuit. Design a circuit in which this drawback is avoided.
5. Explain why an EHV mains supply for 15,000 volts d-c is not an economical proposition.
6. Compare Latour and Schenkel voltage-doubler circuits for EHV supplies.
7. For an EHV supply of 16 kv, compare the single-rectifier and voltage-doubler circuits.
8. Assuming a transformer efficiency of 78 per cent, calculate the power delivered by the power tube in an a-f oscillator type of power supply. The rectifier is a 1B3 and provides 12,000 v, 0.15 ma.
9. Compare a-f oscillator, r-f oscillator, and pulse types of EHV supply.
10. Discuss the advantages and disadvantages of a flyback type of EHV supply.
11. Recommend an oscillating frequency for an r-f type of EHV supply and justify your choice.
12. Recommend an oscillating frequency for an a-f type of EHV supply and justify your choice.
13. Would a flyback type of EHV supply cause interferences with radio receivers?
14. Flyback EHV output could be obtained by using an autotransformer on the primary and a single diode according to classical circuitry. It could also be provided by a voltage doubler fed directly from the plate of the line output tube. Compare the two methods.

15. Design a projection tube EHV supply for 50 kv and 1 ma, using a tube handbook to select tubes.

16. Using a tube handbook, design an a-f type of EHV supply for an electrostatic tube which needs 3000 v, 60 μ amp and 1000 v, 25 μ amp.

17. Discuss the effects of a variation of EHV output on the picture for a flyback type of supply.

18. When an EHV supply of the flyback type has poor regulation, explain what the results are on the picture.

19. It is suggested that the vertical sweep output stage could be made to produce EHV. Investigate this suggestion.

20. Discuss the apparent effect on the picture when a strong percentage of ripple is present in a flyback type of EHV supply.

Chapter 20

AUXILIARY CIRCUITS

Although they are not basic to the fundamental operation of a receiver, auxiliary circuits are always added to the design to improve performance and facilitate the consumer's use of the receiver. Some of the more common circuits which fall into this class are picture AGC, noise elimination, and blanking circuits.

20-1. Types of auxiliary circuits

A number of auxiliary circuits may be included in a television receiver either to improve performance or to facilitate the use of the receiver.

The most common auxiliary circuit is automatic gain control (AGC) of the picture receiver, which is incorporated in one form or another in almost all receiver diagrams. Its function is to correct for signal amplitude variations, whether they occur from station to station or during reception of one station.

Noise eliminating circuits are another fairly common feature; they are mainly used in the synchronizing section to avoid tearing of the picture in the presence of noise.

The immunity to noise of those FM receivers incorporating some form of amplitude limiter makes it unnecessary to include anti-noise circuits in the sound circuits of the receiver.

Noise eliminating circuits may be used, however, in the video section of the picture receiver, in which case they generally do not use a separate tube. The working conditions of one v-f tube may be adjusted so that it acts as a limiter.

detection is automatically applied between grid and cathode and controls the anode voltage which is used as the AGC. The tube has no applied plate voltage. The negative voltage on the anode is due to electrons that are emitted by the cathode shooting through the grid and reaching the plate.

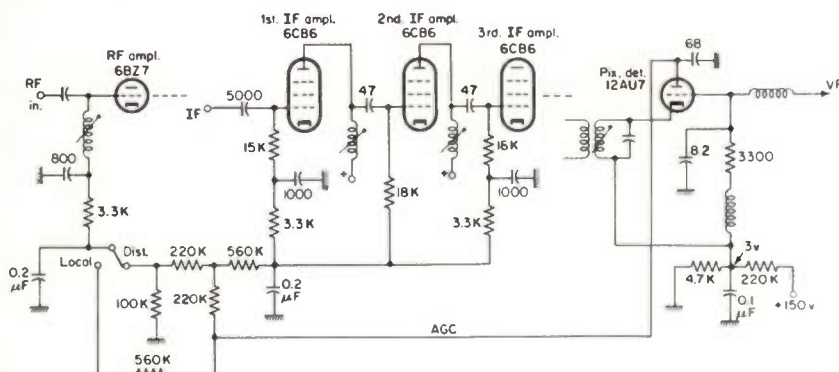


Figure 20-3. Complete AGC system using a triode detector.

A small positive voltage, on the order of 3 volts, is applied to the grid and cathode circuit and acts as a delay voltage, so that no AGC voltage appears on the anode for weak stations and the receiver then works at maximum gain.

Also, a commutation is provided for the AGC on the r-f amplifier. On local stations it receives the full AGC, and thus the risk of saturation by a strong signal is eliminated. On distant stations, it receives only a fraction of the AGC voltage and sensitivity is increased.

20-3. Peak detector

Properly, the AGC voltage should depend on signal level and nothing else. There is an easy reference value, proportional to signal level, in the video

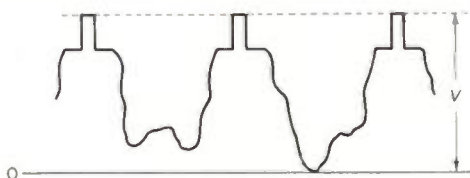


Figure 20-4. Peak value of a v-f signal.

waveform. This reference is the level of the bottom of the sync pulses or in other words, the peak amplitude V of the signal shown in Fig. 20-4 under the condition that it is in d-c form, e.g., when its d-c component is present.

Such a signal can be found in a number of places in a receiver, such as the grid of a v-f stage where d-c restoration takes place, or where the d-c component has been transmitted, or also at the grid of a grid detection separator.

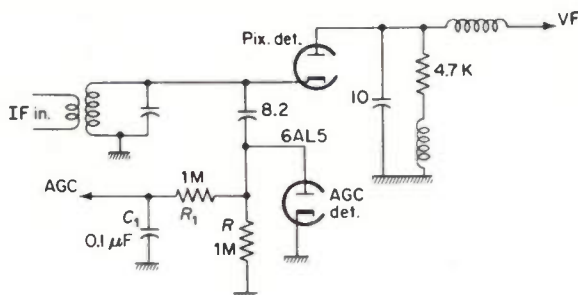


Figure 20-5. Separate AGC detector.

It is common practice to produce it separately simply by using a peak detector—that is, a diode with a high load resistance. Figure 20-5 illustrates how the principle works. A double diode is used; one half is the picture

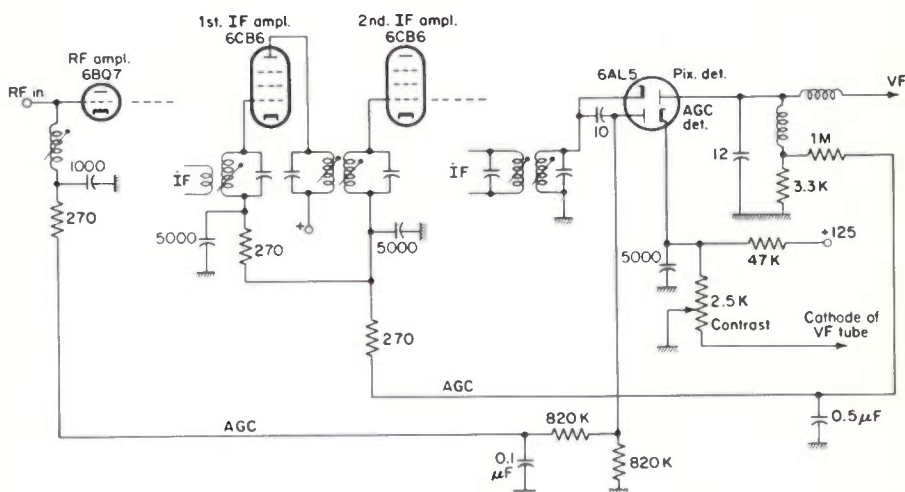


Figure 20-6. Double AGC system with contrast adjustment.

detector and the other half is the AGC detector, which is of the parallel-diode variety. The peak voltage appears across R and is filtered by R_1C_1 as before, with the same effects regarding the speed of action of the AGC.

Both systems may be combined, as illustrated in Fig. 20-6. The d-c voltage appearing across the load of the picture detector controls the first two i-f amplifiers.

A separate detector is used for AGC and provides the control voltage for the r-f amplifier. The cathode of the AGC diode is set at a positive voltage through a resistance bridge between +125 volts and ground. In this way the AGC diode starts to conduct and delivers an AGC voltage only when the r-f voltage on the anode is higher than the d-c delay voltage on the cathode. The delay voltage is adjustable and is, in fact, combined with the contrast control, which modifies the cathode resistance of the v-f amplifier tube so that the delay is automatically a function of the contrast setting—that is, of the strength of the station received.

20-4. Separator voltage

The standard grid detection sync separator aligns the v-f signal on its grid, so that the d-c negative voltage developed across the grid resistance is proportional to v-f signal amplitude and can be used as AGC voltage.

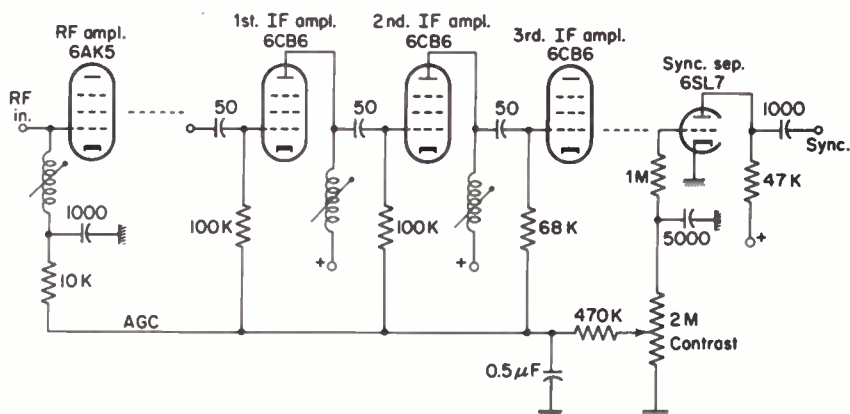


Figure 20-7. AGC voltage on separator grid.

A diagram based on this principle is shown in Fig. 20-7. A triode separator is used. A fraction of its negative grid voltage is used for AGC, after filtering by a long time-constant circuit. It controls directly the r-f amplifier and the first three i-f amplifiers.

Adjusting the amount of AGC voltage used provides a convenient contrast control as indicated.

20-5. Noise sensitivity

In gain control systems where the average d-c component is used, noise pulses do not modify the AGC voltage appreciably unless they are exceptionally dense and of large amplitude. For this reason, when the d-c voltage

produced by the picture detector is used, noise is rarely troublesome as far as the AGC circuit is concerned.

It is a different matter when a separate diode is used as a peak detector for AGC. When a noise pulse rides on the signal, as at X of Fig. 20-8, the instantaneous peak level rises to V_1 instead of the normal V , and the diode load capacitor charges up to V_1 . After that, the excess charge leaks away slowly because of the long time constant of the load circuit, so that the gain of the receiver is reduced during the entire discharging process—not only during the pulse.

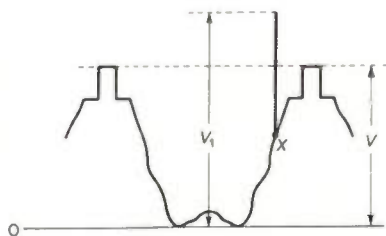


Figure 20-8. Noise pulse riding on the v-f signal.

The peak detection type of AGC is, therefore, very sensitive to noise pulses.

Even the vertical synchronizing signal is able to upset this AGC because the AGC diode, which conducts on the peaks of the i-f signal, will be switched on for a longer time by the vertical sync pulse. A component at field frequency is thus introduced, which must be eliminated by a long time-constant filter. The same filter will smooth out the effect of noise pulses and must be included. Its disadvantage, as we pointed out before, is that its long time constant, on the order of 0.1 second or more, prevents any quick reaction of the AGC circuit to fast-changing signal amplitude.

20-6. Amplified AGC

The AGC voltage may be amplified by a d-c amplifier stage. This offers no difficulty except that the d-c levels in the circuit must be adapted to the operating conditions. For example, it is very convenient to apply to the r-f and i-f grids a control voltage near ground level, so that no sophisticated bias circuitry is needed. Since the amplified AGC voltage will appear on an anode, this means that the anode load will return to ground.

Anode-to-cathode voltage for the tube, then, must be provided as a negative voltage on the cathode. Fortunately, a low voltage is sufficient for the purpose—20 volts is a typical value.

A simplified circuit appears in Fig. 20-9. The diode peak detector provides the AGC voltage across its load. It is then filtered and amplified by a d-c coupled triode fed —20 volts on its cathode. More exactly, the cathode voltage is adjustable and used as a contrast control.

Notice that because of phase reversal due to the triode, the AGC diode provides a positive voltage.

Strong noise pulses will drive the grid of the triode positive, creating grid

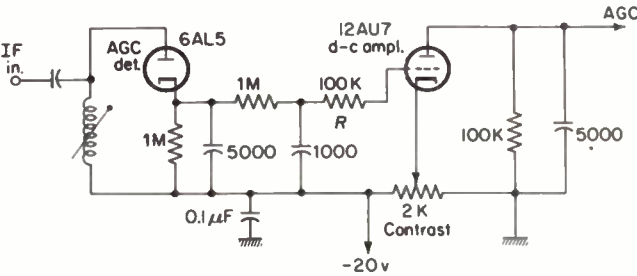


Figure 20-9. Amplified AGC signal.

current. This grid current flowing through the high grid resistance R will cut the tube off, thus suppressing or at least clipping the noise pulse. This is a common arrangement for this purpose.

20-7. Keyed AGC

Fast-acting AGC and a good amount of noise immunity are achieved through the use of a keyed AGC circuit.

The basic idea is that the peak signal voltage which the AGC tube should follow occurs only during horizontal sync pulses. There is then no need for

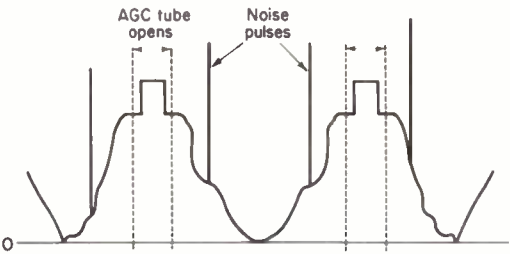


Figure 20-10. Gating the tube to eliminate most of the noise pulses.

the tube to work full time. If it works full time, any noise pulse superposed to picture modulation will upset the operation.

By restricting the working time of the tube so that it covers only the sync pulse (shown in Fig. 20-10), the system becomes immune to all noise pulses that do not occur during this period.

The AGC tube can easily be keyed or gated by a suitable pulse applied to any electrode. The gating pulse itself is readily available in the horizontal time base where it is produced during horizontal flyback—that is, at exactly the right time.

Usually the keying pulse is taken off the horizontal line output transformer and applied to the anode of the AGC tube. Figure 20-11 shows a simplified version of the circuit. The v-f waveform is taken off the plate of the first

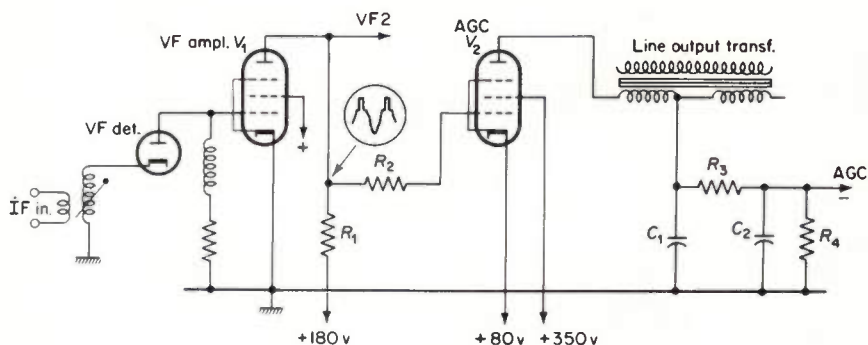


Figure 20-11. Keyed AGC circuit.

v-f amplifier tube V_1 , where the d-c component is present because of direct coupling. It is applied to the grid of the AGC tube V_2 along with the d-c plate voltage of the first v-f tube, so that for correct operation of the tube the cathode and screen must be connected to positive voltages as shown.

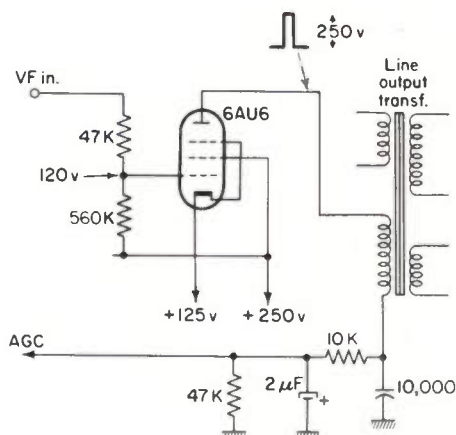


Figure 20-12. Practical gated AGC circuit.

The keying pulse is provided by a separate winding on the line output transformer. It has an amplitude of 400 volts, approximately.

A filter circuit $R_3C_1C_2$ is provided before the anode load resistance R_4 , across which appears the AGC voltage.

Figure 20-14, which illustrates a working practical solution to the problem, is based on the use of the two control grids of a heptode tube. The second control grid is connected for grid detection in the same way as for normal separation, and receives the full v-f signal from the v-f amplifier tube output. Plate and screen voltages are kept low to insure good separation.

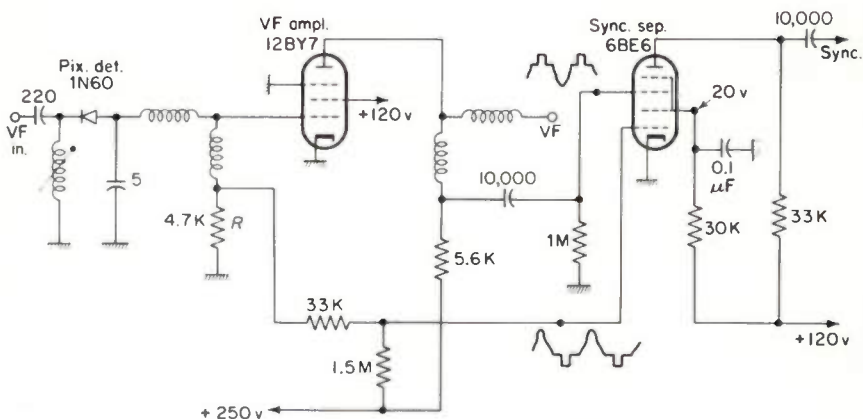


Figure 20-14. Noise elimination at sync separator.

The first grid also receives a v-f signal, but of opposite phase and at low level, coming from the detector output. Grid bias is adjusted so that the sync pulse tips just reach the cutoff voltage. In this way, any noise pulse of amplitude larger than the sync pulses cuts the tube off and prevents the same noise pulse, applied to the third grid, from upsetting synchronization.

Grid bias is fixed by the resistance divider between B+ and ground, which includes the detector load. It could be adjusted by making one of the resistances variable, thus providing a clipping level control.

20-9. Noise opposition

Another circuit for noise suppression is given in Fig. 20-15 in a somewhat simplified version.

The last i-f amplifier tube is adjusted to work normally for the v-f signal, but a large-amplitude noise pulse drives the grid positive and a negative pulse appears in the screen grid circuit. The same noise pulse will appear as a positive pulse at the anode of the first v-f amplifier. By combining both pulses across the same load resistor R , they will cancel each other, or at least give a greatly reduced resulting noise pulse. The sync separator input voltage is taken across R and a good amount of noise immunity is obtained.

LC is series-tuned on the i-f frequency and constitutes an efficient screen grid decoupling circuit for i-f while having almost no effect on the noise pulse.

A similar circuit can be designed by placing a load resistance in the screen circuit of the i-f amplifier, as shown in Fig. 20-16, and adding the voltage

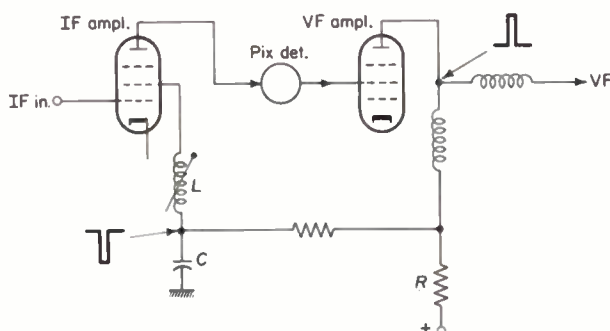


Figure 20-15. Noise elimination by opposition.

across it to the v-f signal at the grid of the sync separator through a suitable resistance divider.

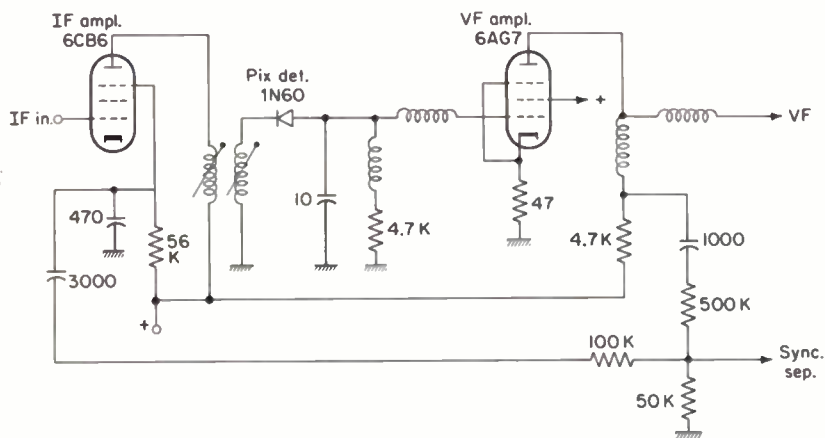


Figure 20-16. Practical diagram of noise opposition circuit.

Other noise suppression circuits have been devised. They are generally based on noise cancellation by opposition or else use the large amplitude of the noise pulses to shut off the signal during a noise burst.

20-10. Blanking circuits

Normally, scanning is not visible on the screen during retrace because the beam should be blanked out by the v-f waveform. Since this is not always the case, a blanking waveform is picked off the sweep sections of the receiver and

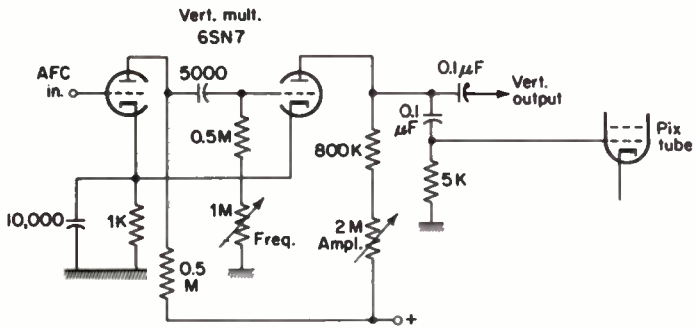


Figure 20-18. Vertical blanking pulse taken across a peaking resistance.

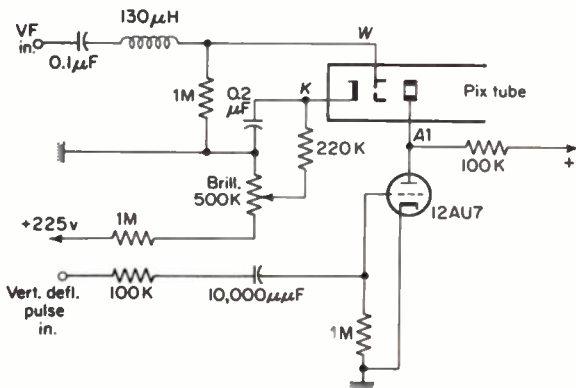


Figure 20-19. Blanking pulse amplifier tube.

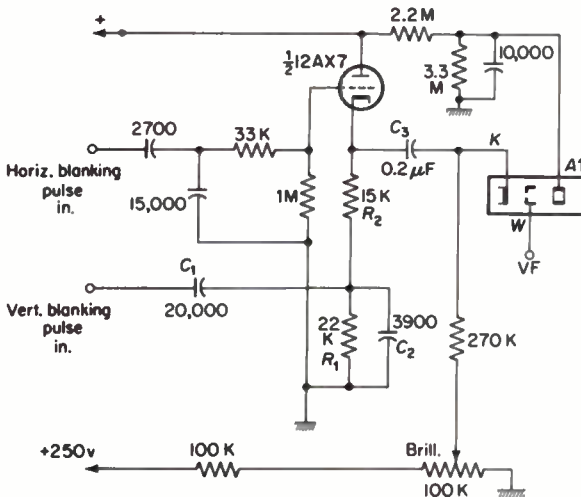


Figure 20-20. Combined horizontal and vertical blanking.

20-11. Contrast control

In a picture receiver automatic gain control, or AGC, takes care of signal level variations, whether they occur from station to station or are due to fading. An additional manual adjustment of gain is necessary, called the *contrast control*, because its visible result is a variation of contrast in the reproduced picture.

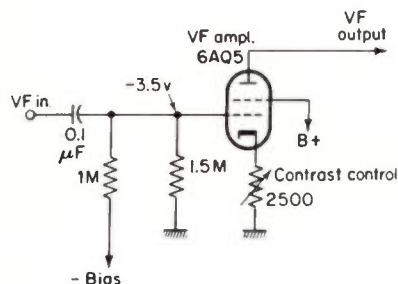


Figure 20-21. Contrast control in cathode circuit.

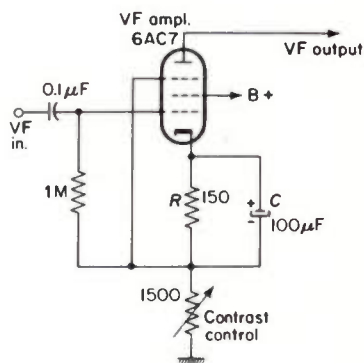


Figure 20-22. Cathode bias and contrast control.

Contrast control may be obtained by adjusting the fixed d-c level of the AGC line, or it may modify the gain of tubes other than those already submitted to AGC.

In this last case, contrast control is conveniently applied to the v-f amplifying section following the picture detector. Figure 20-21 illustrates a simple circuit. A fixed bias of -3.5 v is applied to the control grid of the tube. By adjusting the potentiometer in the cathode circuit, an additional bias is obtained which results in reduced stage gain. Notice that an undecoupled resistance in the cathode circuit results in current feedback which reduces the stage gain.

A variation of the circuit is represented in Fig. 20-22. The grid resistance returns to the top of the contrast control. R and C are the usual bias components, and the voltage drop across R is the grid-to-cathode bias. A variable amount of feedback is provided by the contrast control. Frequently the potentiometer is shunted by a fixed resistance. Also, as shown in Fig. 20-23, it is sometimes shunted by a low-value capacitor. The effect of this capacitor is to decouple the potentiometer and reduce feedback at the highest video frequencies, thus ensuring extra amplification at high frequencies. This is, of course, a form of high-frequency correction, called *cathode peaking*. Notice that here the amount of correction increases when gain is reduced—that is, on local stations, as it should.

Varying the screen grid voltage of the video amplifier will also provide gain control. A simple circuit for this purpose is shown in Fig. 20-24.

An example of a manual contrast control superimposed on AGC is represented in Fig. 20-25. The negative voltage appearing on the grid of the

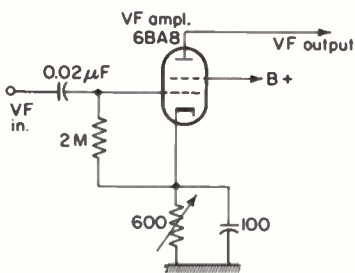


Figure 20-23. Peaking capacitor across contrast control.

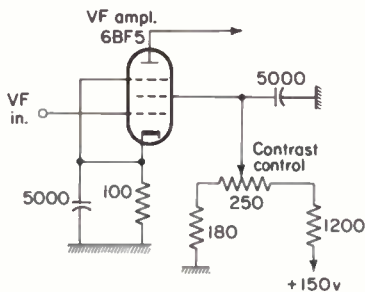


Figure 20-24. Contrast control in screen grid circuit.

sync separator can be used as AGC voltage, as shown previously. By adding a potentiometer, a contrast control results.

Several ideas may be combined. Figure 20-26 shows a circuit in which the action of the contrast control is twofold. Its right-hand section constitutes

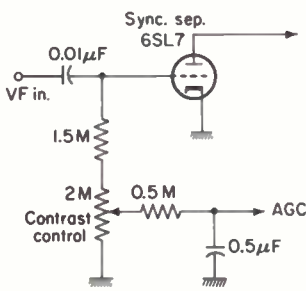


Figure 20-25. Using the negative voltage on the grid of the sync separator.

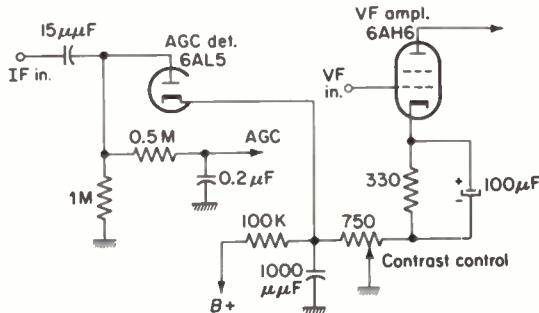


Figure 20-26. Combined contrast control circuit.

a cathode gain control for the video amplifier. At the same time its left-hand section is included in a resistance divider which determines the delay voltage applied to the cathode of the AGC rectifier. When contrast control is advanced, the gain of the video amplifier increases. At the same time, the positive bias on the AGC circuit increases, thus reducing the over-all negative AGC voltage and increasing the gain of the AGC-controlled stages.

One might ask why a simple form of control, such as is found in volume controls, cannot be used for contrast control. The reason, of course, is that due to the large v-f bandwidth, such a control should be frequency-insensitive up to several megacycles. This is difficult to obtain because of unavoidable parasitic capacitances. Nevertheless, Fig. 20-27 shows how the idea has been put to practical use. A low-value potentiometer has been used, and

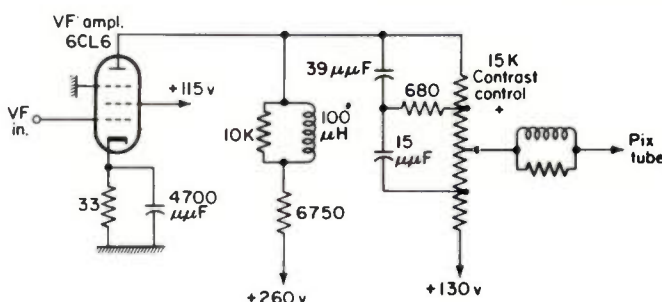


Figure 20-27. Potentiometer type of contrast control.

frequency response is improved by compensating capacitances connected to suitable taps on the potentiometer. The diagram is somewhat simplified for purposes of illustration.

Numerous contrast-control circuits exist. Most of them are based on some method of varying tube gain along the lines of the circuits given in this section.

For contrast control, tubes should ideally belong to the variable-slope variety to provide correct operation and smooth control. This is not always the case, and control circuits using feedback are helpful in this respect because of the linearizing effect they have on the characteristics of the tube.

Problems

1. Compare AGC obtained from the v-f detector and AGC obtained from a separate peak detector.
2. Design an automatic volume control for maximum noise immunity.
3. Design an automatic frequency control for a receiver that has a split i-f sound receiver.
4. Show how the working conditions of a v-f tube can be adjusted so that the tube acts as a limiter.
5. In an AGC circuit using the picture detector d-c voltage, the filter has a resistance of 100 kilohms and a capacitance of 0.1 microfarad. Show how the voltage across the capacitance varies during the vertical sync pulse.

6. In an AGC circuit using the picture detector d-c voltage, compare the circuits using a diode and a triode.

7. Explain in detail the function of the local-distant commutation in the AGC line.

8. Explain in detail the role of the delaying voltage in some AGC circuits.

9. Show that the peak level of the video waveform does not represent the signal level if the d-c component is absent.

10. Since a peak voltage detector is useful for AGC purposes, explain why the picture detector could not belong to that type and fulfill both functions.

11. Compare AGC systems that have long and short time-constant filters.

12. Show how the AGC voltage can be taken off the separator grid.

13. Compare the effect of noise pulses on the peak and average d-c values of the detected signal.

14. Explain in detail how the vertical pulse can upset a peak value AGC system, and draw corresponding diagrams.

15. It is suggested that a gated AGC tube could be keyed by the vertical retrace pulse. Investigate this suggestion.

16. Discuss the possibilities of AGC for a television standard using positive modulation.

17. Design a keyed AGC system in which the tube is gated by its control grid.

18. It is suggested that large-amplitude noise pulses above maximum modulation level could be separated by a clipper, phase inverted, and then added to the v-f signal to produce noise cancellation. Study this idea and propose a circuit.

19. It is proposed to use cathode and grid of the picture tube to eliminate the visible effects of noise pulses on a picture. The cathode would receive the full v-f signal, and the clipped noise pulses would be applied to the grid in such a phase that they would cancel the noise pulses in the cathode signal. Give a suitable circuit and discuss the scheme.

20. Design a combined horizontal and vertical blanking circuit in which blanking pulse mixing is assured by the picture tube.

Chapter 21

COMPLETE RECEIVERS

A complete receiver design is obtained by grouping together all of the sections analyzed in the preceding chapters. In this chapter two commercial designs are given which illustrate the basic principles and show how the various sections are interconnected to obtain a complete receiver. In both cases, a brief circuit analysis points out the main features and links them to the preceding chapters.

21-1. General design

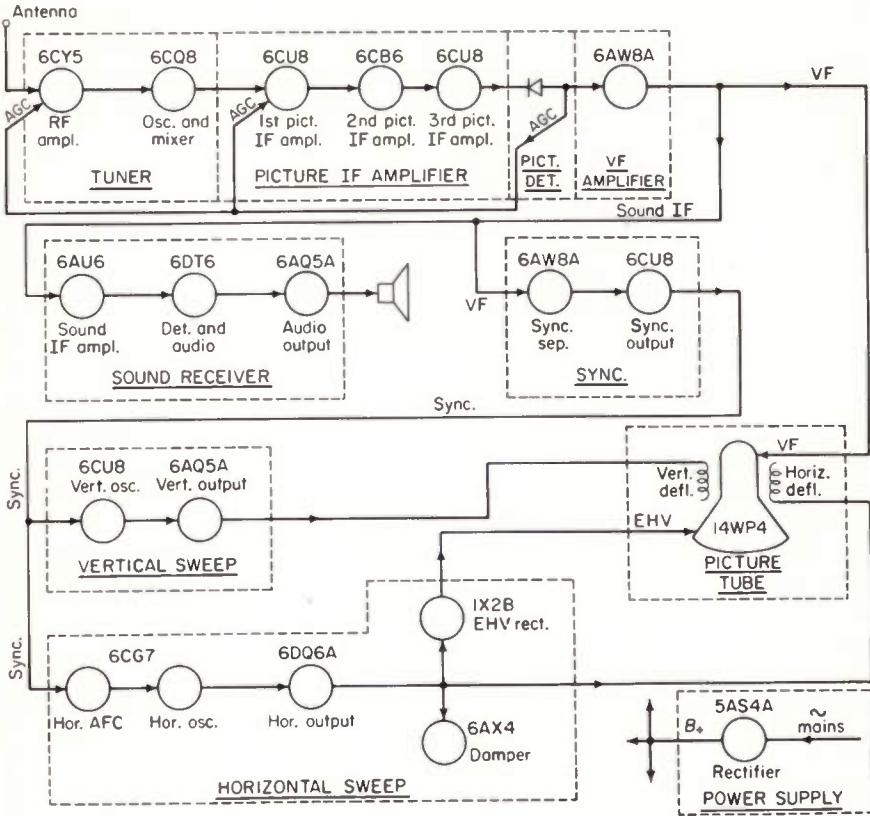
In previous chapters the various sections that make up a complete receiver have been studied separately, and a number of practical stage designs and variations have been given.

To give a better indication of the inter-relations between these sections and the way they are interconnected, two representative and complete commercial receivers will be described. The schematics for both chassis, supplied through the courtesy of the RCA Service Company, are used in a number of different television models. They may receive the VHF band only or both VHF and UHF bands through the addition of a UHF tuner.

The first chassis is a more or less standard design used with 14-inch receivers. The second is a more sophisticated design used with 21-inch receivers.

21-2. 14-inch chassis

Chassis KCS-120 contains 16 or 17 tubes according to whether it covers VHF only or both VHF and UHF. Its block diagram is given in Fig. 21-1. The complete schematic wiring diagram is given in Fig. 21-2, which will be analyzed briefly.



(Courtesy of the RCA Service Company)

Figure 21-1. Block diagram of RCA 14-inch television receiver.

The two tuners have been represented separately. The VHF tuner belongs to the switched-inductance variety and uses a pentode r-f amplifier and a triode-pentode as oscillator-mixer, with capacitive coupling between the oscillator and mixer.

A similar tuner is used for VHF-UHF, with slight modifications and the addition of a continuous-tuning UHF section that uses a triode oscillator and a crystal mixer. Tuning covers 12 VHF channels from 54 mc to 216 mc, and 70 UHF channels from 470 mc to 890 mc. The i-f frequencies are 45.75 mc for picture carrier and 41.25 mc for sound carrier.

The picture receiver section uses three pentodes as i-f amplifiers, with transformer coupling between stages. Besides being applied to the r-f amplifier in the tuner, AGC is also applied to the first i-f amplifying tube, which is connected in series with the second i-f amplifier for the plate voltage supply. The detector is a crystal diode, and the d-c component of detection

is used for AGC purposes. A pentode is used as the v-f amplifier, with a contrast control in its cathode circuit acting simultaneously on the gain through AGC bias. Series-shunt type high-frequency compensation is employed, and since capacitance couplings are used, the d-c component of the v-f signal is not applied to the cathode-modulated picture tube.

The sound receiver is naturally of the intercarrier variety and is fed the 4.5-mc intercarrier beat off the plate of the v-f amplifier. It incorporates an i-f amplifier with the standard grid circuit arrangement and an FM detector which is a variation of the quadrature grid discriminator. Its output voltage is sufficient to drive the audio power output stage directly.

The sync section is fed the v-f voltage taken across the load resistance of the v-f amplifier. The sync separator is a triode grid detector with noise-limiting RC circuits in its grid. It drives a sync amplifier-clipper which is another triode. The full-amplitude separated sync signal is integrated and applied to the vertical sweep section. Approximately one-fifth of the separated sync signal is applied to a triode phase comparator in the horizontal sweep section.

The vertical sweep section contains a triode oscillator and a power output tube, cross-coupled to constitute a multivibrator with standard frequency, amplitude, and linearity controls. An output autotransformer drives the vertical deflection coils.

The horizontal sweep section includes a triode oscillator, AFC-controlled by its grid voltage. The waveform it produces, suitably distorted, drives the power output stage. An output autotransformer provides the driving signal for the horizontal deflection coils, the EHV pulse and the heater power for the EHV rectifier, and the driving voltage for a damping diode. A shunt-type width control is included. The boosted voltage reaches 490 volts. It feeds the horizontal output tube plate, the first anode of the picture tube, the FM detector in the sound receiver, and the vertical oscillator.

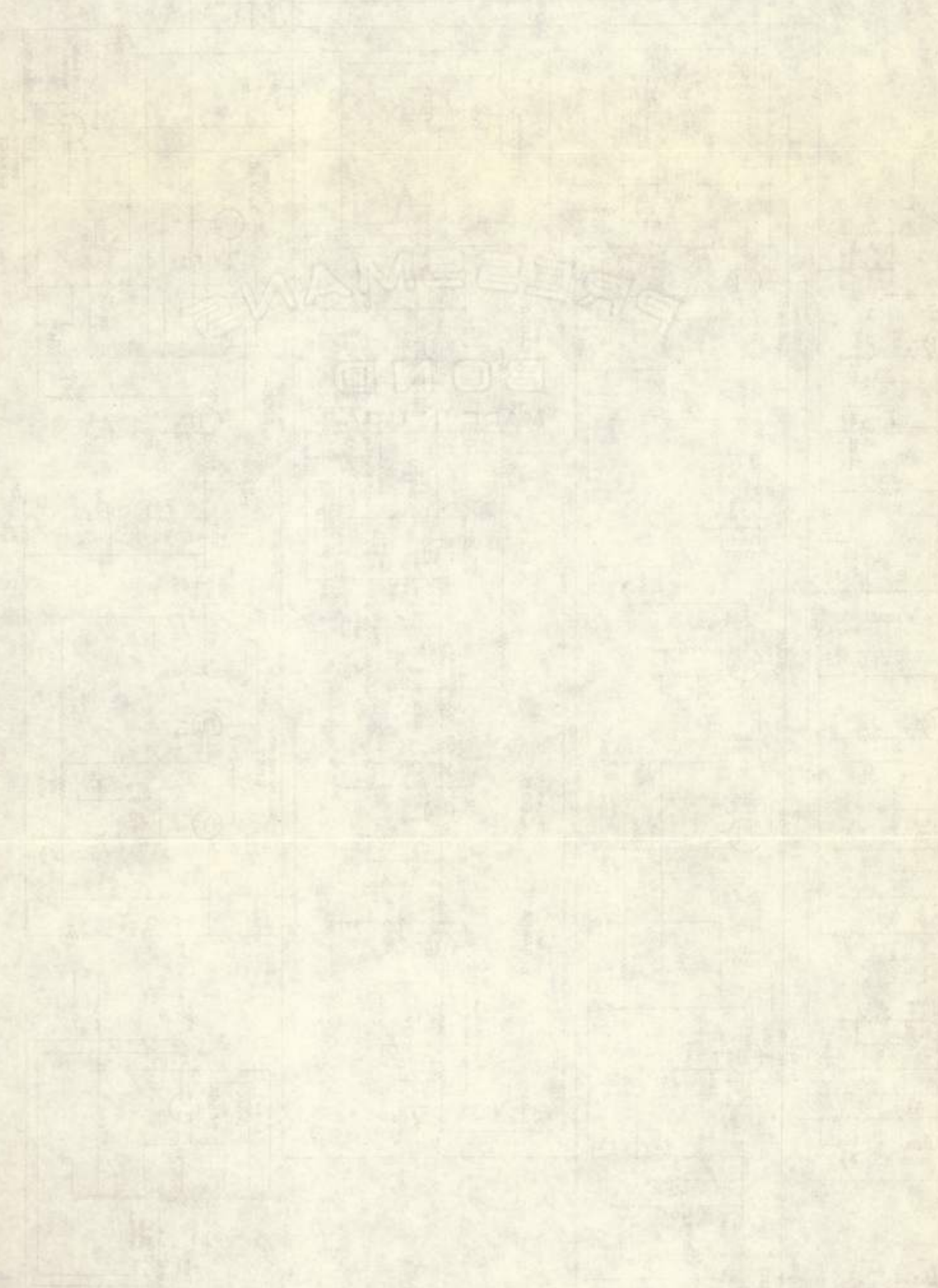
The picture tube is a 14-inch model with fixed electrostatic focusing. It receives 11.5 kv EHV voltage from the horizontal sweep section. Cathode d-c voltage is varied to control brilliance.

The power supply uses a transformer and a full-wave rectifier to provide 245 v plate voltage. Since the rectifier is direct-heated, application of voltage to the other tubes is delayed with a thermal switch in the primary circuit.

A detailed study of the diagram, with identification of all circuit elements and their specific functions, is recommended as an excellent exercise in television receiver analysis.

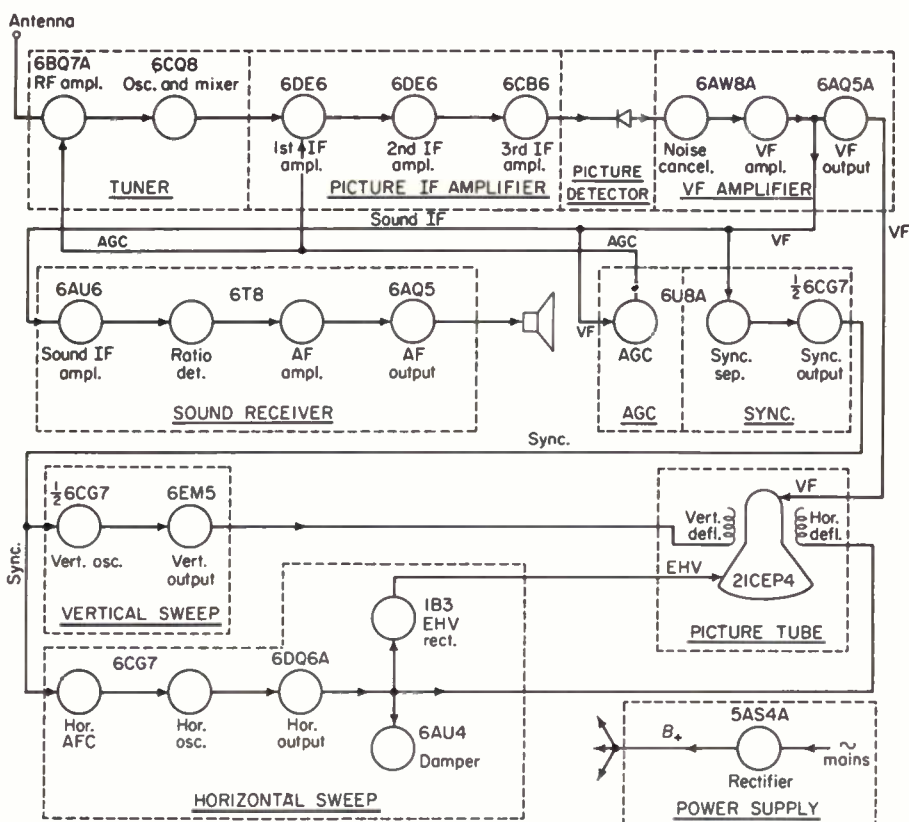
21-3. 21-inch chassis

Chassis KCS-121 is a more elaborate affair including a remote control receiver and transmitter which will not be described here. Figure 21-3 represents its



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(Courtesy of the RCA Service Company.)

Figure 21-3. Block diagram of RCA 21-inch television receiver.

block diagram. The interesting part of the schematic diagram is given in Fig. 21-4.

The VHF tuner is of the switched-inductance type and contains a double-triode cascode r-f amplifier. A triode-pentode is used as oscillator and mixer. The UHF tuner has a triode oscillator and a crystal diode mixer. Tuning is continuous.

Either VHF only or VHF-UHF tuning is found in receivers of this series. The VHF tuners cover the 12 channels from 54 to 216 mc. UHF tuners cover the 70 channels from 470 mc to 890 mc. The i-f picture carrier is on 45.75 mc, and the i-f sound carrier is on 41.25 mc.

The picture receiver is comprised of three pentode i-f amplifiers. The coupling between stages is ensured by bifilar coils, which are equivalent to a single-tuned circuit and are made by winding two wires together to obtain unity coupling. The first two stages are connected in series for d-c plate

voltage supply. AGC voltage is applied to the first i-f stage grid. A crystal diode is used for detection. The v-f amplifier contains two stages: the first is shunted by a noise-cancellation triode whose working level is set by a potentiometer in the grid circuit. Series-type high-frequency compensation is used throughout. The d-c component of the v-f signal is not retained. Contrast control is provided in the cathode of the second v-f amplifier.

The sound intercarrier 4.5-mc signal is picked off the anode of the first v-f tube, and amplified by a sound i-f stage with the usual arrangement of grid circuit. FM detection is performed by a ratio detector. The a-f section consists of a triode voltage amplifier and a power output stage.

The v-f signal taken at the noise-cancellation point in the plate circuit of the first v-f tube is applied to a triode sync separator through the usual grid circuit arrangement. The tube works as a grid detector, and the separated sync signal it provides is further clipped and amplified by a second triode. An integrator circuit feeds the vertical sync pulses to the vertical sweep section. In addition, the sync signal on the plate is fed to a triode AFC control tube in the horizontal sweep section.

The vertical sweep section contains a triode oscillator and a power output tube cross-coupled to work as a multivibrator with usual frequency, amplitude, and linearity controls. An output autotransformer drives the vertical deflection coils. The tap on the autotransformer also feeds a shaping circuit to apply blanking pulses to the cathode of the picture tube.

The horizontal sweep section has a triode sine-wave oscillator with AFC through the grid voltage. The output waveform, suitably distorted, drives a line output tube. An autotransformer couples it to the horizontal coils and feeds the EHV and damping diodes. A shunt width control is provided which at the same time acts as a transformer to provide a blanking pulse with proper polarity. This pulse is applied to the cathode of the picture tube in series with the blanking pulse coming from the vertical sweep section. The EHV voltage reaches 17.5 kv and is fed to the final anode of the picture tube. The boosted voltage measures 515 v. It feeds the plate of the line output tube, the first anode of the picture tube, and the vertical oscillator.

A keyed AGC pentode tube receives on its plate a pulse at line frequency from the line output autotransformer. The v-f signal from the first v-f stage is applied to its grid. The resulting AGC signal, properly filtered, is sent to the i-f amplifying tube in the tuner and to the grid of the first i-f amplifying tube.

The power supply is an a-c type using a transformer and a direct-heated, full-wave rectifier. To delay the application of plate voltage to the tubes until they are warm, a thermal delay switch has been incorporated in the primary circuit.

A detailed analysis of the circuits, with identification of each component and its function, is a very useful exercise that is well worth the effort.

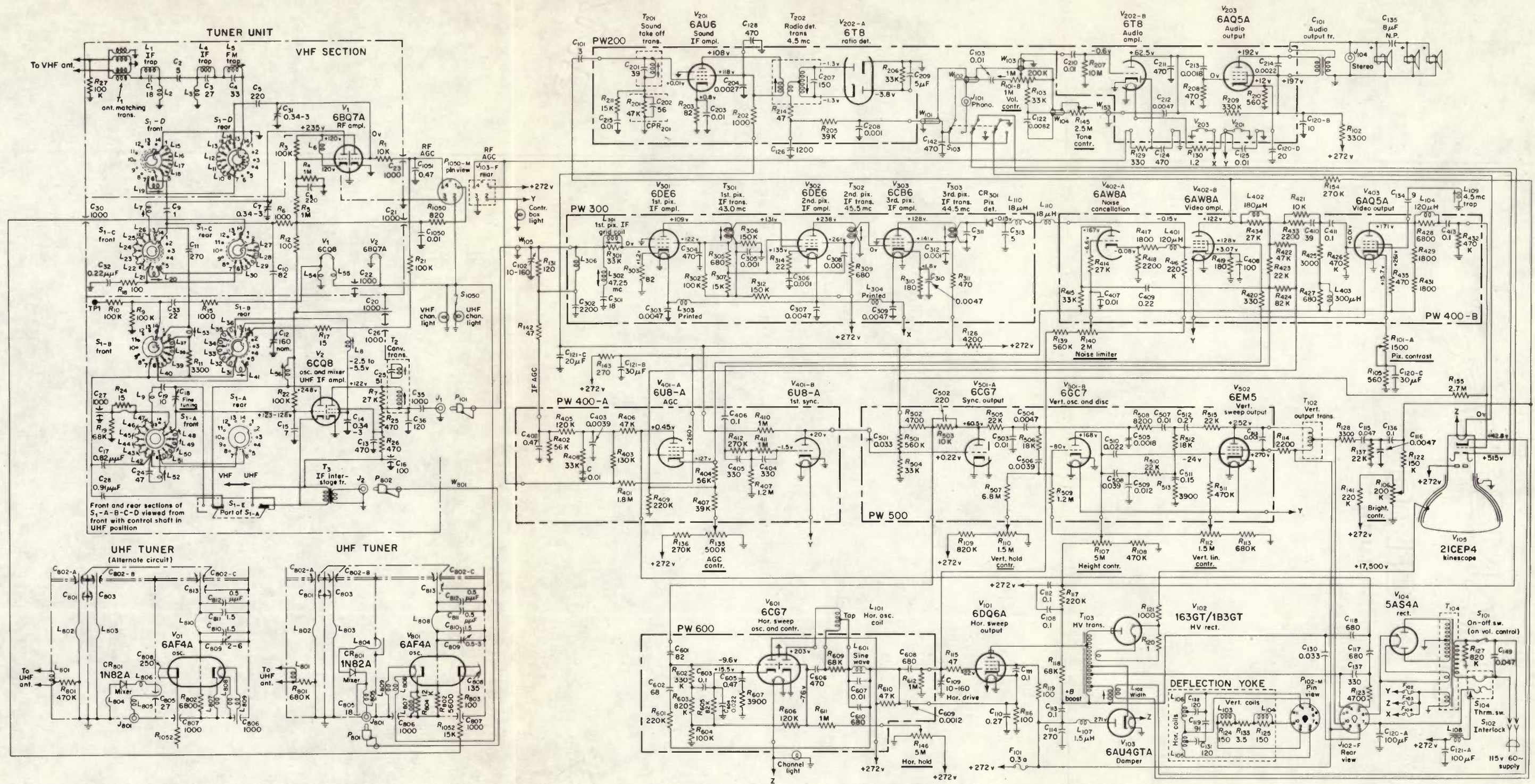
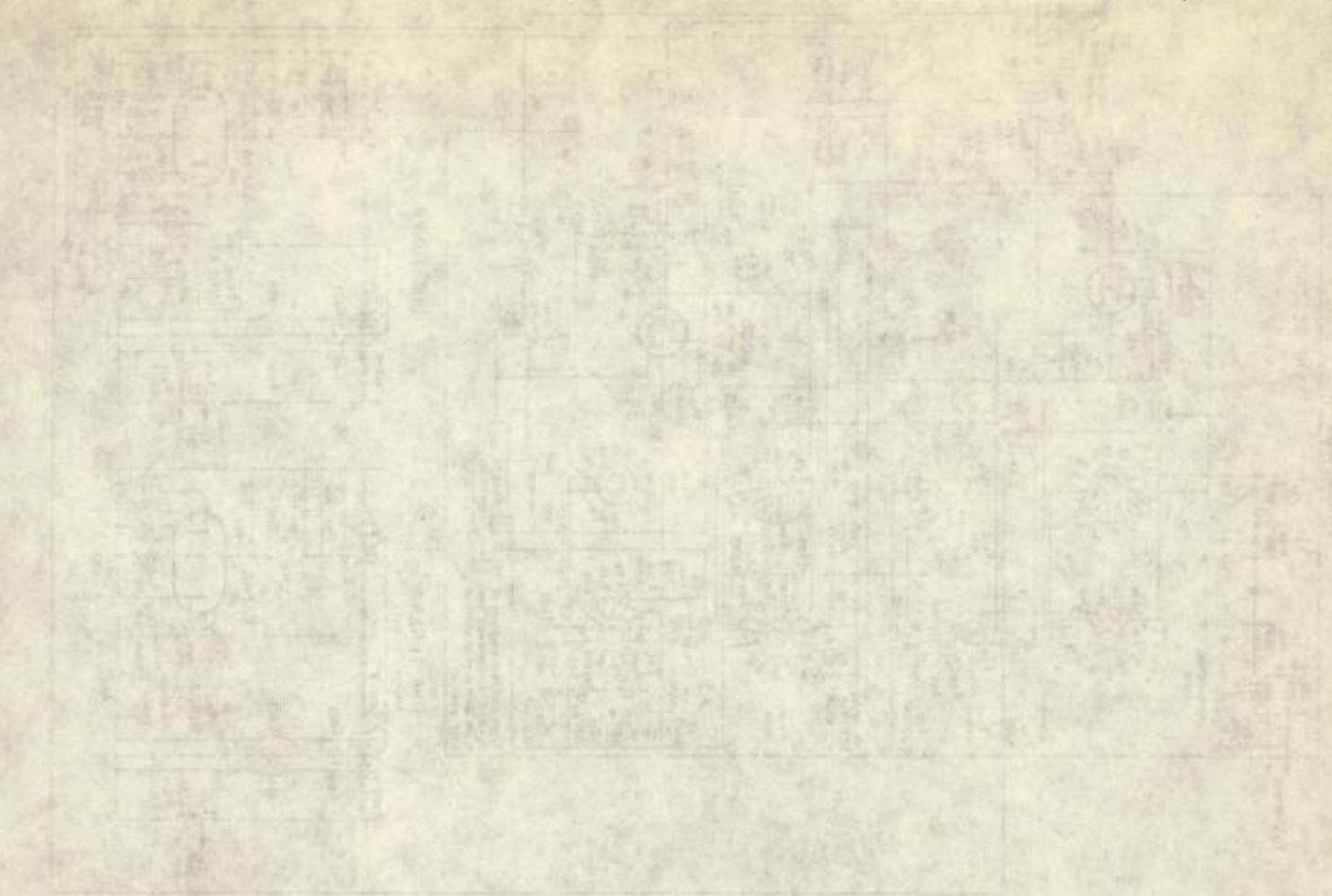


Figure 21-4. Complete theoretical diagram of an RCA 21-inch television receiver. (Courtesy of the RCA Service Company.)



Chapter 22

COLOR TELEVISION

The human eye is an extraordinarily good instrument, but nevertheless it falls short of perfection. Its shortcomings are put to good use in motion pictures and in monochrome television, where persistency of vision is basic to the creation of the illusion of movement.

It is perhaps in color television that the imperfections of human vision are exploited to the utmost. A study of these imperfections has led to the development of color television systems, out of which emerged the NTSC standard. This standard defines a trichromatic system with unbalanced definitions for the three primary colors. It has the advantage of being fully compatible with the older and present monochrome system. In fact, a complete color receiver is derived from the addition of chroma circuits to a monochrome receiver. These circuits do not differ fundamentally from ordinary electronic circuits. The really different part of a color receiver is the picture tube, which is designed to provide a satisfactory color reproduction. Here again, several designs have been developed, out of which emerged the color-points or shadow-mask tube.

22-1. Color vision

A few basic considerations as to the properties of the human eye with respect to color are necessary in order to understand the fundamentals of color television and follow its evolution.

The remanence, memory, or persistence of the eye is a well-known property that is commonly put to good use in motion pictures and monochrome television.

For color vision, the important property is indicated by the experimental

finding that it is possible to reproduce all visible colors by mixing only three fundamental colors. An auxiliary peculiarity is that the eye is much less sensitive to color variations than it is to brilliance or luminosity variations and that its speed of response is lower for color than for luminosity.

In accordance with contemporary technical usage, color will be referred to as *chrominance* and brightness or luminosity as *luminance*.

The eye can resolve or separate black-and-white details if they subtend an angle on the order of one minute of arc. If the same details are in color, they must be much larger in order to be separated. In short, the resolving power of the eye is much better for luminance than it is for chrominance.

Furthermore, since the eye responds more slowly to color variations, color flicker will become noticeable only at frequencies well below that for which black-and-white flicker is objectionable, e.g., around 30 cps.

These two properties of the eye have an important practical result. On the one hand, it is not necessary for color reproduction to have as many details as black-and-white reproduction, as long as luminance variations are correctly reproduced. On the other hand, erratic variations of chrominance will go unnoticed even if they occur at frequencies below field frequency, provided that they are not accompanied by luminance variations.

In an interesting experiment, a surface carrying points of all colors can be clearly seen as such on close inspection. However, if viewing distance is progressively increased, unexpected things happen. First, in order of color disappearance, blues become impossible to separate from grays of the same luminance. Almost at the same time, browns merge with roses and greens with blue-greens. Then reds are confused with grays. Finally, blue-greens disappear and the entire surface looks uniformly gray.

The above order of color disappearance indicates that for small colored areas it is not even necessary to combine three colors to obtain the equivalent visual sensation. Two colors are sufficient, and they are the last colors to disappear in the experiment described. Practical tests have shown that a reddish orange and a bluish green provide the best results.

Lastly, since visual acuity is much smaller for blues than for greens or reds, there is no need to reproduce blue details to as small a degree as green or red details. At normal viewing distance they would not be distinguished anyway.

To sum up, a color television system exploiting at maximum the peculiar properties of human vision would be trichromatic—that is, using three colors—for large details, bichromatic or two-color for medium details, and monochromatic or black and white for small details.

22-2. Chrominance and luminance

In the gamut of electromagnetic wavelengths, the visible spectrum occupies only a small part, roughly between 4000 and 8000 Angstrom units and hence

covering only one octave. The length of the Angstrom unit is equal to 10^{-10} meter.

Inside the limits of this spectrum, color sensation corresponds to wavelength and covers the band from violet to red, going through all the intermediary colors.

Now when a colored surface is illuminated with white light, color sensation is independent of the amount of illumination. However, while retaining the same color, the surface will become more or less brilliant as the illumination is increased or reduced. This points out the difference between brilliance, luminosity, or luminance on the one hand, and color, chromaticity, or chrominance on the other.

22-3. Trichromy

Since every color may be reproduced by a mixture of three fundamental or primary colors, this also applies to pure white. Usually primary colors are a

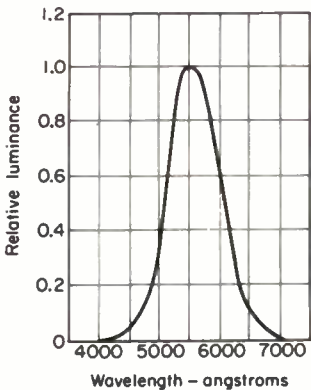


Figure 22-1. Visual sensitivity as a function of wavelength.

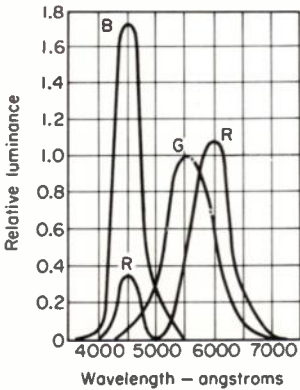


Figure 22-2. Eye response to the three fundamental colors.

red *R*, a blue *B*, and a green *G*. Mixing them in the correct proportions will effectively produce a white.

If the three primary colors *R*, *B*, *G* are kept in a constant ratio but their total intensity is increased or decreased, the resulting color will be a white of increasing or decreasing luminosity. For any other color, the hue is determined by the color components added to white. A pure color containing no white is said to be *saturated*.

The average sensitivity of the human eye as function of the wavelength is given in Fig. 22-1. It exhibits a maximum around 5600 A.

Similarly, the eye's response to the three primaries can be determined experimentally and is given in Fig. 22-2. Notice that the curve for red exhibits

two maxima. If R , B , and G are multiplied simultaneously by a common factor, chrominance does not change—only luminance does. Consequently, the three primary values, R , B , G can be normalized so that

$$R + B + G = 1$$

It then becomes possible to represent all colors with the help of two variables only, say R and G , since $B = 1 - (R + G)$. In addition, the values chosen are such that B , R , and G are always positive—that is, comprised between 0 and 1.

22-4. Maxwell triangle

Any two-variable system may be represented on a plane. With Cartesian rectangular coordinates, the axes can be G and R . Then all possible colors

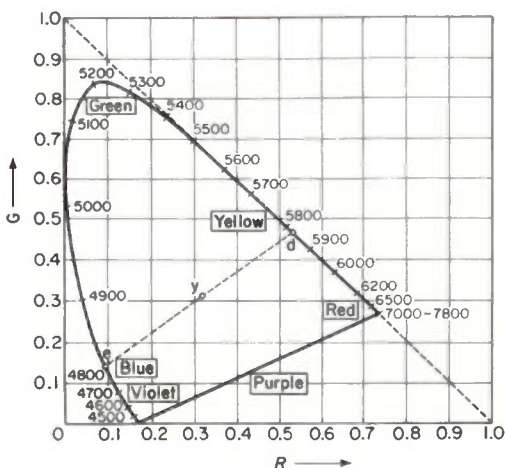


Figure 22-3. Maxwell triangle and visible spectrum limit.

will lie in the triangle limited by the vertical axis for which $R = 0$, the horizontal axis for which $G = 0$, and the diagonal line $R + G = 1$ for which $B = 0$, as shown in Fig. 22-3. This triangle is called the Maxwell triangle.

By reference to the eye's sensitivity curves, all visible spectrum colors can be located in this triangle; they are included between 3800 and 7800 Å. This has been done on the curve. More precisely, pure spectral colors, or monochromatic colors, are distributed on the curve illustrated. The straight line at the bottom linking violets to reds corresponds to purples, which are not spectral colors but are complementary of the spectral colors ranging from 4920 to 5700 Å. Daylight white is situated at the center of the triangle at point Y .

A property of the color triangle is that if any two colors are added, the resulting color lies on the straight line joining the two component colors. When the resulting color is white Y , the two colors are said to be *complementary*. Such would be the case on the figure for blue e at 4800 Å and yellow d at 5820 Å.

The abscissa R and ordinate G of the figure are the chromatic coordinates, frequently called x and y .

22-5. Color reproduction

In the color triangle, white is at the center and the colors become more and more saturated from the center toward the boundary line of the visible spectrum. Now, colors with a high saturation level are rarely found in nature.

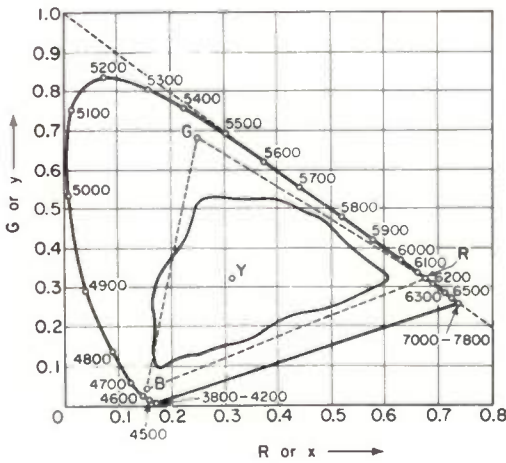


Figure 22-4. Limits of color printing reproduction (continuous curve) and color television reproduction (broken lines).

Moreover, they have low luminance levels and the eye is not very sensitive to variations in their chrominance. This means that a satisfying color reproduction can be obtained without covering the entire range of colors included in the color triangle.

A good example is provided by the range of colors offered by paints, dyes, and printing inks, usually considered as quite satisfactory for everyday use. This range is indicated, as shown in Fig. 22-4, by the irregular solid closed curve inside the color triangle.

Color television, then, would insure satisfactory reproduction if it covered the same area. Actually, it covers a larger area. The three primaries chosen for television are the red, blue, and green indicated limiting the dotted-line

triangle. Except for a small area in the blues, color television can reproduce all colors of the graphic arts, and even more.

By suitable mixing of the television primaries R , B , and G , the white Y can be obtained. The necessary luminances are very different for the three colors. Taking for green a normalized luminance of 1, then red luminance should be 0.5 and blue luminance 0.18, approximately. This can be expressed as

$$Y = G + 0.5R + 0.18B$$

or, better still, normalizing Y to unity:

$$Y = 0.59G + 0.30R + 0.11B$$

22-6. Polar coordinates

Since two variables are necessary to represent the range of colors, they may just as well be plotted as polar coordinates instead of rectangular coordinates.

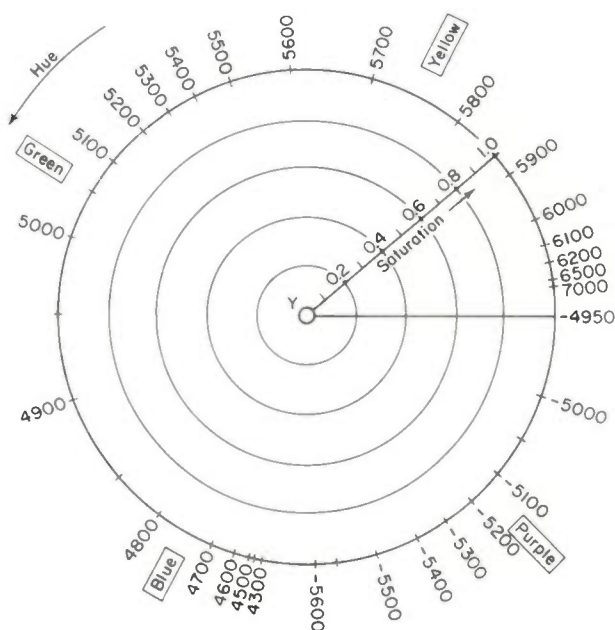
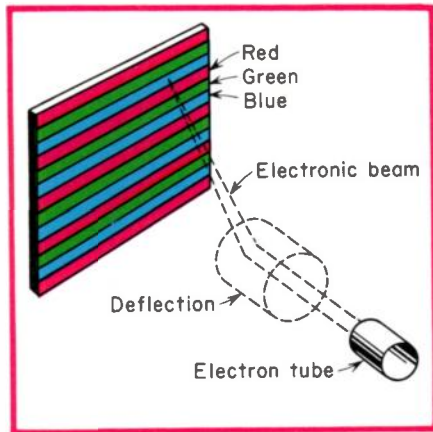


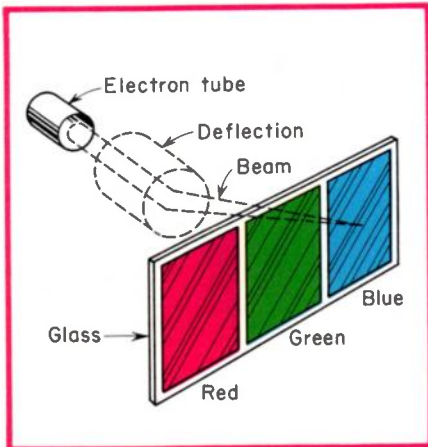
Figure 22-5 Newton color circle diagram.

This has been done and is shown in Fig. 22-5. The center of the diagram is daylight white Y . Pure spectral colors and purples are distributed around a unit radius circle by order of increasing wavelength. For any given color—that is, along any given radius—saturation increases with distance from center.

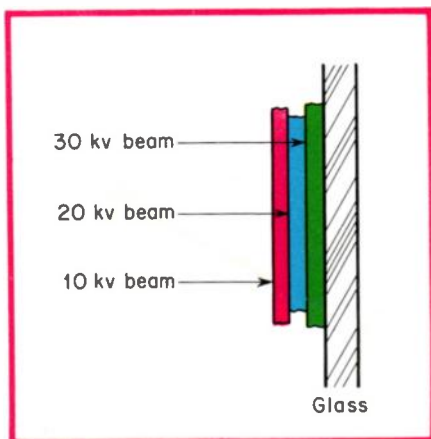
Principle of color lines tube.



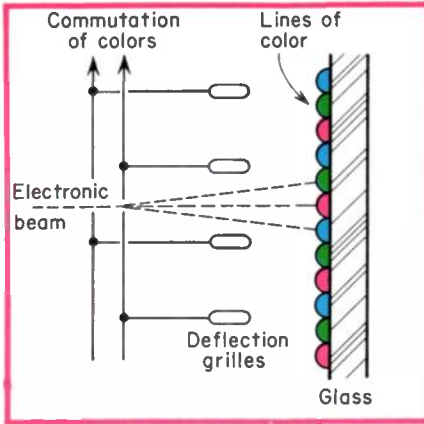
Principle of color areas tube.



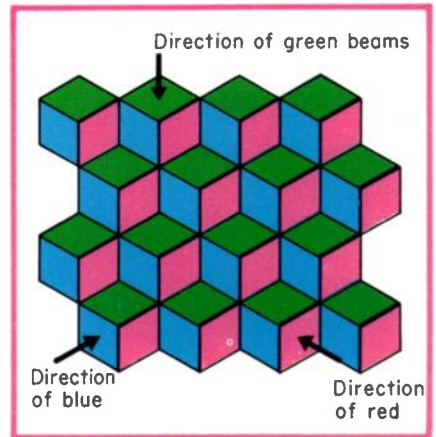
Principle of color layers tube.



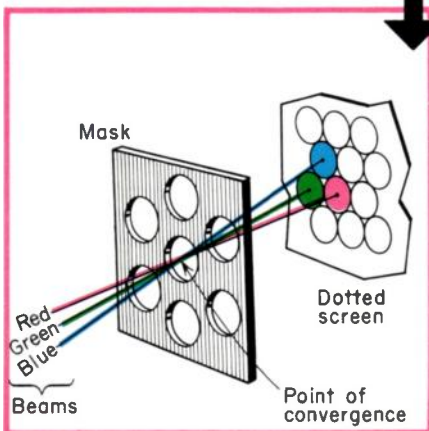
Double grid beam control color tube.

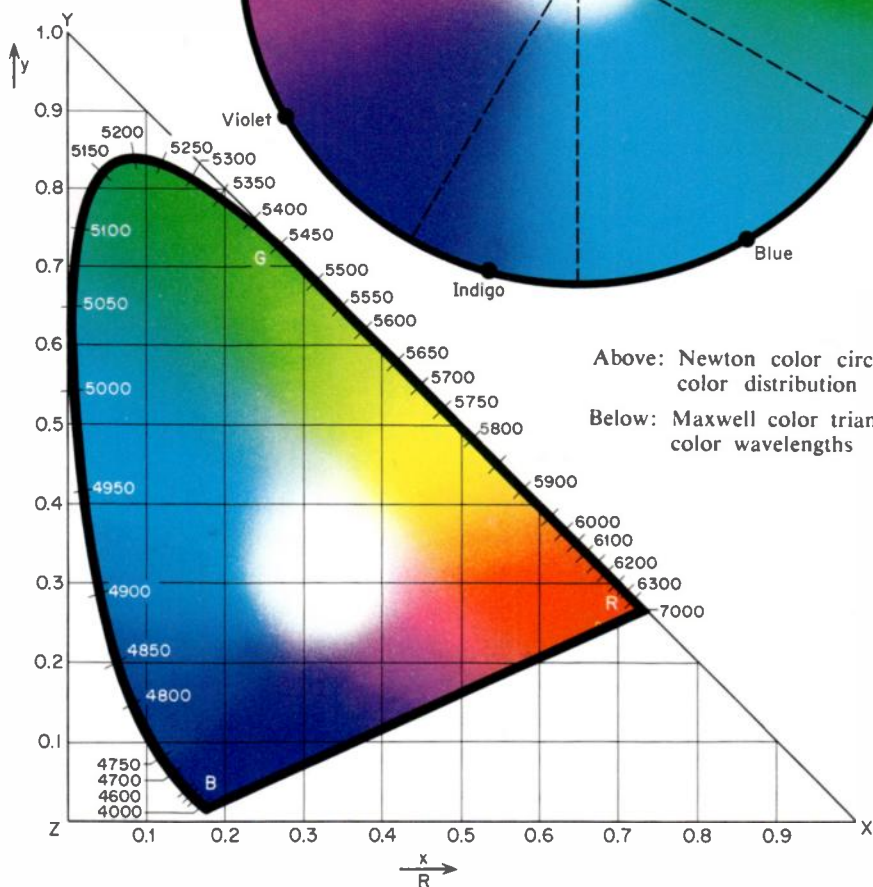
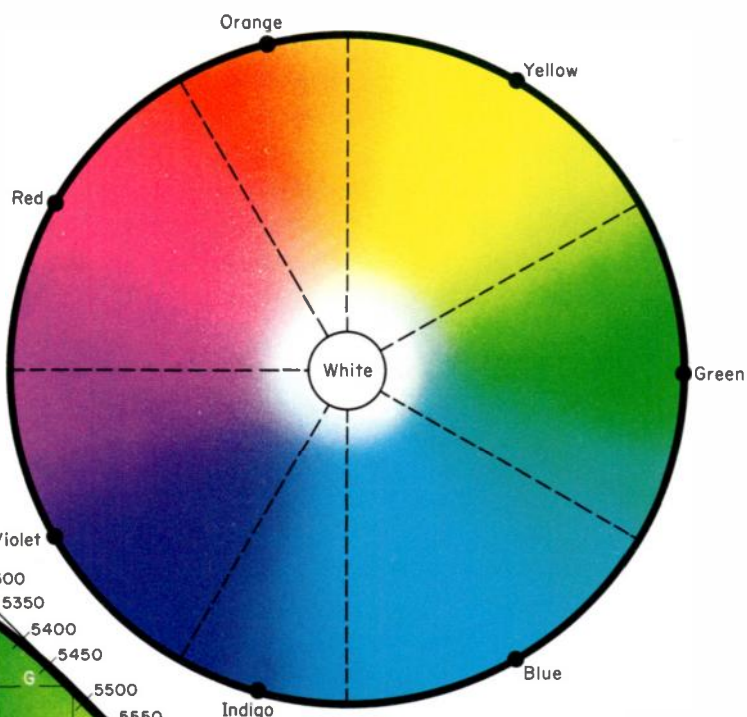


Screen of color pyramids tube.



Principle of perforated mask color points tube.

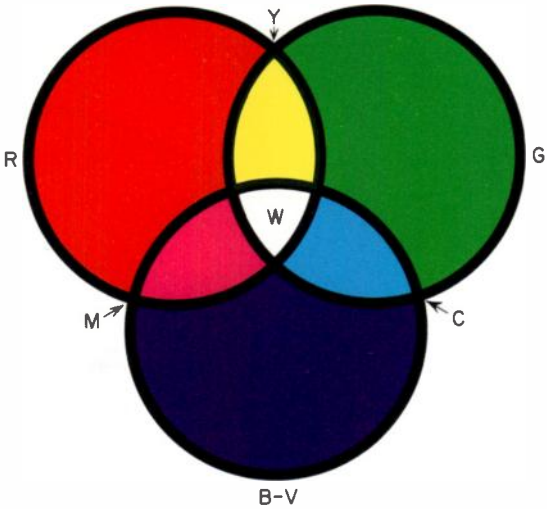




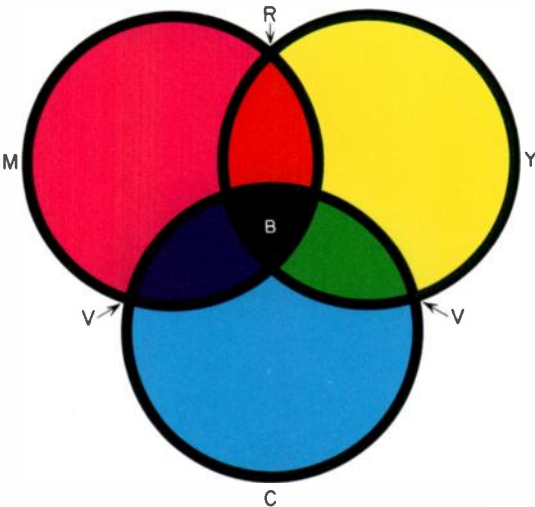
Above: Newton color circle and color distribution

Below: Maxwell color triangle and color wavelengths

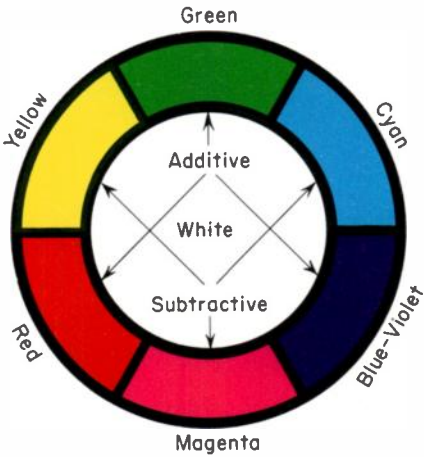
Additive synthesis of three primary colors



Subtractive synthesis of three primary colors



Relations between additive and subtractive primary colors



Also, all colors represented by the opposite radius belong to the complementary color, since mixing both in correct proportions results in white Y . Thus purples have been identified with the wavelength of the complementary color preceded by a minus sign.

The saturation, or chroma, is measured along the radial scale as indicated. Zero saturation is white Y . Unity saturation is a pure spectral color or pure purple.

This polar diagram is called a *Newton circle*.

22-7. Color television

The three fundamental parameters of color vision, then, are luminance, chrominance, and saturation, the latter sometimes called *purity* or *chroma*. All three must be transmitted in a color television system.

Notice that the choice of three primaries corresponds to the minimum number of points delimiting an area on a plane. Actually, several primaries could be used to obtain a better color reproduction by enclosing a larger area of the color triangle, at least with a suitable choice of primaries. Such is the case in quadrichrome printing, where four primaries are used.

A trichrome process, of course, is simpler and provides a generally satisfying color reproduction as explained before; it has been adopted for color television.

The three color pictures can actually be displayed simultaneously on the screen. Such color television systems are said to be *simultaneous*. The three color pictures can also be displayed one after the other in rapid succession, taking advantage of the remanence of the eye to combine them. Such color television systems are said to be *sequential*.

Whatever the system, if the color picture is to have the same definition as a monochrome picture, it is necessary to provide the same details in each primary color picture as would be found in the black-and-white picture. This means that the total number of points in a complete color picture would be three times larger than in a monochrome picture. To transmit three times more points, there are only two solutions: either triple the bandwidth or triple the duration of the transmission.

For example, assume a black-and-white picture containing 120,000 elementary points scanned in $\frac{1}{60}$ of a second. The corresponding v-f bandwidth would be

$$120,000 \times 60/2 = 3,600,000 \text{ cps or } 3.6 \text{ mc}$$

For three primary color pictures each having 120,000 points and transmitted in the same time, the bandwidth would be

$$120,000 \times 3 \times 60/2 = 10,800,000 \text{ cps or } 10.8 \text{ mc}$$

that is, three times larger than previously found for a monochrome picture.

Conversely, the three primary color pictures could also be transmitted in $\frac{1}{20}$ of a second and would occupy

$$120,000 \times 3 \times 20/2 = 3,600,000 \text{ cps or } 3.6 \text{ mc}$$

as for black and white but at the price of a tripled transmission time.

22-8. Typical simultaneous system

The simplest simultaneous system uses a standard television system for each color, illustrated in Fig. 22-6. At the transmitter, a triple optical system with

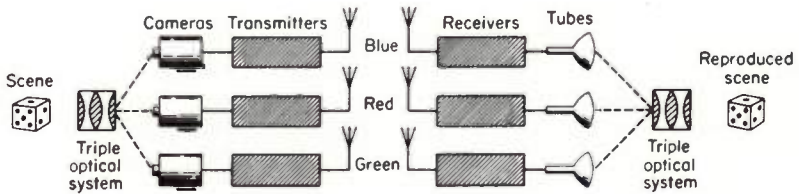


Figure 22-6. Principle of simultaneous system.

appropriate color filters produces three primary color pictures for the three cameras. At the receiver, another triple optical system, with suitable color filters, superposes the three primary pictures to produce a full color picture.

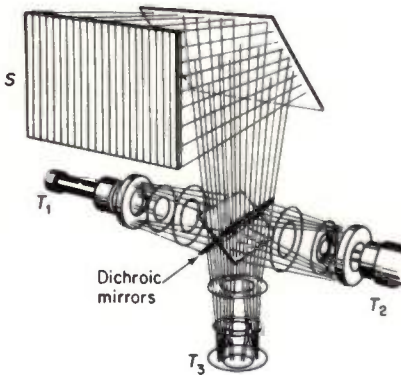


Figure 22-7. Use of dichroic mirrors.

Figure 22-7 shows how dichroic mirrors may be used to superpose on screen S the three primary pictures provided by the three picture tubes T_1 , T_2 , and T_3 . Dichroic mirrors are obtained by coating glass with a very thin metallic film. They have the property of transmitting one of the fundamental colors while reflecting another.

In any case, a simultaneous system as shown is not economical. Evidently, some sections of the system could be common to all three colors, but this would not solve the problem of the triple bandwidth. Furthermore, it may not be too easy to obtain a good, stable superposition of the primary pictures.

22-9. Typical sequential system

A simple sequential color television system uses a rotating disk carrying transparent colored sectors in front of the camera and in front of the receiver. The

two disks rotate in synchronism. All the remainder of the transmission chain is identical with a black-and-white chain, as shown in Fig. 22-8. In the elementary system, each disk carries three sectors—red, blue, and green—each one in use for a picture period, or $1/30$ of a second. There is then sequential transmission of red, blue, and green pictures, and the eye is left with the job of integrating the entire color picture, thanks to visual remanence.

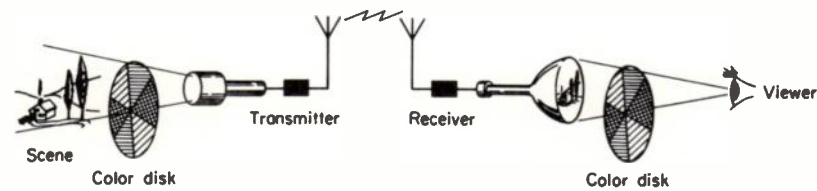


Figure 22-8. Principle of sequential system.

Unfortunately, with the 30 picture-per-second standard, each color occurs only ten times per second, and there is intolerable color flicker and color breakup on fast-changing scenes. An immediate improvement is to switch colors at every field, which doubles the color switching rate. Nevertheless, the over-all result is not satisfying, each color occurring 20 times per second.

Acceptable results are obtained with a color repetition frequency of 48 per second, but then the color standard differs from the monochrome standard.

22-10. Optical systems

Another idea is to share the available picture surface between the three primary pictures. The principle is illustrated in Fig. 22-9. The camera optical system

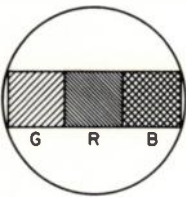


Figure 22-9. Shared area principle.

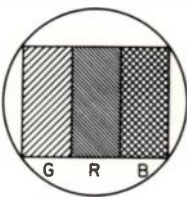


Figure 22-10. Anamorphosis in vertical direction.

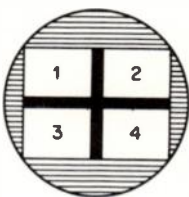


Figure 22-11. Shared area four-color system.

projects a triple picture on the mosaic of the pickup tube through appropriate filters. These three pictures combined are scanned and transmitted as a single black-and-white picture and reproduced as such on a monochrome receiver screen. Then another triple optical system with suitable filters recombines them to provide a full color picture.

The poor utilization ratio of the available surface can be improved by anamorphosis of the picture in the vertical direction, as shown in Fig. 22-10. Also, four fundamental colors can be used with better color reproduction, as shown in Fig. 22-11.

All these systems have the same disadvantage—namely, the difficulty of obtaining good, stable superposition of the primary pictures.

22-11. Compatibility

Another important problem is that of compatibility. A color television system is considered compatible only if its picture can be received as a normal black-and-white picture on an unmodified standard monochrome receiver.

Evidently, this is of paramount importance from the economic point of view of introducing a color system. Public acceptance is insured only if the new color programs can be received as monochrome on the old receivers and if the system does not make the millions of receivers in service obsolete.

Conclusions are easily drawn as to compatibility from the rapid examination of color television systems just performed.

Simultaneous systems are fully compatible if each color transmitter uses the monochrome standard. Tuning the receiver to any of the three channels will provide monochrome reception. The green channel is commonly used because the gray scale it provides is the nearest to the gray scale of a black-and-white transmission.

Sequential systems as described would be compatible if the same vertical frequency could be used; however, this is not the case.

Optical systems may be compatible with certain restrictions. Theoretically they are attractive but are difficult to put to practical use.

Elementary sequential processes, besides being noncompatible, use a mechanical arrangement and as such are doomed as soon as a satisfactory electronic system is developed. They are simple and economical and may be used where noncompatibility is no problem, as, for example, in closed-circuit television.

Simultaneous processes are compatible. In the elementary form shown, superposition difficulties are to be expected. The basic principle can be refined to obtain reductions in bandwidth and cost.

22-12. Sequential systems

All sequential systems rely on vision remanence to integrate three primary color stimuli transmitted sequentially. The three fundamental colors must then be sampled at a rate sufficient for eye remanence to do the rest. The difference in color sampling rate is the distinguishing feature of the three systems—field sequential, line sequential, and point sequential. All these

systems can usefully combine the interlace of monochrome standard with color switching.

In the field sequential system previously touched upon, color switching takes place at field frequency giving the following color sequence:

- picture 1 { even field 1: red
 { odd field 1: blue
- picture 2 { even field 2: green
 { odd field 2: red
- picture 3 { even field 3: blue
 { odd field 3: green

etc.
To avoid flicker, vertical sweep takes place at 72 pictures per second. However, to fit the 6-mc standard bandwidth channel, the horizontal sweep frequency must be reduced to 29,160 cps, which provides 405 lines per picture with attendant reduction in definition.

In the line sequential system, color switching takes place at line rate, giving for lines the same color sequence as for fields in the preceding system. Entire scanning lines appear successively in one of the primary colors, with reversal at each field period.

22-13. Point sequential system

In the point sequential system, color sampling is done at point frequency; Fig. 22-12 illustrates the principle. A triple camera provides three video signals—red, blue, and green—with the help of a triple optical system and

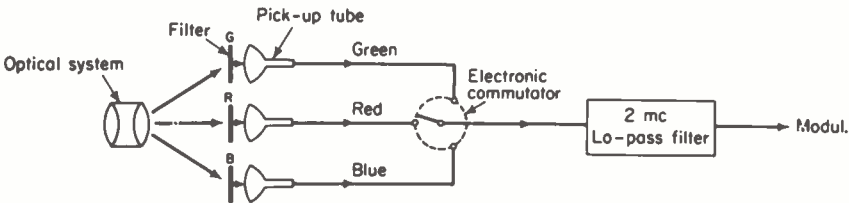


Figure 22-12. Principle of dot sequential system.

suitable filters. Since the available bandwidth of the monochrome standard is 4 mc, the three signals share the available time and are successively switched to the output at the rate of 3,800,000 times per second with an electronic switch. Each signal is on for a short time, and gives a short positive pulse whose amplitude exactly represents the color component amplitude at this precise instant—that is, at this precise point of the picture. The triple series of pulses goes through a 2-mc low-pass filter from which it emerges as a continuous signal containing the color information, shown in Fig. 22-13. It seems

difficult to recover the original color values from the continuous signal, but it will be noted that the effect of the filter on each series of pulses is to transform it into a continuous periodic signal. The sum of the three continuous periodic signals is the complex waveform at the output of the filter.

The interesting point to notice is that when one of the individual continuous signals goes through a maximum, at a point corresponding to the original

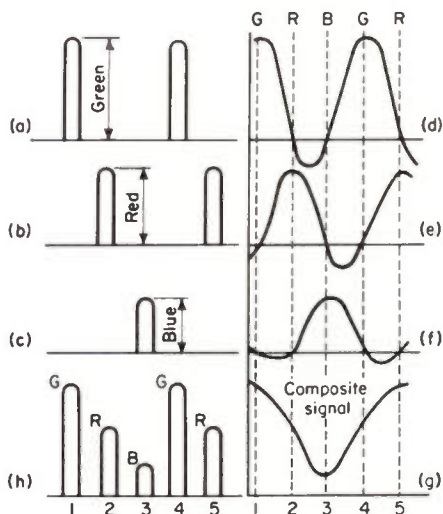


Figure 22-13. Combination of color pulses.

pulse, both the other continuous signals go through zero. This shows that the total signal amplitude at this point is that of the first individual signal only, hence of the first original color pulse.

Sampling the amplitude value of the total signal at points 1, 2, 3, etc., then, gives the amplitudes of the green, red, and blue color signals. This sampling can be done with an electronic switch synchronized with the switch used at the transmitter.

The advantage of the composite signal is that it occupies the same bandwidth as the sampling signal, that is, 3.8 mc, and carries all the information contained in the three color channels. Each color is sampled in such a way that color points are interlaced along consecutive lines and are at the same time submitted to the usual line interlace pattern. This avoids the appearance of stationary parasitic colored patterns, but means that there are four fields for a complete color picture: two, point interlace, multiplied by two, line interlace. There are then 15 color pictures per second, with a vertical frequency of 60 cps.

The maximum detail which can be transmitted corresponds to

$$3.8:2 = 1.9 \text{ mc}$$

22-14. Mixed high frequencies

This last result shows that definition is rather limited. However, there is a way around this difficulty, which takes advantage of the fact that the eye is practically insensitive to color in small areas, and which gives the details of the picture in black and white only.

There are, then, in fact two pictures superposed on the screen. One is a black-and-white picture with full definition and the other a color picture with reduced definition.

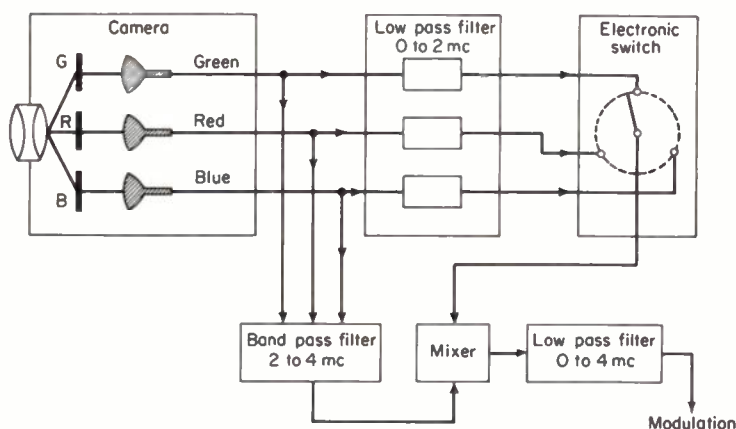


Figure 22-14. Block diagram of color transmitter.

Stated in another way, up to some critical frequency, the video signal is split up into three color components. Above the critical frequency there is no separation of color components and the details are common to all three color channels.

Assuming a 4-mc video bandwidth and a 2-mc critical frequency, the transmitter system may be represented by the block diagram of Fig. 22-14. As before, the three color signals are produced from 0 to 2 mc and sequentially sampled. There is now, however, an extra channel from 2 to 4 mc, bypassing the sampler, and producing a signal which is added to the composite color signal. This system is called *bypassed mixed high frequencies*.

The corresponding receiver block diagram is given in Fig. 22-15.

What is the lower limit of critical or crossing frequency? A frequency as low as 0.1 mc provides usable results. The optimum value seems to lie between 0.5 and 1.5 mc for the standard 6-mc channel.

A high crossing frequency is interesting because it reduces the apparent graininess of the color picture. This grain structure is visible not only in the colored parts but also in the gray or white areas, where it is most objectionable since white is produced by adding the three primary colors.

The last sentence gives a clue to this problem. Since it is the addition of the three color signals which produces graininess, the idea is to proceed in the opposite direction—that is, to subtract a suitable color component from the black-and-white picture.

The resulting mixed-highs modified system is then as follows: A black-and-white signal from 0 to 4 mc is continuously transmitted and applied at the

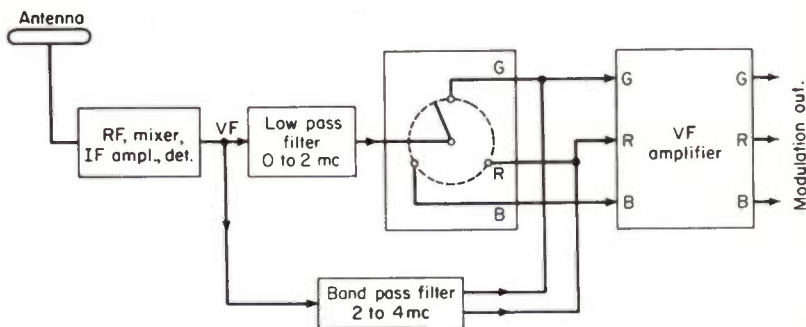


Figure 22-15. Block diagram of color receiver.

receiver to the three color-modulating electrodes. It thus provides gray or white areas without any grain structure.

To obtain colors, a subtractive color signal is transmitted over a reduced bandwidth, say 0 to 2 mc. At the receiver this color signal subtracts the correct components from the white, thus producing the required color.

22-15. Spectrum interlace

Frequency spectrum analysis of a television signal shows that its frequency components appear in clusters, centered around the harmonic of line frequency, shown in Fig. 22-16. Forty-six per cent of the frequency spectrum is never used and the idea of using the available space to accommodate color information arises.

Notice that the clusters all occur around even harmonics of the half-line frequency. Since the half-line frequency is available at the transmitter to produce line interlace, it can be frequency-multiplied by an odd number to provide a new carrier frequency in the v-f band. It will fall, for example, at F_0 , shown in Fig. 22-17 for a multiplying factor of 121—that is, at an empty space in the spectrum.

Moreover, if F_0 is itself modulated with a v-f signal, all resulting frequency components will be spaced from F_0 by an integral multiple of the line frequency and will fall halfway in between the original v-f clusters, as shown in Fig. 22-18. In this figure, the original frequency components of Fig. 22-16 are shown in

continuous lines, whereas the new components due to the modulation of sub-carrier F_0 are shown in broken lines. It can be seen that the two spectra are

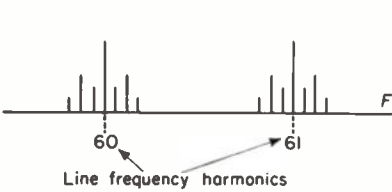


Figure 22-16. Frequency spectrum of a v-f signal.

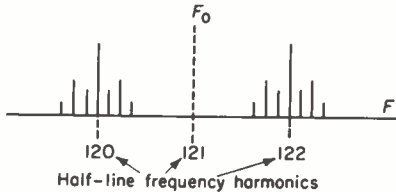


Figure 22-17. Locating the subcarrier between two clusters of frequency components.

distinct and sandwiched one between the other in what is called *frequency interlace*.

An important point is that the polarity of the odd components reverses at each field. With visual remanence playing its usual role, the total visual effect is zero. In other words, the additional modulation does not interfere with the original modulation.

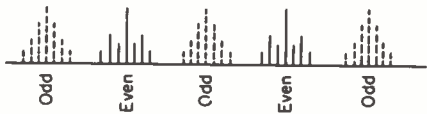


Figure 22-18. The two spectra interlace.

The system can be put to good use in color transmission as illustrated in Fig. 22-19. The monochrome signal is actually the green signal. Red and blue signals are used to modulate two subcarriers inside the v-f band, each of these two modulations being of the single-sideband variety.

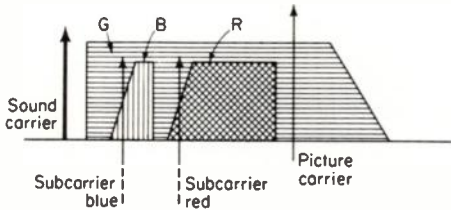


Figure 22-19. Subcarriers disposition in the channel.

Because of the peculiarities of color vision, a bandwidth of 1.5 mc is suitable for red, whereas 0.5 mc is sufficient for blue.

A number of variations of the frequency distribution is possible. Figure 22-19 represents only one of many possible arrangements.

Since only the green component exists at high frequencies, details would appear in green. This is easily avoided by selecting all frequency components above 1.5 mc with the help of a high-pass filter, and applying them to the three

modulating electrodes, green, red, and blue. This is a mixed-highs modulation of a sort.

22-16. NTSC color system

The official color standard currently used is the NTSC color system. It uses the three ideas previously put forward—point sequential scanning, mixed highs, and frequency interlace. However, it adds a new concept, that of the phase-color relation, and exploits to the maximum the peculiarities of color vision. In this system, the white Y and primary colors red, green, and blue are determined on the diagram shown in Fig. 22-4 as

$$Y: x = 0.310, y = 0.316$$

$$R: x = 0.681, y = 0.319$$

$$G: x = 0.197, y = 0.731$$

$$B: x = 0.156, y = 0.032$$

With the three primary colors, red, blue and green, a white Y results from a mixture such that

$$Y = 0.59G + 0.30R + 0.11B$$

As a result, there is no need to transmit the four quantities Y , G , R , and B , since three of them will determine the fourth. For example, Y , B , and R will be transmitted, and G will be easily found at the receiver.

Also, the luminance Y being transmitted, the idea of subtracting a component to obtain color is readily applied by transmitting, for example, $(G - Y)$, $(R - Y)$, and $(B - Y)$.

Then

$$(G - Y) + Y = G$$

will provide the green signal by simple addition of the $(G - Y)$ component to Y , and similarly for red and blue components.

Combining the two preceding remarks, it is sufficient to transmit the luminance Y and the two signals $(B - Y)$ and $(R - Y)$, which are called the *color difference signals*.

22-17. Phase and color

Since the luminance signal Y is transmitted in the usual way over a 4-mc bandwidth, it can be used by an ordinary monochrome receiver and the system is compatible.

Nonetheless, for color reception both $(B - Y)$ and $(R - Y)$ color difference signals must be transmitted, and this seems to require two subcarriers as explained before for frequency interlace.

In the NTSC system, a single subcarrier is used, and color is linked to the phase of the subcarrier, as shown in Fig. 22-20, where phases are measured from the reference line indicated. For example, $(B - Y)$ has a phase of 180 degrees, which means that it is diametrically opposed to the reference. Similarly, $(R - Y)$ lags 90 degrees behind reference.

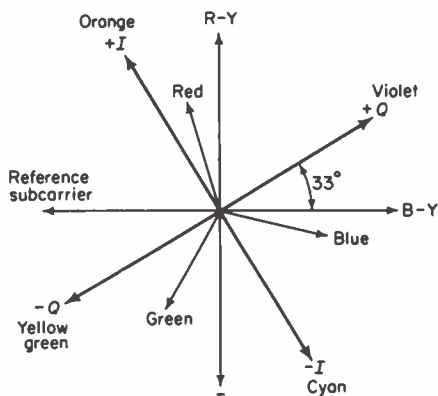


Figure 22-20. Phase relations between colors.

In the same way, all colors can be identified through their phase angle with respect to reference. Red, blue, and green have been indicated. In this way, the *phase* of the subcarrier identifies the *color*, whereas its *amplitude* identifies *saturation*.

22-18. I-Q system

As we have said before, satisfactory color reproduction results if some colors are transmitted over a 1.5-mc bandwidth and other colors over a 0.5-mc bandwidth only. The required bandwidth is essentially determined by the color sensitivity of the eye. Maximum and minimum bandwidths do not occur for primary colors. A maximum bandwidth of 1.5 mc is needed for colors situated on the axis orange-cyan, the *cyan* or *cyanure* being a blue-green. A minimum bandwidth of 0.5 mc is sufficient for colors situated on the axis violet-yellow green, perpendicular to the preceding axis.

Since the origin of phases is arbitrary, for convenience it has been taken on the maximum bandwidth axis, which has been called the *in-phase* or *I-axis*. The minimum bandwidth axis, being perpendicular to the *I-axis*, has been called the *quadrature* or *Q-axis*.

Both *I* and *Q* axes have been indicated as shown on Fig. 22-20 to represent their relative position within the previous system. They lead 33 degrees with respect to the $(R - Y)$ and $(B - Y)$ axes.

In the I - Q system then, the color difference signals I and Q do not correspond to primary colors, but can be used just as well to modulate the subcarrier, with a phase difference of 90 degrees. The result is a single color modulation whose amplitude and phase vary simultaneously as functions of saturation and color.

The advantage of the I - Q system over the $(B - Y)$ - $(R - Y)$ system is that it makes better use of the peculiarities of color vision.

22-19. Color subcarrier

To obtain frequency interlace, the color subcarrier has a frequency which is exactly 455 times the half-line frequency. The exact value is 3.579545 mc,

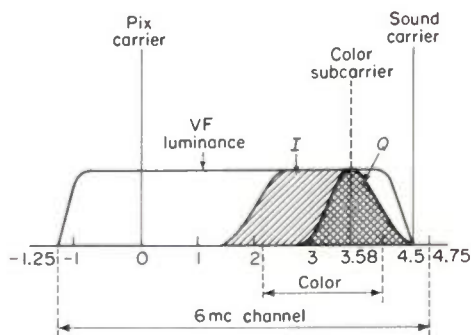


Figure 22-21. Single color subcarrier.

generally written as 3.58 mc for simplicity. On one hand, the subcarrier is modulated by the Q signal with a bandwidth of 0.5 mc, both sidebands being retained. On the other hand, it is simultaneously modulated by the I signal with a bandwidth of 1.5 mc. The full lower sideband is retained, and the upper sideband is limited to 0.5 mc. The resulting frequency spectrum is illustrated in Fig. 22-21.

Notice that the color subcarrier is near the sound carrier and that troublesome interference could result from undesired beats. This is avoided by using the frequency interlace principle once again. The frequency interval between color subcarrier and sound carrier is made equal to an odd multiple of the half-line frequency. In the NTSC standard, the frequency interval between picture and sound carrier is 572 times the half-line frequency. Since the color subcarrier is 455 times the half-line frequency away from the picture carrier, it is separated from the sound carrier by

$$572 - 455 = 117$$

times the half-line frequency—that is, 920 kc approximately.

This takes care of color modulation but does not eliminate the 920-kc beat between sound carrier and color subcarrier. The solution to this difficulty is simple and radical: the color subcarrier is suppressed, only the sidebands being transmitted, with the upper sideband reduced to 0.5 mc. The color then uses a vestigial-sideband, suppressed-carrier system.

Another reason for suppressing the carrier is that it results in a complete disappearance of the color signal in the case of a black-and-white transmission. Effectively, for a monochrome signal there is no I and no Q signal, and since there is no carrier either, the total color signal disappears.

22-20. Color synchronization

As in all suppressed-carrier systems, detection can be accomplished at the receiver only if the color subcarrier is reinserted. The reinserted subcarrier must have a frequency rigorously equal to the frequency of the subcarrier used at the transmitter. This is accomplished by using a 3.58-mc oscillator in the receiver and keeping it in synchronism with the transmitter subcarrier. The synchronizing signal is a short burst of 8 cycles minimum of the subcarrier, inserted at the transmitter in the back porch at black level following the horizontal sync pulse as shown in Fig. 22-22. At the receiver, this burst of oscillation is separated and applied to the local 3.58-mc oscillator to keep it in synchronism.

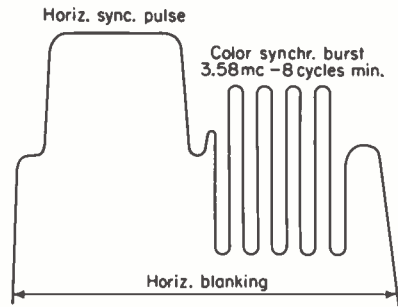


Figure 22-22. Synchronization color burst.

22-21. Color transmitter

A simplified block diagram of a color transmitter is given in Fig. 22-23. A triple camera provides the three primary color signals, red, blue, and green. They are mixed in suitable proportion in a mixer or matrix circuit, simply made of appropriate resistance dividers. Three matrices are used to obtain the Y , I , and Q signals. The Y signal uses the full bandwidth, the I signal is limited by a low-pass filter to 1.5 mc, and the Q signal is limited to 0.5 mc.

A local oscillator provides the 3.58-mc color subcarrier, from which color sync bursts are derived and inserted in the horizontal blanking period.

The 3.58-mc subcarrier feeds a modulator receiving the I signal. The modulator belongs to the symmetrical variety and automatically suppresses the carrier. A sideband filter is included to limit the upper I sideband to 0.5 mc.

The 3.58-mc subcarrier also feeds a Q modulator through a 90-degree phase-shifting network to obtain the desired phase relationship between I and Q axes. The Q signal modulates the subcarrier, and again the modulator is of the symmetrical type which suppresses the carrier. Both 0.5-mc sidebands are retained.

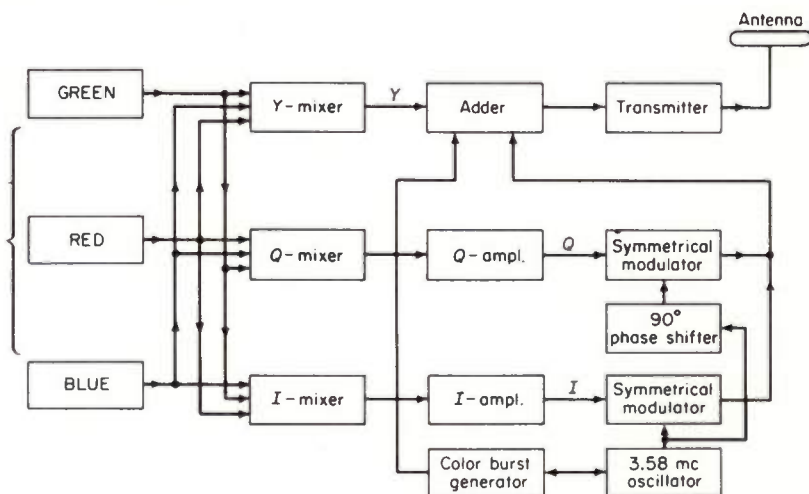


Figure 22-23. Block diagram of a color receiver.

The outputs of the I and Q modulators are combined with the Y luminance signal and the color sync bursts in an adder which provides the full-modulation waveform to the transmitter.

The complete modulation signal then contains:

1. a 3.58-mc color synchronization burst of 8 cycles minimum, occurring during horizontal blanking;
2. the full 4.5-mc Y luminance signal;
3. the I chrominance signal, transmitted vestigial-sideband fashion on a suppressed subcarrier of 3.58 mc; the full lower sideband of 1.5 mc is retained, and the upper sideband is truncated at 0.5 mc;
4. the Q chrominance signal, transmitted with both 0.5-mc sidebands on the same suppressed subcarrier of 3.58 mc.

22-22. I and Q signals

Before studying the receiver, it is necessary to see how the red, blue, and green chrominance signals will be derived from the I and Q signals.

Although it is relatively easy to obtain the relation linking the Y luminance signal to the three primary chrominance signals, that is,

$$Y = 0.59G + 0.30R + 0.11B$$

it is more difficult to find the optimum relations for the 1.5-mc I signal and the 0.5-mc Q signal. In terms of primary colors G , R , and B , these signals are

$$I = -0.28G + 0.60R - 0.32B$$

$$Q = -0.52G + 0.21R + 0.31B$$

In terms of the color difference signals ($G - Y$), ($R - Y$), and ($B - Y$), the calculations are simple enough. From the Y signal value above,

$$R - Y = R - (0.59G + 0.30R + 0.11B)$$

$$\text{or} \quad R - Y = 0.70R - 0.59G - 0.11B$$

Similarly,

$$B - Y = 0.89B - 0.59G - 0.30R$$

$$\text{and} \quad G - Y = 0.41G - 0.30R - 0.11B$$

The last three equations are easily solved in terms of the first members, yielding

$$G - Y = -0.51(R - Y) - 0.19(B - Y)$$

which shows how the ($G - Y$) signal can be obtained from the ($R - Y$) and ($B - Y$) signals.

Combining these results with the expressions for I and Q provides

$$I = 0.75(R - Y) - 0.37(B - Y)$$

$$\text{and} \quad Q = 0.48(R - Y) + 0.41(B - Y)$$

which indicates how the I and Q signals are obtained at the transmitter starting with the color difference signals ($R - Y$) and ($B - Y$).

Conversely, at the receiver the color difference signals are obtained from the I and Q signals by using the following expressions, which are easily obtained by manipulating the two preceding equations:

$$R - Y = 0.62Q + 0.96I$$

$$\text{and} \quad B - Y = 1.11I + 1.72Q$$

from which

$$G - Y = -0.64Q - 0.28I$$

Adding Y to each color difference signal will provide R , G , and B .

22-23. Color receiver

Figure 22-24 shows a simplified block diagram of a color television receiver. Up to the picture detector, the circuitry is identical with that of a monochrome receiver. After detection, two channels are provided, one for the Y luminance signal and the other for chrominance information.

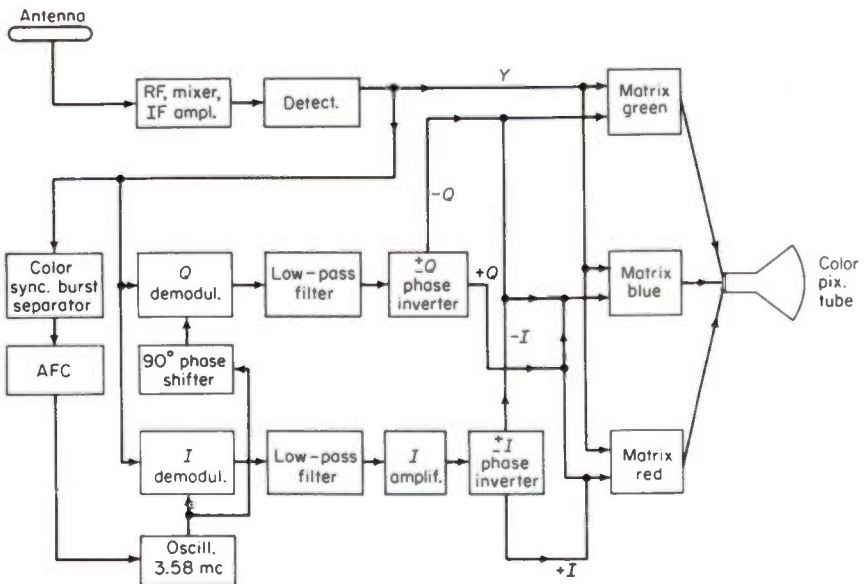


Figure 22.24. I - Q system of color reception.

The color channel first includes a color sync burst separator which extracts the 3.58-mc reference signal from the v-f modulation. This signal controls, through an AFC circuit of some kind, the frequency of the local 3.58-mc oscillator.

The local 3.58-mc frequency is applied to the I demodulator. The resulting I signal goes through a 1.5-mc low-pass filter, an amplifier, and a phase inverter providing $+I$ and $-I$ signals for easy matrixing.

The local 3.58-mc frequency is also applied to the Q demodulator through a 90-degree phase-shifting circuit, in conformity with what has been done at the transmitter. The resulting Q signal goes through a 0.5-mc low-pass filter, an amplifier, and a phase inverter providing $+Q$ and $-Q$ signals for easy matrixing.

Proper combination, or matrixing, of Y , I , and Q signals yields the red, blue, and green signals, applied to the color modulation electrodes.

Of course, auxiliary circuits, sweep sections, sync separator, and sound receiver must be added to this basic diagram, so that a simplified block diagram of the complete color receiver appears as shown in Fig. 22-25.

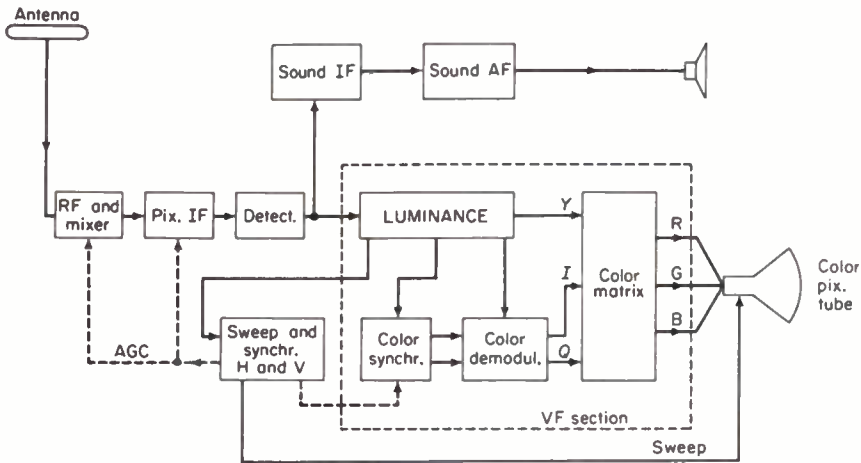


Figure 22-25. Additional stages of a color receiver.

22-24. V-f section

The differences between a color and a monochrome receiver occur only in the v-f section. Figure 22-26 shows in block form a typical diagram of this section in a color receiver. It can be divided into four subsections: luminance or *Y* channel, color synchronization, color demodulation, and color matrix.

The v-f signal coming from the picture detector is applied to a v-f amplifier in the *Y* channel. Its output feeds the AGC and sync separator circuits as usual. It also feeds the output- v-f stage through a delay line.

The function of this delay line is to compensate for phase differences created between luminance and chrominance signals. The *Y* signal leads the *I* signal, which itself leads the *Q* signal. Actually, then, two delay lines are necessary—one to delay the *Y* signal so that it is put back in phase with the *Q* signal and one to do the same for the *I* signal. The order of magnitude of the delays is 1.2 microseconds for the *Y* signal and 0.5 microsecond for the *I* signal. These values vary from one receiver to another. Sometimes it happens that the *I* signal obtained is suitable without the introduction of any additional delay.

The color synchronization section receives the v-f signal from the *Y* channel and applies it to a gated amplifier. The keying signal is the horizontal flyback pulse, so that the amplifier is operative only during horizontal blanking and separates the 3.58-mc color sync burst. In the diagram shown,

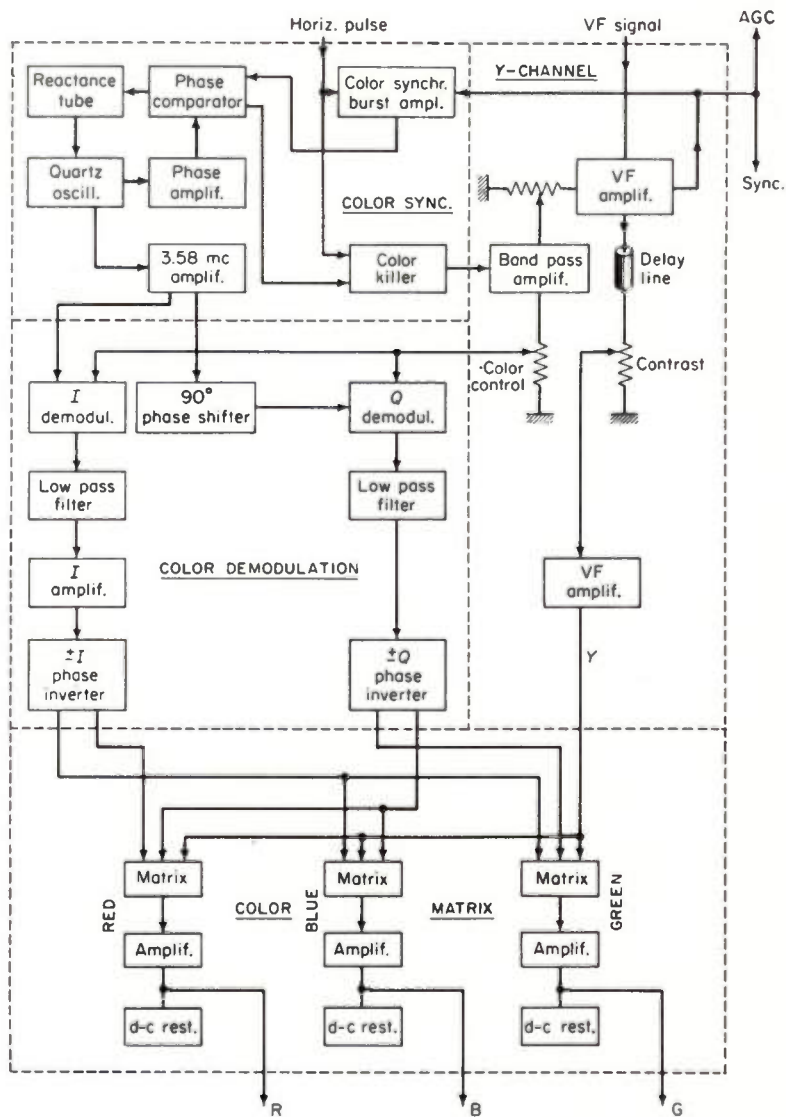


Figure 22-26. Composition of the color section of an I-Q receiver.

the local oscillator is quartz-controlled. Its signal is compared with the color sync burst signal in a phase comparator, whose output controls a reactance tube in parallel with the quartz oscillator. This AFC system provides the 3.58-mc synchronized frequency, which is applied to the demodulators.

The color demodulation section receives the 3.58-mc synchronized frequency and applies it to the *I* demodulator, followed by a low-pass filter, an amplifier, and a phase inverter. On the *Q* side, the 3.58-mc is first amplified, then goes through a 90-degree phase shifter, a demodulator, a low-pass filter, and a phase inverter. The 3.58-mc v-f signal to be demodulated is obtained from the *Y* channel amplifier through a bandpass filter which passes the color subcarrier frequency and the color sidebands only. Adjusting the amount of color signal fed to the *I* and *Q* demodulators constitutes a color control.

The *I*, *Q*, and *Y* signals are fed to the color matrix, which contains, for each primary color (red, blue, and green) a mixing matrix, an amplifier, and a d-c restorer.

A useful auxiliary circuit is the "color killer". When a monochrome transmission is received, v-f components in the color frequency range will go through the color section and create spurious color modulation of the picture. This is avoided by killing the color section for monochrome reception. The bandpass amplifier is normally biased off and does not work, so that the color section is cut off on a black-and-white program. When a color program is transmitted, the color sync burst appears and is used to turn on the bandpass amplifier through the sync color circuit.

For a similar reason, retrace is blanked out by one of the common retrace blanking circuits (not represented on the diagram).

22-25. *I-Q* summary

To help clarify thinking, the various processes leading to color reproduction by the *I-Q* system will now be summed up in a simplified manner:

1. *Tricolor signal.* At the transmitter, a triple color camera produces three separate signals which correspond to the contents of the televised picture in the three primary colors—red, green, and blue.

2. *Luminance signal.* The three signals (red, green, and blue) are mixed together in given proportions to produce the *Y* luminance signal, according to the equation

$$Y = 0.30R + 0.59G + 0.11B$$

The *Y* luminance signal is exactly identical to the v-f signal of monochrome transmissions and is transmitted by the same methods. It can then be received on a black-and-white receiver, which ensures compatibility.

3. *Chrominance signal.* The three primary color signals (red, blue, and green) are again combined in two matrices to produce I and Q signals as follows:

$$\begin{aligned} I &= 0.28G + 0.59R - 0.32B \\ Q &= -0.53G + 0.21R + 0.31B \end{aligned}$$

These signals are then sent through low-pass filters, cutting off at 1.5 mc for the I signal and 0.5 mc for the Q signal.

4. *Complete modulation signal.* The I and Q signals modulate a common 3.58-mc color subcarrier with a phase difference of 90 degrees. The modulators suppress the 3.58-mc subcarrier. The resulting chrominance signal has a phase which varies with the color and an amplitude which is a function of the saturation. It is combined with the Y luminance signal to provide the complete v-f modulation signal.

5. *Color synchronization.* To allow synchronization of the color demodulators in the receiver, a short burst of the color subcarrier at 3.58 mc is inserted in the back porch at black level following the horizontal sync pulse.

6. *Receiver detectors.* In the receiver, the total v-f modulation is detected as usual and provides the Y signal. Synchronous demodulators are used for the I and Q signals. They need a synchronized 3.58-mc carrier, which is produced by a local oscillator locked on the color sync burst.

7. *Color signals.* The Y , I , and Q signals are matrixed together to produce the three color signals (red, blue, and green) as follows:

$$\begin{aligned} R &= Y + 0.62Q + 0.96I \\ G &= Y + 0.64Q - 0.28I \\ B &= Y + 1.72Q - 1.11I \end{aligned}$$

8. *Color picture.* The three color signals are applied to a trichrome picture tube and reproduce the original televised scene in full color.

22-26. $(R - Y)$ and $(B - Y)$ system

The NTSC standard can readily be used in a system more economical than the I - Q system, at the price of some loss of color details. Reference to the color phase diagram will show that a simple 33-degree rotation will transform the I and Q axes into the $(R - Y)$ and $(B - Y)$ axes. Then the outputs of the two demodulators will be the $(R - Y)$ and $(B - Y)$ signals. However, because of the bandwidth of 1.5 mc devoted to the I signal, both $(R - Y)$ and $(B - Y)$ would only contain the I signal in the band comprised between 0.5 and 1.5 mc.

To avoid interferences between the two channels, it is then necessary to limit the bandwidth for both $(R - Y)$ and $(B - Y)$ to 0.5 mc, inside which

are found I and Q signals. This, of course, means that all frequencies above 0.5 mc contained in the I signal are lost, hence the reduction in color details.

The receiver diagram is simpler, since the demodulators provide $(R - Y)$ and $(B - Y)$ directly. A single matrixing will provide $(G - Y)$, and these three signals are applied to the picture tube color grids. The cathodes all receive the Y signal, and the addition of chrominance and luminance signals is performed by the picture tube itself.

The economy is evident with respect to an I - Q receiver. A single wideband amplifier is needed instead of three. Moreover, since the bandwidth is relatively narrow at 0.5 mc, a large gain is easily obtained in the chrominance channel and no extra v-f amplifier is needed. Also, since both color signals are handled by common stages, a single delay line is needed.

22-27. $(R - Y)$ and $(B - Y)$ receiver

Figure 22-27 represents a simplified block diagram of a receiver using the $(R - Y)$ and $(B - Y)$ system. Starting from the antenna, the first stages—r-f, mixer, and i-f amplifiers—are conventional.

The luminance channel contains the luminance detector, the delay line, and a Y amplifier whose gain controls the general contrast. It feeds the sync separator which synchronizes horizontal and vertical sweep sections. There are two extra circuits, the convergence and the voltage-regulation circuits.

A flat-face tube suffers from noticeable loss of focus near the edges of the picture and this is very important, for example, in the case of the color-points tube. To correct for this, a convergence correction is introduced. The correction signals are functions of the scanning signals. They are parabolic waveforms, derived from the scanning waveforms by the convergence circuit and applied to a special set of coils around the neck of the tube.

The EHV output for color picture tubes is on the order of 25 kv and the relatively high current makes it imperative to resort to some kind of EHV output voltage regulation.

Another detector is provided after the i-f amplifier. It feeds both the sound receiver and the chrominance channel. In this channel, a tuned bandpass amplifier drives color demodulators whose output is applied to a simple matrix to obtain $(G - Y)$ as well as $(B - Y)$ and $(R - Y)$. These three color difference signals are applied to the corresponding grids of the picture tube.

In addition, the output of the bandpass amplifier drives a gated amplifier, keyed on by the horizontal flyback pulse and whose output contains only the 3.58-mc reference color subcarrier burst. This burst shock-excites a tuned circuit whose tuning controls the phase, hence the colors produced.

This 3.58-mc signal is applied to a phase comparator which also receives the 3.58-mc of the local color oscillator and holds it in synchronism through a

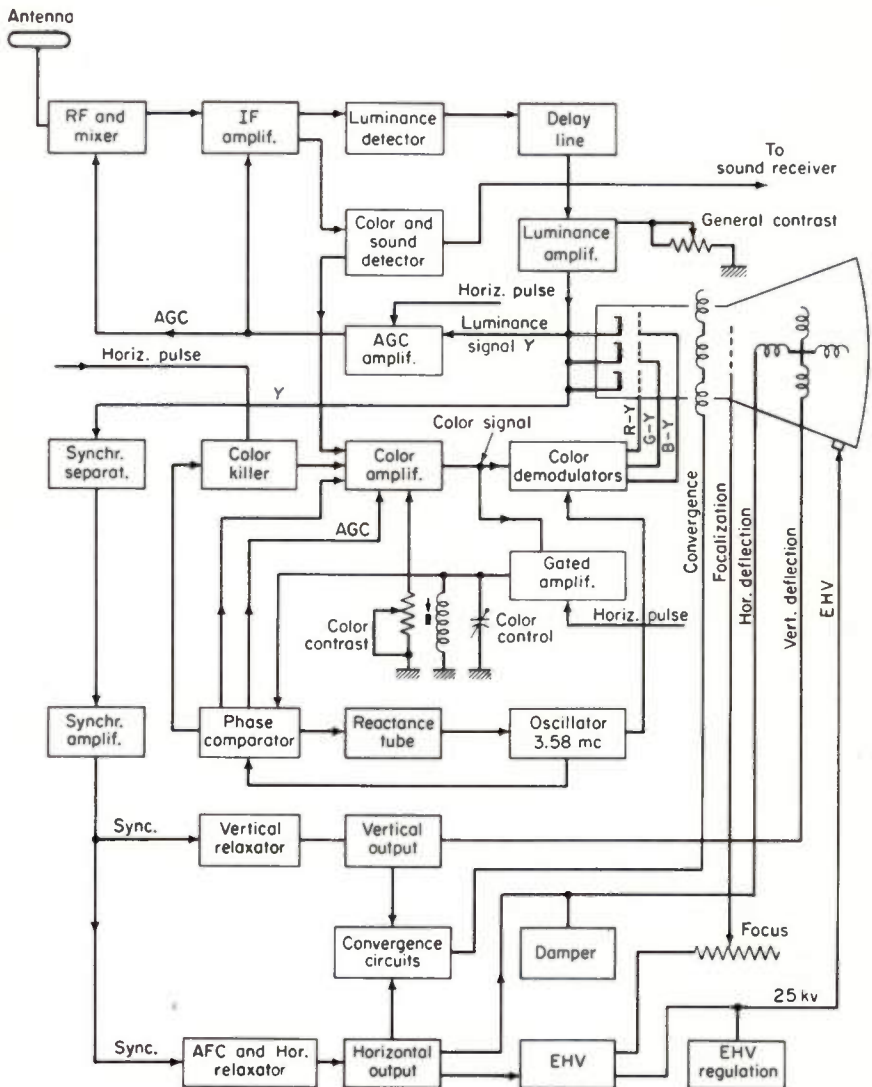


Figure 22-27. Block diagram of a $(B - Y) - (R - Y)$ receiver.

reactance tube. The locked 3.58-mc frequency is used in the color demodulators.

A color-killing circuit is used, as in the $I-Q$ receiver, to make the chrominance section operative only when a color program is received. It unblocks the bandpass amplifier when a color sync burst is present.

As for the $I-Q$ receiver, the basic diagram shown is illustrative only of the technique. Numerous variations are found in both cases.

22-28. *Color tubes*

The combination of the three primary colors necessary for full color reproduction may be either additive or subtractive.

Additive combination occurs, for example, when three beams of colored light are partially superposed. Blue-violet, red, and green suitably mixed will yield a white. The additive process is actually the one used in color picture tubes, which produce themselves the three fundamental colors necessary for full color vision. Blue, red, and green phosphors are used to this effect.

Subtractive combination occurs, for example, when colored filters are placed between a light source and an observer, with partial superposition. With three superposed filters, black will result for a yellow, a magenta purple, and a cyan. This is the process used in printing technique, where the colored inks replace the filters.

The three fundamental additive colors and the three fundamental subtractive colors are evidently different and their relation can be found by referring to the Newton color circle, or by placing them side by side around a circle.

Color picture tubes, using the additive process, must then provide the primary colors of the process—that is, red, green and blue (the latter is more precisely a blue-violet).

Many picture tube designs have been proposed; however, only a few will be examined here.

22-29. *Color-lines tube*

The first designs in the domain of color picture tubes were directly inspired from the standard design of black-and-white tubes; in particular, they use a single electron gun.

In one such design, the screen of the tube carries successive lines of red, green, blue, etc., phosphors. In this way, one of the three primary colors is produced according to the specific line hit by the electron beam.

22-30. *Color-areas tube*

The color areas tube has a screen face divided into three identical areas carrying red, green, and blue phosphors. A single electron beam is swept over the color areas.

A suitable optical system is used to obtain superposition of the three fundamental color pictures.

22-31. *Color-layers tube*

In this tube, the screen carries three superposed layers of phosphors—respectively, green, blue, and red. Electrons reach more or less deeply into

the phosphor layers, depending on their velocity, that is, on the acceleration voltage. They can thus be controlled to excite the fluorescence of one layer or another. For example, under 10 kv electrons will not go past the red layer; under 20 kv they will reach the blue layer; under 30 kv they will go all the way to the green layer.

This process automatically superposes the three primary colors but requires an inconvenient EHV switching.

22-32. Beam deflection tube

This tube is akin to the color-lines tube previously described. Its screen again carries parallel lines of phosphor in the three primary colors (red, green, and blue). However, before reaching the screen, the electron beam goes through a double grid. Control voltages applied to this double grid modify the beam trajectory and direct it toward the suitable phosphor line to produce the desired color.

The color lines may be horizontal or vertical.

22-33. Color-pyramids tube

A new principle is introduced with the color-pyramids tube. It uses three beams, coming from different directions, to excite the three primary colors. The screen carries a large number of small triangular pyramids similarly oriented.

Each face of a pyramid carries a primary color. The angle of arrival of each electron beam is such that it can hit only one side, say the red side, of each pyramid. In this way, each beam controls one color.

A practical difficulty is that the electron beams should hit the screen under widely different angles, so that the three electron guns must be widely separated. A version of the tube, in fact, had three necks at the back of the glass bulb.

22-34. Color-points tube

The color-points tube also uses three electron beams, one for each primary color. Its screen carries triangular groups of three phosphor points (red, green, and blue). Just behind the screen there is a perforated mask, with a small hole for each group of three points. To hit the screen, the electron beam must go through the hole. Since the three electron beams come from different directions, each one of them can hit the screen only at a given place, which it sees through the hole. The phosphor points are distributed in such a way that the red beam can reach only the red points, and similarly for green and blue.

The difficulty is that satisfactory resolution requires a large number of color points, with attendant manufacturing difficulties. Nevertheless, for the time being this type of tube is most commonly used in color receivers. It is also called a *shadow-mask* tube.

The only other type of color tube that has had any extensive application to date is the beam deflection tube, in two versions.

22-35. Deflecting grid tube

In this tube, as we have said before, the screen carries horizontal lines of phosphor in the three primary colors. Actually, the order of colors is blue, green, red, green, blue, etc., a green line always separating red and blue.

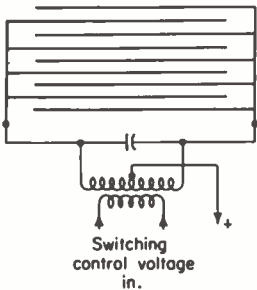


Figure 22-28. Electron beam control by double grid.

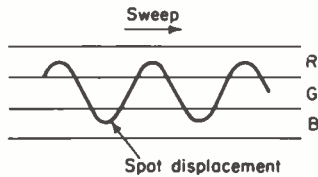


Figure 22-29. Spot movement on the screen.

Just behind the screen, the double grid is made of two series of very fine wires, alternatively connected to two common terminals as shown in Fig. 22-28. A series of wires falls behind the red lines. The other series falls behind the blue lines. The grids are accordingly called *red* and *blue grids*. The double grid deflects the beam with the help of a switching voltage applied to it at 3.58 mc with an amplitude of 440 v. To simplify matters, the double grid is part of a 3.58-mc resonant circuit as shown.

The resulting motion of the beam along the horizontal scanning direction is a wobbled line as shown in Fig. 22-29.

The double grid is connected through a center tap in the tuned circuit to a voltage of 5 kv, whereas the acceleration voltage for the tube is 18 kv. There is then a 13-kv voltage difference between the double grid and the screen, and this produces a strong electrostatic focusing.

Figure 22-30 shows what happens for three different a-c voltage conditions across the double grid. In (a) the red grid is positive and the blue grid negative. The beam is deflected toward the red grid and hits only the red lines, especially since electrostatic focusing makes it strongly convergent.

In (b) the opposite condition prevails. The blue grid is positive and the red grid is negative, so that the beam hits only blue lines. In (c) both grids are at the same potential, and the electron beam hits only green lines.

The superposition of the horizontal sweep displacement and of the small alternative vertical displacement due to the double grid gives rise to the resulting spot movement of Fig. 22-29, where the spot falls successively on lines red, green, blue, green, red, etc.

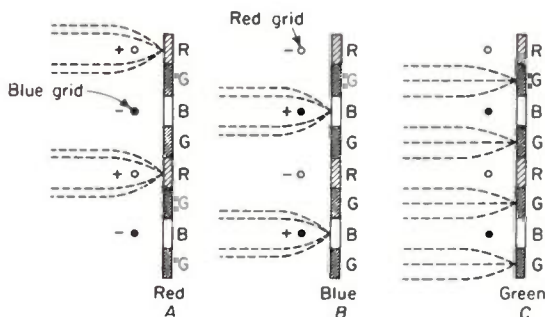


Figure 22-30. Color switching by a double grid.

To obtain a correct color reproduction, the electron beam must be switched in such a way that when it falls on a red line, for example, it is modulated by the red signal, and similarly for blue and green. This switching is done by turning the picture tube on and off with a waveform applied to the cathode. Suitable switching waveforms are represented in Fig. 22-31. They are derived from the 3.58-mc signal. Notice that it must be frequency-doubled to 7.16 mc to switch the green color. Green occurs twice as frequently as red or blue, but is switched on for a duration twice as short, so that color balance is re-established.

Variations of use are evidently possible. For example, switched ($R - Y$), ($B - Y$), and ($G - Y$) signals could be applied to the cathode and the Y luminance signal to the grid. In case of absence of color signals—that is, for monochrome transmission—the bias applied to the tube cathode is such that only the Y signal applied to the grid modulates the tube. If the amplitude of the sinusoidal vertical deflection signal is properly adjusted, the three primary colors red, blue, and green are equally excited and the net result is a black-and-white picture.

22-36. Post-acceleration tube

The post-acceleration tube uses vertical lines of phosphor alternatively blue, green, red, etc., and there is a fine grid just behind the screen. However, the

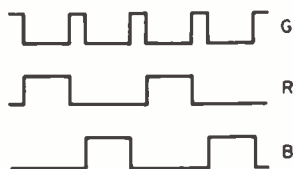


Figure 22-31. Switching signals for the three colors.

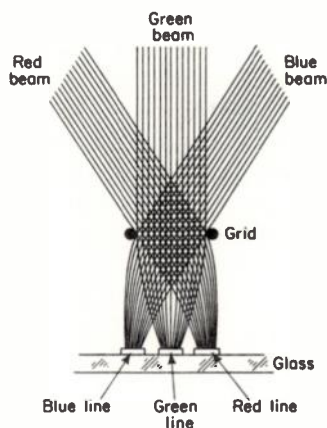


Figure 22-32. Angular separation of the three electron beams.

grid is not double and is not used to switch a beam, but to control and focus three electron beams. There are three electron guns side by side in the same horizontal plane, and they project three electron beams toward the screen. Each beam carries one primary color modulation.

The grid is at a potential of 6000 volts, approximately, and the screen is at 25,000 volts—hence the name of the tube. The electrostatic field between grid and screen has a strong focusing effect on the three beams arriving in slightly different directions. This is illustrated in Fig. 22-32, where the angle between beams has been greatly exaggerated for clarity. Actually, this angle is on the order of one degree at most. Three effects result from the field. One is a focusing effect concentrating the beam; the second is a parabolic deflection which incurves the trajectories; the third is a strong acceleration which projects the electrons on the screen at high speed.

Each beam falls on the screen at a definite position, where the suitable phosphor line has been laid. In fact, the strongly focused beam is smaller than the phosphor line width, so that there is some tolerance there. At the same time, the beam almost cannot hit the wrong line, which means that color purity is excellent. Good focusing is obtained only for a given ratio of screen voltage to grid voltage, so that this must be controlled and some form of stabilized EHV voltage is necessary.

22-37. Color-points tube utilization

The most common color tube is the color-points tube, known under various other names. Its screen carries approximately 600,000 color points in clusters or triads of one green, one red, and one blue point. The perforated mask

carries approximately 200,000 points and is placed behind the screen. There are three electron guns spaced 120 degrees apart. Each has a separate modulation electrode to receive one of the color signals, red, green or blue, and a focusing electrode whose potential is adjusted so that the beam hits the screen with minimum cross section. The three beams are shot through a common focusing system adjusted in such a way that the beams converge at the mask, so that they pass together through the holes. Horizontal and vertical deflection are applied simultaneously to the three beams.

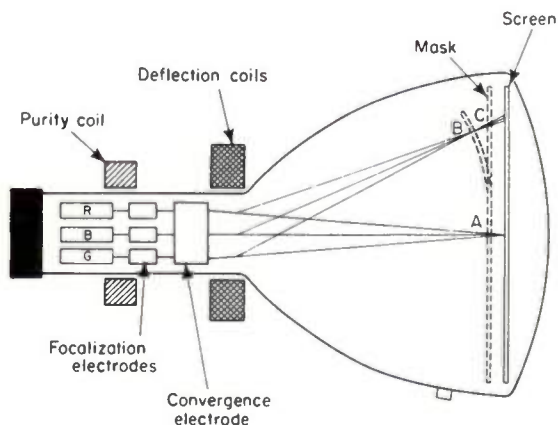


Figure 22-33. Color picture tube and convergence.

To obtain perfect convergence of the beams, besides the common convergence control there is a small individual control for each beam in the form of an adjustable external magnet called the *static convergence control*.

Since the optimum convergence surface is spherical, it is evident that the flat-screen tube is out of convergence on the edges if convergence is correct at center—hence the necessity of adding a dynamic convergence correction for both horizontal and vertical scanings as shown in Fig. 22-33. Correct convergence at center is shown at *A*. Without correction, convergence at the edges occurs at *B*. With correction, it is made to occur at *C*.

Correction waveforms are obtained by partial integration of the horizontal and vertical sweep signals and are applied to the dynamic convergence circuit.

Each electron beam corresponding to one primary color can be individually adjusted in intensity so that color balance is easily obtained. On the other hand, a control common to all three beams readily provides a total color contrast control.

The type of tube used evidently determines, at least partially, the circuits found in the receiver. Since both the color-points tube and the post-acceleration tube are very similar except for the mask or grid, a number of

circuits may be used for both. A few representative examples of the circuitry found in the color section of a receiver will be analyzed, more to illustrate the application of the principles explained than to give a selection of common diagrams in a field which is evolving at a fast rate.

22-38. Chroma-detector and bandpass amplifier

When a separate detector is used for the chrominance or chroma channel, it is connected at the output of the i-f amplifier as is the luminance detector.

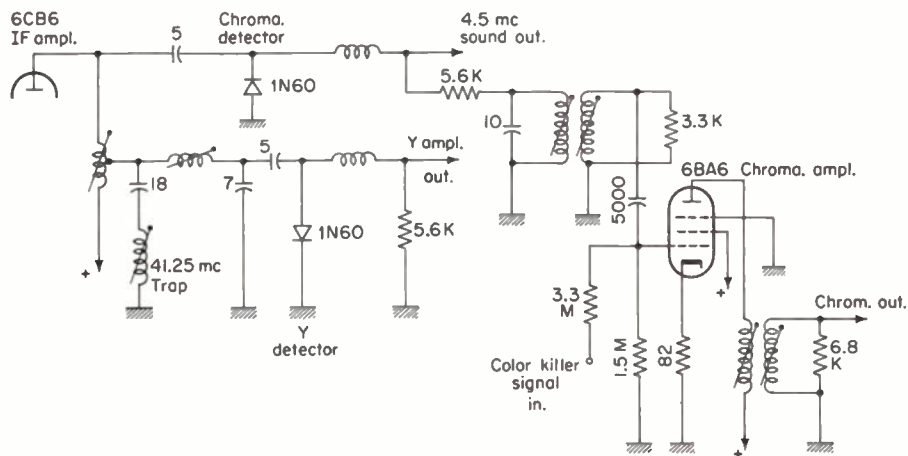


Figure 22-34. Band-pass chroma amplifier.

Figure 22-34 illustrates one example where the chroma detector also provides the 4.5-mc take-off point for the intercarrier sound i-f.

If a single detector is used, it provides the chroma and *Y* signals simultaneously. In both cases, the chroma signal is extracted by a bandpass amplifier tuned on the color subcarrier at 3.58 mc and passing 2 mc for the *I-Q* system or 1 mc for the $(R - Y)$ - $(B - Y)$ system. In the diagram, the color-killer signal is applied to the control grid of the chroma amplifier and cuts the tube off for monochrome reception, thus disabling the entire color section.

22-39. Color demodulators

Synchronous color demodulators frequently use triodes in a more or less elaborate circuit. Basically, the grid receives the local oscillator 3.58-mc signal in proper phase. The anode receives the 3.58-mc color signal, and the demodulated v-f color signal appears across an anode load as shown in Fig. 22-35.

Figure 22-36 illustrates a simple circuit for a narrow-band receiver, that is, a receiver of the $(R - Y)$ -($B - Y$) variety. It uses two demodulators for $(R - Y)$ and $(B - Y)$ signals, combined in a double triode, and simple matrixing provides the $(G - Y)$ signal.

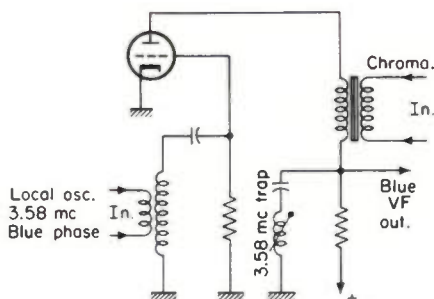


Figure 22-35. Principle of triode demodulator.

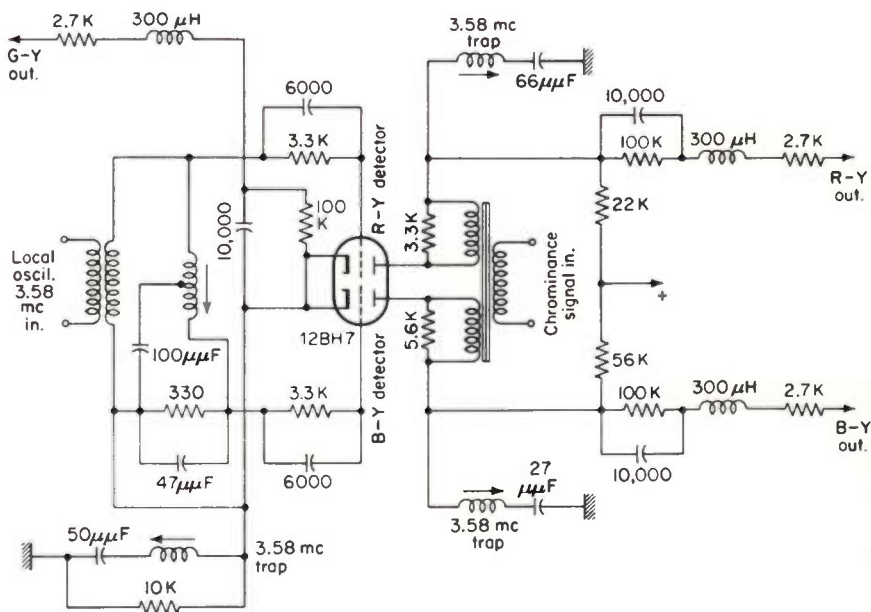


Figure 22-36. Double triode demodulator circuit.

The local oscillator 3.58-mc signal is directly applied to the grid of the $(R - Y)$ demodulator, and through a phase-shifting network to the grid of the $(B - Y)$ demodulator. The chrominance signal is applied to the plates through a double transformer. The demodulated v-f signals $(R - Y)$ and

$(B - Y)$ appear across the anode resistance loads and are available through high-frequency correction networks and insulating resistances.

Matrixing of the $(R - Y)$ and $(B - Y)$ signals is done in the common cathode circuit. The amplitude of the $(R - Y)$ and $(B - Y)$ signals has the correct value for direct mixing because of the different anode load resistances, so that the $(G - Y)$ signal is directly available across the 10,000-ohm cathode load and is picked off through a high-frequency correction network.

In each v-f color output, a 3.58-mc trap eliminates any residual carrier.

V-f color amplifiers, if any, are designed conventionally and for the required bandwidth, according to the system in use.

22-40. Color-burst amplifier

The color sync burst is separated from the v-f signal by a gated amplifier. Figure 22-37 shows a simple diagram. The complete v-f signal is applied to

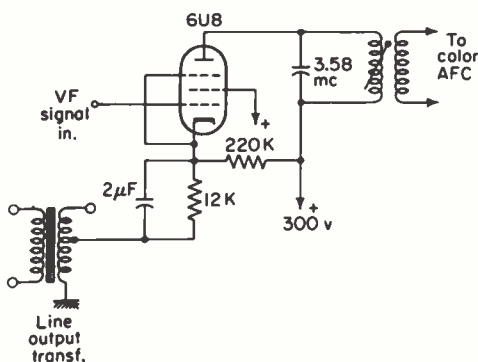


Figure 22-37. Color burst separator amplifier.

the grid of a pentode, normally cutoff by a high positive voltage applied to the cathode.

A negative horizontal flyback pulse provided by the line output transformer turns the tube on during retrace, and the 3.58-mc color sync burst appears amplified in the anode circuit.

22-41. Color oscillator

Numerous diagrams of synchronized 3.58-mc oscillators have been produced. Figure 22-38 represents a somewhat simplified version of a straightforward circuit.

A double-diode phase comparator receives the local 3.58-mc oscillation and the color sync burst coming from the gated-burst amplifier. The resulting

AFC voltage controls a reactance tube connected across a quartz crystal, thus modifying its frequency. The oscillator tube has in its plate a circuit tuned on 3.58-mc which feeds the color demodulators.

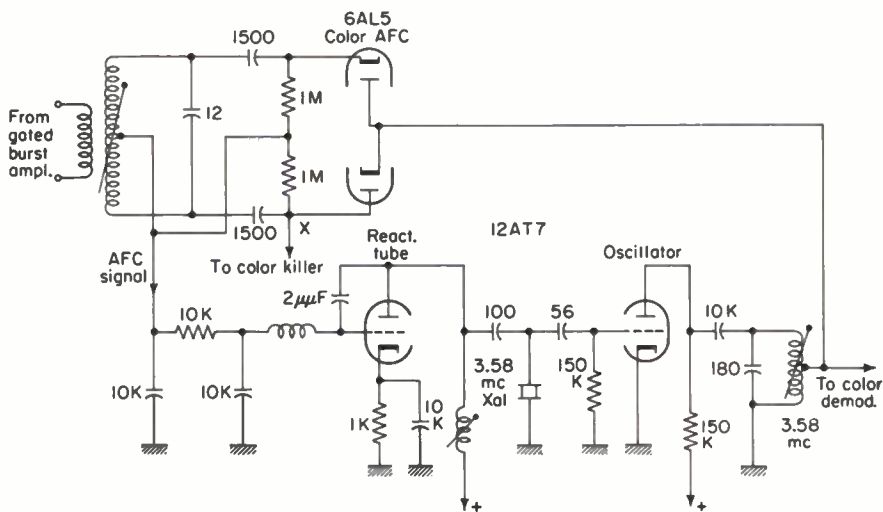


Figure 22-38. AFC controlled oscillator.

22-42. Crystal ringing circuit

A simpler way of obtaining the 3.58-mc local signal is to shock-excite a quartz crystal by the color sync burst. A diagram is given in Fig. 22-39.

The first tube is a gated-burst amplifier that receives the v-f signal on its grid and is keyed on during retrace by a negative pulse coming from the line output transformer. The amplified sync burst shock-excites the quartz crystal into oscillation. Because of its high Q , the crystal will ring for a long time, more than long enough to provide the 3.58-mc signal until the next color sync burst.

The exact crystal frequency is adjusted by the small series capacitance.

A grounded-grid amplifier stage follows. Its plate circuit contains a 3.58-mc resonant circuit providing the signal for the color demodulators.

22-43. Color killer

A simple color-killer circuit is represented in Fig. 22-40. A tube is fed a large positive pulse voltage by the line output transformer, and the plate current flowing through R creates a large negative voltage which is applied to the control grid of the bandpass chroma amplifier and cuts it off, disabling the entire color section.

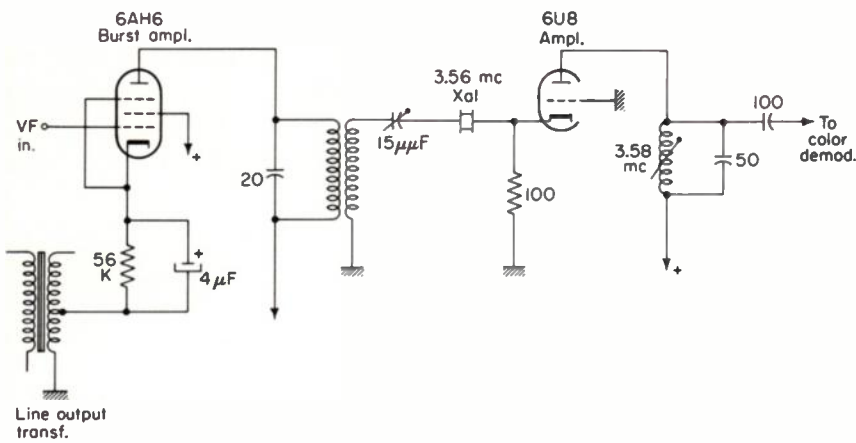


Figure 22-39. Crystal ringing circuit.

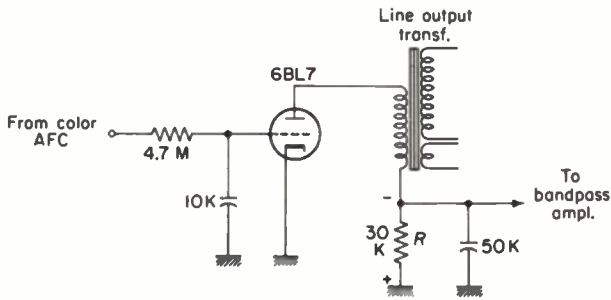


Figure 22-40. Color killer tube.

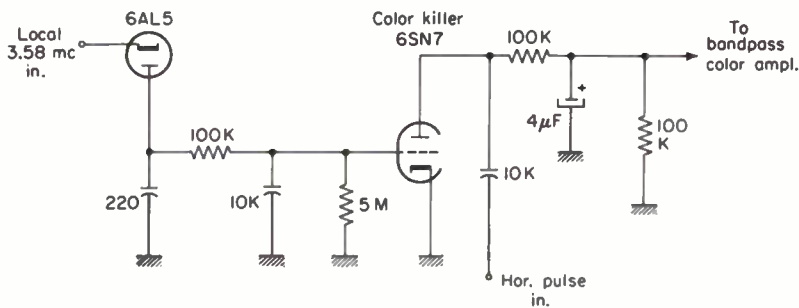


Figure 22-41. Color killer keyed by local color oscillation detector.

When a color program is received, a negative voltage is developed by the color sync burst at point *X* in the phase comparator of Fig. 22-38. This voltage is applied to the color-killer tube and is sufficient to cut it off. Its plate current disappears, and so does the negative bias applied to the control grid of the bandpass amplifier. The color section becomes operative.

With a crystal ringing oscillator circuit, the local 3.58-mc frequency is there only when color is received. It can then be detected and its bias used to control the color-killer tube as before. A diagram to this effect is given in Fig. 22-41. The positive flyback pulse from the line output transformer is applied in parallel to the anode circuit.

22-44. EHV supply

The standard color tube needs an accelerating voltage on the order of 20,000 volts and a focusing voltage on the order of 7000 volts. In addition, the accelerating voltage should have good regulation.

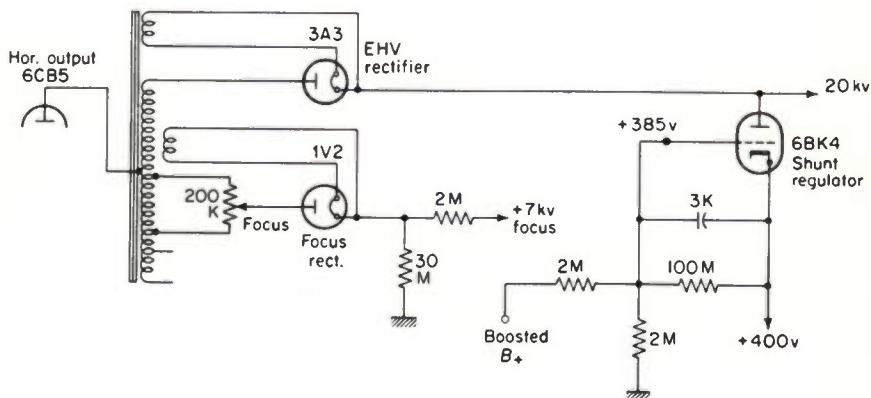


Figure 22-42. EHV supply for color picture tube.

Both high voltages are obtained from the line output transformer in a manner similar to that used in monochrome receivers. The details, of course, differ.

A representative diagram is shown in Fig. 22-42. An EHV rectifier driven by the line output transformer provides 20,000 v accelerating voltage. A special triode shunt regulator tube is used. Variations in EHV current consumption are reflected in a varying boosted B+ voltage, and this voltage controls the grid of the regulator tube so as to keep total EHV current consumption constant. If the picture tube requires less current, the regulator tube will pass more current, and vice versa.

Another EHV rectifier is fed an adjustable pulse amplitude and provides a variable focus voltage. The potentiometer controlling pulse amplitude is the focus control.

22-45. Dynamic convergence

The three beams of the color picture tube must converge in the plane of the mask so as to pass together through the same mask hole. Since the mask surface does not coincide with the spherical ideal deflection surface,

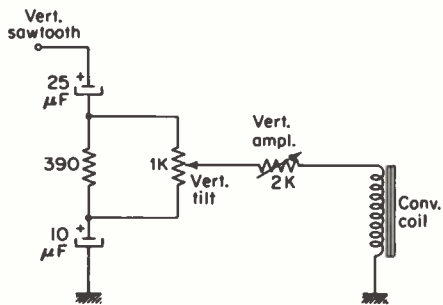


Figure 22-43. Dynamic vertical convergence.

convergence corrections must be introduced for both horizontal and vertical scanning.

The necessary dynamic convergence signals should be parabolic in shape. They are derived from the scanning waveforms.

Figure 22-43 shows how the dynamic vertical convergence signal is derived from the vertical sawtooth. This sawtooth is integrated by two capacitors in series. A potentiometer is inserted between the two capacitors and

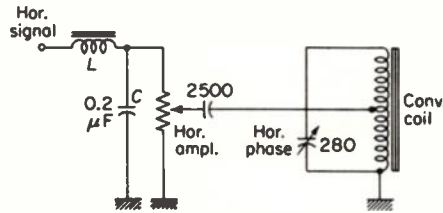


Figure 22-44. Dynamic horizontal convergence.

the position of its slider determines the amount of integration included in the output waveform, which tilts accordingly, hence the name *tilt control*. The integrated signal is fed to the convergence coil through a vertical amplitude control. This convergence coil is an electromagnet slipped on the neck of the tube and acting on one beam.

A similar circuit is used for each beam, so that the neck of the tube in fact carries an assembly of three convergence coils, properly positioned with respect to the three electron guns.

Figure 22-44 shows how the horizontal convergence signal is obtained. The horizontal sweep signal is picked off the line output transformer through an inductor L and integrated on a capacitor C . A fraction of this integrated waveform is taken through the horizontal amplitude control and applied to

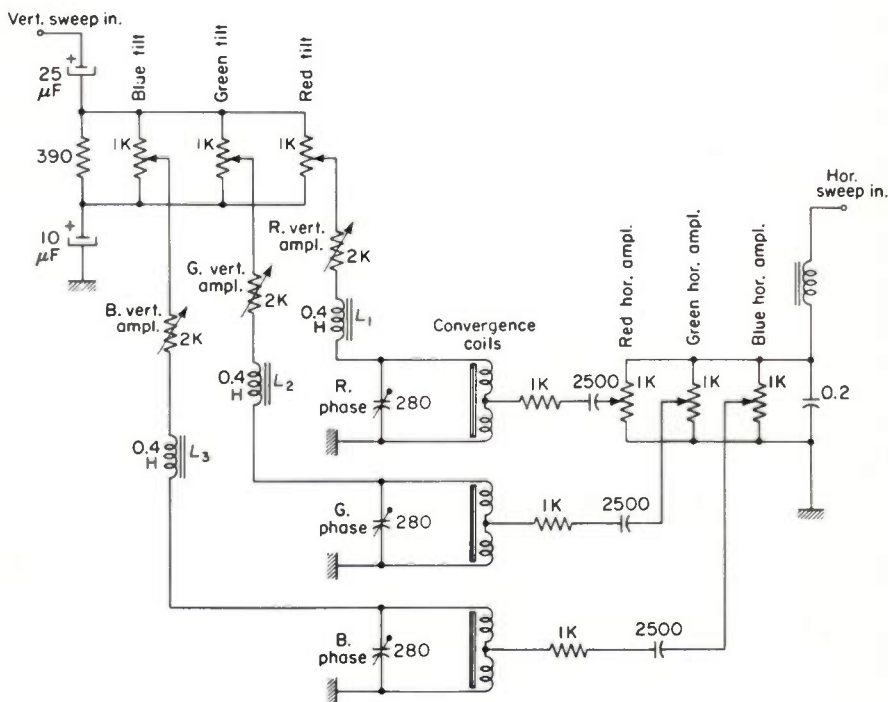


Figure 22-45. Complete convergence circuit.

the convergence coil, which is tuned to resonance by a variable capacitor. Adjusting this capacitor modifies the phase of the convergence signal.

Three identical circuits are provided, one for each electron gun, and an assembly of three convergence coils is slipped over the neck of the tube.

There is nothing to prevent combining the vertical convergence circuits with the horizontal convergence circuits if proper precautions are taken to avoid undesired interdependence. Even the same convergence coils may be used for both horizontal and vertical signals.

A complete dynamic convergence circuit consisting of three of each preceding circuit is illustrated in Fig. 22-45.

Vertical signals are kept off the horizontal circuits by the low-value, $2,500\text{-}\mu\text{f}$ capacitors. Horizontal signals are kept off the vertical circuits by the choke inductors L_1 , L_2 , and L_3 .

Problems

1. Explain how the remanence of the eye is put to good use in black-and-white television.
2. Explain how the remanence of the eye is put to good use in color television.
3. Flicker becomes apparent at a lower frequency for color than for monochrome pictures. Explain why.
4. Assuming the resolving power of the eye to be 1 minute of arc for a black-and-white picture, calculate the optimum distance between a viewer and a 21-inch diagonal screen.
5. Assuming the resolving power of the eye to be 3 minutes of arc for color, calculate the optimum distance between a viewer and a 21-inch diagonal screen (a) for the I - Q system and (b) for the $(R - Y)$ -($B - Y$) system.
6. Discuss the difference between *luminance* and *chrominance*.
7. Compare *trichromy* and *quadrichromy*.
8. Compare *Maxwell triangle* and *Newton circle*.
9. Show how the mixture of a purple and a green can produce a white.
10. Discuss the three fundamental parameters of color vision.
11. Compare additive and subtractive color synthesis.
12. Explain the use of dichroic mirrors in color television.
13. Compare simultaneous and sequential systems from the point of view of the receiver only.
14. Discuss the need for compatibility.
15. Discuss the advantages of the point sequential system with respect to the line or field sequential systems.
16. Explain the need for mixed high frequencies.
17. Discuss the usefulness of frequency interlace.
18. Describe the NTSC color system.
19. Justify the use of I and Q signals.
20. Compare the I - Q and $(R - Y)$ -($B - Y$) systems.
21. Explain why the color subcarrier should be suppressed.

22. Why is it necessary to synchronize the local 3.58-mc oscillator?
23. Establish in detail the mathematical relations between I and Q signals and $(R - Y)$, $(B - Y)$, and $(G - Y)$ signals.
24. Explain the role of a color-killer circuit.
25. Why should synchronous demodulators be used for color signals?
26. From a practical point of view, compare the various types of color picture tubes.
27. Justify the need for static and dynamic convergence adjustments with the perforated-mask type color tube.
28. Describe the operation of a crystal ringing color local oscillator.
29. Discuss the special EHV requirements of a color picture tube.
30. Explain the role of the 3.58-mc phase comparator.

Chapter 23

COMPLETE COLOR RECEIVER

In this chapter a complete diagram of a color television receiver is given, and the differences in design between a color and a monochrome television receiver are explained.

In addition, in the chroma section, we will illustrate how some of the various color circuits described in the preceding chapter are interconnected to ensure correct operation.

A short circuit analysis points out the main features of the design and relates them to the preceding chapters.

23-1. General design

In the preceding chapter we have provided examples of practical circuit design for various sections of a color television receiver; however, no doubt it will be highly useful to analyze a representative integrated circuit diagram to obtain a better idea of the relationships of the various sections.

The chassis chosen for this purpose is the RCA model CTC7, which is found with minor variations in a number of color receivers. It uses the $(R - Y)$, $(B - Y)$ system.*

This chassis provides reception of monochrome as well as color programs. It is equipped with a VHF only or VHF-UHF tuner. It uses the 21-inch, RCA shadow-mask, three-gun color picture tube, and some models have a remote control accessory which will not be described here. A few of the salient characteristics of the chassis are automatic color-killer circuit, regulated EHV voltage, and automatic phase control in the chroma circuits.

* All the relevant data have been obtained through the courtesy of the RCA Service Co.

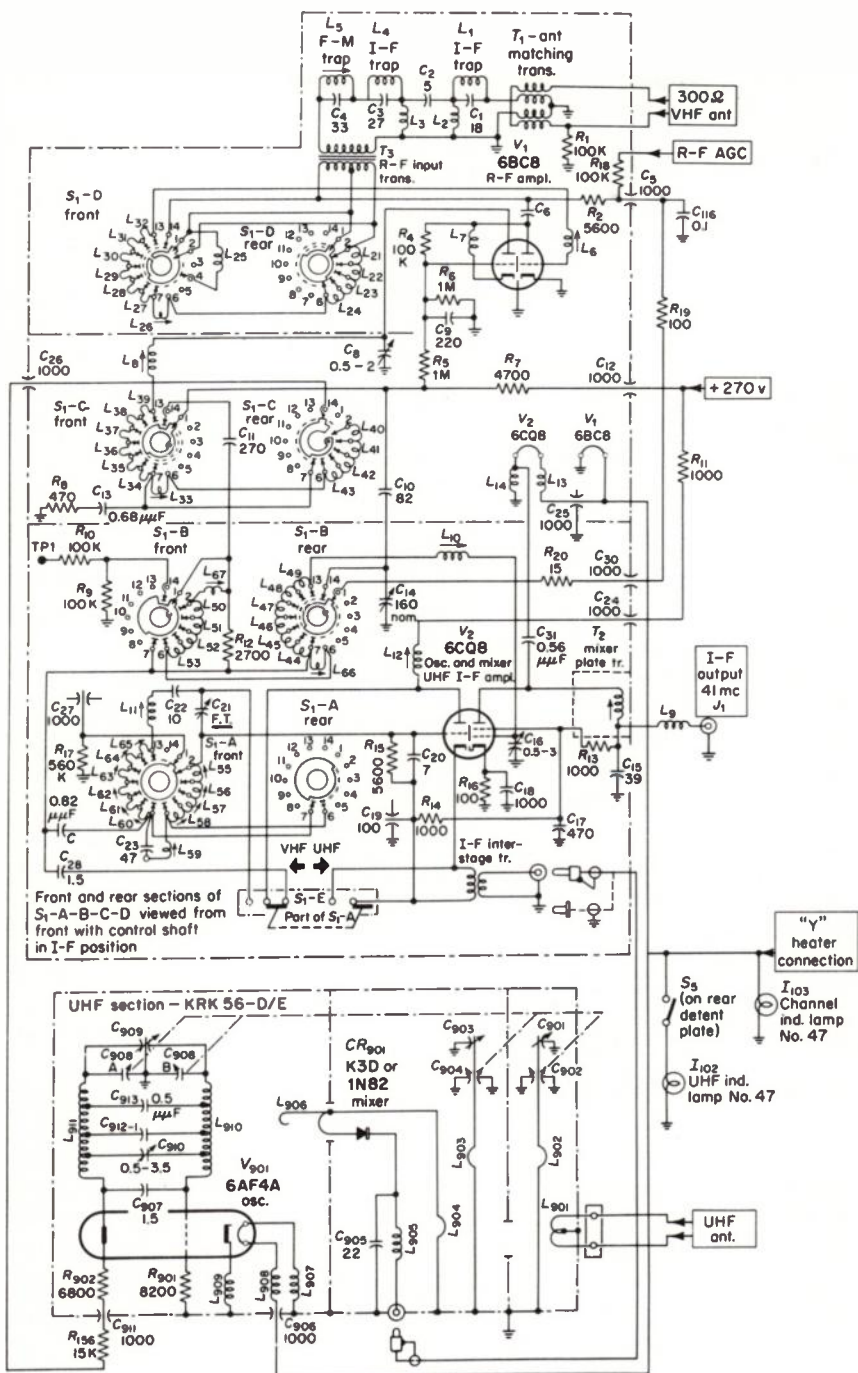


Figure 23-2. Diagram of a combined VHF and UHF tuner.

The vertical sweep section contains an oscillator and a power output tube. The horizontal sweep section is composed of an AFC stage, an oscillator stage, an output tube, a damper tube, and a high-voltage rectifier, plus a high-voltage regulator and a focus-voltage rectifier.

An AGC tube is provided, and convergence circuits are connected to the horizontal and vertical output circuits.

The chroma circuits begin with a bandpass amplifier using two stages. The first stage drives a burst amplifier which feeds a color phase comparator. This comparator controls a 3.58-mc color oscillator through a reactance tube. At the same time, it drives a color-killer detector which controls the color-killer tube. In the absence of color signal, the color killer cuts off the second bandpass amplifier. The signals coming from this amplifier and from the 3.58-mc oscillator are applied to the so-called *X* and *Z demodulators*. The detected voltages are combined to provide the three color-difference signals ($R - Y$), ($G - Y$), and ($B - Y$) applied to the three grids of the picture tube.

Each section will now be described briefly.

23-3. Tuner

Figure 23-2 illustrates a VHF-UHF tuner used with this chassis. For the VHF section it includes a double-triode r-f amplifier in a series cascade circuit, and a triode-pentode oscillator and mixer. Tuning is accomplished by switching inductances.

The UHF tuner uses continuous tuning. It has a triode oscillator and a crystal diode mixer.

The VHF tuner covers the 12 channels from 54 to 216 mc; the UHF tuner covers 70 channels from 470 to 890 mc. The picture carrier i-f is on 45.75 mc, and the sound carrier i-f is on 41.25 mc. With these values, the color subcarrier falls on 42.17 mc.

23-4. Picture i-f amplifier

As can be seen in Fig. 23-3, the picture i-f amplifier uses three pentodes with bifilar coupling coils. Stagger tuning is resorted to in order to obtain the necessary bandwidth. The first two stages are AGC-controlled. The third stage drives two separate crystal diode detectors, one for sound and one for picture. This, added to a thorough 4.5-mc trapping, minimizes the 920-kc beat pattern which might appear on the screen during color programs. This beat results from the sound carrier, which is 4.5 mc above the picture carrier, being mixed with the color subcarrier, 3.58 mc above the picture carrier. The resulting beat frequency is:

$$4.5 - 3.58 = 0.92 \text{ mc}$$

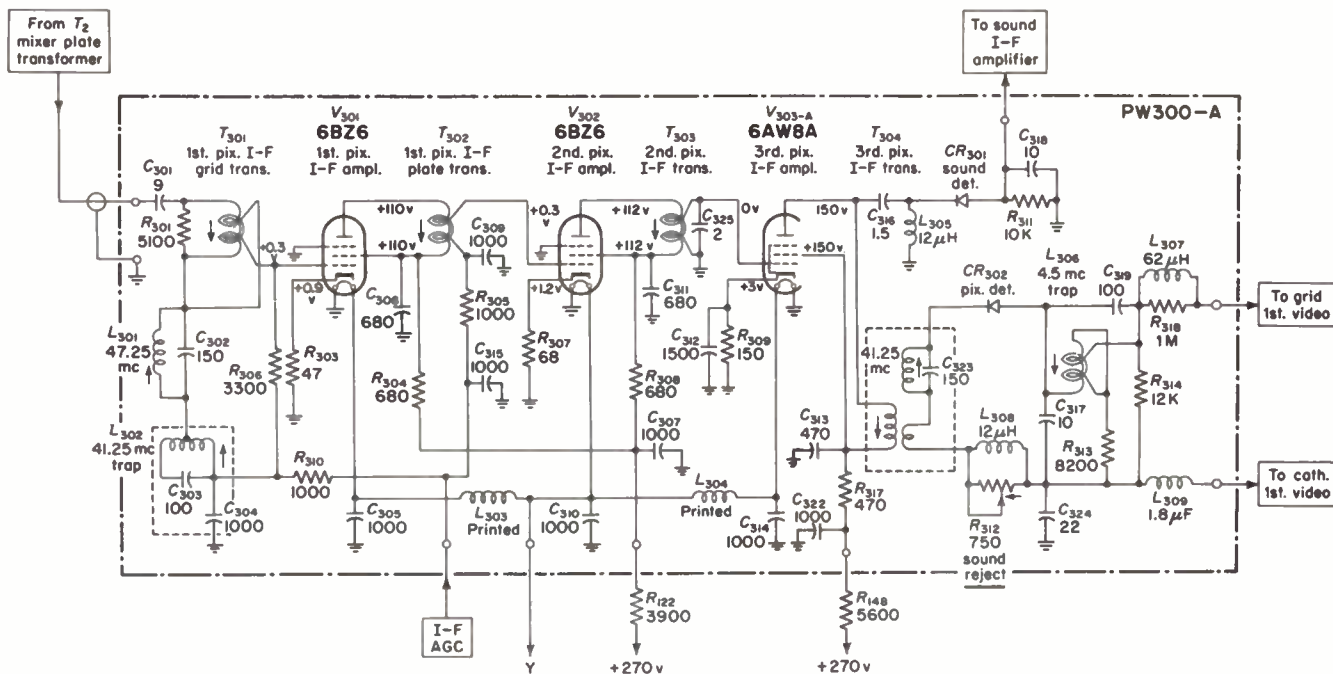


Figure 23-3. Picture i-f amplifier and detector.

23-5. V-f amplifier

The v-f amplifier of Fig. 23-4 has two stages, with series-shunt high-frequency compensation. The first stage is driven between grid and cathode by the signal coming from the picture detector. It feeds the sync separator, the noise inverter, the AGC tube, the bandpass amplifier, and the second v-f stage. This second stage has a contrast control in its cathode and drives the three cathodes of the color tube, with the injection of a vertical blanking pulse coming from the vertical sweep section.

23-6. Sound receiver

The sound receiver section is represented in Fig. 23-5. A pentode is used as a 4.5-mc sound intercarrier i-f amplifier. It drives, through a double-tuned i-f transformer, a quadrature pentode detector having a circuit tuned on 4.5 mc in its suppressor grid circuit.

The demodulated a-f signal is applied to an audio power output tube in which a tone control is provided. An audio output transformer feeds three loudspeakers for quality sound reproduction.

23-7. Synchronization

Figure 23-6 shows the synchronization and noise cancellation circuits. The noise cancellation triode has its cathode connected to the cathode of the first v-f amplifier, and therefore receives the full v-f signal. Its grid is biased in such a way that the tube is cut off and does not conduct for a normal v-f signal. However, when a noise pulse exceeds the maximum v-f signal, it reaches the conducting region of the tube and appears amplified at the plate, with the same phase it had at the input, since the tube is cathode-driven.

At the same time, the noise pulse will appear amplified at the plate of the first v-f amplifier but with opposite phase, since the v-f tube is grid-driven. Adding these two pulses of opposite phase across a common plate load resistance will result in noise cancellation.

The v-f signal, with noise thus eliminated, is applied to a triode separator working as a grid detector, and separated sync pulses appear in the plate circuit. They are further clipped and amplified by a second triode and fed to the horizontal phase comparator. Simultaneously, they are applied to the vertical oscillator through an integrating circuit.

23-8. Horizontal sweep section

The diagram of the horizontal sweep section is given in Fig. 23-7. The oscillator and control functions are combined in one double-triode tube in a so-called *synchroguide circuit*. The grid of the control tube receives the sync

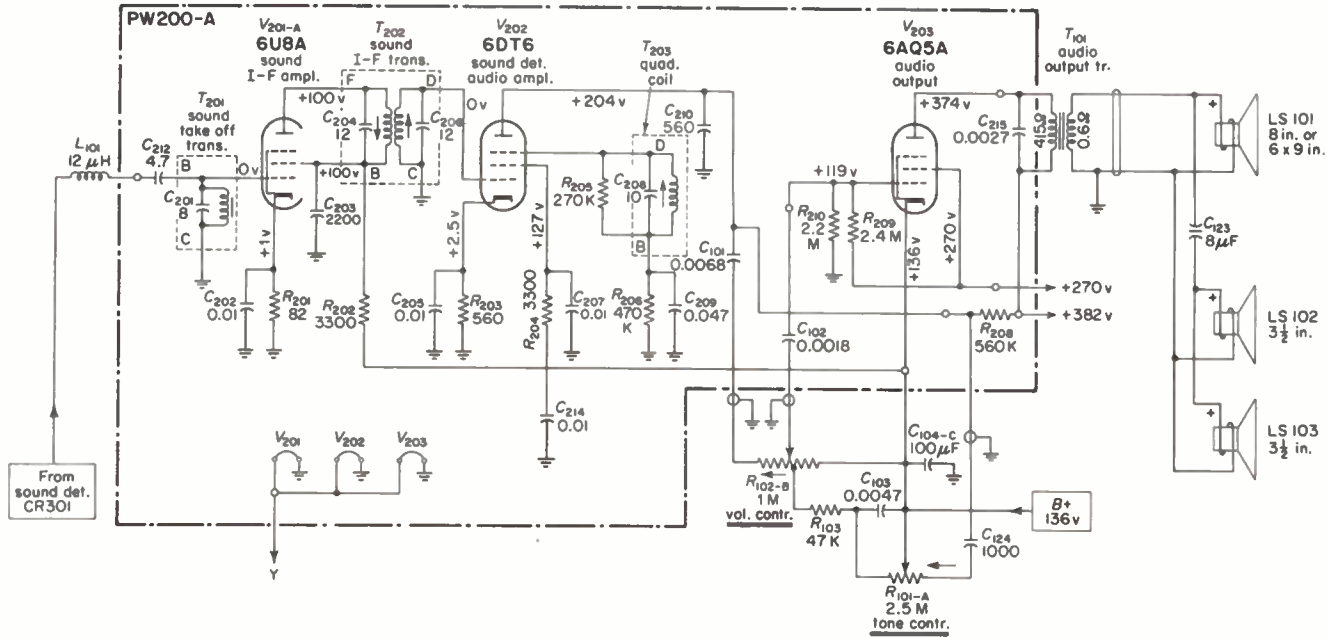


Figure 23-5. Sound receiver.

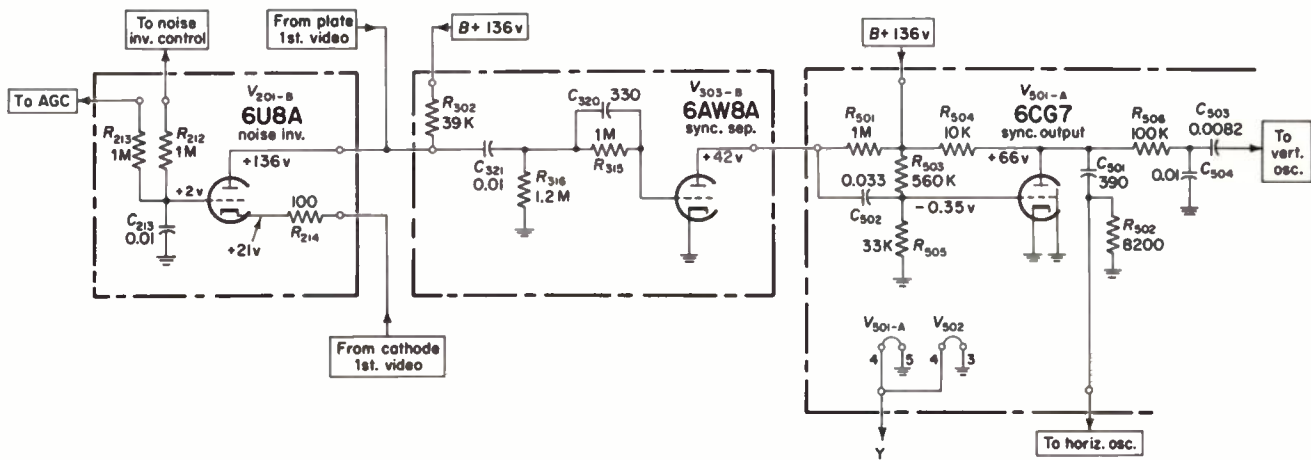


Figure 23-6. Sync separator and noise inverter.

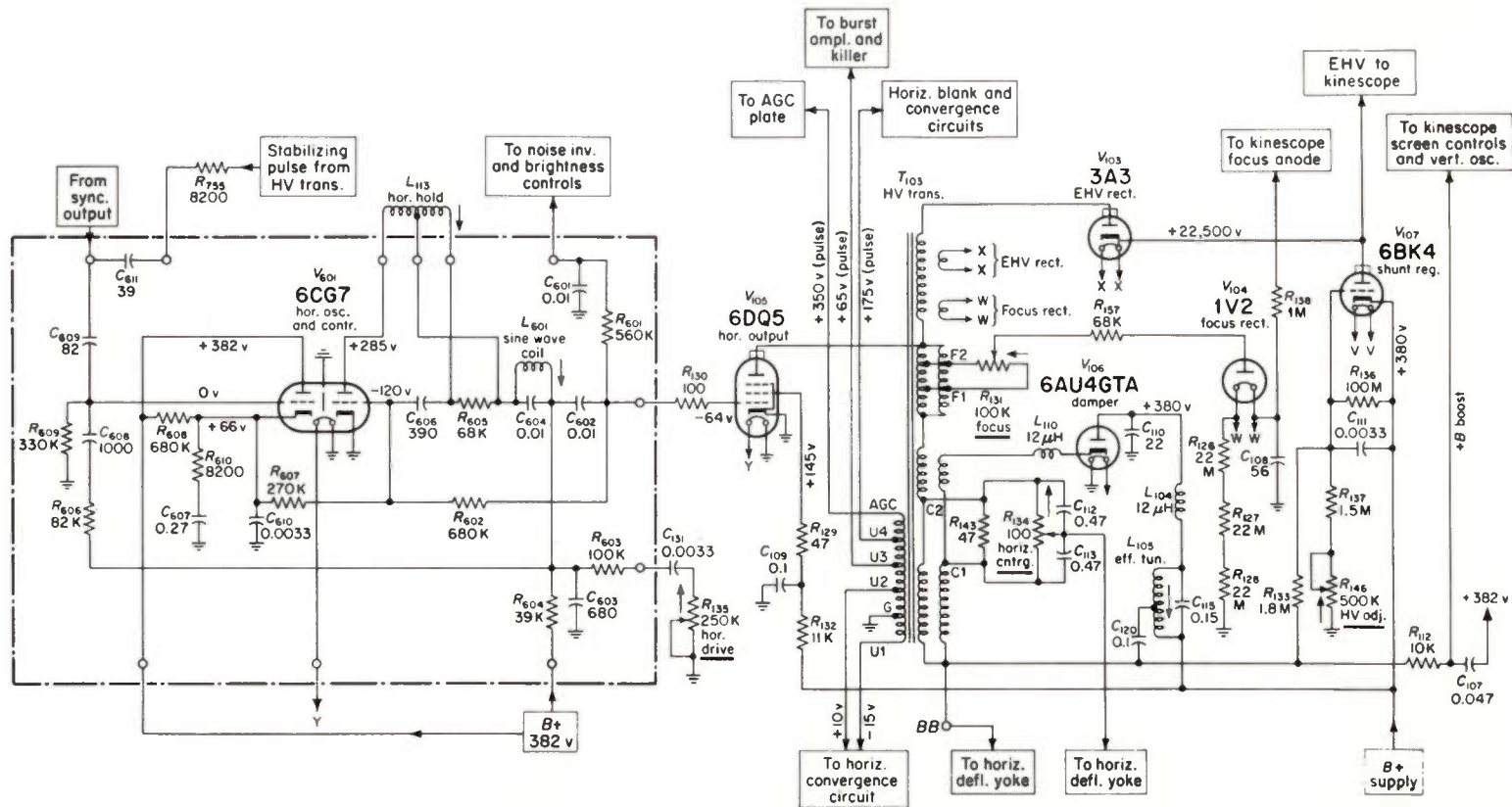


Figure 23-7. Horizontal sweep section and EHV.

signals and the output signal of the horizontal oscillator. The resulting control voltage appears on the cathode and is applied to the grid of the oscillator, which is a plate-to-grid coupling triode oscillator with suitable distortion circuits to provide the grid drive for the power stage.

The plate of the power tube is coupled to a line output transformer which feeds a number of auxiliary circuits, as indicated on the diagram. A centering arrangement is provided in connection with the horizontal coils. A diode is used for damping, and a linearity adjustment is included in its plate circuit. The EHV section uses a diode rectifier to provide 22.5 kv for the final anode of the picture tube. This voltage is regulated by a triode in a shunt-type regulator circuit. Any variation in the EHV current required by the picture tube is reflected as a variation in boosted voltage. This boosted voltage is applied to the grid of the regulator tube, whose plate current varies with the boosted voltage in a manner opposite to the picture tube current variation. Since both regulator tube and picture tube are fed by the same EHV supply, total current consumption remains constant and so does the EHV output.

Focus voltage is obtained through a tap on the line output transformer, with control of the voltage by a potentiometer. The pulse is rectified by a diode and the resulting d-c voltage is applied to the focus electrode.

23-9. Vertical sweep section

Figure 23-8 illustrates the diagram of the vertical sweep section.

The oscillator is a triode that has crossed plate-to-grid couplings with the output stage to work as a multivibrator. The usual frequency, amplitude, and linearity controls are provided.

The power output stage drives the vertical deflection coils through a transformer and provides a vertical blanking pulse added to the video signal at the plate of the second v-f tube. The vertical output transformer has a bifilar secondary winding to allow for an electrical centering control.

23-10. AGC circuit

The automatic gain control circuit represented in Fig. 23-9 is a classical keyed-tube arrangement using a triode. The v-f signal is applied to the grid, and the tube is turned on during horizontal blanking by a pulse coming from the line output transformer. The negative d-c voltage appearing at the plate is filtered and used for AGC purposes in the tuner and picture i-f amplifier.

23-11. Chroma circuits

Up to now, with the exception of the EHV supply, the diagram of the receiver does not differ much from that of a monochrome television receiver. The following sections, however, properly belong to the color circuitry.

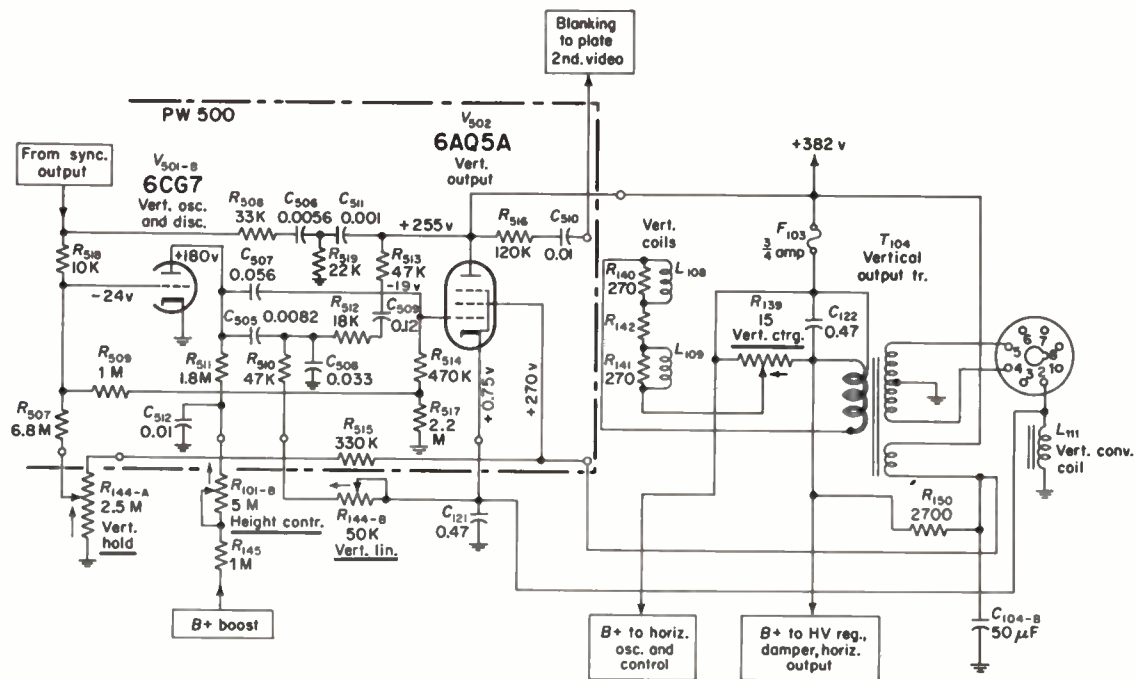


Figure 23-8. Vertical sweep section.

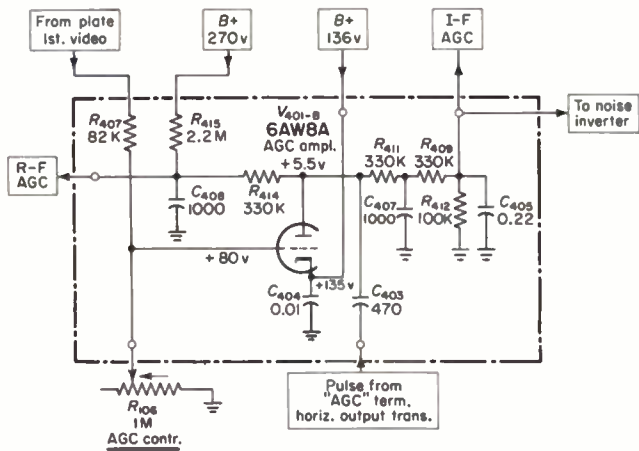


Figure 23-9. AGC circuit.

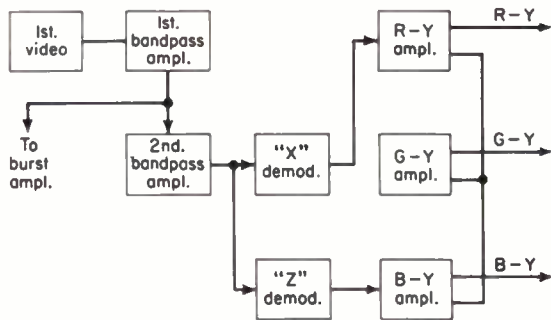


Figure 23-10. Block diagram of the chroma circuits.

Figure 23-10 is a block diagram of the chroma circuits, which illustrates the interconnection of the various functions and facilitates an understanding of the schematic diagram of Fig. 23-11.

The v-f signal coming from the plate of the first v-f tube is applied to a bandpass amplifier, having a bandwidth on the order of 1 mc, centered on a frequency of 3.58 mc, and using two tubes with a gain control which is the color saturation control. The amplified color signal is applied to the grids of two synchronous triode demodulators working along two axes 57.5 degrees apart and designated as *X* and *Z* demodulators. The *X* detector feeds an (*R* — *Y*) amplifier triode which drives the red grid of the picture tube. The *Z* detector feeds a (*B* — *Y*) amplifier triode which drives the blue grid of the picture tube. The signals appearing across the common cathode load of the (*R* — *Y*) and (*B* — *Y*) amplifiers are directly applied to the cathode of a

($G - Y$) amplifier. The phase relations are such that a ($G - Y$) signal results at the plate of the ($G - Y$) triode amplifier and is applied to the green grid of the picture tube.

Horizontal blanking is obtained through a pulse coming from the line output transformer, amplified by a triode, and cutting off the color-difference amplifying triodes during retrace.

23-12. Color synchronization

A block diagram of the color synchronization circuits is given in Fig. 23-12. It clarifies the interconnections among the various stages of the theoretical

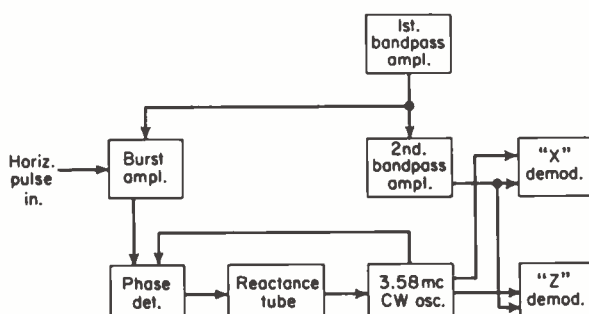


Figure 23-12. Block diagram of color synchronization.

diagram of Fig. 23-13. The main function of this section is to produce a 3.58-mc signal synchronous with the signal used at the transmitter.

Color synchronization is transmitted as a short 3.58-mc burst during horizontal retrace, following the horizontal sync pulse.

The composite color signal is applied to the grid of a burst amplifier tube, which is normally cut off. A pulse from the horizontal output transformer unblocks the tube during line retrace, so that only the amplified color sync burst appears in the plate. This reference 3.58-mc signal is applied through a tuned transformer to a double-diode phase comparator.

Now a pentode is used as a crystal-controlled 3.58-mc local oscillator. The signal it produces is transmitted through a tuned transformer to the double-diode phase comparator. The resulting d-c output voltage of the phase comparator is applied to a reactance tube which controls the frequency of the quartz oscillator and keeps it in synchronism with the reference 3.58-mc transmitted signal.

The synchronized 3.58-mc signal goes through a CW driver transformer to be applied, with the proper phase, to the X and Z demodulators.

The color-killer detector is a triode used as a diode. It receives a fraction

of the local 3.58-mc signal and also the color sync burst, so that a negative d-c voltage is developed by the diode and is applied to the grid of the color-killer triode, which it cuts off. The plate voltage of the killer tube is used as bias on the grid of the second bandpass amplifier. With the killer tube cut off, the second bandpass amplifier operates normally.

In case of a monochrome transmission, there is no color sync burst. The color-killer detector does not provide any bias to the killer tube, which saturates and brings the second bandpass amplifier grid bias beyond cutoff, thus disabling the whole chroma section. An advantage of this arrangement is that noise pulses cannot cause the second bandpass amplifier to operate.

23-13. Picture tube

The circuits associated with the color picture tube are reproduced in Fig. 23-14.

The screen controls are used to balance the amount of red, green, and blue light produced by the screen. In connection with the background controls of

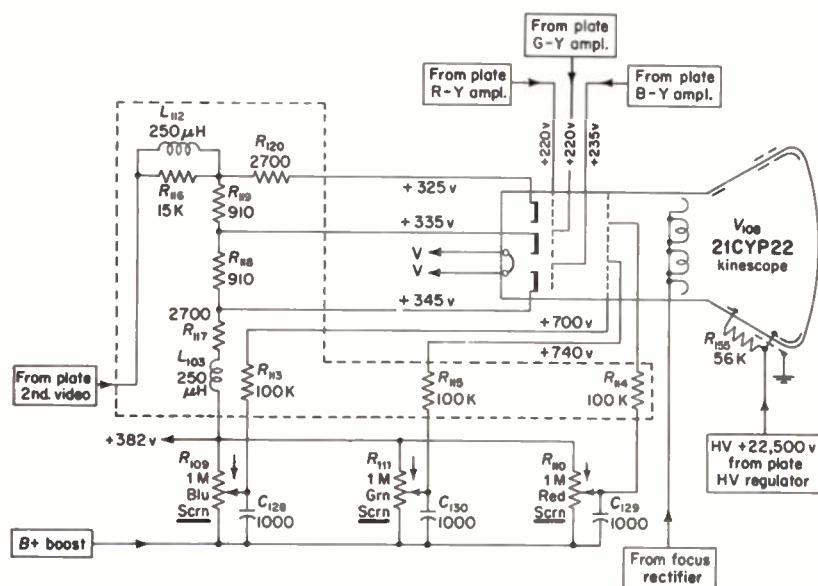


Figure 23-14. Color picture tube and associated circuits.

Fig. 23-11, they ensure that a monochrome picture will be received in black and white without detrimental coloring.

A number of auxiliary components are used with the tube. They are the deflection yoke, the convergence yoke, the lateral beam-positioning magnet, the purity magnet, the edge purity magnets, and the Z purity coil.

23-14. Convergence

Static convergence at the center of the screen is obtained by adjusting the convergence magnets on the neck of the picture tube.

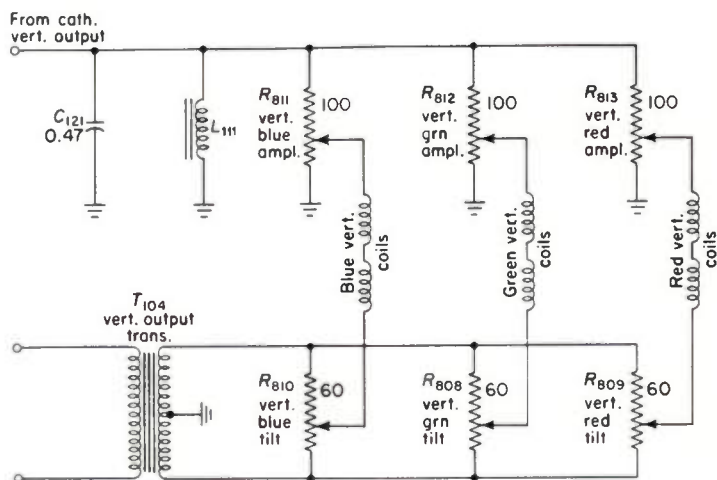


Figure 23-15. Vertical convergence.

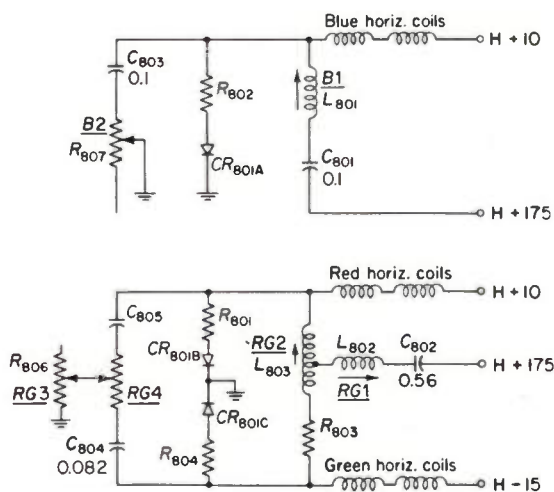


Figure 23-16. Horizontal convergence.

Dynamic convergence all over the screen is obtained by means of a triple convergence coil, through which flow currents derived from vertical and horizontal sweep currents.

The vertical convergence section is shown in Fig. 23-15. The current

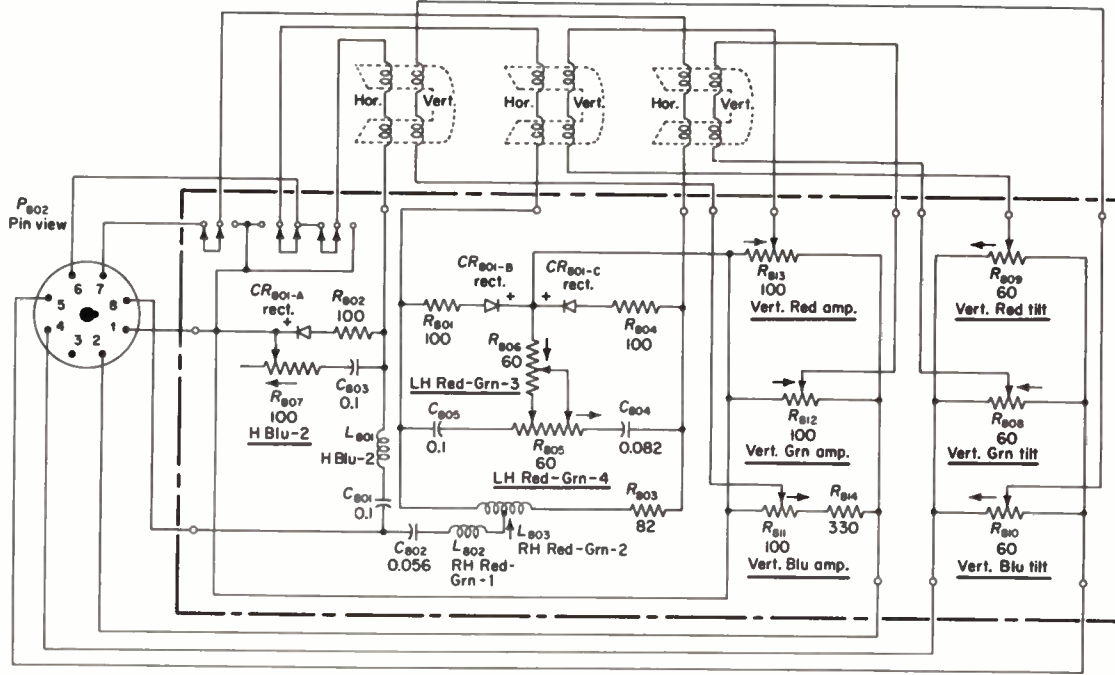


Figure 23-17. Complete convergence circuits.

coming from the cathode of the vertical output tube is controlled by a potentiometer and flows through each coil. In addition, another current coming from the secondary of the vertical output transformer adds or subtracts from the cathode current, depending on the position of the tilt control. Adjustment of the amplitude and tilt control for each color will ensure vertical convergence on the screen.

The horizontal convergence section is shown in Fig. 23-16. Reference to Fig. 23-7 will show that the convergence signal is derived from three taps on a separate winding of the line output transformer, with a grounded intermediary tap. In the blue convergence circuit, for example, the variable inductor adjusts the shape and amplitude of the convergence current. Shape control is further exercised by a potentiometer included in a time-constant circuit. The inductor controls the convergence at the right side of the screen, whereas the potentiometer controls the convergence at the left side of the screen. Similar controls are found for the red and green circuits, except that they are combined in opposition.

Figure 23-17 represents the complete convergence circuits and shows how vertical and horizontal convergence have been combined by using a set of two triple coils.

23-15. Power supply

A heavy power transformer, associated with two full-wave rectifiers in parallel and filter circuits, provides the required 380 and 270 volts of plate

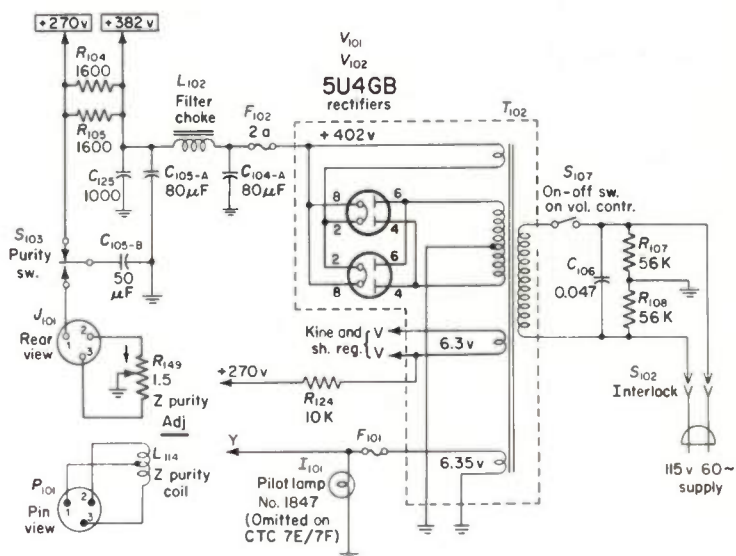
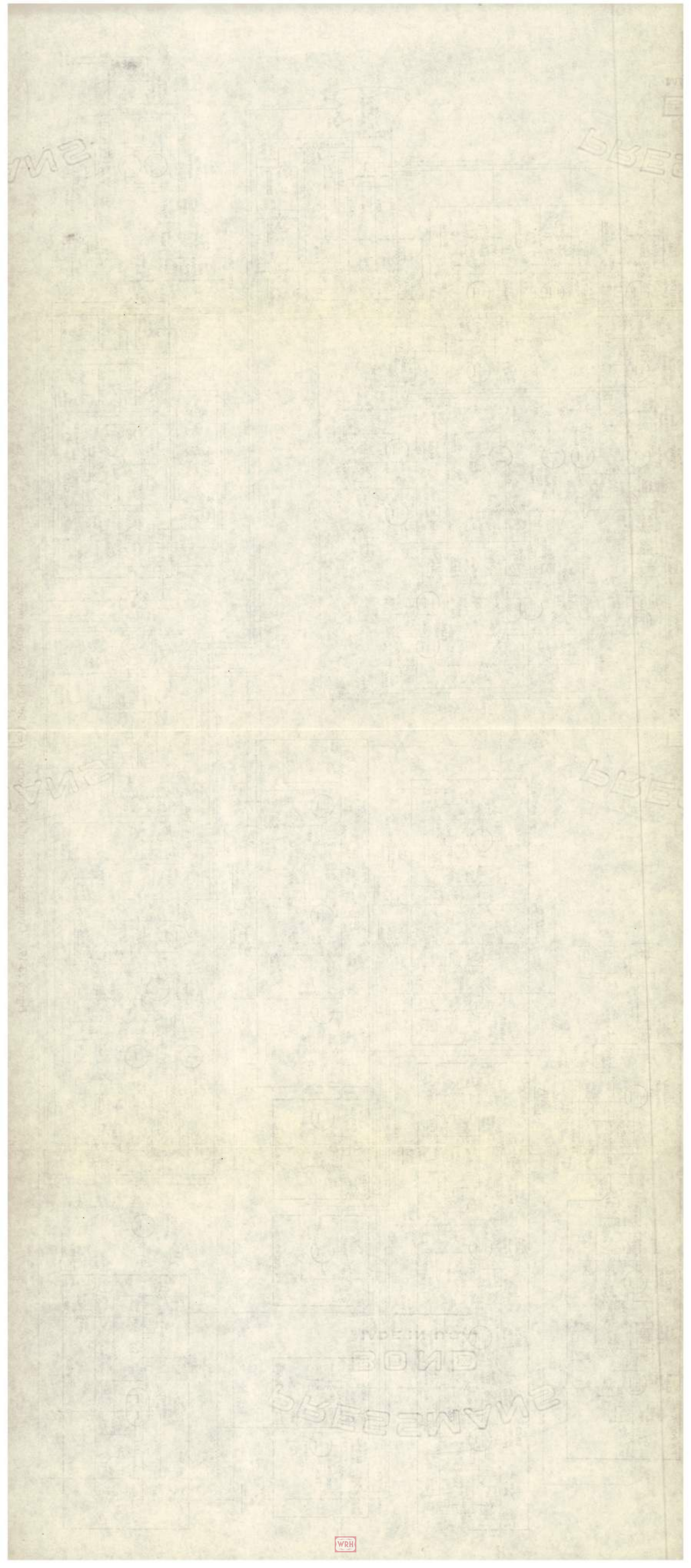


Figure 23-18. Power supply.



voltage supply, as well as the heater power, as shown in Fig. 23-18. The Z purity coil, whose function is to correct for axial parasitic magnetic fields, is energized as seen by discharging a 50-microfarad capacitor through the coil when the purity switch is depressed. Normally, the capacitor is part of the filter circuits.

23-16. Complete diagram

Except for the tuner, already illustrated in Fig. 23-2, the complete diagram of the color receiver is given in Fig. 23-19.

All the sections previously described will be easily identified. A number of them are actually separate units mounted on a printed-circuit subchassis.

The monochrome part of the receiver occupies the left-hand side of the diagram; the color part is confined to the right-hand side.

It is instructive and highly recommended to follow the various signals through the circuits and to identify the components and their functions.

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