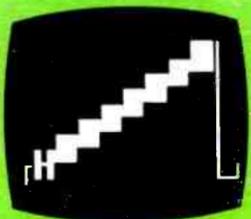
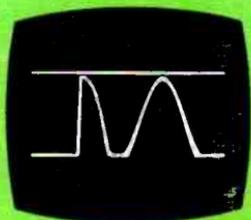


# Television Broadcasting Systems Maintenance



# **Television Broadcasting**

## **Systems Maintenance**

**by**

**Harold E. Ennes**

**Howard W. Sams & Co., Inc.**  
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## Preface to the Second Edition

In preparing this second edition, two major changes were made to the first edition. First, references to typical equipment have been updated, and, second, the maintenance of camera chains has been included. Thus the maintenance of signal sources except tape and disc sources has been added. Tape and disc equipment is covered in a separate volume, *Television Broadcasting: Tape and Disc Recording Systems*.

Three basic reference texts should be considered a prerequisite to this book. They are (1) *Workshop in Solid State*, (2) *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*, and (3) *Digitals in Broadcasting*, all published by Howard W. Sams & Co., Inc. Reference 3 is required for all digital applications in broadcasting, since this subject is not covered in other volumes of this series. Reference 2 is an absolute must before undertaking the theory and practice of equipment and system maintenance.

HAROLD E. ENNES

## Preface to the First Edition

Maintenance procedures for television systems recently have started to become standardized. While there are still many variations in techniques, these are steadily approaching systematized formal procedures. In this volume, the latest techniques are combined with a complete analysis of system malfunctions. A certain amount of fundamental systems theory has been included where it is considered important in maintenance practice.

The television system is treated as a collection and integration of units for selecting, combining, and transmitting video and audio signals. Comprehensive discussions are included for all units from the studio switcher inputs to the transmitter output. Signal sources and technical production facilities are covered in *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*, also published by Howard W. Sams & Co., Inc.

This volume covers not only the visual units of the telecasting system, but the aural portions as well. Major differences in the sound portions (as contrasted with standard fm systems), such as diplexed audio, stl's, and special considerations in measurement of the performance of the aural transmitter are treated accordingly.

The author is indebted to the following stations and manufacturers for important contributions to this book: WBAL-TV; WBBM-TV; WISH-TV; WTAE-TV; Dynair Electronics, Inc.; General Electric Co.; RCA Corp.; and Tektronix, Inc. Special thanks are due L. A. (Al) Pierce of WBBM-TV and Dana Pratt of RCA.

HAROLD E. ENNES

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## Establishing Measurements Standards

Television systems maintenance procedures have started to become more standardized only recently, partially as a result of universally accepted "standard" test signals (Chapter 2). Although there still are many variations in techniques, and we can make no revelation of "standard practice" in the field, we are steadily approaching systematized formal procedures.

Whatever specific technique is used, a "standard" starting point must be established for comparison. That is the basic purpose of this chapter. To emphasize this point, consider the measurement of corner-to-corner resolution of a camera with the aid of a studio test chart. It is perfectly valid for the operator to observe the test chart as reproduced on his monitor and make the adjustments necessary to obtain best corner focus consistent with good overall focus and shading, since he is concerned with a qualitative ratio rather than an absolute quantity. The maintenance department, however, is charged with the responsibility of maintaining performance between limits set at the lower end by FCC standards and at the higher end by the chief engineer of the particular station. This higher end is usually limited (as it should be) only by the performance specifications of the equipment installed. If corner resolution for a given center resolution and gray scale falls below normal, as compared to a value based on previous experience with a given pickup tube and camera, the maintenance engineer must first know the characteristics of the monitor he is using before a valid statement can be made (not considering other and more precise testing methods). It is entirely possible for a monitor to exhibit a difference of 50, 100, or more lines in resolving power (either plus or minus) between the corner and center of the raster.

This chapter is concerned with the importance of proper understanding and calibration of the oscilloscope, which becomes the primary "standard" of the maintenance department. Because of the predominant use of the Tektronix oscilloscope in stations across the country, this unit will be referred to most often in specific applications.

## 1-1. THE TV BROADCAST OSCILLOSCOPE

The television-broadcast, or "professional," oscilloscope differs in several respects from the more conventional service-type instruments. These differences may be summarized as follows:

1. *Horizontal (X-Axis) Sweep Calibration.* In the service scope, horizontal-sweep calibration is normally in terms of repetition frequency. In the professional scope, horizontal-sweep calibration is in terms of time per division. Thus we are using a direct unit of *time* for a given distance of spot travel across the screen, and this is termed the *time base*. For example, if the sweep is calibrated for one microsecond per division and one cycle of a sine wave occurs per division, the *frequency* is:

$$\frac{1}{1 \times 10^{-6}} = 1 \text{ MHz}$$

Therefore, although we can think of the time base in terms of relative sweep speed, we specify it by a term which is actually the reciprocal of speed: time per division. This permits direct measurement of the time between events or between portions of a waveform. For example, horizontal sync width is 4.8  $\mu$ s; horizontal blanking width is 11  $\mu$ s; "front porch" width is 1.6  $\mu$ s; etc.

2. *Vertical (Y-Axis) Bandwidth.* The tv service-type scope normally has a passband (to the 3-dB down point) of around 5 MHz. The tv broadcast scope, particularly that used at the studio, has a passband to 30 or 50 MHz. (This does not apply, however, to the Tektronix Type 524 scope which for many years was the broadcast "standard" scope, and which has a vertical-amplifier bandwidth of 10 MHz. For this reason, Section 1-2 is devoted entirely to special techniques required for the Type 524 scope when used with modern studio gear.)
3. *Retrace Blanking.* The sweep system in the service-type scope generates a *blanking* pulse to eliminate the beam retrace from right to left on the cathode-ray tube (crt). All Tektronix professional scopes employ an "unblanking" pulse which starts concurrently with the sawtooth deflection voltage and ends at

the conclusion of the linear portion of the sawtooth. The beam is extinguished at all times except for the duration of the unblanking pulse.

Fig. 1-1 presents a block diagram of a professional tv scope. The basic functions can be outlined by describing the waveforms as follows:

*Waveform 1.* This is the input waveform applied to the vertical amplifier.

*Waveforms 2 and 3.* The push-pull output of the vertical amplifier is fed to a signal delay line. The amplitude is determined by the setting of the volts/division control on the vertical amplifier. The purpose of the delay line will be pointed out later.

*Waveform 4.* When the sync switch is in the internal sync position, a sample of the input waveform (after amplification) is fed to a trigger circuit which generates this voltage spike at some selected point on the display waveform, determined by the setting of a trigger-level control on the front panel. This pulse is used to *start* the "run-up" portion of the time-base sawtooth. An external sync source may be used if desired.

*Waveform 5.* After a slight delay from the initial trigger caused by run-up multivibrators and saw-former processing, the time-base generator develops this sawtooth that has a rate of rise determined by the time/division control. Thus the waveform rises through a given number of volts each unit of time.

*Waveform 6.* The unblanking pulse is generated in the time-base generator coincident with the start of the run-up portion of the sawtooth and is applied to the grid of the crt. Thus the beam is switched on during its left-to-right travel and is switched off during right-to-left retrace.

*Waveforms 7 and 8.* Push-pull sawtooth waveform; the positive-going waveform is applied to the right-hand horizontal-deflection plate of the crt, and the negative-going sawtooth is applied to the left-hand plate. Thus the beam is swept to the right through a given number of graticule divisions during each unit of time as determined by the setting of the time/division control.

*Waveforms 9 and 10.* The push-pull delay-line signal output is applied to the vertical-deflection plates. Note from the timing diagram that even though the leading edge of the input waveform is used to actuate the trigger circuit, the time-base sawtooth and unblanking operations require a measurable time interval:  $t_1$  to  $t_2$  for triggering, and  $t_2$  to  $t_3$  for sweep and unblanking functions. In practice, this amounts to about  $0.15 \mu\text{s}$ .

Note from the timing diagram of Fig. 1-1 that a considerable portion of the leading edge of the pulse would be lost on the crt screen

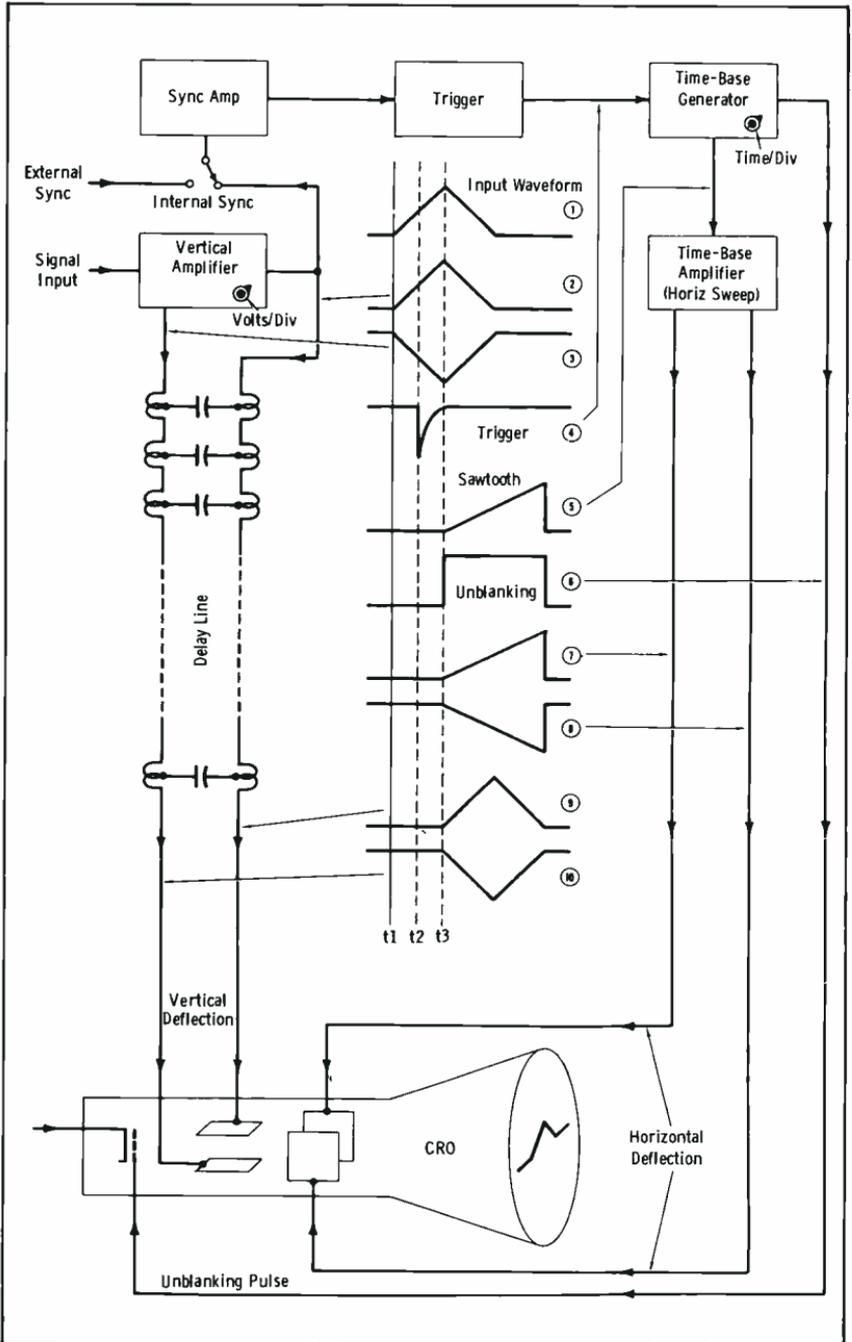
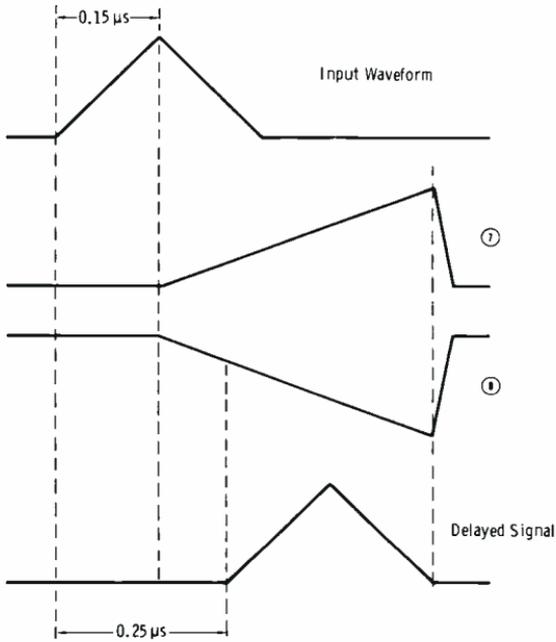
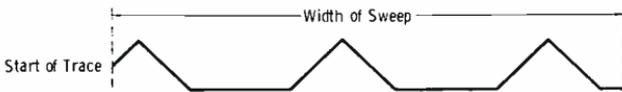


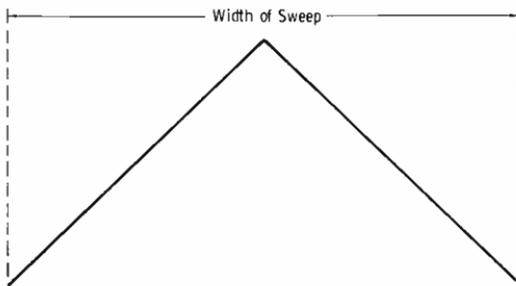
Fig. 1-1. Basic block diagram of professional tv oscilloscope.



(A) Signal and sweep waveforms.



(B) Undelayed display of pulse train.



(C) Display of pulse with delay.

**Fig. 1-2. Purpose of signal delay.**

without signal delay. Through the use of a signal delay of about  $0.25 \mu\text{s}$  (Fig. 1-2A), the entire leading edge of the input waveform is displayed. To completely understand this, observe Fig. 1-2B. If a delayed signal is not provided, the time base can be adjusted to display following pulses in a pulse chain for observation; only the start of the sweep is affected. But with signal delay (Fig. 1-2C), a single pulse can be presented completely across the entire sweep.

Keep in mind also the time-base concept. Note that a  $0.15\text{-}\mu\text{s}$  loss of trace would not be of any importance on a 60-Hz sine waveform, since  $0.15 \mu\text{s}$  is of no consequence in a waveform of  $16,667\text{-}\mu\text{s}$  duration. But for any pulse with a rise time approaching  $0.15 \mu\text{s}$ , this factor is quite important.

## 1-2. SPECIAL TECHNIQUES FOR THE TYPE 524 OSCILLOSCOPE

Although the Tektronix Type 524 oscilloscope is now obsolete, it still is often used as either a primary or an auxiliary test device. Many stations have retained it for use where the built-in marker frequencies provide an advantage (for example, in photographing horizontal-rate waveforms). Therefore, it is desirable to cover the use of the Type 524. This coverage will apply as well to the use of any 10-MHz scope.

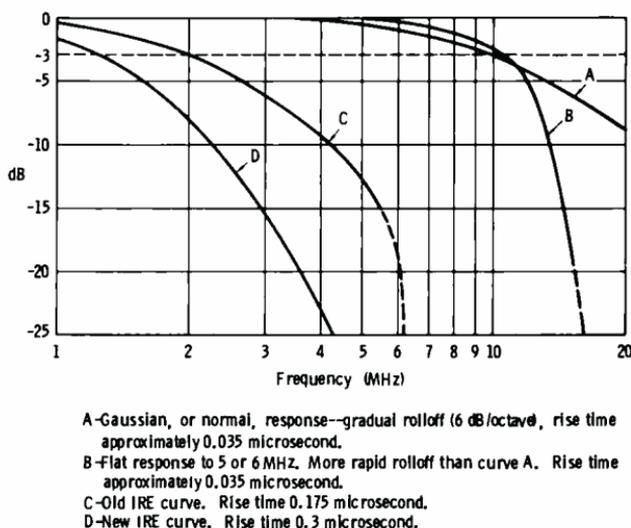
### The 524 Personality

Every oscilloscope has its own "personality" as observed on the crt. Getting acquainted with the characteristics of the individual scope is the initial step in calibration of the instrument.

The two basic types of information displayed by the cro are *amplitude* and *time*. *Waveshape* is not really a third basic type of information; it is simply the relationship between amplitude and time. From the interpretation of a waveshape, you will obtain low-, medium-, and high-frequency response; phase distortion; gray-scale response; and the various factors included in transient response. The amplifier within the scope itself becomes the standard which must be considered in measurements.

Fig. 1-3 shows four response curves which concern the user of the Type 524 scope. The Gaussian curve (A) has a rolloff suited for best transient response. This curve might be recognized by users of the Tektronix 524AD scope as being that obtained with the RESPONSE switch in the Normal position. (This scope has three switchable responses: normal, flat, and IRE.)

The flat response curve (B) has a relatively flat response to 5 or 6 MHz. Since the gain-bandwidth product has a fixed upper limit, the rolloff becomes more rapid above this value.



**Fig. 1-3. Oscilloscope response curves.**

The old IRE response curve (C) was originally adopted for comparative level checks. The new IRE curve (D) adopted in 1958 is more indicative of true luminance levels.

In general the applications of the various response curves are as follows:

1. *Normal (Gaussian rolloff)*. Most suitable for waveform analysis, particularly where transient response becomes a major factor.
2. *Flat (to 5 or 6 MHz)*. Most suitable for single-frequency response runs (or keyed sine-wave burst signal) to avoid a scope correction factor in the readings.
3. *IRE*. Most suitable for checking, comparing, and adjusting amplitude levels. Some existing equipment, such as scopes and master monitors with an IRE position, use the old curve. To avoid the inevitable arguments resulting from various interpretations of peaks of the higher-frequency signal components, the new curve should be adopted as soon as possible. This is of prime importance in color telecasts where luminance levels are critical.

### The Bandwidth–Rise-Time Product

Engineers are quite familiar with the gain-bandwidth product of an amplifier. Of more importance to the user of any given gain-bandwidth amplifier is the bandwidth–rise-time product, since this becomes his standard of measurement.

This relationship is stated as follows:

$$(BW)(RT) = k$$

where,

BW is the bandwidth in megahertz (to the  $-3$ -dB point),

RT is the rise time in microseconds (measured between 10 and 90 percent of peak value),

k is a factor between 0.3 and 0.5, depending on the type and amount of high-frequency compensation.

The limit of factor k is that the overshoot on the leading edge of a pulse must be less than 3 percent. In fact, a system has an equivalent bandwidth and rise time only within the limits of 3-percent overshoot.

The most typical value for k is 0.35, and the equation may be expressed in three possible ways:

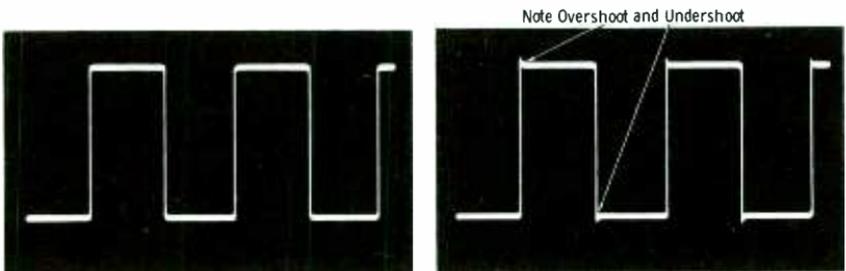
$$(BW)(RT) = 0.35$$

$$(BW) = 0.35/(RT)$$

$$(RT) = 0.35/(BW)$$

Table 1-1 lists rise times for bandwidths from 1 to 10 MHz within the preceding limitations. In Fig. 1-3, it may be noted that the  $-3$ -dB point of either the normal or flat curve falls in an area which safely indicates a bandwidth of 10 MHz. It can be shown from pulse theory that rise time is proportional to the *area* under the amplitude-frequency response curve; hence, changing from one response to the other does not appreciably affect the rise time. Figs. 1-4A and 1-4B illustrate the difference in overshoot of a 75-kHz square wave (rise time  $0.02 \mu\text{s}$ ) with the Type 524 scope set for normal and flat response, respectively.

It is the shape of the curve that actually is being changed when video peaking coils are adjusted. Leading and trailing transients of a rapid transition in picture content must be adequately controlled



(A) Normal response.

(B) Flat response.

Fig. 1-4. Response of Type 524 scope to square waves.

**Table 1-1. Bandwidth and Rise Time;  $k = 0.35$ ,  
Overshoot Under 3%**

BW (Megahertz)	RT (Microseconds)
1	0.35
2	0.175
3	0.1166
4	0.0875
5	0.07
6	0.058
7	0.05
8	0.0437
9	0.039
10	0.035

by the maintenance personnel. Hence, complete familiarity with the scope amplifier characteristic is necessary.

A good square-wave generator with reasonably short rise time and a flat-topped response curve completely free of wrinkles is desirable. The Tektronix Type 105 generator with a rise time of 0.02 microsecond is one example. It is important to remember, however, that to measure the exact rise time of a pulse, the vertical amplifier of the scope must have a rise time of no more than one fifth that of the pulse to be measured. The rise time of a scope with a 10-MHz bandwidth is about 0.035 microsecond. The specified rise time of this square-wave generator is 0.02 microsecond. To measure this exact rise time, it would be necessary for the scope amplifier to have a rise time of 0.004 microsecond or less. A pulse with a short rise time is necessary for transient-response checks in terms of overshoot or undershoot.

**NOTE:** Although the square wave has limited application in modern systems maintenance, its basic characteristics provide a foundation for more advanced understanding of system waveforms and interpretation. The scope itself is checked for frequency and transient response with a square wave.

The total rise time of a pulse through a series of cascaded stages is equal to the square root of the sum of the squares of the individual stage rise times (assuming overshoots of less than 3 percent). When, for example, an amplifier with a rise time of 0.02 microsecond is feeding an amplifier with a rise time of 0.04 microsecond, the total rise time is:

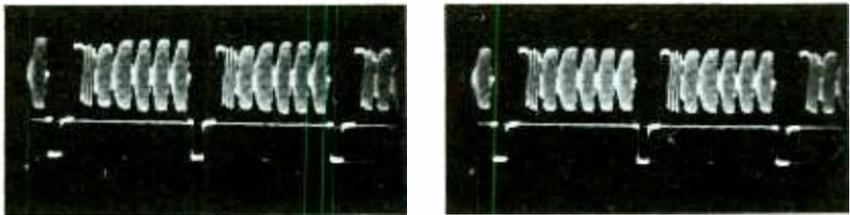
$$RT_t = \sqrt{(0.02)^2 + (0.04)^2} = \sqrt{0.002} = 0.045 \mu s$$

Understanding this relationship will enable the maintenance engineer to estimate closely the condition of his test equipment, even

though an extremely wideband scope is not available, provided he is certain of the scope characteristics. It also emphasizes the better-known premise that an amplifier output must be compared directly to the scope display at the amplifier input, properly terminated, rather than any assumed condition. Each time the test signal is transferred to another stage or amplifier with different cabling, capacitances, etc., it is important to check the input display at the point of connection so that the output can be interpreted properly in terms of the actual input signal.

### Scope Probes

Due to capacitive loading effects, the direct scope probe is severely limited in application to tv equipment maintenance, even when applied directly across 50- or 75-ohm terminations. A direct probe should *never* be used where frequency response or transient response is a factor; therefore, it is limited to certain applications where the IRE response is used. Fig. 1-5A shows the display of a



(A) Direct probe.

(B) 10:1 probe.

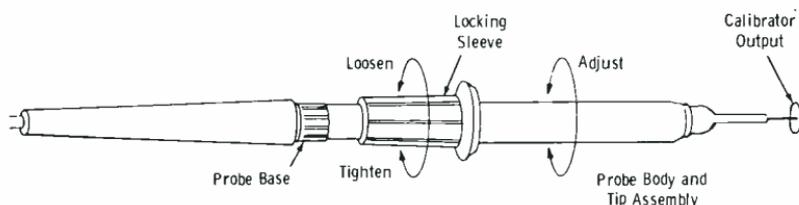
**Fig. 1-5. Scope response to keyed sine-wave burst.**

keyed burst signal taken with a direct probe across a 75-ohm termination. Actually, the display obtained depends on the length, type, and condition of the cable used; one probe could show a decided rolloff of higher frequencies, whereas another could indicate a rolloff at low frequencies. Similarly, pulses would have varying rise time and overshoots, depending on the duration and repetition rates. Fig. 1-5B is the display obtained with a 10:1 isolation probe across the same termination. Use the direct probe only where IRE response is used.

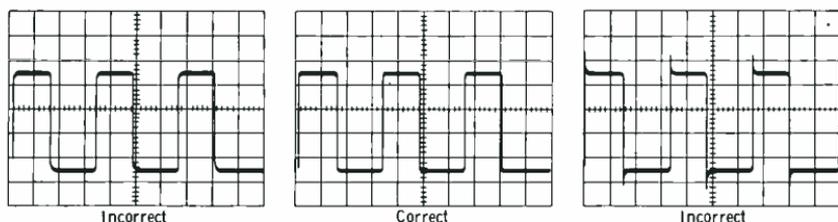
For most applications, the 10:1 capacitance divider probe should be used. For a scope with an input impedance of 1 megohm shunted by 40 pF, the simplest 10:1 probe consists of a series-connected 9-megohm resistor shunted by a trimmer capacitor of 3 to 12 pF. When the probe is connected to the scope, the input impedance from the probe tip becomes 10 megohms shunted by approximately 12 pF. The trimmer capacitor is adjusted so that the RC product of the

probe is equal to the RC product of the scope input, thus making the voltage division independent of frequency. This is done by touching the probe to the scope calibration-pulse output or a square-wave generator set to about 1 kHz, and adjusting the trimmer so that the leading edge of the pulse is not rounded on the top (undercompensated) and does not overshoot (overcompensated) (Fig. 1-6). This adjustment should be checked often and must always be checked when the probe is used with a different scope, even though it is one of the same make and model.

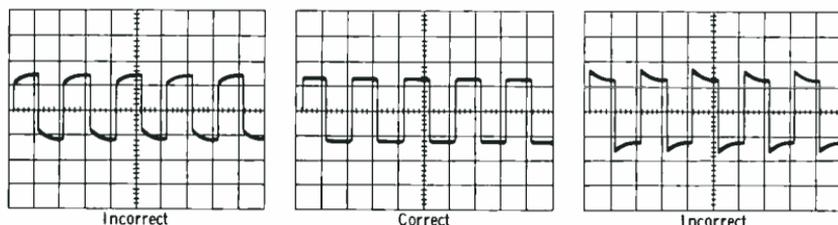
An important point to remember when using the 10:1 probe is that a preamplifier normally is used ahead of the main vertical amplifier in such scopes as the Tektronix Type 524AD. This is necessary since most performance tests are made on standard 1-volt peak-to-peak signals across 75 ohms, and the 10:1 voltage division makes the extra gain necessary. It is, therefore, imperative that single-frequency



(A) Probe adjustments.



(B) Waveforms from line-frequency scope calibrator.



(C) Waveforms from 1-kHz scope calibrator.

Courtesy Tektronix, Inc.

**Fig. 1-6. Compensation of Tektronix Type P6006 probe.**

response runs and square-wave response tests of the scope be made at 1-volt levels in 75-ohm terminations. This is so that the scope attenuator settings are the same as when equipment checks are made.

### Calibration Procedures

A logical step-by-step calibration of the Type 524 scope can be outlined as follows.

1. *The Square Wave*—The point of emphasis at this time is two-fold:

- A. To ensure that the square-wave generator and the scope are performing within specifications.
- B. To establish the back-to-back response of the square-wave generator and the scope. Thus, when the generator is feeding the system and the scope is looking at the system output, the standard against which the measurement is made has been established.

NOTE: In recent years, new prefixes have been added to the old list of standard prefixes. We will be using these where applicable in this book. Table 1-2 lists the new prefixes along with some of the old ones.

Table 1-2. Prefixes

Prefix	Definition	Power of Ten Multiplier
atto	millionth of millionth of millionth part	$10^{-18}$
pico*	millionth of one millionth part	$10^{-12}$
micromicro**	millionth of one millionth part	$10^{-12}$
nano	thousandth of a millionth	$10^{-9}$
micro	millionth of one part	$10^{-6}$
milli	thousandth of one part	$10^{-3}$
centi	hundredth of one part	$10^{-2}$
deci	tenth of one part	$10^{-1}$
deka	ten	$10^1$
hecto	one hundred	$10^2$
kilo	one thousand	$10^3$
mega	one million	$10^6$
giga	one billion	$10^9$
tera	one trillion	$10^{12}$

\* Preferred prefix.

\*\* Obsolete prefix.

Assume that we have a Tektronix Type 105 square-wave generator, with specified 20-nanosecond rise time, and a Tektronix Type 524 scope, with specified 35-nanosecond rise time. (Incidentally, this means that the scope RESPONSE switch must be in the Normal position, *not* the IRE position.) What rise time should you measure

with this combination (Fig. 1-7)? Set the pulse repetition rate at 100 kHz.

$$RT = \sqrt{20^2 + 35^2} = \sqrt{1625} = 40 + \text{ns, or } 0.04 \mu\text{s}$$

If the combination of square-wave generator and scope meets this specification with no more than 3-percent overshoot, you are certain that both units are in proper operating condition. The probe is not used in this test.

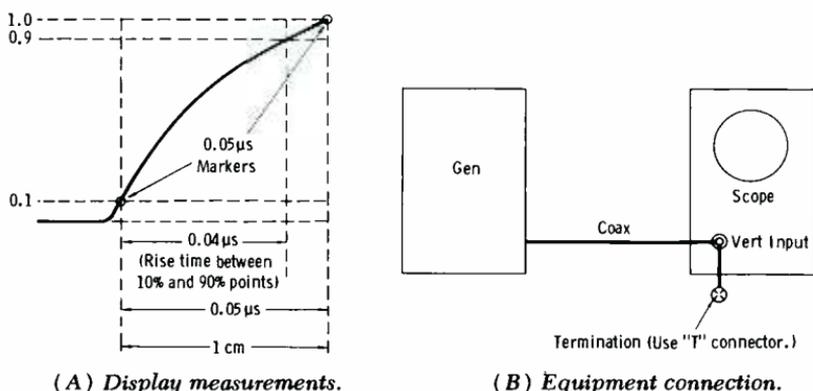


Fig. 1-7. Measurement of rise time.

Assume you do not have overshoot but your measurement indicates a rise time somewhat greater than  $0.04 \mu\text{s}$ ; which is at fault, the generator or the scope? The long way to go about this is to run complete sweep-response checks on the scope. There is a short-cut method that may be used if the square-wave generator is capable of putting out at least 15 volts (peak-to-peak) in 75 ohms. The Tektronix Type 105 is capable of doing this. NOTE: In accordance with specifications, it may be necessary to use a 93-ohm cable with 93-ohm termination. Although this is a manufacturer's specification, all of these units the author has used will reach a 15-volt amplitude in 75 ohms. You do not lose rise time unless you go to a higher impedance than 93 ohms.

See Fig. 1-8. This method involves making a direct connection from the generator to the top vertical deflection plate of the crt in the Type 524 scope; the vertical amplifier is bypassed. The resulting vertical deflection sensitivity is approximately 15 volts/centimeter.

- A. The original test cable which comes with the Type 105 consists of a short piece of 93-ohm cable with a UHF connector on one end and a 93-ohm termination on the other, which incorporates banana jacks. Use this cable if it is available; if it is not, make up a cable as shown in Fig. 1-8.

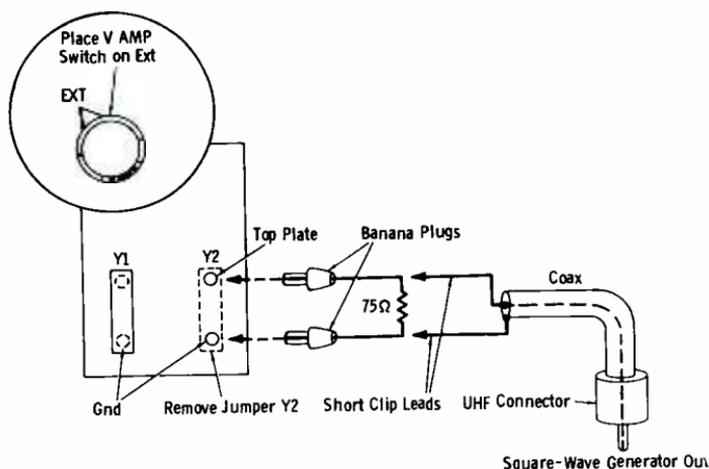


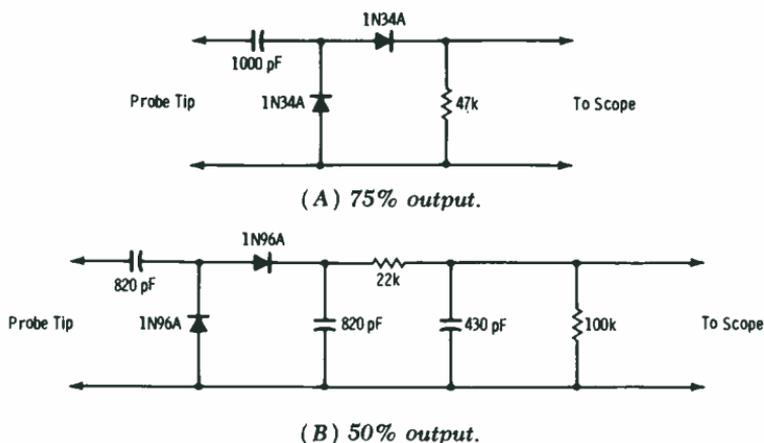
Fig. 1-8. Method of direct connection to crt.

- B. Remove jumper Y2 from the rear panel of the scope. Place the v AMP switch in the Ext position. This capacitance-couples (with an internal blocking capacitor) to the crt top deflection plate, and connects 1-megohm resistors to the vertical amplifier to retain the function of the vertical-position controls.

The above procedure provides a low-capacitance connection directly from the generator to the crt. With the generator amplitude control in the maximum clockwise position, you should get a 1-centimeter deflection on the cro. Expand the time base to  $0.05 \mu\text{s}/\text{cm}$  or greater. You can now measure the exact rise time, which for the Type 105 should be  $0.02 \mu\text{s}$ . Regardless of the type of square-wave generator you have, you can use this procedure by correlating your manufacturer's specifications with the techniques described.

If you follow this procedure, you know in which unit (generator or scope) trouble exists. The Tektronix scope itself is checked and aligned with a square-wave generator. If changing tubes (in whichever unit is at fault) and going through the manufacturer's instructions fail to correct the problem, never hesitate to call on the manufacturer's field service. Broadcast-type test equipment is highly specialized. Remember that broadcasting is a federally regulated service, and the FCC does not excuse faulty test equipment in proof-of-performance failures. Your basic test equipment *must be standardized*.

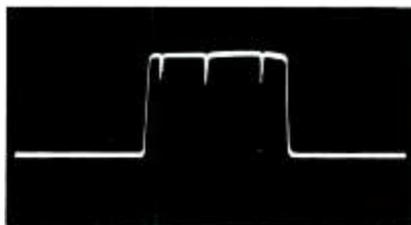
2. *Video Sweep (Detected)*—Terminate the video sweep generator directly at the generator output connector in 75 ohms. Use a video detector probe (Fig. 1-9). The probe in Fig. 1-9A delivers approximately 75 percent of the actual peak-to-peak signal, and the higher-isolation probe in Fig. 1-9B delivers about 50 percent. Adjust



**Fig. 1-9. Peak-to-peak video-sweep detector probes.**

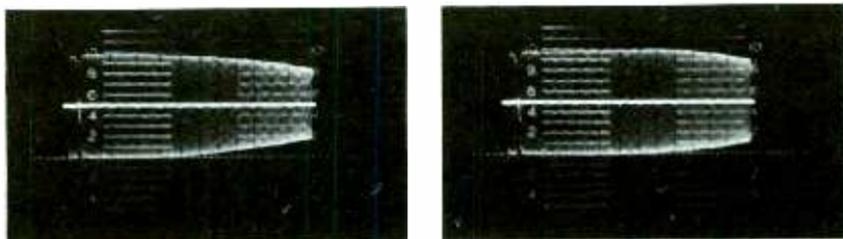
the output amplitude for 1 volt, which will give a reading of approximately 0.75 volt with the probe of Fig. 1-9A or 0.5 volt with the probe of Fig. 1-9B. Also, adjust the scope gain to provide a convenient display (Fig. 1-10). This permits checking the flatness of the sweep generator itself, since the detected sweep envelope does

**Fig. 1-10. Detected video-sweep markers at 1, 5, and 10 MHz.**



not depend on the high-frequency response of the scope. The Tektronix scope may be used on any response position; or a scope very limited in response can be used, provided it has reasonably good low-frequency square-wave response. If the video sweep generator cannot be made perfectly flat, as observed on the scope, the deviations must be plotted as a correction factor for equipment checks and scope calibration.

3. *Video Sweep (Wideband)*—(This should be observed only after determining the flatness of the sweep generator as in Step 2.) Although this method is used only in very special cases (and with extreme care), the rf envelope may be observed directly without detection as a “quickie” check on scope-amplifier response (Fig. 1-11). This check, however, is valid only if the probe to be used for



(A) Scope on Normal response.

(B) Scope on Flat response.

**Fig. 1-11. Typical undetected video-sweep display (to 10 MHz), 524 scope.**

equipment checks is used on the scope and a signal of the same amplitude is employed so that the scope compensated attenuator is at the same setting as that to be used. It is good engineering practice to run these checks with all probes in stock, and through the scope preamplifier as well as directly to the vertical-amplifier input. Use varying levels from the sweep generator to permit use of convenient scales on the scope with different attenuator settings. This will pinpoint any attenuator position that might be incorrectly compensated. An attempt to employ correction factors for different attenuator settings becomes both cumbersome and inaccurate in system measurements. Normally there will be some correction factor when using the preamp and when feeding the vertical amplifier directly. Plot these responses either on a graph or by tabulation in peak-to-peak values. Normally the detector probe is employed when video sweep is used. The wideband display provides a quick check of scope response to single-frequency sine waves or in similar applications, such as keyed sine-wave bursts (Fig. 1-5) which normally go up to about 4.2 MHz (described in Chapter 2).

**4. Single-Frequency Sine-Wave Checks**—To confirm the scope frequency-response curve, we must employ single-frequency runs over the spectrum concerned, normally out to 10 MHz. In the above steps we have measured the flatness of the sweep generator itself, and made an initial observation of the scope-amplifier response in the normal and flat modes of operation. The use of single-frequency sine waves will now confirm the video-sweep response from a different source, the sine-wave generator. The purpose here is to assure that the scope will show identical frequency and transient response (rolloff characteristic) regardless of the slight differences in the complex source impedances of various test generators.

If your sine-wave signal generator employs a 600-ohm internal impedance, you must make special provisions to feed the normal 75-ohm inputs of the station gear. For example, the Hewlett-Packard Model 650A test oscillator is such a device. This particular unit is provided with a "low-impedance" voltage-divider cable which sup-

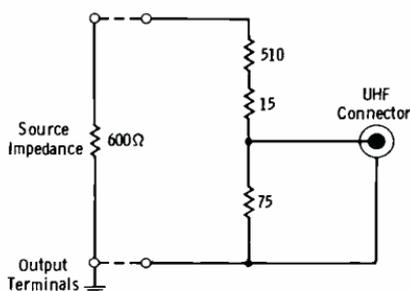
plies the signal across a 6-ohm resistor, and a 75-ohm load can be fed from this with no effect on frequency or transient response. However, it is often desirable to obtain a 1-volt peak-to-peak signal for test purposes, and it is not possible to obtain this amplitude across 6 ohms (from the particular generator discussed).

The best way to modify this type of generator for video use is shown in Fig. 1-12. Note that the total resistance is 600 ohms; the UHF connector (which can be mounted on the front panel) is across 75 ohms of this total. The UHF connector must feed a 75-ohm termination.

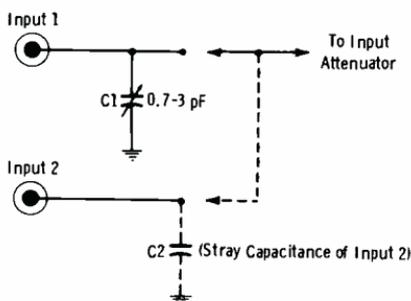
When you plot the frequency response of your scope from this source, you should obtain a curve which indicates less than 2 percent difference from the video-sweep response indication. If this result is not obtained, your system adjustments will not agree between various signal sources.

NOTE: The Hewlett-Packard Model 650A was replaced by the Model 653A. This test oscillator provides either a 75-ohm unbalanced or 124-ohm balanced sending-end impedance. It can furnish up to +11 dBV across 75 ohms, where 0 dBV = 1 V p-p.

The input attenuator of the Tektronix Type 524 scope is a compensated network so that the input time constant is made equal for all positions of the VOLTS/CM switch to get the same frequency and transient response. In later models (Type 524AD), there is also



**Fig. 1-12. Modification of Hewlett-Packard Model 650A test oscillator for 75 ohms.**



**Fig. 1-13. Swamping capacitance in input circuit of Tektronix Type 524AD oscilloscope.**

a small trimmer capacitor directly across input 1, but none across input 2. See Fig. 1-13. Let C2 represent the stray capacitance across input 2, which has a longer run to the input attenuator than does input 1. The small trimmer across input 1 is adjusted in practice to equal C2. This is done as follows:

1. With the input selector set on position 2, and with the 10:1 attenuator probe connected to this input (VOLTS/CM switch on the 0.15-0.5 position), connect the probe to the terminated output of a fast-rise-time square-wave generator. Adjust the probe compensating trimmer to get a flat top without overshoot, with the scope response switch set in the normal position.
2. Step 1 has correctly compensated the probe. Now connect this probe to input 1 and set the selector switch to position 1. Looking at the same square wave with the standardized probe, adjust the trimmer across input 1 in the scope to get the same correct square-wave response.

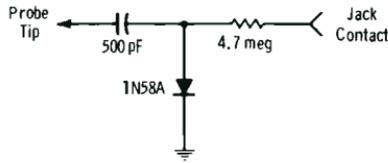
Actually, the input capacitance represented by C2 and matched by C1 serves as "swamping capacitance" in conjunction with the compensating trimmers and resistors for each position of the input attenuator (VOLTS/CM switch). If the RC networks are not correct, as could happen if any of the resistors should change value for any reason, the sending-end complex impedance will influence what your scope tells you considerably. For example, if the square-wave transient response is different for different settings of the VOLTS/CM switch, the compensating circuits are definitely in need of adjustment. Follow your specific instruction manual carefully in the "front-end" alignment.

However, it is possible that you can arrive at what seems to be correct input-attenuator compensation, and still not get close correlation between the scope response indicated by video sweep and the response indicated by single-frequency sine-wave measurements. This can be caused by either or both of the following:

1. One or more resistors in the input attenuator may have changed just enough that adjustment of associated trimmers in the network for a particular position of the attenuator will appear to pass the square wave properly, but the RC ratio is not optimum. This is usually indicated if it is necessary to remove the small trimmer (usually 0.7-3 pF) across input 1 to "match" input 2, or, conversely, if more capacitance is needed across this trimmer.
2. The output-level meter of the sine-wave generator may not be "flat" across the video spectrum to 10 MHz.

Few stations have a wideband vtm with known response to 10 MHz with which to check the internal output-level meter of a signal generator. Fig. 1-14 shows the schematic of a conventional rf detector probe for use with vtm's of 10-megohm input resistance. This circuit develops a dc output voltage proportional to the rms value of a sinusoidal waveform. For example, a reading of 1 volt dc indicates an rms value of 1 volt. (To convert to other values, see Table

**Fig. 1-14. Example of rf probe for use with vtvm.**



1-3.) The circuit shown will extend the response of a meter to well beyond that required for the video spectrum of 10 to 20 MHz. Use it as follows:

1. Plug the regular vtvm probe into the tip jack of the detector probe, and set the AC-OHMS/DC switch of the vtvm probe to the ac position. This bypasses the 1-megohm resistor in the vtvm probe, which is normally in series with the 10-megohm input resistance of the meter, and provides a "direct connection" to the vtvm through the detector probe. If desired, you can make the detector probe an entirely separate item.
2. Set the vtvm meter switch to the -dc position. Connect the detector probe to the terminated output of the sine-wave generator.
3. Adjust the signal-generator amplitude to obtain a reading of 0.25 volt on the vtvm at a reference frequency of 200 kHz. Note the reading of the output-level meter on the generator. Now repeat this procedure at spot frequencies (those you will use, for example, on proof-of-performance runs), adjusting the generator amplitude (if necessary) to maintain the 0.25 volt. Compare this level to the output-level-meter reading at each frequency, and record any difference in reading. This is the correction factor to use when feeding equipment during tests.
4. For example, if the generator meter reads -0.5 dB (relative to the reference-frequency level) at 5 MHz, and the external vtvm shows no deviation from reference level, then you know that at 5 MHz you must set the generator amplitude 0.5 dB lower than you did at 200 kHz, to maintain the same actual output level.

**Table 1-3. Sine-Wave Multiplying Factors**

Given Value	Multiplying Factor to Get			
	Average	RMS	Peak	Peak-to-Peak
Average	—	1.11	1.57	3.14
Rms	0.9	—	1.414	2.828
Peak	0.637	0.707	—	2.0
Peak-to-Peak	0.32	0.3535	0.5	—

5. This also holds true when running the scope response with spot-frequency sine waves to compare with the video-sweep indication of response. Never try to have the scope connected to the generator output at the same time as the vtvm, since the added capacitance of the vtvm arrangement will upset the scope response. *Always* calibrate the generator meter first. If your signal generator does not have an internal meter, you must set the level with the vtvm first for each frequency, then feed the scope and remove the meter. Generators which employ a built-in output-level meter isolate the meter from the output terminals by means of a known attenuator pad.
6. If your generator has a means for calibrating the frequency response of the output-level meter, by all means do so in accordance with the instruction manual. This will contribute immensely to your confidence in measurements, and it obviously permits a great saving of time by eliminating the necessity to correlate correction factors

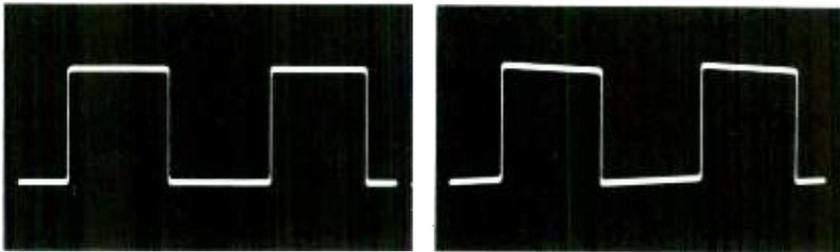
Remember that if the vtvm reading is in rms volts only, to get the peak-to-peak value multiply the rms reading by 2.828. Thus the 0.25 volt used in the above procedure results in an actual peak-to-peak value of 0.7 volt. Conversely, to get the rms value of a peak-to-peak voltage, multiply the peak-to-peak reading by the reciprocal of 2.828, which is 0.3535. (See Table 1-3.)

When your scope and test generators are standardized as we have been discussing thus far, the scope will indicate a reliable story of system performance regardless of the signal source. You are ready for tests without the many pitfalls that so often occur due to lack of attention to these extremely important and necessary details in the use of an oscilloscope with 10-MHz response.

5. *Low-Frequency and Transient Response*—Determine the rise time and percent of overshoot of the square wave as read on the scope, both through the preamp and the main amplifier, at the frequencies normally used. Unless a generator with short rise time is available, higher-frequency square waves (above 75 kHz) are not particularly useful, because for response checks at the higher frequencies the rise time of the pulse must be less than the rise time of the amplifiers to be checked. Keep in mind the discussion associated with Fig. 1-4. A 60-Hz square wave fed to the Tektronix Type 524AD (dc position) should have an absolutely flat top, as shown in Fig. 1-15A. Fig. 1-15B shows the normal amount of tilt introduced by the input coupling capacitor when the scope is in the ac mode of operation. Remember that the last two (highest) gain positions of this scope are ac only, since the preamp is used on these positions. An adjustable grid time constant (low-frequency compensation con-

trol) is used in the preamp, which should be adjusted according to the manufacturer's instructions. On any scope employing either external or plug-in preamps, always include these units for all scope calibration procedures.

It is possible for a wideband scope amplifier to exhibit a leading-edge overshoot as a result of a vacuum-tube defect known as cathode interface. This low-frequency phase shift results from series-resist-



(A) Scope in dc mode.

(B) Scope in ac mode.

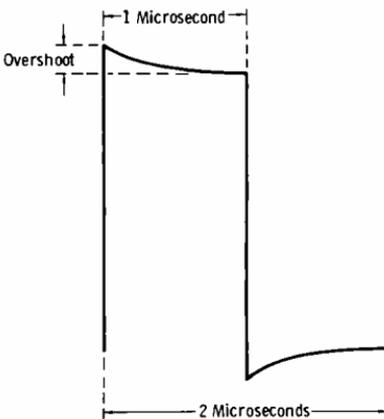
**Fig. 1-15. Response patterns for 60-Hz square wave.**

ance and capacitive-bypassing effects of a chemical interface layer that forms between the sleeve and oxide coating of the cathode. Since some tubes have been known to develop this characteristic in less than 500 hours of operation, the scope should be checked about every two months for this type of tube defect. The procedure is as follows:

- A. Adjust the frequency of the square-wave generator to 500 kHz. The waveform should have a rise time of 0.2 microsecond or less.
- B. Adjust the time base so that several cycles of the square wave are displayed. If an overshoot with a duration of 0.2 to 0.6 microsecond appears (Fig. 1-16), one or more tubes in the vertical amplifier may have cathode interface. (Overshoot duration, or time constant, is the time required for the overshoot to decay to the final flat-top value.) A 500-kHz square wave completes one cycle in 2 microseconds; thus it has a pulse width of 1 microsecond, as shown in Fig. 1-16. The overshoot duration normally is between 20 and 60 percent of the total pulse width when cathode interface is present. As a double check, plug the scope into a variable autotransformer and increase the line voltage to the upper limit allowed. If cathode interface is present, the increased tube heater voltage will reduce the overshoot, and a decrease of line voltage will increase the overshoot. When this occurs, it is best to replace

all tubes in the vertical amplifier with new ones; then substitute the old tubes one at a time while observing the square wave. Discard any tube that tends to show this effect. Leave the line voltage at the lower limit (105-108 volts) to emphasize the effect of cathode interface.

6. *The Type 524 Time Base and Markers*—The Tektronix Type 524 scope has quite accurate internal marker generators which are very useful in certain waveform measurements, particularly at the horizontal rate. More recent professional tv oscilloscopes do not have this feature.



**Fig. 1-16. Square-wave distortion caused by cathode interface.**

In view of the importance of pulse frequencies and pulse durations in telecasting, it is rather startling to realize that a great many stations invest in expensive oscilloscopes with no thought whatsoever of a secondary frequency standard. Actually, in many applications the scope is worth only as much as its accuracy, and a suitable secondary frequency standard is an indispensable partner. The accuracy of the time base and the marker generators (where used) should be checked at least once a year, or whenever it appears necessary. For example, if two scopes are available, they should be checked against each other for any discrepancy.

The most convenient standard from an operational point of view (and for scopes with sweep calibrated in time rather than frequency) is a Tektronix Type 180A or the later Type TG 501 marker generator. When this unit is available, or when it can be rented or borrowed, the sweep can be calibrated in a minimum of time and with maximum accuracy by following the instructions. Otherwise, an oscillator of known accuracy, such as the Hewlett-Packard Model 650A or the later Model 3320 (digital readout), the Signal Corps

BC-221, or any good crystal standard, may be used. Obviously, the secondary standard itself should be checked occasionally, and for this purpose a WWV receiver will be found invaluable (see following sub-subsection 7).

Most heterodyne frequency meters employ a stable crystal oscillator which is used for calibrating the frequency of a variable oscillator. (The crystal normally includes a trimmer capacitor for WWV calibration.) This crystal oscillator produces harmonics which permit calibration of the test equipment at various frequencies. The points of calibration are termed *crystal check points*, and the frequencies at which they occur are given (usually in colored type) in a calibration book, which is also used to interpolate the dial reading. Fig. 1-17 shows a typical setup for calibrating the frequency meter. Assume that the calibration book shows a crystal check point at a frequency of 2000 kHz. The dial setting of the meter is adjusted to correspond to the number given in the calibration book for 2000

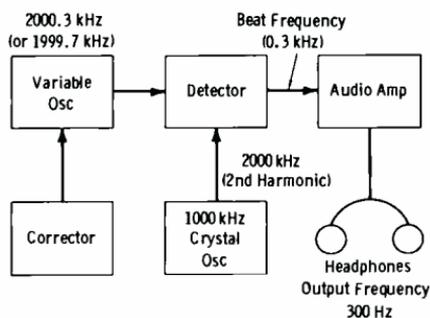
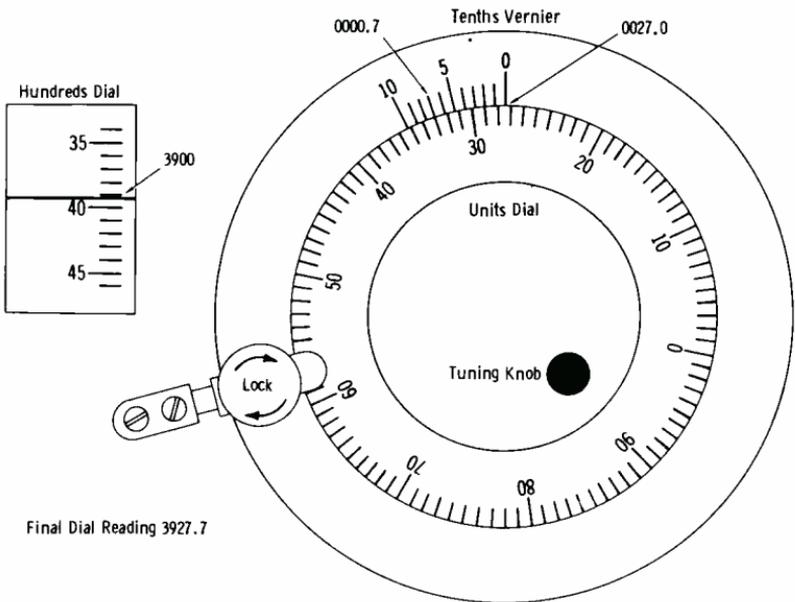


Fig. 1-17. Typical setup for calibrating a heterodyne frequency meter.

kHz. With the crystal switch thrown to the on position, the output of the variable oscillator beats with the output of the crystal oscillator; if there is a difference between the two frequencies, a beat note (within the audio range is sufficiently close) is produced. In Fig. 1-17 the frequency of the beat is 300 Hz, indicating that the frequency of the variable oscillator is 300 Hz higher or lower than the crystal frequency. The variable oscillator is now adjusted to the exact frequency by means of the corrector knob, which is adjusted until zero beat is obtained.

Fig. 1-18 represents the dial of a typical heterodyne frequency meter. The dial setting shown is read in the following manner: The long, thin line marked on the window of the hundreds dial indicates the approximate reading of the dial. Since it is between 3900 and 4000, the exact dial reading must be between these numbers. The reading on the units dial is read directly below the zero on the tenths vernier; it is between 27 and 28. To obtain the exact reading to the nearest tenth, the tenths vernier must be read. The tenths

value is obtained by finding the line on the vernier scale which most nearly coincides with a line on the units dial. The value of 0.7 coincides with 33 on the units dial; therefore, the exact reading for the dial setting shown in Fig. 1-18 is  $3900 + 27 + 0.7$ , or 3927.7. The frequency corresponding to this number must be obtained from the calibration book for the frequency meter.



**Fig. 1-18. Dial of a typical heterodyne frequency meter.**

Whenever the observed dial setting falls between two consecutive dial settings listed in the calibration book, it is necessary to interpolate to find the exact corresponding frequency. The dial reading shown in Fig. 1-18 lies between the numbers 3925.5 and 3927.9, as taken from the calibration book, and the frequencies corresponding to these dial settings are 3669 and 3670 kHz. The difference between the listed dial settings is to the difference between the dial reading and the next higher listed dial setting as the corresponding difference between the listed frequencies is to the difference between the unknown frequency and the next higher listed frequency. This is shown in the formula:

$$\frac{2.4}{0.2} = \frac{1 \text{ kHz}}{x \text{ kHz}}$$

These differences are easily found by employing a simple tabulation scheme as follows:

$$\text{Diff 1 kHz} \left\{ \begin{array}{l} \text{FREQUENCIES} \\ 3669 \text{ kHz} \\ \text{Unknown kHz} \\ 3670 \text{ kHz} \end{array} \right\} \text{Diff } x \text{ kHz}$$

$$\text{Diff 0.2} \left\{ \begin{array}{l} \text{DIAL SETTINGS} \\ 3925.5 \text{ (Listed)} \\ 3927.7 \text{ (Dial reading)} \\ 3927.9 \text{ (Listed)} \end{array} \right\} \text{Diff 2.4}$$

Solving for  $x$  in the formula gives a frequency difference of 0.083 kHz. This figure is then subtracted from the higher listed frequency, producing 3669.917 kHz as the frequency corresponding to the dial reading. The last two digits can be discarded for all practical purposes, giving a reading of 3669.9 kHz.

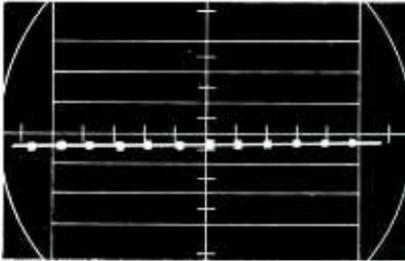
When a suitable frequency meter is not available, crystals for the desired test frequencies can be purchased and used to construct a simple series crystal probe for the scope. This is done by placing the crystal directly in series with a scope probe and connecting it to the signal-generator output. Adjust the signal generator to obtain maximum output on the scope, which will be indicated as a very sharp "jump" in the amplitude of the scope display when the crystal frequency is reached. The accuracy of the setting depends on the accuracy of the crystal used, but it is generally suitable for most test procedures. A possible exception is the region around 4.18 MHz, which is the rolloff point for color standards at the transmitter video output. A slight inaccuracy here can result in a false response characteristic in this critical region. This method, of course, is just as accurate as any other if a means exists to check the frequency error of the crystal against WWV. The most common WWV frequencies used are 2.5, 5, and 10 MHz.

The scope time base normally employs a low-frequency adjustment and several adjustments affecting the faster sweeps. A sine wave from the oscillator may be fed to the scope vertical amplifier, and the sweep time per centimeter may be calculated from the formula:

$$\text{Sweep time/cm} = \frac{\text{Cycles/cm}}{\text{Osc freq}}$$

For example, on the Tektronix Type 524AD scope, a 1-MHz signal should show one cycle/cm when the SWEEP TIME switch is set on 1 microsecond and the multipliers are set to 1.0. The accuracy of the marker generators should then be checked (and adjusted if necessary) against the properly calibrated time base.

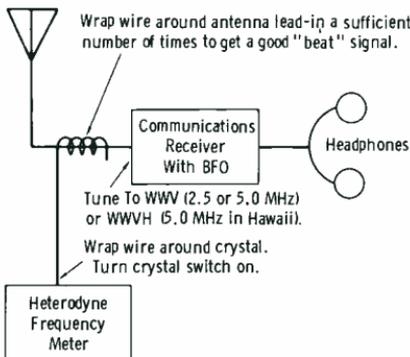
It is also important to be familiar with the sweep linearity of the scope. Usually a slight amount of nonlinearity is present in the



**Fig. 1-19. Scope sweep with 1-microsecond markers.**

524 scope. Fig. 1-19 illustrates a display with 1-microsecond markers; it can be observed that the linearity is reasonable to about 4 cm each side of center, allowing for a slight amount of parallax. Note, however, that in this case 10 microseconds is not indicated by exactly 10 cm of deflection. Nonlinearity is of no importance when markers are present, but instances occur (for example, when setting vertical-sync serration width relative to leading edge of horizontal sync) when it is cumbersome to attempt the use of markers. Excessive sweep nonlinearity usually can be improved by selecting horizontal-amplifier tubes for balance while observing markers on the trace, as in Fig. 1-19. It is most important, however, to determine what portion of the sweep is the linear region.

7. *Using WWV or WWVH for Calibration*—Mention has been made of checking frequency-standard calibration against WWV or WWVH transmissions. These transmissions are maintained by the National Bureau of Standards and are receivable throughout the United States. Difficulty may be experienced in some locations at night (because of skip effects and fading), in which case daytime use is mandatory. A good antenna system may be required together with a good communications receiver, preferably one incorporating a bfo.



**Fig. 1-20. Method for setting frequency of crystal.**

Fig. 1-20 illustrates one satisfactory method of calibrating the crystal in a secondary frequency standard against WWV or WWVH transmissions. Tune the receiver to 2.5 or 5 MHz (10 MHz if necessary for good reception), using the receiver bfo if it is required for accurate dial setting. Wrap a wire around the receiver antenna lead-in a sufficient number of times to obtain a good beat signal. A low-pitched audio growl indicates a frequency difference in the audio range. Adjust the trimmer capacitor across the frequency-meter crystal for zero beat.

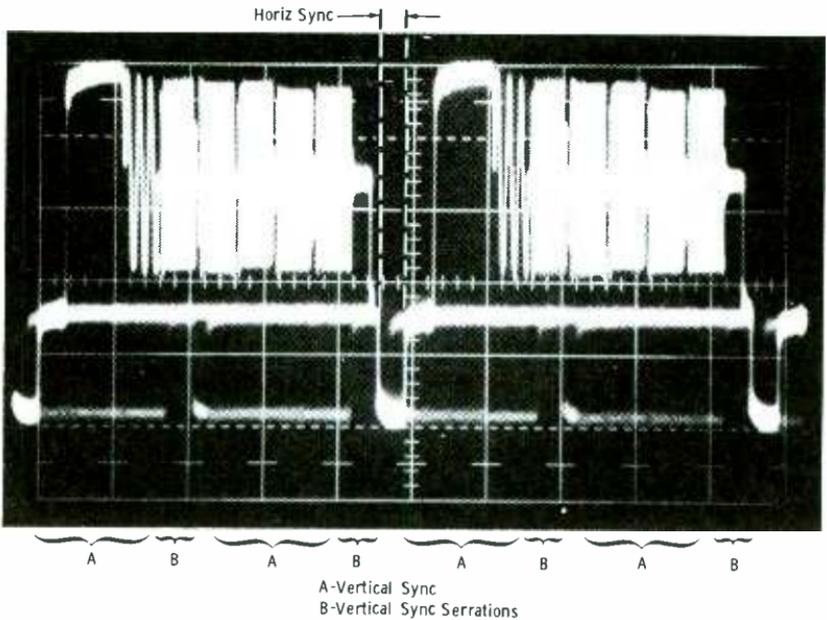
Always observe any precautions that may be spelled out in the instructions for the specific meter used. Many stations are in possession of the military Type BC-221 meter, accompanied only with the calibration book. These users should find the preceding information most helpful.

**IMPORTANT NOTE:** Sweep should be calibrated at periodic intervals on even the most recent scopes after considerable use. The same technique as that described above for the Type 524 applies, except that internal marker frequencies are not available.

8. *The Type 524 Delayed Sweep*—To display what is commonly termed “two lines” at a horizontal rate on the scope, the time base is adjusted to 7875 Hz (one-half the line rate), or 127  $\mu$ s. It is desirable to spread these two lines over the full ten-centimeter scale width. Therefore, the time base on the Type 524 scope is adjusted to 12.7 microseconds per centimeter by setting the sweep time to 10  $\mu$ s/cm and the sweep multiplier to 1.27, with the result that two lines at a horizontal rate appear across the graticule 10-cm scale. See Fig. 1-21.

In actuality, conventional horizontal-rate scope sweep results in a pattern which contains all the lines (at a horizontal rate) of the field. See following Section 1-4 for details. It is, however, possible to observe only a single line of the field on the Tektronix Type 524 scope by placing the TRIGGER SELECTOR switch to the Delayed Sweep sector. By rotating the SWEEP DELAY control, any line or lines may be observed (Fig. 1-22). This sweep is obtained internally from the composite television signal by establishing a coarse time delay from a vertical-sync pulse (from the sync-separator circuit) and then actually triggering the sweep from a selected horizontal-sync pulse. Since the scanned line interval is 63.5 microseconds, if the time base is adjusted to 6.35 microseconds per centimeter, a single line will occur in the full-scale 10 centimeters of the scope graticule. (Fig. 1-22 illustrates this feature.)

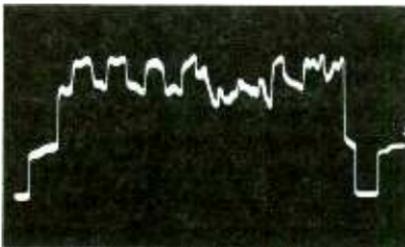
When the SWEEP DELAY control is used in this manner, the sweep is triggered only 30 times per second. The resulting display is correspondingly dim, and when much ambient light exists, the screen



**Fig. 1-21. Conventional horizontal-rate display of multiburst pattern.**

should be viewed through the hood provided for this purpose. A typical display like Fig. 1-22 requires a  $\frac{1}{2}$ -second exposure of 3000 ASA Polaroid film (at  $f/5.6$ ) compared to a  $\frac{1}{30}$ -second exposure for a display like Fig. 1-21.

The particular line being observed on the cro may be identified by connecting a spare video monitor to the LINE INDICATING VIDEO output jack at the rear of the scope. The picture on the monitor is brightened during the time of the scope sweep gate as determined by the selected time base. The SWEEP DELAY control is rotated until the desired line of the picture signal is selected (and brightened on



**Fig. 1-22. Scope display of a single line.**

the monitor). Thus, the amplitude of test-chart bandwidth wedges may be measured relative to gray (100-percent) areas.

NOTE: More recent scopes have a much brighter single-line display than does the Type 524, but they normally lack the external picture-monitor feed. This is provided in modern installations by the "line strobe output" of waveform monitors such as the Tektronix Type 529 or the 1480 series.

9. *Amplitude Calibration*—The absolute accuracy of the signal amplitude is not quite as important as time-base accuracy, provided the same scope is used in setting levels throughout the system. This is true since slight differences in levels are arbitrarily adjusted to give proper modulation of the transmitter or are adjusted for a given level in terminal equipment of the telephone company. However, a reasonable accuracy is desirable. EIA standards call for a picture line-amplifier standard output (black-negative polarity) of 1 volt peak-to-peak within 0.05 volt.

Mercury cells are available which are capable of maintaining a rather precise potential of 1.35 volts unloaded over periods of 30 months or more when used as secondary voltage standards. Eight of these cells in series provide a standard of 10.8 volts within 1 percent, and this standard can be used to calibrate dc meters. The pulse calibration circuit of most scopes employs a dc check point which requires a voltage scale that normally can be calibrated by the 10.8-volt dc reference. Or, the dc reference can be used to check the scope calibration directly if a chopper is available.

The properly calibrated scope can then be used to check the ac meter scales. The peak-to-peak value displayed multiplied by 0.3535 gives the rms meter value (assuming a sinusoidal waveform).

### 1-3. THE MODERN SWEEP TIME BASE (DUAL TRACE)

More modern scopes than the Type 524 not only employ much greater vertical-amplifier bandwidth, but generally allow dual trace presentations and more sophisticated methods of expanding portions of a given trace. These scopes provide extreme accuracy of pulse-width or rise-time measurement, and expanded display of any given line in either field of the composite television signal waveform.

The vertical amplifier has a bandwidth (to the  $-3$ -dB point) of between 30 and 50 MHz. In addition, display of two signals at one time is achieved by either one of two basic methods:

- A. Dual-beam scope.
- B. Dual trace by electronic switching of a single electron beam between two signal inputs. Since this type is the most prevalent

in broadcast service, we will concentrate on the switched beam method.

Electronic switches function in either of two ways:

- A. Rapid switching during sweeps. This method is termed the *chopped* mode.
- B. Synchronous switching during sweep retrace intervals. This method is termed the *alternate* mode.

The chopped mode is most useful at low frequencies, such as television field-rate signals. The alternate mode is used for displays at line frequency (15.75 kHz) and higher rates.

When two bright traces are displayed by the chopped mode, chopping-waveform transients appear as faint lines connecting the two traces. Some scopes turn off (blank) the crt beam during the chopping transition intervals to prevent the appearance of these lines in the display. This is accomplished in the Tektronix Type 545 and Type 547 scopes by a **CHOPPED BLANKING** toggle switch on the rear of the scope. More recent scopes do not require this switch.

### Sweep Magnification

Portions of sweeps may be magnified by increasing the gain of the horizontal amplifier. This is the simplest and least costly way to meet the need. See Fig. 1-23. A five-times sweep magnifier is available on most professional scopes. Thus, if the observed pulse is 1 microsecond in duration and the scope time base is set for 1  $\mu\text{s}/\text{division}$  (A in Fig. 1-23), the five-times sweep magnifier will spread the 1- $\mu\text{s}$  pulse over five divisions for close examination (B in Fig. 1-23). This type of magnification is subject to a certain amount of error and jitter, depending on pulse width and the required time base to be magnified.

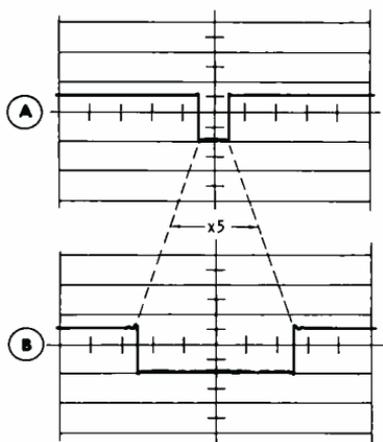


Fig. 1-23. Conventional sweep magnification.

### Delaying and Delayed Sweeps

Much greater magnification and accuracy, as well as greater freedom from jitter, are achieved by delaying-sweep methods using two linear calibrated sweeps. A typical arrangement is as follows:

1. Two separate sweeps, A and B, are provided. Either one may be used independently.
2. A selector switch for A or B sweep is provided. In the delayed-sweep operation, time base B becomes the basic sweep. The selector switch may then be placed in the B Intensified by A position, and a DELAY TIME MULTIPLIER is adjusted to intensify the trace at any given part of the B sweep.
3. The selector switch may then be placed in the A Delayed by B position. The selected portion of the B sweep will be displayed across the entire graticule, depending on the setting of the control for A sweep time per division.

See Fig. 1-24. In this example, the signal contains a vertical-interval test waveform on the last few lines of vertical blanking in each field. It is desired to examine this interval. The procedure is as follows:

1. See waveform A of Fig. 1-24. Time base B is adjusted to display slightly over one field so that the vertical-interval test (VIT) signal is readily observed. (VIT signals are described in Chapter 2.)
2. See waveform B of Fig. 1-24. With sweep A adjusted to a much shorter interval than sweep B, the DELAY TIME MULTIPLIER is adjusted to intensify the VIT interval with the selector switch placed in the B intensified by A position. The width of this brightening pulse (hence the total display to appear in the next step) is determined by the time base of sweep A.
3. See waveform C of Fig. 1-24. The sweep-selector switch is now set in the A Delayed by B position, and the selected portion of waveform B is displayed.

Fig. 1-25 illustrates the fundamental sweep processing for the delayed-sweep scope. The initial signal trigger is taken as the start of the event, or time zero ( $t_0$ ). This is the start of the undelayed sweep, which, for the delayed-sweep mode of operation, becomes the *delaying sweep*. The delaying-sweep voltage ramp is applied to a delay-pickoff stage. This stage is essentially a voltage comparator where the potentiometer sets the voltage threshold of the pickoff. The potentiometer is accurately calibrated and is termed the *delay-time multiplier* (DTM). The voltage comparator produces a trigger pulse at  $t_1$ , and this pulse starts the delayed sweep and initiates the

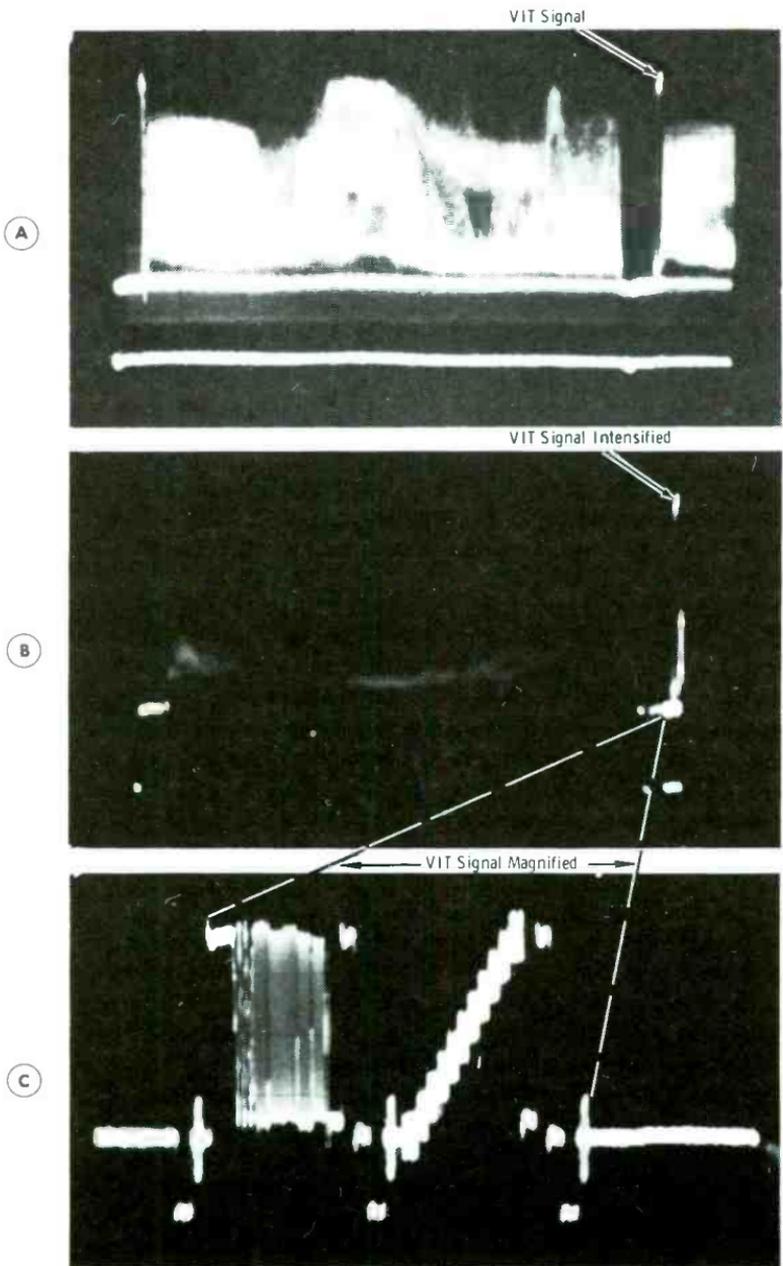
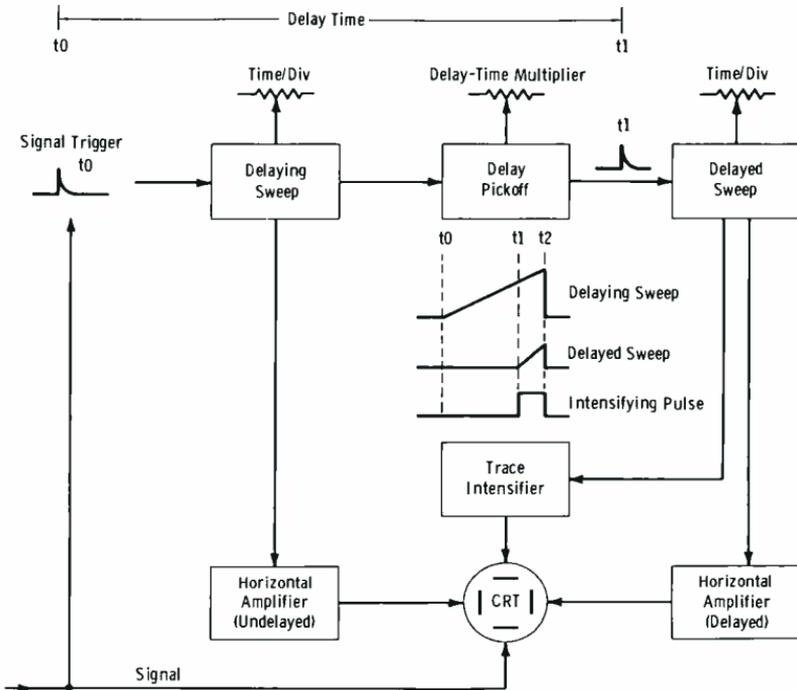
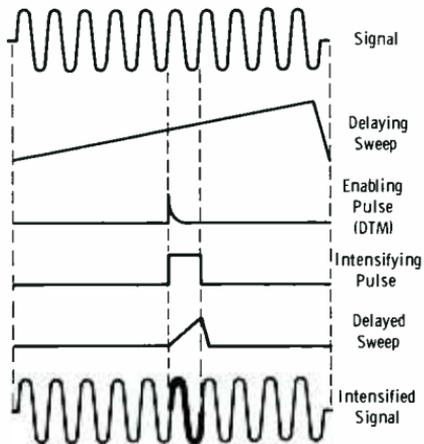


Fig. 1-24. Magnification of VIT signal with delayed-sweep scope.



**Fig. 1-25. Basic block diagram of delayed-sweep scope.**

leading edge of the intensifying pulse. Observe that the time base of the delayed sweep, which determines the width of the delayed sweep, also determines the width of the intensifying pulse. This is the interval from  $t_1$  to  $t_2$ . The resulting signal and sweep relationships are illustrated by Fig. 1-26.



**Fig. 1-26. Signal and sweep relationships in delayed-sweep scope.**

With this mode of operation, delay-time accuracy is determined basically by the delaying sweep and the potentiometer which sets the threshold level of the voltage comparator (DTM). The intensifying pulse indicates the time at which the delayed sweep starts with respect to the delaying sweep. Hence, delay time can be determined independently of horizontal-amplifier and crt considerations.

The delaying-sweep scope acts as a magnifier where the magnification power becomes the ratio of the delaying-sweep rate to the delayed-sweep rate. This is stated as:

$$\text{Magnification} = \frac{\text{Delaying-sweep rate}}{\text{Delayed-sweep rate}}$$

For example, if the delaying-sweep time base is set for 1 millisecond (ms) per centimeter, and the delayed sweep is 1 microsecond ( $\mu\text{s}$ ) per centimeter:

$$\begin{aligned} \text{Magnification} &= \frac{1 \text{ ms/cm}}{1 \mu\text{s/cm}} \\ &= \frac{10^{-3}}{10^{-6}} \\ &= 1000 : 1 \end{aligned}$$

The practical use of the delaying sweep in time and delay measurements is described in Section 1-4.

#### 1-4. BASIC SCOPE TECHNIQUES

The efficiency and accuracy of television system maintenance depends on complete familiarity with the specific system, the components making up the system, and the proper use of the oscilloscope.

##### Interpretation of Composite Video Waveform

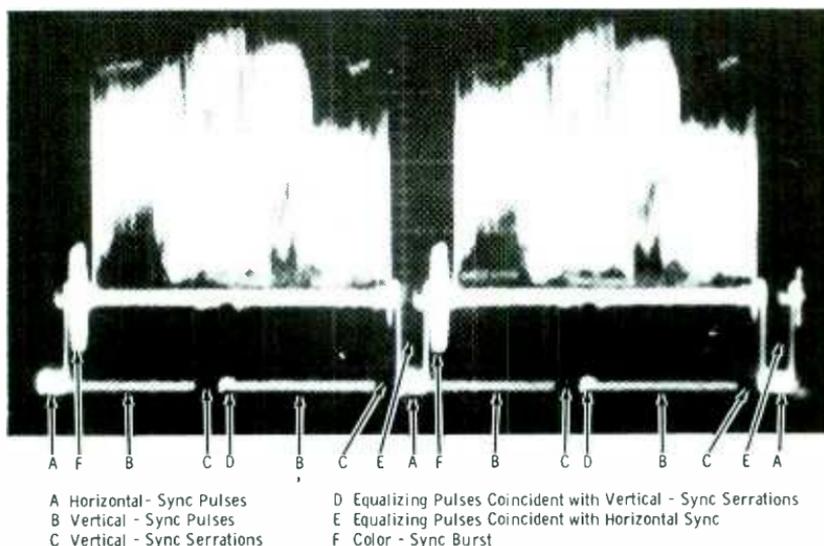
It is important to understand what the oscilloscope "sees" when the composite video signal is observed. The following points are pertinent:

1. When the scope time base is adjusted for viewing the signal at a horizontal rate, both horizontal and vertical pulses are displayed.
2. When the scope time base is adjusted for viewing the signal at a vertical rate, both vertical and horizontal pulses are displayed.

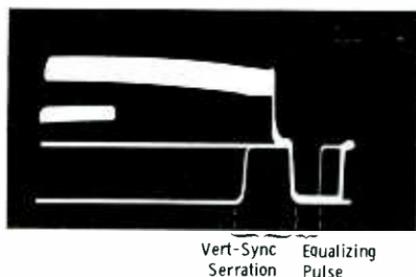
The video waveform, contrary to what is expected, is not displayed as a single line of varying displacement (amplitude) as is represented in drawings of a single video line. With conventional scope sweeps, the actual display of only two picture lines would be im-

possible unless the camera were scanning only two lines over and over. Actually, the scope must be considered as *scanning the video signal*, and, although during any single sweep of the scope beam (at the horizontal rate) two horizontal lines will have been scanned in the camera and traced out on the waveform screen, *all the lines of the picture* are present at the video input terminals to the scope. The effect of persistence of the fluorescent screen and the human eye, combined with the fact that all the lines of the picture are traced out on top of each other during successive sweeps of the scope beam, results in a pattern similar to that shown in Fig. 1-27.

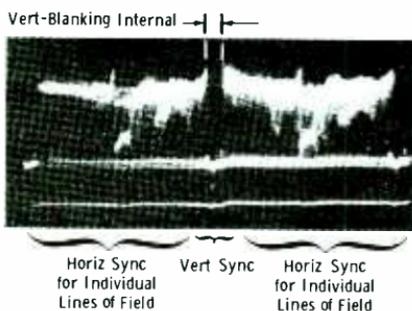
The line appearing at blanking level in Fig. 1-27 is in need of explanation. It should be remembered that at the end of each field (262.5 lines) the vertical retrace takes place, and the vertical-blanking pulse is present to blank out the kinescope beam during this time. Since the waveform-monitoring scope beam is not blanked out during this time, and since it is still sweeping the screen, the vertical-blanking voltage is traced out in the horizontal direction as shown. It is a relatively heavy line because the vertical pedestal is some 18 to 21 horizontal lines in duration (per field). Thus the beam is swept over a wide portion of the screen at a horizontal rate. The long interrupted line at the sync-tip level is formed by the vertical-sync pulses placed atop the vertical pedestal. They are of longer duration than the horizontal-sync pulses, and they are serrated in form.



**Fig. 1-27. Horizontal-rate display of "two lines."**



**Fig. 1-28.** Sweep expanded to show appearance of equalizing pulse and vertical-pulse serration.



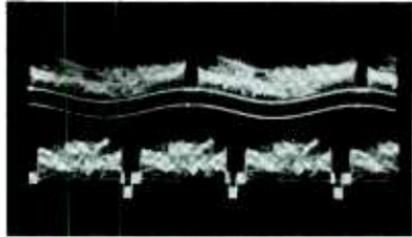
**Fig. 1-29.** Scope display with repetition rate of sweep set so that waveform includes two fields.

Notice that the equalizing pulse is also visible, although the trace produced is faint. This is because the pulse is at a high frequency (31.5 kHz) relative to the time base and exists only at 60-Hz intervals. Fig. 1-28 shows the equalizing pulse more clearly with expanded sweep. The leading edge of each *alternate* equalizing pulse coincides with the leading edge of a horizontal-sync pulse. Since the width of the equalizing pulse is one-half that of the horizontal-sync pulse, it appears to split the sync pulse in half. The leading edges of the remaining equalizing pulses *do not* coincide with horizontal sync, but they are coincident with the endings of the serrations of the vertical-sync pulses, as shown in Fig. 1-27.

When the monitor sweep is set at 30 Hz (33,333 microseconds), the *field-frequency* waveform is displayed as in Fig. 1-29. A single sweep now takes place in  $\frac{1}{30}$  second, the lines of one field appear to the left of the vertical pedestal, and the following field appears to the right. The line at the extreme bottom is composed of the *horizontal-sync* pulses for the lines of the fields above. Since there are approximately 262 such pulses for each field (one at the end of each line, 262.5 lines per field), the pulses appear as a horizontal line across the screen. The serrated vertical-sync pulses are also on this line immediately under the vertical pedestal, but they are hardly distinguishable as such because of their short time duration in ratio to the scope sweep of  $\frac{1}{30}$  second. In this case, the heavy line at the blanking level is made up of the horizontal pedestals for the lines in the fields above.

To emphasize the fact that vertical pulses are apparent in horizontal-rate cro displays (and vice versa), the upper trace in Fig. 1-30 shows the appearance of 60-Hz hum in the video signal viewed at the vertical rate. The lower trace in Fig. 1-30 is the same signal viewed at the horizontal sweep rate. Note the thickened lines in the horizontal-sync area. The same kind of display results for any

**Fig. 1-30. Signal with large 60-Hz hum component. Upper trace: vertical-rate display. Lower trace: horizontal-rate display.**



vertical “tilt” or shading; the horizontal-sync tips and porches are thickened an amount proportional to the vertical defect.

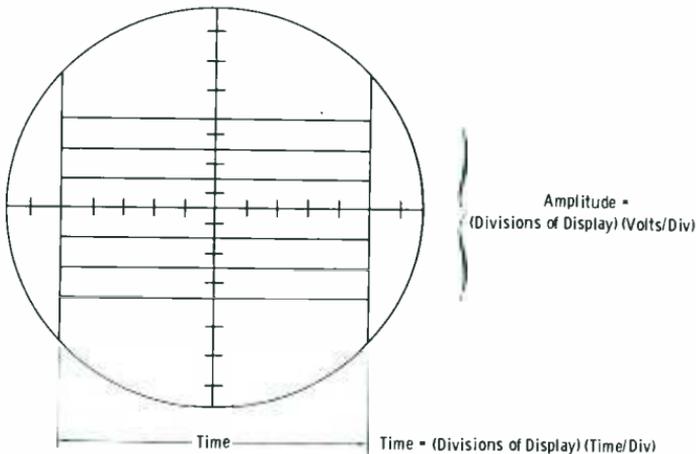
When a television line-rate defect, such as bad tilt (shading), exists at a 15,750-Hz rate, the vertical-rate cro sweep will reveal this as thickened vertical-sync traces.

### The Basic Scope Graticule

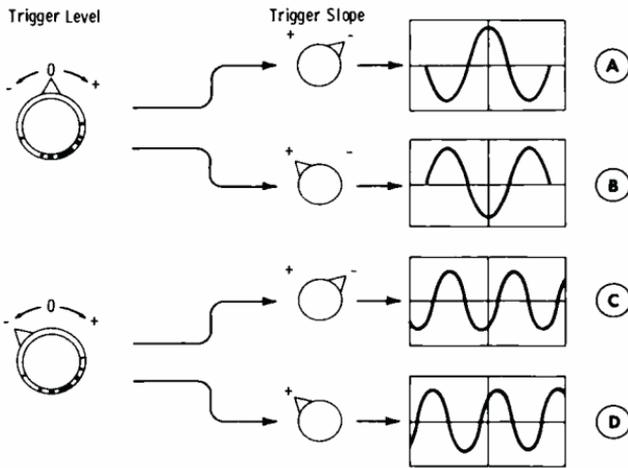
The basic scope graticule is illustrated in Fig. 1-31. There are usually six divisions vertically and ten divisions horizontally. The divisions are normally in centimeters (cm). The vertical axis expresses amplitude of the signal as the product of the number of divisions and the volts/division setting. The horizontal axis expresses time as the product of the number of divisions and the time/division setting of the time base.

### Trigger-Level and Trigger-Slope Controls

Observe Fig. 1-32 in the following analysis. Waveform A represents the display when the TRIGGER LEVEL control is centered and the TRIGGER SLOPE control is set on negative. The crt trace starts in the



**Fig. 1-31. Basic oscilloscope graticule.**



**Fig. 1-32. Effect of trigger-level and trigger-slope controls.**

center of the negative slope of the waveform. Waveform B represents the display when the TRIGGER LEVEL control is centered and the TRIGGER SLOPE control is set on positive. The crt trace starts in the center of the positive slope of the waveform.

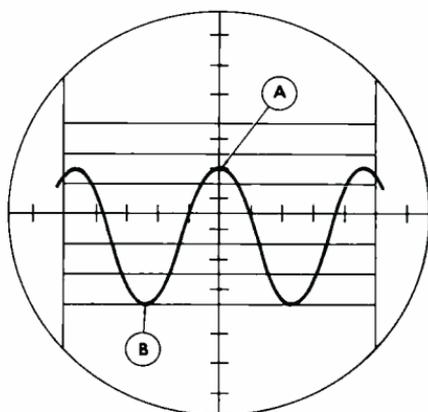
Waveform C of Fig. 1-32 shows the display when the TRIGGER SLOPE control is set on negative as for waveform A, but the TRIGGER LEVEL control has been rotated counterclockwise toward the negative trigger level. The crt trace starts at a more negative point of the negative slope than for waveform A. For waveform D, the TRIGGER SLOPE control is set on positive as for waveform B, but the TRIGGER LEVEL control has been rotated counterclockwise toward the negative trigger level. The crt trace starts at a more negative point of the positive slope than for waveform B.

Conversely, if the TRIGGER LEVEL control were rotated clockwise toward the positive trigger level, waveform C would start at a more positive point of the negative slope, and waveform D would start at a more positive point of the positive slope.

### Measuring Peak-to-Peak Levels

The scope input selector switch should be set on the ac position. Use the horizontal-position control to set the positive peak on the vertical center line of the graticule as in Fig. 1-33. This peak is point A of the waveform. With the vertical-position control, set the maximum negative excursion to a lower reference line, as at point B of Fig. 1-33. In this example, there are 4.5 divisions. Thus if the volts/cm switch is set on 1 volt/cm, the peak-to-peak voltage is 4.5 volts (if a direct probe is used).

**Fig. 1-33. Peak-to-peak voltage measurement.**



In practice, the  $\times 10$  probe is normally used. This must be considered in the voltage measurement. The relationship is:

$$\text{Volts (p-p)} = \left( \frac{\text{Vertical deflection}}{\text{in centimeters}} \right) \times \left( \frac{\text{Volts/cm}}{\text{switch setting}} \right) \times \left( \frac{\text{Probe atten}}{\text{factor}} \right)$$

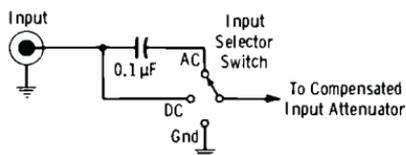
Thus, in the example of Fig. 1-33, assuming that the volts/cm switch is set to 0.05 and a  $\times 10$  probe is used:

$$\text{Volts} = 4.5 \times 0.05 \times 10 = 2.3 \text{ volts (approx)}$$

**IMPORTANT NOTE:** The oscilloscope has a **VARIABLE VOLTS/CM** control in addition to the fixed **VOLTS/CM** switch. When it is necessary to measure actual voltages, always be sure that the variable control is in the calibrate position. The basic purpose of this control is to provide variable gain for obtaining an exact number of divisions on the graticule when *comparing* a voltage with a fixed and known voltage. Usually a small warning lamp is illuminated when this control is in any position other than calibrate.

### DC-Component Measurements

See Fig. 1-34 for a typical input selector circuit. When it is desirable to measure the dc level at a given point on a waveform, the selector switch should be placed first in the ground position. When the voltage is expected to be positive relative to ground, use the vertical-position control to position the trace on the bottom line of the graticule. (The sweep selector can be placed in the automatic mode to obtain a free-running trace in the absence of a signal.) Then place the input selector switch in the dc position with the probe at the pertinent point of the circuitry being checked.

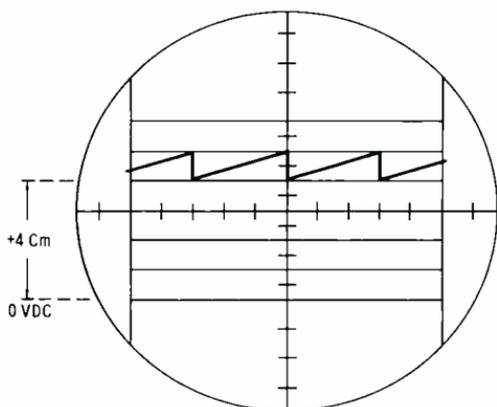


**Fig. 1-34. Typical scope ac-dc-ground selector arrangement.**

In the example of Fig. 1-35, assume the volts/cm switch is set on 0.1 and that a  $\times 10$  probe is used. Since there are +4 cm between ground reference and the bottom of the signal:

$$\text{Volts} = 4 \times 0.1 \times 10 = +4 \text{ volts dc}$$

Since the deflection is 1 volt/cm, the sawtooth waveform is 1 volt p-p riding on a dc level of +4 volts.



**Fig. 1-35. Measurement of dc signal reference.**

This procedure is quite useful in large-signal transistor-circuit testing, since both the dc and ac components are conveniently checked by one measurement. "Large-signal" application simply means that the ac signal is a significant portion of the applied dc voltages in the circuit. "Small-signal" application means that the ac component may be extremely small compared to the applied dc. For example, the signal component in the above example could be in the millivolt range. In this case, the input selector is placed in the ac mode after the dc component has been measured, and the vertical gain is increased to a range suitable for signal voltage measurement.

The above emphasizes one important point: the scope is normally the most accurate vtmv you can use. This is particularly true for the relatively low voltages encountered in solid-state circuitry.

**CAUTION:** Never exceed the input voltage rating of your particular instrument. This is normally 600 volts maximum, which includes the

signal p-p value plus the dc potential applied. This maximum situation seldom presents itself, particularly when the  $\times 10$  probe is used, except in certain picture and waveform monitor circuitry.

### Time-Difference Measurements

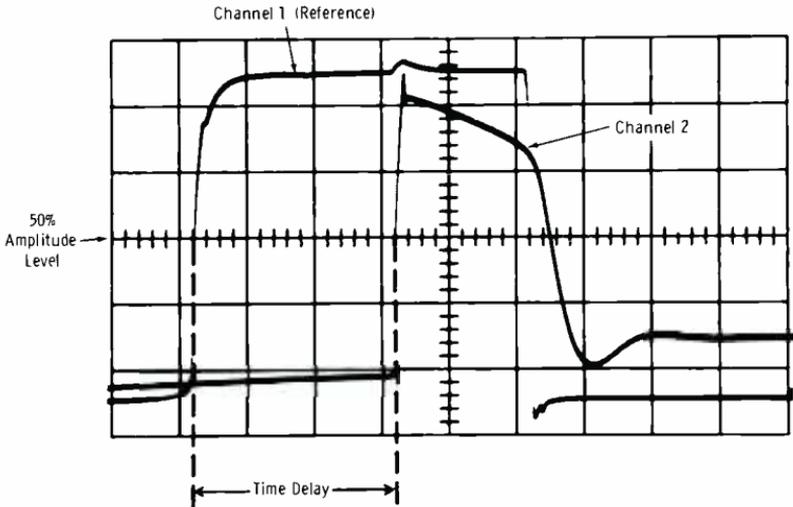
The accurately calibrated sweep rate and dual-trace feature of modern scopes allow rather precise measurement of the time difference between two events. But there are certain very important precautions to be observed.

The normal method of dual-trace presentation on the scope is not suitable for measurement of time differences between two signals. The reason is that the scope vertical-amplifier internal trigger is the trigger source. Thus the signals applied will not be displayed in their true time relationship because triggering occurs from the applied signal as each channel switches on.

To measure time difference, both traces must be triggered by only one signal, the reference signal. To do this, a small coaxial cable is run from the channel 1 trigger output connector to the scope external trigger in connector. The trigger selector switch is placed on ext + or ext -, whichever gives a stable trigger for the waveforms observed. Proceede as follows:

1. Set the mode switch to either chop or alt. The chop mode is most suitable for low-frequency signals and the alternate mode is used for higher-frequency signals.
2. Set both the channel-1 and channel-2 volts/cm switches so that the expected voltages applied to the input connectors will provide suitable vertical deflection.
3. Apply the reference signal to channel 1, and apply the other signal to channel 2. Use coaxial cables or probes having equal delay (same length).
4. Set the oscilloscope time-base controls for a calibrated sweep rate which will allow accurate measurement of the distance between reference points of the two waveforms.
5. Measure the horizontal distance between the appropriate points of the reference waveform and the channel-2 waveform (see Fig. 1-36).
6. Multiply the measured distance by the setting of the oscilloscope time/cm switch to obtain the apparent time interval.
7. To obtain the actual time interval, divide the apparent time interval by the amount of sweep magnification, if sweep magnification is used, and by 1 if no sweep magnification is used. The formula is as follows:

$$\text{Time delay} = \frac{(\text{Time/cm switch setting}) (\text{Distance in cm})}{\text{Sweep magnification}}$$



Courtesy Tektronix, Inc.

**Fig. 1-36. Measurement of time difference between waveforms.**

For example, assume that the time/cm switch setting is  $2 \mu\text{s}/\text{cm}$ , the magnifier is set for  $5\times$  magnification, and there is a horizontal distance of 3 cm (as shown in Fig. 1-36) between the leading edges of the waveforms. Then, substitute these values in the preceding formula:

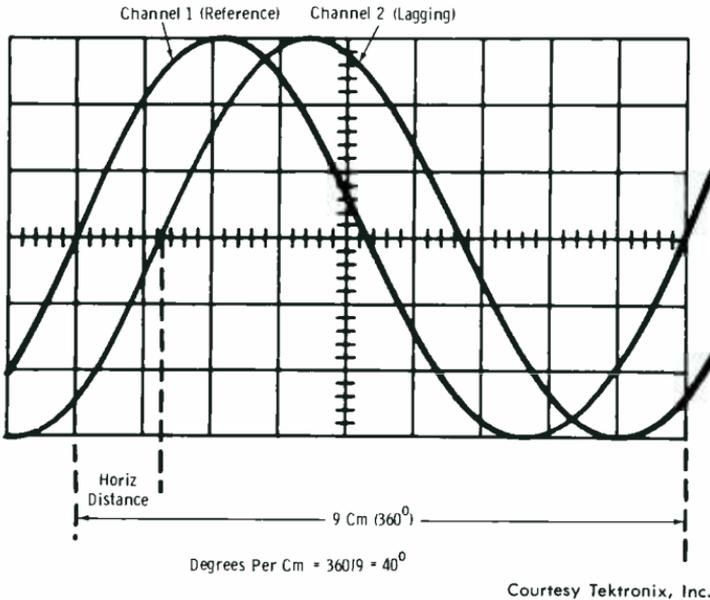
$$\text{Time delay} = \frac{(2 \mu\text{s}/\text{cm}) (3 \text{ cm})}{5} = 1.2 \mu\text{s}$$

### Phase Measurements

Phase comparison of two signals of the same frequency can be made using the dual-trace feature. To make the comparison, proceed as follows:

1. Follow the first three steps of the procedure outlined above for time-difference measurements.
2. Set the oscilloscope sweep rate to obtain a display of less than 1 cycle of the waveform.
3. Adjust the variable volts/cm controls for each channel so the waveform amplitudes are equal and fill the graticule area vertically. Reset the volts/cm switches, if necessary, to obtain equal-amplitude waveform displays. (Equal amplitudes are used to make comparisons easier.)
4. Use the position controls to center the waveforms vertically; that is, they should occupy an equal distance each side of the graticule center line.

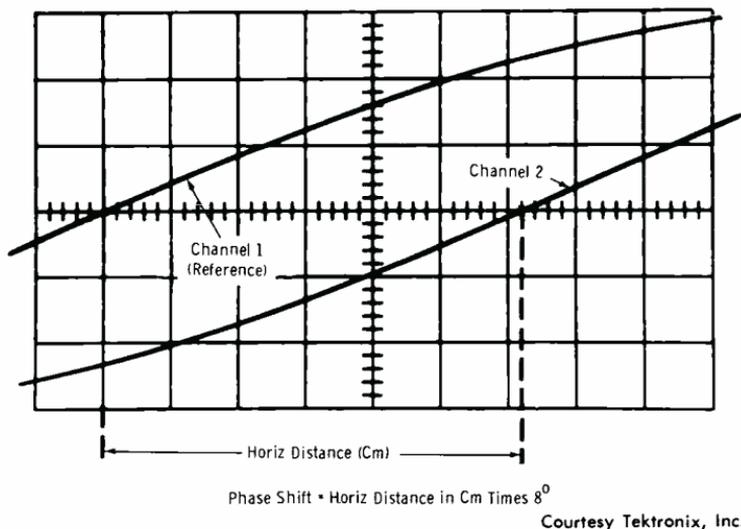
- Turn the oscilloscope variable time/cm control counterclockwise until 1 cycle of the reference signal occupies 9 cm horizontally. Use the trigger-slope and triggering-level controls to trigger on the reference waveform at any desired point. Each cm on the graticule now represents  $40^\circ$  of 1 cycle (see Fig. 1-37).



**Fig. 1-37. Measurement of phase difference between waveforms.**

- Measure the horizontal distance, in cm, between corresponding points on the waveforms. Note the distance and whether the channel 2 waveform is leading or lagging (Fig. 1-37).
- Multiply the measured distance by  $40^\circ/\text{cm}$  to obtain the amount of phase difference.

For more precise measurements, increase the previous sweep rate, but do not change the setting of the oscilloscope variable time/cm control. However, you must consider this increase in your calculations. For example, if you increase the sweep rate by a factor of 5, and then measure the distance between waveforms, each centimeter will represent  $8^\circ$  ( $40 \div 5$ ) of a cycle. Thus, phase differences up to  $80^\circ$  can be measured more accurately. When preparing to make the measurement, position the waveforms horizontally to points where the graticule markings aid in determining the exact distance. Fig.



**Fig. 1-38. Measurement of phase difference with increased sweep rate.**

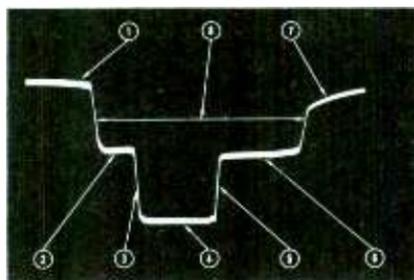
1-38, for example, shows how the phase difference of the channel-2 waveform can be found with this method.

Note that in measuring time difference between two waveforms, the 50-percent point of the waveform normally is used to avoid errors that result from trying to compensate for differences in rise time or overshoots.

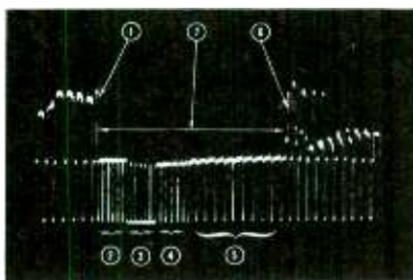
## 1-5. TERMINOLOGY

It is becoming increasingly important that all tv engineers "speak the same language." Fig. 1-39 illustrates two scope displays and the descriptive terminology associated with their analysis. This terminology is approved by AT&T and local telephone companies. Terminology used in describing specific troubles is presented where appropriate in later chapters.

It is common practice to call out levels from blanking toward white and blanking toward sync as "100 over 40," "85 over 30," etc.; the numbers refer to IEEE units in each direction from blanking level. When calling out keyed sine-wave burst levels to the telephone company for frequency-response checks, adjust the level from blanking to peak reference white for 100 units on the scope, and read each burst frequency in units occupied by the individual burst. Reference white is established by a pulse immediately following blanking for use in this adjustment (Chapter 2). Always be sure



1. Picture voltage at right side of image.
2. Front porch.
3. Leading edge of sync.
4. Tip of sync.
5. Trailing edge of sync.
6. Back porch.
7. Picture voltage at left side of image.
8. Horizontal-blanking interval.

(A) *Horizontal.*

1. Picture voltage at bottom of image.
2. Leading group of 6 equalizing pulses.
3. Serrated vertical-sync pulse.
4. Trailing group of 6 equalizing pulses.
5. Horizontal-sync pulses during vertical blanking.
6. Picture voltage at top of image.
7. Vertical-blanking interval.

(B) *Vertical.***Fig. 1-39. Nomenclature of sync waveforms.**

your waveform monitor or scope is calibrated so that 140 IEEE units = 1 volt p-p.

The setup level (units between blanking and maximum picture black) should be called out only when picture content includes a reference black rather than intermediate shades of gray only. This level should be a minimum of 5 and a maximum of 10 units under this condition. It is standard operating practice at some stations to run zero setup on all camera controls and to insert a fixed 5-percent setup at the line-output processing amplifier.

All level checks not involving frequency response should be made with the scope response on the IRE position. If the scope is on wide-band response, the small-energy overshoots of the higher-frequency components will be apparent. If these overshoots are held below 100 units and the scene suddenly changes to one of much lower-frequency content, the operator must adjust his gain to bring the overall level up. Since the luminance content is largely in the middle- and lower-frequency range, the operation results in a needless shift of apparent contrast in the home receiver. To avoid this result, the IRE response curve was standardized for the purpose of "riding gain" on the video and for checking levels of normal signals between the studio and the transmitter, or the studio and the telephone company.

**NOTE:** The narrow-band IRE response characteristic is described in Harold E. Ennes, *Television Broadcasting: Equipment, Systems, and*

*Operating Fundamentals* (Indianapolis: Howard W. Sams & Co., Inc., 1971).

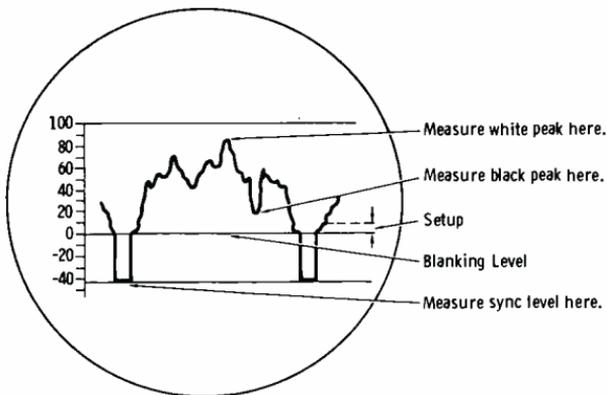
See Figs. 1-40 and 1-41 for specific application of the following terms relating to video levels.

**BLACK PEAK:** The point of maximum excursion of the picture signal in the black direction at the time of observation (Fig. 1-40).

**BLACKER-THAN-BLACK:** The amplitude region of the composite video signal below reference black level in the direction of the synchronizing pulses.

**BLANKING LEVEL:** The level of the front and back porches of the composite video signal (Fig. 1-40).

**BOUNCE:** Sudden variation of level of the picture signal.

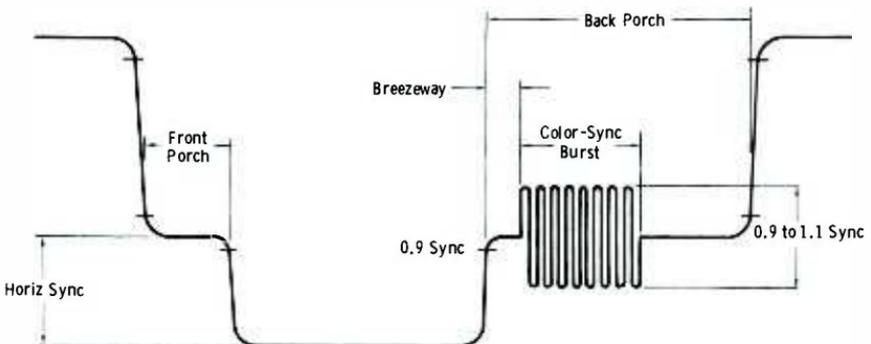


White Peak = 86 (Meaning 86 IEEE scale units, or 86 percent of the range between blanking and reference-white levels.)

Black Peak = 13 (Meaning 13 IEEE scale units, or 13 percent of the range between blanking and reference-white levels.)

Sync Level = -43 (Meaning 43 IEEE scale units, or 43 percent of the range between blanking and reference-white levels, but negative with respect to blanking level.)

**Fig. 1-40. Use and interpretation of IEEE standard scale.**



**Fig. 1-41. Details of color-sync region.**

**BREATHING:** Amplitude variations similar to bounce but at a slow, regular rate.

**BREEZEWAY:** The interval between 0.9 amplitude on the trailing edge of sync and the start of the first color-burst sine wave. (Fig. 1-41).

**CLIPPING:** Sharp cutoff of the video signal peaks. It may affect either the white (positive) peaks or the black (negative) peaks. The sync amplitude may be affected on a composite signal.

**COLOR-SYNC BURST:** The eight to ten cycles of color subcarrier on the back porch of horizontal sync. This burst must have a nominal amplitude equal to sync-pulse level. A level of 0.9 to 1.1 times sync level is tolerable (Fig. 1-41).

**COMPRESSION:** An undesirable decrease in the amplitude of a portion of the composite signal relative to that of another portion. This term defines a less than proportional change in output of a circuit for a change in input level. For example, sync-pulse compression means a decrease in the percentage of sync relative to that at the sending end.

**DISPLACEMENT OF PORCHES:** Any difference between the level of the front porch and the level of the back porch.

**OVERSHOOT:** Excessive response to a unidirectional signal change. Sharp overshoots are sometimes referred to as "spikes."

**PEAK-TO-PEAK:** The amplitude (voltage) difference between the most positive and the most negative peak excursions of the signal. (Abbreviated p-p.)

**PEDESTAL LEVEL:** See "blanking level," now the preferred term.

**POLARITY OF PICTURE SIGNAL:** This term refers only to the polarity of the *black* portion of the waveform as it appears on the cro with respect to the white portion of the signal. It is standard that outputs of camera chains, distribution amplifiers, and terminal equipment be black negative, which is standard polarity for the transmitter input to produce a positive image to the viewer. This term does not refer to the picture as it appears on the monitor in terms of a "positive" or "negative" image.

**REFERENCE BLACK LEVEL:** The level corresponding to the specified maximum excursion of the signal in the black direction.

**REFERENCE WHITE LEVEL:** The level corresponding to the maximum excursion of the luminance signal in the white direction.

**SETUP:** The separation between blanking and reference black levels.

**SYNC LEVEL:** The level of the tips of the sync pulses (Fig. 1-40).

**VIDEO-IN-BLACK:** A condition in which the black peaks extend through reference black level as observed on the cro. More often termed "loss of setup."

**WHITE PEAK:** The maximum excursion of the picture signal in the white direction at the time of observation (Fig. 1-40).

### EXERCISES

- Q1-1. An oscilloscope amplifier is specified as having a rise time of 35 nonoseconds. What is the rise time in microseconds?
- Q1-2. With the rise time specified in Q1-1, what is the frequency response (to the  $-3$ -dB point) in (A) megahertz and (B) gigahertz?
- Q1-3. A schematic diagram shows a coupling capacitor of 470 picofarads (470 pF). What is the value in (A) microfarads ( $\mu\text{F}$ ) and (B) micromicrofarads ( $\mu\mu\text{F}$ )?
- Q1-4. Two types of video-sweep detector probes are shown in Fig. 1-9. How can you check your particular detector probe without disassembly to trace the circuitry?
- Q1-5. Assume your detector probe has a 25-percent attenuation factor. You desire to feed a video sweep signal of 0.5 volt peak-to-peak amplitude to an amplifier input. What detected amplitude should you read?
- Q1-6. Why is correct frequency calibration of the test oscillator particularly important in the region from 4.1 to 4.2 MHz?
- Q1-7. See Fig. 1-33. If the volts/cm switch is set on 0.02 and a  $\times 10$  probe is used, what is the peak-to-peak voltage?
- Q1-8. In a dual-trace scope, can the internal trigger of the scope be used in measuring the time difference between two waveforms?
- Q1-9. See Fig. 1-36. Assume the time/cm switch is set for  $1 \mu\text{s}/\text{cm}$  and that no sweep magnification is used. What is the time difference?
- Q1-10. See Fig. 1-35. Assume the time/cm switch is set for  $10 \mu\text{s}/\text{cm}$  and that no sweep magnification is used. (A) What is the sawtooth duration? (B) What is the sawtooth frequency?

## Television-System Waveforms

The television broadcast system is unique in many respects, and a keen insight into this uniqueness is required of the system maintenance engineer.

### 2-1. THE SYSTEM CONCEPT

It must be emphasized that the overall system concept is complete only when the "average receiver" is included in the analysis. Much of the processing carried out at the transmission end (picture sources, studio and transmitter circuitry) is necessary to compensate for the characteristics of the home receiver.

#### Basic Modulation and Demodulation

Fig. 2-1A illustrates conventional double-sideband (dsb) amplitude modulation. This modulation produces two sidebands, symmetrically located on either side of the carrier. For example, if the carrier is varied in amplitude at a rate of 4 MHz, three frequencies result: the carrier frequency, an upper sideband frequency 4 MHz above the carrier, and a lower sideband frequency 4 MHz below the carrier.

The demodulated output of two sidebands of 100 percent and a carrier of 100 percent is shown in Fig. 2-1B. Note that this type of modulation would require a bandwidth of 8 MHz for a 4-MHz video signal, plus 0.5 MHz more for the sound carrier, plus a 0.5-MHz guard band, for a total of 9 MHz.

Each set of sidebands contains the same information (although in inverted order of frequency). Thus all the essential picture information can be conveyed by using only one of the sidebands, and the very scarce spectrum space can be conserved.

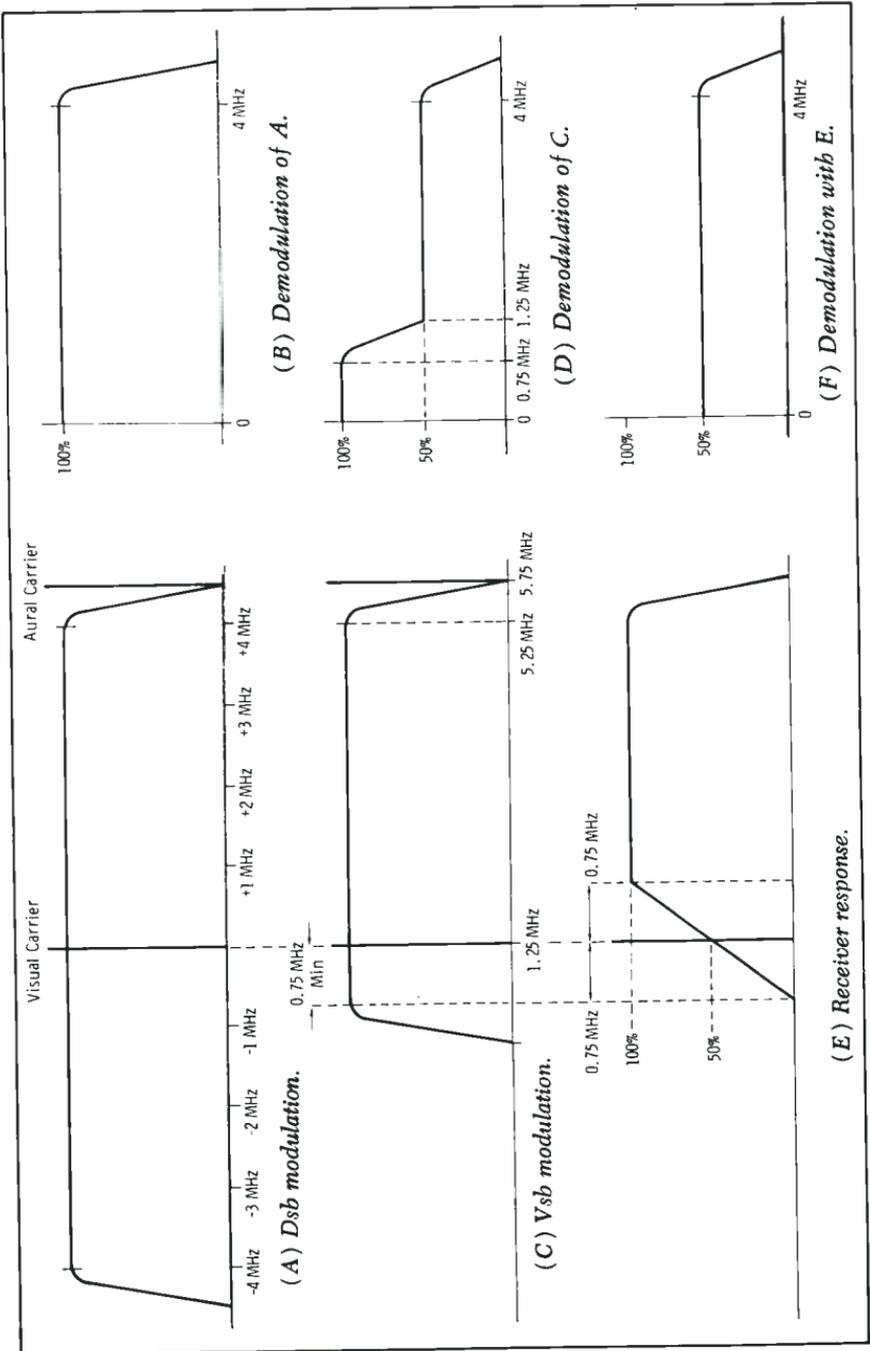


Fig. 2-1. Ideal modulation-demodulation characteristics.

When one set of sidebands is completely eliminated, the associated phase shift around the filter cutoff produces excessive distortion adjacent to the carrier frequency. Therefore, a "vestige" of the lower sideband is retained in vestigial-sideband transmission, as depicted in Fig. 2-1C. At a frequency of 1.25 MHz below the carrier frequency (designated as zero on the drawing), the sideband energy must be at least 20 dB down. The response must be within 2 dB to 0.75 MHz below the carrier frequency. Thus a rather sharp rolloff occurs between 0.75 and 1.25 MHz in the lower sideband.

Fig. 2-1D illustrates the resulting diode response in the transmission line following the vestigial-sideband filter, when used. (When low-level modulation is used, linear amplifiers following the modulated stage are tuned to obtain the vestigial-sideband characteristic, and no filter is necessary.) Note that since double sidebands occur for modulating frequencies up to 0.75 MHz, 100-percent response is demodulated. Modulating frequencies above 1.25 MHz are carried in single sideband only; hence the demodulated response for these frequencies is reduced 6 dB, or 50 percent.

The required receiver amplitude-response curve is shown in Fig. 2-1E. The receiver is tuned so that the if amplifier response is positioned with the visual carrier frequency at the 50-percent point of a straight line from minimum response at 0.75 MHz below to full response at 0.75 MHz above the carrier frequency. The resulting demodulated response is depicted in Fig. 2-1F. The receiver tuning has attenuated the picture carrier by 50 percent on the linear slope of Fig. 2-1E. This provides the equivalent of one set of sidebands of 100-percent amplitude and a carrier of 50-percent amplitude, producing one-half the demodulated output of double-sideband transmission. Since frequencies above 0.75 MHz are transmitted single sideband, the response is again 50 percent of that for double-sideband transmission. Therefore the demodulated response is flat from zero to 4 MHz (actually 4.2 MHz), as shown by Fig. 2-1F.

It should be observed that with proper receiver tuning the receiver demodulation is the same whether the transmission is dsb or vsb. With a double-sideband transmission, frequencies in the lower sideband greater than approximately 0.75 MHz are ignored, and the end result is the same as for vestigial-sideband transmission. Most test equipment for feeding a receiver an rf signal has dsb modulation, but this is of no consequence.

The vsb transmitted signal occupies  $1.25 + 4.5 = 5.75$  MHz (Fig. 2-1C) as compared to the 9 MHz required for dsb transmission. Adding the aural channel and a guard band gives the authorized 6-MHz channel width. This allows an increase of almost 50 percent in the number of assigned tv channels over that possible if dsb transmission were employed.

The required filter shaping of both the upper and lower sidebands, however, presents problems which must be handled by the maintenance technician or engineer. We will consider these in the following subsections.

### Effect of Modulation-Demodulation System

Fig. 2-2 shows the sideband graph of vsb transmission with the addition of the I and Q chroma placement in the upper sideband. The response must be within plus or minus 2 MHz. A very sharp rolloff must then occur so that at 4.5 MHz (position of the aural carrier) the visual response is down at least 20 dB.

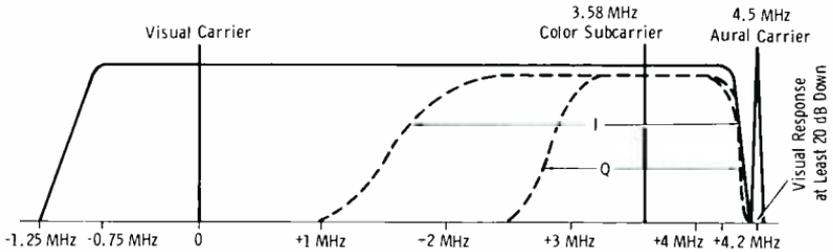


Fig. 2-2. Vsb signal with chroma subchannel.

Fig. 2-3 translates these transmitted components to the IF frequencies in the receiver. The frequencies are inverted because the tuner oscillator operates 45.75 MHz *higher* in frequency than the transmitted picture carrier frequency. The receiver incorporates a *sound trap* so that the response at the sound carrier frequency is attenuated at least 50 dB. This is necessary to minimize 920-kHz beats between the sound modulation and the video information.

The effect of the restricted bandwidth (4.2 MHz), the sharp roll-offs at both ends of the vsb transmission band, and the sharp

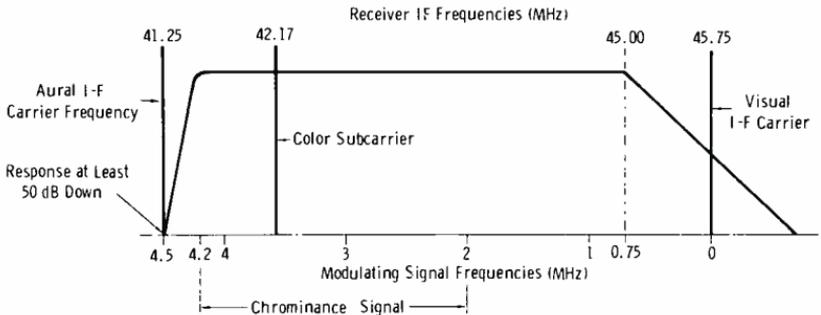
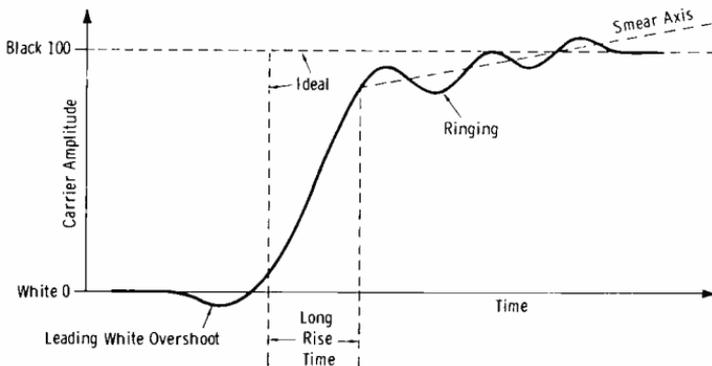


Fig. 2-3. Ideal receiver overall tuner and IF response.



**Fig. 2-4. Average demodulated response to step transition in the absence of transmitter video predistortion.**

sound-carrier trap in the receiver combine to result in an overall step response as depicted in Fig. 2-4. Under ideal conditions (without transmitter video predistortion), the demodulated step response results in a lengthened rise time (loss of resolution), a leading overshoot (white before black), and a smear axis resulting in black following black. Picture white (minimum carrier) and picture black are arbitrarily assigned values of zero and 100, respectively, in the drawing.

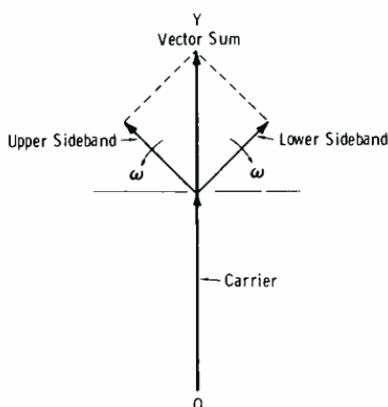
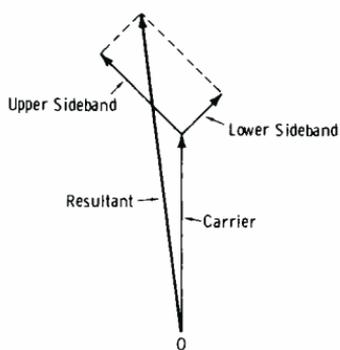
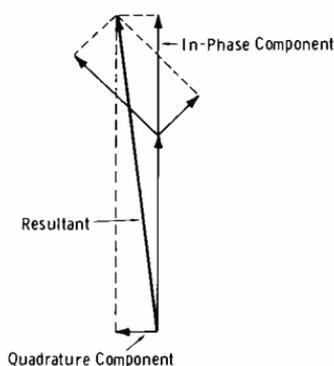
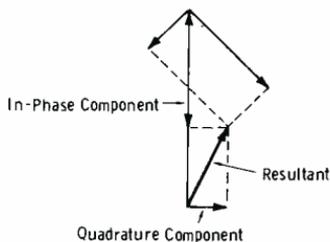
The above facts simply mean that the basic vsb transmission and reception process results in amplitude and phase distortion. For this reason, amplitude and phase precorrection are used at the transmitter to minimize these distortions in the reproduced picture. Transmitter precorrection is covered in Chapters 10 and 11.

**NOTE:** Many color receivers do not employ the type of curve illustrated in Fig. 2-3. The chrominance subcarrier frequency may be placed at the 50-percent response point as the picture carrier is. Extra amplification is then provided for chroma. This allows use of an if bandwidth which need not exceed the conventional 3-MHz range.

Note also that the chroma information runs into the sound-carrier cutoff region. The FCC specifies that the response up to 4.2 MHz be within 2 dB, and that *no sudden amplitude changes even within the 2-dB limit* shall occur. Sharp peaks or dips within this sideband region are common causes of saturation and chroma-phase errors. This can occur at either the transmitter or receiver.

### Quadrature Distortion

For comparison purposes, see Fig. 2-5A. Double-sideband modulation results in equal-amplitude sideband vectors rotating in opposite directions with constant angular velocity. The resultant ampli-

(A) *Dsb transmission vectors.*(B) *Vsb transmission vectors.*(C) *In-phase and quadrature components.*(D) *Components during carrier decrease.***Fig. 2-5. In-phase and quadrature components of vsb transmission.**

tude at any instant is the vector sum of the sideband vectors added to the carrier vector. At the instant represented by Fig. 2-5A, the vector positions are such that the carrier is increased in amplitude. If the sideband vectors were in quadrants 3 and 4, the carrier amplitude would be decreased. In any case, with equal-amplitude sidebands the resultant always lies along line OY.

With vsb transmission, only signals up to 0.75 MHz are radiated as double sidebands. Fig. 2-5B shows the effect of adding unequal sideband vectors to the carrier vector. Note from Fig. 2-5C how the resultant of Fig. 2-5B can be resolved into an in-phase and quadrature component. The quadrature component is  $90^\circ$  out of phase with the carrier.

This type of distortion can be made negligible, provided that both the transmitter and receiver demodulators can be made *skew symmetrical*. See Fig. 2-5D. If the *relative amplitudes* of the upper and lower sidebands can be held the same on carrier decrease as on carrier increase, the skew (quadrature) component will have the same amplitude in both cases. The vectors of Fig. 2-5D are simply those of Fig. 2-5C rotated 180°. With skew symmetry, phase pre-correction processing may be carried out at the transmitter to minimize quadrature distortion (Chapters 10 and 11).

### Resolution

It has been shown that restricting the visual-transmitter bandwidth to 4.2 MHz, along with rapid rolloff at both limits of the bandwidth, results in a demodulated signal which exhibits loss of resolution, ringing, and phase distortion that causes leading and trailing smears on sharp signal transitions.

There are two resolution factors for a television picture. Vertical resolution is independent of system bandwidth, and horizontal resolution is directly related to system bandwidth. Horizontal and vertical resolution factors are developed in a companion volume in this series.<sup>1</sup> Table 2-1 lists values of horizontal resolution for several bandwidths and the corresponding rise times. Vertical resolution is approximately 340 television lines.

### The Transfer Curve

The linearity of the signal-transfer curve ideally should be such that the picture-tube brightness varies exactly in accord with the

**Table 2-1. Bandwidth, Rise Time, and Horizontal Resolving Power**

Bandwidth (MHz)	Rise Time ( $\mu$ s)	TV Lines
1	0.35	80
2	0.175	160
3	0.1166	240
4	0.0875	320
5	0.07	400
6	0.058	480
7	0.05	560
8	0.0437	640
9	0.039	720
10	0.035	800

<sup>1</sup> Harold E. Ennes, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals* (Indianapolis: Howard W. Sams & Co., Inc. 1971).

variations in brightness of the original scene. Starting with the receiving terminal, the picture tube does not have a linear relationship of luminance output to signal-grid voltage swing. Not complicating the situation with factors such as stray light on the face of the picture tube, or of picture-tube phosphor characteristics, this relationship is:

$$Y = E^n$$

where,

Y is the reproduced brightness (luminance),

E is the signal-grid drive voltage above picture cutoff,

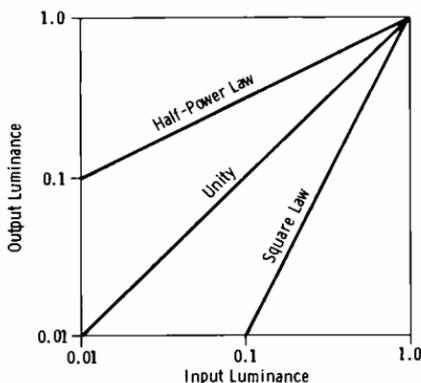
n is an exponent (gamma) which in practice lies between 2 and 3.

For example, if the exponent is 2, a square-law relationship exists in which luminance output increases as the square of the applied signal voltage. Thus blacks are compressed and whites are stretched.

At the transmitting end, the pickup tube seldom has a gamma of unity, although the Plumbicon comes close with an average gamma of 0.95 in many tubes. Even if the pickup device is strictly linear, some correction is necessary for the picture tube.

The resultant overall characteristic of an uncompensated system is black compression because the kinescope is more nonlinear than the pickup tube. This is equally true for the vidicon or Plumbicon.

Fig. 2-6 is a simplified log-log plot of three transfer curves. Unity is the desired characteristic. If the picture tube were a strictly



**Fig. 2-6. Log-log plot of transfer characteristics.**

square-law device (exponent 2), the pickup tube would need to have an exponent of 0.5 to result in a linear overall brightness transfer to the picture-tube screen. It is normally assumed that a color picture tube has an exponent of 2.2, although this obviously varies in practice.

The vidicon has a relatively constant gamma of 0.65 over the normal beam-current operating range. The average kinescope gamma is at least 2, which means that the picture-tube highlight brightness

increases approximately as the square of the applied video above cut-off. Then assuming all other units of the system have unity gamma, the overall system gamma is:

$$(0.65)(2.2) = 1.43$$

or greater than unity. The amount of gamma correction necessary for a vidicon to obtain a unity exponent is:

$$1/1.43 = 0.7$$

so that the product of the system gamma (1.43) and the gamma correction (0.7) is 1, or unity.

The Plumbicon (or any of the equivalent lead-oxide vidicon pickup tubes) has a gamma of 0.95. In this case, the overall system gamma is:

$$(0.95)(2.2) = 2.09$$

so that the amount of gamma correction necessary with the Plumbicon (or equivalent) pickup tube to obtain a unity exponent is:

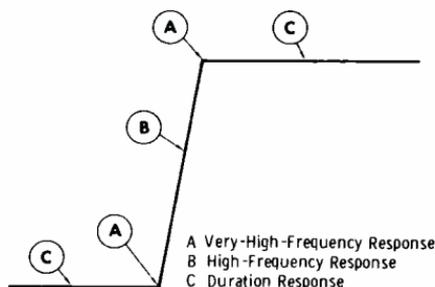
$$1/2.09 = 0.478 \text{ (approx 0.5)}$$

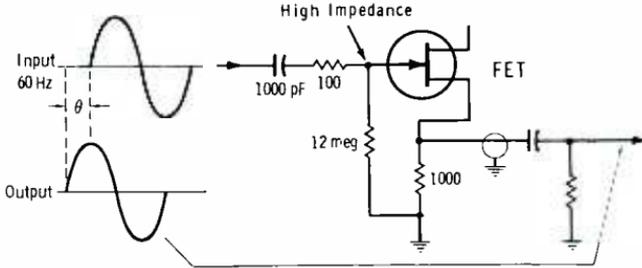
This gamma correction is made in each studio pickup device, such as studio and film camera chains. The remainder of the system must then be made as nearly linear as possible to retain the proper relationship of picture-tube and pickup-tube brightness transfer.

## 2-2. SINE-WAVE vs PULSE RESPONSE

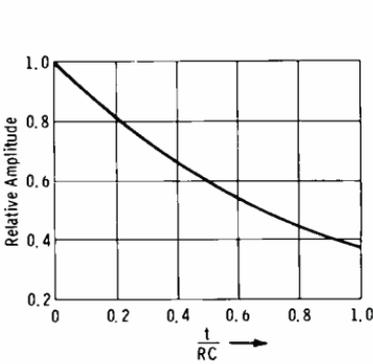
Fig. 2-7 points out the basic relationship between frequency response and pulse characteristics. The rise (and fall) time and the shape of the pulse corners depend on high-frequency response and the shape of the passband response curves. The duration response is (ideally) dc in character. Thus the ideal pulse amplifier passes frequencies from dc to very high frequencies, with no phase distortion. We will look first into the duration response ( $t_d$ ), which depends

**Fig. 2-7. Pulse characteristics affected by frequency response.**

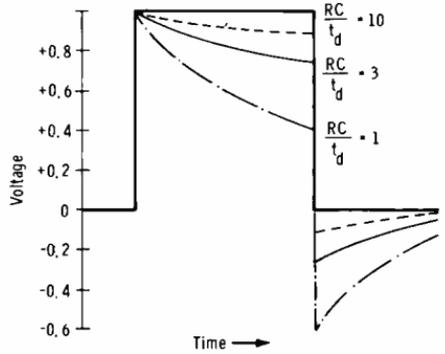




(A) FET source follower circuit.



(B) Output voltage.



(C) Square-wave tilt.

Fig. 2-8. Time constant.

on  $t/RC$ , or time divided by the  $RC$  product. This is the low-frequency characteristic in practice.

Fig. 2-8A illustrates a source-follower FET circuit that has a sine-wave response at 60 Hz down only 1 dB relative to 1 MHz. Since the relatively small value of capacitance results in a leading phase shift ( $\theta$ ), this is the only significant difference in the output compared to the input.

The output voltage as a function of  $t/RC$  is shown in Fig. 2-8B. At the instant the pulse is applied ( $t = 0$ ), the output voltage is 1 times the input voltage. As  $t$  increases, the factor  $t/RC$  increases, and the output voltage decreases until at  $t/RC = 1$ , the output voltage drops to 0.37 of the applied voltage. (This is basic theory which can be found in most basic texts.)

Since the pulse durations required in a tv system are known, it is most convenient to use the reciprocal of the above relationship in thinking of practical  $RC$  coupled circuits. Fig. 2-8C is a plot of the output voltage of a pulse in relation to the  $RC/t_d$  ratio. Note that it is necessary to have an  $RC$  product of 10 times the pulse duration

( $t_d$ ) to avoid more than a 10-percent tilt over the duration of the pulse. It is obvious that the time-constant problem becomes severe in any practical circuit when the duration of the field is 16,666  $\mu$ s (the reciprocal of the field rate of 60 per second).

The time constant (TC) is in seconds when R is in ohms and C is in farads or R is in megohms and C is in microfarads. The time constant is in microseconds ( $\mu$ s) when R is in ohms and C is in microfarads or R is in megohms and C is in picofarads. The second relationship ( $\mu$ s) is most useful for tv circuits.

For example, a 0.1- $\mu$ F coupling capacitor and a 1-meg input resistance result in a time constant of 0.1 s, or 100,000  $\mu$ s. This is not 10 times the field duration. The TC value in practical circuits is limited by the stability factor (motorboating, large capacitances to ground, etc.); this is why either negative feedback to flatten the lows (as well as the highs) is used, or a low-frequency boost circuit is employed. In amplifiers incorporating clamping circuits, the low-frequency characteristics are almost entirely dependent on proper operation of the clamp pulse former and clamping circuit.

See Fig. 2-8A again. For all practical purposes, the time constant of this circuit is  $1000 \text{ pF} \times 12 \text{ meg} = 12,000 \mu\text{s}$ . This is an  $RC/t_d$  ratio of less than 1 for the field duration in a television signal. Thus, although this circuit will pass a 60-Hz sine wave with negligible distortion, note the effect on a 60-Hz square wave (or pulse), as shown in Fig. 2-8C, for an  $RC/t_d$  ratio less than 1.

For a system to pass a pulse with the same rise time and shape as the input pulse, the bandwidth of the system must be:

$$BW = \frac{1}{2RT}$$

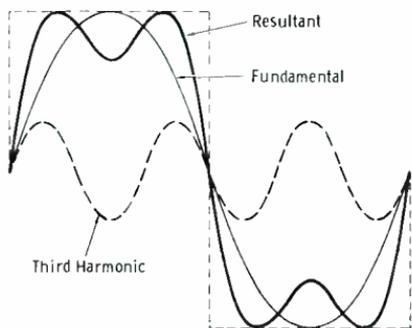
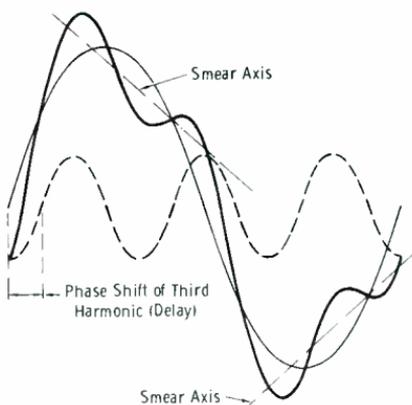
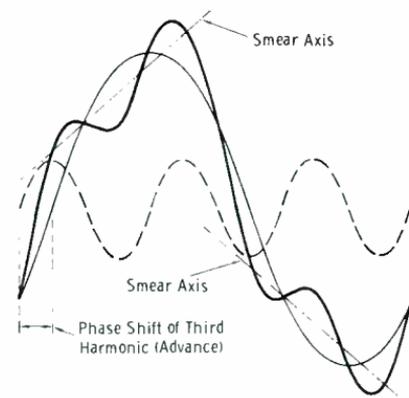
This says the bandwidth must be equal to half the inverse of the rise time. For a pulse with a rise time of 0.02  $\mu$ s:

$$BW = \frac{1}{2(0.02)} = \frac{1}{0.04} = 25 \text{ MHz}$$

(Since the rise time is in microseconds, the result is in megahertz.)

Note that we have used a different relationship for conversion of rise time (RT) to bandwidth (BW) than in Section 1-2. That relationship is based on the assumption that overshoot will be under 3 percent, for this condition the k factor is 0.35, and the bandwidth is 0.35 divided by the rise time. For a rise time of 0.02  $\mu$ s, this calls for a bandwidth of 17.5 MHz. The rise time is not affected materially, but a slight amount of overshoot occurs with a rise time of 0.02  $\mu$ s when the bandwidth of the channel is 17.5 MHz.

The relationship which states that the bandwidth must be equal to half the inverse of the rise time is based on Fourier analysis.

(A) *Tendency to form square wave.*(B) *Harmonic delayed relative to condition in A.*(C) *Harmonic advanced relative to condition in A.***Fig. 2-9. Combinations of fundamental and third harmonic.**

The Fourier theorem says that any recurring nonsinusoidal waveform can be shown to be made up of sine waves, cosine waves, or both, of various amplitudes, phases, and frequencies.

Fig. 2-9A illustrates a sine wave combined with its third harmonic. Note that the resultant shows a tendency to start formation of a square wave, by the steepening of the sides. (The "dip" in the center is filled in by higher-frequency harmonics.) As odd-order harmonics are added, this effect increases. A perfect square wave would be composed of an infinite number of odd harmonics. The amplitudes of the added harmonics vary inversely with frequency; thus, the higher the frequency of the harmonic, the less is its amplitude in relation to the fundamental. Provided the bandwidth is adequate, these harmonics are retained in the original amplitude (and phase) relationship, and the pulse is passed without distortion.

Suppose the third harmonic suffers a different time delay than its fundamental. See Fig. 2-9B. Note that this results in a square-wave

response with a tilt across the top and an undershoot at the trailing edge. The edges of an image will now be dispersed by an amount depending on the phase shift of the harmonics making up the signal.

Fig. 2-9C illustrates the case in which the third harmonic leads the fundamental, resulting in a tilt or smear axis opposite to that of Fig. 2-9B.

*Caution:* The reader should consider that the square wave of Fig. 2-9 represents a very high-frequency signal, so that leading and trailing edge effects are being discussed, and not the duration (dc) component of Fig. 2-8C. An amplifier with good dc response but a rapid cutoff at a higher frequency will affect the edges but not the flat top of a long-duration pulse (*except* at the edges). (Review Fig. 2-4.)

A gaussian response curve, while essential in oscilloscope vertical amplifiers (Section 1-2), is not found in television camera chains or in video distribution amplifiers. The reason is the limitation of rise time in a series of amplifiers forming a cascaded system. The rise time of the original waveform is reduced by the square root of the sum of the squares of the amplifier rise times.

Thus, if we pass a signal through two identical 10-MHz gaussian-response amplifiers:

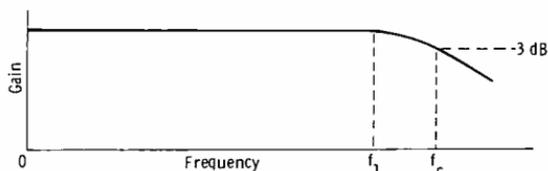
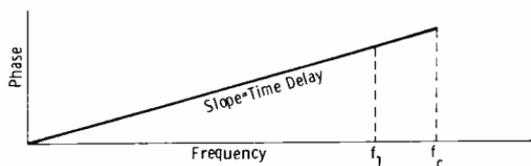
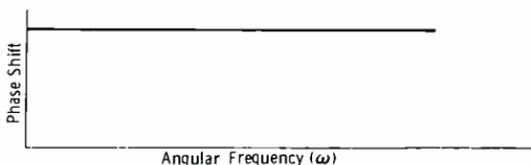
$$\begin{aligned} RT &= \sqrt{0.035^2 + 0.035^2} \\ &= \sqrt{0.002450} = 0.05 \mu\text{s (approx)} \end{aligned}$$

This is a 40-percent increase in rise time as a result of passing the signal through just two cascaded 10-MHz gaussian amplifiers. In practice, many video amplifiers are cascaded in forming a complete system.

It can be seen that the practical video amplifier must have a flat frequency response up to and including the highest anticipated frequency, with a relatively rapid rolloff beyond this frequency. It can be shown from pulse theory that rise time is proportional to the *area* under the amplitude-frequency response curve; hence, cascading such amplifiers does not appreciably affect the rise time. However, such an amplifier will *not* reproduce a step transition at the output free of overshoot, ringing, or other transient distortions.

Fig. 2-10A reviews the amplitude-frequency response curve of the gaussian amplifier. Since an oscilloscope must display waveshape as it actually exists without any error introduced by the scope amplifier itself, this type of response curve is used for frequency- and transient-response measurements.

The ideal phase-frequency characteristic is one in which the phase angle increases linearly with frequency (Fig. 2-10B), maintaining the same time delay for all frequencies (Fig. 2-10C). Departure

(A) *Gaussian rolloff.*(B) *Phase response.*(C) *Envelope delay.*(D) *Time response.***Fig. 2-10. Ideal response of oscilloscope vertical amplifier.**

from this characteristic results in phase distortion most noticeable as trailing white edges following black edges on a gray or white background.

If the phase characteristic departs from the ideal straight slope and bows upward away from the frequency axis, the increasing slope with frequency is an indication that the time delay is increasing with frequency. (This is shown in Fig. 2-11B.) The shifted components making up the pulse now add in such a manner that a leading-edge overshoot and trailing-edge undershoot occur on the passed pulse. Conversely, if the phase characteristic curve bows toward the frequency axis (also shown in Fig. 2-11B), the decreasing slope with frequency indicates a decrease in time delay with frequency. The shifted pulse components now add to produce a leading-edge undershoot and trailing-edge overshoot. It should be noted in passing that

one type of phase distortion may be compensated by an equal and opposite phase correction; a lagging phase shift is corrected by an equal leading phase shift, and vice versa.

Fig. 2-10D shows the ideal time response to a step transition through the gaussian-response amplifier. If the rise time of the pulse to be observed does not exceed  $0.35/\text{bandwidth}$  (Section 1-2), no overshoot or undershoot is added to the pulse display.

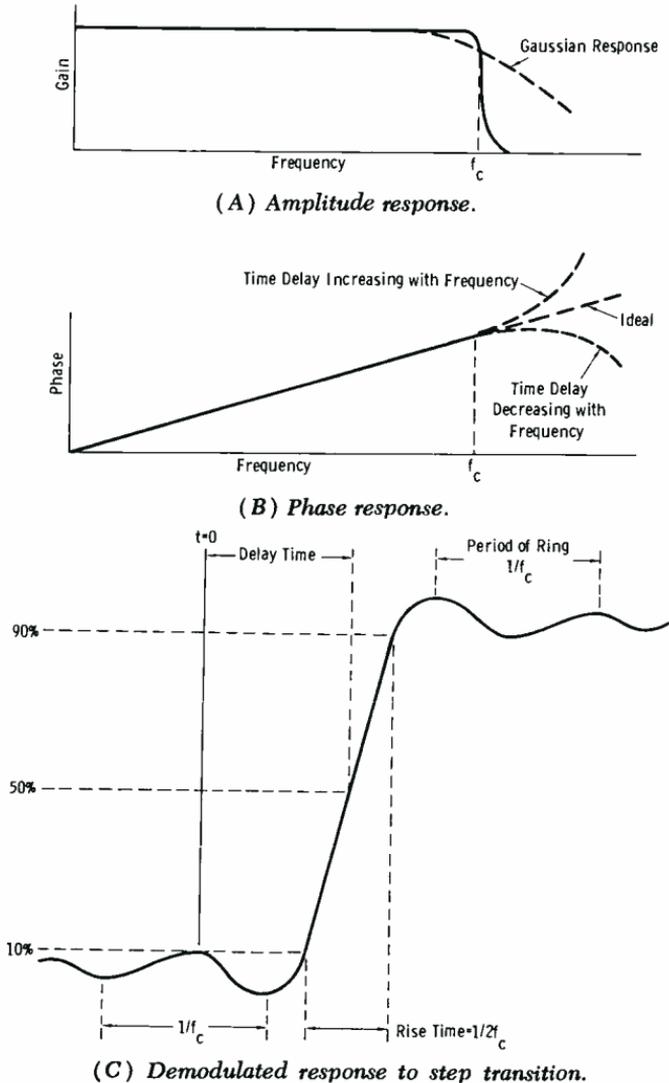


Fig. 2-11. Overall response of practical television system.

Fig. 2-11A shows the video-system amplitude response as compared to that of a gaussian-response amplifier. Remember that the transmitter has a very sharp cutoff between 4.2 and 4.5 MHz. In addition, the receiver employs a sound trap with even sharper cutoff.

Fig. 2-11B shows the phase response already discussed. Phase shift cannot be proportional to frequency around and above the cutoff frequency of the system. A pulse (step transition) requires transmission of the higher-order harmonics, which are actually above the passband intended, to be free from waveshape distortion.

Fig. 2-11C shows the resultant demodulated step-transition response, along with the frequency relationships. Since the cutoff frequency ( $f_c$ ) of the overall tv system is 4.2 MHz, the ideal overall rise time (between 10 and 90 percent points) is  $1/2f_c = 0.12 \mu\text{s}$ , and the period of ring is  $1/f_c = 0.24 \mu\text{s}$ .

The ringing *amplitude* depends on the step-transition rise time for a given amplifier bandwidth and rolloff characteristic. The distribution of ringing (leading and trailing) is an indication of direction and degree of phase shift. Late arrival of high-frequency components causes most of the ringing to occur on the trailing edge of a pulse, whereas early arrival of high-frequency components causes most of the ringing to occur at the leading edge. Fig. 2-12 shows a phase-corrected signal, indicated by the even distribution of ringing at the leading and trailing edges.



Fig. 2-12. Ideal pulse response of overall tv system.

### 2-3. DEVELOPMENT OF THE $\text{SIN}^2$ PULSE

Thus far, we have learned that the sine-wave response of a video amplifier does not provide a complete story of the amplifier performance for a video signal. Likewise, a step transition (or a square-wave signal) is not a particularly useful signal for evaluation unless the exact rise time of the pulse is properly correlated with the intended passband of the amplifier. The very important *transient response*, which accounts for the degree of picture ringing, smearing, or streaking, requires a precise analysis method to assure valid tests.

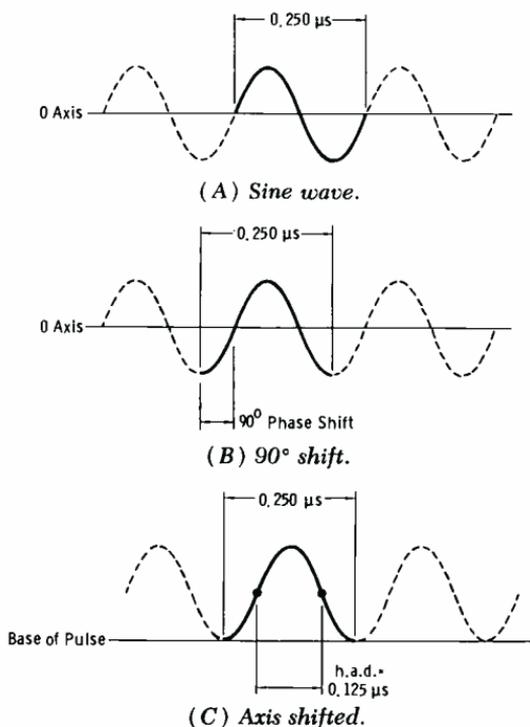
The fact that pulse rise times are not related to the actual picture transmission spectrum is the reason why we will be concerned with the sine-squared ( $\text{sin}^2$ ) pulse. The basic usefulness of this test sig-

nal is in complete system testing for linear distortion, which is covered in Chapter 8, Part 1.

First of all, it is necessary to understand what a picture element is. A picture element is determined by the available bandwidth. Our complete tv system is fixed by FCC standards, which allow the visual transmitter only about a 4-MHz bandwidth for the picture signal. One cycle occurs in a time equal to the reciprocal of the frequency; therefore:

$$1 \text{ cycle at } 4 \text{ MHz} = \frac{1}{4(10^6)} = 0.250 \text{ microsecond}$$

This means that a black-to-white transition of a vertical bar with a width representing 4 MHz will occur in 0.250 microsecond. But black is one picture element and white is one picture element. Therefore, a picture element of a 4-MHz system is 0.125 microsecond (one alternation of the complete cycle). In the  $\sin^2$  technique, a time duration of one picture element is given the symbol T, and a time duration of two picture elements (for the system bandwidth under test) is symbolized by 2T.



**Fig. 2-13. Derivation of sine-squared pulse.**

A basic method of explaining the sine-squared pulse is shown in Fig. 2-13. In Fig. 2-13A, notice the conventional continuous sine wave at a frequency of 4 MHz; one cycle of this wave occurs in a time interval of  $0.250 \mu\text{s}$ . You realize from fundamental theory that any phase shift of a continuous sine wave is measured only by laborious methods not suitable for routine testing of transmission facilities. Also, the amplitude-frequency characteristic of a system simply shows the amplitude of the continuous sine wave relative to a reference frequency, unless you are equipped to measure the phase relative to a known reference.

Observe Fig. 2-13B. If we shift the waveform  $90^\circ$ , we have one complete cycle of a 4-MHz cosine wave, starting and finishing at its negative peaks. Now if we add a dc component of such value as to raise the negative peaks to the zero power line, we have the T pulse of a 4-MHz system (Fig. 2-13C). As shown in Fig. 2-13C, the half-amplitude duration (h.a.d.) is  $0.125 \mu\text{s}$ , equivalent to one picture element for a 4-MHz bandwidth. Fig. 2-14 shows that the significant

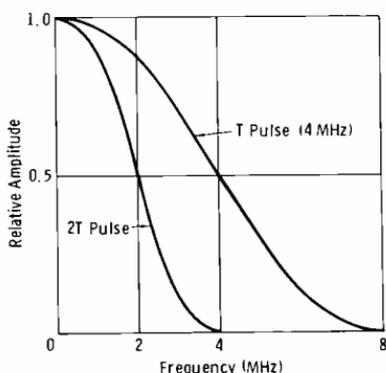
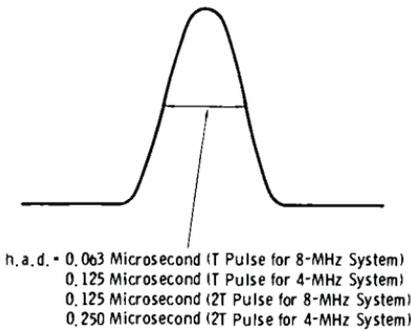


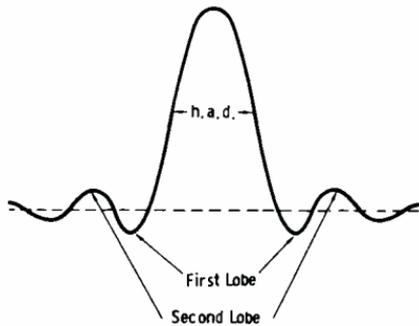
Fig. 2-14. Frequency spectrum of sine-squared pulse.

energy spectrum of the T pulse is 50 percent (6 dB) down at 4 MHz, and there is practically no energy beyond 8 MHz. The 2T pulse (h.a.d. of  $0.250 \mu\text{s}$ ) is 50 percent (6 dB) down at 2 MHz, and there is no significant energy beyond 4 MHz. Thus the system can be checked with a pulse that essentially duplicates actual picture conditions and which provides known frequency content upon which to base judgment of system performance. Please note that any similarity to the sine wave no longer exists; a *pure* sine wave has no harmonic content at all.

Fig. 2-15 shows the preceding definition in terms of T and system bandwidth for 4-MHz and 8-MHz systems. This test pulse is similar to an actual scanned picture element where black is represented as a dc component and the pulse simulates a black-to-white leading transition and a white-to-black trailing transition.



**Fig. 2-15. Sin<sup>2</sup> pulse in terms of T and bandwidth.**



**Fig. 2-16. Nomenclature of pulse overshoots.**

Fig. 2-16 shows the terminology used with a pulse that has passed through an amplifier or (more usually) a complete system. The first lobe is a negative overshoot, and the second lobe is a positive overshoot, preceding and following the pulse.

The sin<sup>2</sup>-pulse generator normally also generates a window signal following the pulse, so that an amplitude reference to low frequencies is established. This is described in Section 2-5.

## 2-4. DEVELOPMENT OF THE 20T AND 12.5T PULSE

The tv transmitter and demodulator combine to form a 4-MHz (approx) low-pass filter. The 2T pulse for a 4-MHz system (h.a.d. = 0.25  $\mu$ s) has practically no energy at the high end of the band, and therefore does not reveal errors that occur around the high-energy color-subcarrier region. The T pulse for a 4-MHz system (h.a.d. = 0.125  $\mu$ s) has an energy spectrum up to twice the cutoff frequency and therefore has high energy content in the color-subcarrier region. However, this pulse is also distorted by an "ideal" 4-MHz low-pass filter because of the energy beyond the usable upper range. The usefulness of the T and 2T pulses is confined to indicating transients as pointed out in practical applications in Chapters 8 and 11. It will suffice at this time to understand that the 2T pulse is a sensitive indicator of transmission distortions up to 60 or 70 percent of the nominal upper video-frequency limit.

The 20T pulse shifts measurement emphasis from determining the ability to reproduce transients to determining (A) the gain *difference* between the high and low ends of the video frequency spectrum and (B) the relative *delay* time between the high and low ends of the video frequency spectrum. Essentially, the 20T pulse is a signal at the frequency of the color subcarrier, modulated by a sine-squared pulse. The h.a.d. of the modulating pulse was chosen so that the

sum of the subcarrier frequency and the highest spectral frequency of the pulse does not exceed the upper video band limit of 4 MHz.

If we take the color-subcarrier frequency, rounded off to 3.6 MHz, we can see that an added 0.4 MHz (400 kHz) takes us to the upper limit of 4 MHz. Then, the h.a.d. of the modulating pulse is  $1/0.4(10^6) = 2.5 \mu\text{s}$ , which is 10 times that of the 2T pulse and 20 times that of the T pulse. The modulating envelope (20T pulse) produces a frequency spectrum from 60 Hz to 400 kHz. A second spectrum extends 400 kHz above and below the subcarrier (Fig. 2-17). The subcarrier normally is *not* locked to the line repetition rate so that the envelope shape is more clearly defined on the scope trace.

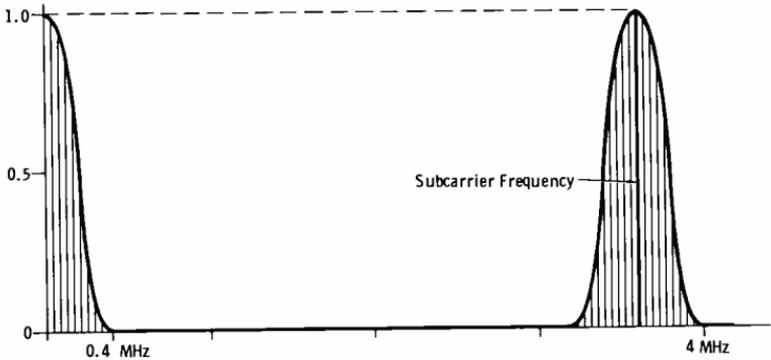


Fig. 2-17. Spectrum of modulated 20T pulse.

More recently, the modulated  $12.5T \sin^2$  pulse has been developed. The h.a.d. is  $12.5T = 12.50(0.125 \mu\text{s}) = 1.56 \mu\text{s}$ , and this pulse is extremely sensitive to delay distortion. The frequency spectrum of the  $12.5T$  pulse extends to  $1/(1.56 \mu\text{s}) = 640 \text{ kHz}$ . In practice, both the  $12.5T$  and  $20T$  modulated pulses are used, depending on the choice of the user.

## 2-5. BASIC TEST SIGNALS

The characteristics of modern tv transmission test signals are always the same and are fixed by definition for each type, just as is the case for the VU meter in audio applications. From this standpoint, they suggest a step in the right direction for obtaining "standard" test signals.

**NOTE:** The following descriptions cover each "standard" type of test signal as generated by the source, so that the exact specifications can be made clear. This serves as a reference for the reader when certain distortion characteristics are mentioned in other parts of the text. The actual results of such distortions on the individual test signals are detailed in Chapters 8, 10, and 11.

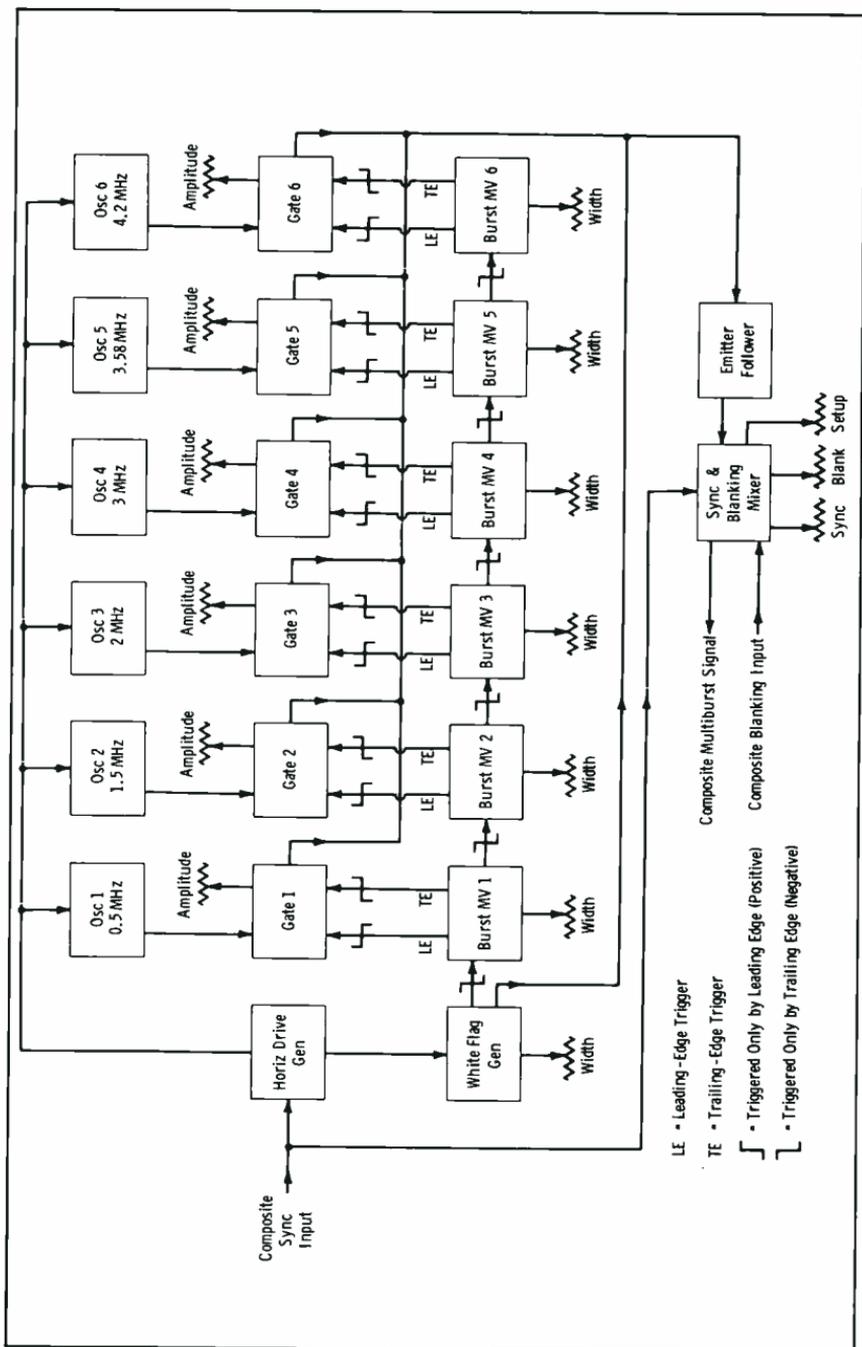


Fig. 2-18. Basic block diagram of multiburst generator.

### The Multiburst Signal

The multiburst signal consists of a white-flag reference pulse, followed by six bursts of individually keyed sine waves in ascending frequency order. This signal is useful as a "quickie" system check for rapid visual presentation of amplitude-frequency response, usually to 4.2 MHz.

The basic block diagram of a multiburst generator is shown in Fig. 2-18. The unit normally receives composite sync and composite blanking from the local sync generator. For field use, self-contained sync usually is available.

Horizontal drive is derived from incoming sync and used to trigger and synchronize six individual (and continuously running) sine-wave oscillators at specified frequencies. It also is used to generate a white-flag pulse at the beginning of active horizontal sweep.

The white-flag pulse width normally is adjusted to about  $8 \mu\text{s}$ . The trailing edge of this pulse triggers burst multivibrator MV1 on; this multivibrator is set to produce a pulse width of about  $7 \mu\text{s}$ , as are all following multivibrators. The leading edge of the pulse from MV1 gates the signal from oscillator 1 on, and the trailing edge gates the same signal off. The trailing edge also turns MV2 on, and so on. Thus, the output signal consists of the white flag and six sine-wave bursts on a time-shared basis during each line interval.

Fig. 2-19 gives the specifications of the multiburst signal. The frequencies shown are those normally used. The amplitudes are given both in IEEE units (where 140 IEEE units = 1 volt) and in voltage units.

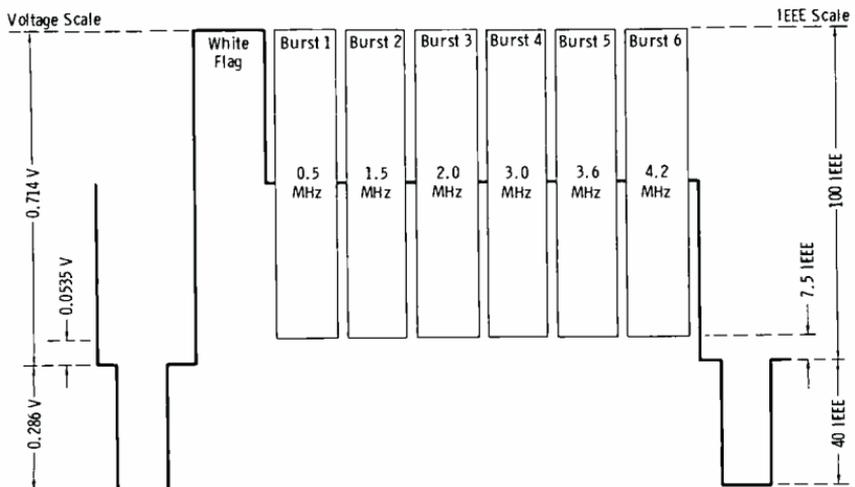


Fig. 2-19. Standard multiburst signal.

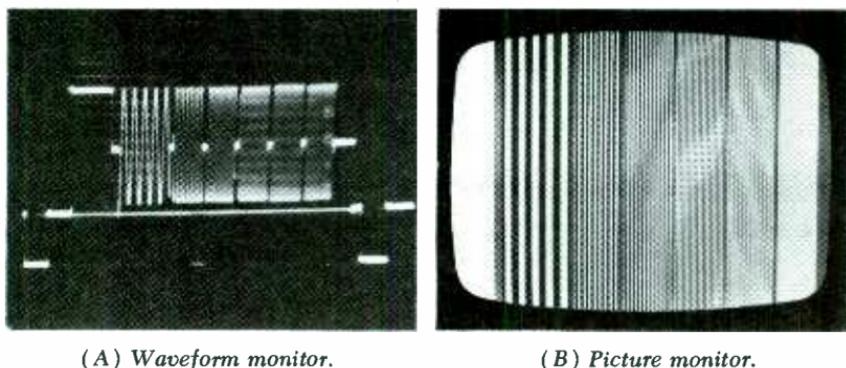


Fig. 2-20. Displays of multiburst signal.

Fig. 2-20A shows the cro trace of the standard multiburst signal at the output of the generator. Fig. 2-20B shows the appearance of the signal on the picture monitor. Moiré in Fig. 2-20B results from halftone processing of the photograph for printing.

### The Color-Bar Signal

The generation of the standard color-bar signal has been adequately covered elsewhere<sup>1</sup> and will not be repeated here. For reference purposes, Fig. 2-21 provides a review of the basic specifications of the split-field standard color-bar signal. The following notes apply:

1. Solid (bold) lines in the idealized displays indicate levels of the luminance signal.
2. Shaded areas indicate color-subcarrier envelope levels.
3. Measurements are made by using a standard IEEE scale in which reference white equals 100 units and sync equals -40 units.
5. (A) Amplitude tolerance of all luminance values at origination point is  $\pm 2.5$  IEEE units.  
(B) Amplitude tolerance of all peak-to-peak subcarrier values at origination point is  $\pm 2.5$  IEEE units.  
(C) Phase tolerance of colors at origination point is  $\pm 2.5^\circ$ .
6. Nominal values of luminance and peak-to-peak chrominance amplitudes for fully saturated color bars, 75 percent of full amplitude, using 10 percent and 7.5 percent setup are as in Table 2-2.

<sup>1</sup> Harold E. Ennes, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals* (Indianapolis: Howard W. Sams & Co., Inc., 1971).

7. The duration of each of the primary (and complementary) bars is to be  $\frac{1}{7}$  of the active portion of a scanning line within a tolerance of  $\pm 10$  percent.
8. The rise and fall times of the luminance-signal component shall not exceed 0.2 microsecond as measured at the point of generation.

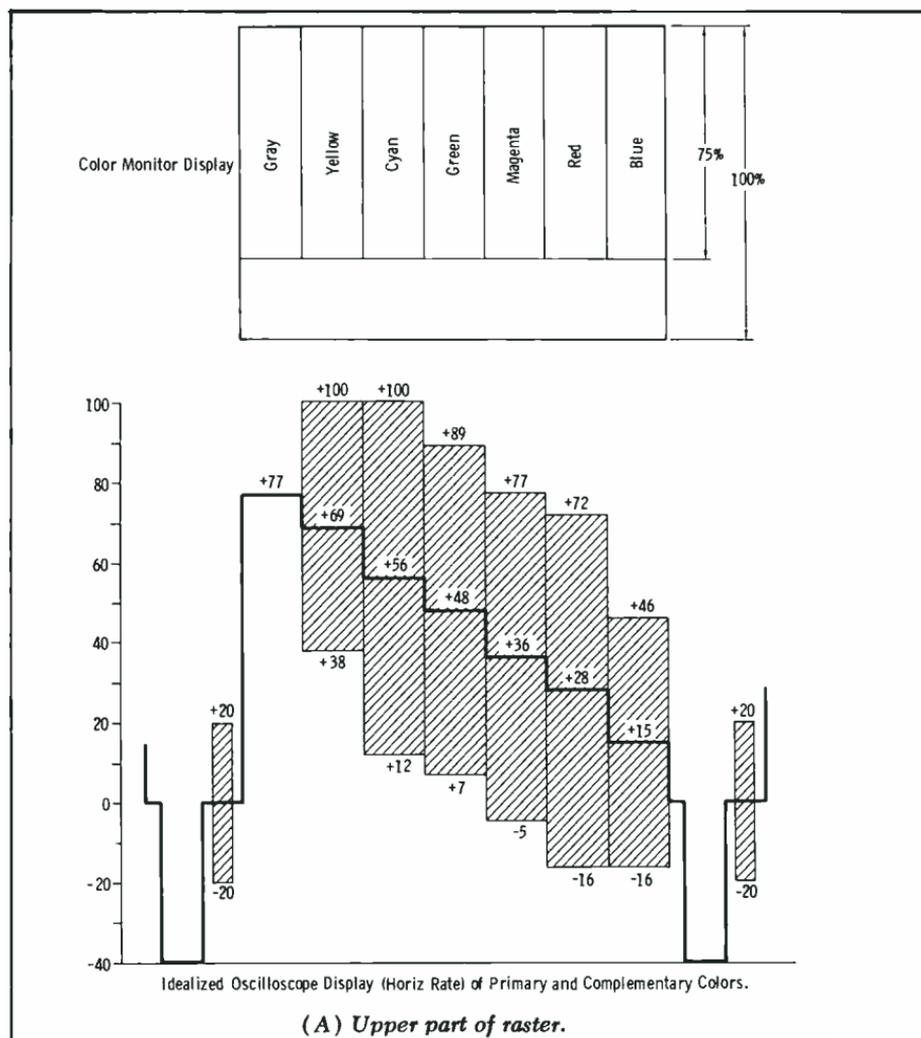
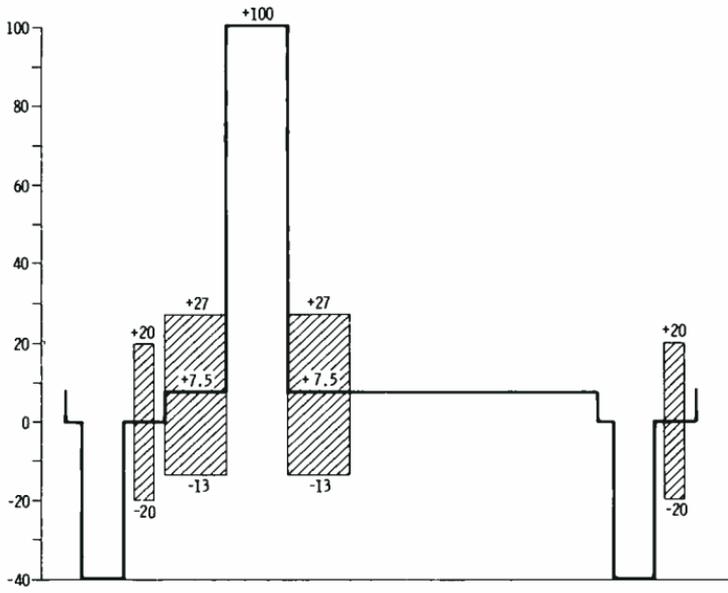
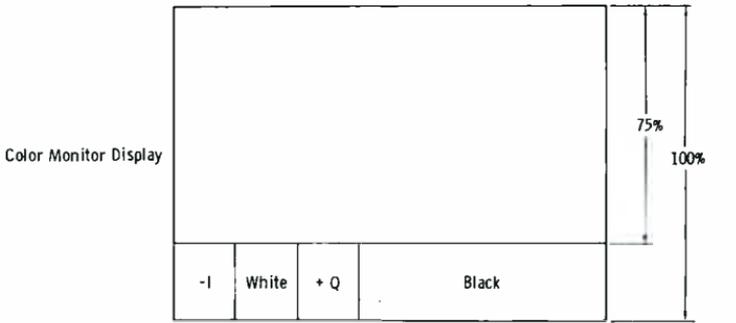


Fig. 2-21. Standard encoded

NOTE: Many older color-bar generators are adjusted to 10-percent setup rather than the revised 7.5-percent setup. Table 2-2 specifies the difference in peak-to-peak amplitudes of chroma in IEEE units for the two conditions.

**The Gray-Scale (Stairstep or Sawtooth) Signal**

The gray-scale test signal is used to explore the entire region from black (or blanking) to white by means of a linear variation of



Idealized Oscilloscope Display (Horiz Rate) of I and Q Test Signals and Reference White Bar

(B) Lower part of raster.

Reference—RETMA Engineering Committee TR-4 on Television Transmitters, EIA Standard for Encoded Color Bar Signals: RS-189, p. 3; Revised RS-189, pp. 1453B: 3a-4:5/66 and 1453B:4a-4:5/66.

color-bar signal with 7.5 % setup.

amplitude with time. It is useful in measuring nonlinear distortion in the system transfer characteristic.

Either a stairstep with well controlled rise-time steps, or a line-duration sawtooth is suitable for such measurement. A stairstep signal of ten steps repeated every line is shown in Fig. 2-22A. Small nonlinearities throughout the gray scale are more easily observed on this type of signal than is possible with the plain sawtooth signal, since each of the ten steps falls exactly on a graticule line when the IEEE scale is used.

**Table 2-2. Amplitudes for Split-Field Color-Bar Signal**

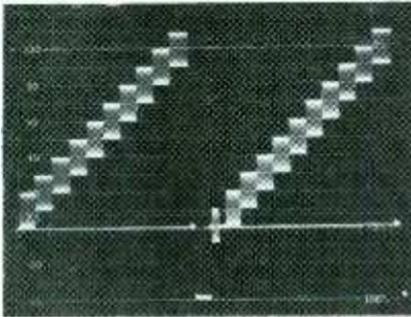
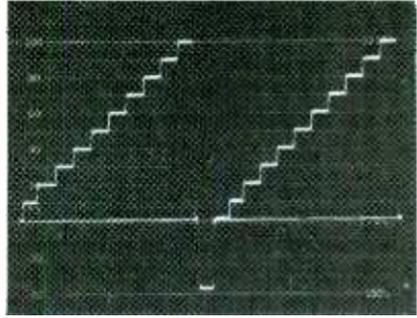
	10% Setup		7.5% Setup	
	Luminance	Peak-to-Peak Chrominance	Luminance	Peak-to-Peak Chrominance
Setup	10		7.5	
White	100		100	
Gray	77		77	
Yellow	70	60	69	62
Cyan	57	86	56	88
Green	50	80	48	82
Magenta	38	80	36	82
Red	30	86	28	88
Blue	17	60	15	62
-I	10	40	7.5	40
-Q	10	40	7.5	40

Sensitivity of measurement for either the stairstep or sawtooth is increased by superimposing 20 IEEE units of 3.58-MHz subcarrier as shown in Fig. 2-22B. Any nonlinearity then results in amplitude variations of the pulses when observed through a high-pass filter. This technique also allows measurement of nonlinearity at the color-subcarrier frequency relative to low-frequency steps. Such distortion is termed *differential gain*. Fig. 2-22C illustrates the high-pass waveform of Fig. 2-22B at the output of the generator.

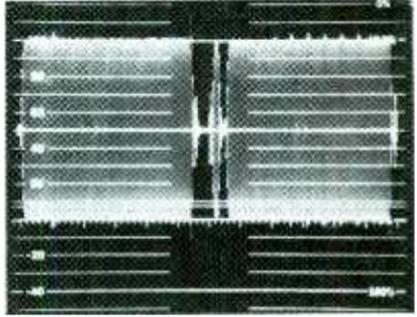
Gray-scale test signals must be able to convey information at low and high frequencies over every possible picture value likely to be encountered. This picture value is interpreted not only by amplitude, frequency, and phase response of the system but also on a widely varying duty cycle. Duty cycle in pulse work simply correlates the pulse duration with the pulse-repetition frequency (prf):

$$\text{Duty cycle} = \text{pulse duration} \times \text{prf}$$

(A) Stairstep-signal waveform.



(B) 3.58 MHz superimposed.



(C) Through high-pass filter.

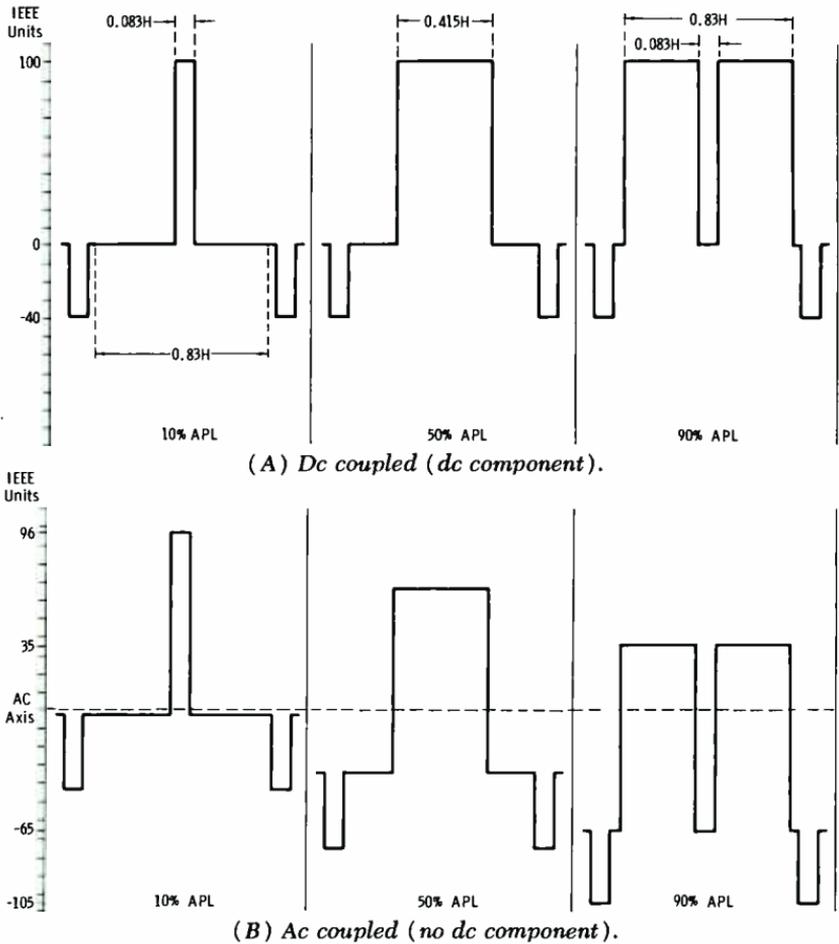
Courtesy Tektronix, Inc.

**Fig. 2-22. Waveform-monitor displays of standard stairstep signal.**

For television, this effect is most appropriately termed *average picture level* (APL), and the amplitude, frequency, and phase response of the system must be held within tolerable limits over the gamut of APL's encountered in practice.

Even experienced engineers sometimes forget that a 1-volt peak-to-peak video signal must be transferred through amplifiers capable of handling twice this range with little degradation (Fig. 2-23). Although the dc component is restored at such points as blanking insertion, sync insertion, gamma-correction stages, transmitter modulator, etc., practically all stages in between, as well as distribution and processing amplifiers, are ac-coupled. Table 2-3 tabulates IEEE units to APL for ac-coupled amplifiers when the average ac axis is arbitrarily assigned zero units. Note that the total excursion of the signal is 201 units instead of the normal 140 units.

Waveform monitors such as those used in master-monitor positions employ clamping circuits to hold blanking level at the reference graticule line. Some scopes designed for waveform monitoring (such as the Tektronix Type 529) allow switchable operation, either clamped or unclamped. Even though the monitoring cro is clamped,



**Fig. 2-23. Variation of signal excursions with APL.**

"bounces" of a momentary duration will occur upon drastic scenic changes in APL, and this is normal. Most scopes employed for routine testing and servicing do not use dc restorers or clamping circuits.

Fig. 2-24 illustrates a sine wave of 15,750 Hz, keyed with sync and phased as shown to represent one cycle of transition (black to white to black) per line. A 3.58-MHz wave is superimposed at 20 IEEE units. This signal is sometimes used by the Long Lines Radio & TV Division of AT&T for testing purposes.

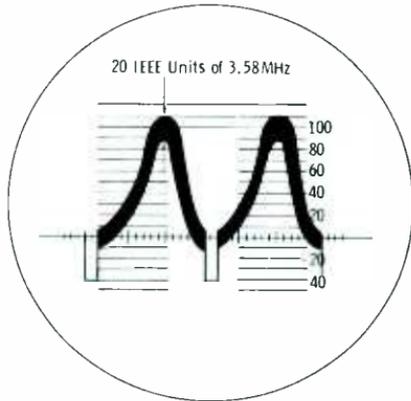
With the sine-wave test signal or sawtooth test signal with APL (Fig. 2-25), the data are processed so that the results reported for each APL condition correspond to the levels shown in Table 2-4. In practice, the 50-percent APL condition is obtained with either

**Table 2-3. IEEE Units to APL for AC-Coupled Amplifiers With AC Axis at 0 IEEE Units**

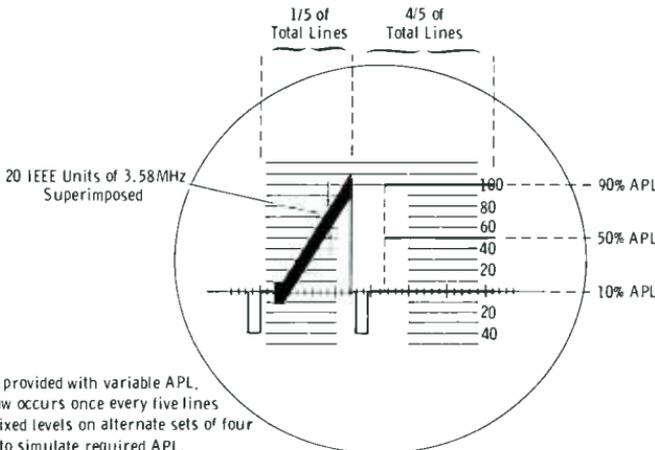
APL	10%	50%	90%
White Tip	96	65	35
Blanking Level	-4	-35	-65
Sync Tip	-44	-75	-105

Total excursion of signal (-105 to +96) = 201 IEEE units.

**Fig. 2-24. Line-period sine-wave signal with 3.58-MHz signal superimposed.**



signal repeated on successive lines at the normal transmission level. With the presence of sync pulses with 40 IEEE units peak amplitude, the 15,750-Hz sine-wave signal requires an amplifier with a capability of 184 IEEE scale units as shown by Table 2-4 for tests to 10-



When provided with variable APL, the saw occurs once every five lines with fixed levels on alternate sets of four lines to simulate required APL.

**Fig. 2-25. Sawtooth signal with 3.58-MHz signal superimposed.**

percent APL. For the sawtooth (and stairstep) wave with horizontal and vertical blanking accounted for, the corresponding peak amplitude is 201 IEEE scale units. Note that in any case the 20 IEEE units of superimposed 3.58-MHz signal results in an additional 10 scale units over the peak low-frequency level.

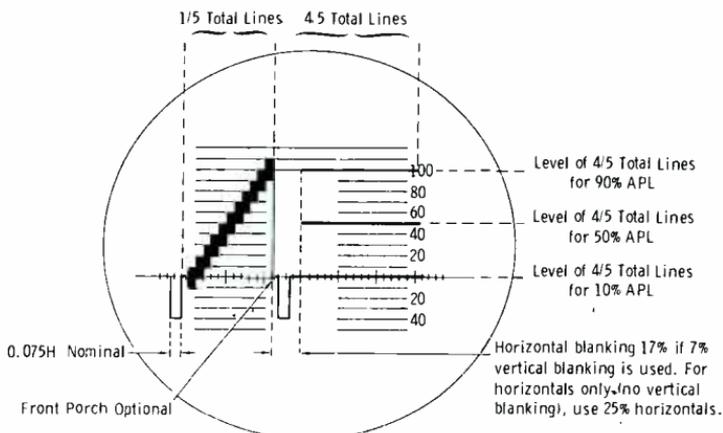
**Table 2-4. Significant Range (IEEE Units) (Relative to Blanking Level in Sawtooth, or Base of Sync on Sine Wave)**

APL	Sine Wave & Sync Pulse (No Blanking)	Sawtooth Wave With Sync Pulse and Blanking
10%	84-184	61-161
50%	53-153	30-130
90%	23-123	0-100

NOTE: 3.58 MHz to be retained at 20 IEEE units for all levels of low-frequency test signal.

Fig. 2-26 presents the specifications for the stairstep-generator test signal, when provided with variable APL. This generator sometimes includes provisions for inserting horizontal sync only. In this case, the blanking width (which is adjustable internally) should be set for 25 percent of a line period (15.8 microseconds). If fed through equipment where composite station blanking is inserted, the normal station blanking pulse should cover the test generator blanking output (11.4  $\mu$ s max horizontal with 7-percent vertical blanking).

All of the waveforms discussed (sine wave, sawtooth, and stairstep) inherently have a 50-percent duty cycle when presented line for line. The required variation in APL is obtained by presenting this 50-percent APL signal for only one-fifth of the total active scan-



**Fig. 2-26. Stairstep signal with 3.58-MHz signal superimposed.**

ning time in each field period. The remaining four-fifths of this time is occupied by a constant, low-frequency level which is set at blanking for 10-percent APL, at 50-percent of reference white for 50-percent APL, and at reference white level for 90-percent APL.

Fig. 2-27 illustrates the output of a staircase generator with variable APL adjusted to equal 10-percent APL (upper trace). The lower trace shows the same signal passed through the high-pass filter. The generator output should show zero differential gain. The

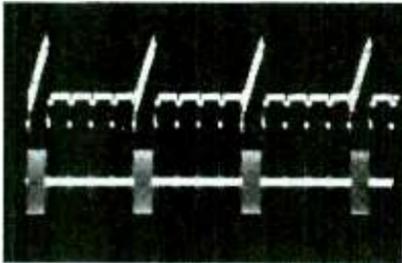


Fig. 2-27. Output of staircase generator with variable APL set to 10%. (Upper trace: generator output. Lower trace: through high-pass filter.)

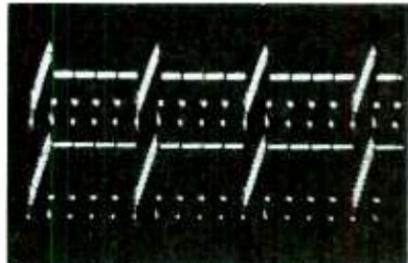


Fig. 2-28. One-line-in-five staircase signal for 50% APL (upper trace) and signal of similar composition for 90% APL (lower trace).

upper trace of Fig. 2-28 shows the adjustment for 50-percent APL; the lower trace is the adjustment for 90-percent APL. The direct output of the generator should indicate zero differential gain at all three levels of APL.

NOTE: Most recent staircase generators simply use the regular line-for-line (repeated every line) signal for the 50-percent APL measurement, as in Fig. 2-22B.

### The $\text{Sin}^2$ -Window Signal

Development of the  $\text{sin}^2$  pulse has been described. This signal is normally accompanied by a half-line and half-field *window* pulse, which is sometimes termed a *bar*.

Fig. 2-29 shows a basic block diagram of such a generator which also includes the modulated 20T pulse. The timing-circuit driver (monostable multivibrator) is triggered from the leading edge of sync and generates a rectangular pulse of about 16- $\mu\text{s}$  duration. The trailing edge of this pulse initiates the operation of the pulse and window timing circuit, which positions the pulse and window leading and trailing edges relative to sync. Blanking pulses are used to inhibit the timing-circuit action during field blanking. The output of the impulse generator is an 18-ns "spike" which becomes the T pulse after shaping in the T-pulse shaping network. A switch is

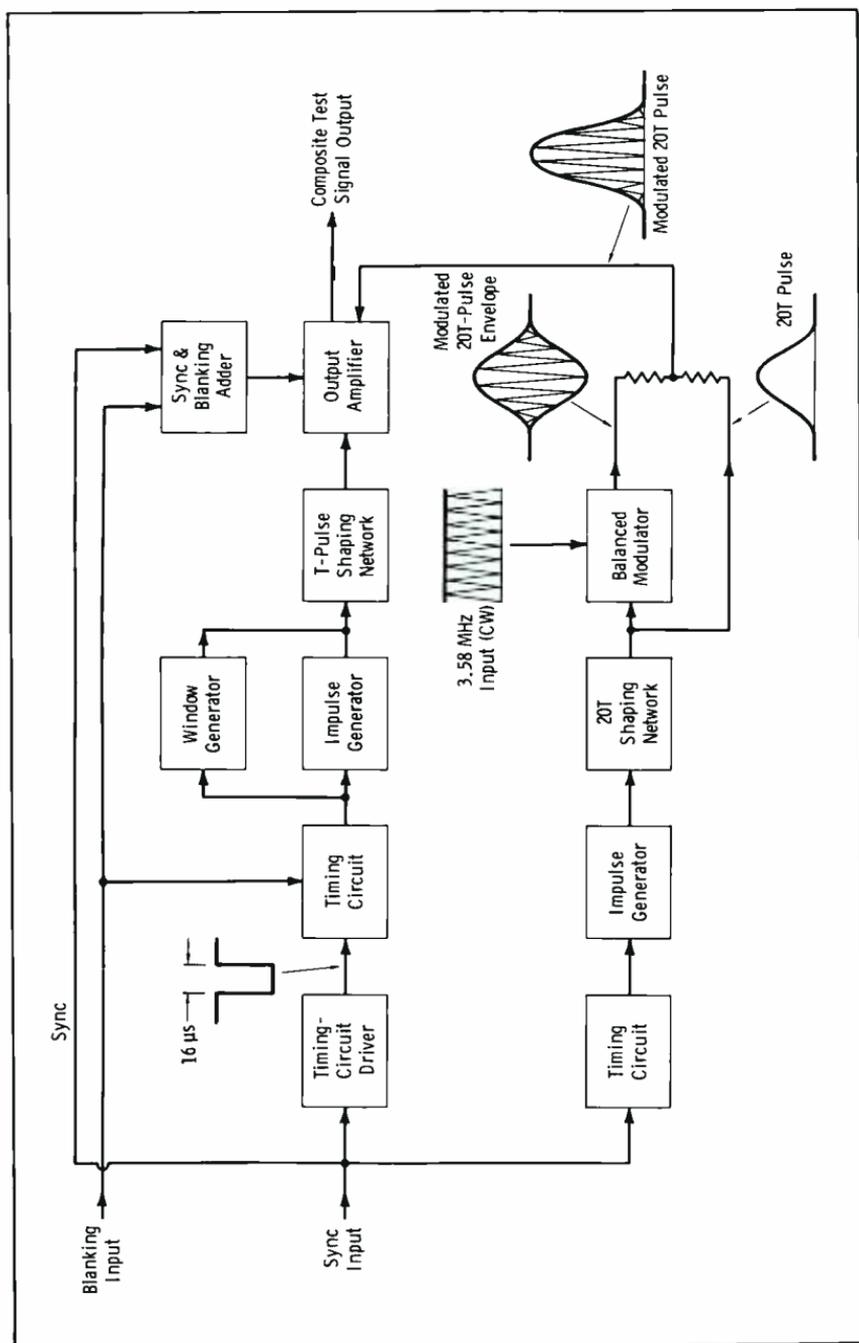
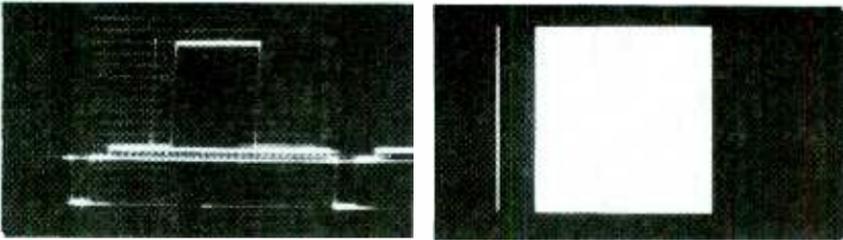


Fig. 2-29. Basic block diagram of  $\sin^2$ -pulse and window generator.

normally provided so that either  $2T$ ,  $T$ , or  $T/2$  pulses are available. Note also that the leading and trailing edges of the window signal, since they pass through the same shaping filter, have the same rise and fall times as the associated  $T$  pulse.

The  $20T$  pulse is shaped by appropriate  $\sin^2$  filters and applied to a doubly balanced 3.58-MHz modulator in a manner similar to that in which chroma information modulates the color subcarrier in an encoder. Thus both the 3.58-MHz carrier and the original  $20T$  pulse are cancelled, and the output is only the product, or the modulated sidebands of the carrier. This produces the modulated  $20T$ -pulse envelope shown in Fig. 2-29. Finally, the original  $20T$  pulse is linearly (resistively) added to the modulated pulse, producing the symmetrical pulse with a base line.

The horizontal-rate display of the conventional pulse-window waveform at the generator output is illustrated in Fig. 2-30A. The video-monitor display (Fig. 2-30B) consists of a thin line followed by the white "window" of about one-half of the active duration and one-half of the active field duration.



(A) Waveform at generator output.

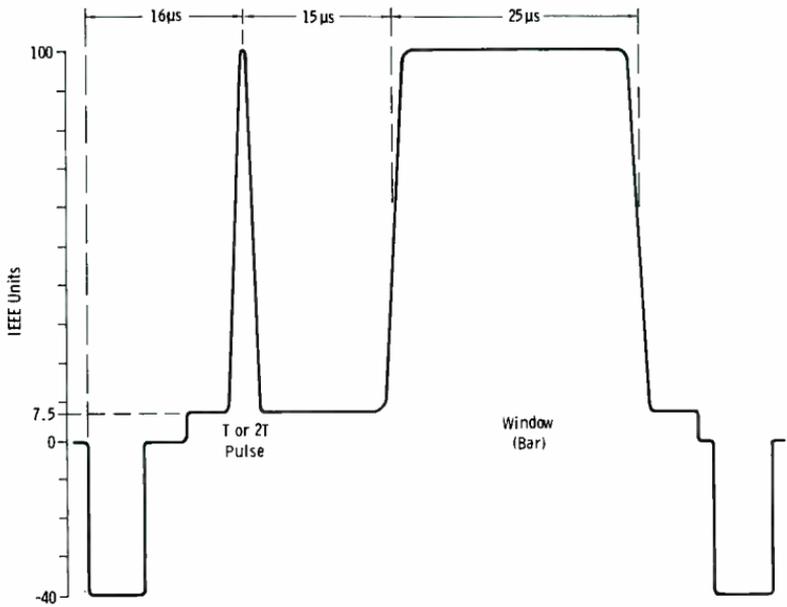
(B) Display on picture monitor.

**Fig. 2-30. Displays of  $T$  pulse and window.**

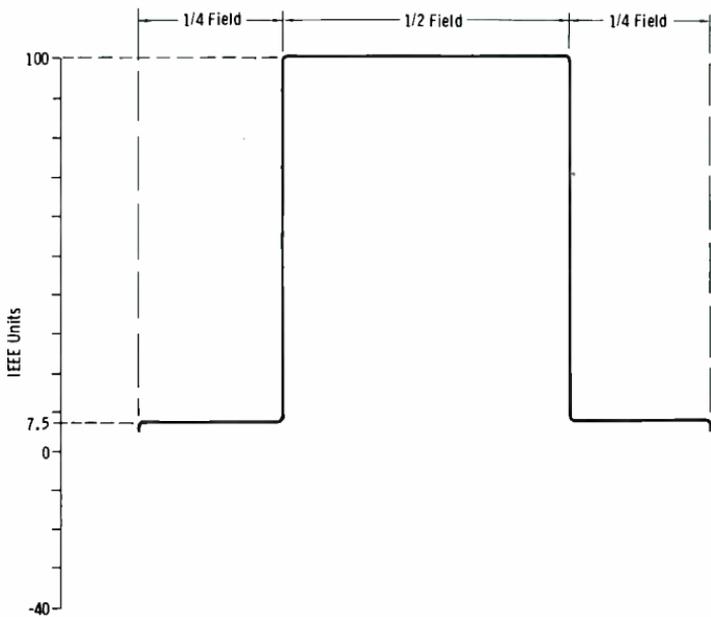
Fig. 2-31A gives the line-rate specifications of the standard pulse-bar signal, with relative timing from the leading edge of horizontal sync. Fig. 2-31B gives the field-rate specifications of the same signal.

Fig. 2-32 illustrates the addition of the modulated  $20T$  pulse to the composite test signal. Fig. 2-32A is the display of two consecutive lines in which the window occupies one line and the pulses are contained in the following line. In Fig. 2-32B, the pulses and window are generated in each single line. In some generators, the positions of the pulses are interchanged; i.e., the  $T$  or  $2T$  pulse precedes the modulated  $20T$  pulse as shown in Fig. 2-33.

The type of display shown in Fig. 2-33 is convenient for one of the tests associated with this signal; the top of the window provides a reference for comparing the relative amplitudes of the window and pulses. When all the pulses and the windows are in a single line, the scope must be double-triggered; that is, it must be triggered



(A) Line rate.



(B) Field rate.

Fig. 2-31. Pulse-window signal displays.

from successive sync pulses. The two-consecutive-line signal (Fig. 2-32A) eliminates the need for double-triggering, since a repetitive sweep automatically provides the double-triggering display. However, the two-consecutive-line signal has the disadvantage of being subject to error from frequency distortion because of the large difference in APL between the two separate lines (window on one line and pulses on the other).

A convenient method for triple-triggering for a still more effective display is suggested by the Australian Broadcasting Commission; this method is illustrated in Fig. 2-34. The unit strips sync pulses from the incoming signal and uses them to fire delay multivibrators (Fig. 2-34B) for producing cro trigger pulses (Fig. 2-34C). Trigger 1 displays the pulse, trigger 2 starts the leading-edge bar display, and trigger 3 starts the trailing-edge display. The multivibrator delay between times 2 and 3 can be made variable so that the pulse height can be used as a "pointer" to detect line tilt should the height of the bar vary along its length. When the bar height is constant along its length, there is no separation along the top of the resulting waveform (Fig. 2-34D); the waveform shown in this example indicates tilt by the separation of the top lines.

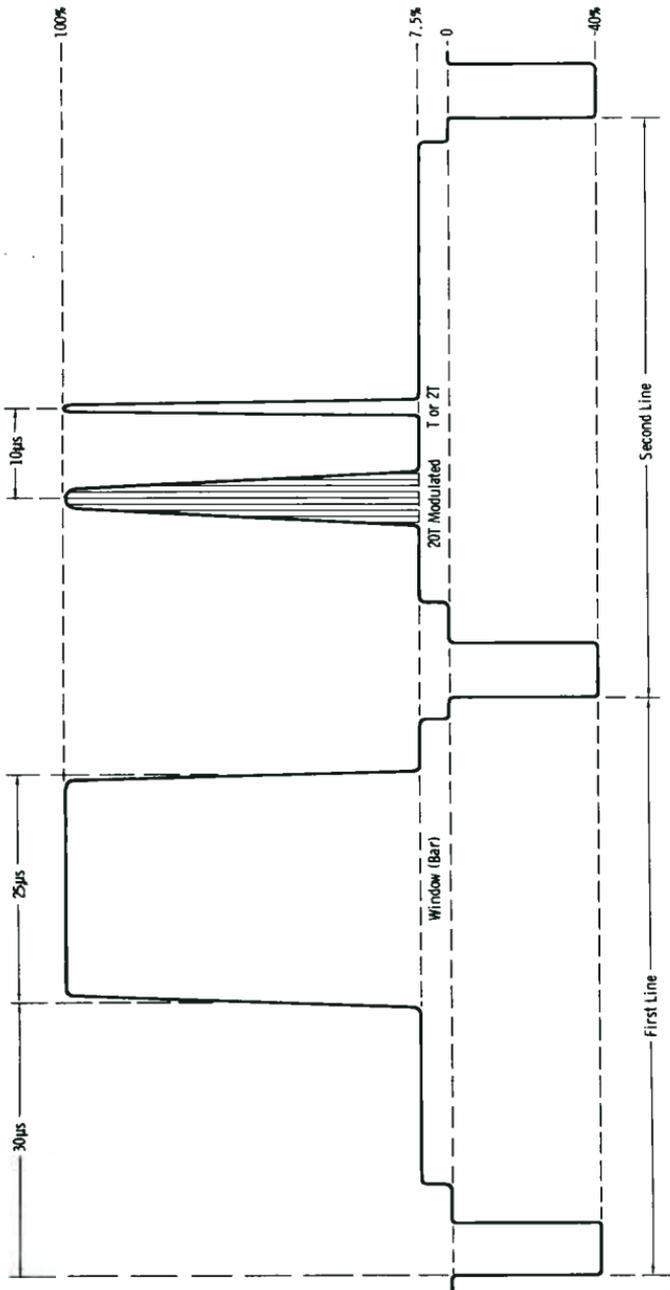
NOTE: Some waveform generators have a special cro-trigger output for either double- or triple-triggering requirements. Many of these, however, are custom-built at the time of this writing.

Use of the pulse-window signal in practice involves a special graticule to indicate certain *K-factors*, particularly for routine testing to provide a quick observation to go/no-go quality. This is discussed in applicable portions of the text in the following chapters.

### The Vertical-Interval Test Signal (VITS)

In the standard television signal there is, immediately following each 9H vertical-synchronizing interval, a series of 11 or 12 horizontal lines which carry no video information. In a receiver or monitor, these lines provide an interval which ensures that the vertical retrace will be completed before any video information is received. The lines are intended to perform no other function and normally appear on a monitor as a gray band across the top of the picture; usually, this band is adjusted so that it is behind the mask. While these are actually horizontal lines, they are conveniently considered as a part of the vertical synchronizing and blanking interval.

By using suitable keying equipment, it is possible to introduce information onto one or more of these blank lines to be transmitted to a specific destination. If desired, this information can be blanked out at the receiving point before the signal is put on the air, or, since it is not normally visible on a receiver, it may be broadcast. Tele-



(A) Consecutive-line display.

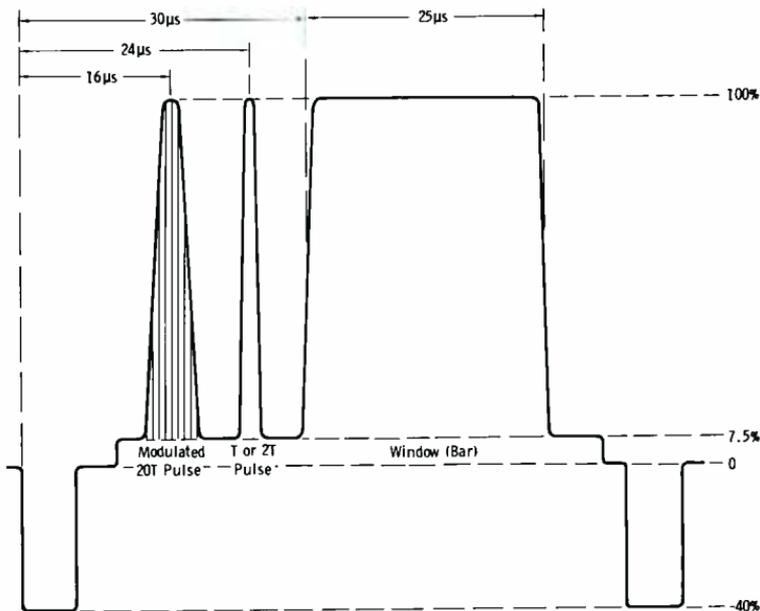
Fig. 2-32. Displays of 20T pulse,

vision test signals are transmitted on certain of these lines, and these signals are designated *vertical-interval test signals*, or VITS. The usual appearance of such test signals on an underscanned picture monitor or receiver is shown in Fig. 2-35. The major networks and the Bell System, working through the Network Transmission Committee, are still working toward the development of standard vertical-interval test signals and tolerances for various transmission systems.

The advantage, of course, of introducing these signals in the vertical interval is that a system can be checked, in service, while a program is actually being transmitted. To this end, it is desirable that all these signals be available for nearly simultaneous observation, and this is the objective.

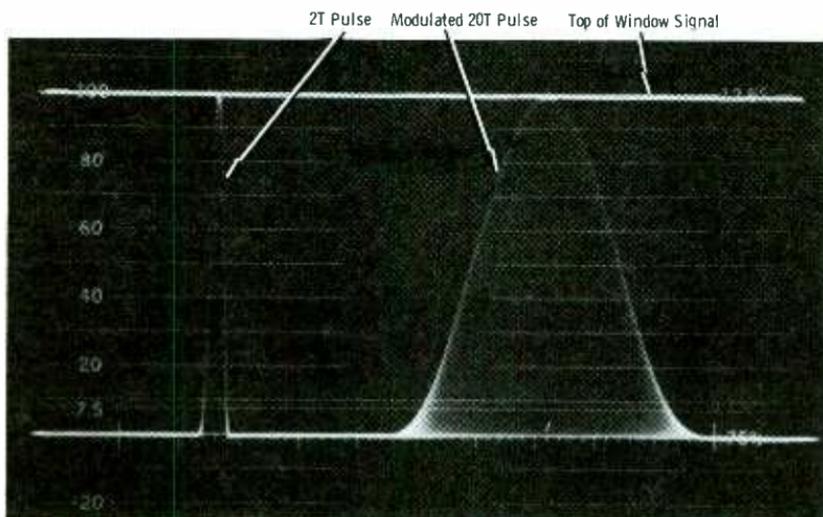
*Test Signals*—While rigid specifications and standards for VITS have still not been established at the time of this writing, the following excerpts from the Network Transmission Committee Engineering Report No. 5 serve as an indication of the types of signals and purposes intended:

It is the objective of the television broadcast industry to provide to home viewers television program signals of the highest possible quality consistent with the state of the art and the economic factors involved. This level of quality is determined, at any given time, by standards established within the industry and judged to be proper



(B) One-line display.

T pulse, and window.



Courtesy Tektronix, Inc.

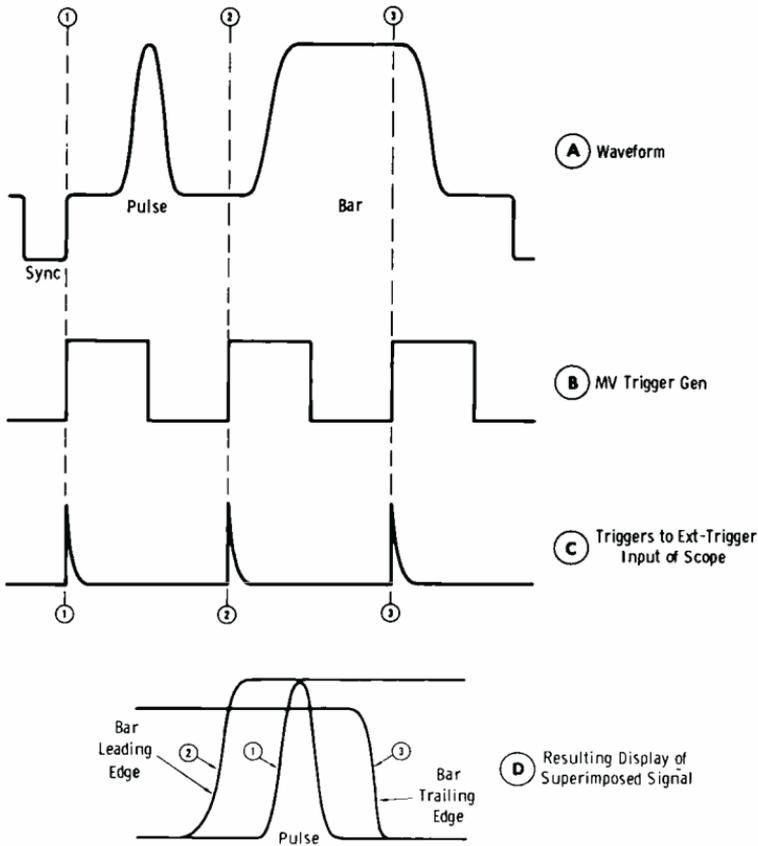
**Fig. 2-33. 2T pulse and modulated 20T pulse displayed on waveform monitor.**

and adequate and in conformance with the rules of the Federal Communications Commission. Since the television art is still developing, the standards are subject to review from time to time and to change when necessary, particularly as dictated by technical advances.

The overall system from source to home receiver contains many interconnected but nevertheless independent elements. Obviously, when an overall quality objective is established, each element in the system must be designed so that, within its normal operating range, it is compatible with the overall objective and with the capabilities of all the other elements.

A standard performance can most conveniently be stated in terms of permissible deviations from perfect transmission. Permissible deviations for an overall system having been stated, it is implied that each link in an overall system will be assigned a portion of the overall tolerances. This allocation must be a matter of mutual agreement, based on engineering judgment of the capabilities of each link. Then, if, because of inherent technical limitations, one element requires a disproportionate share of the overall tolerance on any stated characteristic, this can be balanced by assigning a smaller share to a more favorably disposed element.

As the television art progresses, so also will progress be made in test instrumentalities, methods, and techniques. Therefore, not only are quality standards subject to change, but also the terms in which they are defined and the methods by which they are measured.



Reference—Laboratory Report No. 25; Engineering Div., Australian Broadcasting Commission.

**Fig. 2-34. Method of triple-triggering oscilloscope.**

The performance objectives which are given in this report have been derived in accordance with the foregoing. They are neither design objectives nor maintenance requirements. They are a statement of the expected transmission performance of intercity network facilities which should be achieved a high percentage of the time, derived from current practice and stated in terms of tolerances on specific test signals which are presently accepted as providing significant results.

The following summary of performance objectives is given to provide a source of ready reference for the actual numerical values covered in the report. This summary states the requirements for each test signal as observed at any receiving point under the conditions stated in detail in the body of the report.

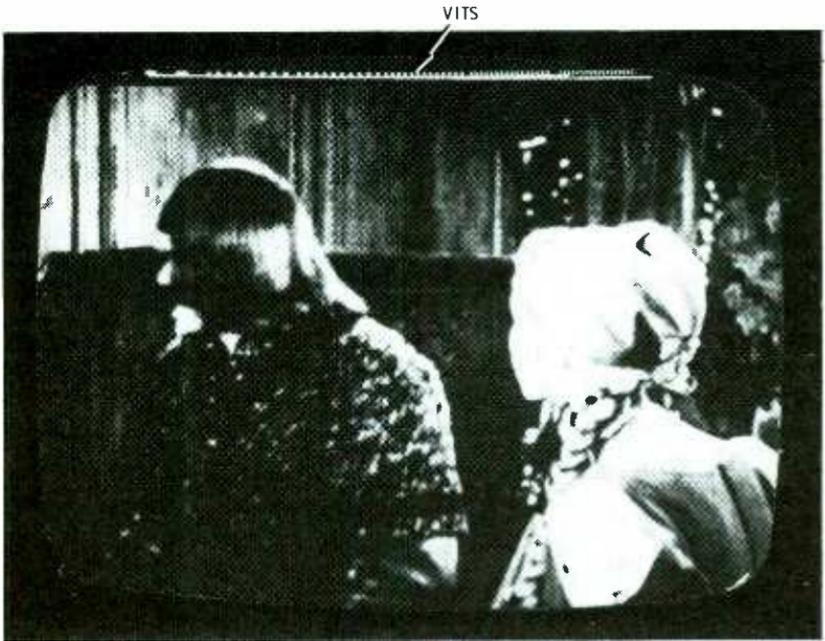


Fig. 2-35. Appearance of VITS on vertically underscanned monitor or receiver.

### MULTIBURST

With the white flag adjusted to 100 IEEE divisions:

1. Each burst shall be between 80 and 95 IEEE divisions in amplitude (approximately +6%, -11%).
2. The color-sync burst shall be between 36 and 44 IEEE divisions.

### STAIRSTEP

Differential gain shall not exceed 15 percent.

Differential phase shall not exceed  $5^\circ$ .

### T SINE-SQUARED PULSE AND BAR

The slope across the top of the bar shall not exceed 4 percent.

The peak-to-peak amplitude of the overshoots, either leading or following the sine-squared pulse, whichever is larger, shall not exceed 25 percent.

Detailed checking of transmission facilities by methods similar to those used for lineups is too time-consuming to be useful for rapid operational checks. For this reason, considerable effort has been

devoted to the development of a group of test signals which would give a reliable indication of the transmission quality of a system by means of viewing the received test signals on an oscilloscope. Study work is still continuing, both in Europe and in the U.S., on the development of a meaningful signal or group of signals. The performance objectives given in this report are based on observation of the following test signals which are currently being used in the U.S.:

- (A) Multiburst
- (B) Stairstep with 3.58 MHz superimposed on each step
- (C)  $T \sin^2$  pulse and bar

Each of these signals may be transmitted either full-frame or in the vertical interval.

A detailed specification for each of the test signals is given in Figs. 2-36 and 2-37. The differences between the VITS and the full-frame signals are not significant so far as test results are concerned, but they are pointed out here and in the illustrations.

On each line of the VIT signal, space is provided for a reference white signal and a reference black signal. The broadcasters may or may not elect to transmit this signal on any particular program. The full-frame signals may be obtained from the same test-signal generator and have the vacant space of the reference signals, or they may be obtained from separate generators which will fill in that space as shown in Fig. 2-37.

The VITS sine-squared pulse and bar signal occupies a single line, so the second signal is properly referred to as a bar. When used as a full-frame signal, the active signal is transmitted only in the middle half of each field, so the bar signal becomes a window signal. It is general practice, however, to refer to the signal in either case as the pulse and bar signal.

Signal generators capable of generating signals that meet these specifications, and keying equipment for inserting VITS, are commercially available. It shall be the responsibility of the transmitting office to maintain this equipment and supervise its adjustment frequently enough to ensure maximum accuracy of the transmitted signal at all times.

Vertical-interval test signals will be transmitted only from network sources. The Bell System has no plans to provide for the insertion of VITS at any of their test rooms. At any location where the networks originate VITS, permanent arrangements will be provided such that the signals will be applied at the interface between the final amplifier at master control and the transmitting local channel. In this way, no broadcasters' terminal gear will be reflected in network facility tests. If the black reference level and white reference level signals are not applied at this same point, supervision should be suf-

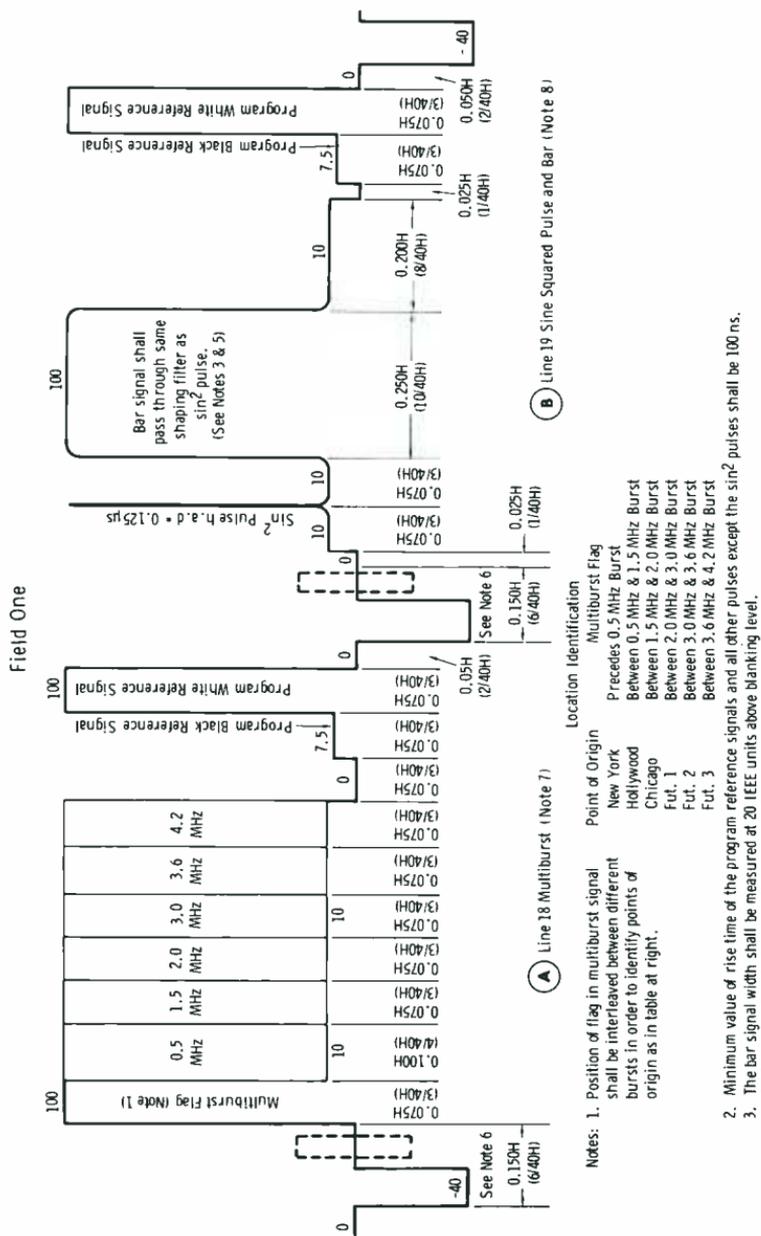
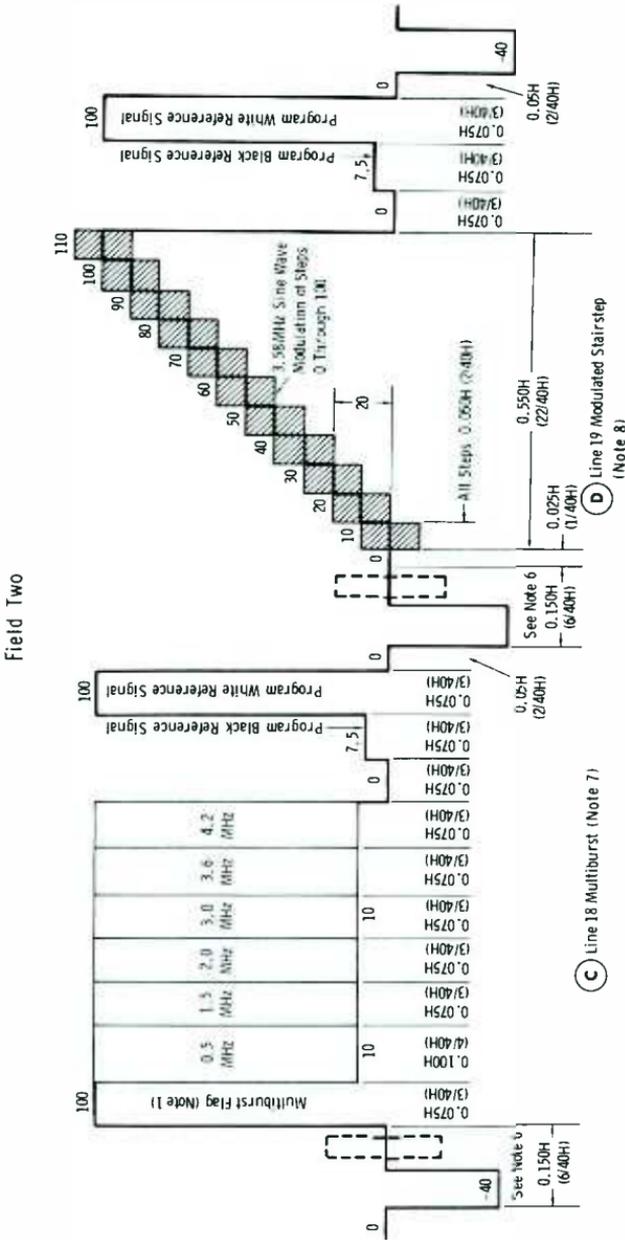


Fig. 2-36. Specifications for



From Network Transmission Committee, Engineering Report No. 5

vertical-interval test signals.

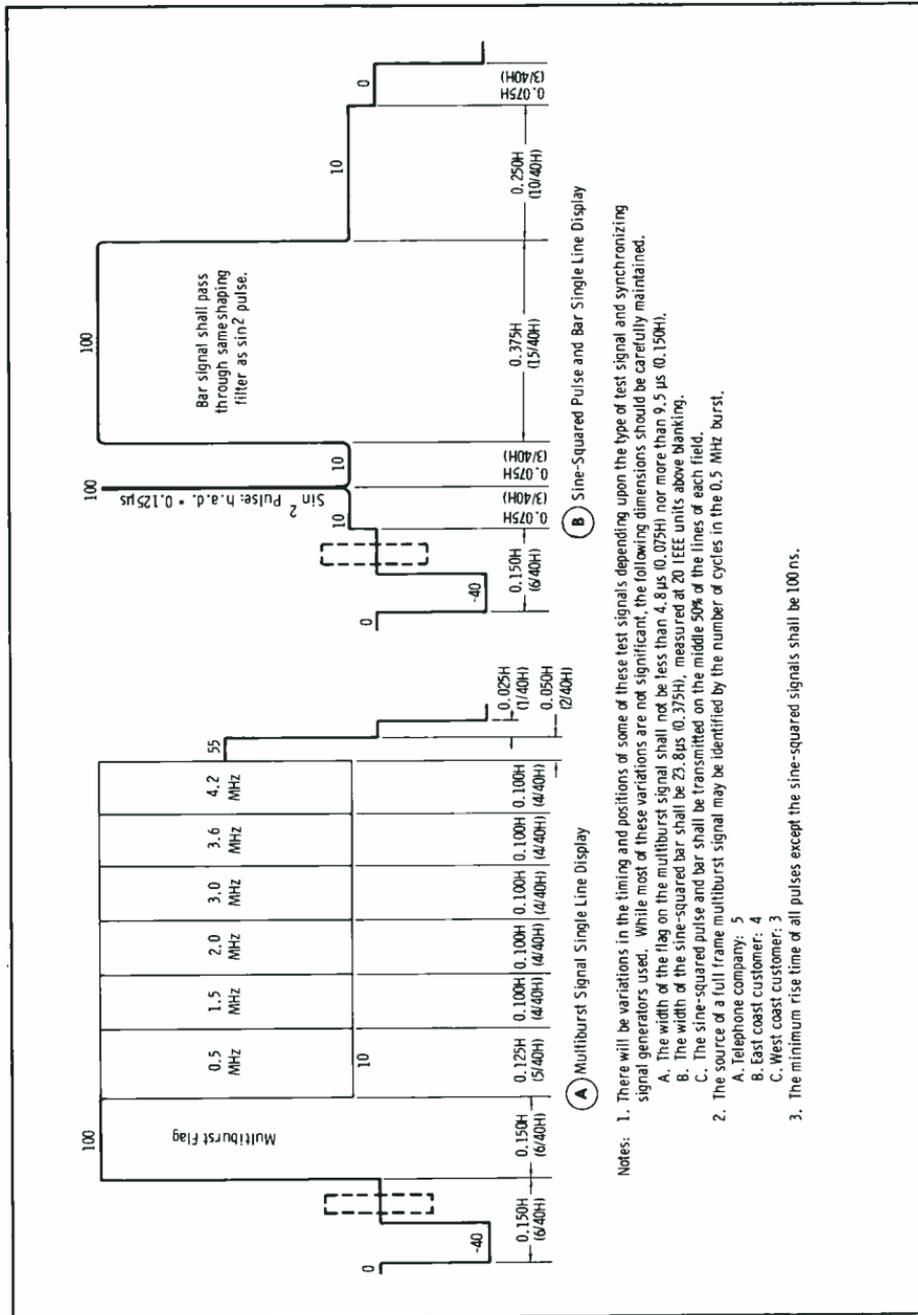
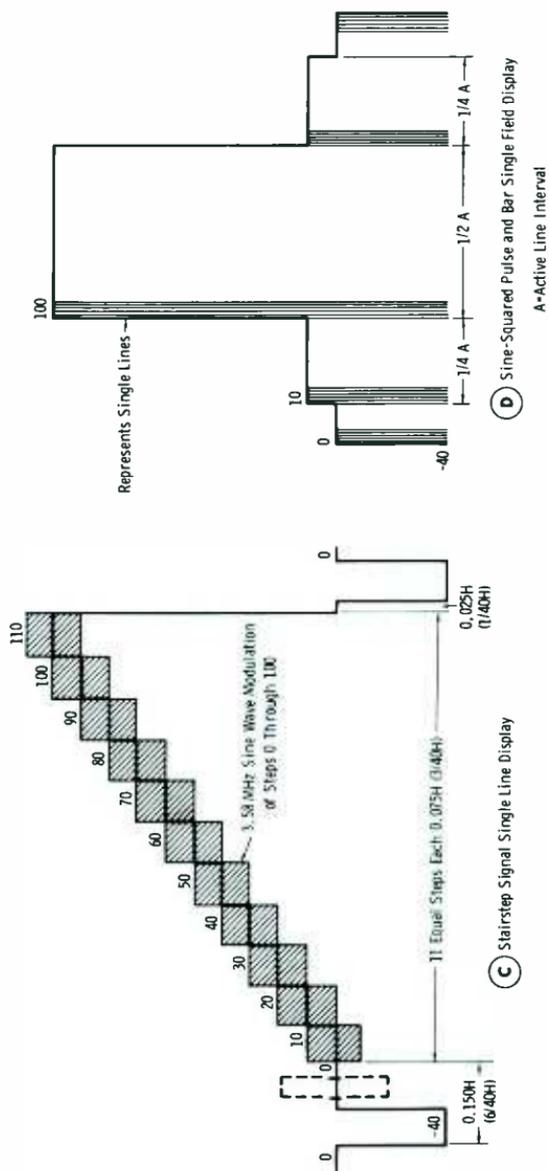


Fig. 2-37. Specifications for



ficient to ensure that levels are properly maintained at that point.

Full-frame test signals may be originated at any point where signal generators are available. The operating personnel have full freedom of choice of the points at which these signals may be connected to a network, but that choice must be made with full knowledge of the transmission conditions. If the signals are to be used to check the performance of overall networks or of network sections, they shall be applied at designated points where the level is normally 1 volt peak-to-peak and the facility is equalized flat. At the broadcasters' locations this would be the same point at which the VITS are inserted. In Bell System test rooms, this would be over standard transmit test trunks or at the transmitting-terminal video patch jacks in the television operating center.

If the signals are to be used as an aid in sectionalizing a trouble condition, they shall always be applied at a point that is normally equalized flat. Level adjustments to compensate for applying the signals at a point where the normal signal level is other than 1 volt peak-to-peak are permissible. No attempt should be made, however, to compensate for applying the signals at other than a flat-equalized point because this cannot be done with sufficient accuracy. Other methods must be used in such situations.

At receiving locations, the same general principles shall be applied. At station locations, fixed arrangements should be provided for making observations of network performance. The connection to the network facilities should be made directly at the interface with Bell System facilities, before any processing amplifier, if one is used, and through an isolation or distribution amplifier which will permit terminated measurements to be made. A monitoring point after the processing amplifier may also be useful.

In Bell System test rooms, observations shall be made over standard monitoring or test trunks or at the output of distribution amplifiers at designated section terminals where the level is normally 1 volt peak-to-peak and the facility is equalized flat. For sectionalizing trouble conditions, observations should preferably be made on a terminated basis and at points that are normally equalized flat. Adjustments may be made for measurements made at points where the normal signal level is other than 1 volt peak-to-peak, but no correction for nonflat equalization should be attempted.

NOTE: Techniques for the measurement of VITS are covered in Chapter 8 of this book.

## 2-6. THE VERTICAL-INTERVAL REFERENCE (VIR) SIGNAL

You will note from the description in Section 2-5 that the VIT signal is inserted at the network originating station between the

final output of master control and the Telco lines. Thus the VIT signals are designed to give diagnostic information primarily through network interconnection facilities. Obviously, the technician can observe the incoming network signal on his scope and compare it with the studio or transmitter output signal to evaluate his own transmission facilities in terms of degradation of the VIT signal, but this is a secondary consideration.

The VIT signal does not give any information at all about the quality or characteristics of the original signal. The VIT information may indicate perfectly acceptable transmission characteristics, yet the picture itself can have improper chroma amplitude or phase, improper setup, or improper sync amplitude. The *vertical-interval reference (VIR)* signal is used to "certify" the original signal as to luminance and chrominance, chrominance amplitude, and chrominance phase relative to burst.

The VIR signal is normally inserted at the output of the studio control room in which the program is originated and where all feeds involved in the particular program are adjusted for the desired color. Once the VIR signal is added in this manner, it will serve as a reference or certification for the program signal. The VIR signal was originally inserted on line 20 because the VIT signals were placed on lines 18 and 19. Effective November 15, 1975, the VITS were moved to lines 17 and 18, and the VIRS is now inserted on line 19 of both fields.

Fig. 2-38 shows the composition of the VIR signal. The first 24  $\mu\text{s}$  of active line interval is a chrominance reference bar consisting of chrominance subcarrier of the same phase as the color burst,  $-(B - Y)$ . The reference bar has an amplitude of 40 IEEU units (0.286 V) peak-to-peak, and it is set on a pedestal of 70 IEEU units (0.5 V). The purpose of the phase reference is to allow the phase of

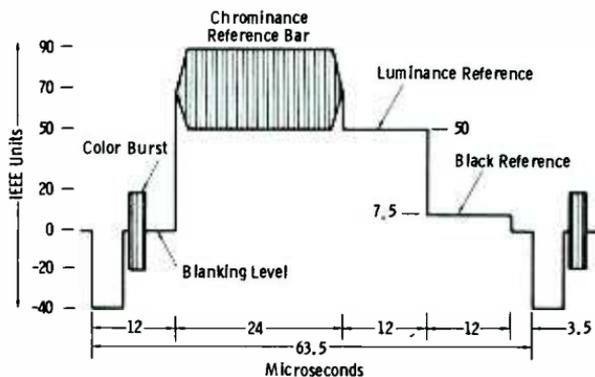


Fig. 2-38. The VIR signal on line 19.

the color burst to be corrected if necessary to agree with the phase of the chrominance reference bar.

The chrominance reference bar also provides a reference for the chrominance subcarrier amplitude. After the luminance amplitude has been correctly adjusted (see next paragraph), the reference bar should be 40 IIEEE units peak-to-peak. The purpose of this amplitude reference is to allow the chrominance subcarrier amplitude, and therefore the saturation, to be correctly adjusted and also to allow the amplitude of the color burst to be correctly adjusted.

Following the 24- $\mu$ s chrominance reference, there is a 12- $\mu$ s luminance reference (gray level) of 50 IIEEE units (0.357 V). This is followed by a 12- $\mu$ s black reference at 7.5 IIEEE units (0.0536 V). The purpose of these two reference levels is to allow the correct adjustment of luminance amplitudes and, if necessary, of the setup level and sync amplitude.

The rationale for this particular format of the VIR signal is as follows:

1. The signal does not use the full excursion from blanking to reference white so that a small amount of signal compression will not affect the reference to any great extent.
2. The reference bar of chrominance subcarrier is set on a pedestal of 70 IIEEE units, which is about the average luminance level of flesh tones. The level of 40 IIEEE units peak-to-peak is the amplitude of the color burst that occurs at blanking level. If there is nonlinear distortion in the system, adjustments can be made to minimize the distortion in the most sensitive area.
3. The luminance reference at 50 IIEEE units is also the level of the lower edge of the subcarrier chrominance reference bar. This allows a quick check of the relative levels of luminance and chrominance.
4. The black reference at 7.5 IIEEE units allows the normal setup level to be re-established.
5. The chrominance reference bar is placed at the beginning of the line so that the phase is least affected by any velocity errors in video tape recorders.

When the color program has once been certified at the studio output, either manual or automatic adjustment can be made at the receiving station for any transmission deficiencies that affect hue, saturation, luminance levels (including setup), and sync amplitude. Manual correction is accomplished with a processing amplifier that provides individual controls for chrominance amplitude and phase, luminance amplitude, setup, and sync amplitude. Automatic correction is provided by terminal gear that "looks" at the VIR signal and

automatically corrects all operating parameters necessary to recertify the program signal.

### EXERCISES

- Q2-1. When a receiver is properly tuned for vsb reception, where is the visual carrier frequency placed on the overall response curve?
- Q2-2. For the receiver in question Q2-1, what would the demodulated response be if the transmission were dsb?
- Q2-3. What is the expected resolution (horizontal and vertical) in tv lines at the output of a color transmitter?
- Q2-4. If the pickup tube at the studio had a strictly linear relationship of voltage output to brightness input, would gamma correction still be necessary?
- Q2-5. What test signal most clearly reveals amplitude nonlinearity?
- Q2-6. What test signal most clearly reveals field-rate distortion?
- Q2-7. What test signal most clearly reveals line-rate distortion?
- Q2-8. What test signal most clearly reveals transient response of a system?
- Q2-9. What test signal(s) most clearly reveals the ratio of high-frequency to low-frequency (line or field rate) response?
- Q2-10. What test signal most clearly reveals both relative chroma-to-luminance gain and chroma-to-luminance delay?

## The Camera Chain

Fig. 3-1 illustrates the wide variance in fundamental television camera-chain systems. Because of the predominance of color telecasting, color systems will be emphasized in this coverage.

### 3-1. THE BASIC CAMERA CHAIN

The FCC standards are based on the work of the National Television Systems Committee (NTSC) as originally filed with the FCC on July 21, 1953. We will often use the term "NTSC color" because the specific details of the system are far more inclusive than those spelled out in the FCC standards.

Since most colors can be duplicated by mixing correct amounts of three properly selected primary colors, it follows that a color tv system can be based on the transmission and reception of images in the three primary colors. The first step is accomplished in the television camera. The camera generates three different signals from the information contained in the image of the scene. There are three general types of color cameras, as follows:

1. Three cameras are operated from a single set of controls so that the view televised by each is identical. In front of each camera lens is placed a red, green, or blue filter. While the view imaged by all three cameras is identical, the light reaching the light-sensitive plate of each camera contains only the components passed by its respective filter. The three camera heads operate as one camera, and this camera produces a signal corresponding to the image in each of the primary colors. The brightness, or luminance, information is a function of the combination, and therefore the three images must be accurately registered to obtain a reasonably sharp picture, whether repro-

- duced in monochrome or color. This is the most widely used type of color camera today.
2. Other cameras use a similar system, except that four pickup tubes are involved: three for the primary colors, and the fourth for luminance only. In this type of camera, only the luminance channel must have wide bandwidth; the three color channels can be relatively narrow-band. With this type of camera, since the color channels carry very little brightness information, the effect of misregistration of images is slight on *monochrome* receivers.
  3. A "convertible" system makes use of only two channels in the camera. One tube is used for a full-bandwidth luminance signal, and the second tube is used in the alternate red/blue channel. The fields are sequenced mechanically through a rotating red and blue filter wheel synchronized to the vertical rate of the main synchronizing generator. The green signal is obtained (before encoding) through subtractive matrixing of the red/blue and luminance signals. Special processing is required to convert the sequential color to the NTSC color signal.

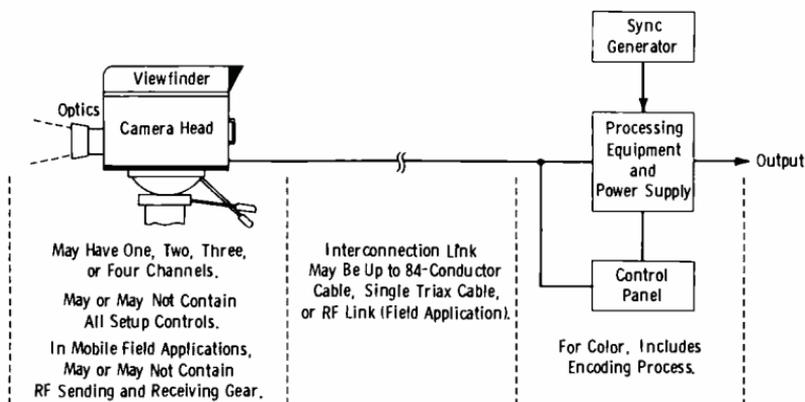
In any of the basic types of camera mentioned above, we will be concerned with three signals, Y for luminance, and I and Q for chrominance. *Caution:* You will find a rather common application of the letter M (for "monochrome") to designate luminance information. In this book, the letter Y will be used for luminance, because in NTSC-FCC color specifications, the voltage of the composite color signal (which obviously contains both luminance information and chrominance information) is designated by the symbol  $E_M$ .<sup>1</sup>

You should remember that in most types of color cameras employed in the studio, the camera chain is concerned only with the three primary colors, red, green, and blue (RGB). Transmission primaries Y, I, and Q (containing the color-difference signals  $R - Y$  and  $B - Y$ ) are generated in the color encoder. Thus camera setup adjustments and maintenance procedures are normally concerned only with obtaining the correct amplitudes and processing in the RGB channels so that the resultant feed to the color encoder is correct.

The development of highly portable and mobile miniature color cameras used for "electronic news gathering" (ENG) has brought about a complete change in design philosophy. These cameras normally have self-contained sync generators and encoding systems.

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<sup>1</sup> It is imperative that the reader have a color training background equivalent to that in *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*, by Harold E. Ennes (Indianapolis: Howard W. Sams & Co., Inc.)



**Fig. 3-1. Components of basic camera-chain systems.**

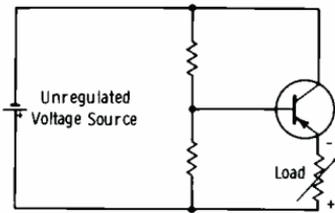
Some of these units, designed for either field or studio applications, carry this same design philosophy into the studio. When this is done, separate control units are used in the control room for the purpose of “matching” and balancing the cameras to each other. Thus, the maintenance technician is faced with a wide variety of camera-chain system concepts.

In this text, camera adjustment and maintenance will be treated as an RGB process only, and it will be assumed that the control unit, power supplies, and color encoder are separate units as is most common with larger studio camera chains. By clearly defining the subject area, the discussion will be appropriate for any system.

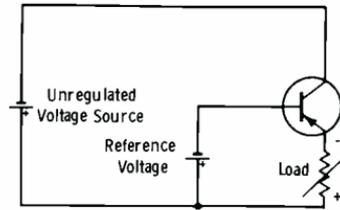
### 3-2. CAMERA-CHAIN POWER SUPPLIES

Before we explore the maintenance and troubleshooting techniques for the video signal path, it is pertinent to become well acquainted with the one common unit upon which all actions of the camera chain depend, the power supply. Camera-chain power supplies are normally rack-mounted units with means of distribution to rack equipment, to the camera control console, and, through the camera cable, to the studio camera. Electronically regulated supplies are universal.

Most regulated power supplies employ transistors and (in some cases) zener diodes. The circuit shown in Fig. 3-2A is a common version of the basic emitter-follower (or common-collector) circuit. The power-supply load (symbolized by a variable resistance) is placed in series with a transistor. The impedance of this transistor is controlled automatically in such a way that it tends to compensate for impedance changes (or current changes) in the load, thus main-

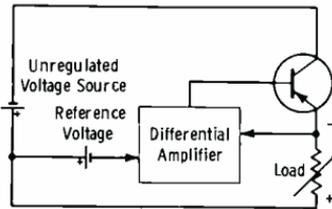


(A) *Stabilized against load variations only.*



(B) *Stabilized against load and source variations.*

(C) *Further stabilization by error amplification.*



**Fig. 3-2. Basic transistor voltage regulators.**

taining an essentially constant voltage across the load. This action may be explained by noting that the voltage drop across the emitter-base junction of a transistor is usually negligible in comparison with the supply voltage (at least over reasonable operating ranges), so the emitter tends to remain near the potential established by the voltage divider in the base circuit. Since the base current is only a small fraction of the emitter (or load) current, the base voltage is not altered significantly by changes in load current, provided the resistors in the voltage divider are not too large.

An alternative approach to the explanation of the regulating action of Fig. 3-2A is to point out that the output impedance of an emitter follower is inherently low, and it approaches the impedance of the emitter-to-base junction alone as the base impedance decreases to zero. The output impedance never decreases to zero, however, so the regulation never becomes perfect with this simple circuit.

Note that a pnp regulating transistor is more conveniently placed in series with the negative side of the load, rather than the positive side as would be the case with most vacuum-tube regulators. The transistor itself must, of course, be capable of handling the maximum load current. In practical transistorized power supplies, it frequently is necessary to mount the large series regulators on radiators or other types of heat sinks to keep the temperatures of the transistor junctions within safe limits.

While the simple circuit in Fig. 3-2A is reasonably effective in stabilizing the output voltage against load variations, it does not

remove variations caused by voltage changes in the unregulated source. This is because the voltage on the base of the transistor is changed in proportion to the unregulated voltage. The circuit in Fig. 3-2B overcomes this problem through the use of a separate, stabilized reference voltage source at the base. Although a battery symbol is shown, the reference-voltage source in a practical circuit could be a *reference diode* (zener diode), which is a semiconductor diode with enough reverse bias to operate in the *breakdown region*. A diode operated in this manner behaves very much like the glow tube, or gaseous voltage regulator; that is, the voltage drop across the device is essentially independent of the current over a rather wide range.

The degree of regulation attainable with the circuit in Fig. 3-2B is determined by the emitter-to-base impedance of the transistor itself, which might be of the order of a few ohms. Even better stabilization (or lower output impedance) can be provided by the use of additional gain in the control circuit to supplement the gain of the regulating transistor itself. Such an approach is illustrated in simplified form in Fig. 3-2C. The voltage across the load may be compared with a stabilized reference voltage in a differential amplifier, which can be designed with enough gain to make the voltage variation at the load as small as required.

In a reverse-biased junction diode (zener diode), at a certain value of reverse-bias voltage the current increases rapidly while the voltage across the diode remains essentially constant. This *breakdown voltage*, which may be any value between 2 and 60 or more volts, depends on the construction of the diode. The zener diode is used in transistorized regulated power supplies to hold an element of the transistor at a given reference voltage.

A basic diagram of a constant-voltage regulated power supply is shown in Fig. 3-3. A regulated reference voltage is obtained from a full-wave rectifier; a bridge rectifier supplies the series-regulator transistor, which receives at the base a feedback voltage from a comparison circuit. This voltage is an error signal of such magnitude and polarity as to change the conduction of the series regulator, hence changing the current through the load resistor until the output voltage ( $E_o$ ) equals the voltage across the voltage control ( $R_{EO}$ ). Note that since the series regulator is in the positive side of the circuit, an npn transistor is used.

The difference between the two voltage inputs to the differential amplifier (Q1A and Q1B) is held at zero by feedback action. Thus the voltage across summing resistor  $R_s$  is held equal to the reference voltage. The current through  $R_s$  and the output-voltage control ( $R_{EO}$ ) is termed the *programming current*,  $I_p$ . The input impedance of the differential amplifier is high, and essentially all of the current

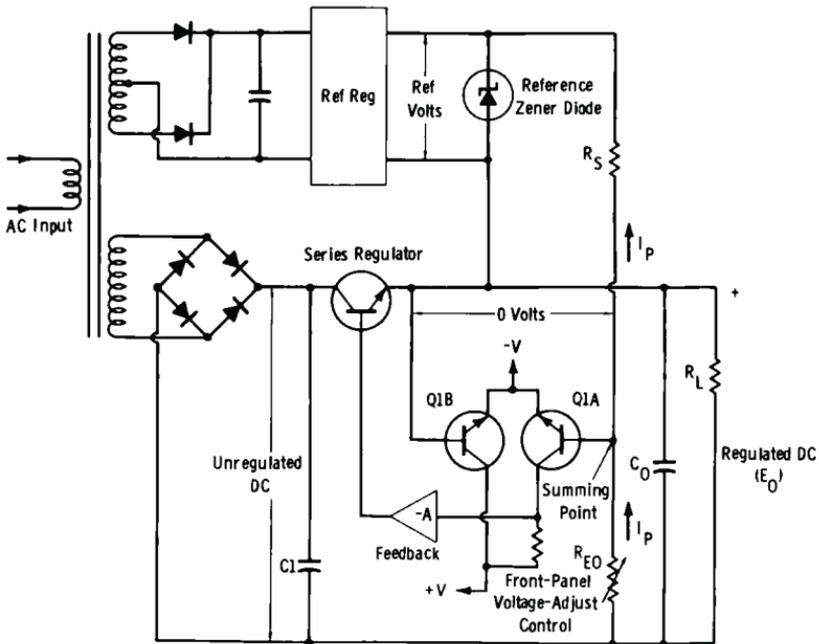


Fig. 3-3. Basic regulated power supply for constant voltage.

( $I_p$ ) through  $R_S$  also passes through  $R_{EO}$ . Because the voltage across  $R_S$  is constant,  $I_p$  is constant. Since  $R_{EO}$  is variable, the output voltage is directly proportional to the resistance of this control. Thus, the output voltage is the same as the voltage drop across  $R_{EO}$  and will become zero if this control is reduced to zero ohms. This variable control is sometimes in series with a fixed resistor to hold the output voltage to a given minimum value.

Most reference supplies for units employing npn power transistors, as in Fig. 3-3, are referred to the positive output (or positive sensing, to be described) as a circuit common. The reference auxiliaries for supplies using pnp power transistors are referred to the negative output terminal.

The differential amplifier contains matched transistors (Q1A and Q1B) placed in thermal proximity or contained in a single "chip." This markedly improves the drift performance of the supply.

We can follow the feedback regulating action by assuming that the regulated dc output has momentarily increased. We will regard the positive output terminal as "common"; then the output-voltage increase causes the summing point to become instantaneously more negative. The resultant decrease in current through Q1A causes its collector to become more positive; hence, a more negative voltage is applied through the inverting feedback amplifier (-A) to the

base of the series regulator. The resultant decreased conduction of the series regulator reduces the output voltage by an amount proportional to the momentary increase, and the error voltage between the bases of Q1A and Q1B is reduced to zero.

Camera-chain regulated power supplies are never as simple as the basic diagram of Fig. 3-3. For even more moderate power outputs, the dissipation requirement for the series-regulator circuit is such that multiple transistors sometimes are used in parallel to provide adequate power-handling capacity. Most recent power supplies employ some means of *preregulation* in the rectifier path. The purpose of a preregulator is to allow the rectifier output to change in coordination with the output voltage so that minimum voltage drop occurs across the series regulator, and power dissipation is reduced to a small value in all series regulator elements. Silicon controlled rectifiers (SCRs) usually are used in the preregulator so that firing time can be controlled for the required conduction angles.

Another feature found in modern supplies is the use of two extra wires between the supply and the load (Fig. 3-4). This results in optimum regulation at the load terminals rather than at the power-supply output terminals, compensating for the IR drop across the resistance of the wire. The current through the sensing lead is so small that, in spite of the resistance of these leads, the voltage drop is negligible. This automatic arrangement eliminates the need for an adjustable tapped-transformer switch in the camera head to compensate for a change in cable length.

A block diagram of a power-supply module incorporating preregulation and remote sensing is shown in Fig. 3-5. Basic analysis by blocks is as follows:

**Bridge Rectifier:** The bridge rectifier provides full-wave rectification of the ac from the power transformer.

**Switching Preregulator:** The switching preregulator switches on and off twice during each half cycle to charge the filter capacitors. It maintains across the series regulator a small voltage drop that

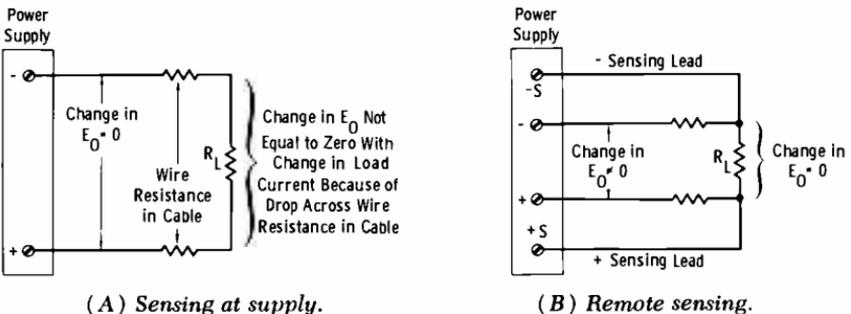
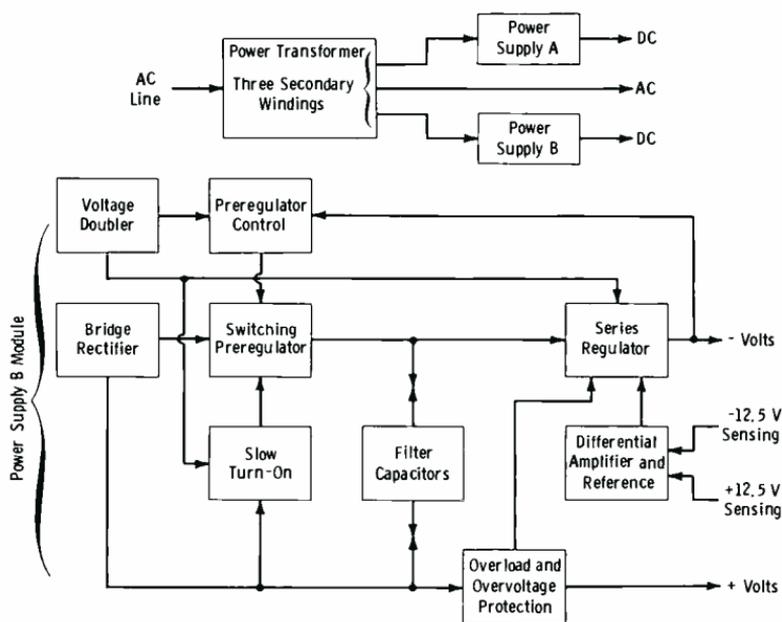


Fig. 3-4. Principle of remote sensing.

varies little with ac line-voltage changes. Thus, it is possible to reduce the power dissipated in the series regulator to the minimum required for adequate ac ripple filtering.

**Preregulator Control:** The preregulator control determines the conduction angle of the switching preregulator by comparing the full-wave rectified output from the bridge rectifier with the output from the series regulator.



Courtesy RCA

**Fig. 3-5. Block diagram of power-supply module.**

**Slow Turn-On:** The slow turn-on circuit causes the output voltage to increase from zero to the rated output in about one second. This gradual build-up of dc reduces possible damaging current and voltage surges within equipment that obtains power from this unit, as well as high current surges within the supply itself.

**Filter Capacitors:** The filter capacitors are charged twice during each half cycle through the switching preregulator. During the time interval between charges, the energy that has been stored is discharged through the series regulator to the load.

**Voltage Doubler:** The voltage doubler provides the current source for the control-signal input to the series regulator. The use of the voltage doubler greatly reduces ripple on the dc output and minimizes the change in output voltage with changes in the ac line input.

The voltage doubler also provides power to operate the switching preregulator.

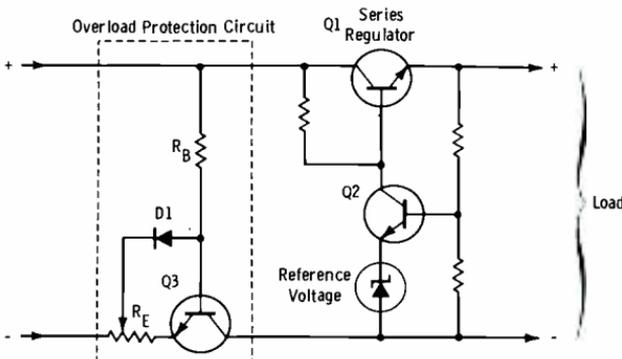
**Series Regulator:** The series regulator filters out almost all of the ripple that appears across the filter capacitors. It compensates for changes in line voltage and load current; it also compensates for moderate voltage drops across the resistances of connector contacts and intrarack wiring.

**Differential Amplifier:** The differential amplifier and reference provide the control signal for the series regulator. A differential amplifier rather than a single-ended amplifier is used because of its inherent temperature stability. Since the supply voltage must be extremely accurate, a precision reference is used instead of a voltage-adjustment control.

**Overload and Overvoltage Protection:** The overload-protection circuit prevents component damage, especially to transistors, because of an overload or short-circuited output. This circuit does not have to be reset after an overload; the output returns to its correct voltage as soon as an overload or short circuit is removed. The overvoltage-protection feature prevents the series regulator and load from being damaged (as a result of excessive power dissipation) in case the switching preregulator should develop a short circuit.

Fig. 3-6 illustrates the basic idea of overload protection. The value of  $R_B$  is such that, during normal regulator operation, Q3 is saturated. This places negligible resistance in series with the negative return. Potentiometer  $R_E$  is adjusted so that it produces sufficient voltage drop to cause D1 to conduct if the load should develop a short or a specific value of overload (low resistance, excessive current). Conduction of D1 reduces the bias on Q3 so that it appears as increasing series resistance in the regulator circuit.

Relatively high voltages are required for the pickup tube or tubes in the camera head. The well-regulated low voltages required for

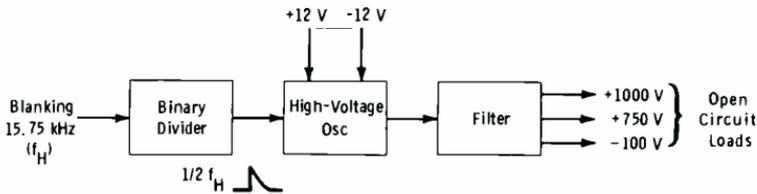


**Fig. 3-6. Basic overload-protection circuit.**

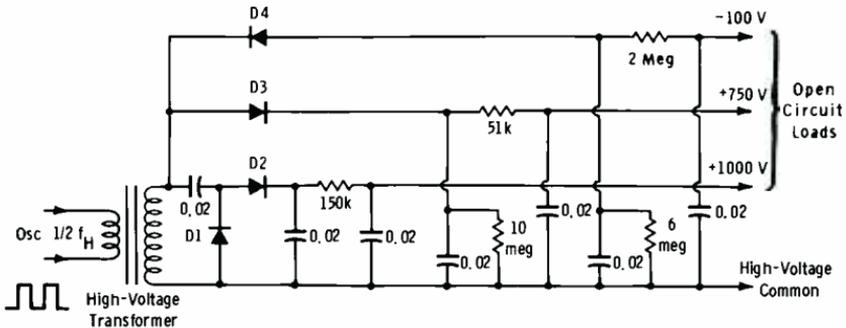
transistor circuitry (such as plus 12 and minus 12 volts) are used to supply a square-wave oscillator. Usually, this oscillator is line-locked to one-half the horizontal scanning frequency so that any radio-frequency interference (RFI) components that should escape the heavy filtering do not appear as beat patterns in the picture.

A block diagram of the basic high-voltage supply is shown in Fig. 3-7A. This supply normally is located in the camera head. The binary counter receives either camera horizontal-drive pulses or separated horizontal-sync pulses from the composite camera blanking signal. Each input pulse is differentiated to obtain a trigger for the binary stage. Thus, the square-wave oscillator is synchronized to one-half the horizontal-scanning rate. This prevents any ripple or transients from appearing in the active scan interval. The square-wave oscillator is driven from cutoff to saturation. Since the voltage supply to this stage is extremely well regulated, excellent regulation of the generated high voltage is obtained.

The source of the high voltages is a step-up torroid transformer at the oscillator output. All "low sides" of the voltages are tied together for a common reference to ensure tracking (Fig. 3-7B). The +1000 volts is developed by doubler D1-D2. The +750 volts is provided by half-wave rectifier D3, and the -100 volts is provided by D4. The current requirement is extremely small, allowing simple



(A) Block diagram.



(B) Filter and voltage divider.

Fig. 3-7. Example of camera-head high-voltage supply.

RC filtering. The slight alteration in output voltage because of the load current is a negligible factor, since all voltages to the pickup tube(s) track from the common reference.

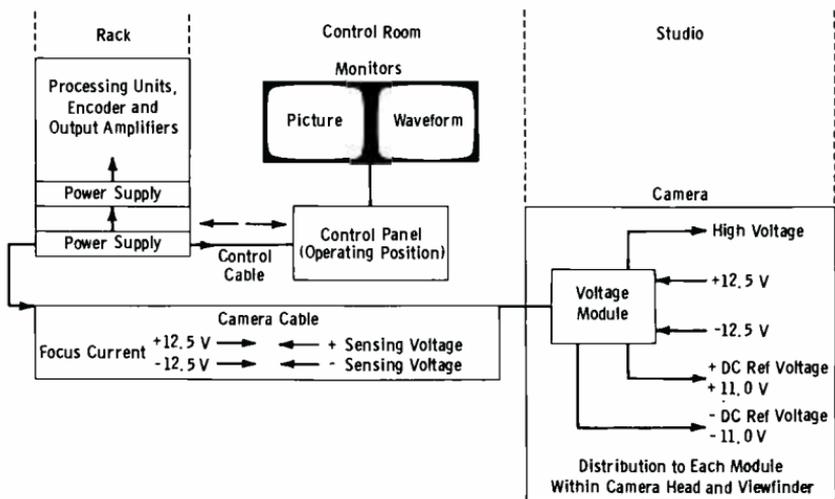
### 3-3. CAMERA-CHAIN POWER DISTRIBUTION

Typical voltage distribution for a modern color camera chain is shown in simplified form in Fig. 3-8. The main power supplies are rack mounted, and all operating and fixed voltage distribution is by way of the control cable from the rack to the control panel and the camera cable from the rack to the camera head.

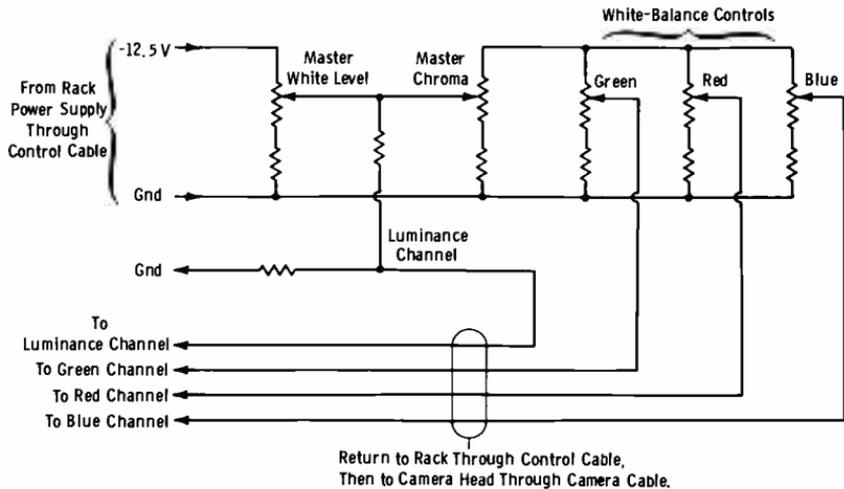
Fig. 3-9 represents a portion of a typical control panel for a four-color camera; voltage distribution is shown for the operating controls designated. The master white-level control is common to the three color channels. Individual color-channel controls are level controls to permit proper white balance on a neutral gray-scale (chip) chart. Distribution is then made to the respective circuitry in the camera head.

Power distribution for a 3-channel camera is essentially the same without the fourth (luminance) channel. In the 3-channel (RGB) camera, luminance is derived from the three color channels in the color encoder.

As shown in Fig. 3-8, the camera head normally contains a voltage module that receives the regulated voltage and serves as a main distribution point for all other modules in the camera. Fig. 3-10 illustrates basic reference-voltage generators in the voltage module



**Fig. 3-8. Typical voltage distribution in color camera chain.**

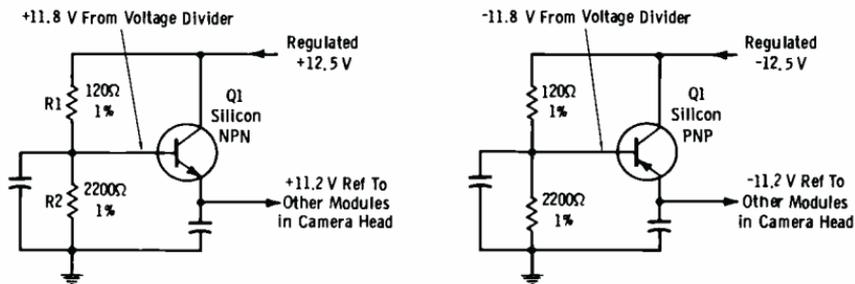


Other controls normally are: Master Black Level  
 Black Balance Controls  
 Registration (Horiz and Vert Centering Controls, Each Channel)  
 Lens Iris Control

**Fig. 3-9. Portion of control panel for four-tube color camera.**

for supplying other modules in the camera. Note that in Fig. 3-10A, the regulated +12.5 volts is divided to +11.8 volts at the base of transistor Q1. Since the transistor is silicon npn, the emitter is at  $11.8 - 0.6$ , or +11.2 volts for reference-voltage distribution. The circuit for the negative reference-voltage generator (Fig. 3-10B) is identical, except that a silicon pnp transistor must be used.

Circuitry in all other individual modules is decoupled from both the main power supply and the camera-head voltage module by decoupler circuits. Basic negative-voltage decoupler circuits are shown in Fig. 3-11. (Positive decouplers are identical except that npn transistors are used.) The circuit of Fig. 3-11A employs a regu-



(A) Positive voltage.

(B) Negative voltage.

**Fig. 3-10. Basic reference-voltage generators.**

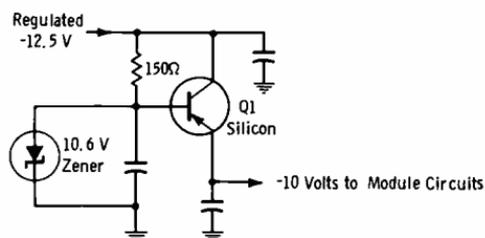
lating zener diode for the base-voltage reference. The circuit of Fig. 3-11B uses the common reference-voltage supply from the camera-head voltage module.

### 3-4. POWER-SUPPLY MAINTENANCE

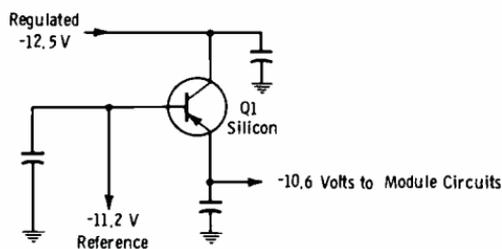
Power supplies, in addition to the overvoltage and overcurrent protection circuitry described previously, often include thermal relays and fuses.

*Thermal-Relay Shutdown:* The thermal relay opens the input circuit only when the power-supply output current exceeds the current rating specified for the operating *ambient temperature*. When the temperature decreases to normal, the thermostat will reset automatically. If this occurs often, forced-air cooling may be required.

*Shutdown From Blown Fuse:* Fatigue failure can occur as a result of mechanical vibrations combined with thermally induced stresses that weaken the fuse metal. Many times, fuse failures can be caused by a temporary condition, and replacing the blown fuse will make the fuse-protected circuitry operative again. Never replace a fuse with the unit turned on (power applied). The resulting temporary loose connection before solid contact is made may open the fuse again through no fault of the equipment. Always inspect fuse holders for tightness and cleanliness. Never substitute a fuse with a rating higher or lower than the original rating.



(A) Reference from zener diode.



(B) Reference from generator.

Fig. 3-11. Basic module decoupler circuits.

When a power supply shuts down from causes other than a thermal relay or blown fuse, it is necessary to determine whether the fault is internal to the supply or in the load. Suitable dummy loads should be made up and kept available so that this problem can be solved readily. Fig. 3-12 illustrates the use of a dummy load for such checks. The plus and minus sensing terminals should be connected to their respective outputs for internal sensing. Automotive-type lamps, such as the type 1073 (1.8 amperes at 12.8 volts), serve as excellent substitute loads. As many as needed should be paralleled to approach the maximum or nominal load of the supply. For example, if the nominal load is 6 amperes, three such lamps in parallel are quite suitable.

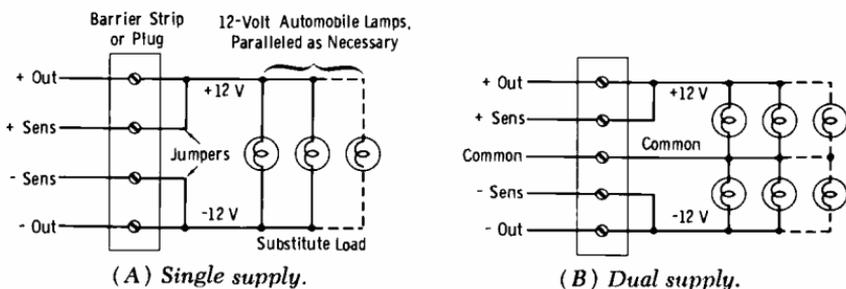


Fig. 3-12. Dummy loads for power supply.

A typical power supply of this kind might have a nominal load of 6 amperes, an overload-protection limit of 10 amperes, and an overvoltage-protection limit of 16.5 volts. With the dummy load substituted for the normal load, an internal fault will cause the lamps to burn brighter than normal for an instant, and then the supply will shut down. The higher-than-normal voltage is caused by lack of regulation in the supply. If the lamps light at the normal output voltage and remain lighted, the trouble is obviously in the load. In the case of the dual power supply of Fig. 3-12B, usually only one of the supplies or loads will be faulty. Thus, if the trouble is internal, one bank of lights will glow brightly and then go out. If all lamps remain lighted, the trouble is isolated to the load.

In the case of a 280-volt supply, a power resistor of the proper value to bring the power-supply output current near maximum normally is used as a dummy load. For example, if the specifications call for a maximum load current of 1.5 amperes, two 500-ohm, 200-watt resistors connected in parallel should be used. Such dummy loads should always be available and ready to connect to the barrier strip or receptacle at the power-supply output.

Older tube-type regulated supplies without modern protective circuitry do not shut down automatically in case of trouble. The output

voltage simply wavers in value around nominal voltage. Such a supply sometimes can be out of regulation with almost no noticeable shift on an external voltmeter. This condition is best checked with an oscilloscope set on ac input at high sensitivity. The change in dc on the coupling capacitor in the scope will show up as a "bouncing" line on the free-running scope trace, usually with a large ripple component.

In the case of intermittent or unusual camera problems which seem to "chase themselves back" to some kind of power-supply problem, use of the scope is mandatory. However, certain precautions *must* be observed to make such tests valid. The importance of proper connection of load and monitoring leads to the power-supply output terminals cannot be overemphasized, since the most common errors associated with the measurement of power-supply performance result from improper connection to the output terminals. Failure to connect the monitoring instrument to the proper points will result in measurement of the characteristics not of the power supply, but of the power supply plus the resistance of the leads between its output terminals and the point of connection. Even connecting the load by means of clip leads to the power-supply terminals and then connecting the monitoring instrument by means of clip leads fastened to the load clip leads can result in a serious measurement error. Remember that the power supply being measured probably has an output impedance of less than 1 milliohm, and the contact resistance between clip leads and power-supply terminals will, in most cases, be considerably greater than the specified output impedance of the power supply.

All measuring instruments (oscilloscope, ac voltmeter, differential or digital voltmeter) must be connected *directly* by separate pairs of leads to the monitoring points. This is necessary in order to avoid the mutual coupling effects that may occur between measuring instruments unless all are returned to the low-impedance terminals of the power supply. Twisted pairs (in some cases shielded cable will be necessary) should be used to avoid pickup on the measuring leads.

Care must be taken that the measurements are not unduly influenced by the presence of pickup on the measuring leads or by power-line frequency components introduced by ground-loop paths. Two quick checks should be made to see if the measurement setup is free of extraneous signals:

1. Turn off the power supply, and observe whether any signal is observable on the face of the crt.
2. Instead of connecting the oscilloscope leads separately to the positive and negative sensing terminals of the supply, connect

both leads to either the positive or the negative sensing terminal, whichever is grounded to the chassis.

Signals observable on the face of the crt as a result of either of these tests are indicative of shortcomings in the measurement setup.

In measuring the input voltage, it is important that the ac voltmeter be connected as close as possible to the input ac terminals of the power supply so that its indication will be a valid measurement of the power-supply input, without any error introduced by the IR drop present in the leads connecting the power-supply input to the ac line-voltage source.

Use an autotransformer of adequate current rating. If this precaution is not followed, the input ac waveform presented to the power supply may be severely distorted, and the rectifying and regulating circuits within the power supply may be caused to operate improperly.

A regulated power supply beginning to "slip" in performance usually has an amount of ripple greater than that specified by the manufacturer as the maximum. Fig. 3-13A shows an incorrect method of measuring peak-to-peak ripple. Note that a continuous ground loop exists from the third wire of the input power cord of the power supply to the third wire of the input power cord of the oscilloscope. This path is through the grounded power-supply case, the wire between the negative output terminal of the power supply and the scope, and the grounded scope case. Any ground current circulating in this loop as a result of the difference in potential ( $E_G$ ) between the two ground points causes an IR drop that is in series with the scope input. This IR drop, normally having a 60-Hz (line-frequency) fundamental, plus any pickup on the unshielded leads interconnecting the power supply and scope, appears on the face of the crt. This resulting noise signal can easily have a magnitude much greater than the true ripple developed between the output terminals of the power supply, and it can completely invalidate the measurement.

The same ground-current and pickup problems can exist if an rms voltmeter is substituted in place of the oscilloscope. However, the oscilloscope display, unlike the meter reading, tells the observer immediately whether the fundamental period of the signal displayed is 8.3 milliseconds ( $1/120$  Hz) or 16.7 milliseconds ( $1/60$  Hz). Since the fundamental ripple frequency present at the output of a supply is 120 Hz (as a result of full-wave rectification), an oscilloscope display showing a 120-Hz fundamental component is indicative of a "clean" measurement setup, whereas the presence of a 60-Hz fundamental usually means that an improved setup will result in a more accurate (and lower) measured value of ripple.

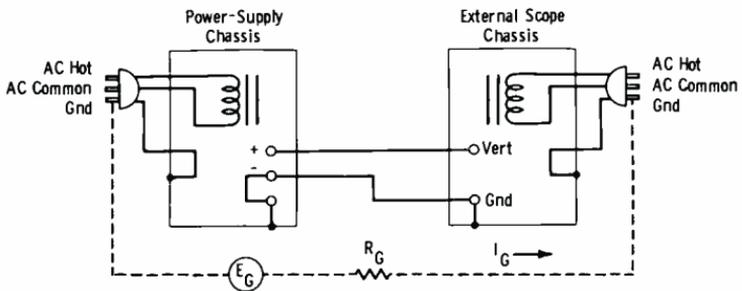
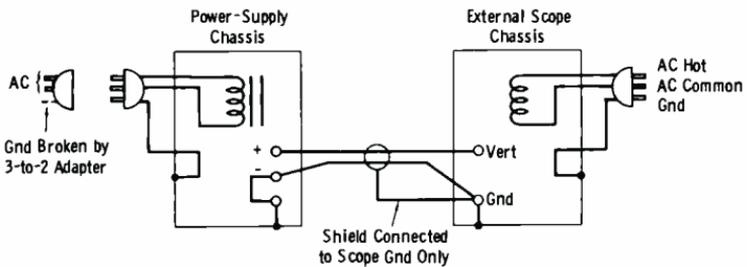
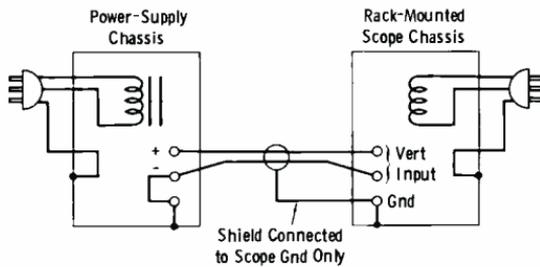
(A) *Incorrect method.*(B) *Single-ended scope.*(C) *Differential scope.***Fig. 3-13. Ripple measurements with oscilloscope.**

Fig. 3-13B shows a correct method for using a single-ended scope to measure the output ripple of a constant-voltage power supply. The ground-loop path is broken with a 3-to-2 adapter in series with the ac line plug of the power supply. Notice, however, that the power-supply case still is connected to ground through the power-supply output terminals, the leads connecting these terminals to the scope terminals, the scope case, and the third wire of the scope power cord.

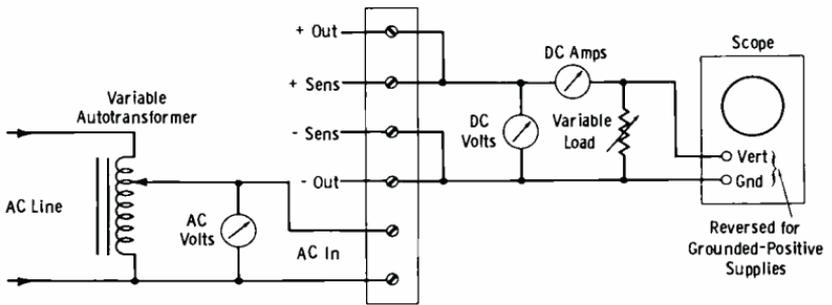
Either a twisted pair or (preferably) a shielded two-wire cable should be used to connect the output terminals of the power supply

to the vertical-input terminals of the scope. When a twisted pair is used, care must be taken that one of the two wires is connected both to the grounded terminal of the power supply and the grounded input terminal of the oscilloscope. When shielded two-wire cable is used, it is essential for the shield to be connected to ground at one end only so that no ground current can exist in this shield and induce a noise signal in the shielded leads.

To verify that the oscilloscope is not displaying ripple that is induced in the leads or picked up from the grounds, the plus scope lead should be touched to the minus scope lead at the power-supply terminals. The ripple value obtained when the leads are in this position should be subtracted from the value obtained in the actual ripple measurement.

In most cases, the single-ended scope method of Fig. 3-13B will be adequate to eliminate extraneous components of ripple and noise so that a satisfactory measurement may be obtained. However, in more stubborn cases, or in measurement situations in which it is essential that both the power-supply case and the oscilloscope case be connected to ground (e.g., if both are rack-mounted), it may be necessary to use a differential scope with floating input, as shown in Fig. 3-13C. If desired, two single-conductor shielded cables may be substituted for the shielded two-wire cable with equal success. Because of its common-mode rejection, a differential oscilloscope displays only the difference in signal between its two vertical-input terminals, thus ignoring the effects of any common-mode signal introduced because of the ac difference in potential between the power-supply case and scope case. Before a differential-input scope is used in this manner, however, it is imperative that the common-mode-rejection capability of the scope be verified by shorting together its two input leads at the power supply and observing the trace on the crt. If this trace is a straight line, the scope is properly ignoring any common-mode signal present. If the trace is not a straight line, the scope is not rejecting the ground signal and must be realigned in accordance with the manufacturer's instructions until proper common-mode rejection is attained.

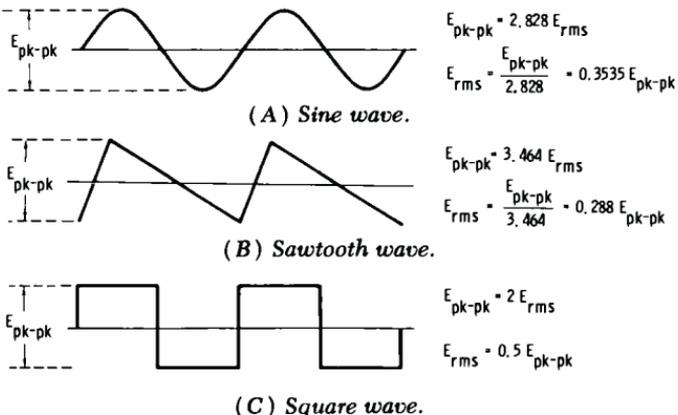
The complete hookup for checking a regulated power supply for comparison to manufacturer's specifications is shown in Fig. 3-14. Most modern 12.5-volt regulated supplies maintain the rated output voltage within plus or minus 1 percent up to maximum rated load with an ac line input of from 90 to 130 volts. The most severe test is at maximum output current with minimum line-voltage input. The ripple voltage is normally around 5 millivolts peak-to-peak minimum for a 12.5-volt supply. Note that with the variable load, the overcurrent-protection circuitry can be checked conveniently in this same setup.



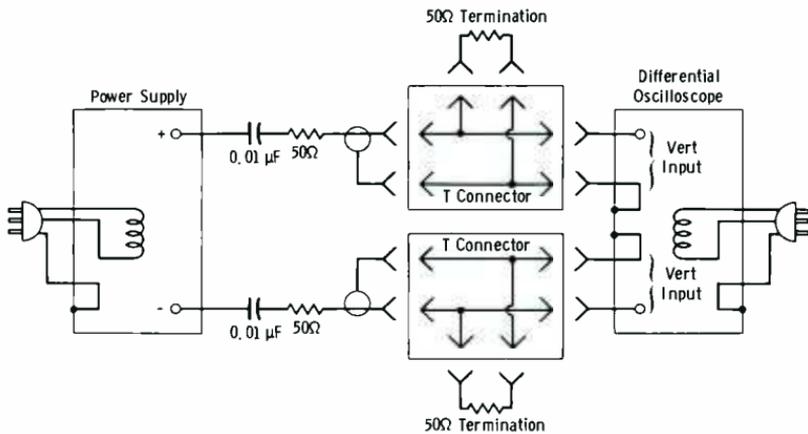
**Fig. 3-14. Setup for checking power-supply performance.**

The peak-to-peak ripple voltage and waveshape may be measured readily with the oscilloscope, but some manufacturers give the ripple specification in terms of rms voltage. Fig. 3-15 shows the relationship between the peak-to-peak and rms values of three common waveforms. The output ripple of a dc power supply usually does not approximate the sine wave of Fig. 3-15A; in many cases the output ripple has a waveshape that closely approximates the sawtooth of Fig. 3-15B. In this case, the rms ripple is 1/3.464 of the peak-to-peak value displayed on the oscilloscope. The square wave (Fig. 3-15C) is included because this waveshape has the highest possible ratio of rms to peak-to-peak values. Thus, the rms ripple and noise present at the output terminals of a power supply cannot be greater than one-half the peak-to-peak value measured on the oscilloscope. In most cases, the ripple waveshape is such that the rms value is between one-third and one-fourth of the peak-to-peak value.

When a high-frequency spike measurement is being made, an instrument of sufficient bandwidth must be used; an oscilloscope



**Fig. 3-15. Conversion of peak-to-peak to rms.**



**Fig. 3-16. Arrangement for measuring noise spikes.**

with a bandwidth of 20 MHz or more is adequate. Measuring noise with an instrument that has insufficient bandwidth may conceal high-frequency spikes that are detrimental to the load. The test setups illustrated in Figs. 3-13A and 3-13B generally are not acceptable for measuring spikes; a differential oscilloscope is necessary. Furthermore, the measurement concept of Fig. 3-13C must be modified if accurate spike measurement is to be achieved.

The Hewlett-Packard Company suggests the following procedure for checking their regulated supplies for noise spikes:

1. As shown in Fig. 3-16, two coaxial cables must be substituted for the shielded two-wire cable.
2. Impedance-matching resistors must be included to eliminate standing waves and cable ringing, and capacitors must be used to block the dc current path.
3. The lengths of the test leads outside the coaxial cables are critical and must be kept as short as possible; the blocking capacitor and the impedance-matching resistor should be connected directly from the inner conductor of the cable to the power-supply terminals.
4. Notice that the shields at the power-supply ends of the two coaxial cables are not connected to the power-supply ground, since such a connection would give rise to a ground-current path through the cable shield, resulting in an erroneous measurement.
5. The measured noise-spike values must be doubled, since the impedance-matching resistors constitute a 2-to-1 attenuator.

The circuit of Fig. 3-16 also can be used for the normal measurement of low-frequency ripple and noise. Simply remove the four

terminating resistors and the blocking capacitors, and substitute a higher-gain plug-in preamplifier in place of the wideband plug-in module required for spike measurements. Notice that with these changes, Fig. 3-16 becomes a two-cable version of Fig. 3-13C.

It may happen that a camera chain loses transistors in certain modules on a rather consistent basis. Sometimes this trouble is attributable to the power supply. Modern regulated supplies have a slow turn-on and also a certain time constant for turn-off to prevent excessive transients from damaging delicate transistors. Transistors actually can be damaged upon turn-off of the power supply; this damage obviously does not show up until the equipment is turned on again.

The checking for on and off transients must be done with an oscilloscope, but many pitfalls exist in such measurement. All power-supply output terminals have a small amount of inductance, and the load can be either capacitive or inductive. The best that can be done without laboratory-type equipment and setups is to compare the on and off transients with those of another camera chain that has not exhibited a problem from this cause. When such a check does reveal a power supply with excessive on or off transients, all filter time constants and all "overshoot" or transient protection circuitry should be checked thoroughly.

NOTE: For stability in video levels, the associated power supply should have very low internal resistance, theoretically zero. (This is never attained in practice.) The internal dc output resistance may be found as follows:

$$R_o = \frac{\Delta V_o}{\Delta I_L}$$

where,

$R_o$  is the dc output resistance,

$\Delta V_o$  is the change in output voltage.

$\Delta I_L$  is the change in output current.

For example, if the output voltage changes 0.1 volt with a load-current change of 5 amperes:

$$R_o = \frac{0.1}{5} = 0.02 \text{ ohm} = 20 \text{ milliohms}$$

The operating condition of a zener diode (or any other type of diode) is most reliably checked with an oscilloscope and the simple associated circuitry shown in Fig. 3-17. The upper trace in Fig. 3-18 illustrates a typical curve obtained by this method. If desired, the forward trace may be eliminated by means of an added silicon diode, as shown by the dash lines in Fig. 3-17. In this case, a curve such as the lower trace illustrated in Fig. 3-18 results.

**Fig. 3-17. Test setup for checking semiconductor diodes.**

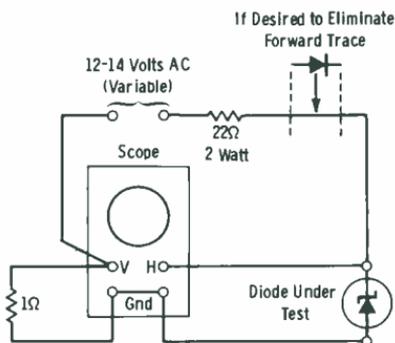
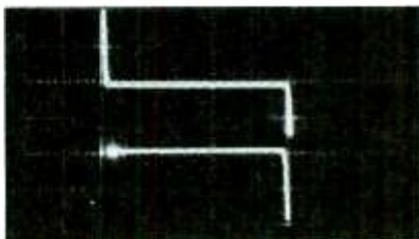


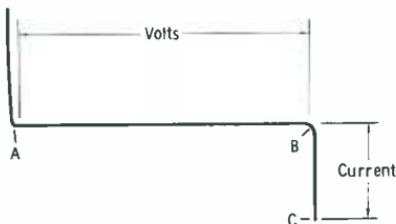
Fig. 3-19 is an interpretation of the trace in Fig. 3-18. As the variable ac voltage (Fig. 3-17) is increased from zero, the voltage is traced horizontally (A-B) along the scope graticule, which may be calibrated in volts/centimeter. When the zener breakdown voltage is reached, the horizontal trace should remain the same length

**Fig. 3-18. Scope traces showing diode operation in test circuit.**

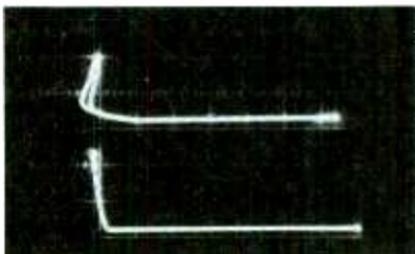


as the current curve increases. If desired, the vertical scale (B-C) may be calibrated in milliamperes/centimeter. Care should be taken not to exceed the maximum current specification (wattage = voltage applied across the diode times the diode current) of the zener diode being tested.

The same test circuit should be employed to check a regular diode at the operating potential encountered in the circuit in which it is used. Some diodes (such as the common 1N34) have a natural *hysteresis loop*, as shown by the upper trace in Fig. 3-20. This loop



**Fig. 3-19. Interpretation of diode trace.**



**Fig. 3-20. Traces with and without hysteresis.**

should remain stable without jitter or erratic "looping" as the voltage is varied around the normal operating level. Other diodes (such as the Type 1N279) do not reveal a loop (lower trace in Fig. 3-20). There should be no instability of the trace as the voltage is varied around the normal operating level.

### 3-5. CAMERA-CHAIN TROUBLESHOOTING

It is assumed that the user of this book is well acquainted with the basic philosophy of circuit design and camera-chain operation and setup as described in the preceding volume of this series, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*. This knowledge is mandatory before maintenance procedures can be undertaken.

In this section, we will consider primarily the RGB (video) and deflection circuitry. Problems in the color encoding unit are considered in a following section.

#### The Video Path

A typical video path for one channel of a color camera employing test pulses for gain calibration is shown in Fig. 3-21. The sequence of functions is different for every manufacturer, but the basic function of each block is the same regardless of the sequence. These functions can be outlined briefly as follows:

*Preamplifier.* This stage is used to convert the video signal current from the pickup tube to an amplitude sufficient for satisfactory processing of the signal.

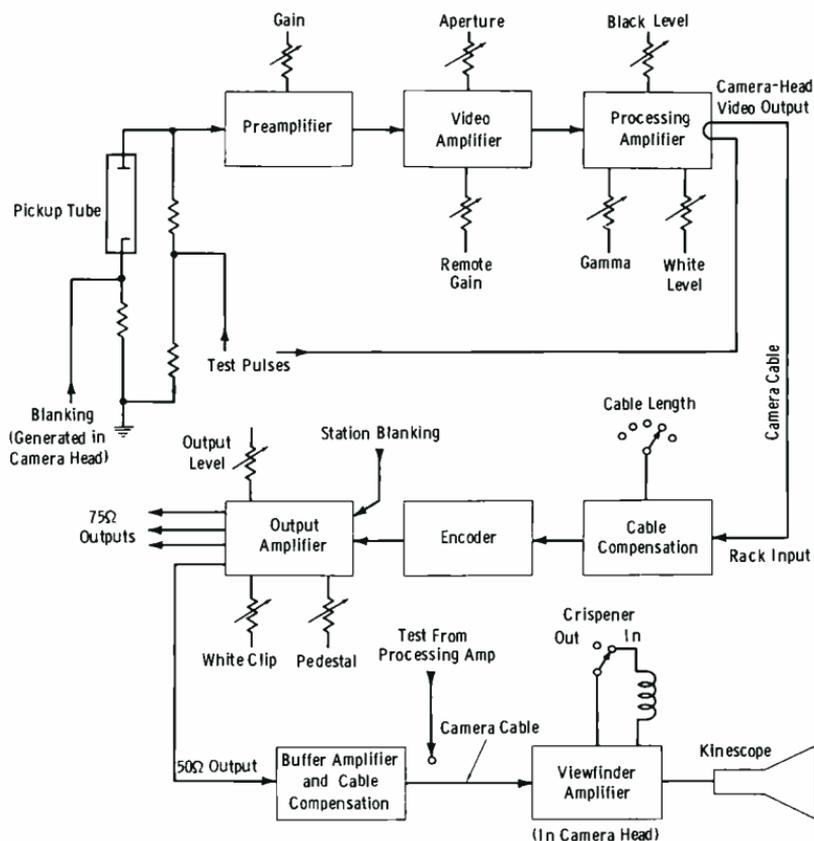
*Video Amplifier.* In the example of Fig. 3-21, the stage includes aperture correction and remotely controlled gain.

*Processing (Proc) Amplifier.* This amplifier involves voltage amplification to boost the signal voltage to a level suitable for gamma correction. It includes manual level sets for black and peak white.

*Cable Compensation.* This section corrects for camera-cable losses at all frequencies across the video passband, for cable lengths of 200 to 1000 or 2000 feet.

*Output Amplifier.* This amplifier provides multiple 75-ohm outputs for distribution of the video signal to the system, and a 50-ohm output to feed back to the viewfinder through the camera cable. It also contains circuitry to limit peak black and peak white signal excursions to preselected values under overload conditions. There is provision for insertion of system blanking, and sync insertion is optional at this point.

*Buffer Amplifier.* This element of the system provides isolation for signal feed to the viewfinder. It includes cable compensation for the viewfinder coaxial line.



**Fig. 3-21. Typical video path for camera channel.**

**Viewfinder Amplifier.** This section amplifies the video signal to a level suitable to drive the display kinescope. Also, it normally employs a switchable "crispener" circuit as a focusing aid for the cameraman.

Conditions requiring emergency maintenance can be classified broadly into three categories:

1. Erratic video level
2. Erratic black level
3. Erratic horizontal and/or vertical deflection

A dead camera chain or a combination of all three of the above conditions generally points to the common unit for the entire chain—the power supply. Maintenance of power supplies was covered in Section 3-4.

**Erratic Video Level (Black Level Constant)**—Review Fig. 3-21. The first step is to check the camera chain with test pulses, when

these are provided. If the pulse inserted at the preamplifier input is erratic relative to the reference pulse (inserted at the camera output end of the cable), then the trouble is obviously in the preamplifier, video amplifier, or processing amplifier in the camera head. If both the inserted and reference pulses are erratic, the problem is in the control console or control-room rack equipment. Test points normally are provided at each module output for quick scope observation. If the pulses are stable but video with the pickup tube looking at a scene is erratic, the problem lies in the pickup-tube circuitry.

In a multiple-channel color camera, trouble of this nature is usually existent in only one of the channels at a time. If the trouble is exhibited in all channels, then it occurs at a point after combination of all channels—in the encoder. In this case, the B, R, and G monitoring points will show no level changes on an individual channel at the encoder input.

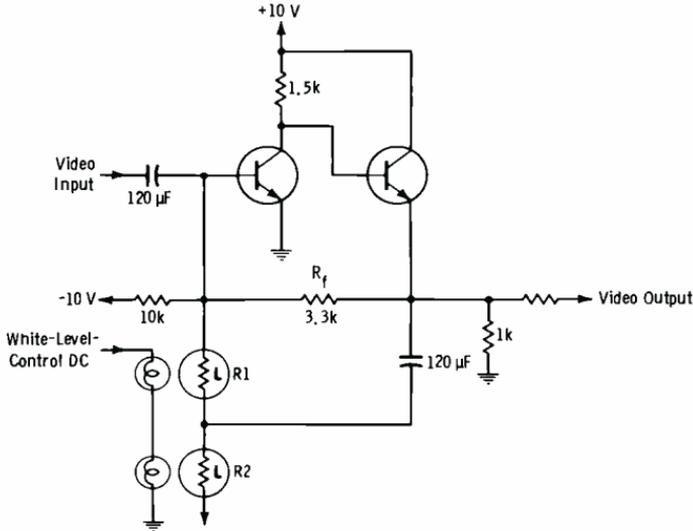
Erratic video is often traced to the remotely controlled video-gain-control stage. Fig. 3-22A shows a typical gain-controlled video stage employing a feedback pair. The feedback resistor ( $R_f$ ) is shunted by a network that includes photoconductive resistance devices R1 and R2. The filament brightness of the small lamps determines the resistance of their associated cadmium-sulphide cells, and therefore the total value of feedback impedance. Voltage is supplied to the lamps from the master white-level control at the remote-control panel.

If this control should fail, you can, in an emergency, substitute a fixed resistor of about 1600 ohms for the defective cell. This will provide a fixed gain, not variable for that channel at the remote point. You can shunt the cell temporarily with the resistor for a quick check to determine whether this is the source of the trouble. If you do not have a direct replacement, it may be necessary to experiment with the value of the substitute resistor to obtain proper control of the output level without overloading a stage prior to the final output-level control in the control room. The temporary resistor should have about one-half the value of the feedback resistor used.

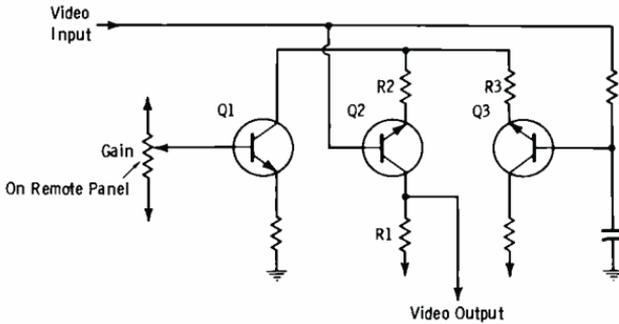
Another popular type of remote video-gain control is shown in Fig. 3-22B. This simple and stable circuit depends on the principle that the small-signal emitter resistance ( $r_e$ ) is approximately equal to  $26/I_E$ , where  $I_E$  is the emitter current in milliamperes. By control of the current in the constant-current supply (Q1), the currents in Q2 and Q3 are controlled; hence, the value of  $r_e$  in these transistors (and therefore the stage gain) is varied.

All plug-in modules are susceptible to plug and receptacle contact problems. Intermittent deflection or fluctuations in video black or white level are often the result of dirty or otherwise faulty contacts.

If the equipment manufacturer recommends a particular cleaner for these contacts, by all means use it on a regular basis. Otherwise, a good tuner cleaner ordinarily is satisfactory.



(A) Photoconductive control element.



(B) Variable emitter-current source.

Fig. 3-22. Typical remote video-gain controls.

*Erratic Black Level (Peak-to-Peak Video Level Constant)*—Now we will consider the case in which the overall amplitude is erratic, but the peak-to-peak video remains constant while the pedestal changes. Again, the first step is to check the test pulses. This problem sometimes results when a clamp-pulse width or timing adjustment is on the edge of the proper setting. The maintenance engineer should familiarize himself thoroughly with all such adjust-

ments for the equipment in the installation with which he is concerned.

Check first for the existence of clamp pulses. Then, using video of varying APLs, check for the effectiveness of clamping at the clamped point. Sometimes the black-level control pulses are fed into a video stage prior to the clamp and are timed (by adjustment or design) to occur within the clamp interval. Check these points on a dual-trace scope.

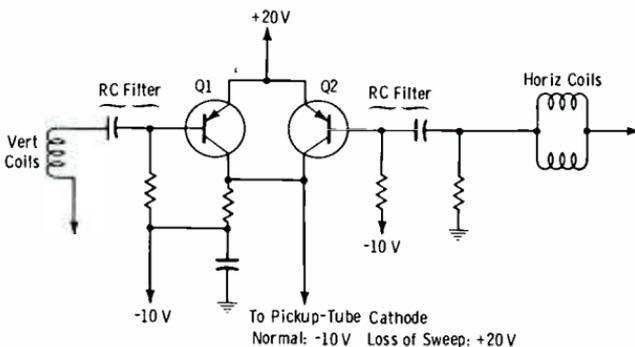
Obviously, the wide variety of circuitry employed in different camera chains prevents establishment of a set pattern of testing. The maintenance engineer must study and analyze his particular instruction-book descriptions. Any point that is not clear is sufficient reason to contact a factory representative for clarification.

*Vertical White Spike on Left of Picture*—This is a problem common to many makes and models of cameras. Whenever you see a large vertical spike just following the end of horizontal blanking (start of picture on left of raster), check for any clamp-pulse position (or width) control or horizontal-drive width control that may be used in the camera head. This control, when used, must be adjusted for a “clean” end to horizontal blanking without a noticeable spike as indicated on the waveform monitor.

### Deflection Problems

You will recall that if either the horizontal or vertical deflection fails entirely, the pickup tubes will be biased off and no video will be obtained. When this condition exists, the test pulse is passed, but no video is obtained from the pickup tubes. In this case, the first place to check is the tube protection stage (Fig. 3-23).

Transistors Q1 and Q2 sample the outputs of the vertical- and horizontal-deflection yokes, respectively, and control the voltage appearing at the pickup-tube cathode. The biasing arrangement shown causes conduction of both transistors in the absence of a base



**Fig. 3-23. Example of protection circuit for pickup tube.**

signal. Under this condition (transistors saturated), the common collector voltage goes to the emitter voltage of +20 volts, cutting off the pickup-tube beam. When both deflection signals are present, both transistors cut off, and the common collector voltage goes to -10 volts. Since this point is connected to the pickup-tube cathode, beam current can exist.

In the absence of either deflection signal, the respective transistor conducts (saturates), and the common collector goes from -10 to +20 volts, again cutting off beam current. This specific circuit treats the retrace or flyback duration as absence of deflection signals and triggers the beam to cutoff, thus providing blanking. In this case, blanking and pickup-tube protection are provided by the same circuit.

The best way to check this circuit is to use the dc position on an oscilloscope with a deflection sensitivity of 10 volts per centimeter, and observe the collector voltage of Q2. If a steady positive voltage exists, sweep failure has occurred (unless there is a shorted transistor in the protection circuit).

See Fig. 3-24 for a review of a typical deflection arrangement in a three-channel camera. The master size control affects all channels, and it is sometimes not used except in four-channel cameras. In this case, it is set to adjust the scanning size of the luminance tube.

Note that the green size control becomes the master for all chroma tubes, and individual controls are provided for red and blue. Linearity controls (not shown) for red and blue are variable resistors in series with the deflection coils. This arrangement is for the purpose of facilitating registration.

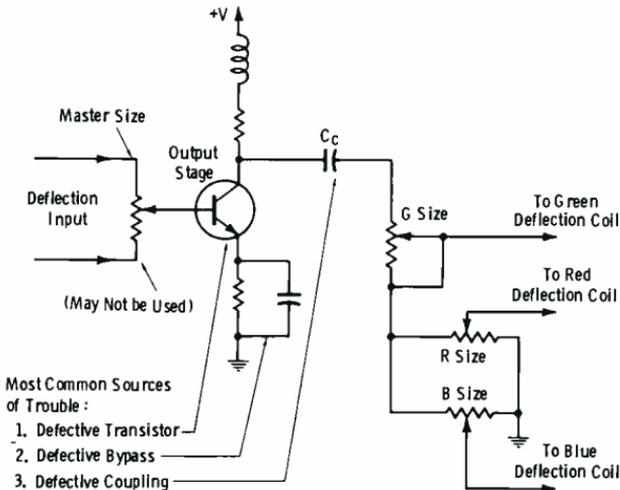
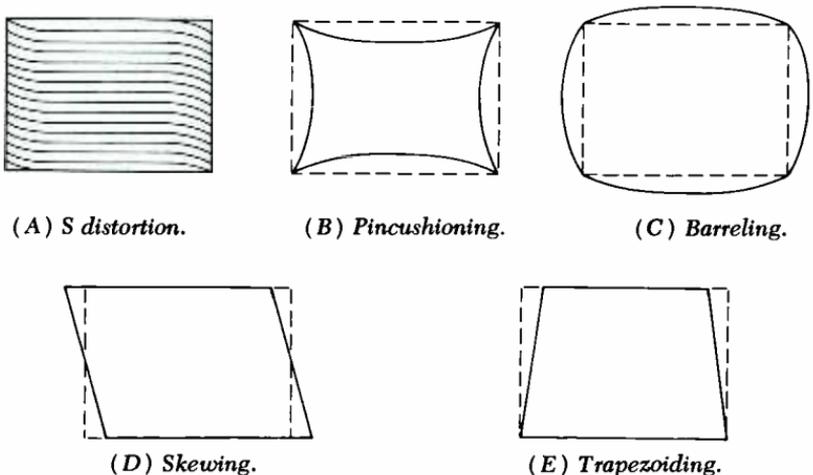


Fig. 3-24. Basic deflection arrangement for three-channel camera.

If all channels show erratic deflection, the problem obviously is located prior to the multiple-yoke takeoff: in coupling capacitor  $C_c$ , in the deflection output stage, or ahead of this stage. If the erratic deflection is in one channel only, troubleshooting is isolated to the corresponding network.

### Yoke Maintenance

Modern yoke assemblies are so designed that very little trouble occurs in this section of the camera head. This is not true of older cameras, many of which are still in use. The type of distortion produced by yoke problems is termed *geometric distortion*. The five general types of geometric distortion that can occur in either the camera or picture monitor are illustrated in Fig. 3-25.



**Fig. 3-25. Types of geometric distortion.**

Fig. 3-25A shows *S distortion*. You pan the camera along horizontal lines and observe the departure from straight lines in the reproduction. This departure is the result of a nonuniform axial field in the pickup tube; such a field imparts nonuniform rotation to the electrons in the scanning pattern. In practice, *S distortion* results from improper adjustment of pickup-tube potentials, from stray magnetic fields, or from a magnetized yoke.

*Pincushioning* (Fig. 3-25B) normally results from improper distribution of windings in a picture-tube (monitor) deflection yoke. It is quite common in low-cost yokes, or it may result from an attempt to substitute a different yoke than that intended for the particular kinescope. *Barreling* (Fig. 3-25C) may result from the same causes.

*Skewing* (Fig. 3-25D) can occur in either the pickup tube or the monitor kinescope. It results when the horizontal- and vertical-deflection coils are not perpendicular to each other. Color cameras employ skew controls (either mechanical or electrical). Excessive skew can be caused by a magnetized yoke or stray magnetic fields.

*Trapezoiding* (Fig. 3-25E) can be introduced into either the pickup tube or the display monitor. It occurs when one set of deflection coils is not symmetrically placed with respect to the axis of the other; the axes of the horizontal- and vertical-deflection coils should effectively bisect each other. Trapezoiding also can be caused by a defective capacitor or resistor network used as built-in compensation for the difference in effective capacitance of the two sides of the coil.

If you encounter a bothersome degree of geometric distortion such as that in Fig. 3-25A, 3-25D, or 3-25E, you may have a magnetized yoke. First, note whether the type or shape of distortion changes with a change of location of the camera or monitor. If it *does* change with location, you have a stray magnetic field. If it *does not* change with location, you have a magnetized yoke, or something in the camera or monitor is strongly magnetized. If this is the case, proceed as follows:

1. Disconnect the *focus coil* (*not* the deflection coil) leads from the terminal board. If no terminal board is used for the focus-coil leads, simply disconnect the camera cable, and locate where these leads go on the camera-cable receptacle.
2. Attach the output of a variable autotransformer (with switch off) across the focus-coil leads.
3. Set the autotransformer arm on 115 volts, and plug the cord into an outlet.
4. Turn the variable autotransformer on. Reduce the voltage to zero in about 5 seconds.
5. Turn the autotransformer off, remove the autotransformer leads, and restore the focus coil to normal operation. The yoke should be demagnetized.

If the above method does not work, it will be necessary to use the longer procedure of removing the entire yoke from the surrounding shield and demagnetizing it with a degaussing coil of the type used for color picture tubes and receivers. Use the same coil on the entire camera. The small hand-type degaussers used for magnetic audio-recording heads are not effective in this procedure.

#### **Measuring Detail Contrast (Resolution)**

The fact that one person can "see" 600 lines of horizontal resolution but someone else can "see" only 500 lines on the same test-

pattern reproduction is not meaningful. There are too many variables involved, including eyesight, psychology, and the condition of the display device. There is only one way to put resolution on an absolute and measurable basis: the use of the line-selected sweep (Sections 1-2 and 1-3, Chapter 1).

The conventional horizontal-rate scope sweep results in a pattern that contains all the lines of the field. It is, however, possible to observe only a single line of the field on, for example, the Tektronix Type 524 scope by turning the TRIGGER SELECTOR switch to the delayed-sweep sector. By rotation of the SWEEP DELAY control, any line or lines may be observed. This sweep is obtained internally from the composite television signal by establishing a coarse time delay from a vertical-sync pulse (from the sync-separator circuit) and then actually triggering the sweep from a selected horizontal-sync pulse. Since the scanned line interval is 63.5 microseconds, if the time base is adjusted to 6.35 microseconds per centimeter, a single line will occur in the full-scale 10 centimeters of the scope graticule.

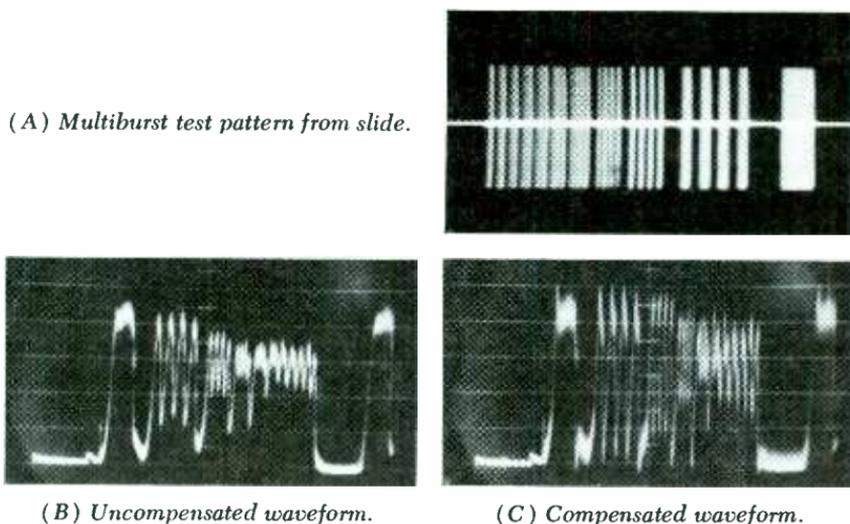
When the SWEEP DELAY control is used in this manner, the sweep is triggered only 30 times per second. The resulting display is correspondingly dim, and, when much ambient light exists, the screen of the Type 524 scope should be viewed through the hood provided for this purpose.

The single-line display on more modern oscilloscopes is much brighter than on the Type 524, and no viewing hood is necessary. However, the line-indicating video output is not provided. To obtain this feature, it is necessary to employ certain waveform monitors such as the Tektronix Type 529 or 1480 series. Use the picture-monitor output with line strobe added to feed the external picture monitor.

One of the main reasons for the line-strobe technique is to evaluate the resolution capabilities between various areas of the pickup tube. For example, a test pattern is available with wedges representing 300 tv lines in each corner. This permits direct comparison of center-to-corner resolution.

The particular line being observed on the cro may be determined by connecting a spare video monitor to the LINE INDICATING VIDEO output jack at the rear of the Type 524 scope (or the line-strobe output of the waveform monitor). The picture on the monitor is brightened during the time of the sweep gate. The SWEEP DELAY control is rotated until the desired line of the picture is selected (brightened on the monitor). Thus, the amplitude of test-chart bandwidth wedges may be measured relative to gray (100-percent) areas.

Fig. 3-26A shows the picture-monitor display of a multiburst test-pattern slide (film chain) with the strobe line indicating the point



**Fig. 3-26. Strobe-line technique of measuring detail contrast.**

of observation on the waveform monitor. (The scope time base in this instance was adjusted to five tv lines to make the indicating line apparent in the photo.) The highest burst frequency in this particular test pattern is 8 MHz, which corresponds to 640 tv lines (Table 2-1, Chapter 2). Fig. 3-26B shows the resulting waveform presentation with the aperture-correction circuitry switched off. Fig. 3-26C shows the waveform with the aperture-correction circuitry switched on; note the increased response at mid and high frequencies.

The amount of aperture correction that can be employed is limited because high-frequency noise is increased along with high-frequency picture-signal content. For example, note the increase in noise at the black level of Fig. 3-26C compared to that in Fig. 3-26B. Where amplitude controls are incorporated for the amount of aperture correction inserted, the noise amplification becomes the limiting factor, as observed on a good picture monitor.

When the studio camera employs amplitude controls in the aperture-correction circuitry, the EIA Resolution Chart is used in the studio under normal lighting. In Fig. 3-27, the delayed sweep is adjusted to the 300-line resolution point on the chart. In all modern cameras, amplitude controls for aperture (and horizontal and vertical aperture compensation units termed "image enhancers") can be adjusted for 100-percent response relative to picture white at 300 lines resolution. Many cameras are capable of 100-percent response at 400 tv lines or more before noise begins to become a limiting factor.



**Fig. 3-27. Picture monitor fed from line-strobe output of waveform monitor.**

NOTE: In Fig. 3-27, the delayed sweep time base was adjusted to five tv lines, rather than a single line, so that the brightening pulse would have good visibility in the photo.

Bear in mind that a 100-percent response is not necessary for the eye to “resolve” a given line number on the picture monitor. For example, if 600 lines has an amplitude response of around 50 percent of white reference, the 600-line wedge is readily apparent.

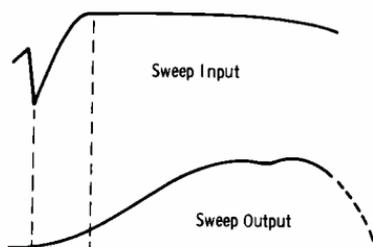
### Adjustment of High-Peaker and Phase Controls

The high-peaker and phase controls are closely associated with resolution problems. Pickup-tube output-circuit capacitance, which results from the tube-yoke combination as well as the input capacitance of the first preamplifier stage, causes a loss of high frequencies with attendant phase shift. This is corrected by the *high-peaker* adjustment normally in the preamplifier assembly, which corrects the frequency loss and introduces a phase shift opposite to that caused by shunt capacitances. The round shape and finite size of the scanning beam results in *aperture distortion*, which is a loss of high-frequency response *without* simultaneous phase shift. Therefore, the aperture-correction circuitry is a phaseless boost of high frequencies.<sup>2</sup>

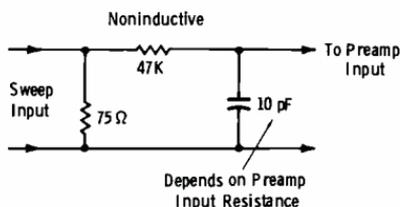
Since the high-peaker (and, for that matter, any “peaking” control in an amplifier) introduces phase shift, over-or-under compensation results in streaking in the picture.

Some video amplifiers employ an incompletely bypassed emitter resistor (small value of capacitance) to compensate for frequency-phase nonlinearity caused by circuit and wiring capacitance. A video sweep through such an amplifier is distorted (Fig. 3-28A) and is meaningless. There is practically no response below about 2 MHz. This is where the term “high peaker” originates, but it is a misleading term.

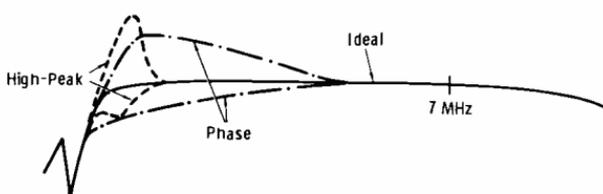
<sup>2</sup> This subject is covered in the preceding volume in this series, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*.



(A) Effect of partially bypassed emitter network.



(B) Example of pickup-tube simulator circuit.



(C) Effects of phase and high-peak controls.

**Fig. 3-28. Modifications in sweep response.**

To see the actual effect on frequency response, you would need to feed a video-sweep signal through a network that simulates a pickup-tube circuit. An example of this is illustrated in Fig. 3-28B. Fig. 3-28C shows the amplifier output with the video-sweep signal fed through the simulator; note that the so-called high-peaker affects only the very low end of the video sweep. Built-in frequency-phase compensation circuitry affects a higher frequency range up to about 5 MHz. Sometimes (in older amplifiers), "phase" correction controls are provided; note that their frequency response is as indicated in the drawing of Fig. 3-28C.

The above should emphasize in your thinking that phase correction circuitry (when used) has a relatively short time constant that will affect short trailing smears. The high-peaker circuitry has a much longer time constant (lower frequency) and affects relatively long streaking of trailing edges.

See Fig. 3-29 for a simplified diagram of high-peaker circuitry incorporated with a gain-set adjustment, as employed in some RCA cameras. One output from the emitter of Q2, the degenerative video feedback signal, is coupled through C4 and R4 to the gate terminal of the FET. A second output from Q2 is coupled through C5 and R5 to a peaking and gain-selection network. Capacitor C1 is adjusted to compensate for losses caused by pickup-tube target capacitive loading at the input to the input amplifier. Gain-selection switch S1 sets the current gain at 50 when in position 2, at 100 when in

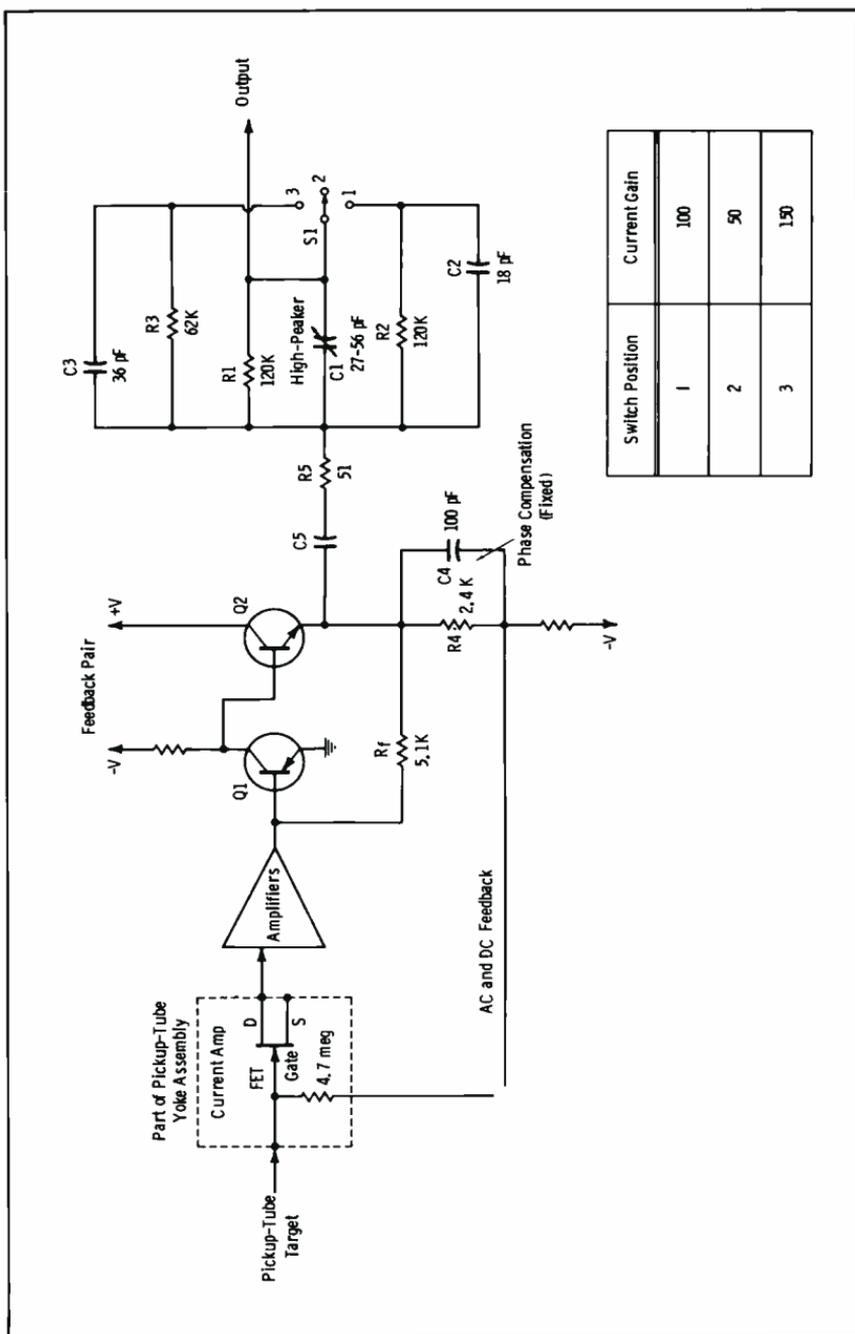


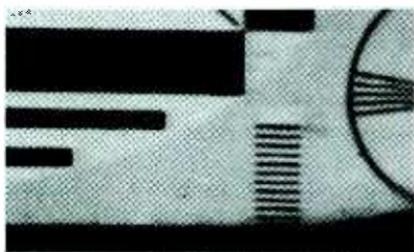
Fig. 3-29. Simplified diagram of video preamplifier used in RCA cameras.

position 1, and at 150 when in position 3. The high-peaker trim requires adjustment when the gain is changed.

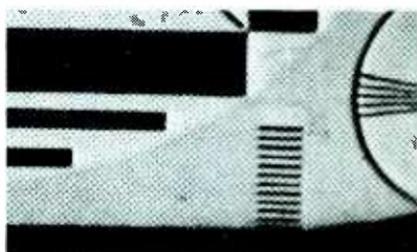
Note that when S1 is in position 2, the output is through R1 and high-peaker trimmer C1. When S1 is in position 1, R2 is paralleled with R1. Since the resistances of R1 and R2 are equal, the current gain is doubled. When the switch is in position 3, R3 is paralleled with R1; the value of R3 is such that the current gain is triple that for position 2.

Capacitors C2 (across R2) and C3 (across R3) are compensating capacitors to help maintain the proper frequency-phase response for the various switch positions. However, as stated before, the high peaker (C1) normally must be readjusted when the switch position is changed.

Final adjustments of the high-peaker and phase controls are made with the camera looking at a test pattern. The horizontal bars in the standard test pattern serve as a reference for adjusting preamplifier controls. Fig. 3-30A shows slight *positive streaking* (black after black or white after white), and Fig. 3-30B shows *negative streaking* (white after black or black after white). Fig. 3-30C illustrates the appearance of lettering under severe negative-streaking conditions. A condition this severe may originate in the preamplifier or in the following video amplifiers that incorporate clamping circuitry.

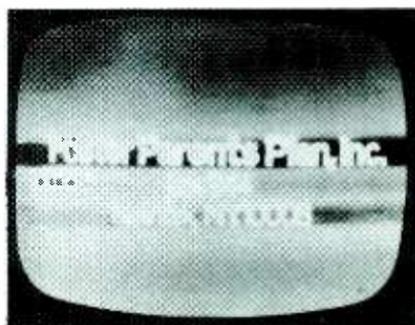


(A) *Slight positive streaking.*

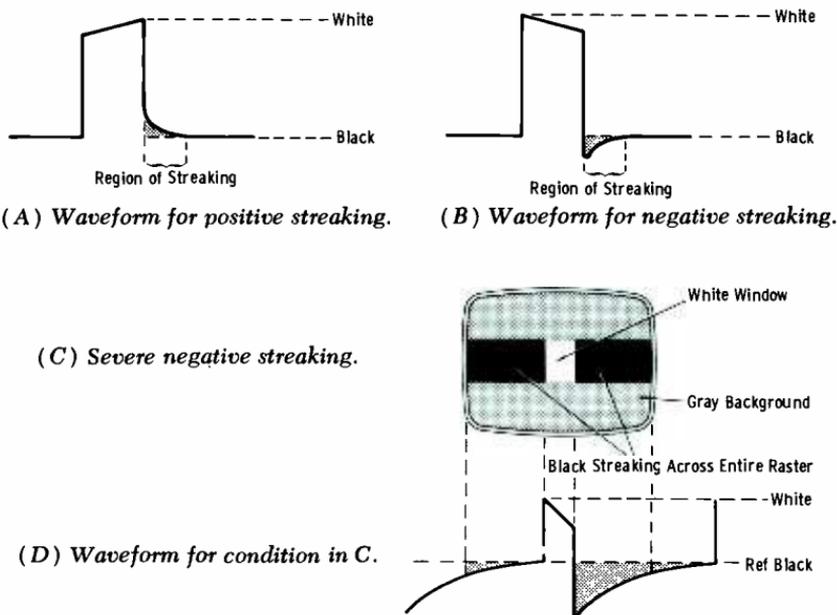


(B) *Negative streaking.*

(C) *Severe negative streaking.*



**Fig. 3-30. Examples of streaking.**



**Fig. 3-31. Conditions of streaking.**

It may be asked how “anticipatory” streaking can occur before (for example) a white area in the scene, causing streaks all the way from left to right on the raster as in Fig. 3-30C. In Fig. 3-31A, the build-up of low-frequency response causes a gradual decline toward black after the white signal, resulting in white streaking of the white image. In Fig. 3-31B, the loss of low-frequency response causes an overshoot into the black region on the trailing edge of the signal, causing black streaking after the white image. These waveforms are representative of “short streaking” (as in Figs. 3-30A and 3-30B). Control of such streaking normally is within the range of high-peaker and phase controls in the preamplifier.

Fig. 3-31C illustrates a white window on a gray background (analogous to the lettering of Fig. 3-30C) with severe negative streaking. The reason for this may be made clear if we consider more than one line of information, as in Fig. 3-31D. A severe loss of low-frequency response not only affects the shape of the white signal (ideally a flat-topped pulse), but also affects the base line as shown. Thus, the black streaking occurs all the way across the raster, not just following the white signal.

Many of the most recent cameras have what are called “low” and “medium” frequency adjustments in the pickup-tube preamplifier. These are set by training the camera on a multiburst test pattern

in the studio (or a multiburst test slide for film chains). The resulting signal consists of flat-topped pulses at varying spacings to represent different frequencies (review Fig. 3-26A). The controls are adjusted so that the tops of the pulses are flat without overshoot or undershoot (similar to the patterns concerned with adjusting the scope 10:1 probe—review Fig. 1-6, Chapter 1).

### The Iris Control Servo

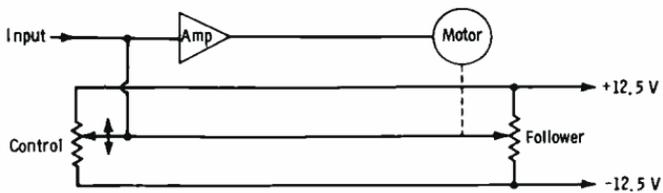
Most modern cameras employ remote control of the lens iris from the operating panel. Servo functions for this purpose generally are located in a servo module of the camera head. Automatic iris control may also be provided, particularly on ENG-type cameras.

Fig. 3-32A represents the basic servo function. The motor is geared to a follower potentiometer which turns with the motor rotation. The voltage applied to the follower also is applied to a control potentiometer. When the net voltage at the amplifier input is zero (equal and opposite voltages from the two potentiometers), the motor cannot receive power. Assume the control is turned toward a positive voltage. The amplifier now has an input that is amplified and applied to the motor. When the motor rotates, the follower rotates in the opposite voltage direction, in this case toward a negative voltage. When the amplifier input again receives equal and opposite voltages (net zero), the motor stops. Bridge circuits normally are involved in practice.

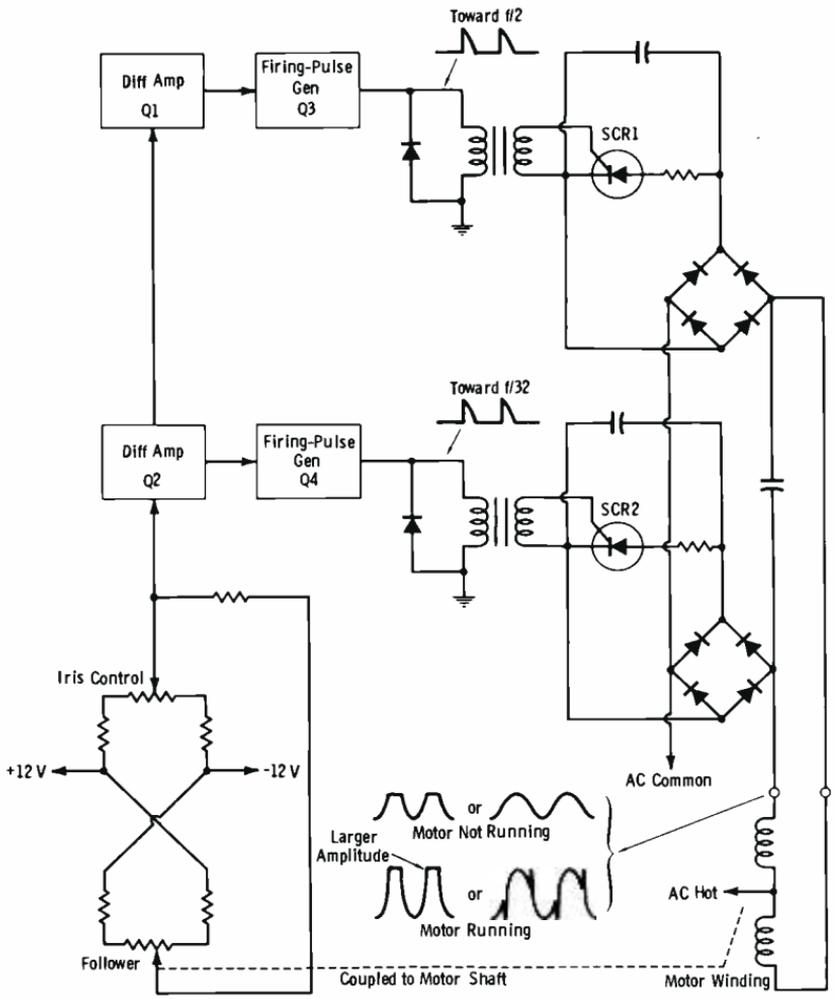
Fig. 3-32B is a functional block diagram of an iris servo. When either of the silicon controlled rectifiers is "fired," its associated diode bridge circuit is able to pass ac. When this happens, the circuit of one winding of the two-phase motor is completed to ground directly through the bridge, and the circuit of the other winding is completed to ground through the bridge and a capacitor. The direction in which the motor runs depends on which winding is in series with the capacitor, and this in turn depends on which bridge is conducting (and therefore which SCR is fired).

When the inputs to the differential amplifier are equal, neither bridge conducts. A change at the Q2 base causes an opposite reaction on Q1. The change occurs because of operation of the iris control; one extreme of this control is the maximum stop opening, and the opposite extreme is the minimum stop opening. An opposite rotation of the iris control causes the opposite SCR to fire, and the motor turns in the reverse direction.

The waveforms on the drawing give a clue as to how to signal trace in probing for troubles. The two types of signals depend on the circuitry involved; these are the most prevalent. Remember, you will always see the 60-Hz power waveform on the scope at the point indicated whether the circuit is completed to ground or not.



(A) Basic action.



(B) Typical circuit.

Fig. 3-32. Principle of lens-iris servo.

The clue to completion of the circuit for the motor is the increase in the amplitude clipping point or the presence of the SCR "spikes" on the waveform.

There are many types of automatic iris controls in highly mobile, miniaturized cameras. One of the most popular types employs a voltage-controlled unijunction oscillator whose frequency is determined by the arm of the follower potentiometer. This feeds a digital up/down counter. The direction of count is set by the polarity of the follower-pot arm, and the counter output is decoded by a digital-to-analog converter and used as the reference for one input to a voltage comparator (operational amplifier). The other input of the comparator is semi-peak-detected video. The comparator output controls the iris motor.

### Checking Gamma Circuitry for Film Cameras

Regardless of the type of circuitry employed, an amplifier containing gamma correction is most conveniently checked with the aid of a stair-step generator. The following is a complete step-by-step procedure that can be used regardless of the method in which gamma correction is obtained:

1. Observe the linearity-generator output directly with the scope; check for proper linearity of the steps. If the steps from the generator are not perfectly linear, take this condition into account when checking the amplifiers.
2. Remove the camera-signal input from the control chassis, and feed the stairsteps into this point. Increase the amplitude of the stair-step signal until clipping starts to show at the output. This tells you the maximum peak-to-peak signal the control (processing) amplifiers can handle without compression. Keep a record of this amplitude for future reference. Be sure the gamma switch *is on unity* (no gamma correction). Remove any aperture correction employed.

In a vidicon film-camera chain, fixed values of gamma correction normally may be switched into or out of the circuit. Most units have a switch with three positions: unity, 0.7, and 0.5. The 0.5 gamma is often necessary for films originally processed for theater projection in order to fit the wide dynamic range of the film gray scale into the range that can be transmitted without severe compression of low grays and blacks. The vidicon has no knee and will not compress whites (if the beam current is sufficient to discharge the highest highlight.)

3. For accuracy in checking linearity without danger of erroneous measurements as a result of excessive levels, reduce the input level to one-half the value recorded in Step 2.

4. When observing linearity of steps at the output of the control chassis, be sure the pedestal control is adjusted with sufficient setup so that the bottom (black) step is not compressed. If the black steps are still "pulled down," check any transient-suppressor controls for incorrect adjustment. These circuits affect the black region.
5. Now place the black-stretch (gamma) switch on the position to be checked. Fig. 3-33 shows the values of output steps you should obtain (with linear input) for the two most common values of correction.

As an example of the mathematical relationship involved, assume gamma is 0.5, and figure where the step for a 0.1-volt input step should be:

$$(0.1)^{1/2} = \sqrt{0.1} = 0.316$$

For the 0.2-volt input step:

Input	Output (Gamma = 0.7)	Output (Gamma = 0.5)
1.0	1.0	1.0
0.9	0.93	0.94
0.8	0.855	0.89
0.7	0.78	0.84
0.6	0.70	0.77
0.5	0.615	0.70
0.4	0.526	0.632
0.3	0.43	0.547
0.2	0.324	0.447
0.1	0.2	0.316
0		

**Fig. 3-33. Gamma correction.**

$$(0.2)^{1/2} = \sqrt{0.2} = 0.447$$

To make computations for the 0.7-gamma correction, it is necessary to use logarithms. For the 0.1-volt step, it is necessary to find the value of  $(0.1)^{0.7}$ . The method is as follows:

$$\begin{aligned} \log N &= 0.7 \log 0.1 \\ &= (0.7)(-1) = -0.7, \text{ or } -1 + 0.3 \end{aligned}$$

Since  $\log N = -1 + 0.3$ ,  $N$  is 0.1 times the antilogarithm of 0.3:

$$N = (0.1)(2.0) = 0.2 \text{ (approx)}$$

6. If the step-voltage response is distorted or more than a few percent in error with the gamma correction in, the reference diodes (when used) are the most likely source of trouble. A slight departure from theoretical step response is normal in circuits employing diodes that are biased to conduct at various levels of video. This is because the resulting response is in incremental steps from black to white rather than a smooth curve.

**NOTE:** It was pointed out previously that the camera chain might employ high peakers and phase-correction circuits with small trimming capacitors across a cathode or emitter resistance. These circuits may cause some tilt on the stairsteps, but normally this will not prevent an accurate measurement. If excessive distortion is present, it is a simple matter to bypass such a cathode or emitter correction circuit with a capacitance of about 1.0  $\mu\text{F}$ .

7. Now run the gray-scale slide or test loop through the complete film chain. This will give a good indication of the operation of the camera head as it responds to the black-to-white pattern viewed by the vidicon. If any compression (with gamma correction removed) is now present, either the trouble is in the camera head itself or the peak-to-peak video level from the camera is excessive. Check this level at the input to the camera control chassis. Always keep this input level below that found in Step 2.

Always remember this: The black stretch of gamma-correction circuitry applies to a *linear input signal*. This is the basic function of the amplifier. Then, with the pickup tube looking at a *logarithmic* gray scale (standard), the output *should be linear*.

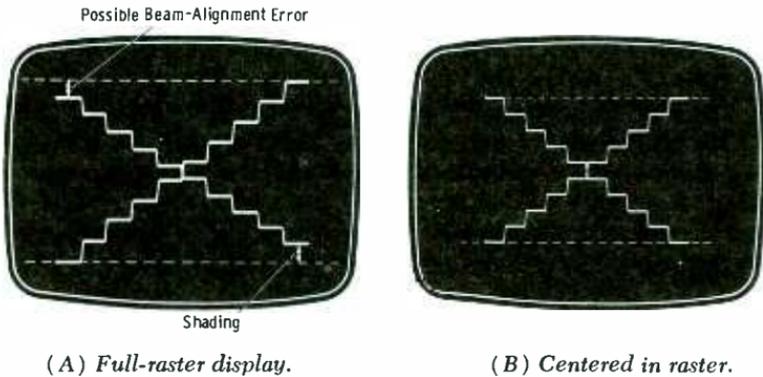
### Checking Gamma Circuitry for Live Cameras

Gamma circuitry is best checked with the camera looking at the crossed gray scale in the studio. The scope will reveal the range of control, which should be checked against the specification sheet

for the camera involved. Obviously, if the gamma correction range of any one channel is low, color balance with the other channels cannot be obtained. In an emergency, you can operate without gamma correction in the channels required to obtain balance.

A troublesome gamma stage is most often the result of faulty transistors or diodes. It is best to make direct substitutions of parts as a check.

Be certain you can distinguish between a gamma fault and some other fault. Fig. 3-34A shows one possible cro presentation when the crossed gray scale is scanned at full raster. Note that the ascending luminance steps are linear, showing that gamma is proper when a logarithmic chart is observed.



**Fig. 3-34. Waveforms for crossed gray scale.**

The mismatch at black level should be corrected with the horizontal shading control. The white levels will still be unequal.

Be certain that the incident-light meter shows equal illumination at the lower left and upper right white chips. Also, be sure the *reflected* light is the same. Sometimes a white chip can be dulled by imperfections or dirt.

The next thing to check is beam alignment. The color camera normally has a "rock" circuit, which is simply a 30-Hz multivibrator; its signal is applied so as to result in a split field if the beam is not properly aligned. Aim the camera at a registration chart, and adjust both the vertical- and horizontal-alignment controls until the images on the viewfinder are superimposed.

You may find that, with the particular pickup tube involved, it is possible to obtain good beam alignment around the central area but not at the corners and edges. This is just one of the factors that can cause unequal white levels at the two extremes of the pickup area. The presence of this effect can be verified by zooming out or moving away from the chip chart so that it is in the center of

the raster as in Fig. 3-34B. (The chart should be displayed against a neutral background.) If the levels become even, it is likely that a beam-alignment problem (or possibly black-level shading) is present.

**IMPORTANT NOTE:** Always check the manufacturer's specifications for the specific pickup tube involved. Some tubes have a photocathode or target sensitivity specification that can vary as much as 20 percent between left and right areas and still meet standards.

### 3-6. THE COLOR ENCODER

The setup and maintenance of the color encoder requires an extensive background in color encoder theory. This theory, as well as initial setup of the encoding system, was covered in Chapters 2 and 10 of an earlier volume in this series, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*. It would be pertinent to review those chapters before or simultaneously with the following discussion.

#### "Distortions" in NTSC Color

In the consideration of color-transmission "distortions," it is necessary to consider the transmission and reception processes together; in fact, it is impossible to consider one without the other. For example, consider the matter of gamma correction. The pickup device must actually be made nonlinear in gray scale to match the "average" color picture tube. The system *following* the pickup device must be made as strictly linear as possible, so that all of the amplitude-versus-brightness correction can be made in the pickup device, and complicating factors are not introduced after this correction.

The point to be emphasized here is that the system handling video information can be carrying a nonlinear function (from the pickup device) and that it is normal for this function to appear nonlinear. Nonlinearity of a certain kind is required (in the case of gamma correction) because of picture-tube characteristics, or (in other cases which we will point out) because of receiver or monitor design. In the cases just cited, you should understand that these nonlinearities are *not* "distortions."

#### The Basic NTSC Color Problem

The number one problem is *phase sensitivity*. For accurate reproduction of a hue, the resultant color phasor must be held within a tolerance of  $\pm 5^\circ$ . Hue error becomes noticeable at  $\pm 10^\circ$  from the proper position relative to burst. The transmission and reception process, to maintain such a tight tolerance, must operate with a signal time-base accuracy of a few nanoseconds.

Color information (hue and saturation) is affected by various types of amplitude and phase shifts in transmitting and receiving equipment. In addition, hue shifts can occur as a result of multipath transmission errors. Burst and subcarrier-sideband phase shift caused by variations in topography over different transmission paths can require the viewer to readjust his receiver with every station change. Although this can be annoying to the viewer, it is not a factor under the control of the station operator. However, it is necessary to adjust the station color gear to precise operating parameters (and this is possible!) and to *minimize*, as much as possible within the limitations of the equipment, any amplitude or phase distortion. When you receive complaints from viewers, you rest easy when you can prove the performance of your system to qualified technical agents.

In the case of multipath-reflection errors, if you are transmitting an accurately adjusted color picture, the viewer can readjust his hue control from the setting for a different station and still get a good color picture. Far worse is the situation in which there is a *differential phase* error. In this case, different colors can be shifted by different angles, depending on luminance. The viewer then can only attempt to strike a "happy medium" with his hue control.

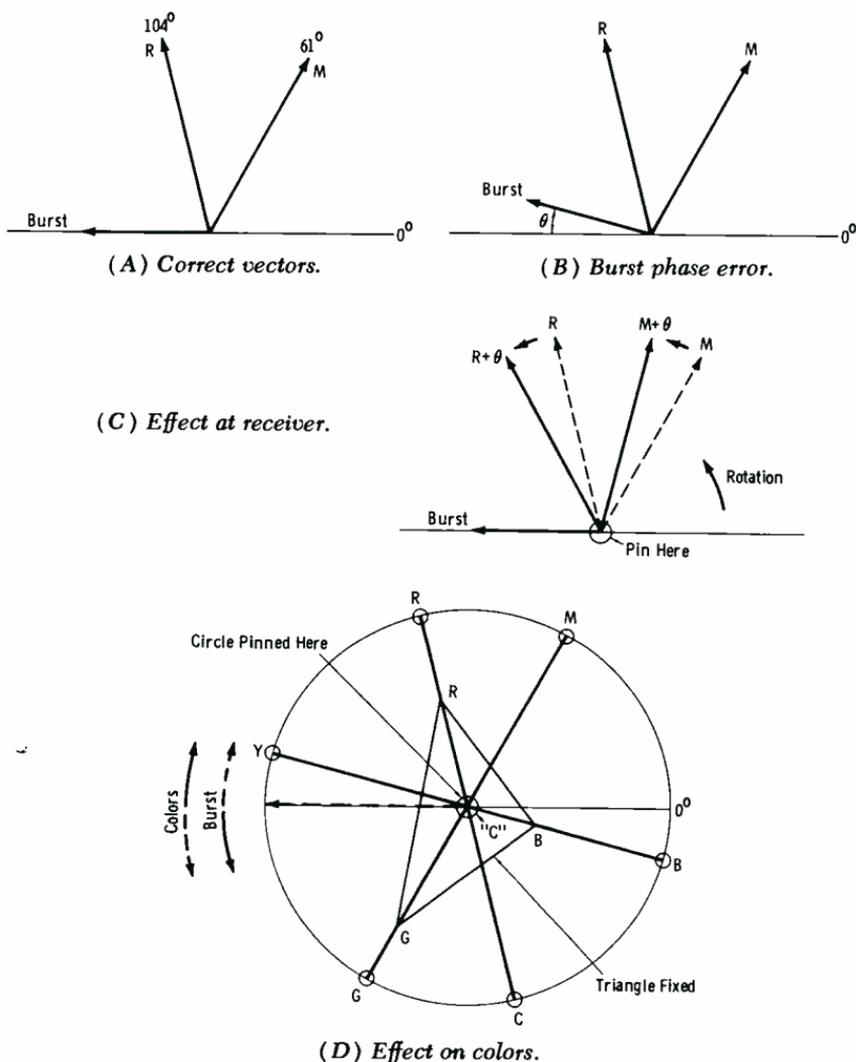
### Burst Phase Error

Remember this: The color monitor or receiver will "insist" that the burst phase not be in error. This results from the fact that the subcarrier-sideband information is synchronously demodulated with the burst as a reference. The angles clockwise from burst for various colors are as in Table 3-1.

**Table 3-1. Color Phase With Respect to Burst**

Color	Phase Angle (Degrees)
Burst	0
Yellow	12.9
Red	76.6
Magenta	119.2
Blue	192.9
Cyan	256.6
Green	299.2

Fig. 3-35A shows the proper vectors for burst, red, and magenta. Now we assume we have the burst-phase error ( $\theta$ ) shown in Fig. 3-35B. The color receiver requires the burst phase shown in Fig. 3-35A; to visualize this, put an imaginary pin at the center dot, and rotate the vectors as in Fig. 3-35C to place the burst at the correct phase. The other vectors become  $R + \theta$  and  $M + \theta$ , which represent a red shifted toward yellow and a magenta shifted toward red.



**Fig. 3-35. Effect of burst phase error.**

The effect of burst-phase error is to rotate all hues in the direction opposite to the burst-phase error. If the burst error in Fig. 3-35B had been counterclockwise, it would have been necessary to rotate the vectors clockwise to return the burst to the proper position. Red would then tend to go toward magenta, magenta would tend to go toward blue, etc., all around the color gamut.

You can visualize this most conveniently by using the color triangle as in Fig. 3-35D. Note carefully that this triangle has been

turned around from the position normally presented, to fit it into the NTSC phase diagram. The center of the circle is at illuminant C, where the color vectors collapse to zero value. Place the imaginary pin for the circle here; the color triangle must remain fixed in the position shown. If the burst slips *clockwise* by some angle, all reproduced colors shift *counterclockwise* by the same angle (and vice versa for counterclockwise slip of burst phase).

Note again the interdependence of the encoding and decoding processes. The hue control at the receiver is an operational adjustment. It has, in most modern receivers, a range of at least  $\pm 70^\circ$ , whereas the tolerance is only  $\pm 10^\circ$  for the *overall* transmission system. This tells you that any receiver in normal operating condition should be able to have its hue control adjusted to obtain proper colors from your station. Your responsibility at the sending end is to assure that the burst phase is as nearly correct as possible so that (theoretically) receivers need not be adjusted for proper color reproduction from different stations. Remember that if you *do* transmit a burst phase error, the receiver exhibits a "locked phase error" so that colors reproduced are not exactly the same as in the original scene. Since the viewer does not have a direct comparison, he is not aware of this as long as he can obtain good flesh tones. But a super-market sponsor will complain if his meat is blue rather than red!

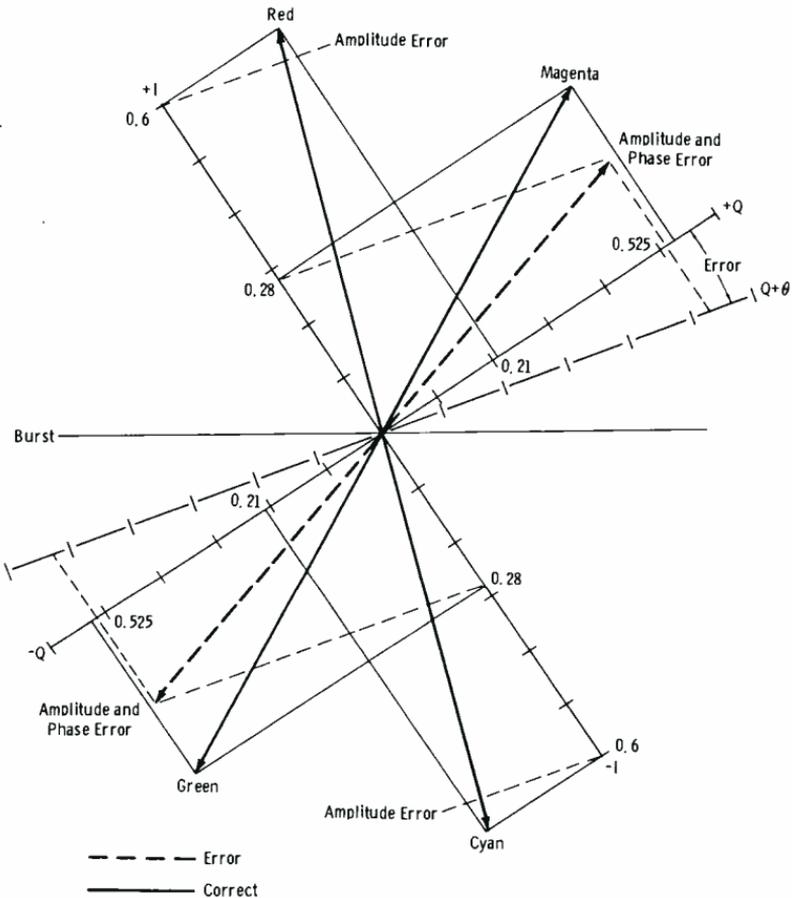
### Quadrature Distortion

Quadrature distortion results from cross talk between the I and Q video information. It has more possible causes than most of the other types of color distortion. It does not involve an operational control at the receiver, but the receiver *can* cause this effect if the circuits, especially the quadrature-transformer adjustment, are faulty.

A symptom of quadrature distortion is color displacement; in severe cases, a girl's red lips can be in the middle of one cheek. In the more usual case, there is color fringing (not caused by camera misregistration or a misconverged picture tube) at the edges of color transitions.

The most obvious type of quadrature distortion occurs when I and Q are not phased exactly  $90^\circ$  in the encoder. Fig. 3-36 illustrates the case in which Q lags I by more than  $90^\circ$ . For simplicity, only the basic colors in the first and third quadrants are plotted. Remember the polarities of I and Q in each quadrant. These are:

- Quadrant 1 is bounded by +Q and +I (same polarity of I and Q).
- Quadrant 2 is bounded by +I and -Q (opposite polarities of I and Q).
- Quadrant 3 is bounded by -Q and -I (same polarity of I and Q).



**Fig. 3-36. Q lagging I more than 90°.**

Quadrant 4 is bounded by  $-I$  and  $+Q$  (opposite polarities of  $I$  and  $Q$ ).

The relative amplitude values of  $I$  and  $Q$  for the colors plotted are as developed in Table 3-2. Note that colors with large amounts of  $Q$  are affected more than others. For example, red and cyan have relatively small amplitude and phase errors; green and magenta have larger amplitude and phase errors. By adjusting the hue control on the monitor or receiver, you cannot get a good red and cyan *simultaneously* with a good green (green will be yellowish) or magenta (magenta will be bluish).

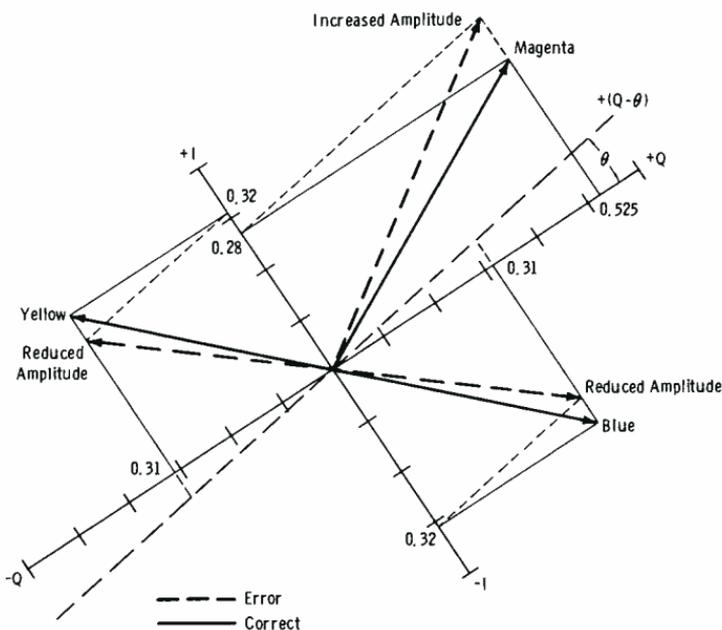
Now if you will take the trouble to plot yellow and blue (second and fourth quadrants), you will find these increased in amplitude. For a  $Q$  lag greater than  $90^\circ$ , colors in the first and third quadrants

Table 3-2. Color System Relationships for Primaries and Complements

Transmitted Color	$E_G$	$E_R$	$E_B$	$E_Y$	$G - Y$	$R - Y$	$B - Y$	$Q$	$I$
Green	1	0	0	0.59	0.41	-0.59	-0.59	-0.525	-0.28
Yellow	1	1	0	0.89	0.11	0.11	-0.89	-0.31	+0.32
Red	0	1	0	0.3	-0.3	0.7	-0.3	+0.21	+0.60
Magenta	0	1	1	0.41	-0.41	0.59	0.59	+0.525	+0.28
Blue	0	0	1	0.11	-0.11	-0.11	0.89	+0.31	-0.32
Cyan	1	0	1	0.7	0.3	-0.7	0.3	-0.21	-0.60

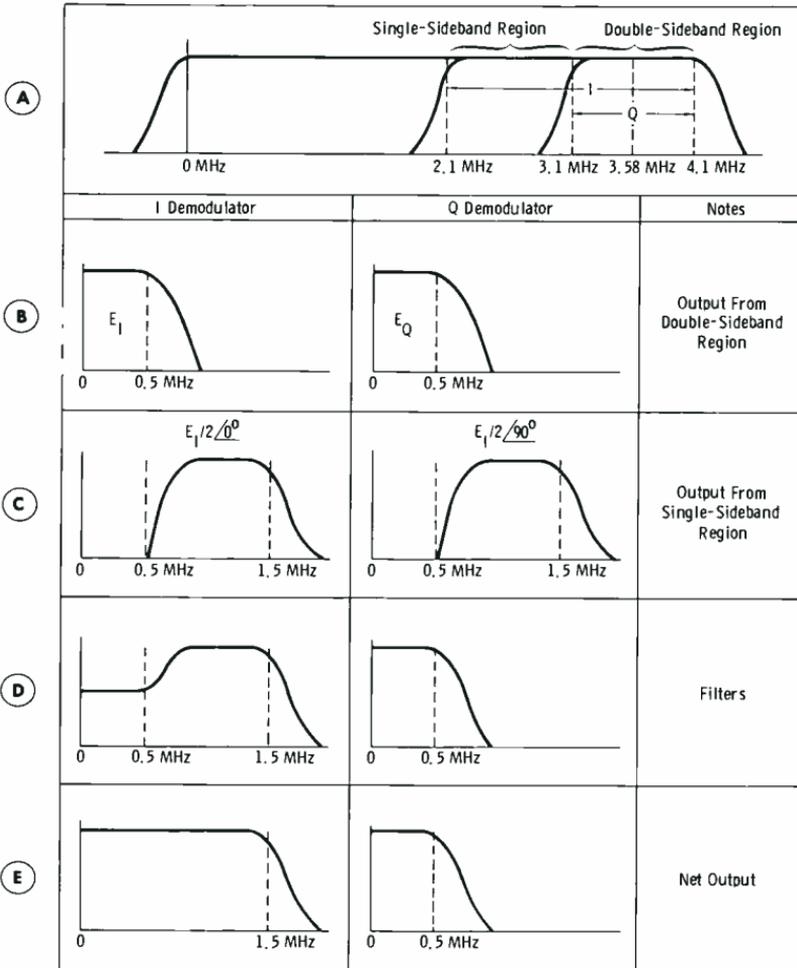
are reduced in amplitude; those in the second and fourth quadrants are increased in amplitude. In each quadrant, the phase error is in the direction of the  $Q$  phase error.

In Fig. 3-37,  $Q$  lags  $I$  by less than  $90^\circ$ . For simplicity, only magenta is shown in quadrant 1. Note that it is now increased in amplitude, whereas yellow and blue are reduced in amplitude. In either case, the phase error is in the direction of the  $Q$  error. Since the receiver separates signals with a  $90^\circ$  relationship,  $Q$  will cross-talk into  $I$  and vice versa when the phase angle between them is different from  $90^\circ$ .

Fig. 3-37.  $Q$  lagging  $I$  less than  $90^\circ$ .

Before going further, be sure to grasp the fundamentals of chrominance-signal transmission and reception for NTSC color. See Fig. 3-38 and the following analysis.

Fig. 3-38A represents the signal as transmitted. The I and Q components are double sideband in the region shown. The upper



**Fig. 3-38. Response of I and Q demodulators.**

sideband of the wideband I chrominance is cut off at the transmitter to achieve a 20-dB rolloff at the sound carrier frequency. A portion of the lower sideband of the I chroma constitutes single-sideband information; no Q chroma exists there.

The outputs of the I and Q demodulators are equal in the double-sideband region (Fig. 3-38B). Over the single-sideband region, the

voltage output of the I demodulator is one-half that which occurs in the double-sideband region (Fig. 3-38C). Note also that a one-half-I voltage, shifted in phase by  $90^\circ$ , appears at the Q-demodulator output. This is  $E_I$  at its single-sideband frequencies of about 0.6 to 1.5 MHz.

The output of the I demodulator is boosted by 6 dB above 0.5 MHz to recover the gain lost in the single-sideband region (Fig. 3-38D). The Q demodulator is limited in bandwidth to 0.5 MHz.

The filtering and relative gain action of Fig. 3-38D results in voltages  $E_I$  and  $E_Q$  free of cross talk (Fig. 3-38E). This assumes, of course, that I and Q are actually being transmitted in the proper quadrature relationship.

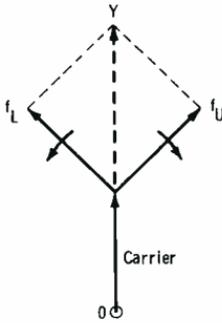
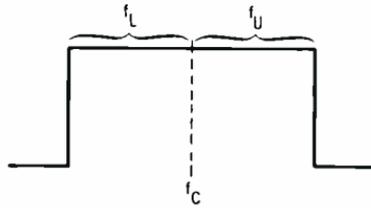
“Narrow-band” color receivers demodulate on the R – Y and B – Y axes. These receivers use the same bandwidth for all chrominance components. With all chrominance channels of the same bandwidth, delay equalization is unnecessary, and no cross talk occurs in a properly aligned receiver of this type. Again, this assumes that the I and Q chroma signals are being transmitted in the proper quadrature relationship. Also, there are other causes of lack of quadrature than misadjustment of the Q lag.

To investigate other causes of quadrature distortion, study Fig. 3-39. Fig. 3-39A shows the usual double-sideband representation of a modulated carrier. If, for example, the highest modulating frequency is 0.5 MHz, upper sidebands ( $f_U$ ) extend 0.5 MHz above  $f_C$ , and lower sidebands ( $f_L$ ) extend 0.5 MHz below  $f_C$ . This can be represented by equal-amplitude phasors (vectors) rotating in opposite directions as in Fig. 3-39B. The resultant amplitude is the vector sum of  $f_U$  and  $f_L$ , added to the carrier vector so that the resultant always lies along carrier line OY.

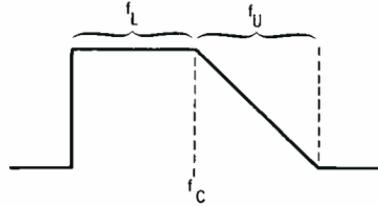
In Fig. 3-39C, the amplitude-frequency response exhibits a rapid rolloff above the carrier frequency. Thus  $f_U$  is severely attenuated (Fig. 3-39D). The resultant vector no longer lies along line OY, but contains a quadrature component, as shown in Fig. 3-39E. If  $f_C$  is the color subcarrier frequency, this type of sloping response will result in cross talk in both the I and Q detected signals in the video-frequency range of 0 to 0.6 MHz. It makes no difference whether I-Q or color-difference demodulation is used; the result is the same, cross talk among all the colors.

Fig. 3-39F shows this as applied to transmitter envelope response. A rapid rolloff too close to the color subcarrier frequency of 3.58 MHz will result both in desaturation of colors and in quadrature cross talk. Fig. 3-39G shows rapid variations in response around 3.58 MHz. Although the FCC allows a  $\pm 2$ -dB variation, the Rules further state that this variation must be substantially smooth. Sudden dips or peaks must be avoided for good color transmission.

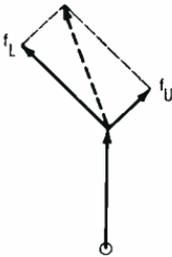
(A) Double-sideband signal.



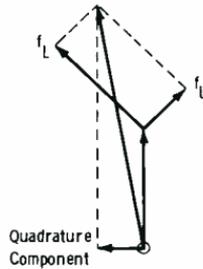
(B) Double-sideband vectors.



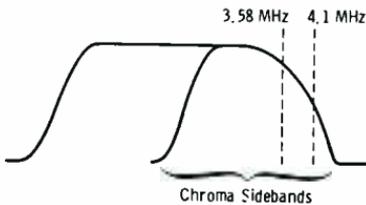
(C) Modified response.



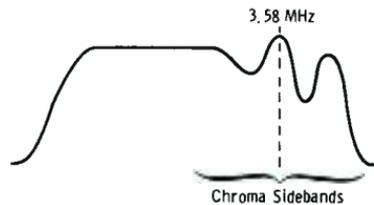
(D) Modified vectors.



(E) Quadrature component.



(F) Excessive rolloff.



(G) Uneven response curve.

Fig. 3-39. Causes of quadrature distortion.

Envelope-delay distortion at the transmitter is a major contributing factor in color misregistration. This subject is covered in Chapters 10 and 11.

### Effect of Carrier Unbalance

In a doubly balanced modulator, the carrier is suppressed so that only the sidebands remain. If this suppression is not perfect, the carrier appears in the output, and a condition known as *carrier unbalance* exists. Under this circumstance, the carrier adds itself vectorially to all vectors present in the encoder output. To visualize this, study Fig. 3-40A, which represents a carrier unbalance in the

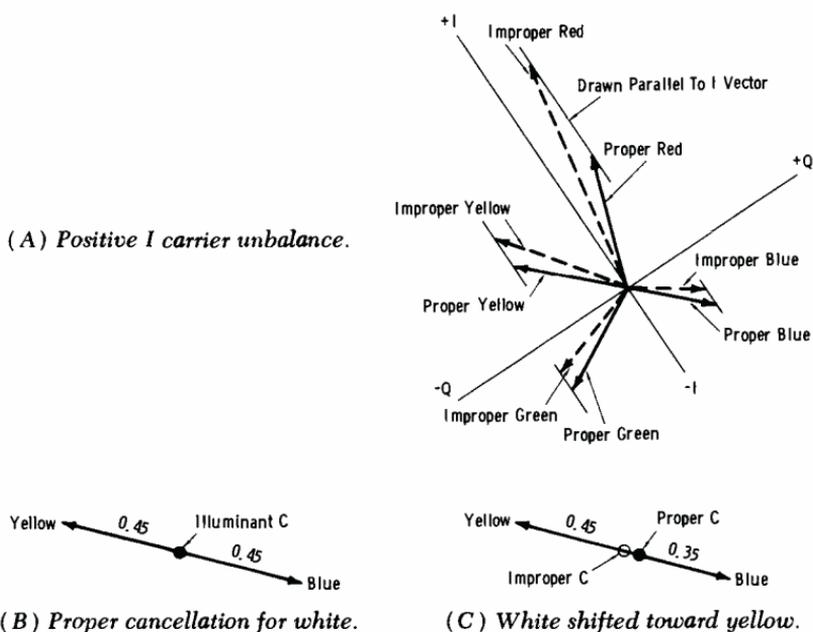


Fig. 3-40. Effect of carrier unbalance.

positive direction of the I modulator. Since the unbalance occurs in the I modulator, a new line, parallel to the I axis, is drawn from the proper color vector to the new vector representing the amount of carrier present. Since the unbalance is in the positive direction, the new vector is toward the +I axis. The resultant colors are shifted toward the orange axis of the +I vector, as well as being changed in amplitude.

Now see Fig. 3-40B. Recall that primary and complementary colors are opposite in phase. If both yellow and blue have the same amplitude, the result of their vector addition is illuminant C, or white. This is the proper complementary relationship. But note from

the vectors of Fig. 3-40A that the blue amplitude has been reduced and the yellow amplitude has been increased. You see the result in Fig. 3-40C: white or gray areas become colored because of incomplete cancellation of the subcarrier. Remember that "white" or "gray" can occur only during an interval of zero subcarrier.

Carrier unbalance shifts all hues (as well as whites and grays) in the direction of unbalance. A positive I unbalance shifts toward orange; a negative I unbalance shifts toward cyan. A positive Q unbalance shifts toward yellow-green; a negative Q unbalance shifts toward purple.

### Effect of Video Unbalance

Recall that double balancing of the modulators means that both the carrier and the modulating video are balanced out, leaving only the sidebands of the subcarrier frequency. If I and Q video suppression is not complete, the condition is known as *video unbalance*.

See Fig. 3-41A. The outputs of the I and Q channels for the indicated color bars are shown. Fig. 3-41B shows the result of a positive Q video unbalance. Note that the axis for all colors with plus values of Q is shifted in the positive direction. However, the actual peak-to-peak values of these colors remain the same. The net result is that the unwanted video signal is added to the *luminance* signal after the chroma signal is combined with the luminance signal, and the gray scale of the picture is distorted. Note that the effect of a positive Q video unbalance is to brighten reds, blues, and purples, and to darken yellows, greens, and cyans.

For a negative video unbalance, the colors with negative amounts of Q would be shifted upward. In this case, reds, blues, and purples would be darkened, and yellows, greens, and cyans would be brightened.

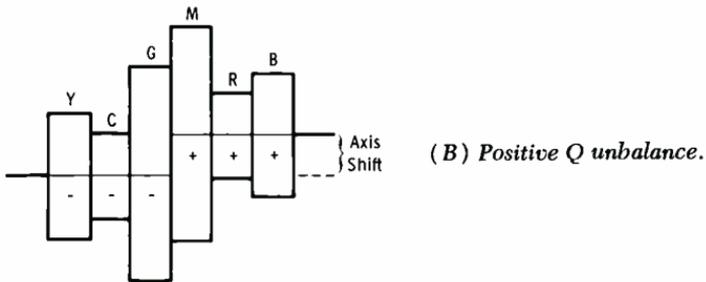
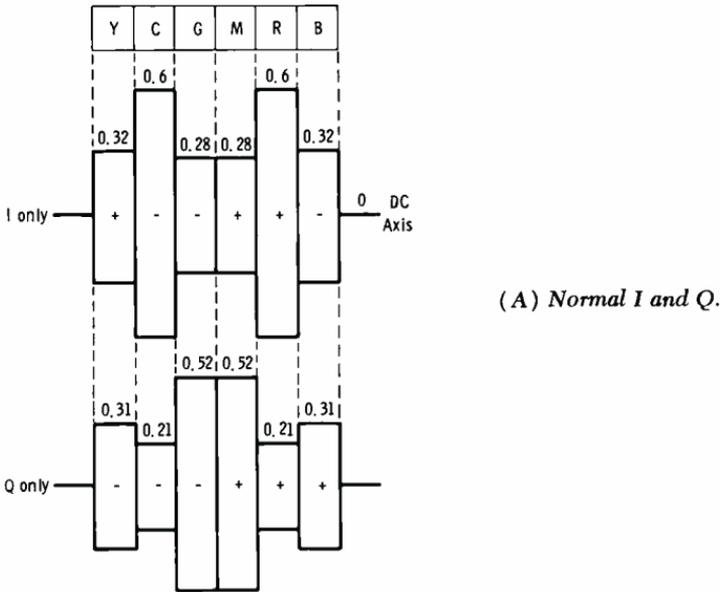
### Effect of Chroma Gains and Gain Ratio

The transmission paths from encoder input to receiver matrix must maintain a constant ratio for Y, I, and Q. A variation of gain in any one of the paths results in loss of color fidelity.

The noncomposite luminance level for a color-bar pattern (Fig. 3-42A) is 0.7 volt to peak white. When chrominance gain is correct (Fig. 3-42B) and of the proper I-to-Q gain ratio, and chrominance is added to the luminance signal of Fig. 3-42A, we have the following condition (see Fig. 3-42C):

1. Bars 1 and 2 overshoot by 33 percent.
2. Bars 5 and 6 undershoot by 33 percent.
3. Green (bar 3) just touches black level.

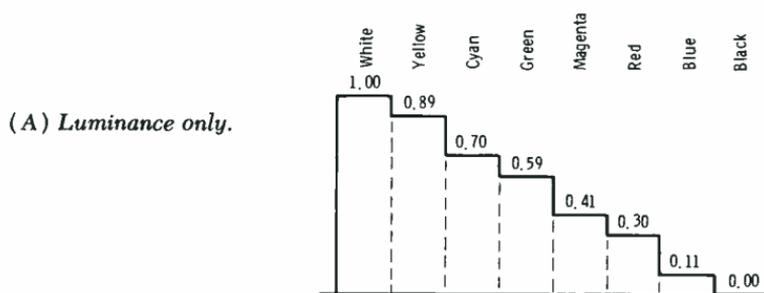
The above assumes that 100-percent bars are used and that no blanking (pedestal) is inserted in the signal (zero setup).



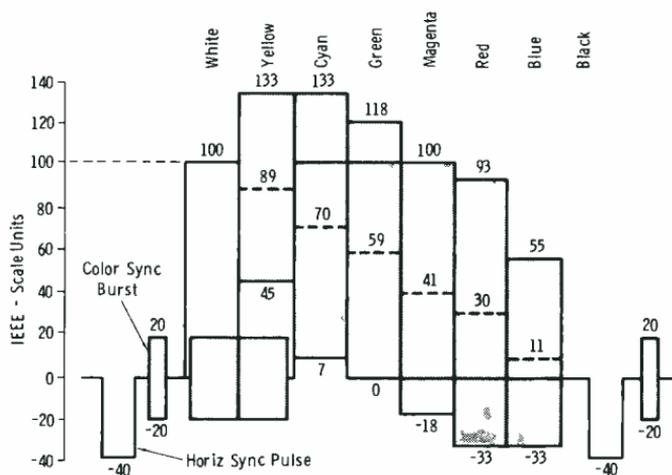
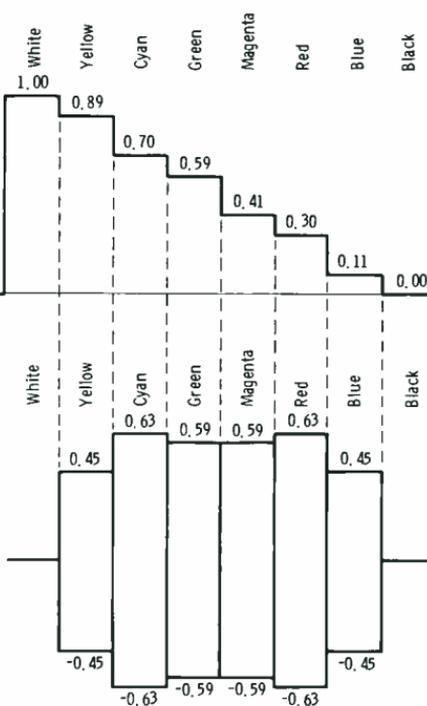
**Fig. 3-41. Effect of video unbalance.**

Now see Fig. 3-43. This is the same presentation as Fig. 3-42B except that the values of I and Q are shown for each color bar. Suppose that the ratio of I gain to Q gain is not correct. As you would expect, a deficiency of I gain would reduce the saturation of colors in the orange-cyan gamut, leaving greens and purples practically unaffected. Conversely, a deficiency of Q gain would reduce saturation of greens and purples without practical difference in the orange-cyan region. By noting the relative I and Q levels making up each color as in Fig. 3-43, you can understand how the pattern of Fig. 3-42C would show these deficiencies on a scope:

1. If the I gain is high relative to the Q gain, bar 2 will be higher than bar 1, and bar 5 will be lower (greater undershoot) than bar 6.



(B) *Chrominance only.*



**Fig. 3-42. Composition of 100% color bars, zero setup.**

2. If the Q gain is high relative to the I gain, bar 1 will be higher than bar 2, and bar 6 will have greater undershoot in the black region than will bar 5.

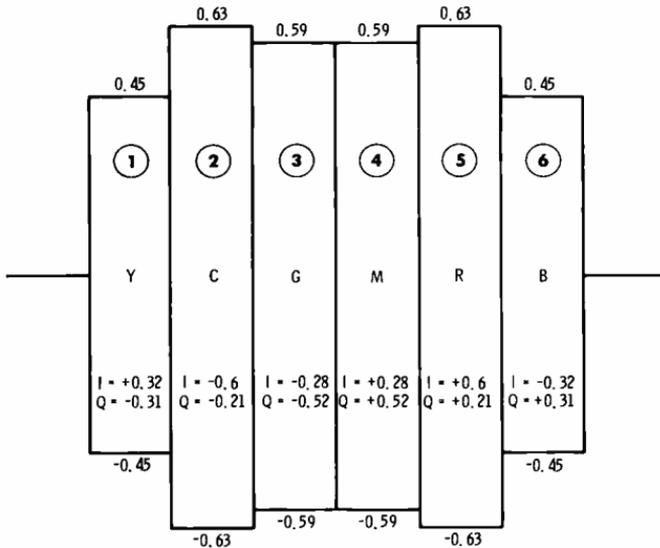


Fig. 3-43. I and Q chroma amplitude ratios.

### Effect of Differential Gain

*Differential gain* means that the gain of the 3.58-MHz chroma information is not constant with brightness level. This results in a change of saturation sensation with brightness.

Fig. 3-44A represents a stair-step signal with a superimposed 3.5-MHz sine wave. Fig. 3-44B shows the same signal observed through a high-pass filter to eliminate the low-frequency steps; a system with strictly linear amplitude response will result in sine waves of equal amplitudes for each step, as shown.

Fig. 3-44C represents the type of nonlinearity in which black regions are compressed and white regions stretched. Normally, this condition will be apparent also on the steps, as shown, and would be evident when the signal is passed through a low-pass filter to observe the steps only. However, it is possible to have linear low-frequency amplitude response and nonlinear high-frequency amplitude response. This is why the use of low- and high-pass filters is convenient in observing test signals of this type. Fig. 3-44D illustrates the opposite type of nonlinearity, and the same conditions apply.

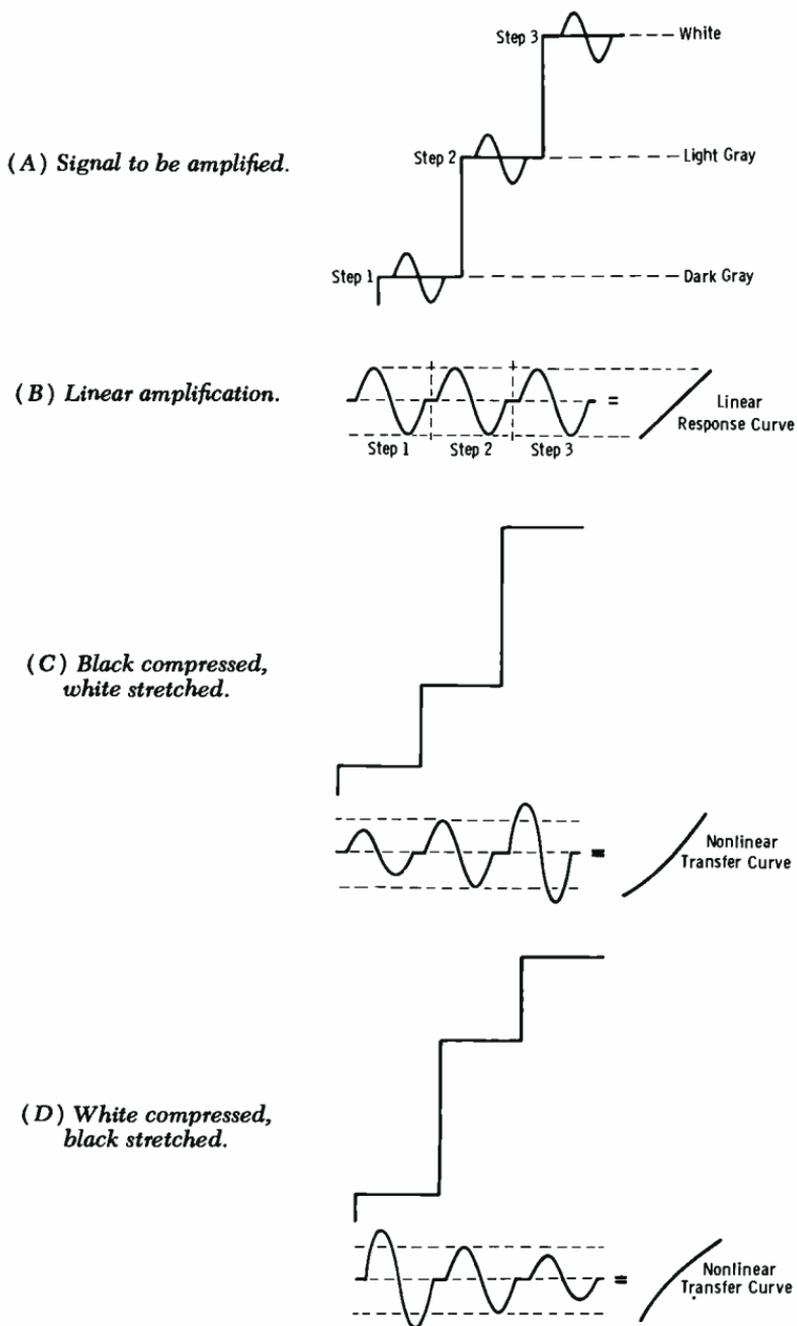


Fig. 3-44. Effects of differential gain.

The transfer curves of Figs. 3-44C and 3-44D are unusual. Generally, you will find that amplitude nonlinearity occurs either at one end or the other, or both ends, with a relatively linear middle response. This tells you that those colors near the white or black extremes normally will be most susceptible to saturation changes, particularly with highly saturated colors.

### Effect of Differential Phase

Fig. 3-45A shows the same signal as that pictured in Fig. 3-44A. Although the sine waves may look the same on each step, a phase displacement can occur with brightness level, as shown by Fig. 3-45B. This error is termed *differential phase*.

Remember that the subcarrier phase carries *hue* information. A low-brightness yellow should be the same as a high-brightness yellow. This is not to say that the two yellows would appear the same on the receiver. But the point is, the observed color should be yellow and not (for example) green or red as the brightness of the yellow component changes.

In practice, the effect of differential phase is judged best in the yellow and blue areas (the two extremes of the luminance scale). A system introducing as much as  $10^\circ$  of differential phase can result in a monitor or receiver adjustment that gives proper reproduction of a high-luminance hue such as saturated yellow, or a low-luminance hue such as saturated blue, but not both simultaneously. One or the other will be off-color. When this defect is accompanied by more than 10 percent differential gain (as often occurs), the error becomes quite noticeable.

### Checking Subcarrier Countdown and Color Lock

The station synchronizing generator is covered in Chapter 4. We will discuss here only those characteristics that involve the

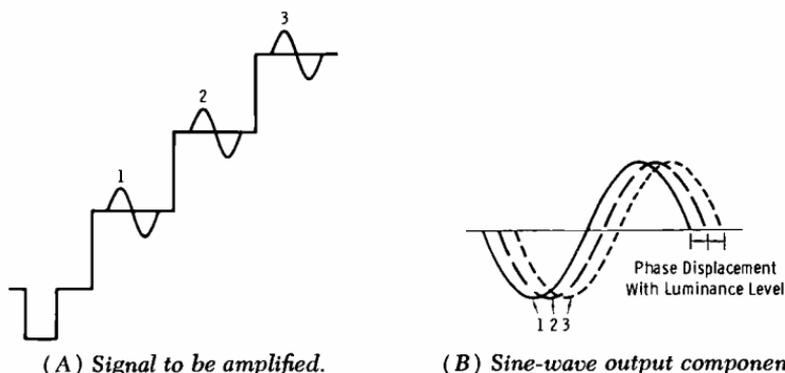


Fig. 3-45. Differential phase.

subcarrier feed to the color encoder. Many late-model sync generators (digital type) have a minimum of adjustable controls.

The following steps constitute a final check on the countdown of the subcarrier generator:

1. Connect the vertical input of the scope to the 3.58-MHz generator output.
2. Use external trigger for the scope horizontal sweep and trigger from the 31.468-kHz master oscillator in the sync generator.
3. The pattern on the scope should lock to a steady trace with no tendency to drift or jump. Temporarily remove the external trigger to be certain the scope is under external trigger control. Critically adjust the trigger-amplitude control on the scope for a "clean" trace.
4. Reversing the above connections to the scope should also produce a steady trace.

For a final check on color lock of the sync generator, these steps may be followed:

5. Place the sync generator on internal crystal control, and check all counters in the generator for proper centering.
6. Place the sync generator on free-run operation. Observe any vertical output (composite sync, blanking, or vertical drive) with the scope on 60-Hz trigger. Adjust the master-oscillator frequency for a very slow drift of the trace from left to right on the scope. Check the sync-generator afc operation with 60-Hz line lock.
7. Place the sync generator under color control. Connect the scope vertical input to the 3.58-MHz output of the color-subcarrier generator. Trigger the scope horizontal sweep with horizontal drive from the sync generator. The resulting pattern should be perfectly stationary with no erratic jumps.
8. If this condition does not exist, recheck the color-subcarrier generator as in steps 1 to 4. If the subcarrier generator is not at fault, the trouble is in the afc (or apfc) circuitry of the sync generator.

### Setting the Color-Subcarrier Frequency

There are four quite satisfactory methods for setting the subcarrier frequency. Of course, only Method 1 gives positive assurance that the frequency tolerance established by the FCC is being satisfied by the station.

*Method 1: Frequency Standard and Primary Measuring Service—*With this method, the station may or may not have a color-subcarrier frequency meter. The procedure is to contact the frequency-measur-

ing service by telephone and transmit a composite color signal for purposes of measuring the subcarrier burst frequency. If the station has a subcarrier-frequency meter, adjust it to agree with the measuring-service reading, and then adjust the subcarrier frequency for zero reading. Or, the service can "talk you in" to the proper frequency adjustment.

*Method 2: Vectorscope*—If the station is affiliated with a network, or if the output of a good color receiver can be fed to the alternate input of the vectorscope, you can make use of a rapid and convenient way of adjusting the color-subcarrier frequency. In this method you must assume, of course, that the network or the received station is using the proper subcarrier frequency. Feed the external composite color signal to the B input of the vectorscope and feed the local composite color signal to the A input. Place the vectorscope on external 3.58-MHz lock from the subcarrier generator (normal operation in station use). Place the input switch in the "A shared with B" position. The local burst, locked to the local subcarrier generator, will be stationary, and the signal being used as a standard will have its burst rotating at a rate dependent on the frequency difference. Adjust the local subcarrier frequency until the "standard" burst is as nearly stationary as possible. You cannot obtain a steady lock unless you have color genlock facilities.

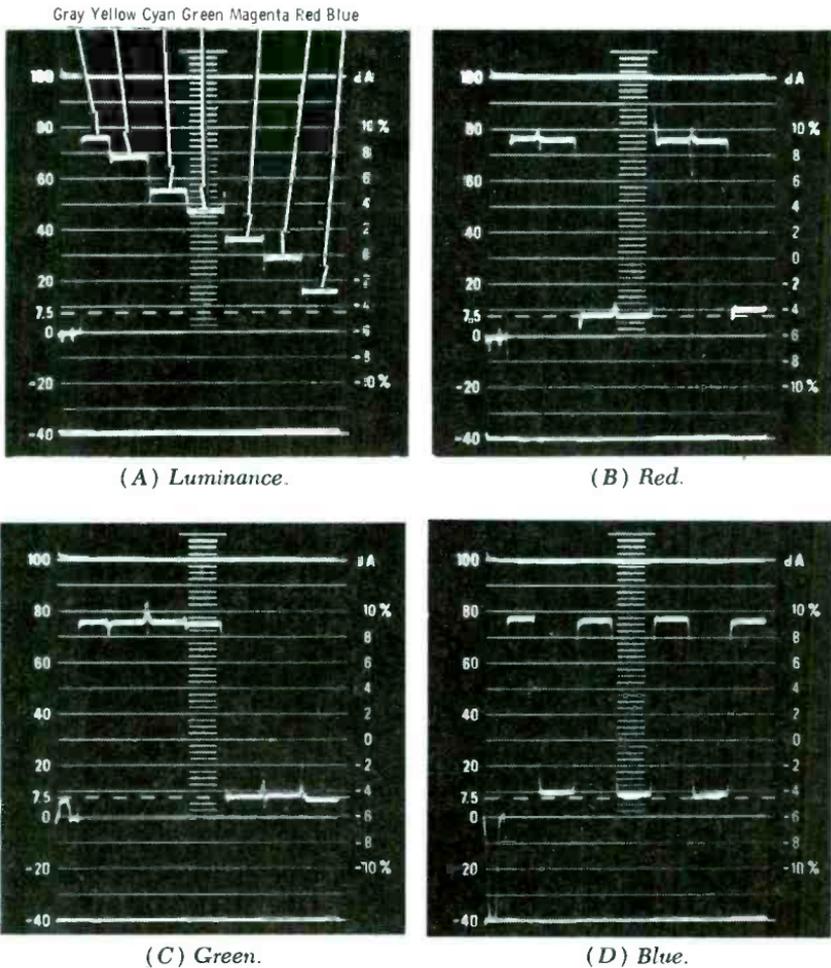
This is an extremely quick and convenient procedure if a signal-selector switch is incorporated with the B vectorscope input, so that you can "punch up" the reference frequency at a moment's notice.

*Method 3: Oscilloscope*—If you do not have a vectorscope, the next best procedure is to feed the reference composite color signal to the external trigger input of an oscilloscope. Trigger the scope at the horizontal frequency. Observe 10 to 12 cycles of the local 3.58-MHz subcarrier on the scope triggered by the reference signal. Adjust the subcarrier frequency to obtain a stable trace.

*Method 4: Dot Crawl*—This method should be used only in an emergency when you do not have any other means of checking the subcarrier frequency. With a good stop watch, time the travel of a single dot from the bottom to the top of the raster on an *underscanned* monitor. The dots are quite apparent at color-bar transitions. Adjust the subcarrier frequency until this time is between 8 and 8.1 seconds.

### Use of the Vectorscope

The use of the vectorscope in setting quadrature and burst phase was covered in Chapter 10 of the preceding volume in this series, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*, and will not be considered here. The same is true for



Courtesy Tektronix, Inc.

Fig. 3-46. Line-sweep presentations of Y, R, G, and B decoded video.

adjustment of the breezeway and the number of color-burst cycles. That chapter should be reviewed at this time.

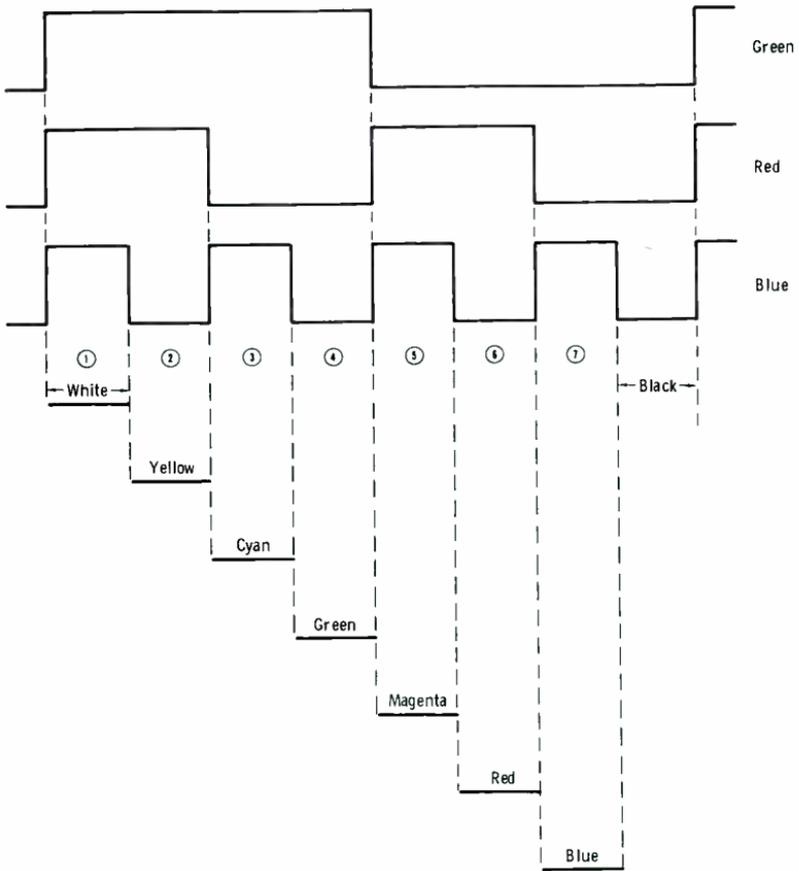
Use of the vectorscope in differential gain and phase measurements is more appropriately covered under complete systems measurements in Chapter 8 of this book.

### R, G, B, and Y Observations

The vectorscope provides a luminance channel that permits the separation and display of the luminance (Y) component (Fig. 3-46A) of the composite color signal. The Y component also can

be combined with the outputs of the chrominance demodulators for R, G, and B displays at a line rate ( Figs. 3-46B, 3-46C, and 3-46D). Amplitude measurements of color-signal components can be made with an accuracy of 3 percent.

A great deal of valuable information can be obtained by this line-sweep presentation of Y, R, G, and B. To take full advantage of such measurements, see Fig. 3-47 and Table 3-3. These review the color-bar-generator output pattern.



**Fig. 3-47. Color-bar generator outputs.**

Note from Fig. 3-46 that the decoded R, G, and B video signals for an encoder that is properly set up have the same maximum amplitude and minimum amplitude per step. The latter falls on the 7.5-percent setup level.

Table 3-3. Color-Bar Pattern

Steps	Hue	Primaries	Common Factor
1	White	R+G+B	Green
2	Yellow	R+G	
3	Cyan	G+B	
4	Green	G	
5	Magenta	R+B	Red
6	Red	R	
7	Blue	B	Blue
8	Black	0	

All encoders, whether four- or three-channel cameras are used, convert R, G, and B to luminance for color bars. In the three-channel systems, the same is true for the camera signals. This is termed *derived* luminance.

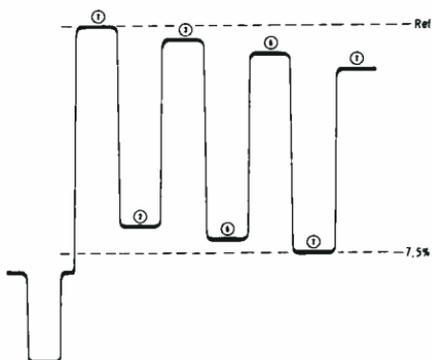
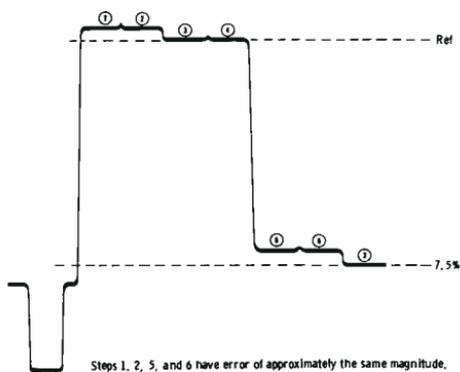
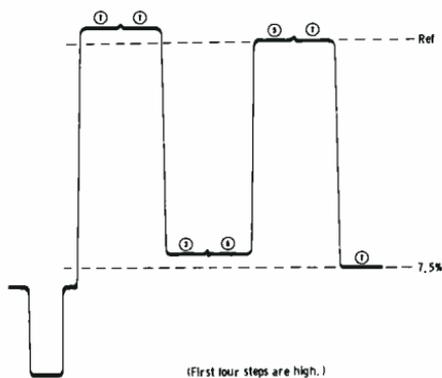
Fig. 3-48 shows errors that can be displayed by line-sweep presentations of decoded R, G, and B video on the vectorscope. Errors of this type occur in the matrix that converts the R, G, and B signals to luminance. There may be a change of actual values of matrix resistors, or there may be an improper setting of an adjustable value.

Fig. 3-48A shows decoded red video. Note that the first four steps are incorrect in value and (from Table 3-3) that green is the common factor. Therefore, an error in the green portion of the luminance matrix results in approximately the same error in the first four steps of the *red* or *blue* decoded video display.

NOTE: It is helpful for the technician to memorize the actual sequence of the color-bar pattern: gray, yellow, cyan, green, magenta, red, and blue, and also the primaries used to form yellow, cyan, and magenta.

Fig. 3-48B shows decoded green video. Steps 1, 2, 5, and 6 have an error of approximately the same magnitude. Note from Table 3-3 that red is now the common factor, and an error in the red part of the luminance resistive matrix is indicated.

Fig. 3-48C shows a type of distortion more likely to occur in the video distribution and switching system following the encoder. Remember that the decoded output of the blue channel is made up of only 11 percent luminance. Thus, the majority of the signal contributing to decoded blue is chrominance. Therefore, the decoded blue signal is most sensitive to errors in the luminance-to-chrominance ratio. The downward slope shown in Fig. 3-48C indicates that the chrominance is low in relation to luminance.



**Fig. 3-48. Luminance errors in resistive matrix of three-channel encoder.**

A downward slope can indicate either that luminance is excessive or that chrominance is deficient. Bear in mind that the first step (gray or white) is all luminance and zero chrominance. Thus, noting the amplitude of the first blue bar on the left indicates whether the luminance signal or the chrominance signal is in error. Then the *slope* of the decoded blue indicates the ratio of luminance to chrominance. An upward slope indicates that chrominance is excessive in relation to luminance.

### 3-7. PICKUP TUBES (SENSORS)

Camera pickup tubes are often called "sensors" in modern terminology, and we will use this term to cover the types of modern tubes used in color cameras. The antimony trisulphide vidicon was first developed, and it is still very common in film-chain (telecine) use. The more recently developed lead-oxide vidicon is used primarily in studio and portable-mobile cameras for "live" pickup. There are many variations in both types, but maintenance is fairly well established and almost identical for all versions.

A "bad" antimony trisulphide vidicon (telecine) as a general rule will show excessive *target lag* before it deteriorates in resolution. Target lag is image retention under vertical or horizontal movement of the images. Such a vidicon can be retired to a slide (only) camera if such is used, since no image movement (except slide change) is involved. After considerable extended use in the slide camera, the vidicon will go "soft" and lose resolving power.

There is one difference between the two basic types of vidicons when the problem is a deteriorating thermionic cathode. In the antimony trisulphide type, if the beam current is slightly excessive, a split image results. In the lead-oxide type, the beam is simply insufficient for the desired highlight operating characteristics. This problem can, of course, be caused by low filament voltage or incorrect cathode voltage, and these should be checked.

Dark current in lead-oxide vidicons is extremely low. In some cameras, a type of vidicon is used that employs a light bias which introduces a small amount of dark current in each tube to neutralize the lag of the three sensors.

Troubles peculiar to the lead-oxide type are:

1. *Dark-current breakdown.* When the barrier layer of the photoconductor breaks down, an extremely high dark current results. Breakdown usually starts around the periphery of the scanned area and then moves into the scanned area, particularly in the corners of the picture, where it produces objectionable flare. At the time of this writing, there is no cure for this defect.

2. *Blue slump.* This is noted when a tube begins to develop a negatively burned-in pattern from specular highlights. If a tube begins to slump, the effect can be readily observed in the blue channel of the camera by exposing the tube to light reflected from a blue background. Sensitivity as observed on a waveform monitor tends to slump first in the central part of the picture. As the problem becomes worse, the blue signal will be fairly high when the camera lens is first uncapped, but it will drop rapidly to a lower equilibrium level.

If you are using a normal target voltage of 45 volts, you can increase this to the maximum of 50 volts, which might eliminate the slump for a short period of time. Another way of extending the life of the sensor is to interchange it with the red tube in the camera. This is possible in many cases because the red sensitivity of the sensor seems to increase as blue slump develops.

A slumping characteristic can sometimes occur in the green or red channel; in this case the sensor has a "spongy" response. When the lens is first uncapped, the red signal will be slow in building up to its equilibrium value. There is no remedy in the red or green channel other than to try raising the target voltage to the maximum 50-volt value.

Spots in either type of vidicon are always a problem. In the antimony trisulphide type, there is no cure other than replacement. In the lead-oxide type, certain types of spots can be reduced in contrast or made to disappear entirely by slightly reducing the target voltage. This can be done unless the sensitivity drops drastically or an increase in burn-in is noticed.

Always remember to operate spare vidicons for several hours at monthly intervals, or at no longer than three month intervals. Following this practice helps to prevent the formation of gas within the sensor envelope.

Characteristics of the *Saticon* (Selenium-Arsenic-Tellurium) photoconductor are essentially the same as those of the lead-oxide type, except that a higher degree of tube-to-tube uniformity results from the fact that the manufacture of this target can be controlled more carefully than the manufacture of the lead-oxide target. Increasing image stickiness will result at higher operating temperatures. There is no special doping or construction of Saticons for red, green, or blue channels, and any tube can be used in any channel. An exception to this is the RCA practice of sorting Saticons into red, green, and blue solely on the basis of sensitivity. Interchanging of tubes into other color channels can be done in RCA cameras at the expense of sensitivity.

### 3-8. GENERAL PREVENTIVE MAINTENANCE

Preventive maintenance in general consists of:

1. Lubrication according to manufacturer's specifications
2. Cleaning of all equipment and optics at periodic intervals
3. Performance checks at periodic intervals.

The extent of the periodic interval is determined by the manufacturer's recommendations or by experience with a particular installation.

#### Cleaning

In general, cleaning procedures can be scheduled at daily, monthly, or longer intervals as follows.

*Daily*—Wipe and clean all exposed surfaces of the camera. Use only lens tissue on turret lenses or zoom-lens front surfaces. When necessary, use a cotton ball dipped in isopropyl alcohol, and use a light circular motion over the optical surface. Wipe the surface dry with a wad of lint-free cheesecloth or diaper cloth. If warranted, clean the faceplates of pickup tubes in the same manner. **WARNING:** Turn the power off when doing this.

*Monthly*—Inspect all exposed color optics and front-surface mirrors (film multiplexers) in the optical path. If necessary, clean dichroic surfaces, but use only pure grain alcohol or a 1-to-6 ratio of benzene and grain alcohol. Use a cotton ball and diaper cloth as described above. On front-surface mirrors, use isopropyl alcohol diluted with a small amount of distilled water (to prevent rapid evaporation), and wipe dry with diaper cloth *before* the solution on the mirror surface evaporates.

**NOTE:** It is very important, as a final step in lens and optics cleaning, to use a camel's-hair "static" brush to remove dust that inevitably falls on the surfaces because of static deposits.

*Every Three Months*—This interval depends entirely on the cleanliness of the operating area. In the case of portable and mobile use, these operations should be performed monthly (or more often).

Clean all air filters in the blower path of the camera and in the forced-air paths of the cooling systems of power supplies, etc. Use only the cleaning solvent that is recommended in the instruction book for the particular installation.

With a low-pressure air blower, blow out the camera head, control panels, and rack equipment to eliminate dirt, dust, and general contamination. Wipe clean with a soft paint brush and cloth.

**IMPORTANT NOTE:** At this time, it is imperative to make a thorough *visual inspection* (when possible) of component parts. Observe resis-

tors for discoloration or signs of heating. Inspect capacitors for signs of bulging or leakage. Inspect terminal boards for cracks and loose connections. Be sure all fuse mounts and fuse caps are tight. The time spent here is invaluable in preventing future breakdowns, which is the basic reason for performing preventive maintenance.

### Lubrication

Lubrication of modern camera chains normally is scheduled about every six months. For operation under extreme conditions of dust, heat, and humidity, lubrication procedures must be carried out more often. *Caution:* Some of the older camera chains employ fan and blower motors that require lubrication every 200 hours. Always check the instruction book for a particular system, and work the lubrication schedule into a check-off sheet at the recommended intervals.

Items that require lubrication include: turret shaft and detent rollers; turret gears and bearings; camera-head counterbalance linkage; focus, iris, and zoom mechanisms (which usually involve multiple gears, bearings, and shafts); and blower and fan motors. It is quite important to check the manufacturer's specifications for the type of oil or grease to use on the items involved.

It is important not to overlubricate. Always wipe off excess oil or grease, and use any special applicators that are supplied with a camera chain by the manufacturer.

### Plugs and Receptacles

There are two main types of plugs and receptacles used to interconnect the various components. The first type of plug is used with a coaxial line and consists of a metal shell with a single center pin that is insulated from the shell. When the plug is inserted into the receptacle, this pin is gripped firmly by a spring connector. There is a knurled metal ring around the plug; this ring is screwed onto the corresponding threads on the receptacle. The insulation in these plugs is heavy in order to withstand considerable voltage.

The second type of plug is used for connecting multiconductor cables. The plug usually consists of a number of pins that are insulated from the shell. The pins are inserted into a corresponding number of female connectors in the receptacle, although in some cases the plug has the female connectors in it and the male connectors are in the receptacle. This type of plug usually has two small pins or buttons that are mounted on a spring inside the shell and protrude through the shell. When the shell is properly oriented and placed in the receptacle, one of these pins springs up through a hole in the receptacle, locking the plug and receptacle together. To remove the plug, the other pin is simply depressed.

Connections between all plugs and their cables are made inside the plug shell. The cable conductor may be soldered to the pin, or there may be a screw to hold the wire to the pin. Remove the shell if it is necessary to get at these connections for repair or inspection. If there is a clamp holding the cable to the shell, loosen the clamp screws. Usually there are several screws holding the shell; these are removed and the shell is pulled off. In some cases, it is found that the shell and plug body are both threaded; then the shell may simply be unscrewed.

*Inspect:*

1. The part of the cable that was inside the shell for dirt and cracked or burned insulation.
2. The conductor or conductors and their connection to the pins for broken wires; bad insulation; and dirty, corroded, broken, or loose connections.
3. The male or female connectors in the plug for looseness in the insulation, damage, and dirt or corrosion.
4. The plug body for damage to the insulation and for dirt or corrosion.
5. The shell for damage such as dents or cracks and for dirt or corrosion.
6. The receptacle for damaged or corroded connectors, cracked insulation, and improper electrical connection between the connectors and the leads.

*Tighten:*

1. Any looseness of the connectors in the insulation. If tightening is not possible, replace the plug.
2. Any loose electrical connections. Resolder if necessary.

*Clean:*

1. The cable, using a cloth and cleaning fluid.
2. The connectors and connections, using a cloth and cleaning fluid. Use crocus cloth to remove corrosion.
3. The plug body and shell, using a cloth and cleaning fluid. Use crocus cloth to remove corrosion.
4. The receptacle, with a cloth and cleaning fluid if necessary. Use crocus cloth to remove corrosion.

*Adjust* the connectors for proper contact if they are of the spring type.

*Lubricate* the plug and receptacle with a thin coat of petroleum jelly if they are difficult to connect or remove. The type of plug with the threaded ring may especially require this.

### 3-9. THE ENG CAMERA

Electronic news gathering (ENG), also termed electronic journalism (EJ), must employ color cameras of the miniature variety to compete with the usual hand-held film camera formerly used. This miniaturized construction presents two basic areas of trouble, heat and service accessibility. Television sensors such as Plumbicons and Saticons are very sensitive to heat. Transistors and ICs must be provided with adequate heat-sinking arrangements.

Accessibility for maintenance is generally provided by plug-in circuit modules. An example is illustrated in Fig. 3-49A, which shows the left side of an RCA TK-76 camera with the cover removed. (The module partially removed contains a temperature-compensated crystal oscillator for sync.)

Circuit functions for the modules are normally divided as follows:

Video preamplifiers, one for each tube and mounted at the yoke assembly (not visible in Fig. 3-49)

Input processing amplifier

Output processing amplifier

"Automatics" circuit board

Horizontal and vertical contour (aperture correction) board, generally containing 1H and 2H delay lines.

Sync board (usually with genlock capability)

Deflection board

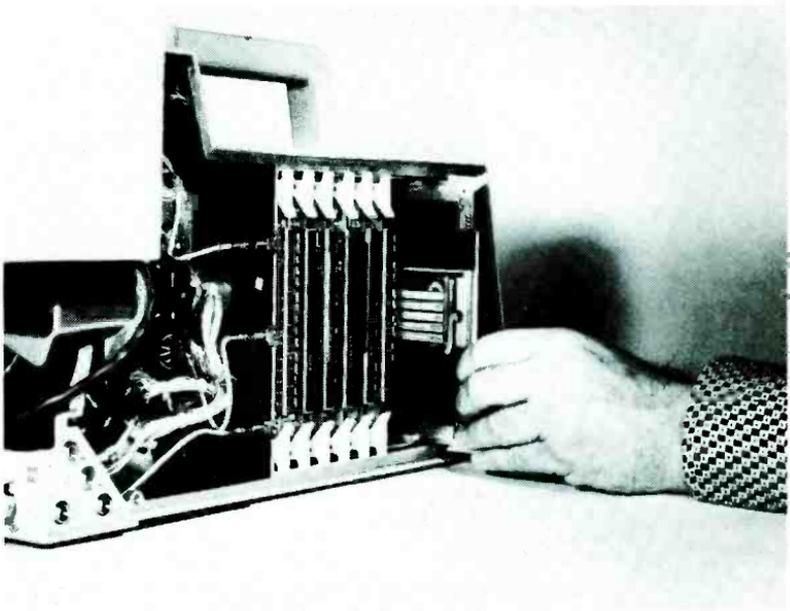
Luminance board

Chroma board (including color encoder)

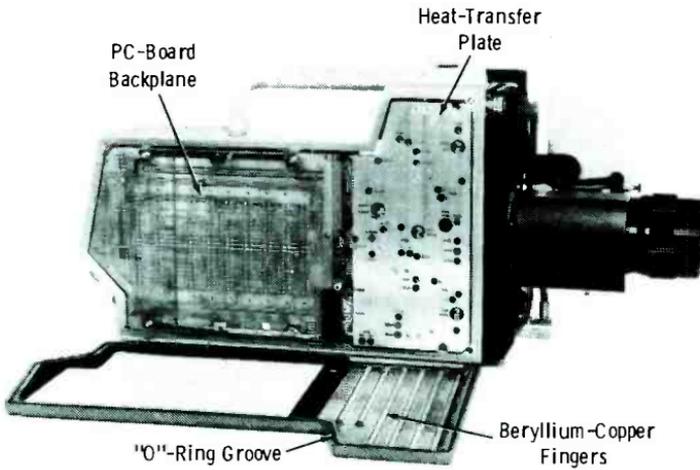
Power-supply board (low and high voltage, powered by external 12-volt battery belt)

When a problem occurs, the particular module involved is generally apparent to the experienced maintenance person because of the symptoms of the fault. Since two or more cameras are normally available, it is a simple matter to substitute for a suspected module with one from a working camera to verify the diagnosis. The faulty module may then be serviced when time permits. Modules may be placed on extenders for servicing in the camera.

In the RCA TK-76 camera shown in Fig. 3-49, the pickup-tube and yoke assemblies are thermally isolated and shock-mounted from the main camera body. Fig. 3-49B is a view of the right side showing how the cover interfaces with a heat-transfer plate by means of beryllium-copper contact fingers. This serves as a heat exchanger to prevent heat from entering the pickup-tube assembly. Camera setup adjustments are accessible through and identified on the plate. Both the right and left side covers are aluminum cast with



(A) Left side.



(B) Right side.

Courtesy RCA

**Fig. 3-49. Side views of RCA TK-76 ENG camera.**

a slight bow so that "O" ring gaskets will seal properly around their entire perimeter for waterproofing.

Registration and the color encoder are normally set up with the camera at home base (studio), where a vectorscope is available. When field pickups occur in locations of widely varying temperatures, it is also wise to include a vectorscope in the mobile van.

The sync generator quite often includes no adjustments whatever; it might be a single IC with clock inputs and composite sync and blanking, horizontal and vertical drive, and burst-flag outputs. Failure here is generally catastrophic; it either works or it does not work. If you are fortunate enough to have a spare camera or spare boards along, down time is negligible. Of course, the same can be said of any specific board in the camera.

You should always go through the parts lists of cameras and be sure to stock important spares. Find out what components, if any, might be hard to locate in local stores so that you can stock these items in advance. Many times, this simple procedure can prevent ill feelings between the user and the manufacturer when the complaint is, "Parts are hard to find and extremely slow in delivery." The same can be true, of course, for studio camera chains.

The preventive maintenance schedules given in Section 3-8 are extremely important to the well being of ENG cameras that take a lot of bumps and, in general, suffer much greater temperature extremes than studio camera chains.

## EXERCISES

- Q3-1. What is the purpose of the preregulator section of a regulated power supply?
- Q3-2. What is remote sensing, and why is it used in modern regulated power-supply systems?
- Q3-3. Why are oscillators in camera high-voltage supplies usually synchronized to the horizontal scanning frequency or to one-half this frequency?
- Q3-4. If D1 (Fig. 3-6) should open, what would happen?
- Q3-5. What is the proper amplitude of the color-sync burst?
- Q3-6. Does the color-sync burst occur following every horizontal-sync time in the complete composite color signal?
- Q3-7. How many cycles of color-burst signal should be present?
- Q3-8. What are the attenuation-vs-frequency requirements for a color transmitter?
- Q3-9. If you apply a linear stairstep signal to the first amplifier of a camera head, should the signal at the output of the color encoding system be linear?

- Q3-10. If you apply a linear staircase signal to the input of an encoder, should the signal at the output of the encoder be linear?
- Q3-11. What is the basic problem in handling NTSC color?
- Q3-12. What factors other than transmitting-system deficiencies can result in receiver color problems?
- Q3-13. How would the following colors be reproduced in a properly adjusted monitor or receiver if the burst in the encoder were at  $+160^\circ$  from zero reference on an NTSC polar diagram? (Assume all other adjustments to be correct.)
1. Yellow
  2. Cyan
  3. Green
  4. Magenta
  5. Red
  6. Blue
  7. White
- Q3-14. What could cause white to be contaminated with a certain hue?
- Q3-15. What would cause "washed-out" color in a properly adjusted receiver?
- Q3-16. What is the major criterion in the adjustment of a receiver for "good color"?
- Q3-17. What could cause a human face to be reproduced with unnaturally ruddy complexion and very dark red lips? Assume a properly adjusted receiver.
- Q3-18. What would cause yellows to be "washed out" while blues are reproduced well?

## The Synchronizing Generator and Pulse Distribution

Fig. 4-1 reviews the basic functions of any sync generator. All modes of operation involve control of a master voltage-controlled oscillator (vco) operating at twice the horizontal rate. In all modes except free-run, the master vco is phase-locked to a reference frequency in a feedback control system.

It is obvious, then, that the "heart" of any sync generator is the vco and automatic frequency-phase control (afpc) loop. It is pertinent at this time to review the basic circuitry involved.

### 4-1. THE VOLTAGE-CONTROLLED OSCILLATOR (VCO)

A simplified schematic diagram of one type of voltage-controlled oscillator is presented in Fig. 4-2. Transistors Q2 and Q4 form a cross-coupled multivibrator; Q2 and Q4 alternate with each other in the cutoff and saturated condition. The off time of Q2 is governed by C2 and R3 for a given base voltage. The off time of Q4 is similarly governed by C1 and R7. However, both of these off times are also dependent on the dc voltage applied to the junction of R3 and R7 through Q3.

Fig. 4-3 illustrates how this voltage is used to control the oscillator frequency. In waveform A, at t1 the base of Q2 is at a certain negative reference voltage, which we will call the voltage existing for the free-running position of the selector switch (Fig. 4-1). The common emitter return of the vco is near ground potential through the frequency-control module. Since the base of Q2 is negative with respect to the emitter, Q2 is in saturation and Q4 is cut off. Therefore, from t1 to t2, the collector of Q2 is near zero volts (pnp transistor). At t2, C1 has discharged through R7 so that Q4 saturates.

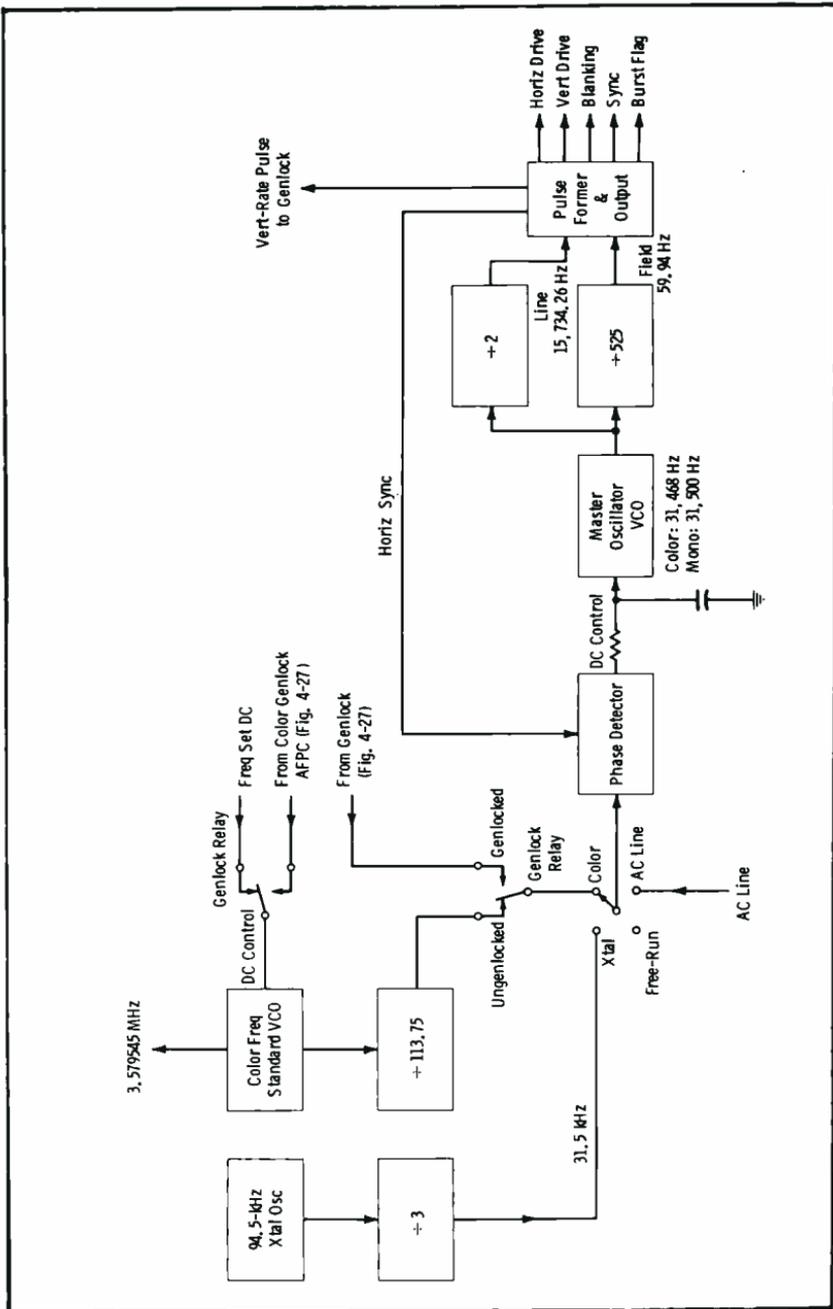


Fig. 4-1. Simplified block diagram of sync generator.

The resulting positive-going pulse applied to the base of Q2 cuts that transistor off, sending its collector to  $-12.5$  volts as shown (waveform B). From  $t_2$  to  $t_3$ , C2 discharges through R3, and when the reference base voltage is intercepted, conditions reverse for the time  $t_3$  to  $t_4$ . Remember that Q2 supplies one alternation of a complete cycle, with Q4 supplying the opposite alternation.

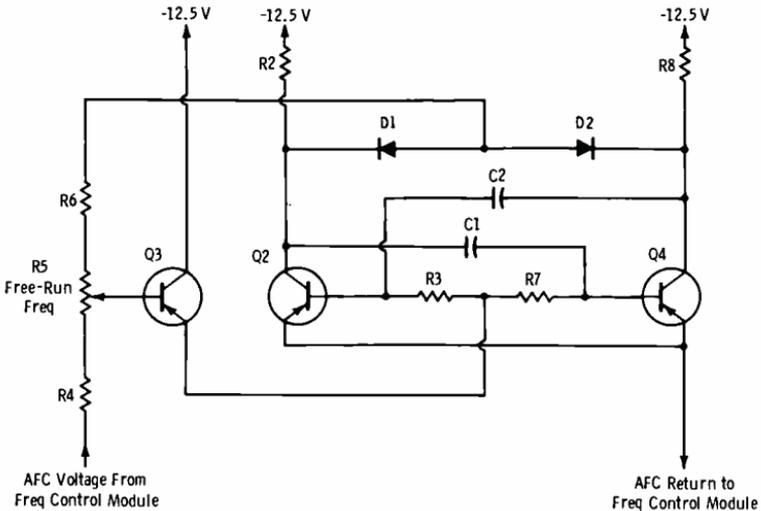


Fig. 4-2. Voltage-controlled oscillator.

Now let us examine what happens when the base voltage of Q2 and Q4 are made less negative (more positive) through Q3. As shown in waveforms C and D of Fig. 4-3, the discharge curve intercepts the cutoff point at an earlier time, increasing the frequency of the vco. Conversely, as shown in waveforms E and F, when the reference base voltage is made more negative, a longer time is involved in intercepting the cutoff point, decreasing the vco frequency.

Another type of vco employs a varicap (voltage-variable capacitance diode) as shown in Fig. 4-4. The varicap is always operated in the reverse-voltage mode, and the capacitance is dependent on the magnitude of the reverse voltage. The LC product is made to correspond to the desired free-run operating frequency by adjusting R1 with the afc voltage grounded. Application of the afc voltage then controls the frequency of the oscillator. Troubles in the varicap type of vco are normally cleared by direct replacement of the diode or transistor, providing the proper operating voltages exist.

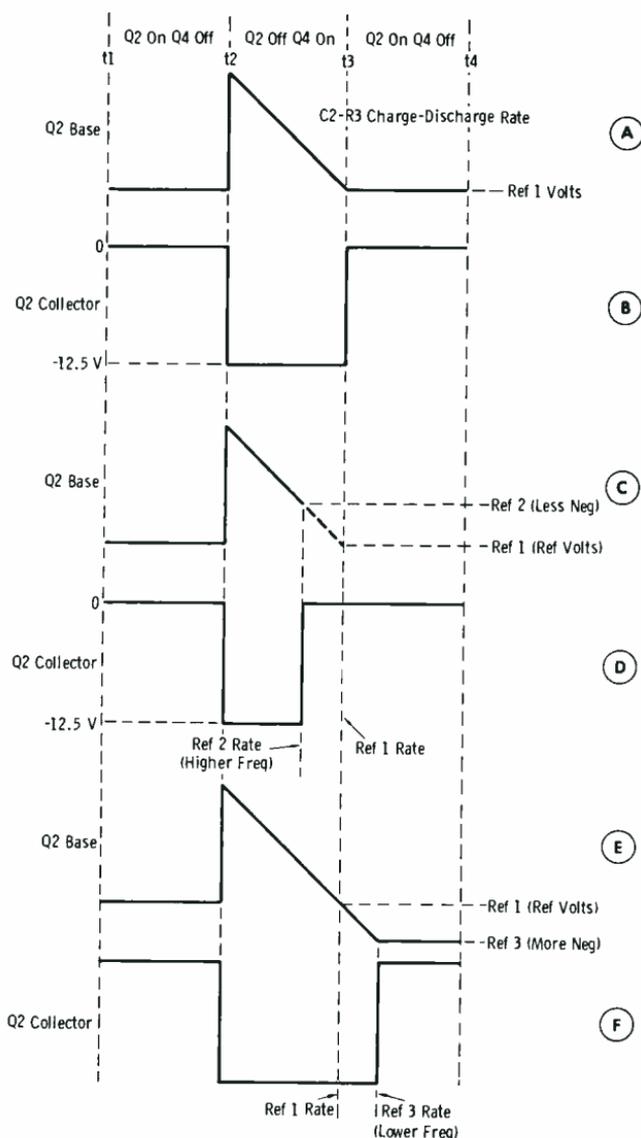


Fig. 4-3. Waveforms in voltage-controlled oscillator.

## 4-2. PHASE CORRECTION

Automatic frequency control (afc) and automatic phase control (apc) are both forms of *error detection* with an associated *corrective function*. The "error" must be related to some known standard.

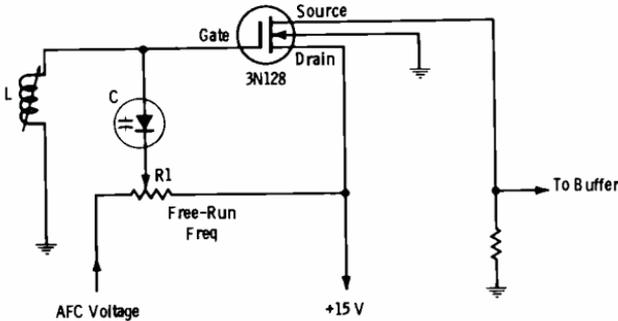
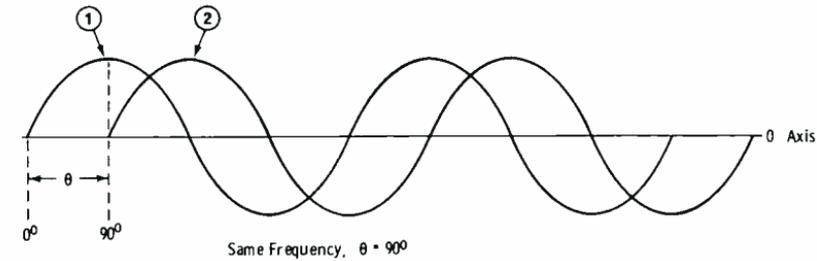


Fig. 4-4. Varicap type of vco.

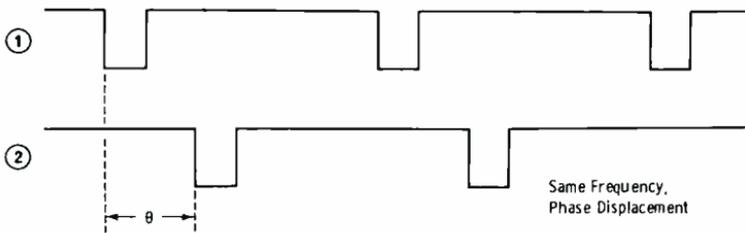
Fig. 4-5A shows two sine waves of identical frequency but with a  $90^\circ$  phase difference. If we take waveform 1 as the reference, waveform 2 *lags* waveform 1 by  $90^\circ$ . If we take waveform 2 as the reference, waveform 1 *leads* waveform 2 by  $90^\circ$ . Either sine wave could be the reference or "standard," and the other controlled in frequency and phase. The proper phase relationship might be the  $90^\circ$  shown, zero, or some other phase difference.

Next, look at Fig. 4-5B. We have two pulses of identical frequency (same pulse period), but with a phase displacement. Now go on to Fig. 4-5C. In waveform 1 we have the reference, or standard, pulse period. This might be obtained from the local sync generator, a color-subcarrier pulse (or countdown from the color subcarrier), etc. We form a sawtooth from this standard pulse as in waveform 2. Waveform 3 is a pulse from an oscillator to be controlled. It may or may not have the correct frequency at this time; we know from the drawing that it definitely is not in phase with the reference.

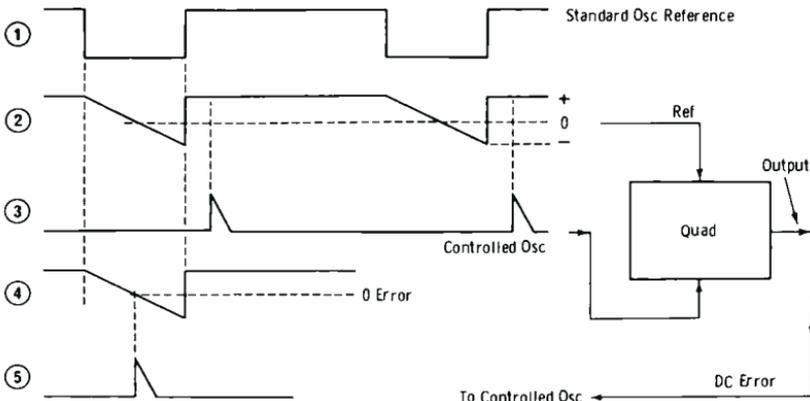
We feed waveforms 2 and 3 into an error detector such as a diode *quad* (to be described shortly). The quad conducts for the duration of the pulse in waveform 3, and at the moment we can see the resultant dc output will be a positive voltage. (The quad is sampling at a time when waveform 2 is maximum positive.) If this error voltage is fed back to the controlled oscillator, it should serve to change the oscillation *frequency* in the proper direction until the sampling occurs near the center of the reference sawtooth (essentially zero error voltage). This condition is shown by waveforms 4 and 5. Thus the controlled oscillator is locked in frequency *and* phase. The proper system phasing is obtained by the position of the reference pulse; this could be at the leading edge of sync, trailing edge of sync, a certain position within vertical sync, etc. This entire process is termed the *sampling pulse technique*.



(A) Phase relation of sine waves.



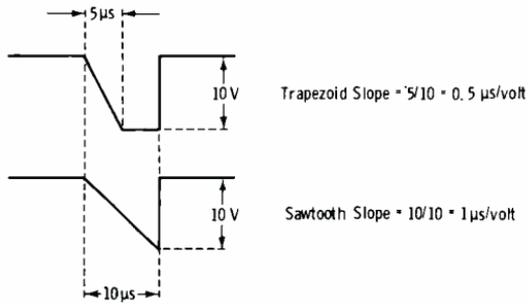
(B) Phase relation of pulses.



(C) Error detection and correction.

Fig. 4-5. Fundamentals of error detection.

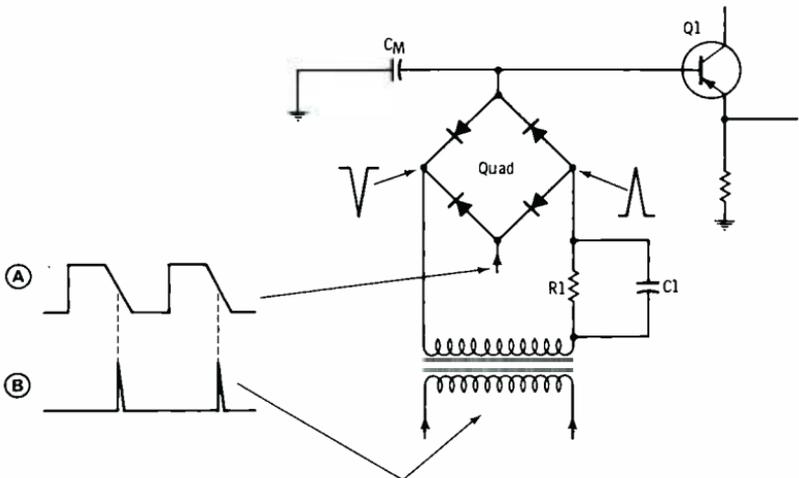
In Fig. 4-6, we see how to measure loop gain, and the difference between a sawtooth and trapezoid derived from the same reference pulse period. In the example given, we have two nonvariables; the pulse duration is 10 microseconds, and the amplitude is 10 volts. The trapezoid slope (hence loop gain) is twice as great as for the sawtooth. If the level change is completed in 5 microseconds, the slope is  $0.5 \mu\text{s/volt}$ ; that is, it takes just one-half microsecond to change one volt, or five microseconds to change 10 volts. The slope



**Fig. 4-6. Sawtooth and trapezoid of same duration.**

of the sawtooth is more gradual (lower loop gain). This waveform requires 10 microseconds to change 10 volts; thus the sensitivity is  $1 \mu\text{s/volt}$ . We will find some systems use first the lower loop gain (sawtooth) to establish frequency control, then the trapezoid to exert a tighter phase control after the frequency has been stabilized. In any case, the sampling pulse (derived from the controlled oscillator) tends to hold the oscillator frequency (and phase) at the point where sampling occurs at zero error voltage (center of slope).

The function of the sampling diode quad is shown in Fig. 4-7. At the instant of the sampling pulse (waveform B), the diode bridge is driven into conduction, connecting the trapezoid (waveform A) to memory capacitor  $C_M$ . This capacitor charges (or discharges) to the dc potential existing at the time of the sample. Capacitor  $C_1$  charges during the pulse time, and the time constant of  $R_1$  and  $C_1$



**Fig. 4-7. Circuit of phase detector.**

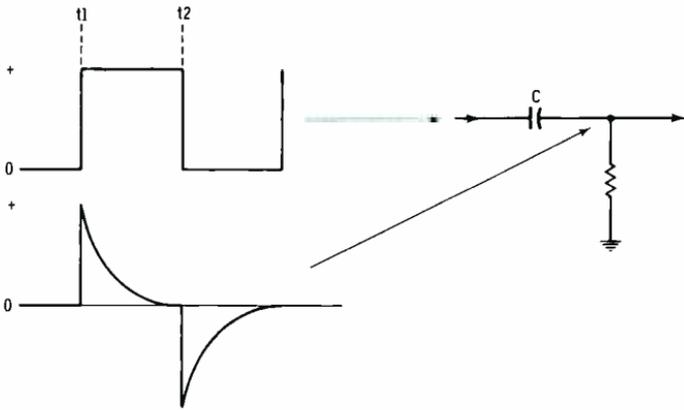
holds the quad nonconducting during the interval between sampling pulses. Capacitor  $C_M$  has no discharge path except through the relatively high impedance of Q1 and the diode bridge. Therefore it "remembers" the condition existing at the time of the previous pulse, and its charge will change only if the next pulse samples at a different dc level than existed for the preceding pulse. By dc coupling through Q1 back to the controlled oscillator (from which the sampling pulses are derived), the frequency-phase relationship is maintained.

The reader may now apply the above fundamentals to the vco of Fig. 4-2. Sampling occurs in the frequency-control module. In the free-run mode, the reference is a fixed dc voltage, and the free-run frequency control is R5 of Fig. 4-2. In the crystal mode, the reference is the countdown from the crystal oscillator. In the color mode, the reference is the countdown from the subcarrier oscillator, or the subcarrier burst from a remote signal in the color genlock mode of operation. In the ac-line mode, the reference is a pulse generated from the ac power-line frequency.

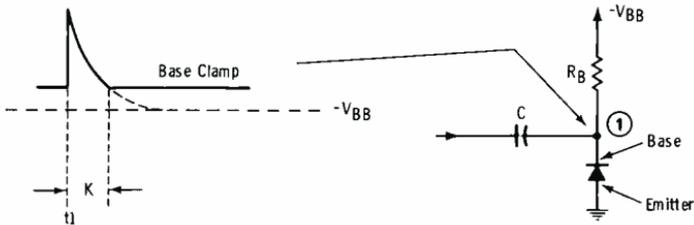
### 4-3. OUTPUT-PULSE WIDTH CONTROLS

Each of the five output pulse trains from the pulse-output module (Fig. 4-1) is generated by a bistable multivibrator. At one input of the multivibrator, triggers produce the leading edges of the output pulses, and at the other input, triggers produce the trailing edges. The leading-edge (le) and trailing edge (te) triggers are derived from the common vco so that precise relations are maintained between the pulses and during the vertical-sync interval.

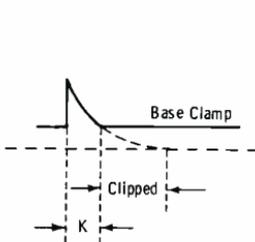
The delay mechanism for the timing relationships is, in many cases, a *boxcar* circuit. The maintenance technician must become thoroughly familiar with this circuit function. The first step is to analyze the content of Fig. 4-8. Fig. 4-8A can be used to review the conventional action of a short-time-constant RC circuit on a square-wave input. Prior to time  $t_1$ , there is no charge on the capacitor and no current in the resistor. At  $t_1$ , a positive-going voltage is applied in the form of a square wave. The capacitor begins to charge, but there is no voltage across the capacitor until it is charged. Therefore, at  $t_1$  the entire voltage must appear as a voltage drop across the resistor. As the capacitor becomes charged, the voltage appearing across it is subtracted from the applied voltage, and the voltage across the resistor is decreased. Since the time constant (RC product) is short relative to the input pulse duration, the capacitor charges quite rapidly, and the voltage across the resistor quickly falls to zero ( $t_2$ ). At this time, the negative-going pulse reverses the previous action, and the conventional differentiated



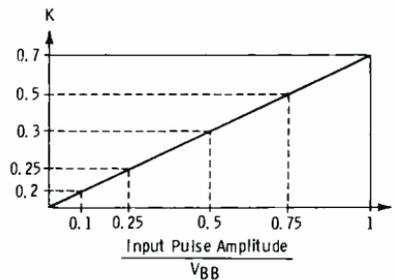
(A) Differentiating network.



(B) Equivalent of junction.



(C) Smaller pulse amplitude.



(D) Graph of K factor.

**Fig. 4-8. Action of a forward-biased base-emitter junction.**

waveform with twice the peak-to-peak value of the applied pulse results.

Now go to Fig. 4-8B. Here is a diode which simulates the base-emitter junction of a pnp transistor. It is forward biased by the current from  $V_{BB}$  through  $R_B$ . The time constant,  $CR_B$ , is assumed the same as in Fig. 4-8A. Prior to time  $t_1$ , point 1 is held near ground potential by the forward-biased diode junction. At time  $t_1$ , application of the positive pulse reverse biases (opens) the diode

junction, and the capacitor starts to charge toward negative potential  $-V_{BB}$  through  $R_B$ . The action continues only to the point at which the diode again becomes forward biased and the potential falls to the small forward voltage drop across the junction. Note that any further negative-going excursions of the waveform are clamped out by the forward-biased junction.

This action can be related to a very important *K factor*, which determines the width of the pulse at the transistor base. This *K* factor depends not only on the  $RC$  product (which determines the shape of the charge-discharge curve), but also on the amplitude of the input pulse. Observe Fig. 4-8C, where the input pulse is smaller in amplitude than in Fig. 4-8B. Now the base clamp occurs earlier, and the *K* factor becomes smaller in value. The width of the pulse at the base of the waveform is less than in Fig. 4-8B.

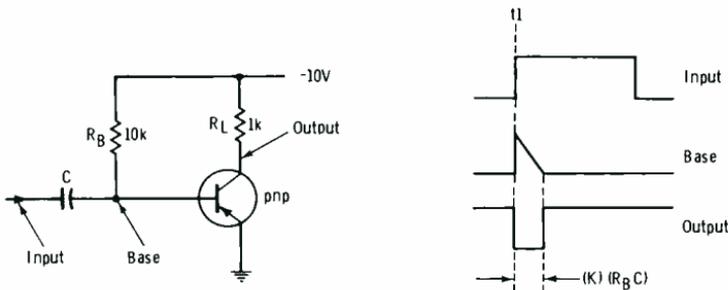
An approximate relationship between the *K* factor and the input-pulse amplitude is shown in the graph of Fig. 4-8D. If the input-pulse amplitude is equal to or greater than  $V_{BB}$ , a *K* factor of 0.7 results. If the ratio of the input-pulse amplitude to  $V_{BB}$  is unity (one), the base pulse width can be considered to be:

$$\text{Width} \approx (K)(RC) = 0.7(RC)$$

This relationship is only approximate, but it is useful.

Apply this principle to the circuit of Fig. 4-9A. First of all, a boxcar circuit can be recognized by the ratio of  $R_B$  to  $R_L$ . This is seldom more than a ten-to-one ratio. Since the *beta* of a transistor is always more than 10, the connection shown in Fig. 4-9A assures that the transistor is saturated with no input pulse because the base is returned to  $-10$  volts and the emitter is tied directly to ground. In this state, the collector is essentially at ground potential (closed switch).

At time  $t_1$  (Fig. 4-9B), a positive input pulse is applied, reverse biasing the base-emitter junction and cutting off the transistor (as-



(A) Diagram.

(B) Waveforms.

Fig. 4-9. Inverting circuit for narrowing positive pulses.

suming the pulse supplies sufficient cutoff current). This action sends the collector voltage to  $-10$  volts (switch open). The capacitor starts to charge through the base resistor toward  $-10$  volts, but the base voltage is clamped at the point where the input junction again becomes forward biased. The transistor is again saturated, and the collector voltage is again near zero. The width of the output pulse is approximately the  $K$  factor times the product  $CR_B$ .

Note that a boxcar is held in saturation under quiescent operating conditions, and that the coupling time constant ( $RC$ ) is usually made very short compared to the input-pulse duration.

Now go to Fig. 4-10. Here we have the same circuit and pnp transistor as in Fig. 4-9, but the input pulse is a *negative-going* pulse. At the base of the transistor, the pulse transferred by the coupling capacitor would normally consist of a negative-going excursion followed by a positive-going excursion. However, the low forward impedance of the junction provides a clamping action which prevents negative-going excursions. Since the transistor is already saturated, negative excursions have no effect. At the end of the input pulse, the positive-going excursion drives the transistor toward cutoff, and  $C$  begins to charge through  $R_B$  toward  $-10$  volts. Note that the output pulse occurs at the *trailing edge* of the input pulse. The width of the output pulse is  $(K)(R_B C)$ . Obviously, if  $R_B$  is made variable, the pulse width can be adjusted as desired.

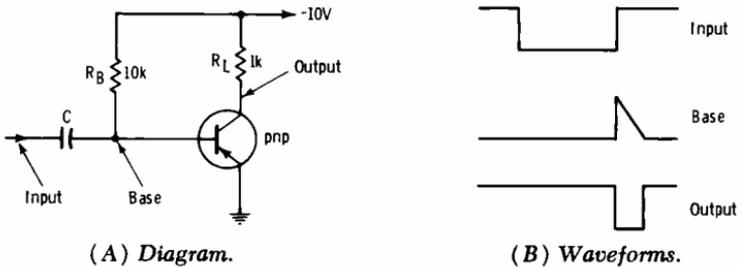


Fig. 4-10. Noninverting circuit for delaying negative pulses.

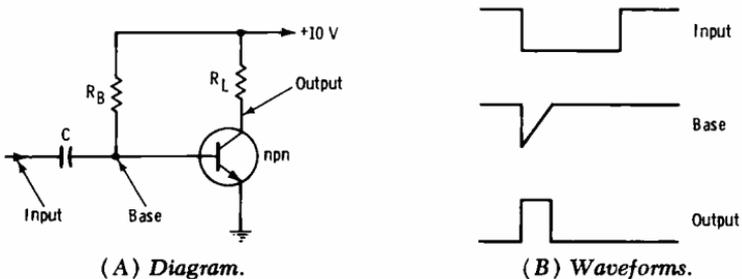
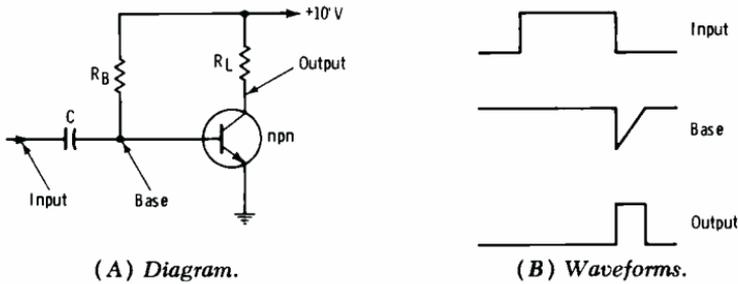


Fig. 4-11. Inverting circuit for narrowing negative pulses.

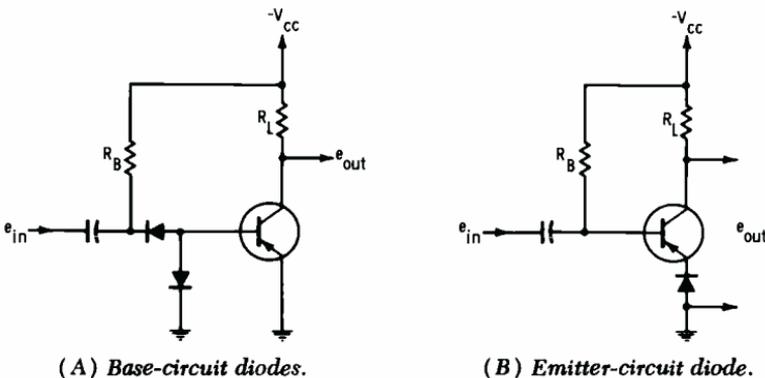


**Fig. 4-12. Noninverting circuit for delaying positive pulses.**

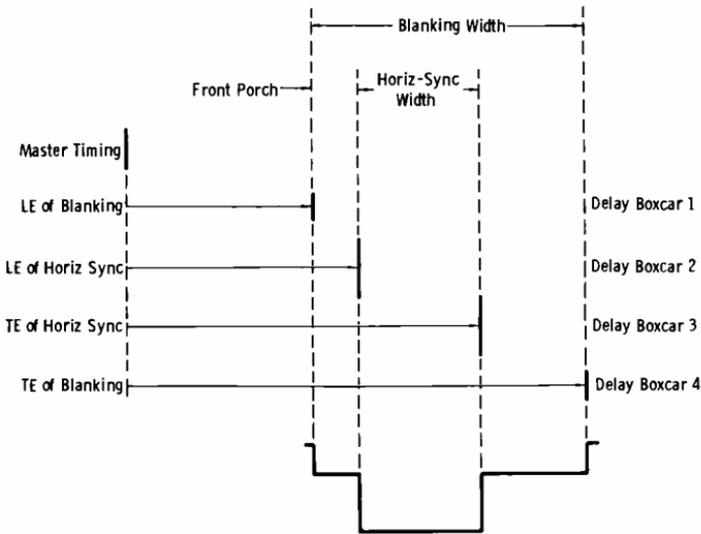
Using the foregoing principles, the reader should be able to analyze the circuitry of Figs. 4-11 and 4-12. When we know the type of transistor (pnp or npn) and the polarity of the input pulse, we know whether the circuit is intended as an undelayed pulse generator or a delayed pulse generator. Also bear in mind that the initial pulse narrowing is done by the differentiator circuit ( $R_B$  and  $C$ ) and that the resulting output-pulse width depends on this RC product as well as the K factor. The K factor is determined by the ratio of the input-pulse amplitude to the base supply voltage. To drive the transistor completely to cutoff, the pulse must have sufficient current to overcome the forward bias current.

Variations of the basic boxcar circuit are found. The variations are methods of immunizing the timing function from emitter-to-base breakdown effects in the transistor. Fig. 4-13A involves two diodes in the base circuit. The circuit of Fig. 4-13B employs a single diode in series with the emitter.

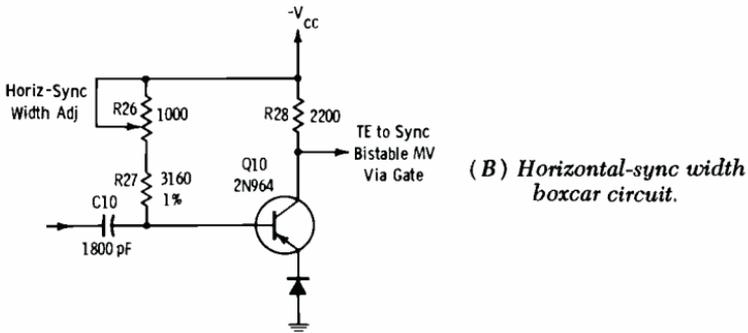
A simplified example of the use of boxcars in timing techniques is shown in Fig. 4-14A. Boxcar 1 receives the master timing pulse from the vco and generates a delayed trigger for the start of blanking. Boxcar 2 receives the output pulse from boxcar 1 and generates



**Fig. 4-13. Variations of basic boxcar circuit.**



(A) Relationship of horizontal sync and blanking.



(B) Horizontal-sync width boxcar circuit.

Fig. 4-14. Example of boxcar timing technique.

a delayed trigger which starts horizontal sync. Therefore, the width of the output pulse from boxcar 2 determines the front-porch width. Boxcar 3 receives the output from boxcar 2 and generates a delayed pulse for the trailing edge of sync; hence boxcar 3 determines horizontal-sync width. Boxcar 4 receives the output of boxcar 1 and generates a delayed trigger for the trailing edge of blanking. Hence the width of this pulse determines the horizontal-blanking width.

Fig. 4-14B illustrates the boxcar which sets the horizontal-sync width, termed boxcar 3 in Fig. 4-14A. With R26 set at midrange (500 ohms), the RC time constant is  $1800 \times 0.00366$  microseconds, and we expect a pulse width of  $(K)(RC) = (0.7)(0.00366)(1800)$

= 4.6  $\mu$ s. In practice, R26 is adjusted to obtain a horizontal-sync width of 4.76 (or 4.8)  $\mu$ s.

#### 4-4. SYNC-GENERATOR ADJUSTMENTS

Most of the more recent solid-state sync generators have no front-panel adjustment controls, since the stability of these units is extremely good compared to that of the former tube-type circuitry. Some controls, generally located on the circuit boards within the various modules, are provided. We will treat the entire subject of generator adjustments as if tube-type equipment were being employed, since the controls (when provided) in solid-state generators perform the same functions.

Proper adjustment of the station sync generator depends on the technician's familiarity with his oscilloscope, the trouble taken to maintain marker and time-base standards accurately, and the technique of adjustment. The reader should be thoroughly familiar with Chapter 1 of this book.

##### Frequency Adjustment

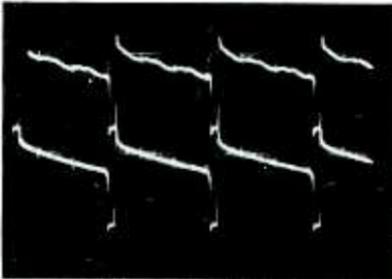
Proper operation of the counters and pulse-gating circuits will usually result if the master oscillator is within 5 percent of the 31.5-kHz nominal frequency. It is entirely practical to adjust this frequency within 1 percent or less, greatly extending the stability factor over a period of time with changing transistor and component characteristics.

Sync generators provide some means of controlling the master-oscillator frequency in any one of the following conditions:

1. Free-running for test or emergency use.
2. Crystal control. Usually used for test setups, but may also be used when the power-line frequency is unreliable, as in some remote locations, or when standby power sources, such as emergency power generators, are in use. (Monochrome only.)
3. Line lock to local 60-Hz power line. This is the most common operating condition for monochrome pictures.
4. Genlock. Uses stripped-off sync pulses from network or remote signals to control the local oscillator (and field counters), allowing superimposition of local signals over the network or remote signal.
5. "Color lock" for color programs. In this case, the color subcarrier generator is the source of the 3.579545-MHz subcarrier frequency, and a countdown from this source locks the 31.5-kHz sync-generator master oscillator (now slaved to the subcarrier countdown).

When a crystal circuit is employed, the crystal frequency is usually 94.5 kHz, or three times the master-oscillator (mo) frequency. The oscillator acts as a 3/1 divider when placed under crystal control. The crystal provides an accurate means of adjusting the mo frequency even though the remaining circuits may not be functioning. Therefore, it serves as a convenient servicing aid. The best way to adjust the free-running frequency of the mo when a crystal circuit is included can be outlined as follows:

*Step 1:* Place the low-capacitance scope probe at the 31.5-kHz test point. With the switch in the crystal position, the waveform should resemble the upper trace in Fig. 4-15. Since the crystal frequency is 94.5 kHz, division by 3, indicated by three sine-wave cycles, gives 31.5 kHz. If a different indication occurs, the mo frequency-adjustment control should be rotated for proper frequency.

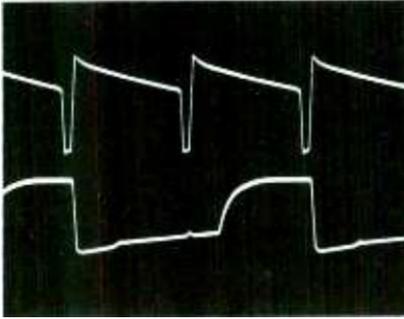


**Fig. 4-15. Waveforms at 31.5-kHz test point (upper, crystal control; lower, free-running).**

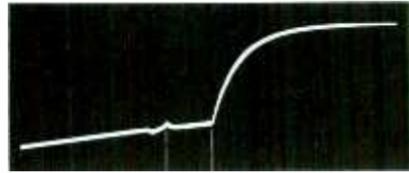
In the RCA Type TG-2A generator, a capacitor is used as a coarse adjustment, and a time-constant potentiometer is used as a fine (about 3 percent) adjustment. Use the capacitor for setting the exact frequency with the time-constant control in midposition. This same procedure is followed for similar types of mo circuitry in other generators.

The procedure in Step 1 assures the proper frequency under crystal control. Step 2 sets the free-running frequency to 31.5 kHz within close limits; this step assumes use of a triggered scope. Step 3 is an alternate method for use when triggered sweep is not available. Step 4 describes a vernier adjustment that can be used for precise setting, providing the counters are functioning properly.

*Step 2:* Expand the scope sweep to display one cycle at 31.5 kHz with the switch in the crystal position, and place the leading edge of the pulse on a vertical graticule line. Switch the mo control from the crystal to the off (free-running) condition; if the pulse shifts from the reference position, adjust the mo frequency until no shift occurs on a change between the crystal-control and free-running



**Fig. 4-16. Master-oscillator waveform (upper trace) and divide-by-2 waveform (lower trace).**



Measure this distance.

**Fig. 4-17. Lower waveform of Fig. 4-16 displayed with scope set for expanded sweep.**

modes. A waveform similar to the lower trace in Fig. 4-15 should be observed when the mo is free-running.

*Step 3:* Fig. 4-16 illustrates the basis for a method of setting the mo frequency when a scope with triggered sweep is not available. The upper trace in Fig. 4-16 is the waveform of the 31.5-kHz mo, and the lower trace is the output of the divide-by-two (15.75-kHz) circuit showing the characteristic division pip of the alternate pulses. Expand the waveform in the lower trace as in Fig. 4-17 so that the distance between the tip of this pip and the steep side of the pulse immediately to the right occupies a given number of graticule lines. If this width in the free-running position is different from the width with crystal control, adjust the mo frequency so that no change occurs between the crystal and free-running modes.

*Step 4:* With the mo in the free-running mode, observe any 60-Hz signal (such as at the counter output, vertical drive, or blanking) with the scope trigger selector in the line position. This reveals any slip with respect to the line frequency. Adjust the mo frequency so that the trace becomes stable. A slight drift back and forth may occur, but no sudden changes should exist. The mo frequency is now within a small fraction of 1 percent of the desired value, and placing the oscillator in line lock control should immediately stabilize the trace. If this does not occur, or if the trace becomes noticeably unstable on-line lock, the afc circuitry needs service.

Normally, Step 4 is the only step necessary for routine checks of the mo frequency. However, the previous steps should be carried out occasionally to check the crystal control, and, of course, they are necessary if trouble is experienced in the counter stages.

Some generators that do not include a crystal for test purposes have a 94.5-kHz ringing circuit in the master oscillator to sharpen wavefronts and provide a reasonable degree of short-term stability.

(Long-term stability is provided by the afc circuit.) Step 4 normally is used to adjust the frequency of this type of mo, but any indication of faulty 60-Hz pulses always poses the question of whether the trouble is in the counters or the master oscillator. By use of the low-capacitance scope probe, the waveform across the ringing circuit may be observed. The mo frequency should be adjusted so that the firing pulse occurs at approximately the 45° point of the third sine-wave cycle.

An alternate method is the old standby of employing a 31.5-kHz frequency standard and some form of comparison. The beat method is subject to error in pulse circuits because of the high harmonic content. The frequency-standard signal may be displayed on the scope, with the scope triggered externally by the same signal. Without the external trigger, observe the mo output, and adjust the circuit for the required frequency and maximum stability of trace.

Whenever a sync generator becomes erratic in operation, the first step should be to change the mo control from the line-lock position (in monochrome) or the color-lock position (in color) to the free-run or crystal-control position. If the pulses stabilize, the mo frequency may be out of range of the afc, or the afc circuitry may be operating erratically.

NOTE: Setting the color subcarrier frequency, as well as a final check for color lock, was covered in Chapter 3.

### Checking the Countdown

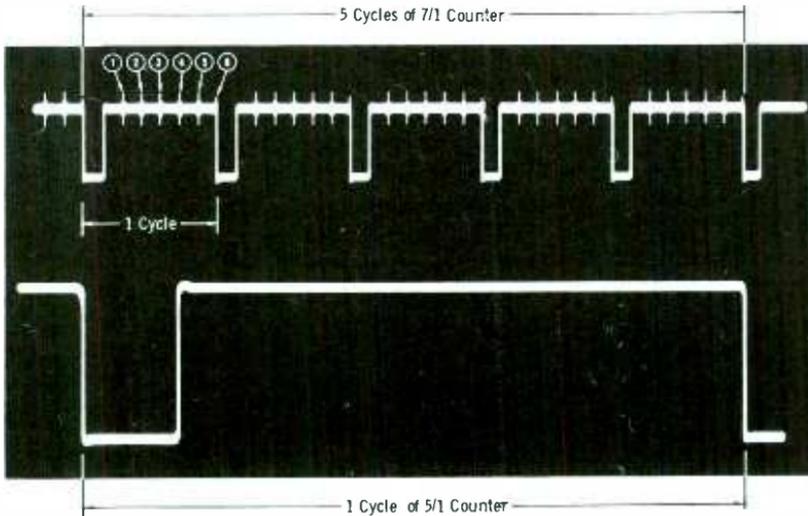
Some of the older-type sync generators employing step-by-step counters featured continuous display of counter-circuit action by means of a small scope on the front panel. In more recent generators, this feature has been omitted in the interest of overall simplification. Several types use binary counters with plug-in modules or replaceable strips which are not serviced by station personnel. In the RCA Type TC-2A generator, the first counter (7 to 1) contains a variable adjustment, whereas the remaining counters use fixed constants with no adjustable controls. Later solid-state sync generators have no adjustable controls at all in the counter circuits. On all generators, test points generally are provided at the various counter outputs for the purpose of testing and servicing.

Because of the nature of binary counters that employ feedback pulses to obtain the odd-number countdown (525 to 1), a direct counting check usually is not practical. Trouble in the preceding binary stage is indicated at the first check point where random or unstable pulses exist.

Generators using multivibrator-type counters normally employ four stages as summarized in Table 4-1. The total division is thus

$7 \times 5 \times 5 \times 3 = 525$ . The most convenient method of checking such a counter chain is as follows:

1. Observe the waveform of the first divider circuit by adjusting the sweep to obtain 1 cycle of the waveform shown in the upper trace of Fig. 4-18. The counter should show a division of 7.



**Fig. 4-18.** Five cycles of 7-to-1 counter (upper waveform) and one cycle of 5-to-1 counter (lower waveform).

2. Apply the same signal to the external-sync input of the scope, and adjust the sweep so that five complete cycles are displayed on the screen.
3. Without changing the external sync connection or the sweep, use the scope vertical-amplifier probe to sample the output of the following (5 to 1) counter. Exactly one cycle should now occur in the same interval, as shown by the lower trace in Fig. 4-18.

**Table 4-1.** Sync-Generator Counters

Stage Number	Input Frequency	Division	Output Frequency
1	31,500	7	4500
2	4500	5	900
3	900	5	180
4	180	3	60

Repeat the same procedure for the remaining stages by triggering the sweep from the counter *preceding* the one to be checked and adjusting the scope sweep for the number of cycles that equals the division of the counter to be checked.

An alternate, but less accurate, method is to measure the frequency of the pulses by using the time base of the scope (see Table 4-2). In this case, the engineer must assure himself that the time base is both accurate and linear. This can be assumed to be true for most recent scopes such as the Tektronix Type 545, 547, or 7633. The latter scope includes a digital readout for precise duration measurements.

**Table 4-2. Time Base of Counter Circuits**

Stage	Total Division	Freq	Width of Pulse* in Terms of H	Width of Pulse* in Microseconds
MO	1	31,500	0.5	31.75
÷ 7	7	4500	3.5	222
÷ 5	35	900	17.5	1111
÷ 5	175	180	87.5	5556
÷ 3	525	60	262.5	16,668

\* From leading edge of pulse to leading edge of following pulse.

### Setting Pulse Widths

Pulse widths are set by either of two general methods: (1) variable time constants of triggered multivibrators, or (2) multivibrators (driven type) which receive on and off triggers from adjustable delay lines. The methods of measuring pulse widths are applicable to any type of generator, but the adjustment procedures differ radically among various manufacturers. It is usually important to follow the sequence of adjustments recommended in the instruction manual for the particular generator involved.

**NOTE:** Some digital-type sync generators do not employ adjustments for pulse width. Digital counting and logic circuitry makes both leading and trailing edges of all output waveforms direct responses to clock-generated transitions. Pulse widths are fixed according to industry standards and fall at or near the center range of limits specified by the NTSC, EIA, and FCC, where such standards apply. The technician, of course, must measure pulse widths periodically to assure compliance with the rules. Digital troubleshooting is covered in a companion volume of this series, *Digitals in Broadcasting*.

For convenience, Table 4-3 lists pulse widths in terms of H (as given by the FCC) and in microseconds. The latter form is most convenient for the technician, since he measures microseconds when using the scope. Fig. 4-19 presents a more detailed analysis of the horizontal pulses, including relative amplitudes. Dimension P represents the peak excursion of the luminance signal from the blanking level, but it does not include the chrominance signal. Dimension S is the reference sync level. For a composite video signal with a 1-volt

Table 4-3. Pulse Table

Pulse	Minimum		Nominal		Maximum	
	H	Micro-seconds	H	Micro-seconds	H	Micro-seconds
Horiz Blanking	0.165	10.5 <sup>5</sup>	0.175	11.1 <sup>5</sup>	0.178	11.3 <sup>5</sup>
Horiz Sync	0.07	4.445	0.075	4.8	0.08	5.08
Equalizing	0.45 of Horiz Sync	2.0	0.5 of Horiz Sync	2.4	0.04	2.54
Vert Serration	0.06	3.8	0.07	4.5	0.08	5.08
Vert Blanking	18.375	1167 <sup>5</sup>	19.7	1250 <sup>5</sup>	21	1333 <sup>5</sup>

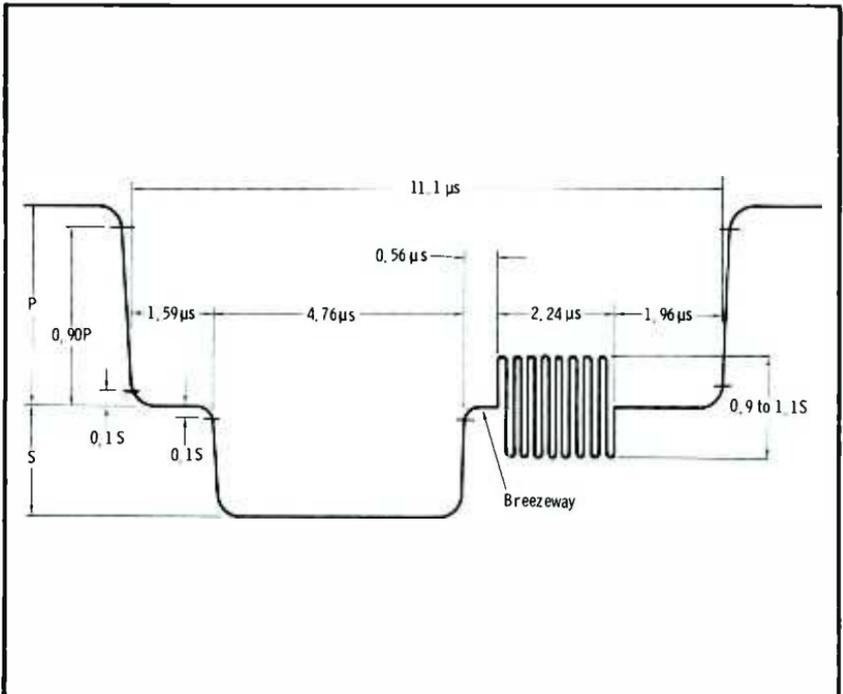
## Notes:

- EIA resolution and ball charts are based on the following values:  
 Horiz Blanking = 11.1 Microseconds or 11.3 Microseconds  
 Vert Blanking = 1250 Microseconds or 1333 Microseconds
- The width of the vertical-sync pulse is set by the tolerance on the width of the vertical serration. An interval of 63.5  $\mu$ s (H) must exist from the leading edge of the last (leading) equalizing pulse to the trailing edge of the first serration.
- The width of the equalizing pulse must be 0.45 to 0.5 of the horizontal-sync pulse width used.
- The duration of vertical blanking is from 18.375 to 21 lines. Although 21 lines (21H) is shown as maximum in the above chart, this is the width of vertical blanking maintained by the networks. This allows vertical-interval test signals to be inserted during vertical blanking with a suitable "guard band" of blanking lines before the start of active line scan.
- Horizontal and vertical blanking must be in the proper ratio to establish the 4:3 aspect ratio.

peak-to-peak amplitude, the video amplitude is 0.714 volt (0 to 100 IEEE units), and the sync amplitude is 0.286 volt (0 to -40 IEEE units).

Fig. 4-20 illustrates the FCC-designated points of measurement on the IEEE scale. With blanking positioned on the 0 IEEE line, these are at +4 and -4 IEEE units. Fig. 4-20 also shows the typical values for setting pulse widths. The interval designated SCH (sub-carrier-horizontal phase) is a recommended standard of adjustment to minimize some program nonuniformity on the receiver. The same standard recommends that nine complete cycles of burst be used. (Reference: EIA TV Systems Bulletin No. 4)

By cross-mixing sync and blanking outputs and adjusting the



	Nominal Microseconds	Tolerance Microseconds
Blanking	11.1	+0.2 -0.6
Sync	4.76	+0.32 -0.315
Front Porch	1.59	+0.13 -0.32
Back Porch	4.76	+0.96 -0.61
Breezeway	0.56	+0.08 -0.17
Burst	2.24	+0.37 -0
Blanking to Burst <sup>†</sup>	6.91	+0.08 -0.17
Sync & Burst	7.56	+0.38 -0.49
Sync & Back Porch	9.54	±0.32

<sup>†</sup> Blanking-to-burst tolerances apply only to signal before addition of sync.

**Fig. 4-19. Details of horizontal pulses. (Drawing shows median values.)**

scope time base to obtain the waveform shown in Fig. 4-21, horizontal-sync, equalizing, horizontal-blanking, front porch, and vertical-serration widths may be adjusted from one display. The markers are  $0.005H$  (0.5% of  $H$ ) for maximum accuracy. (Such markers are provided on the Tektronix Type 524 scope.) Table 4-4 lists the proper number of  $0.005H$  markers for the appropriate intervals. On more recent scopes, such as the Tektronix Type 545, 547, or 7633, the time base is sufficiently accurate to use the graticule lines.

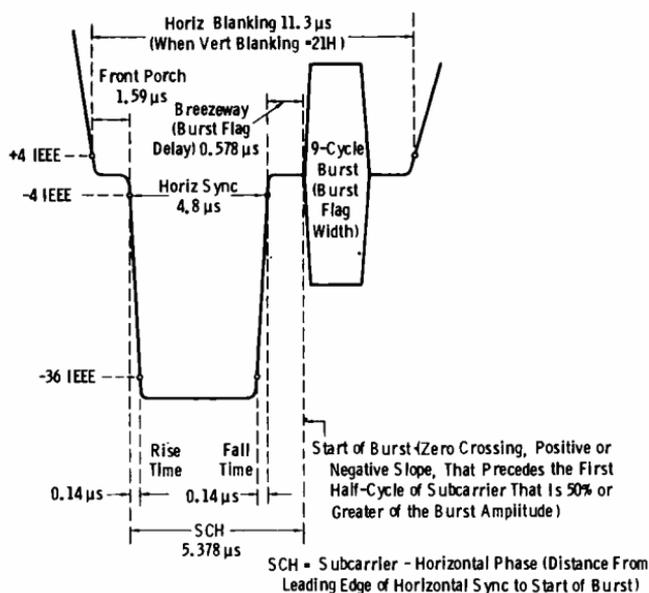
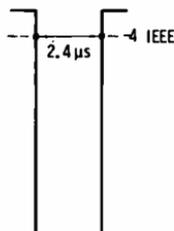
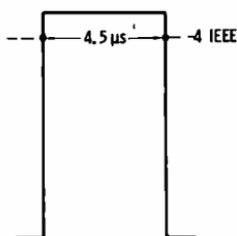
According to Table 4-3, horizontal sync should be adjusted to 4.8 microseconds, or  $0.75H$ . For this width, 15 of the  $0.005H$  markers are present, as shown in Fig. 4-21. The 1-microsecond markers could be used here if desired, as well as for setting the total width of horizontal blanking to 11.1 microseconds; slightly less accuracy would be obtained, however. The width of the vertical-sync serration should be 4.5 microseconds, or  $0.07H$ , which is indicated by 14 of the  $0.005H$  markers. When the time base and sweep linearity are known to be accurate, the graticule lines may be used alone. For example, the vertical serration could be set by adjusting the scope for 2.25 microseconds/centimeter and adjusting the serration to cover exactly 2 centimeters.

The cross-mixing of sync and blanking is not required when a modern dual-trace scope incorporating digital readout is employed. For example, composite blanking can be fed to channel A and composite sync to channel B, with the sweep time base adjusted to a horizontal rate that gives the display indicated by Fig. 4-22. By adjusting the brightening-pulse markers to the desired points on the waveforms, the front porch ( $1.59 \mu\text{s}$  when properly adjusted) is read out on the LED display.

Vertical blanking can be set accurately by using horizontal-drive pulses as markers. When a 1000-ohm resistor is placed from the horizontal-drive test point to the blanking test point, the horizontal signal is attenuated to a convenient marker amplitude superimposed on blanking viewed at the vertical rate on the scope. From Table 4-3, the maximum duration of vertical blanking is  $21H$ . This duration is most generally used in order to allow for VIT signals and signals

Table 4-4.  $0.005H$  Table

Pulse	Minimum	Nominal	Maximum
Horiz Blanking	33	35	35.6
Horiz Sync	14	15	16
Equalizing	6.3	7.5	8
Vert Serration	12	14	16

(A) *Horizontal pulse.*(B) *Equalizing pulse.*(C) *Vertical serration.*

**Fig. 4-20. FCC points of measurement on IEEU scale; blanking at 0 IEEU units, 140 IEEU units = 1 volt.**

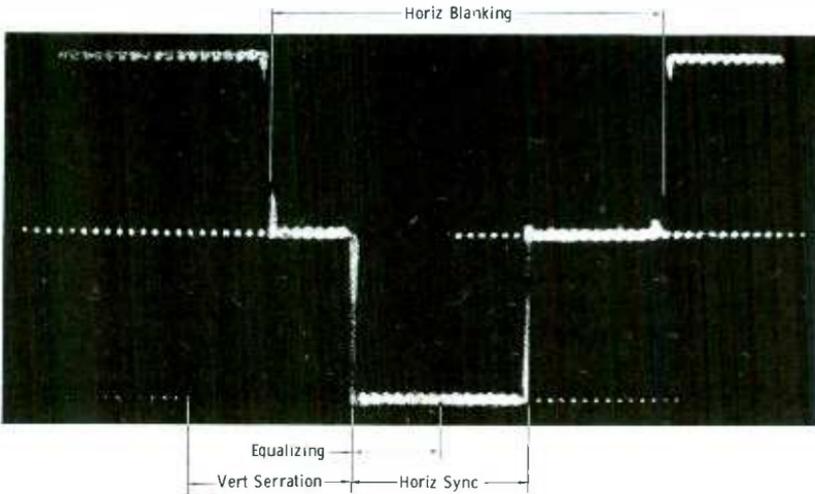


Fig. 4-21. Cross-mixed sync and blanking with 0.005H markers.

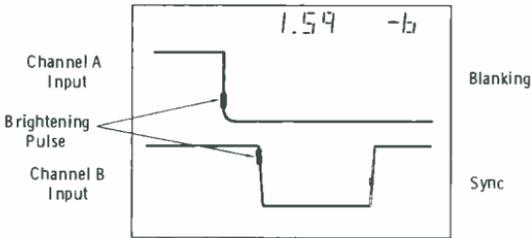


Fig. 4-22. LED display of 1.59 microsecond on digital-readout scope.

required for remote transmitter control. Thus 21 markers indicate the proper width. Note also from Table 4-3 that if a vertical-blanking interval of 21 lines is used, horizontal blanking must be  $11.3 \mu\text{s}$  to retain the proper 4:3 aspect ratio.

The use of markers is not necessary if the scope includes variable delayed sweep. Simply leave sync and blanking cross-mixed (or observe a composite video signal), and select field 1 for display. This field has a full line (H) preceding the first equalizing pulse, and a  $\frac{1}{2}H$  spacing from the last equalizing pulse to the first horizontal-sync pulse of field 2. Note from Fig. 4-23 that the leading edge of this sync pulse is the end of the first 9H interval of vertical blanking, plus the  $0.025H$  vertical front porch (this is  $1.5875 \mu\text{s}$  of front porch). The remaining horizontal-sync pulses may then be counted to arrive at the total vertical-blanking interval.

For color signals, there are additional pulse-width controls for the following:

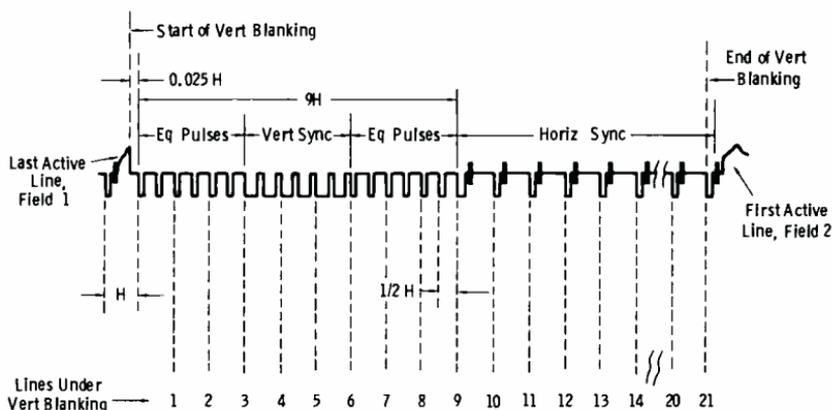


Fig. 4-23. Use of field 1 to set vertical-blanking width.

1. Burst-flag delay control for proper breezeway
2. Burst-flag width control for 8 to 9 cycles of color burst
3. Burst-eliminate control to delete color bursts during the 9H vertical-sync region

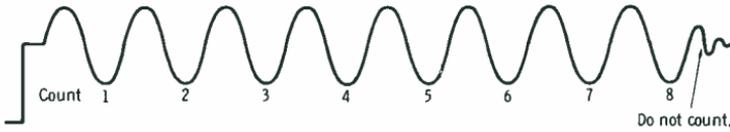
### Adjustment of Burst Key (Flag)

An encoded signal may be used in the adjustment of the burst-key (burst-flag) generator according to the following procedure:

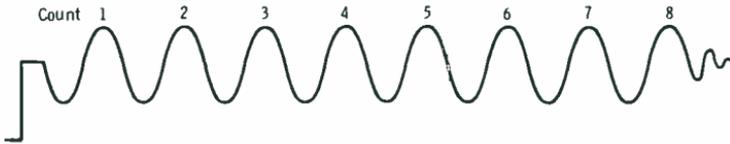
1. Observe the composite color signal (sync and blanking added) at some point after encoding. Trigger the scope externally with horizontal drive to obtain a steady trace.
2. Adjust the burst-key delay control for proper breezeway (Fig. 4-19).
3. Adjust the burst-key width control for 8 or 9 *complete* cycles. Do not count cycles of less than 50 percent of the nominal peak-to-peak value. If the first cycle starts with the positive-going alternation, count the negative peaks (Fig. 4-24A). If the first cycle starts with the negative-going alternation, count the positive peaks (Fig. 4-24B).

NOTE: If you use straightforward internal triggering of the scope instead of external trigger, you can obtain the "interlaced" burst pattern of Fig. 4-25. You should count 16 to 18 peaks for 8 or 9 cycles of burst.

4. Trigger the scope externally with vertical drive, and use the delayed sweep on the scope while observing the vertical interval for one field. The time base should be such as to permit observation of the 9H interval plus a few horizontal-sync pulses



(A) Peak amplitudes.



(B) Polarity of peaks.

Fig. 4-24. Counting of cycles in burst.

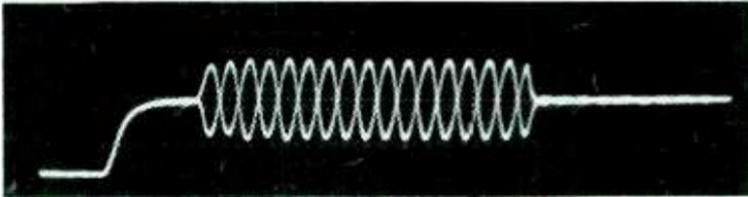


Fig. 4-25. Interlaced color burst.

following the trailing equalizing pulses. Adjust the burst-eliminate control to eliminate all bursts in the 9H interval (Fig. 4-26). Operate the field-shift key on the scope and observe the alternate field. If necessary, readjust the burst-

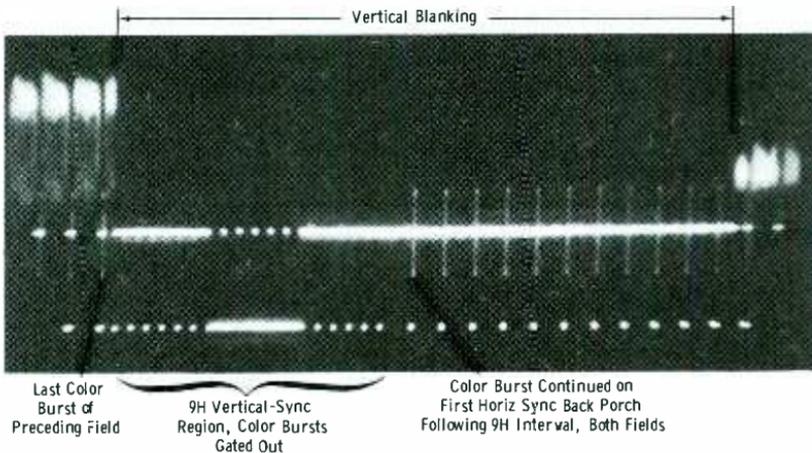


Fig. 4-26. Vertical-blanking interval of composite color signal.

eliminate control to obtain the 9H "key out" on this field. Go back and forth between the two fields and adjust until the 9H elimination is correct for both fields.

### Color Genlock

There are two general applications of genlocking, (1) the slaving of the local sync generator to a network or remote signal source, and (2) locking two or more cameras in the field to one another and to a common reference for mixing purposes. The second case applies only when each mobile camera has an internal sync generator; obviously it does not apply to field setups employing a common sync generator to drive all cameras.

Genlocking the studio sync generator (application 1 above) is no longer necessary when the station employs an automatic field or frame store system working on digital principles. These are covered in a companion volume, *Digitals in Broadcasting*.

Fig. 4-27 shows a basic genlock arrangement that fits into Fig. 4-1.

Checking for troubles in color genlocking requires a good familiarity with the circuitry involved in the particular system. In practice, circuits vary greatly in design; the best we can do is to acquaint you with the principles involved so that you can more readily understand the equipment instruction books.

The basic operation of normal *monochrome* genlock may be outlined as follows:

1. The frequency and phase of locally generated horizontal- and vertical-sync pulses are compared with those of corresponding pulses received from a network or remote pickup.

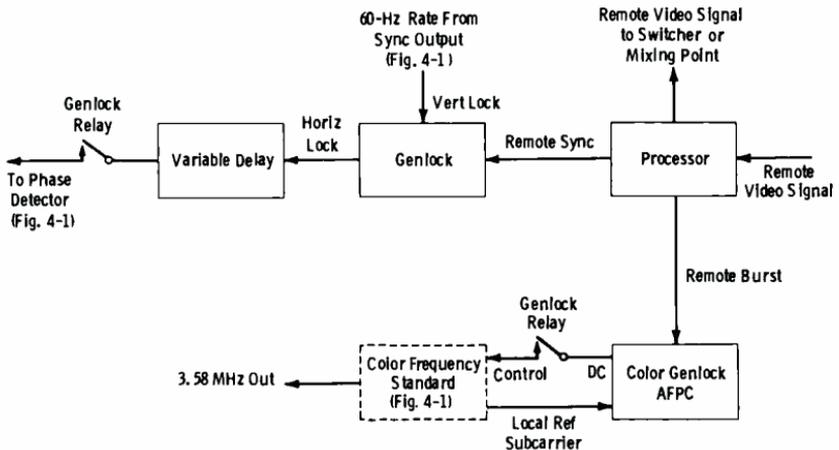


Fig. 4-27. Basic genlock arrangement.

2. From this comparison are derived correcting signals which reflect the frequency and phase difference between the local and remote signals.
3. The correction signals are applied to the appropriate circuits in the local sync generator so that its output pulses are exactly in phase with the incoming pulses from the remote source.

Now assume that the remote signal is a composite color signal, and it is desirable to lap-dissolve or superimpose a local color signal during the program. Since the local sync generator (in genlock) is already locked in frequency and phase to remote sync, the task remaining is to lock the local color-subcarrier oscillator to the exact frequency and phase of the remote subcarrier burst.

The color-genlock circuitry extracts the burst from the remote signal and, through a reference phasing network to obtain proper phasing with local color vectors, forms a sawtooth for use in a phase-detector circuit such as that in Fig. 4-28. In this particular type of system, a sampling pulse is formed from the local subcarrier burst. This pulse is applied to the base of Q1. Diode D1 suppresses any negative-going signal that might appear at the collector of Q1 and prevents ringing caused by inductive kickback. Thus, only a positive-going sample pulse appears at the collector of Q1 to be coupled by T1 to the error-detection circuitry of Q2 and Q3.

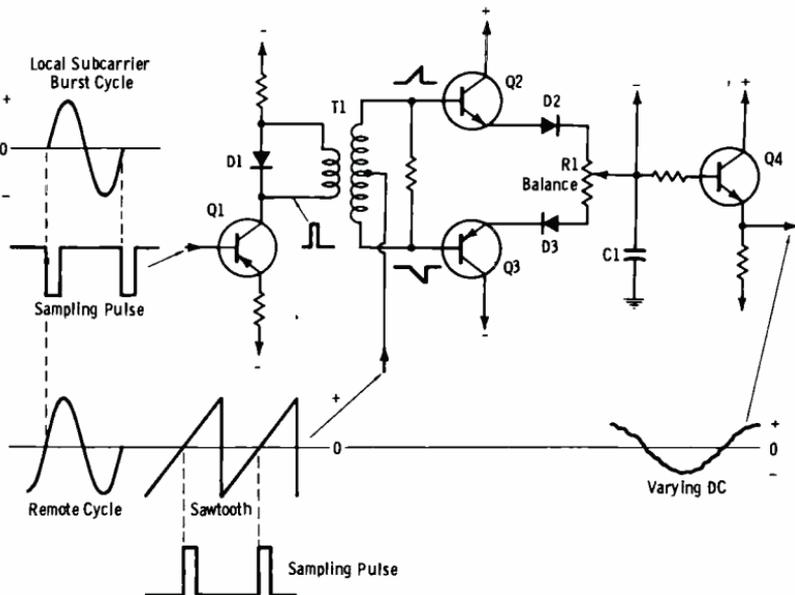


Fig. 4-28. Basic circuit for error detection.

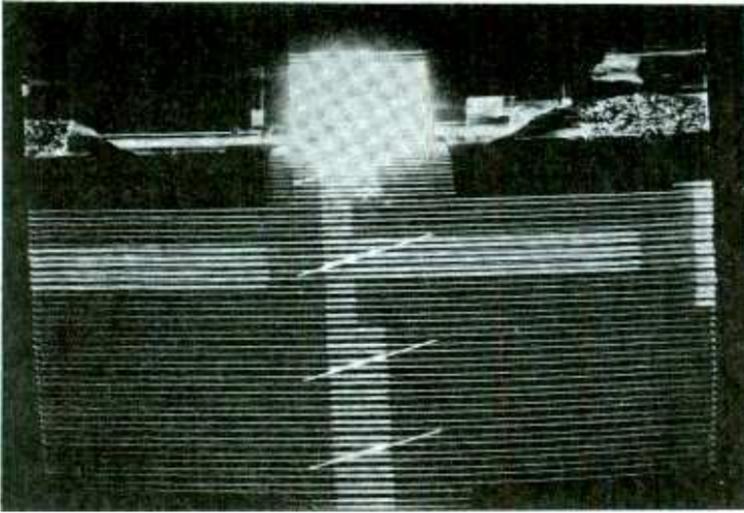
The sawtooth formed from reference burst appears at the center tap of the secondary of T1. This transformer splits the phase of the sample pulse so that pulses of opposite polarity are fed to the base connections of Q2 and Q3. These transistors are normally cut off. When the sample pulses occur, the transistors are driven into conduction by an amount depending on the position (determined by frequency or phase of the local subcarrier oscillator) of the sampling pulse on the slope of the reference sawtooth waveform. The balance control (R1) permits balancing the zero potential at the center of the saw slope. The detected error causes variations in the amplitude and polarity of the charge on C1, and these variations are coupled through Q4 to the frequency-controlling element of the local subcarrier oscillator. This is the conventional pulse-sampling technique that is commonly found in automatic-frequency-control circuitry.

To check a sampling-error circuit, use a dual-trace scope with one input at the Q1 collector (sample pulse) and the other input at the sawtooth connection to the center tap of T1. (Test points are often provided at such connections.) Check first for the presence of both pulses. If one or the other is missing, check back through that particular path. Proper timing is evident if the sample pulse occurs near the center of the sawtooth slope.

#### 4-5. THE PULSE-CROSS MONITOR

Two basic types of pulse-cross monitors are used, the interlaced and the noninterlaced. In both types, the horizontal sweep is delayed sufficiently so that the horizontal-blanking and sync pulses appear near the center of the raster. To display the entire frame (interlaced presentation), it is also necessary that the vertical sweep be delayed to place vertical blanking and sync in the normal active line area of the monitor. Both fields (interlaced) are displayed so that the entire 37- to 42-line vertical-blanking interval is visible. If the monitor vertical-deflection rate is changed to half rate (30 Hz), a single field is displayed with half the number of pulses (non-interlaced presentation). In either case, expansion of the vertical sweep normally is used to allow more critical observation of the pulses.

Fig. 4-29 shows the pulse-cross presentation of a line-output signal on an interlaced monitor. In this case, the video polarity is inverted so that sync is in the white-going direction. Note the convenience as a quick-reference check for the widths of horizontal front porch, sync, and blanking; equalizing pulses; and vertical sync. Vertical blanking is checked conveniently by counting the number of blanking lines. Some stations construct graticules with



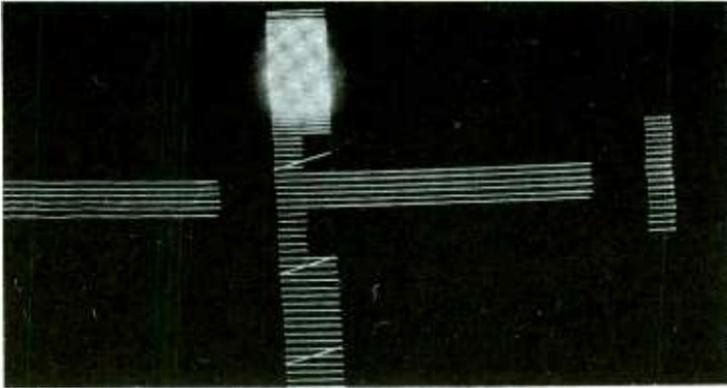
**Fig. 4-29. Interlaced pulse-cross display (composite signal).**

normal pulse widths marked after an accurate check of the generator with an oscilloscope.

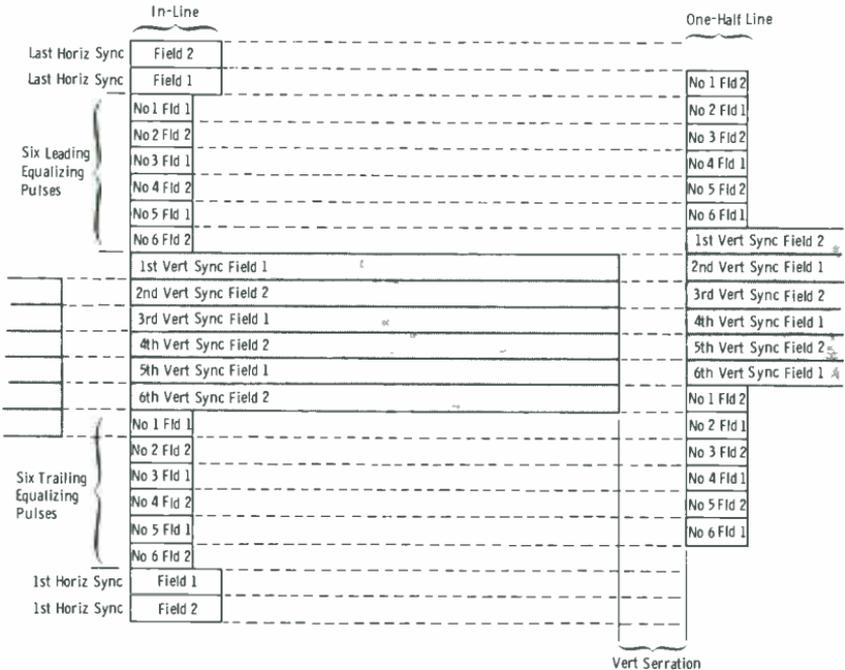
The pulse-cross monitor is extremely useful both as a continuous monitor and as a servicing tool. A  $9 \times 1$  switch panel is used at station WTAE-TV to allow selection of signals from a number of points to feed the monitor, but this switch panel is normally left in the standby-generator position. This permits continuous monitoring of whichever generator is in the standby position (composite sync only), as shown in Fig. 4-30A. Fig. 4-30B represents an expanded view of this display with identification of the pulses.

The "cross" is formed at the position in line with horizontal sync. The reader can readily understand the sequential presentation of the monitor (Fig. 4-30B) if he will mentally move field 2 (Fig. 4-31) to the left one-half line so that the horizontal-sync pulses of both fields are in vertical alignment. Now, observing Fig. 4-30B, note that in-line pulses (those occurring at H intervals) are equalizing pulses 1, 3, and 5 of field 1 and 2, 4, and 6 of field 2, spaced on alternate lines of the display because of interlace of fields. The half-line intervals and the remainder of the presentation should be obvious from following a similar analysis.

With an interlaced type of pulse-cross monitor, loss of interlace, such as could be caused by a vertical-countdown error of  $0.5H$ , is readily apparent, as illustrated by Fig. 4-32. The brightness of the display is greater than normal for a given adjustment because of the double tracing of identical raster lines.



(A) Display on monitor.



(B) Pulse identification.

Fig. 4-30. Pulse-cross display, sync only.

Some master monitors, line monitors, and tape-system monitors provide a single-field (noninterlaced) pulse-cross presentation. Fig. 4-33 shows how to interpret this display. Equalizing pulses 1, 3, and 5 of field 1 occur at horizontal-sync intervals, and equalizing pulses 2, 4, and 6 occur at intermediate intervals. Similarly, vertical-sync

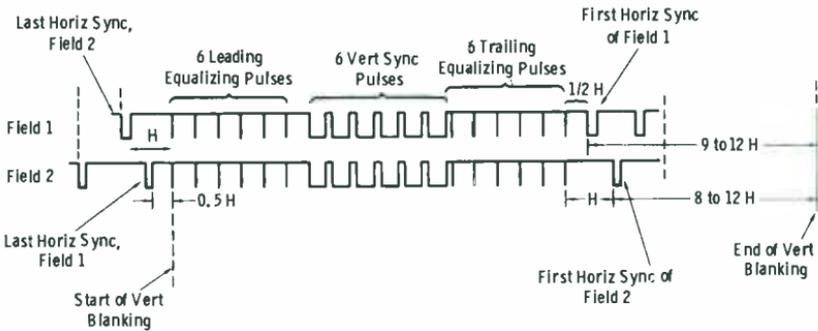


Fig. 4-31. Pulses for pulse-cross display.

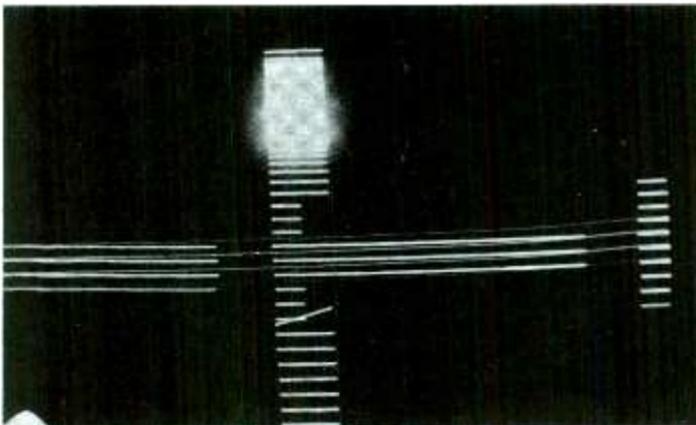
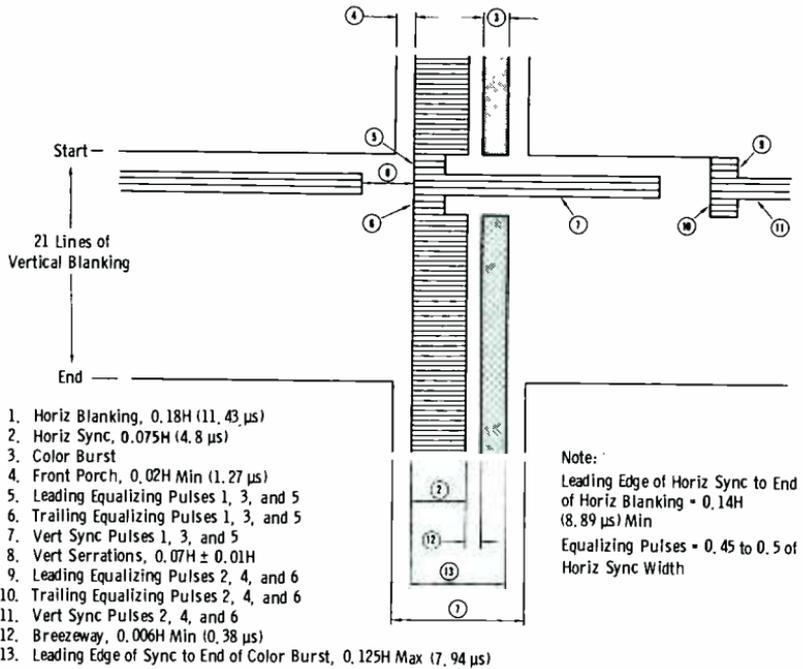


Fig. 4-32. Interlaced pulse-cross display; sync generator has lost interlace.

pulses 1, 3, and 5 of field 1 start in line with horizontal sync, and vertical-sync pulses 2, 4, and 6 start at the half-line interval. Note that this drawing interprets a single-field display of a composite picture signal in which 21 lines (maximum) of vertical blanking will be counted.

#### 4-6. SUGGESTED MAINTENANCE PROCEDURES

Modern solid-state sync generators need little in the way of preventive maintenance. This consists mainly of performance checks about once a month to assure compliance with EIA standards or (at least) FCC standards. Check all pulse widths and the performance of color genlock. Always check for proper operation of the generator-changeover switch. It is good practice to alternate sync generators in daily on-the-air service; for example, generator num-



**Fig. 4-33. Single-field pulse-cross display.**

ber 1 may be used one day and number 2 may be used the next day. In this way, any intermittent problem should be discovered.

For tube-type generators, an adequate preventive-maintenance schedule must be followed. Every 30 days, stability, pulse widths, and compliance with EIA standards should be checked. This includes an overall stability check where possible, and a check (and adjustment if necessary) of the master-oscillator frequency, counter-chain operation, and all pulse widths. It also includes measurement of pulse amplitude, overshoot, and rise times both at the generator output and at all pulse-distribution-amplifier outputs with cables connected as in normal operation.

NOTE: It is not advisable to check tubes on a tube checker at any arbitrary interval. It is better to go through certain performance tests, as outlined below, which should reveal the necessity of checking or replacing tubes. Actually, the best "tube checker" is the sync generator itself.

The overall stability test is a good indication of any deterioration of tubes and/or components. The test is possible if the master oscillator has a vernier frequency adjustment which will allow a frequency variation of 3 to 5 percent from the exact operating fre-

quency. In this test, it is assumed that one of the methods described previously has been used to adjust the master oscillator to 31.5 kHz in the free-running mode with the vernier control at midrange.

Cross-mix sync and blanking at the test points, and observe the resultant on the scope or pulse-cross monitor. The best display is obtained by using delayed sweep on a 200-microsecond-per-centimeter time base and rotating the delay control so that an entire vertical-blanking region plus several lines can be observed. The field-shift key should then be used to observe the alternate field. This procedure is unnecessary if an interlaced pulse-cross monitor is available.

Rotate the vernier frequency control between the extremes of rotation. Maximum stability is indicated if no erratic pulse behavior exists between these extremes. If horizontal pulses become erratic or change frequency, check the divide-by-two circuit (15.75 kHz) with the scope. If this output remains stable, check each following horizontal circuit until the instability becomes evident. If only equalizing pulses become erratic, check the equalizing-pulse gate and any gate amplifiers. If only vertical-sync instability occurs, check the vertical-sync gate. If both equalizing-pulse and vertical-sync instability occur check any 9H gating circuit and/or 3H stop or delay circuits, and particularly the counter chain.

Keeping a record of the rise and decay times of pulses is also a good procedure from the standpoint of anticipating troubles. The time interval of the leading and trailing edges of horizontal- and vertical-sync, equalizing, horizontal-blanking, and horizontal-drive pulses should not exceed 0.003H (0.19 microsecond). Using an expanded scale on the scope, measure the slopes between the 0.1 and 0.9 amplitude points. A record should be kept by the maintenance department so that any deterioration of pulse characteristics will be evident from month to month.

Leading-edge overshoots should be no greater than 5 percent as measured with a scope that has good transient response and with the interconnecting cables in place. When long coaxial runs are involved, an overshoot with ringing sometimes occurs. If the pulse waveshape is important (pulse not used simply as a trigger), the termination should be varied slightly around the nominal value for best transmission. Sometimes it is necessary to provide complex terminations on extremely long cable runs.

#### 4-7. SYNC CROSS TALK

The term "sync cross talk" can be applied either to cross talk within the sync generator itself or to the "windshield-wiper" effect (when on monochrome standards) similar to the horizontal motion

caused by cochannel interference on a home receiver. On color standards, a very slow drift occurs on the interfering signal.

Cross talk within the generator itself usually is caused by leakage of a very small amount of one of the counter frequencies into the master-oscillator waveform. This type of cross talk is evident on driven monitors as a slight horizontal line displacement at the vertical raster edges and on vertical lines in the picture. The frequency of the cross talk can be determined by considering the number of such horizontal line displacements occurring from top to bottom of the raster, as follows:

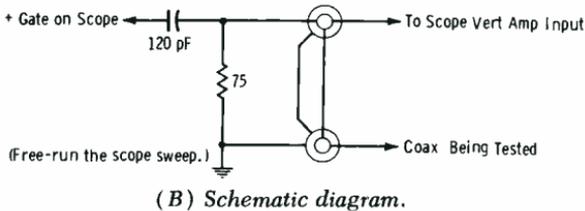
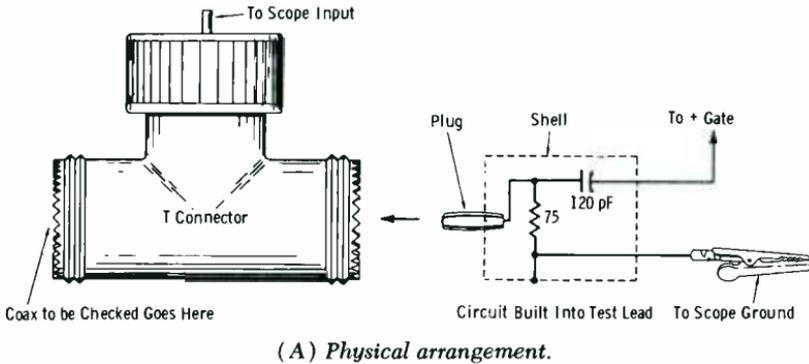
<i>Cross Talk Frequency</i>	<i>Number of Displacements Top to Bottom</i>
4500 Hz	70
900 Hz	14
180 Hz	3

If this effect should be evident on all driven monitors, the counter indicated should be shielded additionally, or wiring should be re-routed until the interference is eliminated.

A much more prevalent type of sync cross talk is that which occurs between signals from two nonsynchronous sources, such as local and network or local and video-tape signals. This trouble is evident as a vertical bar or line moving back and forth horizontally through the picture when a remote source not tied to the local sync generator is observed. When this condition exists, there will also be cross talk between local sync and the signal from the video-tape machine in the nonsynchronous playback mode. Also, in many cases the system is subject to pickup of strong fields of a transient nature, such as radiation from inadequately shielded rotating machinery.

Trouble of this nature is usually the result of ground loops. A ground loop in an otherwise well designed installation is most often caused by an open or intermittent ground at one end of a coaxial cable. Each cable should be disconnected from the sending end (only) and checked with an ohmmeter from center conductor to shield to determine if the termination resistance is obtained at the other end. If the shield is open, there will be no continuity. Always twist both the sending and receiving connectors while making this check so that loose or intermittent (or high-resistance) connections will be evident.

A much more rapid check which can be performed efficiently by one person is possible with the construction of the simple tester shown in Fig. 4-34. Fig. 4-34A illustrates the physical construction, and Fig. 4-34B is the schematic diagram. The two components (resistor and capacitor) can be part of the cable running from the

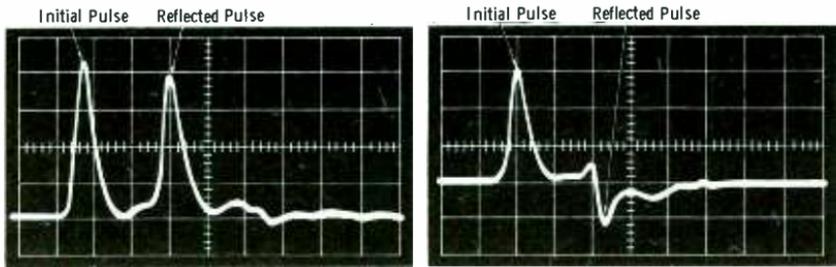


**Fig. 4-34. Tester for coaxial lines.**

plus-gate binding post to the UHF connector. The line to be checked is connected to the opposite side of the UHF T connector. When BNC connectors must be used, the resistor and capacitor should be mounted on a small board external to the connector.

The arrangement shown in Fig. 4-34 is suitable for use with the wide-bandwidth scopes which have sweeps adjusted to cover only the 10-centimeter graticule, such as Types 551, 535, 545, and 547. (The latter scopes normally use BNC rather than UHF connectors.) The procedure is as follows:

1. With the connection shown in Fig. 4-34A, free-run the scope sweep. This produces a positive pulse on the screen from the plus-gate post.
2. Connect the cable to be tested to the UHF T connector (or BNC connector when required). If the cable is terminated properly, no return pulse will appear, and the base line will be perfectly smooth. Slight bumps may be caused by junctions, such as jack fields, splices, etc.
3. If the cable is open, a reflected pulse of positive polarity will appear (Fig. 4-35A). If the cable is shorted, a negative reflected pulse occurs (Fig. 4-35B).
4. The distance to the discontinuity can be determined very closely by the following analysis:

(A) *Open circuit.*(B) *Shorted cable.***Fig. 4-35. Scope displays indicating coaxial-cable defects.**

- (A) The speed of light is 983.5 feet/microsecond  
 (B) Multiply this figure by the propagation factor of the cable used (Table 4-5).  
 (C) Divide by two. (The pulse must travel to the discontinuity, then return.) This is the factor by which the indicated space in microseconds is multiplied to obtain the distance in feet.

**Table 4-5. Coaxial Cable Chart for Tester**

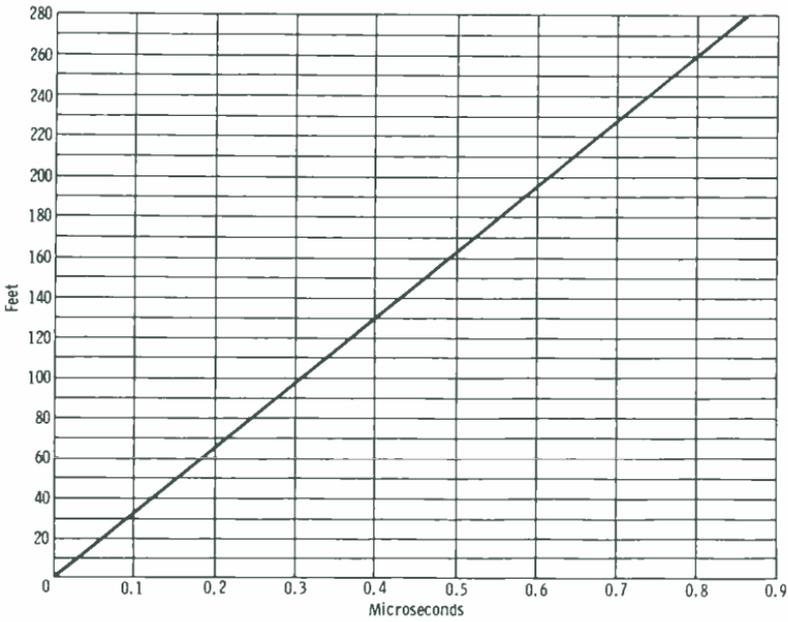
Cable Type	Propagation Factor	Multiplying Factor
Solid Poly	0.66	325
Foam Poly	0.82	404
1/2" Styro	0.89	439
3/4" Styro	0.90	443

The graphs in Figs. 4-36 and 4-37 are convenient to use when the cable has a propagation factor of 0.66, as is common for video runs. In this case, the multiplying factor is 325. Thus, if the indicated delay of the reflected pulse is 0.5 microsecond, the distance to the discontinuity is:

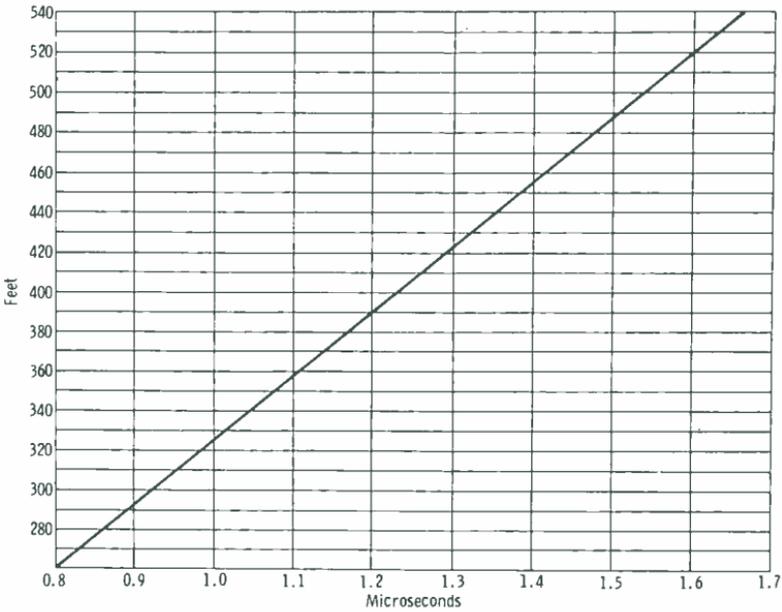
$$325 \times 0.5 = 162.5 \text{ feet}$$

The waveforms of Fig. 4-35 represent a scope sweep time base of 0.1 microsecond/centimeter. Thus, approximately 72 feet, or 0.22 microsecond, to the discontinuity is indicated by these examples.

**NOTE:** When the discontinuity is at a distance of less than approximately 40 feet (causing the reflected pulse to be superimposed on the initial pulse), use a 0.1-microsecond delay line in series with the cable tested, and subtract this amount of delay when making the final computation. A 70-foot roll of RG-11/U or RG-59/U may be used to provide this amount of delay if a delay cable is not available.



**Fig. 4-36. Relation of cable delay and cable length (up to 280 feet) for a propagation factor of 0.66.**



**Fig. 4-37. Relation of cable delay and cable length (260 to 540 feet) for a propagation factor of 0.66.**

(These cables have a delay of approximately 1.5 microseconds per 1000 feet.)

#### 4-8. COLOR SYSTEM PHASING

For the beginner, system phasing is one of the least understood subjects. See Fig. 4-38; it makes no difference which of the sets of vectors, A, B, or C, the receiver or color monitor "sees." In this case, provided the burst phase is correct (which is true for each individual set of vectors), proper color rendition will be achieved. Remember that the synchronous demodulators in the color receiver simply reference to burst phase, so there is no problem at all in system phasing if you simply switch one camera chain with its associated

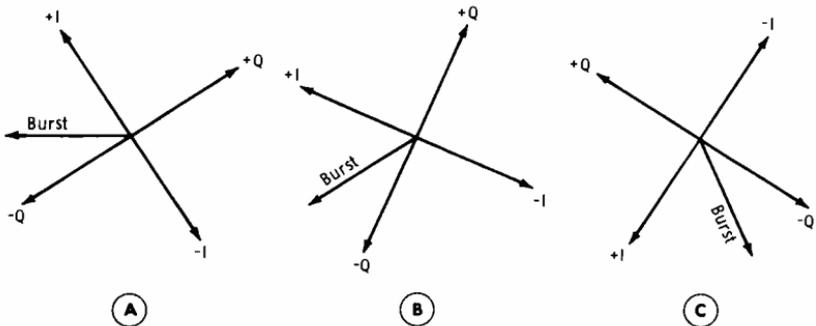


Fig. 4-38. Three sets of color vectors.

encoder on the air at any one time. But the television system will not be limited to this type of color operation. Sooner or later (if not at the beginning), you will be called upon to mix two color signals when a program requires a lap-dissolve operation or special-effects mixing, etc.

Consider the breezeway of Fig. 4-19. All camera encoders sometimes receive the same subcarrier and the same burst flag (Fig. 4-39). Normally all the camera encoders are rack mounted and reasonably close to each other. The nominal value of the breezeway is  $0.56 \mu\text{s}$ ; the tolerance range is  $0.39 \mu\text{s}$  minimum to  $0.64 \mu\text{s}$  maximum.

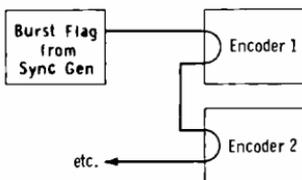


Fig. 4-39. Burst-flag routing to encoders.

The approximate pulse delay for RG-11/U cable is  $1.5 \mu\text{s}$  per 1000 ft. This is  $0.0015 \mu\text{s}$  per foot of cable. We can see that there is only a slight chance of severe timing error in the normal installation. But let us consider a situation for a large installation in which the cable run might be (for example) 100 feet longer to the farthest color unit than to the nearest unit. Further assume the nominal breezeway value of  $0.56 \mu\text{s}$  is used. The delay of the burst-flag pulse to the farthest unit is  $0.15 \mu\text{s}$ , so the encoder will put out a breezeway of  $0.56 + 0.15 = 0.71 \mu\text{s}$ . This value is outside the tolerance. But the same installation would be within tolerance if the breezeway used for the nearest encoder were, say,  $0.45 \mu\text{s}$ . Then the farthest encoder would have a breezeway of  $0.45 + 0.15 = 0.60 \mu\text{s}$ , which is within specifications. It is, however, not good operating practice to have different breezeways when mixing color sources; therefore, it is recommended to use a delay line that provides a delay to the nearest unit equivalent to that encountered for the farthest unit (undelayed burst flag to the farthest encoder) so that the breezeways are very nearly the same. In practice, all pulse distribution of color chains is carried over coaxial cables of the same length. Runs from the pulse-distribution center to the nearest camera chain are "coiled up" and made equal in length to the longest run.

Now let us consider the major source of system phasing errors, the encoded output from each of the color camera chains. You can see that without proper consideration of system phasing, the color receiver would become thoroughly "confused" by receiving such vector systems as A and C of Fig. 4-38 simultaneously. The burst phase would be the vector sum of the mixed bursts and would be incorrect for either of the two pictures.

The period of the 3.58-MHz signal is:

$$\frac{1}{3.58 \times 10^6} = 0.28 \mu\text{s}$$

Therefore, at 3.58 MHz:

$$1 \text{ cycle} = 360^\circ = 0.28 \mu\text{s}$$

Since coaxial cable has a delay of about  $0.0015 \mu\text{s}/\text{foot}$ , then:

$$\frac{360^\circ}{0.28 \mu\text{s}} = \frac{x}{0.0015 \mu\text{s}}$$

$$0.28x = 0.54^\circ$$

and  $x = 1.9^\circ$ , or practically  $2^\circ$ , per foot of cable at 3.58 MHz.

The subcarrier from each encoder can be shifted over a  $360^\circ$  range in each unit for the purpose of matching all color chains through a common switching system. Thus in Fig. 4-40, if switcher 2 is not

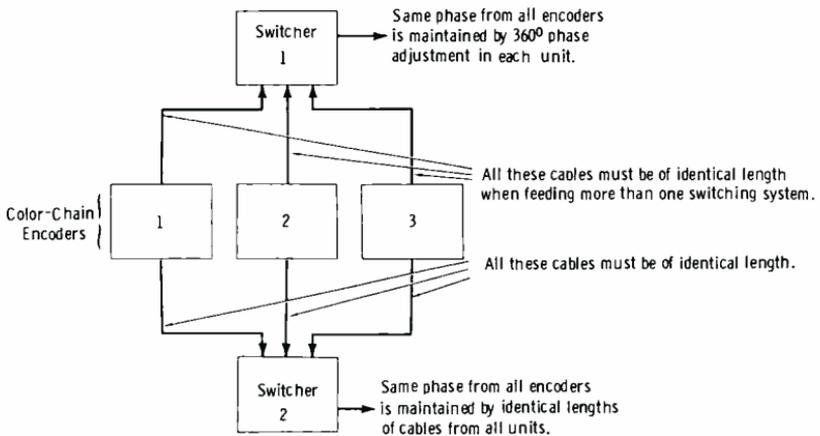


Fig. 4-40. Considerations in system phasing for color.

present, the cables interconnecting the encoders (actually *into* and *out of* distribution amplifiers) and switcher 1 can be any length. Simply adjust the subcarrier phase shifters of encoders 2 and 3 to match the vectors of encoder 1 at the output of switcher 1. This is done conveniently with a vectorscope by using the A-shared-with-B display and matching the absolute burst phases to each other. Another method useful when no vectorscope is available will be described shortly.

If you have more than one switching system, it is necessary to use identical lengths of coaxial cable from each encoder to the additional switcher so that this switcher output will have identical phases from all units. Since there is about  $2^\circ$  of phase shift per foot of cable at 3.58 MHz, to maintain less than  $1^\circ$  of mismatch the cables must be no different in length than six inches. In practice, if they are cut to precisely the same length, you will have no problem with any measurable phase mismatch, with a given type of cable.

If it is necessary to mix two or more switcher-system outputs together, you must have identical length cables in all encoder-switcher interconnections, and identical internal delays in each switching system through all switch banks. Then, identical-length cables to the third switching point must be used.

A "color-approved" type of special-effects system provides a satisfactory color-match (system phase) facility when a vectorscope is not available. In such a system, burst from only one source is carried at a time. In "wiping up" from bottom to top, burst from the bottom bank is being passed. When the midway point is reached, continuing the upward "wipe" deletes the bottom-bank burst and passes burst from the top bank. Thus, if you feed the

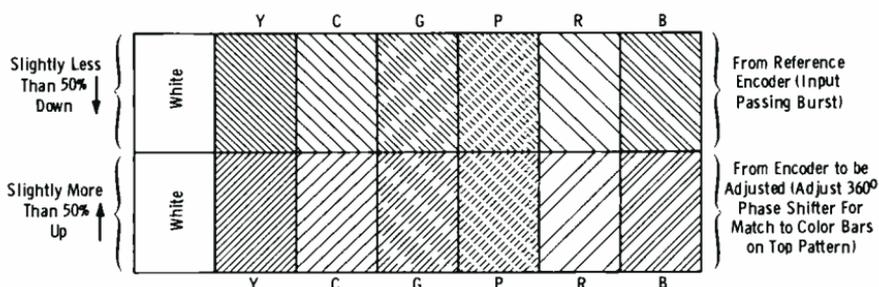


Fig. 4-41. Use of special-effects unit for phase comparison.

reference encoder signal to one bank and use the split-field vertical wipe as in Fig. 4-41, the encoder to be adjusted can be “wiped up” a little over half way, as shown. Since the color monitor receives only one burst (that from the reference encoder to which all others are to be matched), you can adjust the  $360^\circ$  phase shifter of any additional encoder for a perfect color-bar match. This means that the absolute phase for both encoders at a common switch point being observed is identical, and their signals can be mixed without color contamination. Be sure the luminance output levels and chroma output levels are identical from both encoders so that you are not confused by brightness differences in colors. When the adjustment is completed, you should be able to wipe through both sources on the switcher without shift of colors. This assumes that the switcher is properly timed internally (next chapter).

## EXERCISES

- Q4-1. For color operation, when we speak of 31.5-kHz pulses, what is the actual pulse frequency?
- Q4-2. How many leading and trailing equalizing pulses should exist in the vertical interval of composite sync?
- Q4-3. How are “field 1” and “field 2” identified?
- Q4-4. What is the approximate delay of signals in video coaxial cable?
- Q4-5. A foot of video cable represents what shift in color-subcarrier phase?
- Q4-6. Does the burst flag have any effect on the phase of the color-subcarrier sync burst?
- Q4-7. How would you measure the breezeway?
- Q4-8. If you find it impossible to eliminate the flag pulses for the entire 9H interval and you have replaced tubes or transistors to no avail, what are the most likely troubles?

- Q4-9. If you want to use nine cycles of color burst, how many peaks should you count on the interlaced color-burst pattern?
- Q4-10. What is the proper amplitude of the color-sync burst?
- Q4-11. Does the color-sync burst occur following every horizontal-sync time in the complete composite color signal?
- Q4-12. How many cycles of color-burst signal should be present?
- Q4-13. What type of oscillator is universal in sync generators?
- Q4-14. If you notice dots traveling from top to bottom of a picture monitor when viewing a color signal, what does this indicate?
- Q4-15. You are adjusting counter chains before setting the subcarrier frequency. Will this frequency adjustment upset the counter adjustments?
- Q4-16. How do we arrive at the figure 113.75 for the subcarrier count-down?
- Q4-17. If you do not have the same system phase for all color sources at the studio, will receivers show a shift in colors when you switch from one source to any other source?

## Video Switching and Special-Effects Systems

There are three general types of video switchers:

1. Mechanical push-button switching with video on the actual switch contacts. The bank of switches is interlocked to prevent more than one source from being "punched up" at a time. This type of video switcher is used mainly for certain monitoring purposes, such as selecting the source of feed to a picture or waveform monitor.
2. The relay switcher employs remotely controlled, rack-mounted banks of relays; the switch banks are not interlocked, since the interlocking function is in the relay arrangement. This type of switcher was at one time the most commonly employed for studio installations.
3. The most recently developed video switcher is the vertical-interval switcher that uses solid-state switching plates timed to switch video sources in an interval of a microsecond or two during the blanking time following vertical sync. This type of switching system has replaced the relay type in older stations, and it is used almost universally in new stations. It may or may not be tied to computer control.

A popular type of television control-room switching is known as *audio-video* switching. With this method, audio-channel relays may be operated from the video system. Or, an audio console with its associated relays may provide for operation of the video relay or crosspoint bays. Such systems are usually arranged so that either separate or tied-in operation may be used.

When several channels must be switched simultaneously among several outgoing lines, *preset switching* methods are used.

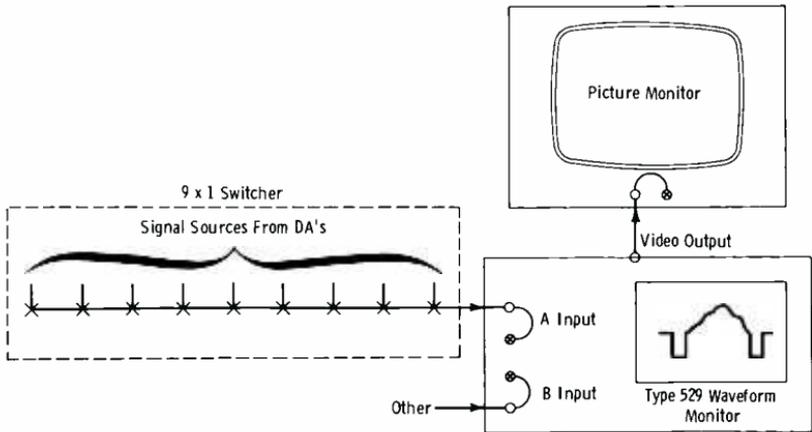
NOTE: It is assumed in the following study of maintenance procedures that the reader has a background in basic switching systems equivalent to the study presented in Chapter 7 of *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*.<sup>1</sup>

**5-1. MECHANICAL PUSH-BUTTON MONITOR SWITCHING TECHNIQUES**

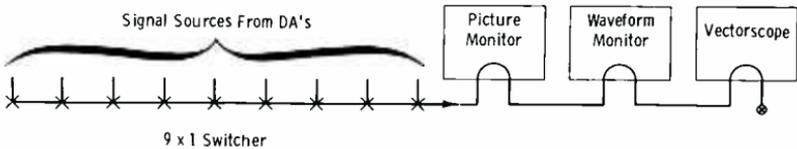
Many picture and waveform monitors in the system are tied to a particular picture source at all times. There are also many instances in which it is advantageous to provide switchable signal sources to a picture monitor, waveform monitor, or vectorscope, or a combination of such monitoring devices. The simple mechanical type of switcher with interlocked push buttons is often used for this application, especially in smaller stations.

In the examples illustrated in Fig. 5-1, a 9 x 1 switcher (9 inputs, 1 output) is employed to select the signal to be monitored. The method of Fig. 5-1A has the advantage of using the video output of

<sup>1</sup> Harold E. Ennes, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals* (Indianapolis: Howard W. Sams & Co., Inc.).

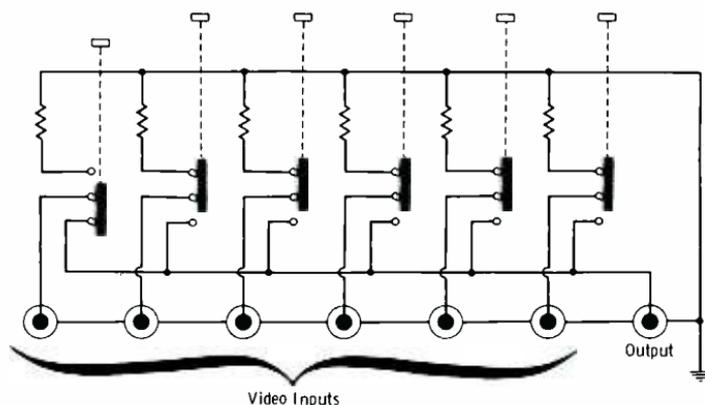


(A) Waveform monitor feeds picture monitor.



(B) Input looped through monitors.

**Fig. 5-1. Typical methods of monitor switching.**



**Fig. 5-2. Basic terminating-type passive video switcher.**

the Tektronix Type 529 or 1480 waveform monitor to feed the picture monitor. This output contains the brightening pulse so that as the DELAY control is varied, the exact lines being observed on the crt are indicated on the picture monitor.

The switchable monitoring system is extremely useful in balancing color camera chains to each other when the picture monitor is a color type. Such a system also provides a quick check of incoming and outgoing lines, switcher inputs and outputs, etc.

The basic types of passive video switchers are terminating, bridging, or a combination of the two. Fig. 5-2 illustrates the simplest form of terminating switcher. Terminating switchers connect a single 75-ohm output to any one of the 75-ohm inputs. The example of Fig. 5-2 is a  $6 \times 1$  arrangement. Note that the first button on the left is depressed, disconnecting the built-in terminating resistor and connecting the input to the output bus which must feed a 75-ohm terminated line.

Bridging-type switchers have a pair of looping connectors provided for each input. Each input is connected to the corresponding switch through a toroidal inductor. A fixed capacitor, in conjunction with this inductor, provides high-frequency compensation to correct for the distributed capacitance of the switch, internal wiring, and a short length of coaxial cable.

Fig. 5-3 shows a typical  $6 \times 1$  bridging switcher that includes indicating lamps for the circuit selected. Note that S1 is the button depressed. The associated tally-lamp circuit is completed through switch contacts 2 and 3. The output bus is connected to input 1 through contacts 5, 6, 11, and 12. Compensating capacitor C1 is lifted from ground when the switch is repressed to select the corresponding input. Note that on the switches not depressed, the associated capacitor is connected to ground.

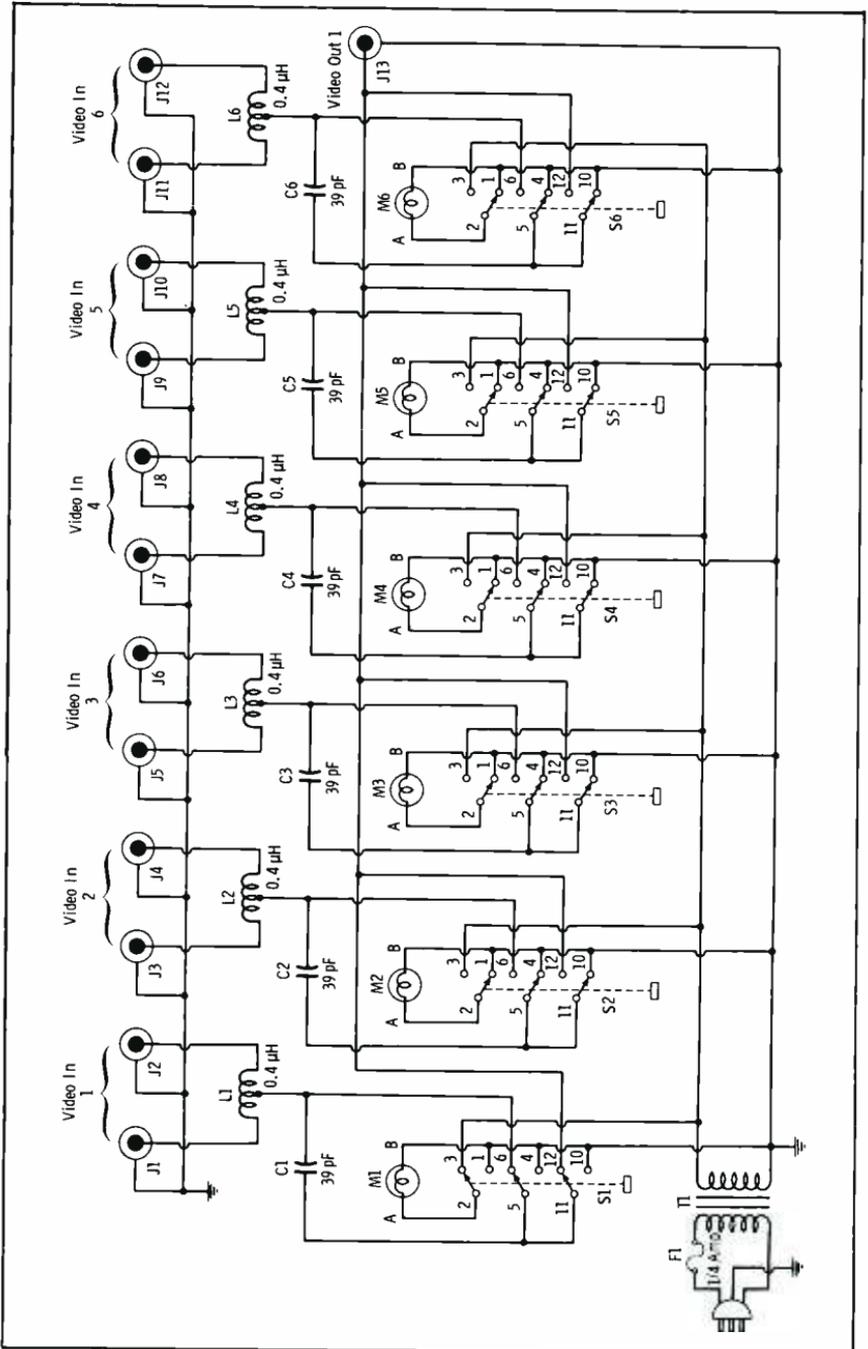


Fig. 5-3. Example of a bridging-type passive video switcher.

A fixed capacitor (39 pF) acts in conjunction with a toroidal transformer (0.4  $\mu$ H) at each video input to provide a near perfect impedance match at all video frequencies. This capacitor is removed when an input button is selected, and the match then is maintained by the capacitance of the interconnecting cable and the input capacitance of the driven device. This close attention to impedance matching prevents loss of picture quality whether color or monochrome, even when large switching matrices are used.

The switcher of Fig. 5-3 is designed to operate into the high-impedance video input of a distribution amplifier, monitor, or other device. A total of 40 pF of external output capacitance is fully compensated by the switcher design; however, tests have established that an additional 100 pF of capacitance can be added at the output without noticeable degradation of a 1000-line picture. This permits operating the switcher up to five feet from the driven device when 75-ohm cable is used, or at greater distances when low-capacitance cable is used.

Fig. 5-4 illustrates a typical application for monitoring facilities. By means of the loop-through connectors, several switchers may receive the same signal, the last switcher providing a termination on the loop-through jack. The monitor should be physically located within a few feet of its associated switcher to avoid excessive capacitance build-up on the bridging input.

Another application for bridging-type passive switchers which provide the loop-through facility is that of using a number of  $6 \times 1$  or  $9 \times 1$  units to preselect the signals which feed a common output switcher. Fig. 5-5 illustrates a vertical line-up of four RCA Type TS-2,  $9 \times 1$  passive video switchers, modified for loop input, loop

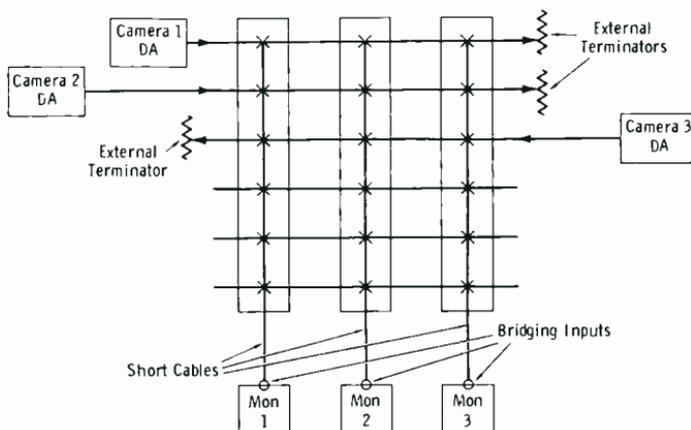
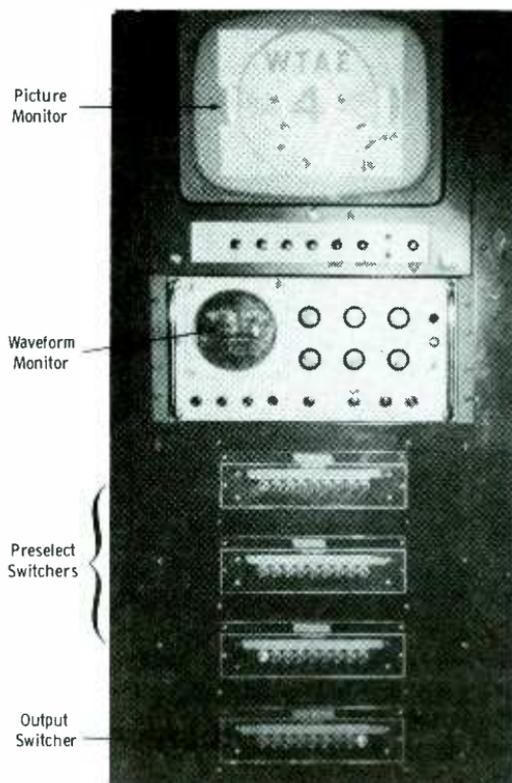


Fig. 5-4. Typical monitoring application of bridging switchers.

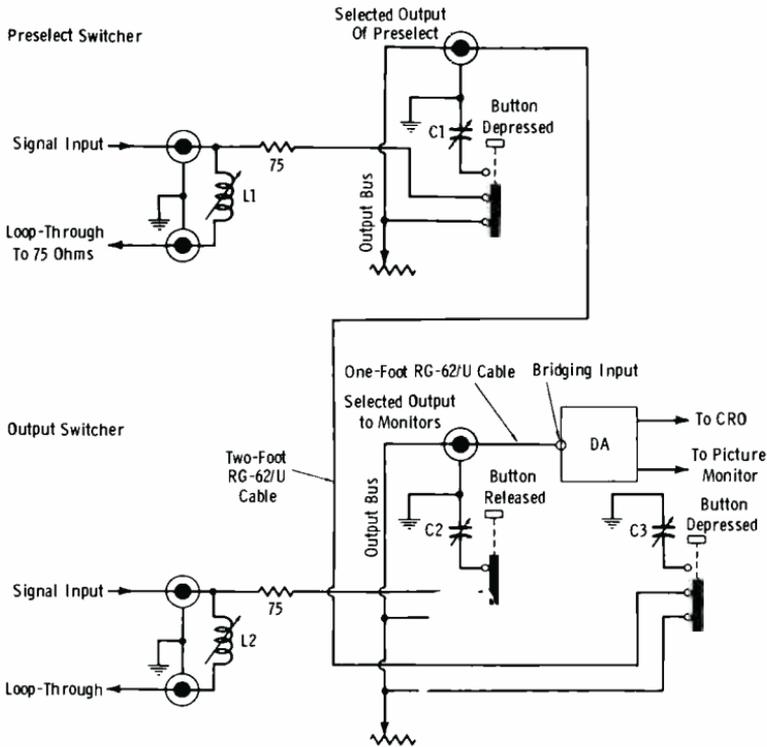


**Fig. 5-5. Installation using three preselect switchers with one output switcher.**

output connections and arranged as shown in Fig. 5-6. Each of the three upper units has its output tied to one of the first three input buttons of the bottom unit. The bottom unit is the monitor-feed output; it feeds a video DA, which in turn feeds the cro and picture monitor. Thus, any source selected on the top switcher appears on switch 1 of the bottom switcher, etc. The output switcher is then able to provide an additional 6 signal sources, for a total of 33.

Observe Fig. 5-6. A single button on the preselect switcher is shown depressed, with the selected output feeding the output-switcher button which is also depressed. Thus the desired preselected signal is being fed by the output switcher to the monitoring facility. Also shown on the output switcher is one of the buttons released, illustrating the contact arrangement in the video path for the released position.

Variable coils L1, L2, etc., are provided for each loop-through position to neutralize the distributed capacitive effects. Capacitors

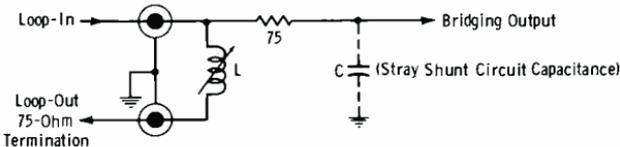


**Fig. 5-6. Fundamentals of switcher arrangement of Fig. 5-5.**

C1, C2, etc., provide the compensating equivalent capacitance of the bridging arrangement when a button is released, which disconnects the bridging output from the associated signal source.

Fig. 5-7 shows the equivalent circuit for a single path. The total capacitance (C) is made up of the following capacitances when the RCA Type TS-2 modified switcher is used.

- TS-2 Input capacitance: 15 pF
- TS-2 Output capacitance: 18 pF
- Two-foot RG-62/U capacitance with connector: 30 pF
- One-foot RG-62/U capacitance with connector: 15 pF
- Video DA input (TA-3): 22 pF



**Fig. 5-7. Equivalent circuit of single path.**

Now note that the signal "sees" the input capacitance of the switcher, the output capacitance of the same switcher, the input capacitance of the output switcher, and the output capacitance of the output switcher, in addition to the capacitances of the RG-62/U and the DA input. Therefore, the actual capacitance is  $15 + 18 + 30 + 15 + 18 + 15 + 22 = 133$  pF.

Since  $R = 75$  ohms and  $C = 133$  pF, the optimum nominal value of coil  $L$  is:

$$\begin{aligned} L &= R^2C \\ &= (75)^2(133 \times 10^{-12}) \\ &= (5625)(133 \times 10^{-12}) \\ &= 0.75 \mu\text{H nominal} \end{aligned}$$

The bandwidth capability of this system to the  $-3$ -dB point is:

$$\begin{aligned} f(-3 \text{ dB}) &= \frac{1}{2\pi RC} \\ &= \frac{1}{6.28(75)(133 \times 10^{-12})} \\ &= 16 \text{ MHz} \end{aligned}$$

The proper adjustment of such a video switching system is as follows:

1. Connect an oscilloscope across the output of a video-sweep generator which sweeps to at least 10MHz.
2. Connect the sweep-generator output to a length of coaxial cable equivalent to the total length of video signal cable including the loop-through path. Say this is 1000 feet. Then use 1000 feet from the sweep generator to the switcher input, and *terminate* the loop-through connector.
3. Adjust the inductor ( $L_1$ ,  $L_2$ , etc.) for the specific input being observed for minimum standing-wave ratio as observed on the scope.
4. Release the button. Adjust the trimmer capacitor ( $C_1$ ,  $C_2$ , etc.) for the specific input being observed for minimum standing-wave ratio.

The coils should be able to tune through the nominal value of  $0.75 \mu\text{H}$ . A typical value for the variable trimmer capacitors is 10-180 pF.

## 5-2. PRODUCTION SWITCHERS

The usual program production switcher consists of two main sections: (1) the switcher-effects system itself and (2) the machine

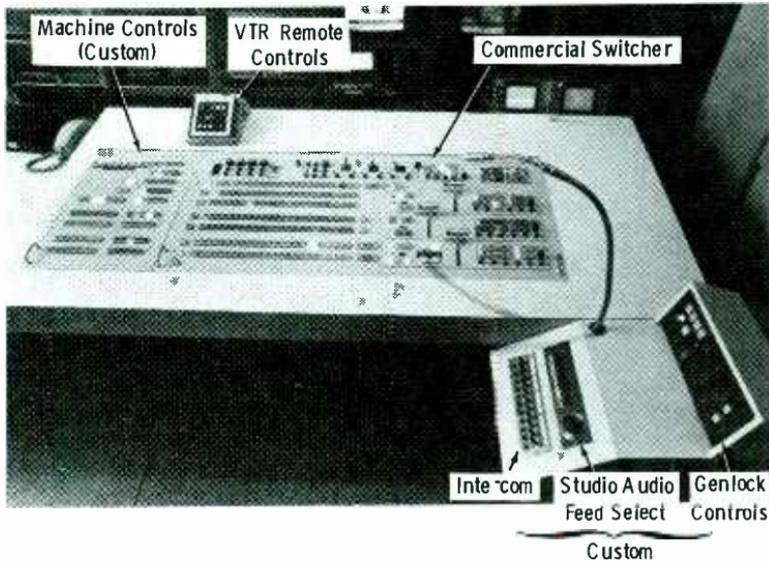
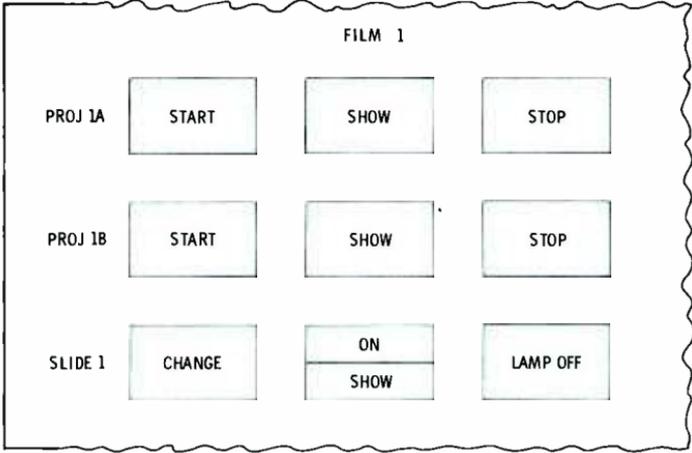


Fig. 5-8. Example of studio switching system.

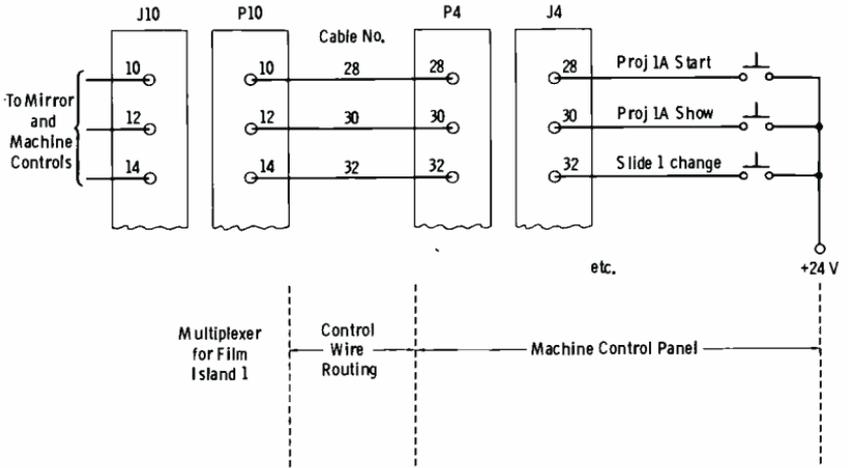
control panel. Fig. 5-8 shows a representative arrangement for the studio switching-system panels at a medium-sized station. Very often, the main switching and effects panel works with commercially designed and fabricated units, while auxiliary equipment may be custom made for the particular station either by the manufacturer or by the station staff. You will find a wide variance in production switching arrangements, depending on the size and activity of each station and the degree of "modernization" that exists.

The machine controls are associated with film and tape equipment. Each film island normally consists of a multiplexer, two film projectors (usually designated A and B), a slide projector, and a camera. Operation of the multiplexer mirrors selects the desired source and routes its image to the camera. The film and slide projectors are normally plugged into the multiplexer unit, through which all control voltages are routed. The maintenance engineer must be familiar with the location of wiring information as represented by Fig. 5-9. Fig. 5-9A shows a small portion of the typical machine control panel, and Fig. 5-9B is an example of the control-wiring information that should be on hand in the maintenance files.

Troubleshooting procedures are obvious. For example, pressing the PROJ 1A START button places 24 volts on the start relay of projector 1A via the path shown in Fig. 5-9B. This automatically turns the projection lamp on. Then after the three-second preroll, pressing the PROJ 1A SHOW button operates the proper mirror in the



(A) Portion of panel.



(B) Control wiring diagram.

Fig. 5-9. Wiring information for machine control panel.

multiplexer to route the light from projector 1A to the camera. When the audio-follow-video switch (not shown) is on, the audio from the projector is automatically switched to the audio line. For automated station breaks, such operations are normally pulsed from the programmed memory in the central control unit of the automation system.

Fig. 5-10 represents an earlier type of video switcher with special effects added to the lap-dissolve amplifier. The very first production



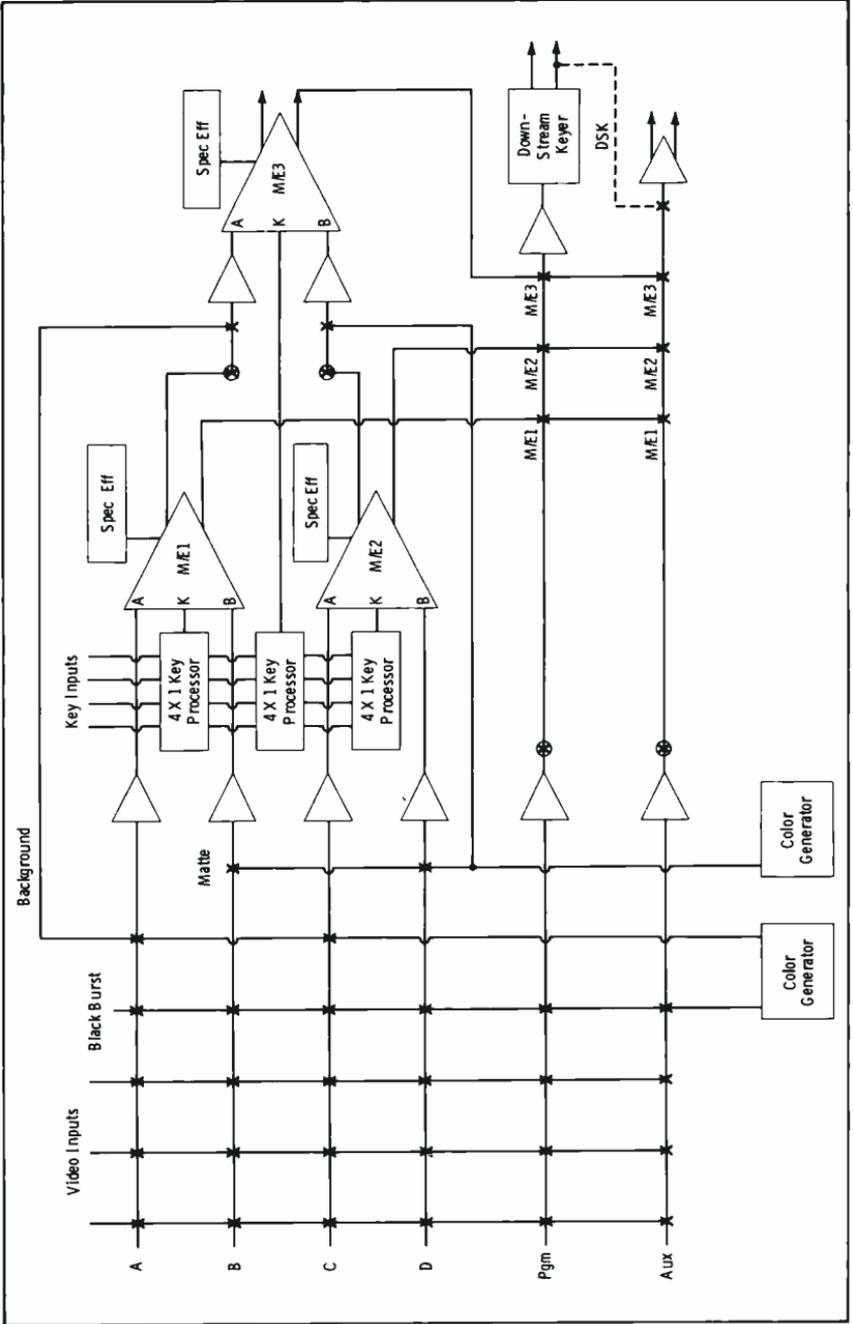


Fig. 5-11. "Fourth-generation" switcher with mixer/effects on all buses.

switchers employed only a lap-dissolve arrangement without any means of keying portions of the signal. Two pictures could be superimposed but not "keyed." Thus Fig. 5-10 actually represents a "second-generation" switcher, many of which are still in use at smaller stations with limited production facilities. This type of switching system required the use of delay lines (actually coiled-up coaxial cables) in the video paths to obtain the same system color phase throughout the various paths used.

Fig. 5-11 shows a much later switching system in which mixer/effects (M/E) amplifiers on all buses eliminate the use of delay lines and provide maximum production facilities. This arrangement allows almost anything to be done that would be desired. Some possible operations are:

1. A-B wipe with border
2. A-B wipe through 100% border
3. Mix or dissolve or wipe to a preset wipe
4. Mix or dissolve or wipe to a key
5. A-B mix behind a chroma key
6. A-B wipe behind a chroma key
7. A-B wipe with borders behind a chroma key
8. Mix or wipe to a quad split behind a chroma key
9. Luminance key over a quad split behind a chroma key

Downstream keyers are provided with program-bus genlock and pulse and subcarrier outputs. Thus it is possible to key over non-synchronous sources. The downstream keyer genlocks the switcher to the incoming nonsynchronous video and provides a source of timing signals for peripheral equipment such as character generators.

### 5-3. SWITCHER-SYSTEM MAINTENANCE

Regardless of the type of control circuitry employed in a switching system, certain basic tests may be used on a periodic basis to assure proper performance of the video-signal paths. Proper level adjustments throughout all the paths of the video switch banks should be made first so that the switcher output indicates the same level from all sources. A fixed signal, such as a calibration pulse of the proper amplitude, is more accurate to use than a fixed video signal, such as a slide pattern.

#### Level Sets

A. Feed a calibration pulse (0.7 volt noncomposite, or 1 volt composite) to a given switcher input. Do not use a single-frequency sine wave, since the level then depends on the frequency response;

this is a separate check. If you have a window generator with the proper switcher input amplitude, this may be used.

B. Check each amplifier in the path for each destination (air, preview, etc.), and adjust the level to the value which should exist at each point (normally 0.7 volt or 1.0 volt, as above). This is to ensure against overloading any one amplifier in the path even though the final output signal may be proper.

C. A final check should be made at the output bus by switching the signal input through every possible path. The following is a general outline of paths involved for all following tests:

Key bank (when used)

Fader A bank through special effects

Fader B bank through special effects

Fader A bank, normal fade or mix position

Fader B bank, normal fade or mix position

Engineering preview, normal

Engineering preview, on air (when provided)

All director preview paths

All additional paths in more modern switchers

### Amplitude-Versus-Frequency Test

Use the procedure shown in Fig. 5-12. If a dual-trace scope is not available, observe the output of the sweep generator (properly terminated) through the detector probe; then observe the output of the switcher for comparison. Fig. 5-13 shows typical video-response curves. The markers are at 5 and 10 MHz. Fig. 5-13A was obtained with internal 60-Hz sweep used on the scope. Many technicians prefer this type of sweep because of the well-defined base line produced. Fig. 5-13B shows the same response as observed with a normal 60-Hz time base. This type of presentation is often preferred because it permits observation of the actual shape of the rolloff; this shape is quite important for good transient response. When a good modern 30 to 50-MHz scope is used, the detector is not necessary.

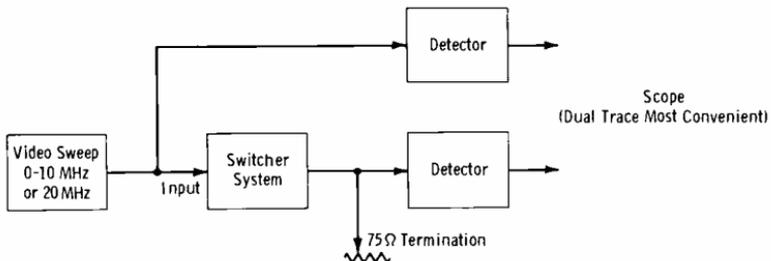
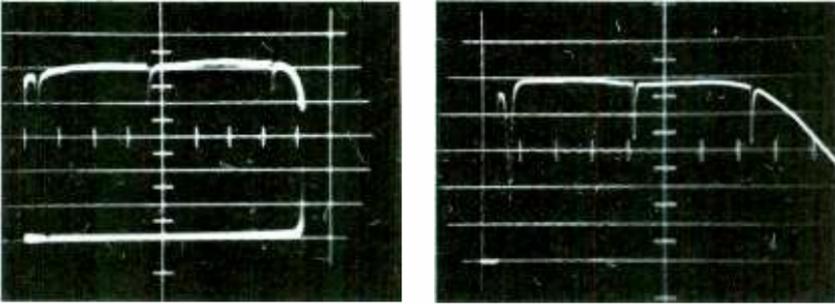


Fig. 5-12. Basic test setup for amplitude-versus-frequency test.

### Cross-Talk Test at 3.58 MHz

A. Feed the local 3.58-MHz subcarrier to input 1 and terminate all other inputs. Adjust the input amplitude to 1 volt peak-to-peak.

B. Operate crosspoint 1 and adjust the signal for an output of 1 volt peak-to-peak.



(A) *Internal 60-Hz sweep.*

(B) *Normal 60-Hz sweep.*

**Fig. 5-13. Typical video response curves for modern switching systems.**

C. Leave the crosspoint on. Measure the cross talk at all other outputs by increasing the scope gain to maximum sensitivity. Normally, the cross talk should be at least 50 dB down from 1 volt (3.2 mV). Check the manufacturer's specifications.

D. Repeat for all inputs.

### Noise Test

A & B. Set up as for the cross-talk test.

C. Disconnect the 3.58-MHz signal and terminate the input.

D. Read the residual noise.

E. Repeat for all inputs.

### Differential Phase and Gain at 3.58 MHz

A. Feed stairsteps with a superimposed 3.58-MHz signal to each input of the switcher in turn.

B. Measure differential gain and differential phase at 10, 50-, and 90-percent APL. (Techniques are covered in Chapter 8.)

### Low-Frequency Tilt

A. Use the setup illustrated in Fig. 5-14. Either a square-wave (60-Hz) or window signal may be used.

B. With the scope on a field-rate time base, measure the tilt of the switcher output waveform compared to the input waveform.

C. Repeat for all inputs. The tilt normally should be less than 2 percent. Check the manufacturer's specifications.

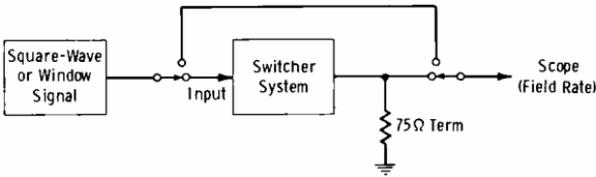


Fig. 5-14. Test setup for low-frequency tilt.

### T-Pulse Test

A. Use the same setup as above, except substitute a T-pulse generator.

B. Check the signal directly ( $T = 0.125 \mu\text{s}$  h.a.d.).

C. Feed the signal through the switcher, and measure the distortion at the output. (Interpretation of the T pulse is covered in Chapter 8.)

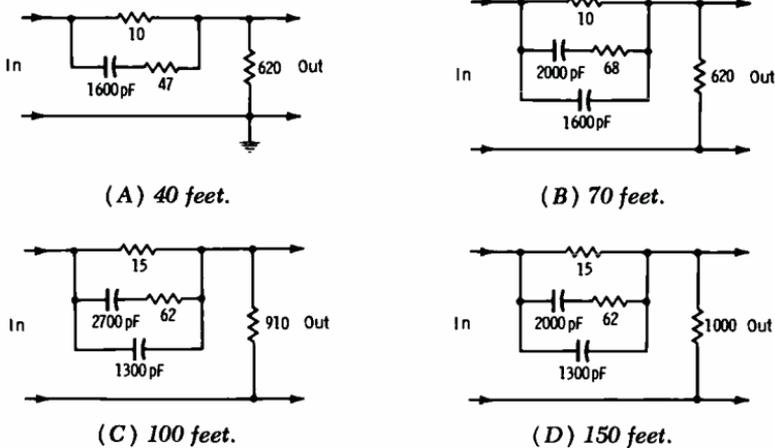
D. Repeat for all inputs and all paths.

### Troubleshooting

One of the most consistently occurring problems in some switching systems is keeping all paths through the switcher properly timed (phased) at the color-subcarrier frequency. Thus, when two color signals are mixed in the switcher, color distortion occurs. We have discussed the importance of assuring that all color sources arrive at the switcher *inputs* in phase coincidence at the burst frequency. All paths within the switching system itself must then maintain the same phase relationship (have the same delay) to the output bus.

A quick and practical check on internal switcher timing can be made by using a color monitor and special effects. Feed the same color-bar signal (common source) to two special-effects inputs, (such as effects fader A and effects fader B), and use the vertical wipe feature while observing the color monitor at the output. As you wipe from bottom to top, observe the color-bar pattern on the monitor for any difference in hue between the two parts of the split screen. Remember that only the burst from the top bank is being used until the wipe reaches just past the 50-percent point; then the burst used is from the bottom bank. Any difference in hue indicates the internal paths of the switcher are not properly timed with appropriate cable lengths.

Sometimes a switcher does not remain timed because of emergency switching from one system to another of amplifiers in the various paths. Also, in some circuitry, an adjustment of a capacitive trimmer for frequency response will affect, to some extent, the 3.58-MHz delay through the amplifier. It is a good idea to mark all amplifiers to show the position when the switcher was initially timed, and interchange only in an extreme emergency. Remember that a



**Fig. 5-15. Typical video equalizer circuits for cable lengths indicated.**

delay equivalent to only one foot of coaxial cable results in  $2^\circ$  of shift at 3.58 MHz.

It is almost mandatory that all long lengths of cable used for delay-matching purposes within the switcher be properly equalized so that a minimum of frequency-compensation adjustment within the amplifiers is required. Fig. 5-15 illustrates typical equalizer values for the cable lengths shown. These networks must be used between a 75-ohm sending-end impedance and a 75-ohm receiving termination. All modern amplifiers employ a 75-ohm internal output impedance.

The only difference between maintenance of the older-type switcher system and maintenance of other units is in the type of control switches. The older system involves a large number of push-button assemblies, in some cases interlocked, and, in the case of remotely controlled relay systems, a large number of video relays. Also usually included in this system are the intercommunication sound circuits to the cameras and production staff, as well as to the relay transmitter on field units.

The most common source of trouble in older switching systems is the presence of minute amounts of dirt or foreign matter on the contacting surfaces. In the case of field equipment, mountings and connections must be tightened and cleaned often. See that all moving parts of the switch assembly move freely without a tendency to bind but with sufficient tension. A flashlight and dental mirror are handy tools to have for inspecting and servicing certain types of switches.

Relays normally are enclosed in a dust-tight cover over the panel mountings, and they should never be serviced except when trouble

is indicated. When necessary, relays should be inspected for correct spacing and proper alignment of contacts. See that connections are tight and that wiring is not becoming frayed. Cultivate the habit of checking coils for signs of excessive heating. Contacts may be cleaned by simply pulling a narrow strip of canvas or linen cloth through the contacts with the points held closed to provide sufficient pressure. At longer intervals, crocus cloth dipped in carbon tetrachloride should be used for this process and then followed by a dry linen cloth.

A trouble commonly encountered in relay-type switchers that employ dc-coupled cathode-follower outputs is a rapid "bounce" which may occur on a switch between signal sources or even on a drastic change of scenic brightness (duty cycle or average picture level). This is often caused by the tube in the output stage, and in stubborn cases it can be remedied only by careful selection of new tubes. When this dc bounce is coupled to a following distribution amplifier that employs dc-coupled feedback circuits, severe fading for two or three cycles may occur until the feedback becomes stabilized.

A good way to check for tube selection in such instances is to employ an adjustable-APL (average picture level) staircase generator, or use two slides with widely different densities in a film chain. The effectiveness of each selected tube can be evaluated critically by means of an oscilloscope connected to the distribution-amplifier output; note the deflection in centimeters as the duty cycle is varied or the two slides are switched alternately. A video monitor used in conjunction with the scope allows correlation of allowable bounce as indicated on the graticule with satisfactory or unsatisfactory visual response on the monitor. Do not use drastic changes, such as open-gated projector to no light; this far exceeds normal operating limits and will usually result in severe bounce in a normal system.

#### 5-4. SPECIAL-EFFECTS SYSTEMS

The special-effects system is an integral part of all modern switching systems. We can consider special effects as a built-in and important part of switchers for modern telecasting. It is usually a part of this system that has made the composite-signal bank and non-composite-signal bank a thing of the past. It is no longer necessary to require noncomposite signals at the switcher inputs. Most recent systems accept either noncomposite or composite signals at the inputs and have automatic sensing circuits to add sync when the signal is noncomposite. Even nonsynchronous sources may be used with automatic circuits that prevent mixing of a signal with a non-

synchronous signal until the original signal is faded completely to black so that the roll-over occurs in black.

NOTE: The fundamentals of electronic special-effects equipment and its operation are covered in *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*.<sup>2</sup>

### Keying Functions

Regardless of the size and complexity of a special-effects system, the basic function is as shown in Fig. 5-16. A circuit performing as an electronic switch provides high-speed changeover between two synchronous video sources. The keying signal which actuates the switch is derived from an internal electronic pattern generator or an external camera source. Usually, another switching bus is located on the operating panel to permit selection of the keying signal from any source, as shown in Fig. 5-17. The external key can work either on luminance differences or on chroma of a specific hue and saturation. In chroma key, the red, green, and blue channels of a specific camera are combined to form a single keying signal (Fig. 5-17). The unit separates a selected saturated color from all other colors

<sup>2</sup> Harold E. Ennes, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals* (Indianapolis: Howard W. Sams & Co., Inc. 1971).

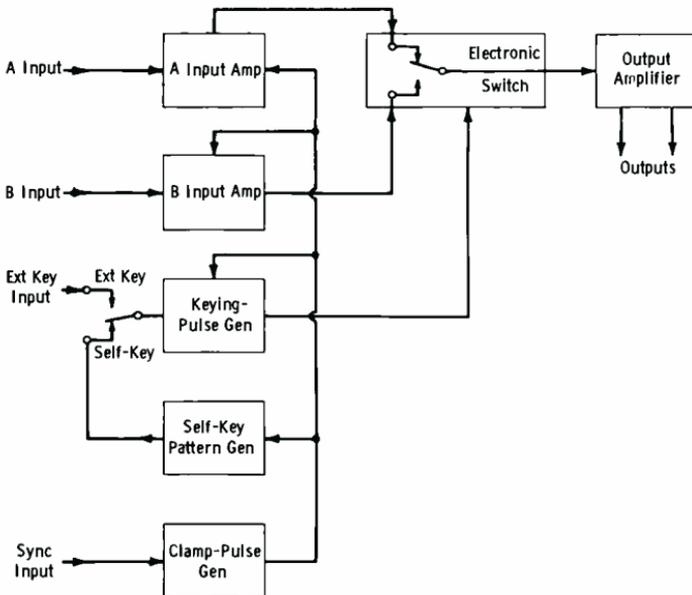
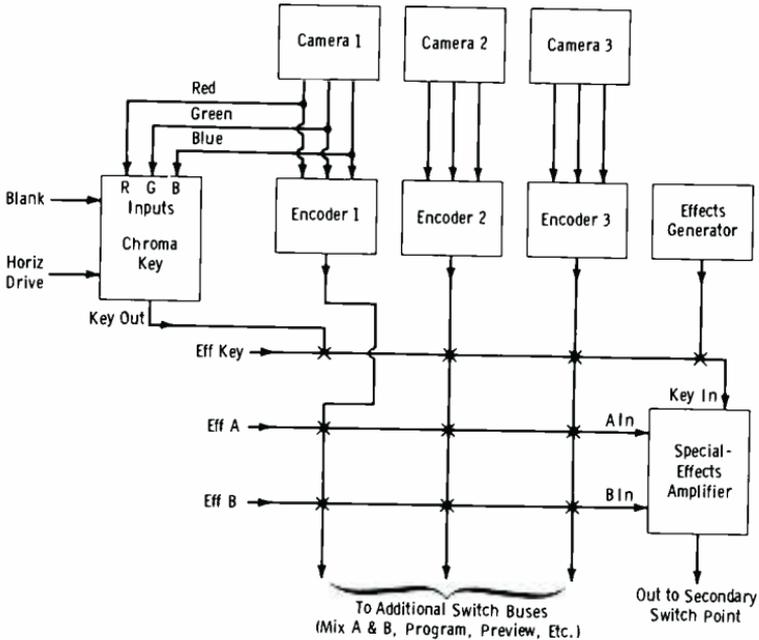


Fig. 5-16. Basic block diagram of a special-effects system.

to form a "hole" in the background so that a second video source is keyed in the background of the subject on the other switch bank.

In general, the internal pattern generator provides the keying waveforms for wipes (horizontal, vertical, rectangular, diamond, "venetian blind," etc.), whereas external key is used for insert or inlay effects. The associated effects push buttons determine relay closures to be made for diode or transistor matrices which fix the pattern generated.



**Fig. 5-17. Use of effects key switching bus.**

The pattern generator basically produces rectangular, sawtooth, triangular, and parabolic waveforms synchronous with the input video signals (local sync generator), and at both horizontal and vertical rates. When the pulses are of single line or field duration, a full-line or full-frame transition wipe is performed. When the rate is at some even multiple of the line or field rate, multiple splits in the frame occur. Multiple frequencies usually provided are 2, 4, and 6 times the horizontal and vertical frequencies.

The reader of this text should already be familiar with the generation of such waveforms as sawtooth, rectangular, and parabolic pulses. The triangle is more unconventional and should be mentioned. This waveform is generated by integrating a square wave that has a 50-percent duty cycle. Thus integrating circuits are em-

ployed to generate horizontal- and vertical-rate triangular pulses from the applied horizontal- and vertical-rate square waves.

### Automatic Sync-Adder Functions

Although most switching systems are now operated with full composite signal inputs, the modern mixing amplifier accepts any combination of synchronous input signals, composite or noncomposite. This is accomplished by clipping and adding sync continuously. In each input channel, before mixing, the signal is clipped just below burst level ( $-20$  IEEE units). For either composite or noncomposite signals, this establishes both the black level at the subsequent mixing or gating point and a definite bottoming point in the event of APL shift as a result of switching ahead of the mixer. Next, an adjustable level of sync is gated into the signal; this is set to produce the required 40 IEEE units of sync regardless of the level of sync on the signal at this point. Thus, the output sync level is correct for inputs ranging from no sync to excessive sync, and it is also independent of the fading action which may occur in some operating modes. In the special case when all inputs are noncomposite, the sync-add feature may be used normally for a composite output (sync inserted after mixing), or it may be disabled by means of a switch on the video output module for noncomposite output.

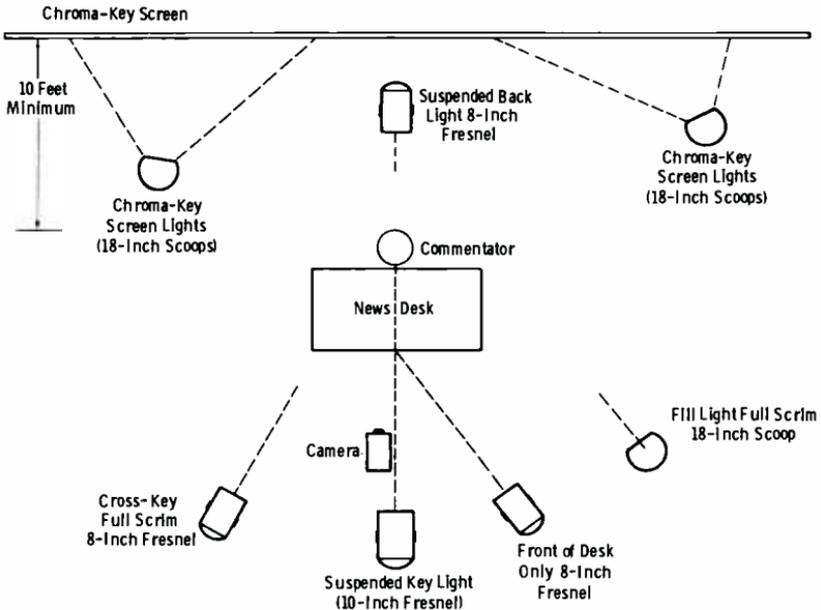
If a dissolve between a synchronous signal and a nonsynchronous signal is attempted, an inhibitor circuit prevents the signals from mixing. No dissolve takes place as the control levers are moved. Instead a fade to black occurs, and when the limit is reached at the opposite end of lever travel, a switch is made to the nonsynchronous signal at full level.

## 5-5. PROBLEMS WITH CHROMA KEY

Modern chroma-key units are stable in performance, and any problems normally result from improper operation of the keying setup. In this context, "operation" means the control of light bounce from the chroma-key backdrop or shadows that cause "tearing."

Provided that the keying gain is properly set, the most common complaint is a noticeable "halo" around the news commentator's shoulders or hair. This is most generally caused by contamination of the foreground with bounce from the chroma-key screen. The chroma-key background screen is normally blue or green, and the foreground must *not* contain this color. If there is excessive reflection from the background into the foreground, this keying problem will occur.

Fig. 5-18 reviews the proper basic chroma-key lighting technique for "clean" key operation. To minimize the possibility of reflections,



**Fig. 5-18. Basic chroma-key lighting arrangement for "clean" keying.**

keep the foreground subject at least 10 feet in front of the chroma-key screen. Use only 75 to 100 footcandles of illumination on the screen and no more than 200 to 250 footcandles on the subject. (This assumes modern color cameras are in use. Some older cameras will require slightly more illumination.) Avoid any shadows within the pickup area of the background by using ample fill light.

## EXERCISES

- Q5-1. Why must coils and capacitors be used in a bridging-type passive video switcher?
- Q5-2. Name the two basic sectional panels of a production switcher.
- Q5-3. On a film island, what is the main control point for control interface?
- Q5-4. When delay lines are necessary as in earlier video switchers, what else is required for these lines?
- Q5-5. If 3.58-MHz cross talk is to be down 60 dB through the switcher, what should be the peak-to-peak signal measured on the scope?

## Video DA's, Processing Amplifiers, and Power Supplies

Since the introduction of all-solid-state circuitry, video distribution facilities have achieved a relatively high level of performance and stability. The cumulative effect of many units in cascade is now being reduced to that of a single unit of former years. Table 6-1 lists typical single-unit video-amplifier characteristics of 1957 as compared to those of 1977.

**Table 6-1. Typical Single-Unit Video-Amplifier Characteristics**

Measurement	1957 Value	1977 Value
Freq Response (dB)	$\pm 0.5$ dB @ 6 MHz	+0 to $-0.1$ dB @ 10 MHz
Differential Gain	3% Maximum	0.2% Maximum
Differential Phase	$1^\circ$ Maximum	$0.1^\circ$ Maximum
Vertical Tilt (Field)	2% Maximum	0.5% Maximum

Except for an occasional faulty component, the performance of individual units has become so good that the majority of system errors can be traced to secondary sources such as interconnecting cables, patch panels, and termination methods. Techniques for system measurements are covered in Chapter 8. In the present chapter, we will cover sufficient circuit analysis to provide a basic understanding of modern video distribution amplifiers (DA's), processing amplifiers, and associated regulated power supplies to assure familiarity with the functions intended. The cabling techniques necessary for interconnections are also included in this study.

NOTE: A video distribution amplifier is a straightforward amplifier intended to accept a single video signal and provide multiple outputs for distribution. Processing units include stabilizing and/or clamping amplifiers, and more complex processing units which regenerate blanking, sync, and color-burst signals. The integrated circuit (IC) operational amplifier is covered later in this chapter.

## 6-1. BASIC PARAMETERS OF SOLID-STATE AMPLIFIERS

There are certain parameters of solid-state video amplifiers that are common to all units of commercial manufacture. If the maintenance technician is sufficiently familiar with these characteristics, he is able to know almost exactly what to expect in waveform and circuit-voltage analysis, regardless of what make of amplifier he is troubleshooting.

Table 6-2 lists the symbols used in this study, and Chart 6-1 gives the basic parameters common to all circuit configurations: common emitter, common base, and common collector (emitter follower).

### The Common-Emitter Amplifier

The common-emitter circuit (Fig. 6-1) is by far the most widely used in amplifier design. The common-base circuit is most often used

Table 6-2. Symbols

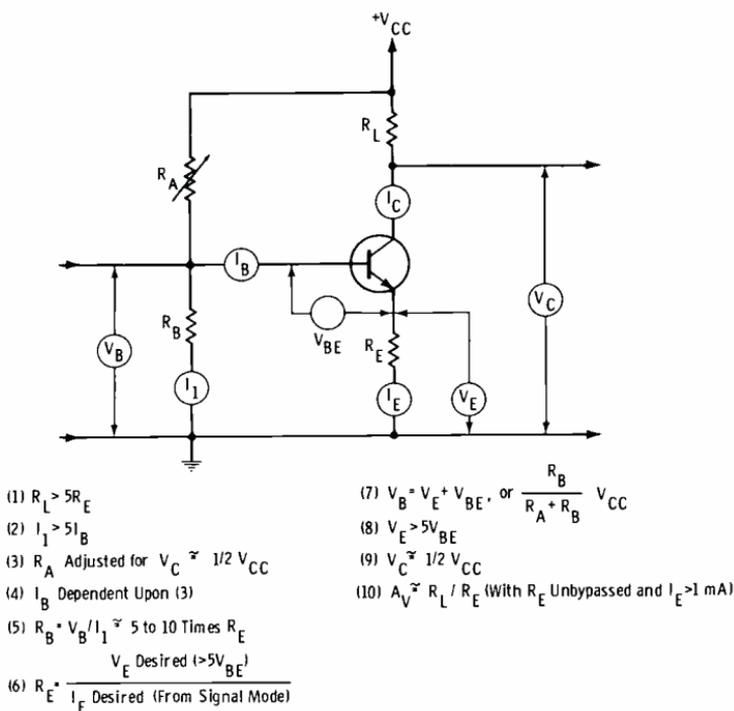
$\alpha$	(Alpha) Common-base short-circuit current gain	$I_E$	Dc emitter current
$A_I$	Current gain	$I_e$	Ac (signal) emitter current
$A_P$	Power gain	$R_E$	Emitter resistor
$A_V$	Voltage gain	$r_e$	Small-signal emitter resistance
$\beta$	(Beta) Common-emitter short-circuit current gain	$R_{EB}$	Emitter-base junction resistance (assume 4 ohms average)
$C_{cb}$	Collector-base capacitance	$R_f$	Feedback resistance
$C_{ce}$	Collector-emitter capacitance	$R_G$	Generator resistance
$C_{in}$	Input capacitance	$R_{in}$	Input resistance
$f_{\alpha fb}$	Alpha cutoff frequency (common base)	$R_L$	Load resistance
$f_{\beta fe}$	Beta cutoff frequency (common emitter)	$R_S$	Source resistance
$f_{max}$	Maximum frequency of oscillation	$r_{tr}$	Transresistance
$Ge$	Germanium	$Si$	Silicon
$g_m$	Transconductance	$V_B$	Base voltage (dc)
$h_{FE}$	Dc short-circuit current gain	$V_{RB}$	Base supply voltage
$h_{fe}$	Ac (signal) short-circuit current gain	$V_{BE}$	Base-to-emitter voltage (dc)
$I_B$	Dc base current	$V_C$	Collector voltage (dc)
$I_b$	Ac (signal) base current	$V_{CC}$	Collector supply voltage
$I_C$	Dc collector current	$V_{CE}$	Collector-to-emitter voltage (dc)
$I_c$	Ac (signal) collector current	$V_E$	Emitter voltage (dc)
$I_{CO}$	Collector leakage current (cutoff current)	$V_{EE}$	Emitter supply voltage
		$Z_g$	Generator impedance
		$Z_{in}$	Input impedance
		$Z_o$	Output impedance
		$Z_s$	Source impedance

Chart 6-1. Basic Parameters Common to all Configurations

<b>Base Current</b>	$I_B = I_E - I_C = \frac{I_C}{h_{FE}} \text{ or } \frac{I_E}{h_{FE}} - I_{C0}$
<b>Collector Current</b>	$I_C = I_E - I_B = \alpha I_E = h_{FE} I_B$
<b>Collector Power</b>	$P_C = V_{CE} I_C$
<b>Emitter Current</b>	$I_E = I_B + I_C$ (Total Current)
<b>Small-Signal Emitter Resistance</b>	$r_e = \frac{26}{I_E}$ where, $I_E$ is emitter current in mA.
<b>Transresistance</b>	$r_{tr} = r_e + R_{EB} + R_E$ where, $r_e$ is the small-signal emitter resistance in ohms, $R_{EB}$ is assumed to be 4 ohms, $R_E$ is the unbypassed external emitter resistance.
<b>Transconductance</b>	$g_m = \frac{1}{r_{tr}} = \frac{I_E}{26}$ where, $I_E$ is emitter current in mA.
<b>Bandwidth</b>	$f_{hre} = \frac{f_{hrb}}{h_{re}}$ , or $f_{hrb} = h_{re} f_{hre}$ where, $f_{hre}$ is the beta cutoff frequency (3-dB point), $f_{hrb}$ is the alpha cutoff frequency (3-dB point), $h_{re}$ is $\beta = \frac{\alpha}{1 - \alpha}$
<b>Upper Frequency Limit</b>	$f_u = \frac{g_m}{6.28 C_t}$ where, $f_u$ is upper frequency limit (unity gain) in MHz, $g_m$ is transconductance in micromhos, $C_t$ is total capacitance in pF.
<b>Input Capacitance</b>	$C_{in} = \frac{g_m}{6.28 f_{hrb}}$

to match a low impedance to a high impedance, and the common-collector (emitter-follower) circuit is most often used to match a high impedance to a low impedance. We will study the latter two circuit configurations a little later.

The design parameters for Fig. 6-1 apply to a stage intended for linear amplification, as revealed by the fact that the quiescent collector voltage ( $V_C$ ) is fixed approximately midway between ground



**Fig. 6-1. Design parameters for common-emitter amplifier.**

and full collector supply voltage ( $V_{CC}$ ). Some form of stabilizing feedback normally is provided, and we will examine this basic characteristic now.

**NOTE:** The reader should have sufficient solid-state background to be able to grasp the following basic analysis. If not, he should undertake a separate study, as presented (for example) in Harold E. Ennes, *Workshop in Solid State*, 2nd ed. (Indianapolis: Howard W. Sams & Co., Inc., 1977).

### Emitter Feedback

Emitter feedback is, very simply, the result of the unbypassed emitter resistor,  $R_E$ . Since the signal induced in useful collector load  $R_L$  must pass through  $R_E$ , the signal is degenerated. This type of feedback is a form known as "series feedback," and it increases the input impedance. Before we go further, it is important to examine in more detail this most common form of feedback stabilization.

First, let us put down the applicable relationships for the circuit of Fig. 6-1 for further reference:

$$V_{BE} = 0.2 \text{ to } 0.3 \text{ V for germanium transistors} \quad (\text{Eq. 6-1.})$$

$$V_{BE} = 0.6 \text{ to } 0.7 \text{ V for silicon transistors} \quad (\text{Eq. 6-2.})$$

$$I_B = \frac{I_E}{h_{FE}} - I_{CO} \quad (\text{Eq. 6-3.})$$

$$I_C = \alpha I_E \quad (\text{Eq. 6-4.})$$

$$I_C = I_E - I_B \quad (\text{Eq. 6-5.})$$

$$I_C = h_{FE} I_B \quad (\text{Eq. 6-6.})$$

$$Z_{in} = h_{FE} r_{tr} \quad (\text{Eq. 6-7.})$$

$$Z_{load} = R_L \text{ in parallel with following-stage input } Z \quad (\text{Eq. 6-8.})$$

$$A_i = h_{FE} \quad (\text{Eq. 6-9.})$$

$$A_v = \frac{R_L}{r_{tr}} \quad (\text{Eq. 6-10.})$$

$$A_v = g_m R_L \quad (\text{Eq. 6-11.})$$

$$A_p = (A_i)^2 \frac{R_L}{\beta r_{tr}} = A_i A_v = \beta \frac{R_L}{r_{tr}} \quad (\text{Eq. 6-12.})$$

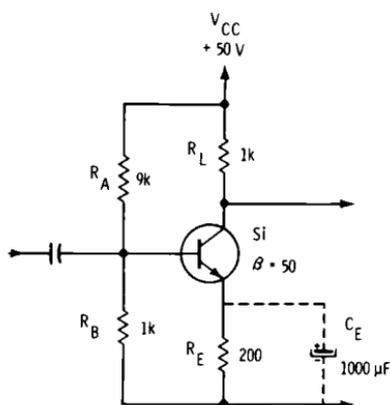
Now study the ten points listed on Fig. 6-1. They show how the design engineer arrives at stable operating parameters for a linear (class-A) amplifier.

In (6), it is stated that the emitter current desired is derived from the intended operating mode. For example, a low-level stage (such as an audio preamplifier) must be operated at very low emitter current (and therefore low base and collector currents) in the interest of low noise level. As the signal level is brought higher and higher in amplitude, more and more dc current must be employed to accommodate the higher peak-to-peak signal swings.

The main problem in stabilization is to control the current gain ( $\beta$ , or  $h_{FE}$ ) of the common-emitter circuit. The actual value of  $\beta$  differs with different transistors of the same type, and it also varies with temperature changes. Consequently, practical circuit design requires a *controlled beta* (controlled current gain) so that this parameter will remain fixed under varying operating temperatures and with necessary transistor replacements. Thus, if a certain transistor type has a "minimum  $h_{FE}$ " of 20, the circuit is designed to limit beta to no more than 20 by means of fixed resistance ratios.

We can now examine a practical single-stage common-emitter linear amplifier and see how such a design is made. See Fig. 6-2. Let us analyze this circuit for practice in determining what we should expect in dc voltage measurements, and to see how it meets the design requirements of Fig. 6-1.

Step 1 is to determine the base voltage,  $V_B$ . We can do this by figuring the current in  $R_B$ , or we can use the second relationship of item (7) in Fig. 6-1. In either case, we get  $V_B = +5$  volts.



**Fig. 6-2. Class-A common-emitter amplifier circuit.**

Then, since the transistor is a silicon (Si) type:

$$V_E = +5 - 0.6 = +4.4 \text{ volts}$$

(Note that this value is greater than 5 times  $V_{BE}$ , as called for in Fig. 6-1.)

Since the emitter voltage is +4.4 volts, the emitter current (from Ohm's law) must be:

$$I_E = \frac{4.4}{200} = 22 \text{ mA}$$

For rapid analysis, let us assume the same current in the collector as in the emitter:

$$I_C = 22 \text{ mA}$$

Then the dc voltage drop across  $R_L$  is:

$$V_{RL} = (0.022) (1000) = 22 \text{ volts}$$

and:

$$V_C = 50 - 22 = +28 \text{ volts}$$

(Note that this value is close to  $\frac{1}{2} V_{CC}$ .)

For the given transistor,  $h_{FE} = \beta = 50$ . Then, from Eq. 6-3 (ignoring  $I_{CO}$ ):

$$I_B = \frac{0.022}{50} = 0.44 \text{ mA}$$

Note that current  $I_1$  is 5 mA, and therefore meets the requirement of being more than 5 times the base current ( $I_B$ ).

With  $R_E$  bypassed ( $C_E$  connected):

$$A_i = h_{FE} = 50 \text{ (from specification sheets)}$$

$$r_{tr} = \frac{26}{22} + 4 = 5 \text{ (approx)}$$

$$A_v = \frac{R_L}{r_{tr}} = \frac{1000}{5} = 200$$

$$Z_{in} = (50)(5) = 250 \text{ ohms (Eq. 6-7.)}$$

Note that since  $R_E$  is bypassed for the signal, the signal is not degenerated. Therefore, the full value of beta is in effect, with the exception of the slight shunting action of bias resistors  $R_A$  and  $R_B$ .

Now we will consider the case in which  $R_E$  is not bypassed by  $C_E$ . The first important change to note is the effect on input impedance  $Z_{in}$  of the transistor alone. When  $R_E$  was bypassed,  $Z_{in}$  was 250 ohms. But when  $R_E$  is not bypassed,  $r_{tr}$  becomes 205 ohms, or approximately 200 ohms. Then:

$$Z_{in} = (50)(200) = 10,000 \text{ ohms}$$

It is evident that the value of  $R_B$  essentially fixes the value of the new input impedance. Since  $R_B$  is only one-tenth of the transistor input impedance, the current gain will be reduced drastically. Let us put this relationship down on record so that we can remember it and use it in practical form.

For practical circuit analysis, the following relationship of currents and resistances is sufficiently accurate:

$$\frac{I_c}{I_{in}} = \frac{R_B}{R_E}$$

This equation says that the emitter current is to the input current as base resistor  $R_B$  is to emitter resistor  $R_E$ . From this basic relationship we can derive a useful equation for quickly evaluating current gain in a degenerative amplifier. Since collector current (for rapid estimates) can be considered to be the same as emitter current, then substituting  $I_c$  (collector current) for  $I_e$  (emitter current) in the above basic equation:

$$\frac{I_c}{I_{in}} = \frac{R_B}{R_E}$$

Note that the ratio of  $I_c$  (collector current) to  $I_{in}$  (input current) is an expression for current gain ( $A_i$ ). So:

$$A_i = \frac{R_B}{R_E} \text{ (for an unbypassed } R_E \text{)} \quad \text{(Eq. 6-13.)}$$

Note also that when an unbypassed  $R_E$  is much greater than  $r_e + R_{EB}$ , voltage amplification ( $A_v$ ) can be expressed:

$$A_v = \frac{R_L}{R_E} \text{ (for an unbypassed } R_E \text{)} \quad \text{(Eq. 6-14.)}$$

This equation is valid so long as  $r_e + R_{EB}$  is very small in value compared to  $R_E$  and can be ignored.

Based on the above equations, the new current gain with  $R_E$  unbypassed is:

$$A_i = \frac{R_B}{R_E} = \frac{1000}{200} = 5$$

and the new voltage gain is:

$$A_v = \frac{R_L}{R_E} = \frac{1000}{200} = 5$$

### Other Basic Negative-Feedback Circuits

Although emitter degeneration (series feedback) is almost universally used, other forms of degenerative feedback, which may or may not be used along with emitter degeneration, are often encountered. We will explore briefly the fundamentals of such circuitry here, and then expand on them as we encounter typical circuits in later discussion.

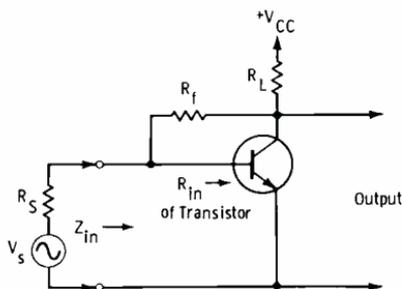


Fig. 6-3. Circuit with shunt feedback.

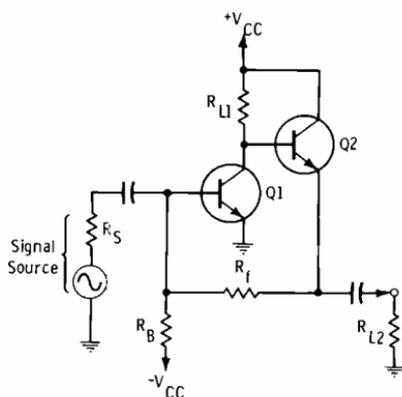
The basic arrangement for collector-base feedback (termed “shunt feedback”) is shown in Fig. 6-3. For most practical circuits, the resistance of  $R_f$  is small enough that we can use these approximate gain formulas for quick-analysis purposes:

$$A_i = \frac{R_f}{R_L} \quad (\text{Eq. 6-15.})$$

$$A_v = \frac{R_f}{Z_{in}} \quad (\text{Eq. 6-16.})$$

The quantity  $Z_{in}$ , the effective input impedance of the stage, is approximately:

$$Z_{in} = \frac{R_L + R_f}{\beta R_L + R_f} R_{in} \quad (\text{Eq. 6-17.})$$

**Fig. 6-4. Diagram of a feedback pair.**

where  $R_{in}$  is the input impedance without feedback.

Note that shunt feedback *decreases* the effective  $Z_{in}$ , whereas series feedback (such as that provided by an unbypassed emitter resistor) *increases* the effective  $Z_{in}$ .

NOTE: The gain formulas given above are adequate for quick analysis of most circuits because  $R_f$  is usually relatively low in value to meet the requirement of "adequate feedback." Somewhat more nearly exact relationships are as follows:

$$A_i = \frac{R_f}{R_f + \beta R_L} \beta$$

$$A_v = \frac{R_f}{R_L + R_f} \beta \frac{R_L}{R_{in}}$$

Fig. 6-4 shows a "feedback pair." This circuit is gaining universal acceptance as a stable, wideband amplifier with inherent ac and dc stability. This type of circuit has low input and output impedance. For quick analysis, its voltage gain can be calculated from the expression:

$$A_v = \frac{R_f}{R_s} \quad (\text{Eq. 6-18.})$$

where  $R_s$  is the output resistance of the signal source. This formula is sufficiently accurate for use in normal applications of the circuit with low input and output impedances. The voltage gain is almost universally held to a value less than 10.

### Feedback Gain-Impedance Relationships

We should be certain of a "sharp focus" on impedance-to-gain relationships. We will see why series feedback affects voltage gain much more than current gain, and why shunt feedback affects current gain much more than voltage gain. These facts always bring up the question, "If you cut current gain, how can voltage gain

remain the same?" Or, "If current gain is not affected, why does voltage gain decrease?" The following discussion should answer these questions.

See Fig. 6-5A. This circuit is without negative feedback. Assume that:

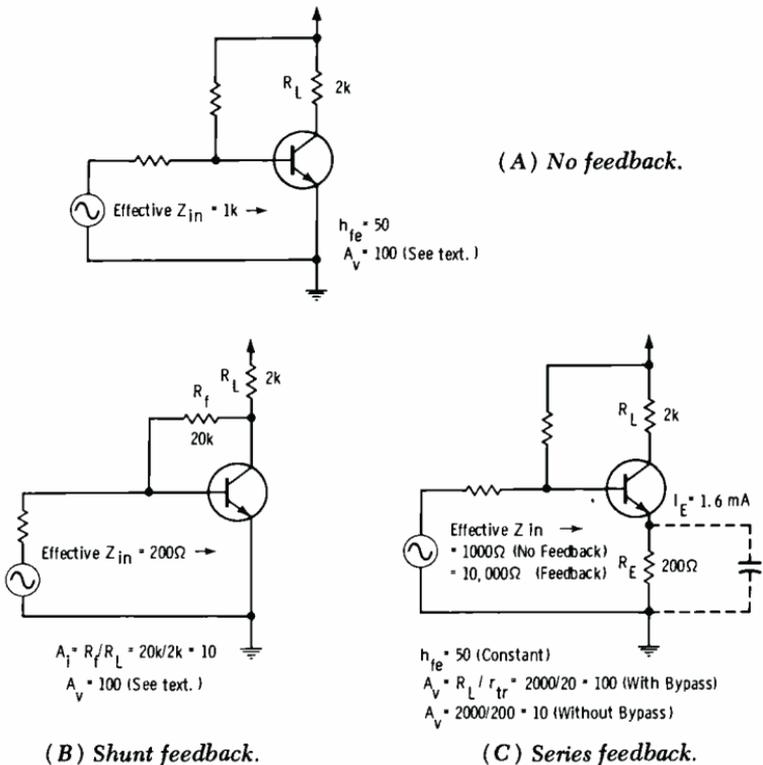


Fig. 6-5. Comparison of shunt and series feedback.

$$A_i = h_{fe} = 50$$

$$\text{Input signal current} = 10 \mu\text{A}$$

$$\text{Effective } Z_{in} = 1000 \text{ ohms (resistive)}$$

$$R_L = 2000 \text{ ohms}$$

Then:

$$\text{Signal current in } R_L = (50) (10 \mu\text{A}) = 500 \mu\text{A}.$$

$$\text{Signal voltage in } Z_{in} = IR = (10 \mu\text{A}) (1000) = 0.01 \text{ volt}$$

$$\text{Signal voltage in } R_L = (500 \mu\text{A}) (2000) = 1 \text{ volt}$$

$$A_v = \frac{V_{out}}{V_{in}} = \frac{1}{0.01} = 100$$

Now see Fig. 6-5B. Shunt feedback is used, and:

$A_i$  is cut to 10 (from 50).

Signal input current is maintained at  $10 \mu\text{A}$  (for reference).

$R_{in}$  is reduced (by shunt feedback) from 1000 ohms to about 200 ohms (from Eq. 6-17).

Now:

Signal current in  $R_L = (10) (10 \mu\text{A}) = 100 \mu\text{A} = 0.1 \text{ mA}$

Signal voltage in  $R_L = (0.1 \text{ mA}) (2000 \Omega) = 0.2 \text{ V} = 200 \text{ mV}$

Signal voltage in  $R_{in} = (10 \mu\text{A}) (200 \Omega) = 0.002 \text{ V} = 2 \text{ mV}$

$$A_v = \frac{200 \text{ mV}}{2 \text{ mV}} = 100 \text{ (the same as for the circuit of Fig. 6-5A)}$$

Note that although the output signal voltage was reduced in amplitude for the same signal input current, the voltage *amplification* remained the same for the circuit of Fig. 6-5B as for that of Fig. 6-5A. The current amplification is limited to 10 (which will be below the minimum beta of the transistor used) so that the voltage output is stabilized for changes in  $h_{re}$ .

As a matter of interest, the voltage gain calculated from the more complete equation given previously is:

$$A_v = \frac{20\text{k}}{2\text{k} + 20\text{k}} (50) \frac{2\text{k}}{1\text{k}} = 90$$

The "quick-analysis" value is within about 10 percent of this result.

Fig. 6-5C shows the series-feedback arrangement of the common-emitter circuit. Whether bypassed or not,  $R_E$  has no effect on the current-gain parameter of the transistor. But we already know that it drastically affects the voltage gain of the circuit.

Assume that  $R_E$  is bypassed and that the signal input is 10 microamperes. Then:

Signal current in  $R_L = (50) (10 \mu\text{A}) = 500 \mu\text{A}$

Signal voltage in  $R_L = (500 \mu\text{A}) (2000 \Omega) = 1 \text{ volt}$

Signal voltage in  $R_{in} = 10 \mu\text{A} (h_{re}r_{tr}) = 10 \mu\text{A} (50) \left( \frac{26}{1.6} + 4 \right)$

$= 10 \mu\text{A} (1000 \Omega) = 0.01 \text{ volt}$

$$A_v = \frac{1}{0.01} = 100$$

Now assume that  $R_E$  is not bypassed (series feedback to signal):

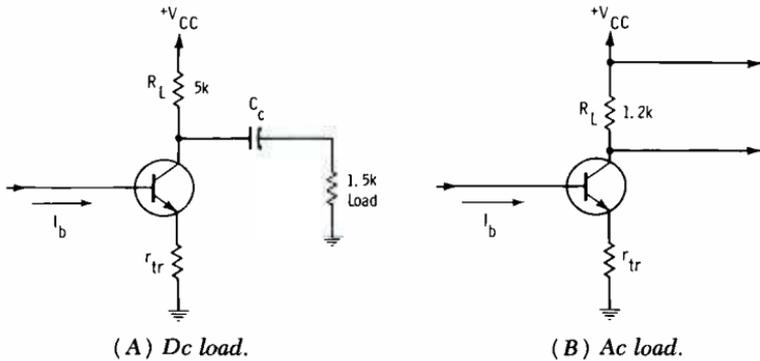
$R_{in} = 50 (20 + 200) = 10,000 \text{ ohms (approx)}$

$V_{in} = (10 \mu\text{A}) (10,000 \Omega) = 0.1 \text{ volt}$

Signal current in  $R_L = 50 (10 \mu\text{A}) = 500 \mu\text{A}$  ( $h_{re}$  not changed)

Signal voltage in  $R_L = (500 \mu\text{A}) (2000 \Omega) = 1 \text{ volt}$

$$A_v = \frac{1}{0.1} = 10 \text{ (same as } \frac{R_L}{R_E} = \frac{2000}{200} = 10 \text{)}$$



**Fig. 6-6. Circuit equivalents for ac and dc gain.**

The quantity  $h_{FE}$  is the dc beta (common emitter), and  $h_{fe}$  is the ac, or signal, beta. This difference can be emphasized by studying Fig. 6-6.

The circuit of Fig. 6-6A is equivalent to a single stage that has a 5k collector load and is feeding a following circuit with a 1.5k input impedance. At low frequencies, the load resistance is effectively 5000 ohms. Assume  $h_{FE} = h_{fe} = 50$ . The output voltage swing for a base-current swing of (for example) 3 microamperes is:

$$V = (3 \mu A) (50) (5000 \Omega) = 750 \text{ mV}$$

At high frequencies (for which the reactance of  $C_c$  can be neglected), the effective load becomes 5k in parallel with 1.5k, or about 1.2k (Fig. 6-6B). Now the output-voltage swing for a 3-microampere base-current swing is:

$$V = (3 \mu A) (50) (1200 \Omega) = 180 \text{ mV}$$

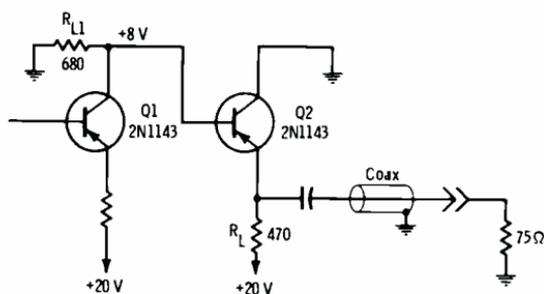
This example simply emphasizes that with  $h_{fe}$  equal to  $h_{FE}$ , the actual voltage swing is much less for  $h_{fe}$  due to the difference in the load the transistor sees.

### The Common-Collector (Emitter-Follower) Circuit

The emitter follower is used extensively in video circuits, primarily to drive 75-ohm distribution lines. It is also common in other applications in which a low output impedance is desired.

Fig. 6-7 shows an output stage driving a 75-ohm coaxial line. The 8 volts on the base is developed across the collector load of Q1. The emitter measures 8.5 volts, so the voltage drop across  $R_L$  is  $20 - 8.5 = 11.5$  volts. Therefore:

$$I_E = \frac{11.5}{470} = 24 \text{ mA}$$



**Fig. 6-7. Typical emitter follower.**

The transistor has a typical  $\beta$  of 50; therefore:

$$I_B = \frac{I_E}{h_{FE}} = \frac{0.024}{50} = 480 \mu\text{A}$$

We can see (Fig. 6-8A) that a current gain (480  $\mu\text{A}$  of base current to nearly 24 mA of collector current) occurs in the emitter follower just as in the common-emitter amplifier. But the *voltage* gain is approximately unity. The output signal is of the same amplitude as the input signal, less a small voltage drop across the base-emitter junction. The voltage gain of this circuit can be explained as follows.

The effective  $R_L$  (the load presented to the signal voltage) in this case is the 75-ohm line termination in parallel with 470 ohms, for an effective  $R_L$  of 65 ohms. The transresistance ( $r_{tr}$ ) is:

$$r_{tr} = \frac{26}{I_E} + R_{EB} + R_L = 1 \text{ (approx)} + 4 + 65 = 70 \text{ ohms}$$

The value of  $g_m$  is  $1/r_{tr}$ , or 0.014. Then:

$$A_v = g_m R_L = (0.014) (70) = 0.98$$

Or, we can say that, since  $R_L$  is much greater than  $r_e + R_{EB}$ :

$$A_v = \frac{R_L}{R_E} \text{ (approx)}$$

In this case,  $R_E = R_L$ , so:

$$A_v = \frac{R_L}{R_L} = 1 \text{ (approx)}$$

In practice, there are many cases in which we cannot measure any amplitude difference between the signals at the base and emitter of a common-collector stage. For this analysis, we can delete  $R_{EB}$  from the  $g_m$  computation. The gain may normally be assumed to be unity for quick circuit analysis. Note also that the dc parameters were

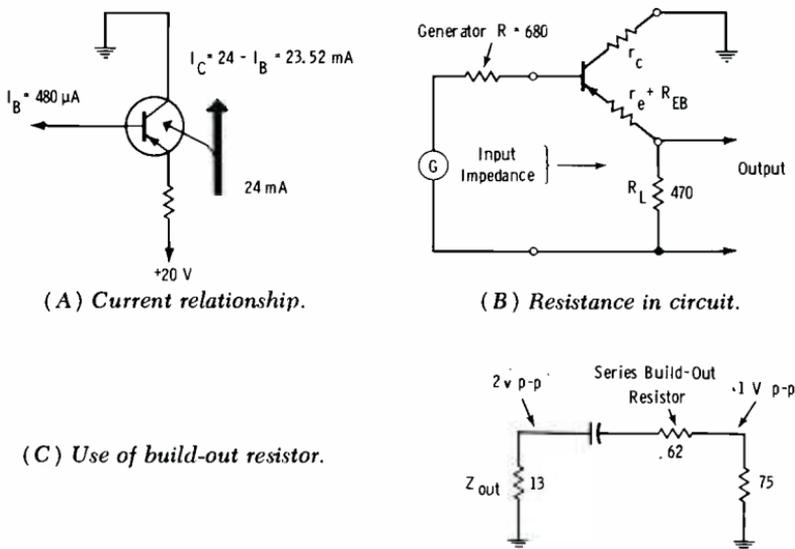


Fig. 6-8. Emitter-follower stage.

used in the above analysis, so we are considering only low frequencies (and dc) at the present.

Study Fig. 6-8B, and observe that  $r_c$  appears in parallel with the sum of  $r_e$ ,  $R_{EB}$ , and  $R_L$ . Since  $r_c$  can be assumed to be 1.5 megohms or more, it need not be considered in quick circuit analysis. What is important is to note that the actual input impedance and the output impedance are somewhat interdependent. This is true because the input load is part of the output, and the output load is part of the input. The high intrinsic value of  $r_c$  does not serve as isolation between input and output as it does in the other circuit configurations; as Fig. 6-8B shows, it now forms a parallel resistance path only.

Since the emitter follower is used primarily as an impedance transformer, the input and output impedances are most important characteristics to know. While conventional treatments result in highly complex formulas, we can use the following rules of thumb that are fairly accurate in practice:

$$Z_{in} = (\beta + 1) R_L \quad (\text{Eq. 6-19.})$$

$$Z_{out} = \frac{R_s}{\beta + 1} \quad (\text{Eq. 6-20.})$$

The input resistance of the stage in Fig. 6-8B (assuming the stage feeds a 75-ohm termination) is:

$$Z_{in} = (50 + 1) 65 = 3300 \text{ ohms (approx)}$$

This resistance has a reasonably small loading effect on the 680-ohm collector load of the preceding stage.

The output impedance is

$$Z_{out} = \frac{680}{50 + 1} = 13 \text{ ohms (approx)}$$

The internal output impedance (sending-end impedance) of an emitter follower can be made quite low, sufficiently low to drive 2-ohm switching buses directly. This low impedance also is the reason why we will usually find a series "build-out" resistor (Fig. 6-8C). For the receiving end (75 ohms) to "see" a 75-ohm impedance, the 13-ohm output would require a series resistance of 62 ohms. Note also that this arrangement forms a voltage divider, and a 2-volt signal across  $R_L$  is dropped to about 1 volt across the 75 ohms.

In practice, as the effective load impedance increases, the input impedance increases. As the generator impedance increases, the output impedance increases. Providing the generator impedance (previous collector  $R_L$ ) is not extremely low (so long as it is greater than 10 times  $R_L$ , as is usually the case), the formulas given above will prove sufficiently accurate for practical purposes.

Figs. 6-9 through 6-12 illustrate the basic applications of the common-collector configuration. In Fig. 6-9, the output impedance of Q1 is essentially the collector load of 1000 ohms. For class-A operation, we expect a quiescent collector voltage of  $\frac{1}{2} V_{CC}$ , or -5 volts. So assuming germanium transistors with a beta of 50, the common-collector line driver, Q2, has:

$$R_L = \frac{470 (75)}{470 + 75} = 65 \text{ ohms (effective load)}$$

$$Z_{out} = \frac{1000}{50 + 1} = 20 \text{ ohms (approx)}$$

$$Z_{in} = (50 + 1) 65 = 3300 \text{ ohms (approx)}$$

Now study Fig. 6-10A. In this case, the emitter-follower line driver is *driven* by an emitter follower. The 1k load is now in the emitter circuit of Q1 rather than the collector circuit. Let us see what happens. (Assume Q1 sees a source resistance of 1000 ohms, and  $\beta$  is 50 for both transistors.)

The output impedance of Q1 is:

$$Z_{out} = \frac{1000}{50 + 1} = 20 \text{ ohms (approx)}$$

Then the output impedance of Q2 is:

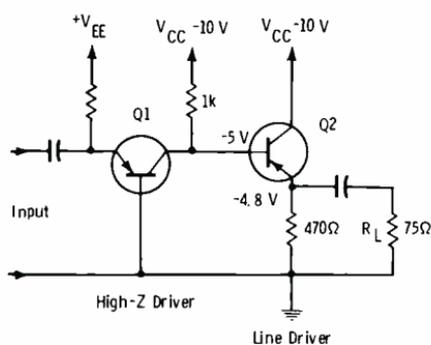
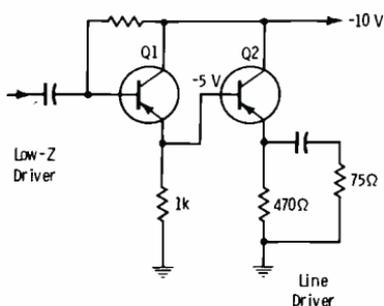
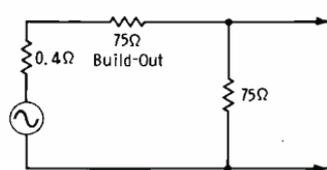


Fig. 6-9. Emitter-follower line driver.



(A) Schematic diagram.



(B) Equivalent of output stage.

Fig. 6-10. Cascaded emitter followers.

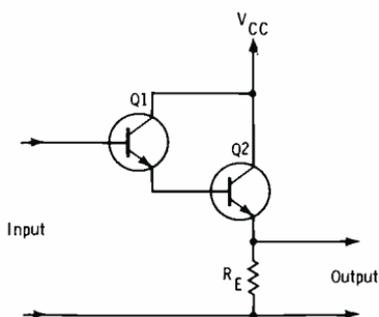


Fig. 6-11. Darlington circuit.

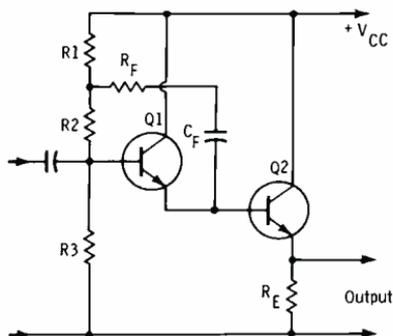


Fig. 6-12. Circuit with bootstrapping.

$$Z_{out} = \frac{20}{50 + 1} = 0.4 \text{ ohm (approx)}$$

The equivalent circuit of the output stage is shown in Fig. 6-10B. Note that to feed a 75-ohm load, a build-out resistance of 75 ohms is needed so that the effective internal output impedance becomes 75 ohms. The build-out and load combine to form a 2:1 voltage di-

vider. We must therefore expect to find the signal voltage at the emitter of Q2 to be twice as great as the signal voltage across the 75-ohm load. This condition is normal. For a 1-volt peak-to-peak signal to appear across the 75-ohm load, a 2-volt peak-to-peak signal must be available at the Q2 emitter.

In solid-state circuit applications requiring a very high input impedance, the emitter follower is quite naturally chosen. The Darlington emitter follower shown in Fig. 6-11 consists of two current amplifiers in cascade to provide large current gain and very high input impedance. The input impedance is raised due to the effective product of the betas ( $h_{re}$ ) of both transistors. This is to say that, since the base currents have been reduced by the product of the current gains, the input impedance is raised accordingly.

The input impedance of the Darlington current multiplier is limited by the shunting effect of the Q1 collector resistance and capacitance. A technique for reducing this shunting effect is to employ positive shunt feedback, termed "bootstrapping." See Fig. 6-12. The value of  $R_F$  is such as to produce at the junction of R1 and R2 a voltage which is in phase with the input voltage. The signal-voltage drop across the bias resistors is thereby reduced, which is equivalent to raising the resistance of the input network. We will find bootstrapping used in other than Darlington circuits for the same reason—to reduce the effective loading of the input biasing networks.

### Bandwidth

Alpha (or  $h_{fb}$ ) and beta (or  $h_{re}$ ) are specified at some low frequency (normally 1 kHz) or, in the case of power transistors primarily for power-supply service, in dc values ( $h_{FB}$  and  $h_{FE}$ ). But as the frequency of operation is increased, the forward current gain decreases. This effect is in common with electron-tube circuitry, and occurs for the same reason: input, output, interelement, and circuit capacitance.

The cutoff frequency of a transistor is that frequency at which the value of alpha or beta drops to 0.707 (3 dB) of its one-kilohertz value. Fig. 6-13 shows typical curves of alpha and beta ( $h_{fb}$  and  $h_{re}$ ) as functions of frequency for a high-frequency transistor.

Note that in any such curves the forward current gain of the common-emitter circuit ( $h_{re}$ ) falls 3 dB from its low-frequency value at a much lower frequency than  $h_{fb}$  does. Why? See Fig. 6-14. Fig. 6-14A shows the common-base circuit with the total input capacitance indicated by dash lines. The input resistance of the common-base circuit is relatively low, in this case ( $I_E = 1$  mA), it is 30 ohms. Now study Fig. 6-14B for the same transistor in a common-emitter circuit. The input resistance in this case is 3000

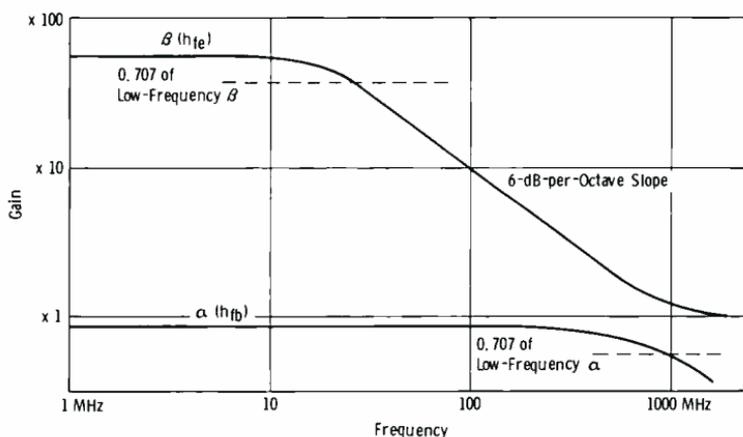


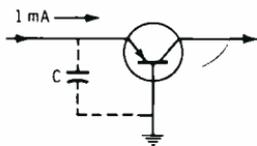
Fig. 6-13. Typical effects of frequency on alpha and beta.

ohms. The same value of capacitance now has a much greater shunting effect; effectively the input capacitance is magnified by the factor  $h_{fe}$ . The same kind of problem exists with respect to the plate-load value in a vacuum-tube circuit: the higher  $R_L$  is made (for more gain), the greater is the loss of high-frequency response because of output capacitance across the plate load. Degeneration in a common-emitter circuit can be used to increase the bandwidth by greatly reducing the intrinsic transistor capacitance (discussed shortly).

NOTE: The actual multiplication factor is  $\beta + 1$ , and we will find this value used in other transistor studies. We have dropped the 1 here since practical circuit analysis normally does not call for such precise values.

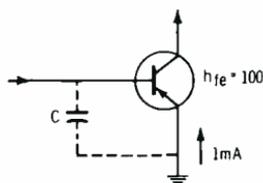
So now we understand why the beta cutoff frequency in Fig. 6-13 is much lower than the alpha cutoff frequency. Here is the basic relationship:

$$f_{hfe} = \frac{f_{hfb}}{h_{fe}}$$



$$R_{in} = \frac{26}{IE} + R_{EB} = 26 + 4 = 30 \text{ Ohms}$$

(A) Common-base circuit.



$$R_{in} = (h_{fe})^2 (r_{tr}) = (100)^2 (30) = 3000 \text{ Ohms}$$

(B) Common-emitter circuit.

Fig. 6-14. Capacitance shunting input resistance.

where,

$f_{h_{fe}}$  is the beta cutoff frequency,

$f_{h_{fb}}$  is the alpha cutoff frequency (normally given in specification sheets)

This formula says that the beta cutoff frequency (point where response is 3 dB down from low-frequency value) is equal to the ratio of the alpha cutoff frequency to the  $h_{fe}$  of the transistor. It relates gain to frequency response (in the common-emitter circuit), just as in tube-type amplifiers.

The gain-bandwidth product ( $f_r$ ) is that frequency at which  $h_{fe}$  goes to unity gain. Beyond the  $\beta$  cutoff frequency,  $\beta$  decreases at approximately 6 dB/octave. (If we double the frequency, we have increased the frequency by one octave. If  $\beta$  decreases 6 dB/octave,  $\beta$  is down 6 dB each time the frequency is doubled.) In this frequency range,  $f_r$  is the product of a given frequency and the measured  $h_{fe}$  at the given frequency. In the example of Fig. 6-13, unity gain (no amplification) occurs at approximately the frequency at which the common-base curve is down 3 dB (alpha cutoff).

Here is a rule of thumb we can use: The gain-bandwidth product (common-emitter circuit) is approximately equal to one-half alpha cutoff without peaking. It is approximately equal to alpha cutoff with double peaking. This approximation takes care of stray and circuit capacitance in practical amplifiers. Actually, we will find practically no wideband amplifiers (for example, in video service) that employ peaking circuits, since modern transistors designed for this type of service have an extremely high value of gain-bandwidth. About the only place we will find compensation circuits (in video applications) is in such units as camera preamplifiers, where pickup-tube coupling circuits require frequency-phase correction. Obviously, amplifiers designed for processing signals (aperture boost, phase correction, etc.) are exceptions. (Some special high-frequency transistors will do better than the one-half alpha rating.)

Let us be sure of the practical application of the gain-bandwidth (GB) relationship. If we have a transistor with an alpha cutoff listed as 500 MHz and we use the rule of thumb mentioned:

$$GB = 250 \text{ MHz (no peaking)}$$

Then the gain for a bandwidth of 10 MHz is:

$$G = \frac{250 \text{ MHz}}{10 \text{ MHz}} = 25$$

This result indicates that the maximum gain we can expect for a bandwidth of 10 MHz is 25 times. The bandwidth we would obtain at a gain of 50 is:

$$B = \frac{250}{50} = 5 \text{ MHz}$$

This result illustrates the familiar rule of thumb that we can halve the gain to double the bandwidth, or double the gain to halve the bandwidth.

Fig. 6-15A shows interelement capacitances of a transistor. Recall that the "width" of the barrier of the pn junction is influenced by the voltage across the junction. Since an electrostatic field is thus generated, we have the equivalent of plates of a capacitor; the spacing between the "plates" is governed by the voltage (and current) of the junction.

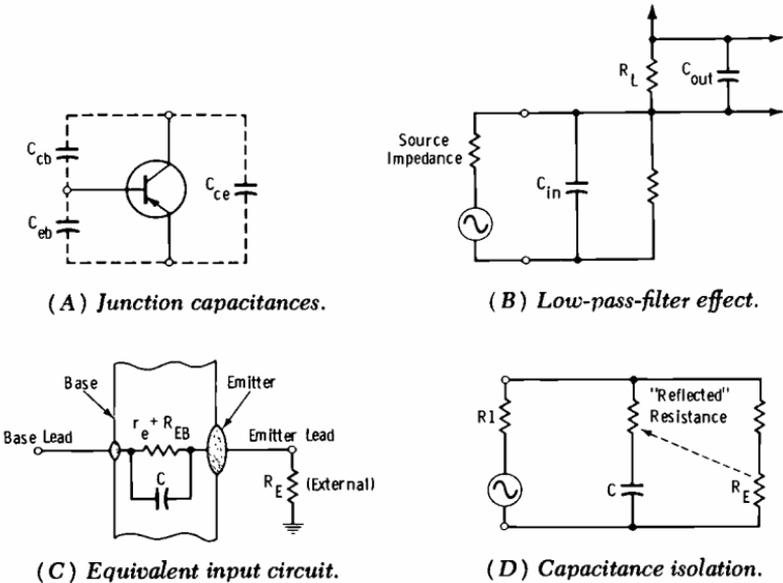


Fig. 6-15. Transistor capacitance effects.

Anything that increases the width of the pn-junction barrier has the effect of spreading apart the plates of the capacitor, resulting in less capacitance. As reverse bias is increased, (greater barrier width), the capacitance decreases. As the reverse bias is decreased, junction capacitance increases. For example, decreasing the collector-to-base voltage ( $V_{CB}$ ) causes  $C_{cb}$  to increase. Also, increased emitter current, most of which is injected into the collector through the base-to-collector junction, increases  $C_{cb}$ . The increased pn-junction current in effect reduces the pn-junction width.

The value of  $C_{cb}$  varies from about 1.5 pF in high-frequency transistors to around 50 pF in audio transistors. The collector-emitter

capacitance ( $C_{ce}$ ) is 5 to 10 times greater than  $C_{cb}$  and is also influenced by emitter current and collector voltage. The emitter-base capacitance ( $C_{eb}$ ), since the pn-junction width is small (forward bias), is higher than  $C_{ce}$ , but its effect is less because the input-generator impedance is normally much less than the high internal collector resistance.

Low-frequency transistors give poorer bandwidth than "high-frequency" transistors (low interelement capacitances) can give. A higher  $R_L$  results in less bandwidth because of the effects of output capacitance. Higher currents decrease bandwidth. Capacitance  $C_{cb}$  forms a small feedback capacitance which causes oscillation at high (rf) frequencies and must be neutralized. (The latter effect normally is not of concern in the case of video amplifiers.)

Always bear in mind the high-frequency parameters we have already studied. We know that if a common-emitter circuit results in a half-power frequency of (for example) 1 MHz, a common-base circuit (same transistor) could give a bandwidth beta times as great ( $f_{hfe} = f_{hcb}/\beta$ ). Assuming  $\beta$  is 50, the bandwidth of the circuit would become 50 MHz. But the input impedance is quite low (reducing the gain of the previous stage by radically decreasing its effective  $R_L$ ), and the current gain is less than one. This is another example of the gain-bandwidth problem.

The basic problem of bandwidth is illustrated in Fig. 6-15B; designers must do what is possible to minimize the effects of the low-pass filters formed by interelement and circuit capacitances. As the source impedance is made higher, the total input capacitance has a greater effect on bandwidth. Also, as  $R_L$  is made higher, the effect is the same as if the total output capacitance is increased, and the bandwidth is decreased. Thus  $R_L$  (which becomes the source impedance for the following stage) must be a compromise between the gain and bandwidth required.

Another effect of  $R_L$  is the "Miller effect": collector-to-base capacitance is multiplied by a factor of  $1 + \text{gain}$ . For example, if  $C_{cb}$  is 6 pF and the stage gain is 10, an effective  $C_{cb}$  of approximately 60 pF appears across the input. So the  $C_{in}$  shown in Fig. 6-15B is the sum of  $C_{eb}$  and the effective  $C_{cb}$ , plus, of course, component and wiring capacitances.

Now study Fig. 6-15C. This diagram gives a highly simplified equivalent to illustrate intrinsic input capacitance across the base-emitter junction. Note that an unbypassed external emitter resistor ( $R_E$ ) appears in series with the input capacitance and reduces the shunting effect (to higher frequencies) of the capacitance.

An unbypassed emitter resistor ( $R_E$ ) decreases the effective input capacitance. A "reflection" of  $R_E$  appears in series with the shunt capacitance as shown in Fig. 6-15D. Input capacitance can be re-

lated to the transconductance ( $g_m$ ) and alpha cutoff frequency ( $f_{\alpha}$ ) as follows:

$$C_{in} = \frac{g_m}{(6.28)(f_{\alpha})}$$

Assume a stage without any form of degeneration has a transconductance of 0.035 mho, and that the alpha cutoff (3dB) is 100 MHz:

$$C_{in} = \frac{0.035}{(6.28)(10^8)} = 56 \text{ pF}$$

Assume an  $R_E$  of 100 ohms (unbypassed) is added. Now we have an emitter factor (call it  $K_e$ ) which is:

$$K_e = \frac{g_m R_E}{1 + (g_m R_E)} = \frac{(0.035)(100)}{1 + (0.035)(100)} = \frac{3.5}{4.5} = 0.8 \text{ (almost)}$$

Since  $K_e$  is a degenerative figure,  $1 - K_e$  times the former  $C_{in}$  is the new input capacitance:  $(1 - 0.8)(56) = (0.2)(56) = 11.2 \text{ pF}$ . Bear in mind, however, that although the effective input capacitance has been reduced for greater bandwidth, the gain of the stage has been reduced also.

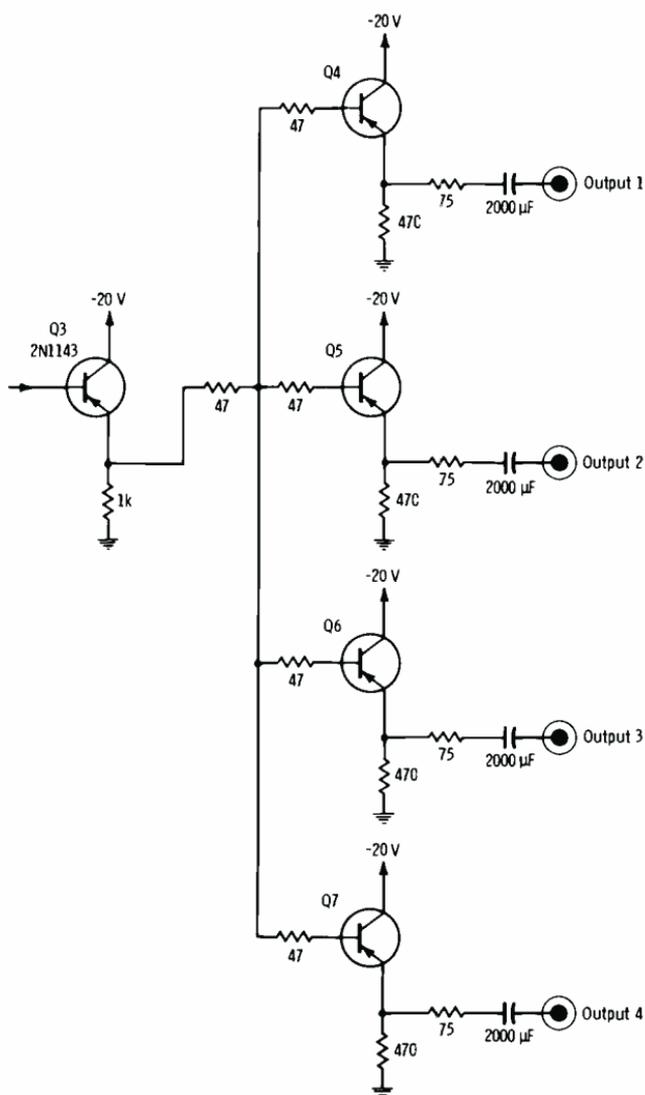
## 6-2. TYPICAL CIRCUITRY IN COMMERCIAL VIDEO DA'S

The straightforward video distribution amplifier (DA) is normally used for distribution of signals from a given source to multiple destinations. Unity gain is ordinarily used, although some limited gain range control is usually provided.

### Coaxial-Line-Driver Output Stages

The emitter-follower stage quite naturally comes to mind as a "natural" for driving 75-ohm distribution lines, because it provides impedance transformation from a high-impedance intermediate stage to a low-impedance output.

Fig. 6-16 illustrates the four-output line-driver configuration of a typical commercial video DA. Note that because Q3 is also an emitter-follower stage, the internal output impedance of transistors Q4, Q5, Q6, and Q7 is extremely low (review the text associated with Fig. 6-10). Therefore, we find the conventional 75-ohm build-out resistors in series with all individual outputs. The 47-ohm resistors between Q3 and the following transistor bases maintain isolation between all outputs even if an extraneous signal is inadvertently applied to any output, and they also protect other transistors if one should develop a short. All outputs, of course, normally reach 75-ohm terminations at the receiving end of the coaxial line, but if any one line is unterminated for servicing or rerouting, there is no effect on any other output.



**Fig. 6-16. Four-output circuit of commercial video DA.**

In practice, we will also encounter the complementary-symmetry circuit of Fig. 6-17 in many video DA's. Although various details may differ, the circuit shown is the basic type of configuration. Note that an adjustable resistor ( $R_A$ ) is shown in this particular circuit; it is adjusted to obtain zero dc volts at the outputs (balanced transistor currents). This permits dc coupling with this type of circuit, eliminating the output capacitor (which must be quite large for

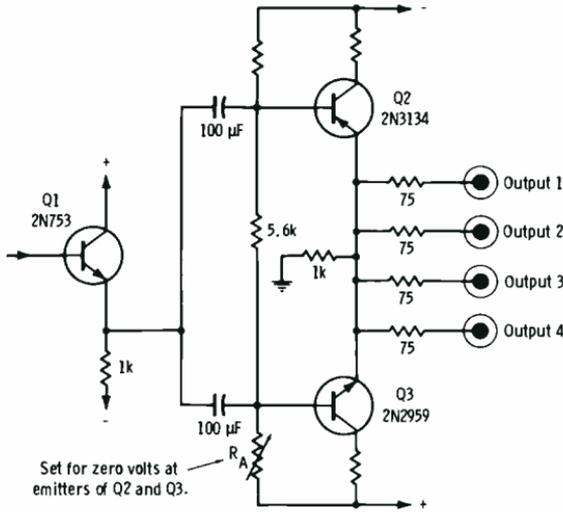
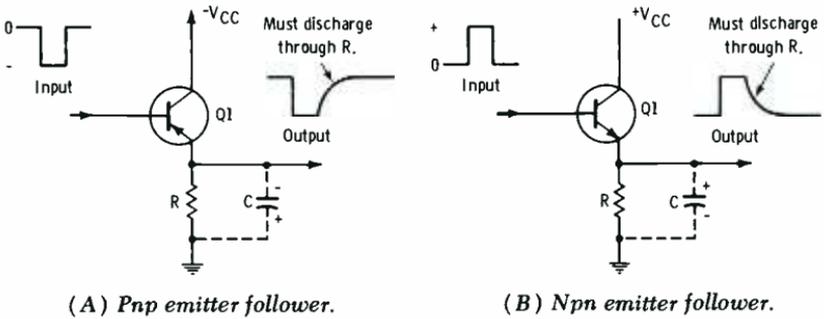
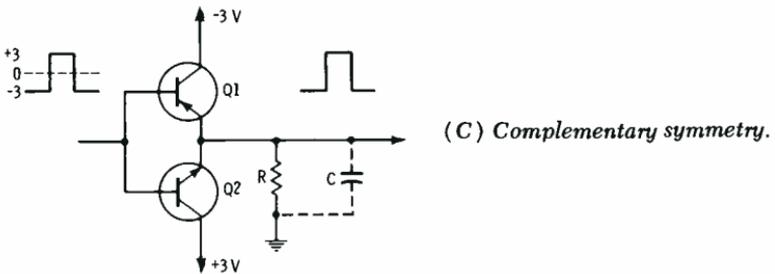


Fig. 6-17. Typical dc-coupled complementary-symmetry output stage.



(A) Pnp emitter follower.

(B) Npn emitter follower.



(C) Complementary symmetry.

Fig. 6-18. Effects of capacitance on pulse amplifiers.

adequate time constant). Sometimes a fixed resistor is used in this position, and it must be checked if a change of components is required, so that the dc output at the Q2 and Q3 emitters is no more than 5 mV.

Fig. 6-18A shows the basic problem in a pnp emitter follower for pulse work. Capacitance C represents the depletion-layer capacitance between the emitter and base, as well as the capacitance of the load. A negative input pulse at the base of Q1 causes conduction, and C is charged through the relatively low resistance of the forward-biased junction with the polarity shown. Removal of the negative drive from the base means that the emitter is held negative with respect to the base, causing cutoff. The capacitance must now discharge through R. We can see that if R is higher in value than the very low charging resistance, the fall time of the output pulse will be longer than the rise time. Following the same line of reasoning, note that the same problem exists for the npn circuit of Fig. 6-18B.

Obviously, the fall time of the output waveform can be reduced by providing a low-impedance discharge path for C similar to the charge path. This is just what is provided by the complementary emitter-follower circuit of Fig. 6-18C. A positive input at the base drives Q2 into conduction, charging C to a voltage almost equal to the input voltage. When the input goes negative, Q1 conducts and discharges C through the relatively low-resistance base-emitter diode of Q1. With matched transistors, rise and fall times are identical, and the output pulse is symmetrical.

When Q1 of Fig. 6-18C is conducting, Q2 is cut off, and vice versa. (Of course, for video amplification, the transistors are operated in their linear region.) So we have a push-pull stage without the need of a transformer or phase splitter. This is one transistor circuit impossible to duplicate with vacuum tubes. Although the circuit is push-pull, it provides a single-in, single-out connection. The output voltage swing has approximately the same amplitude as the input voltage swing.

Another example of a dc-coupled output stage is illustrated in Fig. 6-19. Transistors Q3 and Q4 are connected in a unity-gain feedback amplifier, in which the collector of Q3 drives the base of Q4. Without Q1 and Q2, the output impedance at the junction of Q3 and Q4 would be equal to the output impedance of Q3 acting as an emitter-follower, divided by one plus loop gain. This low impedance is further lowered by a factor of one plus the loop gain provided by Q1 and Q2. This results in an internal output impedance of around 1 ohm. Note from the waveform polarities that the feedback provided by  $R_f$  is positive. Such an amplifier has been termed "potentiometric." The dc output balance control provides

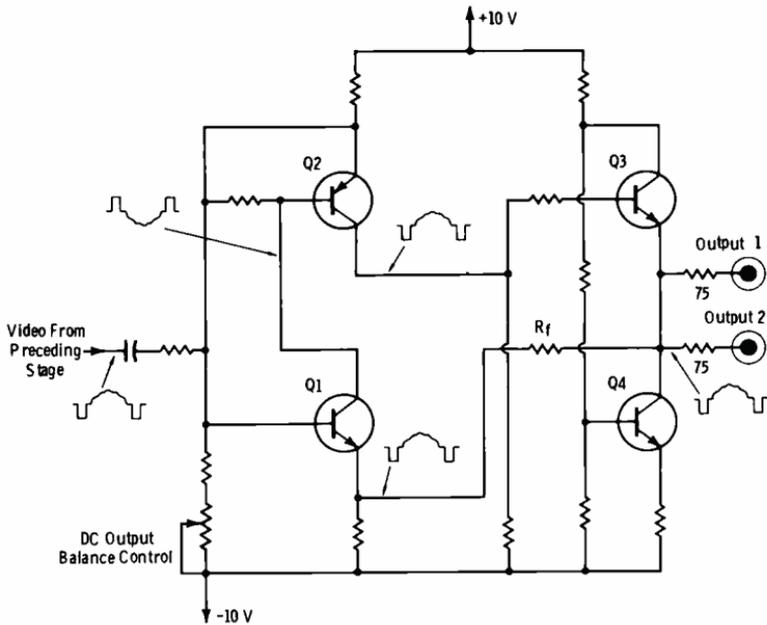


Fig. 6-19. Output stage with dc coupling.

the means for balancing the opposing dc currents of Q3 and Q4 to obtain zero dc volts at the output junction.

### The Intermediate Stage

The intermediate stage of a video DA normally includes the video level (gain) control, and provides for any type of amplitude versus frequency compensation that may be used. These functions are in addition to the necessary basic function of buffering the input stages from the output stages.

Fig. 6-20 presents a simplified schematic diagram of the intermediate stages in the RCA Type TA-23 video DA. Actually, Q1 also serves as the video input stage, and this function will be described later.

The base voltage of Q1 is approximately  $-5$  volts dc. The dc emitter current of Q1 is established by R4; the voltage drop across R4 is determined by the base voltage and the negative supply voltage. The signal gain of Q1 is established by the signal impedance in the emitter circuit. The large capacitor, C1, offers a low impedance down to very low frequencies; therefore, the dominant impedance at signal frequencies is the sum of R2 and R3, which is always small compared with R4. The signal current in the emitter is approximately equal to the base signal voltage divided by the



which R10 sets the dc emitter voltage of Q3 is as follows: (A) The dc collector current of Q1 is set by R4; (B) the dc current through R10 is the same as the dc collector current of Q1 (neglecting the base current of Q2); (C) the potential at the base of Q2 is set at approximately 0.7 volt above ground by normal operation of Q2, the emitter of which is connected to ground through the low dc impedance of L1; (D) the current through R10 having been set, and the potential at its left-hand end having been set, the potential at its right-hand end is thus established. This potential is approximately +9 volts dc.

2. Resistor R8 sets the dc operating current of Q2.
3. Resistor R11 sets the dc operating voltage of Q3.
4. Capacitor C4 bypasses signal currents from the collector of Q3 to ground, thereby eliminating signal voltage from the collector of Q3.
5. The network of C2, R5, R6, C3, and R7 forms an impedance from the feedback-pair input to ground in order to provide adequate phase margin inside the feedback loop. The network has an impedance that decreases with increasing frequency, with a minimum of phase shift. This characteristic reduces the loop gain at high frequencies, by shunting signal currents to ground and thereby removing them from the feedback loop. The minimum-phase-shift characteristic of the network permits its effect to be primarily on the amplitude characteristic without adding to the phase shift inside the loop.
6. Resistor R9 and inductor L1, in the emitter circuit of Q2, serve to raise the input impedance of Q2 at high frequencies, and thereby aid the removal of high-frequency currents from the loop by the network discussed in (5) above. In practice, R9 is adjusted in conjunction with R6 for proper response.
7. Diodes D1 and D2 are for protection purposes; they limit the reverse bias that is possible across the base-emitter junctions of Q2 and Q3. Such reverse-bias conditions may be caused by the loss of one or both power-supply voltages, or by a fault voltage at the input.

### The Input Stage

The input stage must provide a bridging input for a 75-ohm line, and it usually provides some kind of protection from faults such as excessive dc on the line or high ground-loop currents.

Fig. 6-21 shows the input stage of the amplifier intermediate section of Fig. 6-20. The dc voltage for Q1 is developed by a bleeder which runs between the positive and negative supply buses. The positive supply bus in this particular instance is nominally at +36 volts. The negative supply bus is nominally at -36 volts. The bleeder

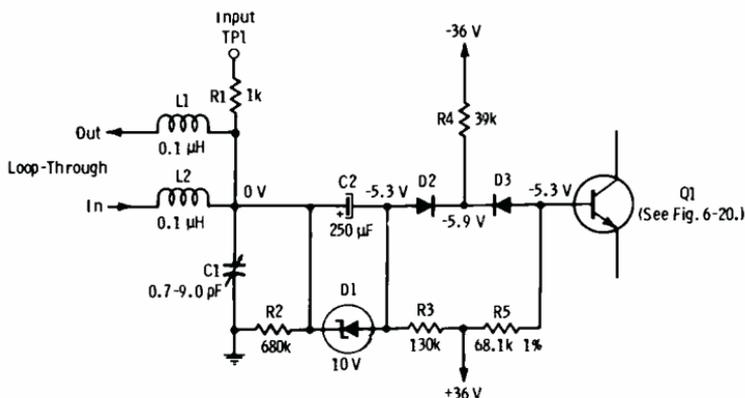


Fig. 6-21. Input circuit of video DA.

which establishes base voltage for Q1 consists of resistors R3, R4, and R5, and diodes D2 and D3. The diodes are normally biased on with only approximately 0.6 volt appearing across each of them. These diodes are part of a protection circuit which prevents excessive positive or negative fault voltages from appearing at the base of Q1. For example, a high positive fault voltage at the input of the amplifier will bias off D3, whereas a high negative fault voltage will bias off D2.

The functions of the other components at the input of the amplifier are as follows: Chokes L1 and L2 and trimmer capacitor C1 form an input-capacitance-compensation network. In particular, L1 and L2, which form the connection links from the two input connectors to the amplifier input, form a T-section, low-pass filter. This filter has an image impedance of 75 ohms. The image impedance may be trimmed to the actual characteristic impedance of a given piece of coaxial cable (with a nominal 75-ohm value) by means of capacitor C1. A feature of this compensation arrangement is that it is contained on the module. When the module is removed from the amplifier frame, a jumper link on the frame connector bridges the two input connectors together, providing continuity in the input line. The switching action of this normalizing jumper link is make-before-break, assuring no interruption of signal transmission along the input line.

Also at the amplifier input is a resistor, R1, which is connected to the input test point. This 1000-ohm resistor minimizes the capacitance loading effect of the long circuit run from the back of the unit to the test point, which is located on the front; it also minimizes loading effects of test equipment, such as an oscilloscope, which may be connected to the test point.

Another resistor, R2, is shunted to ground at the amplifier input. This resistor provides a discharge path for C2 when the unit is removed from service. Diode D1, which is a zener diode, protects C2 from excessive voltages at the input. For example, a high positive fault voltage will force D1 into operation in the breakdown region. The breakdown voltage (10 volts), being less than the voltage rating of C2 (12 volts), limits the voltage across C2 to a safe value. A high negative fault voltage will cause D1 to conduct in the forward direction with a potential of approximately one volt across it. This voltage, which is in opposition to the polarization of C2, is small enough to prevent damage to C2.

As previously mentioned, it is desirable to obtain as high an input impedance as possible in a video DA so that a number of such amplifiers can be "looped through" on the inputs without loading effects on the 75-ohm video line. The number of amplifiers that can be bridged across a single line is then limited by the input capacitance across the 75-ohm terminating impedance.

The field-effect transistor (FET) has a very high input impedance similar to that of a vacuum tube. Fig. 6-22 shows a representative type of video DA (TeleMation TMV-550) using an FET. Signal voltage is provided by amplifier pair Q1-Q2. Field-effect transistor Q1 assures a high input impedance for the video input signal. Negative feedback is applied through R4 to the source of Q1 to prevent distortion and to ensure maximum stability. Gain adjustment is provided by potentiometer R5.

For the conventional transistor, we know there are two major differences in operation compared to the vacuum tube: the input circuit must be forward biased (the tube is actually reverse biased), and the output circuit must be reverse biased. So essentially we have a current-operated device (transistor) compared to a voltage-operated device (tube). This is simply the most convenient way to contrast the two devices.

Now see Fig. 6-23A. In this junction type of FET, a pn junction is employed as the gate (control) electrode. The sole function of this gate is to provide a control element analogous to the grid of a vacuum tube. The FET also has a source (analogous to the cathode) and a drain (analogous to the plate). Note that the gate receives a voltage which is the reverse bias necessary to control majority-carrier current in the channel.

When the gate voltage is increased, the fields set up by the junction barriers cause a reduction in the number of majority carriers flowing through the channel from source to drain (see Fig. 6-24). As the gate voltage is reduced, drain current increases. The area in which the drain voltage has a relatively small effect on drain current (between points A and B in Fig. 6-24) is termed the "pinch-off

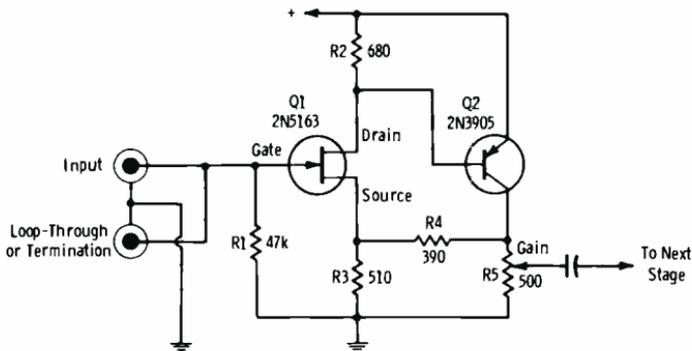
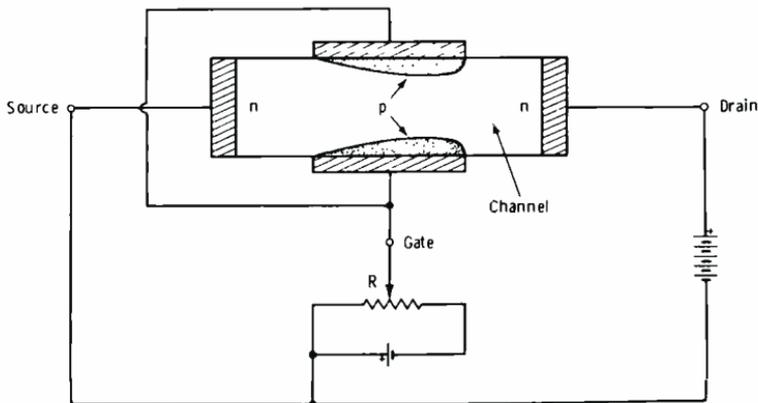


Fig. 6-22. FET input stage of video DA.

region" of operation. As the drain-to-source voltage is increased, or the gate voltage goes to zero bias or a slight forward bias, the drain current can increase excessively, and the FET can be damaged by the resultant heating of the junction (part C of the curve in Fig. 6-24).

Just as with the vacuum tube, the input impedance of the FET in the common-source configuration is very high due to the reverse bias on the gate. The FET is advantageous to use where very high input impedances are required in a single stage of amplification.



(A) Simplified diagram.



(B) Schematic symbols.

Fig. 6-23. Fundamentals of junction FET.

The device is extremely delicate in handling and servicing, but this disadvantage is rapidly being overcome by "protected gate" design.

Schematic symbols for the FET are shown in Fig. 6-23B. Figs. 6-25A and 6-25B present a circuit drawn for an n-channel and a p-channel FET. The gate input impedance is normally around 10 megohms, and a fixed resistor ( $R_G$ ) is used so that the input signal voltage ( $v_i$ ) is developed across a known load for all frequencies concerned.

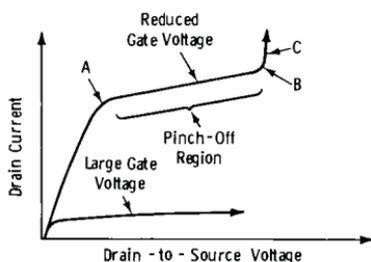


Fig. 6-24. Typical FET drain-current characteristic.

Typically, the signal at the source ( $v_s$ ) will be between 0.6 and 0.75 of  $v_i$  when  $R_S$  is not bypassed. Resistor  $R_S$  supplies the proper bias to the gate and (for class-A operation) fixes the drain voltage at about  $\frac{1}{2}$  of  $V_{DD}$ . The drain signal output ( $V_d$ ) is inverted.

Just as a  $\beta$  of 50 is a good "average" for the conventional transistor, so the transconductance ( $g_m$ ) for the proper operating point of the average FET can be assumed to be 500 micromhos, and  $A_v = g_m R_L$  (approx) without feedback.

NOTE: In an FET,  $g_m$  varies with drain current just as the beta of a conventional transistor varies with collector current. Therefore, it is subject to wide variation, and data sheets must be consulted to determine  $g_m$  at a specified drain current.

The voltage amplification with source feedback (C not connected) is:

$$A_v = \frac{g_m R_L}{1 + g_m R_S} \text{ (approx)}$$

For the circuit of Fig. 6-25A, assume that  $g_m = 500$ . Then the gain without feedback is:

$$A_v = (0.0005)(20,000) = 10 \text{ (approx)}$$

With C not used (source feedback):

$$A_v = \frac{0.0005(20,000)}{1 + 0.0005(3000)} = 4 \text{ (approx)}$$

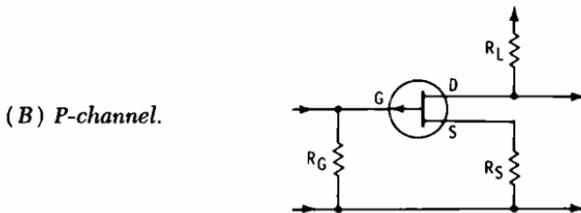
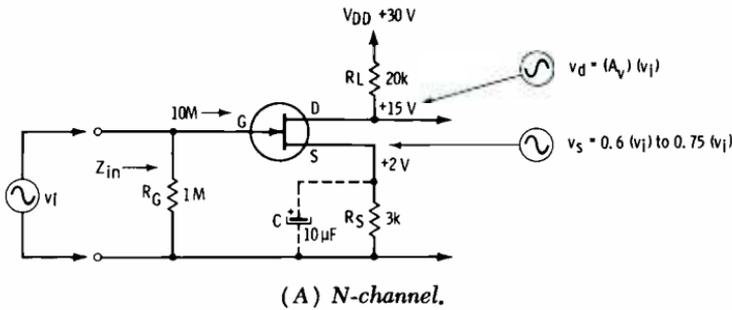


Fig. 6-25. Amplifier circuits using FETs.

Fig. 6-26 shows symbols for three types of metal-oxide-semiconductor FET (MOSFET). These transistors are treated further in applicable portions of following chapters.

FET characteristics may be summarized as follows:

1. High impedances permit use of vacuum-tube biasing techniques.
2. Very good thermal stability.
3. Extremely low feedback capacitance.
4. Low noise figure ( $nf = 3.5$  dB, typical). A noise figure of 5 dB is considered very good for conventional transistors.

### 6-3. UNBALANCED AND BALANCED COAXIAL CABLE DISTRIBUTION

Most coaxial-cable runs within the tv plant use conventional 75-ohm RG-11/U cable, which consists of a center conductor and a shield, the shield being connected to ground at both the sending and receiving terminals (Fig. 6-27). Providing that a good grounding system is installed, such runs of short to medium length cause very little problem. On long runs, however, a difference in ac potential can occur between the shield grounds at the ends of the cable, and a hum component appears in the picture. The video signal works between the center conductor and shield, with opposite-

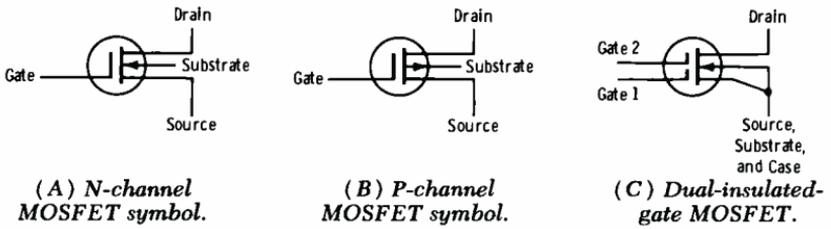


Fig. 6-26. Schematic symbols for MOSFETs.

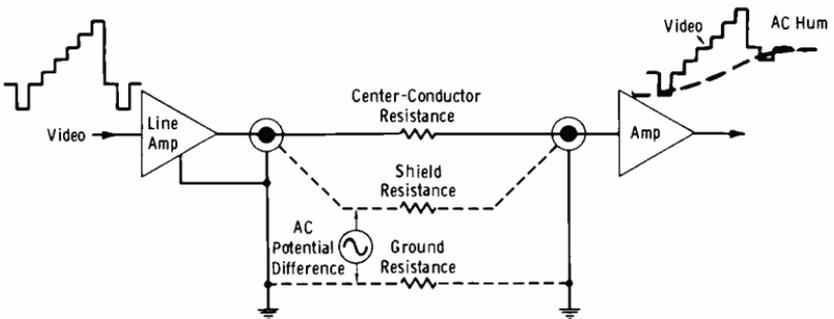


Fig. 6-27. Basic problem with long run of unbalanced video cable.

polarity currents. (If you reverse the shield and ground connections on a monitoring coaxial input, you have a negative-polarity picture.) The ac interfering signal flows from ground back through both the center conductor and the shield.

Fig. 6-28 illustrates a method for reducing the hum problem by using a differential-input amplifier (Section 6-7). The shield is grounded at the sending end, but a large resistance (R) is inserted

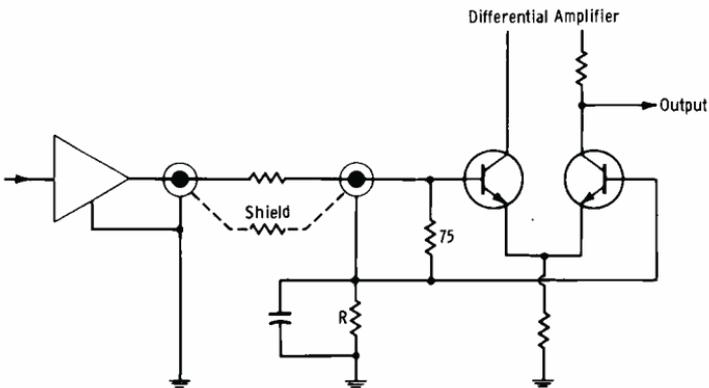


Fig. 6-28. Unbalanced coaxial cable with differential amplifier.

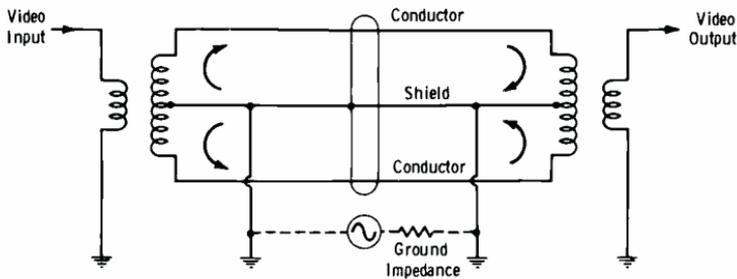


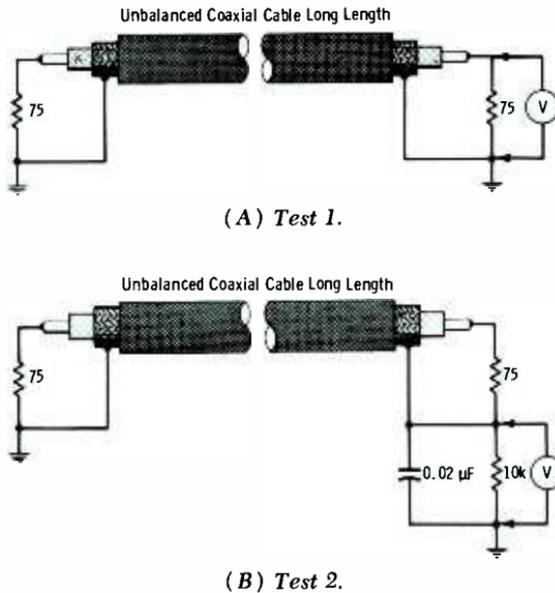
Fig. 6-29. Equivalent circuit of balanced-line transmission.

between the shield and ground at the receiving end. With the differential input, the video signal appearing between the center conductor and shield is amplified in the normal manner. However, the hum component (same polarity on both conductors) is rejected. This provides a satisfactory solution if the ac hum potential is not sufficient to overload the differential amplifier. The same problem exists with a clamper amplifier if the hum component is so high as to overload the clamp circuit. Thus, under some circumstances in relatively large installations, the use of 124-ohm balanced coaxial lines is mandatory.

The balanced coaxial cable is a single cable with twin conductors and a shield. The characteristic impedance of this type of cable is standardized at 124 ohms. Fig. 6-29 serves to illustrate the equivalent circuit. The video signal appears in push-pull across the two inner conductors, and ground currents are carried by the shield. Noise pickup (glitches, etc.) or induced ac hum is applied equally into each conductor and appears in phase. These in-phase disturbances are cancelled at the termination of the run, while the video signal in push-pull, for which the two lines are balanced, is transferred to the load. (For a practical circuit, see Fig. 6-31.) Many of the more recent video DA's provide inputs suitable for the balanced-cable distribution technique.

Dynair Electronics, Inc. provides a practical test to be made to determine whether a balanced cable run should be used, or whether improvements in existing grounding systems will permit the continued use of the unbalanced cable. See Fig. 6-30. Remove all equipment from each end of the cable and terminate as shown for the test. The cable should be dressed in the normal routing being used. For test 1 (Fig. 6-30A), if the voltage is less than 5 millivolts ac, the hum content is low enough to allow use of unbalanced line and a regular DA. If the induced voltage is more than 5 millivolts (46 dB below 1 volt), then perform test 2.

For test 2 (Fig. 6-30B), if the voltage is less than 1 volt ac, the hum content is low enough to allow use of a differential amplifier



**Fig. 6-30. Tests for unbalanced coaxial cable.**

with the unbalanced line (Fig. 6-28). If the voltage is more than 1 volt ac, attempt to reduce it to below 1 volt by improving the grounds on each end of the cable, or by using a parallel grounding wire between the ends of the cable. If this does not reduce the induced voltage to less than 1 volt ac, use balanced line with a video DA that has a suitable input for the 124-ohm balanced line, and use a DA with balanced line outputs at the sending end of the cable.

Fig. 6-31 illustrates a typical differential-amplifier stage for a video DA accommodating 124-ohm balanced input lines. This amplifier has a differential input and a single-ended output. Transistor Q3 provides the constant-current supply for differential amplifier Q1-Q2. Such circuitry is often fabricated in IC's rather than from discrete components (Section 6-7).

Note that Q1 serves as an emitter-follower feed to the Q2 emitter. A negative-going signal at the Q1 base results in a negative-going signal at the Q2 collector. The same negative-going signal at the Q2 base results in an inverted (positive-going) signal at the Q2 collector. Thus in-phase signal inputs to Q1 and Q2 cancel in the collector load of Q2. As described previously, video signal components on the balanced line are push-pull, or 180° out of phase at the Q1 base and Q2 base. Thus the desired signal components add at the Q2 collector and are fed to the following stage minus the noise, hum, and "glitch" in-phase interference signals.

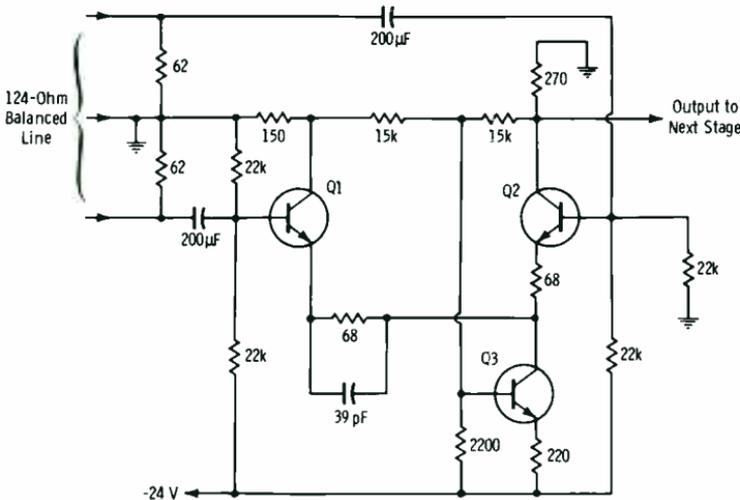


Fig. 6-31. Typical differential amplifier used for balanced line input.

#### 6-4. REMOTE-OPERATED VIDEO GAIN CONTROLS

Video gain controls are sometimes required to be located on remote operating panels, particularly in the case of incoming network or remote distribution amplifiers. These controls take the form of some kind of voltage-controlled gain circuit, usually in the intermediate stages of a video DA.

In vacuum-tube video amplifiers in which the gain is controlled from a remote panel, a variable dc bias control is normally provided to adjust the operating point of a remote-gain-control tube in the unit. Such an arrangement is not satisfactory for transistor circuits. The very small increments of dc bias required would be subject to noise and transients in remote cabling and power supplies. For this reason, we will often encounter the light-controlled gain adjustment shown in Fig. 6-32. The two parallel lamps used for illumination of the photocell have their current controlled from the remote location; this current normally varies from about 70 to 140 mA and is provided by the current source, Q2. For a 20-volt circuit, the lamps are usually 28-volt types, such as the type 1829 or similar bulbs. Thus their life is quite long and reliable. Sometimes the lamp and the photocell are contained in a single plug-in unit about the size of a 14-pin dual in-line IC.

For video service the photocell will, in general, have a resistance which varies from about 70 ohms to 30,000 ohms, depending on the currents in the lamps that illuminate the cell. The lamp current is adjusted from the remote control position. This type of control

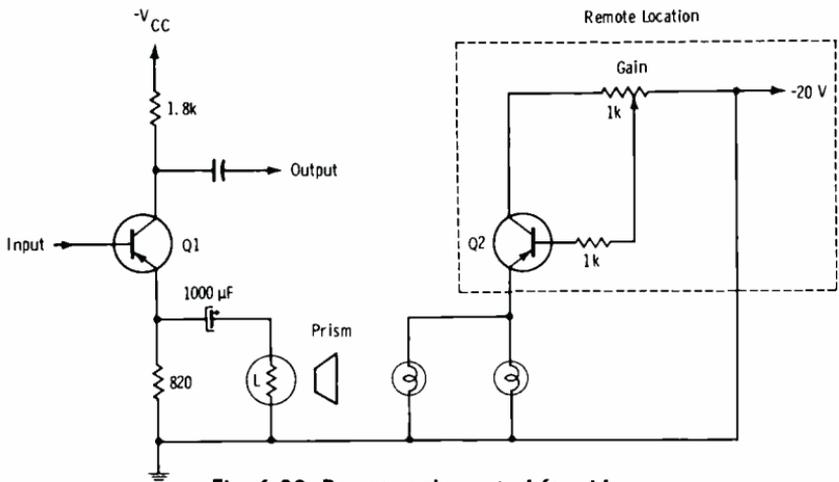


Fig. 6-32. Remote gain control for video.

often is employed in both audio and video applications when the remote-control feature is provided.

Another type of circuit for remote control of gain is shown in Fig. 6-33. The MOSFET provides a voltage-controlled variable attenuator in series with the video path. The forward transfer characteristic is shifted along the gate-voltage axis by a change in the gate voltage, which is controlled by a potentiometer on a remote control panel. For the n-channel MOSFET shown, the signal output voltage decreases (drain-to-source resistance increases) as the gate voltage is made more negative, and it increases as the gate voltage is made less negative. The minimum attenuation is between 1 and 3 dB, and the maximum attenuation possible is about 60 dB. The circuit works into a low-impedance network for minimum capacitive effects at higher frequencies in the video passband.

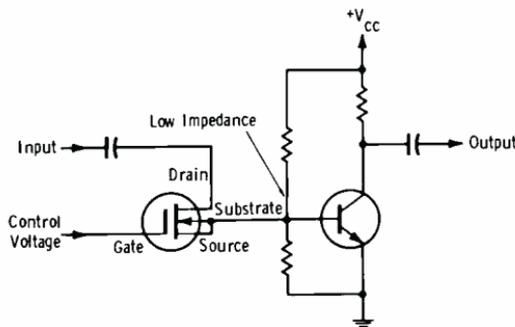


Fig. 6-33. MOSFET voltage-variable resistance element in series with signal path to provide remote gain control.

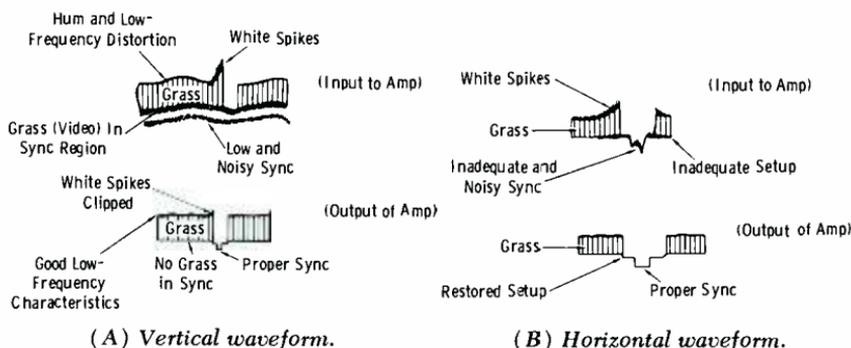
## 6-5. SPECIAL PROCESSING AMPLIFIERS

A processing amplifier is used where it is desirable to remove defects such as hum, noise, loss of setup, improper sync-to-video ratio, etc., from a signal. In some cases it is also required at the transmitter for predistorting the signal to compensate for nonlinearities in the modulation or amplification process.

There is one feature common to any such amplifier, whether it be a stabilizing amplifier, clamper amplifier, or one of the more recent complex processing amplifiers. This feature is the clamping circuit which must be used wherever it is necessary to obtain a stable black reference, or where hum and noise signal components must be minimized.

### The Clamper

One of the primary reasons why clamping is necessary is shown in Fig. 6-34. The clamp removes the hum component or low-frequency distortion. It also provides a black signal reference so that amplification and clipping can be used for elimination of spikes in the white region, removal of noise in the blanking region, and restoration of proper setup, correct video-to-sync amplitude ratio, and "clean" sync pulses.



**Fig. 6-34. Action of stabilizing, clamper, or processing amplifier on waveforms containing indicated faults.**

Keying pulses are derived from the sync portion of the incoming signal or from the sync input itself, properly shaped and amplified, and used to operate the clamping circuits. The keying-pulse shaping circuits which develop the clamping pulses provide a delay in time so that clamping takes place during the portion of the blanking signal that follows the sync interval. This is the "back porch" interval of the blanking pulse. Clamping during this interval is much more effective than attempting to clamp on sync tips, since any compres-

sion of the sync region would tend to defeat the purpose of the clamping circuits.

A clamping circuit accomplishes two things that at first appear to be contradictory: (1) It improves low-frequency video signal response and (2) it eliminates spurious low-frequency pickup, such as 60 Hz ac, from the amplifier response.

To understand how an electronic device can eliminate one low-frequency signal and improve the response to another low-frequency signal requires a study of the clamping circuit, which functions at high efficiency because of the nature of the tv waveform itself. Consider first Fig. 6-35, which shows an ac signal source with a series RC circuit and a switch. Waveform 1 represents the signal voltage at the output terminals when the switch is open. Suppose now that the switch is closed (shorting the output terminals) for the duration of the shaded areas along the axis of the ac signal.

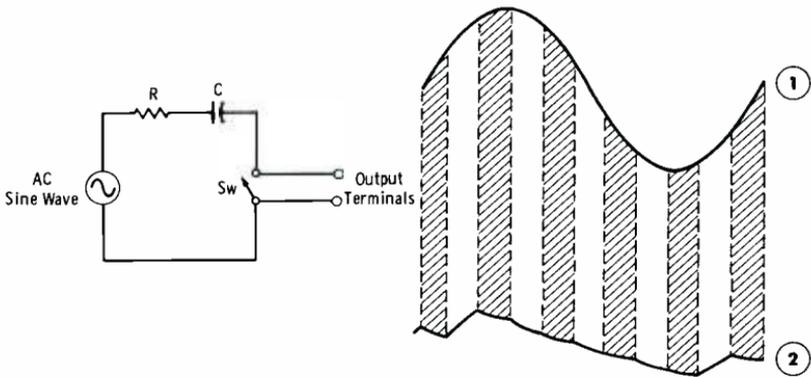


Fig. 6-35. Clamp-circuit theory for sine waves.

Waveform 2 illustrates the severe attenuation of the signal appearing at the output terminals. If the switch were opened and closed at a rate *much faster* than the frequency of the applied ac signal, the output voltage would be practically zero. A clamping circuit is actually an electronic switch that does exactly what was just described. It opens and closes a switch at the horizontal frequency (15,750 times per second) so that any 60-Hz sine wave (such as would occur from a stray field) is greatly attenuated.

Consider now the action of the same circuit when the input waveform is not a pure sine wave, but is broken up by pedestals at a fixed level, such as a video signal with inserted blanking pulses. Waveform 1 in Fig. 6-36 shows the output waveform with the switch open, showing that the same waveform appears as that applied. Now assume that some circuit action occurs which results in poor

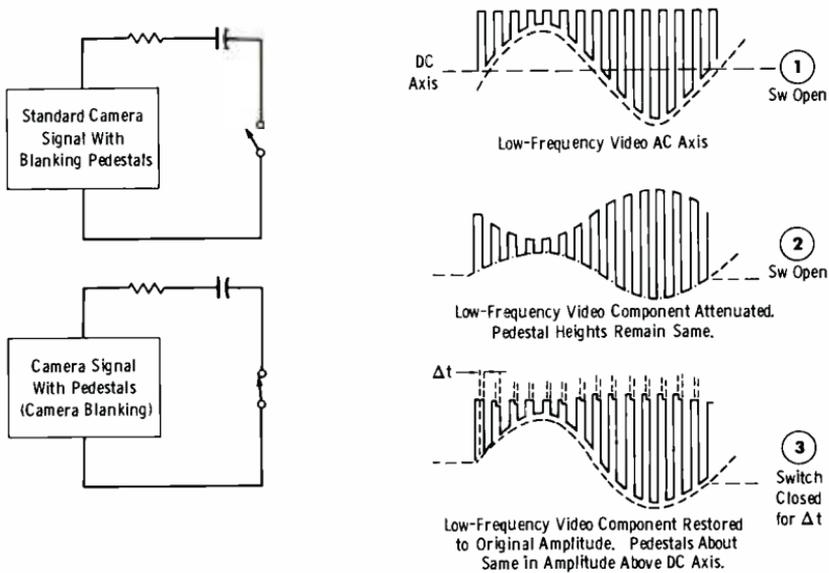
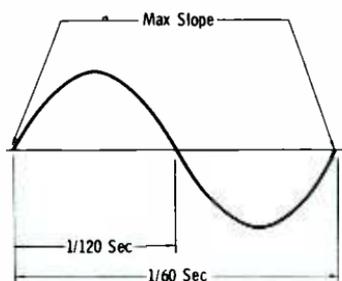


Fig. 6-36. Clamp-circuit operation for tv signal.

low-frequency response. Waveform 2 in Fig. 6-36 shows the resulting waveform when the switch remains open as before. The low-frequency component is attenuated, but the pedestals (fixed levels) remain at the same amplitude. Thus the tips of the pedestal peaks vary from constant level above the dc axis. If the switch is closed for the  $\Delta t$  intervals shown, the output waveform will appear as in waveform 3 of Fig. 6-36, restoring somewhat the original waveform. Should the switch be operated electronically at a rate much faster than the applied waveform, negligible attenuation will result. In effect, the low-frequency video signals will be improved in response. It may be observed that the clamping action depends on having a fixed pedestal level for the duration of time in which the switch is closed. In this way, low-frequency sine waves are severely attenuated, whereas video signals containing fixed pedestals are passed without attenuation, improving low-frequency response without accentuating stray-field response or induced hum.

The effectiveness of line-to-line clamping at 60 Hz may be evaluated by the following analysis: It takes  $1/120$ th second for the hum signal to pass through its amplitude extremes (Fig. 6-37), and  $1/120$  second corresponds to about 131 lines. The amplitude change at the peaks of the hum sine wave is negligible. Therefore, where the slope of the hum signal is maximum, the amplitude change during a single line may be considered to be  $1/131$ , which is 0.00764, or 0.764 percent. Since the signal passes through maximum slope



**Fig. 6-37. Time-amplitude relationship for 60-Hz hum.**

twice in this time, this value is multiplied by 2, which gives approximately 1.5 percent. This percentage may be expressed as 38-dB attenuation of the 60-Hz hum component. Effectiveness at higher frequencies of sine-wave pickup decreases until complete lack of effect is noticed at approximately 2 kHz.

Note that for twice the 60-Hz frequency (120 Hz, or power-supply ripple frequency), the wave passes through maximum slope at a rate two times greater, so the amplitude change becomes 3 percent. This is approximately 30-dB attenuation. At ten times the 60-Hz frequency (600 Hz), the wave passes through maximum slope at a rate ten times as great, for an amplitude change of 15 percent. This is only approximately 16.5-dB attenuation.

The color signal differs from a monochrome signal in two major respects, both of which pose problems in processing amplifiers. First, the addition of the color subcarrier components to the luminance signal causes the resultant color video signals to extend into the blacker-than-black and whiter-than-white regions. Second, a color-synchronizing burst is placed on the back porch, following each horizontal-sync pulse. These characteristics of the color signal give rise to two problems, as follows:

1. *Clipping of burst blacker-than-black excursions.* In monochrome stabilizing amplifiers, the video signal is usually clipped at black level. This removes the sync signal and also any noise spikes or signal overshoots that extend into the sync region. The sync signal is regenerated by amplification and clipping in a separate channel and then added back to the video signal. The purpose, of course, is to restore the sync signal to its original wave shape and amplitude, i.e., to remove any distortion incurred during transmission. In stabilizing amplifiers intended for color, some means must be provided for bypassing the burst and subcarrier components around the clipper so that their blacker-than-black excursions are not clipped off.
2. *Burst distortion.* To ensure that video clipping will automatically occur at black level despite changes in signal level or

average brightness, the signal must be clamped during the back-porch interval. Since it is during this time that the color-sync burst is transmitted, steps must be taken to prevent the clamp action from distorting the burst.

### The RCA TA-9 Stabilizing Amplifier

One of the first "colorized" stabilizing amplifiers to come into common use was the RCA Model TA-9 vacuum-tube amplifier. The two problems, subcarrier clipping and burst distortion, are avoided in this amplifier by passing the composite color signal through a spectrum-separation network, or crossover filter, in which the subcarrier components are separated from the luminance and sync signals. Essentially, this leaves a composite monochrome signal that can be processed in the normal manner.

The block diagram shown in Fig. 6-38 illustrates the major circuit features of the Model TA-9 stabilizing amplifier. Note that the composite picture signal traverses three paths—for chrominance, luminance, and sync. The input signal is first split into two channels, one for picture information and the other for sync. Provision is made for inserting a relay to select either internal or external sync. Use of this relay eliminates the need for the transient suppressor required in many stabilizing amplifiers of older design.

In the sync channel, separation of sync information is accomplished in a high-level clipper. This stage is driven from an automatically gain-regulated amplifier to ensure stable and accurate clipping over a wide range of signal level variations.

A noisy-immunity circuit is used between the clipper and pulse former to provide clamp pulses free from the spurious pulses that might otherwise be formed from noise spikes in the incoming signal. The circuit works by virtue of the fact that spurious noise impulses are normally much narrower than the desired sync pulses. The sync signal delivered to the noise-immunity circuit has previously been doubly clipped so that both the sync pulses and the spurious noise impulses have the same peak-to-peak amplitude. An RC integrating circuit is employed to attenuate the narrow noise pulses greatly so that the sync pulses can trigger the pulse former.

In the picture channel, the signal is again split into two parts, one carrying chrominance information and the other carrying luminance information. The crossover between paths occurs at the color-subcarrier frequency, 3.58 MHz, with a complete null at that frequency in the luminance channel.

The feedback-clamp and clipper circuits are contained in the luminance channel. Here, the purpose of the feedback clamp is threefold: (1) to maintain clipping exactly at black level over long periods of time without readjustment, (2) to set the clipped signal

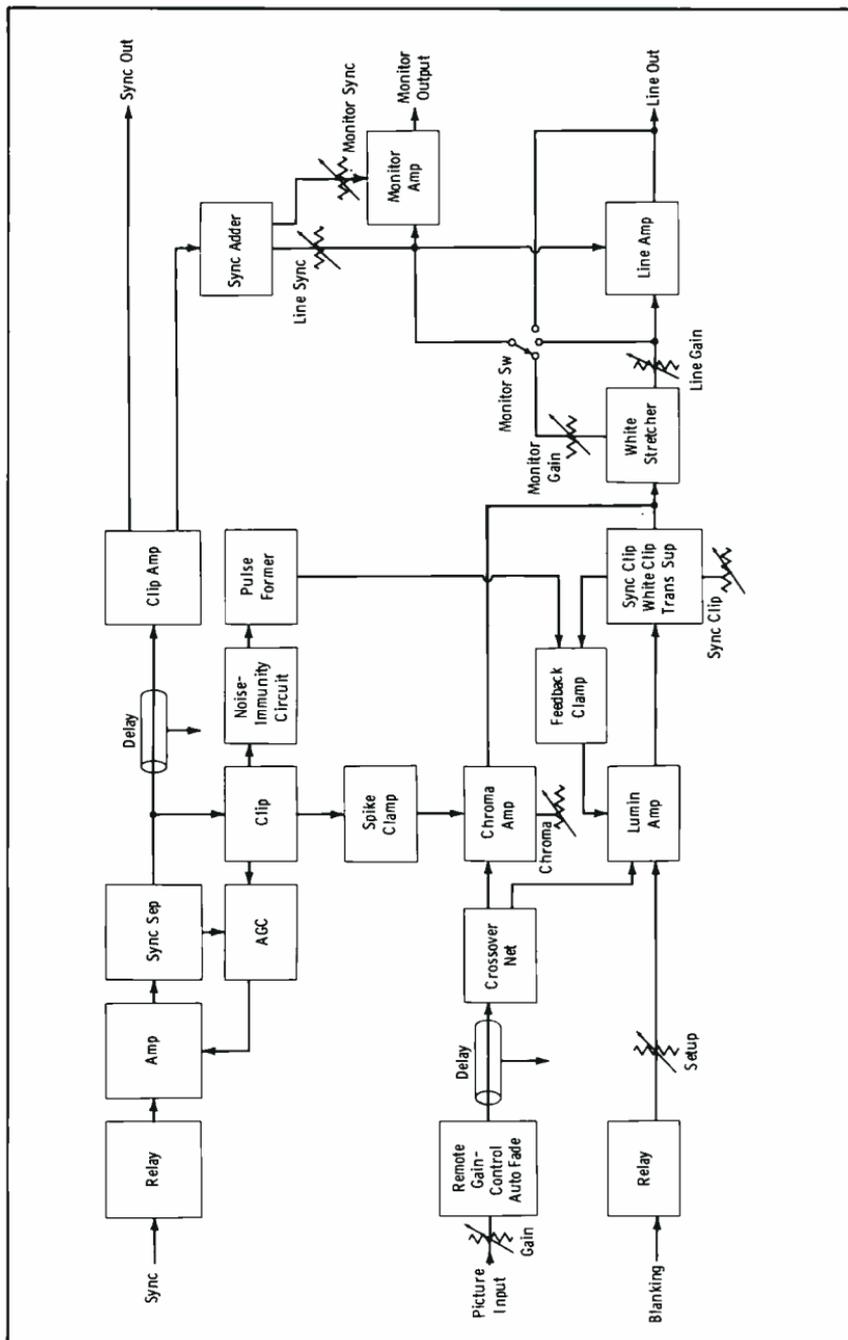


Fig. 6-38. Simplified block diagram of RCA Model TA-9 stabilizing amplifier.

automatically at the proper position on the white-stretcher characteristic, and (3) to provide a high degree of immunity to tube aging and supply-voltage variations. Since the color subcarrier is not present in the luminance channel, sync may be clipped off all the way to blanking level, and back-porch clamping may be performed with full effectiveness without affecting the color burst in the color signal.

Following the clamp stage, where accurate reference level is maintained (for sync clipping, white stretching, etc.), a white-clipper circuit is provided. The purpose of this clipper is to reduce receiver intercarrier buzz caused when the carrier is overmodulated by peak whites. Chroma and high-definition video components may still cause overmodulation, since these components pass through the chroma channel and thus bypass the white clipper. However, the frequency and energy of these components is such that the buzz usually is inaudible.

The chrominance information is passed around the clamp and clipper stages through a two-stage amplifier channel. This allows control over chroma gain and provides proper delay for later recombination of the chrominance signal with the luminance signal. The signals from the chrominance and luminance channels are mixed together and applied to the white-stretch circuit. Here, an adjustable degree of amplitude nonlinearity may be introduced to predistort, or compensate, the signal for later passage through equipment that may cause compression. An example of this requirement is in transmitters which do not contain built-in compensation. A switch is provided to bypass this function when it is not needed. The output composite picture signal is finally formed by addition of the reshaped sync signal to the clamped picture signal.

Differential gain and phase controls are provided to compensate for characteristics of the transmitter, when necessary, as described in Chapters 10 and 11.

The clipper in this unit is a *double* clipper that has the ability to clip the signal at the reference white level (or slightly above) as well as at black level. The white-clipping feature is desirable in cases where signals are marred by impulse noise or excessive highlight "glints," but the same feature can be a source of serious distortion for color signals if inadequate attention is given to proper setting of the white-clip control. Note that the dc restoration for the clipping operation is accomplished with a feedback-stabilized clamp; while the unit may have its share of problems, drift in clipping levels is not likely to be one of them. The white stretcher which follows the clipper is another potential source of trouble. This feature is essential when the stabilizing amplifier is used to drive some types of television transmitters, but some operators tend to forget that

the white-stretch circuit is designed to alter both the differential-gain and differential-phase characteristics to compensate for distortions in other types of equipment. If the white-stretch adjustments are not made carefully with the aid of differential-gain and phase test signals, the final result may be worse than no compensation at all.

The fact that the chrominance information is handled in a separate amplifier which bypasses the clipper represents both an opportunity and a hazard. The opportunity is that of adjusting the gain separately in the chrominance channel to compensate for obvious problems in the chrominance level of the incoming signal. The hazard is that operators tend to forget when they have deliberately increased the chroma gain to compensate for a substandard signal, and fail to restore the gain to normal when the abnormal signal condition no longer exists. If the maintenance engineer receives operator complaints about excessive noise or "spiking" in the TA-9, in the great majority of cases the problem turns out to be nothing more than excessive gain in the chrominance channel.

The spike clamp associated with the chrominance amplifier is necessary to remove the high-frequency spikes associated with the leading and trailing edges of the original sync pulses. These spikes pass through the high-pass section of the crossover filter, but must not be added back to the signal because the reshaped sync pulses provided by the sync channel already have a full share of high-frequency information in their edges. The spike-clamp circuit should not be particularly troublesome, although it could introduce some burst distortion if badly out of adjustment. The instruction book has complete information on the function and adjustment of this circuit.

Occasionally there are complaints about low-frequency instability, or a tendency toward "bounce," in the TA-9. In the majority of cases, this problem results from the use of an improper voltage tap on the power transformer, which delivers ac power both to the heaters and to the built-in bias supply used for the output amplifier stages. If the tap is chosen improperly, the voltage-regulator tube in this bias supply may not "fire" reliably, and the resulting instability in the bias voltage becomes evident in the output signal.

### Details of Solid-State Clamping Circuitry

Fig. 6-39 illustrates the pulse-transformer type of clamping circuit. The base of clamp driver Q1 receives a sharp negative pulse to drive it from cutoff to saturation. The sudden pulse of collector current through the primary of the transformer "rings" this circuit due to the inductive "kickback." But after the first positive excursion, diode D5 shorts out the primary when the collector attempts to swing negative. Note that this polarity would be reversed if an npn transistor were used. Note also the polarity of the resulting pulses

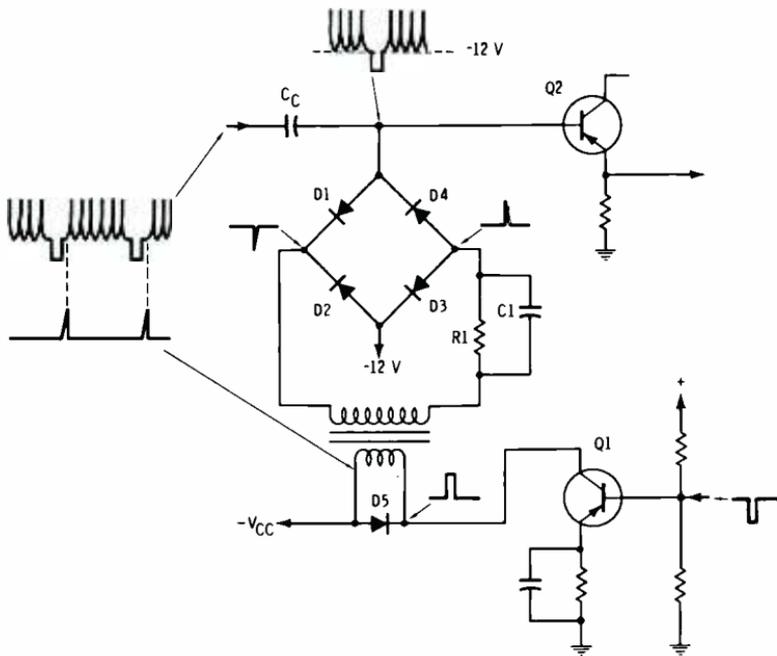


Fig. 6-39. Clamping circuit with pulse transformer.

on the secondary, and how this results in forward-biasing of the quad diode circuit. This forward-biasing closes the "switch," and  $-12$  volts is applied to the base of  $Q2$ . Coupling capacitor  $C_C$  is always small, since it must be charged or discharged quickly (during the approximately  $1.5\text{-}\mu\text{s}$  duration of clamping) to the reference  $-12$  volts. The  $R1C1$  time constant must be long compared to a line interval so that the charge on  $C1$  will hold the switch open (non-conducting) between pulses during the active line (video) interval. As a rule of thumb, the pulse amplitudes at each end of the quad are at least three times the video signal (peak-to-peak) at the clamped base. This relationship will be true for any type of driven clamping circuit.

Now let us go through the circuit of Fig. 6-40 for practice in analysis. Since this is a back-porch clamer, we know that the clamping pulse must occur after the trailing edge of horizontal sync. If we correlate the circuit of  $Q1$  with Fig. 4-12 (Chapter 4), we know how the delay is obtained.

NOTE: Since it is conventional to show grounded points toward the bottom of a schematic diagram, we will almost always see the "inverted" type of drawing when the emitter rather than the collector is connected to the voltage supply.

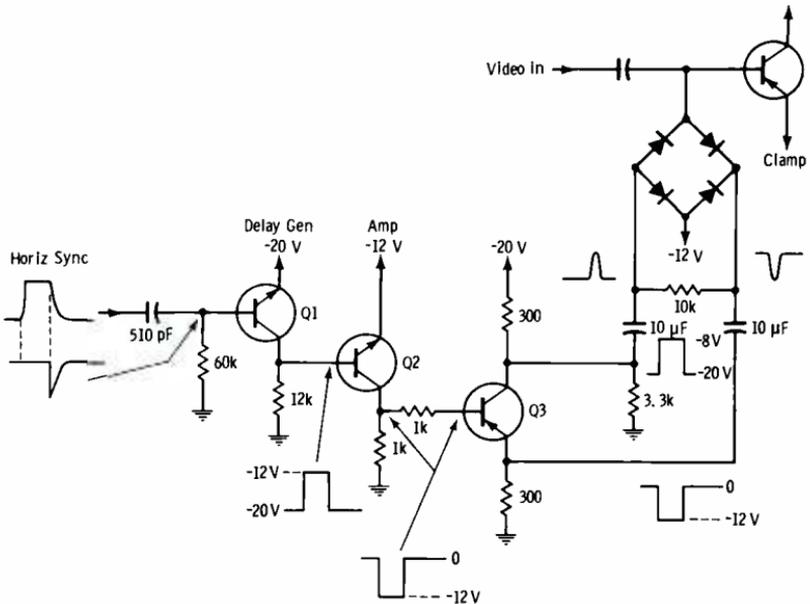


Fig. 6-40. Circuit of back-porch clamper.

Note that Q1 is actually a form of boxcar (saturated), since the emitter is returned directly to  $-20$  volts and the base is grounded through its resistor. The delayed negative base pulse drives Q1 toward cutoff, but the Q1 collector is coupled directly to the Q2 base. The positive-going pulse excursion at this point is "caught" by the  $-12$  volt base-emitter clamp of Q2, and can go no further positive.

We know that prior to the pulse Q1 is saturated and its collector is at  $-20$  volts (switch closed). At this same time, since the emitter of Q2 is at  $-12$  volts and its base is at  $-20$  volts, Q2 is cut off. The positive pulse at the Q2 base drives this transistor on, and since there are no limiting diodes and no emitter resistance for self-bias, Q2 saturates. So we would expect at the Q2 collector a pulse extending from 0 volts (cutoff) to  $-12$  volts (saturation). The unbypassed emitter resistance of phase-splitter Q3 tells us that the input impedance of Q3 will be quite high (very low base current), so we would expect about the same pulse amplitude directly at the Q3 base.

Now analyze Q3, first the interval between pulses, then the pulse interval. Between pulses, the Q3 base is at essentially zero voltage, so the transistor is cut off. During the pulse interval, since the Q3 emitter and collector loads are identical, we will expect essentially unity gain. Therefore, we would expect the Q3 emitter pulse to extend from 0 (cutoff) to  $-12$  volts (unity gain).

The emitter current at the  $-12$ -volt peak is:

$$I_e = \frac{V}{R_E} = \frac{12}{300} = 40 \text{ mA}$$

For a quick analysis, assume the collector current is the same as the emitter current; the collector signal-voltage swing from the pulse is:

$$V_c = (0.04)(300) = 12 \text{ volts}$$

Then we would expect the collector to swing up 12 volts from  $-20$  volts (cutoff), or to  $-8$  volts.

In practice, we realize that the collector current is slightly less than the emitter current ( $I_c = \alpha I_e$ ), and for this reason, we will normally find the collector load resistor of a clamp driver slightly higher in value than the emitter resistance. For a 300-ohm emitter resistance, the collector load would usually be around 330 ohms. Thus the slightly smaller collector-current swing develops a voltage swing equal to that at the emitter.

The above analysis should serve to emphasize an important servicing technique: Always use the dc-amplifier position during scope checks. This mode tells us considerably more about circuit function than the ac-coupled position. For large-signal operation, this method is very convenient.

### Luminance-Chrominance Separation Techniques

Almost all stabilizing amplifiers, and some video-processing amplifiers, separate the luminance information from the chrominance information and process each separately. This permits conventional back-porch clamping on the luminance signal and provides individual circuitry for chroma level, burst level and phasing, and any other processing required before luminance and chrominance are recombined.

The separation of luminance and chrominance information as used in stabilizing amplifiers and other composite color equipment is illustrated by Figs. 6-41 and 6-42. In the processing of Fig. 6-41, L1, C1, and C2 form a parallel circuit tuned to 3.58 MHz. Thus the output across  $R_{E1}$  is low-frequency video and sync only, which is fed to the luminance signal-processing channel. The Q1 emitter is coupled to the Q2 base through a circuit series resonant at 3.58 MHz. Thus the output across  $R_{E2}$  consists only of chrominance information.

The circuit of Fig. 6-42 is a more sophisticated approach which we will encounter in varying degrees in modern composite color-signal units. Transistors Q1 and Q2 are connected in a differential-amplifier circuit. Transistor Q3 isolates (at high impedance) the low-pass circuitry of Q4, Q5, and current-gain driver Q6. The



filtering is accomplished by the shunt capacitors from the bases to ground. Note that the signal phase shift through the low-pass stages is 360 degrees, so the signal at the Q2 base is back in phase with the composite signal at the Q1 base. Thus at low (luminance) frequencies, an amplitude change at the Q1 base is followed by a proportional amplitude change caused by Q2 at the Q1 emitter. Therefore, the Q1 collector signal is the *difference* signal; the luminance information has been cancelled, and only the chroma is passed. Potentiometer R1 is a high-pass balance control which is adjusted to provide the same amplitude of luminance signal at the Q2 base that appears at the Q1 base. If these amplitudes differ, the chroma-signal axis is shifted up or down in step with the luminance amplitude difference. Peaking coil L1 compensates for high-frequency losses.

### Video AGC

Practically all modern video-processing amplifiers incorporate some form of automatic gain control (agc) which maintains proper video amplitude regardless of variation in incoming level. The procedure normally used is to sample the video level by means of a level detector, and to use any variations in amplitude to control the gain of the video amplifier. This requires some form of threshold adjustment so that the gain is not increased on fades to black and the signal does not go to noise.

See Fig. 6-43. When the signal exceeds the breakdown voltage of D1 and the base-emitter voltage drop of Q1, current begins to flow in Q1. The collector of Q1 and the gate of FET Q2 become less negative. However, C2 opposes any change, which tends to increase the response (attack) time. This is the time required for C1 to

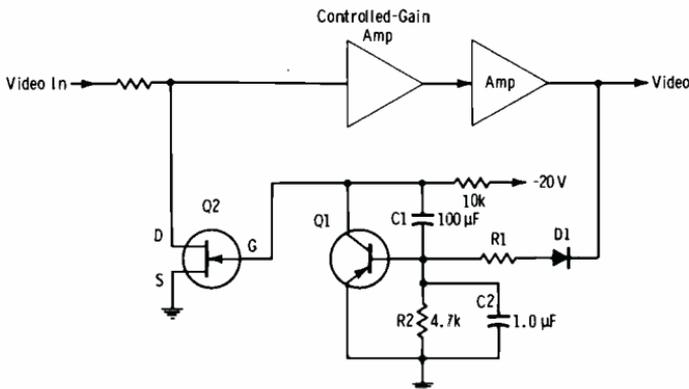


Fig. 6-43. Example of basic video agc circuit.

charge through D1 and R1 to the proper output voltage. As the gate voltage of Q2 does go more positive (less negative), the drain-to-source resistance decreases, and the signal amplitude is therefore reduced. (The FET acts as a shunt resistor and passes more of the signal to ground.) When the video amplitude decreases, the Q1 base tends toward the positive direction to reduce the current. Capacitor C2 again opposes this change until it is discharged through R2 and the base-emitter resistance of Q1. This increases the release time. When the Q1 collector current is finally decreased, the collector of Q1 and the gate of Q2 swing negative, increasing the resistance of Q2 and increasing the video amplitude. The release time is normally made somewhat longer than the attack time so that a complete fade-down to black does not cause a transient resulting from a quick attempt to increase the gain back to normal.

Note that the FET voltage-controlled attenuator of Fig. 6-43 is fundamentally the same as the MOSFET circuit of Fig. 6-33. The major difference is that one is a series-resistance circuit, whereas the other is a shunt-resistance circuit. The basic action remains the same. The gate control voltages, of course, are opposite for similar action on video gain for the two circuits.

### **Additional Features of Video-Processing Amplifiers**

The modern processing amplifier, particularly of the type used in the terminal gear at the transmitter, employs built-in EIA sync generators which supply the video input in case of complete loss of input signal. When an input signal is present, all pulses are regenerated and controlled in amplitude and width within the amplifier. In some units, should equalizing or horizontal-sync pulses be missing, the amplifier will "flywheel" and insert these pulses into the signal from the internal generator, which is locked to the incoming signal.

Processing amplifiers normally provide an automatic bypass to the primary signal path upon failure of the internal power supply or other internal fault.

Controls normally provided are sync gain, video gain, chroma gain, setup level, white clipping level, and unity or variable burst gain. Some units which completely regenerate the burst (separately from chroma processing) incorporate a burst-phase control. Even the breezeway is adjustable in some units so that all output signals are correct regardless of errors in the incoming signal.

Even the modes of video agc are selectable in some units. For example, the RCA processing amplifier provides three switch-selectable modes for the agc system: (1) referenced to the vertical-interval test (VIT) signal, (2) without external reference signal, and (3) automatically referenced to the highest amplitude of both

VIT and picture signals. Control is also provided to eliminate age action. A panel lamp indicates loss of the age function.

## 6-6. DA POWER SUPPLIES

Some video DAs (and practically all video-processing amplifiers) incorporate built-in power supplies. Other DAs have one power supply for each bay of amplifiers, usually eight in number.

Fig. 6-44 illustrates the well regulated power supply of the Tele-Mation TMV-550 video DA. The function is as follows: The ac voltage across the secondary of transformer T1 is rectified by the full-wave bridge rectifier to provide the positive supply voltage for the distribution amplifier. Initial ripple filtering is accomplished by capacitor C1. Voltage regulation is provided by transistors Q1, Q2, and Q3 and zener diode D1. Any tendency of the potential at the cathode of D1 to increase results in a positive transition at the base of Q3 and causes Q3 to increase conduction. The resulting drop in potential at the base of Q2 causes Q2 to decrease conduction, thereby increasing the potential at the base of Q1. The conduction of Q1 decreases, and since Q1 is in series with the power-supply load, the increased voltage drop across Q1 causes the potential at the cathode of D1 to stabilize. Conversely, a tendency of the potential at the cathode of D1 to decrease results in the opposite stabilization process.

NOTE: Large rack-mounted regulated power supplies are detailed in Chapter 3 of this book. Regulators, preregulators, overload protection, and voltage-sensing circuitry are analyzed. Detailed testing procedures with the necessary precautions in hookup are included, and should be referred to if necessary.

## 6-7. INTEGRATED-CIRCUIT LINEAR AMPLIFIERS

Linear integrated-circuit (IC) amplifiers are generally included within the broad family of *operational amplifiers*, often referred to as *op amps*. The term "operational" evolves from the fact that these devices were originally developed for performing mathematical operations in computers, in which case the amplification could be either linear or nonlinear. At the moment, we are concerned with the linear operation mode of ICs.

Three basic types of IC packages and the usual terminal arrangements are shown in Fig. 6-45. Figs. 6-45A and 6-45B are *top* views. The index is either a notch or a dot. Pin numbers increase counterclockwise around the package as seen from the top. The 14-pin arrangement is common, but there may be as few as 8 pins or as

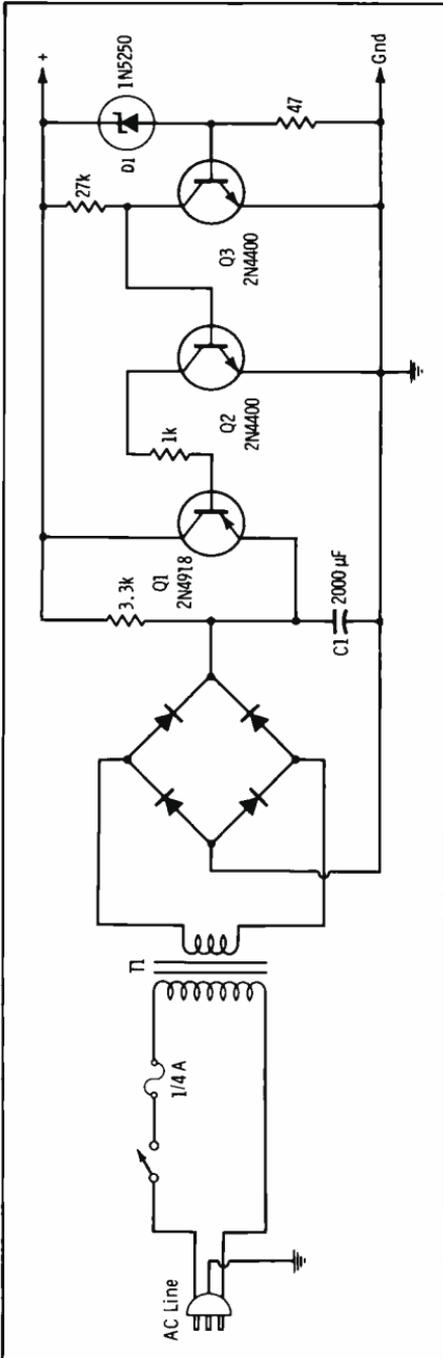
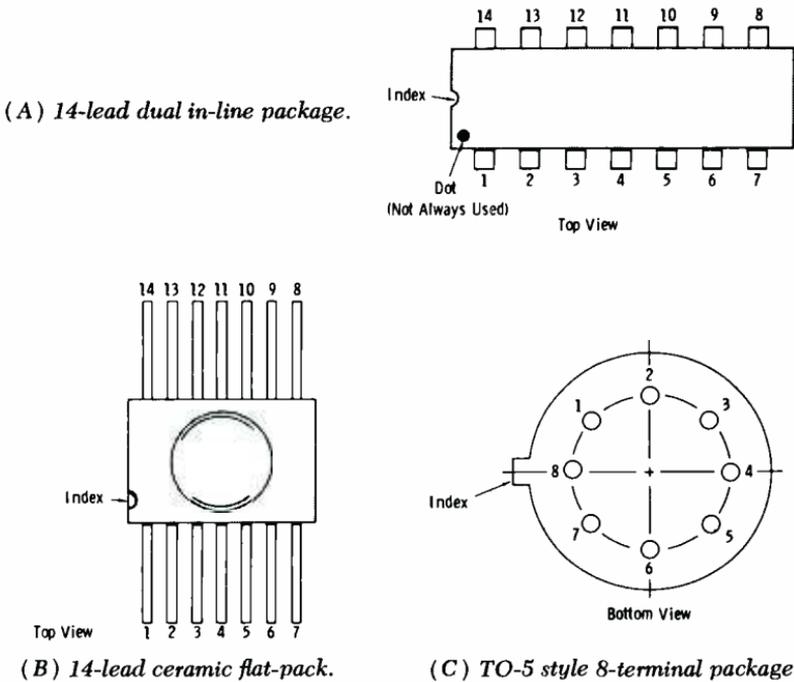


Fig. 6-44. Example of regulated power supply for video DA.



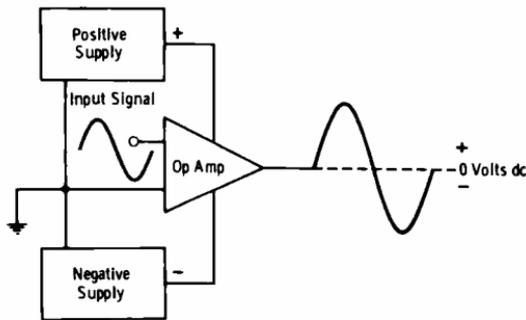
**Fig. 6-45. Basic integrated-circuit packages and terminal arrangements.**

many as 40 pins or more in a single unit. In any case, pin numbering follows the same system as that shown.

Fig. 6-45C shows the TO-5 type of IC package, for which it is conventional to show the *bottom view*. Note that the highest lead number is adjacent to the index tab, and that the numbers progress clockwise as viewed from the bottom. As with other types, the total number of terminals varies.

### Operational Amplifiers

The operational amplifier has two inputs and one output. When signals are applied to both inputs, the device becomes a difference amplifier. If one input is grounded or fixed at a reference voltage, the unit may be used as either an inverting or noninverting amplifier, depending on which input is used for the signal. Usually, the connections are arranged so that the input signal is referenced to ground and the output signal is referenced to zero volts dc with respect to ground. See Fig. 6-46. The zero dc level at the output is established by operating the amplifier between two power sources of opposite polarity to ground, with the output terminal connected to the point that is at zero dc voltage with respect to ground.



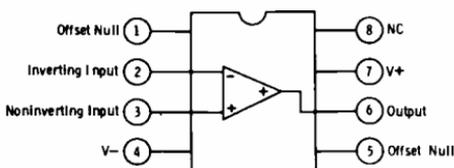
**Fig. 6-46. Method of obtaining dc ground at output of operational amplifier.**

Fig. 6-47 shows the pin configuration of a basic operational amplifier (Type 741). This is an 8-pin chip with two pins used for an offset null network. (The purpose of the offset null is described later.) The minus input terminal is the *inverting input* (output signal inverted from input polarity), and the plus input terminal is the *noninverting input* (output signal same polarity as input). The output terminal in Fig. 6-47 is marked plus, but this terminal is not normally designated with a plus or minus sign. This particular type of op amp has internal phase compensation, but some types have extra pins for external phase-compensation components. Some types have an additional ground pin. (The Type 747 IC is a 14-pin chip with dual 741s.)

The operational amplifier is represented on schematic diagrams by a triangular symbol with terminal numbers shown but no internal details included. Also, the power-supply terminals normally are omitted from a diagram, and only signal-pin connections are shown.

The op amp is a direct-coupled dc amplifier with very high open-loop (no feedback) gain, high input impedance, and low output impedance. Open-loop operation is not used in linear circuits, and the amount of external negative feedback applied determines the gain of the circuit. This is closed-loop operation.

Fig. 6-48 illustrates the four basic applications of linear operation. The feedback resistor ( $R_f$ ) is always between the output terminal and the inverting input terminal so that negative feedback is obtained. In Figs. 6-48A and 6-48B, note that the actual amplification



**Fig. 6-47. Configuration of Type 741 operational amplifier.**

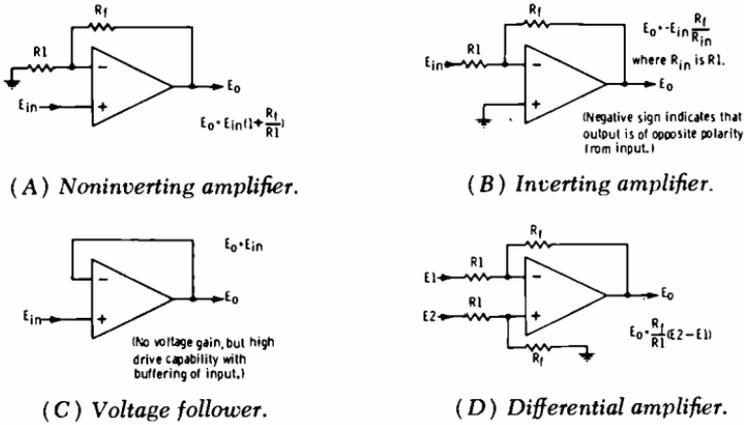


Fig. 6-48. Four basic applications of linear operational amplifiers.

is essentially the ratio of  $R_f$  to input resistance  $R_1$ . Thus, if the ratio  $R_f$  to  $R_1$  is 10, the voltage amplification is 10.

From the foregoing, it would appear that the output voltage can be made as large as desired by increasing the difference between  $R_f$  and  $R_1$ . This is not the case. Because of the dual power supply, which allows the dc output to swing above or below ground, saturation can occur in either the positive or negative direction. The range of linear operation is the voltage range between saturation limits. Increasing the resistance of  $R_f$  finally brings the operating mode to essentially an open-loop characteristic. The open-loop amplifier (infinite resistance between output and input) saturates rapidly with only a few millivolts of input signal.

The voltage follower of Fig. 6-48C is just the opposite of an open-loop configuration; the output is shorted to the inverting input. Thus the output voltage becomes equal to the source voltage applied to the noninverting input, and unity amplification occurs. There is no voltage gain, but the output is buffered and has a high drive capability.

If different signals are applied simultaneously to the input terminals (Fig. 6-48D), the output voltage is the sum of two components, one out of phase with the input signal applied to the inverting input, and the other in phase with the signal applied to the noninverting input. This is a differential-input circuit, and the IC is being operated as a differential amplifier. If the resistors are perfectly matched and  $E_1$  and  $E_2$  are identical in amplitude and phase, zero output occurs. Any difference in amplitude then is amplified as  $E_2 - E_1$ .

In an ideal amplifier, the output voltage is zero when the input voltages are zero. In practice, the output is shifted (offset) slightly.

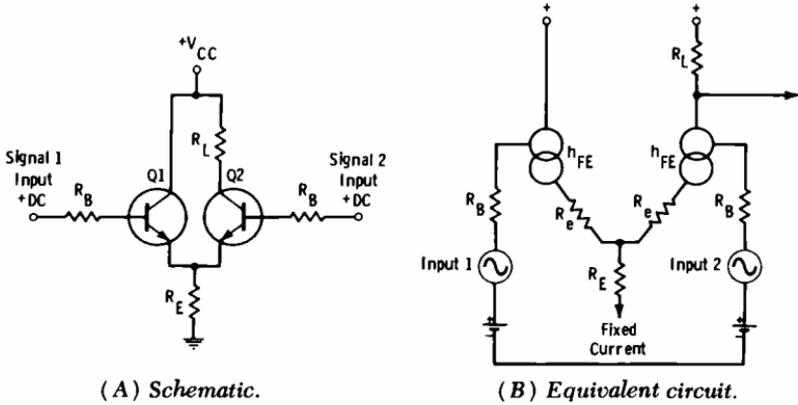


Fig. 6-49. Differential amplifier.

The positive and negative power supplies might be 10 volts each, and the offset might be 200 millivolts. When this has an effect on proper operation, an *offset adjustment* is used. For example, a 10-kilohm potentiometer can be connected between terminals 1 and 5 of a Type 741 IC (Fig. 6-47), with the arm connected to a negative supply. The potentiometer is adjusted to null any offset in the output when the inputs are zero.

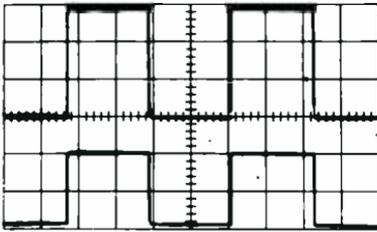
### The Differential Amplifier

Fig. 6-49 shows a basic differential amplifier circuit. The term "differential" in the amplifier name is not related to differential calculus or a differentiating network. The circuit is, quite simply, a difference amplifier. The circuit is used in many applications, including audio and video amplifiers. Let us see why.

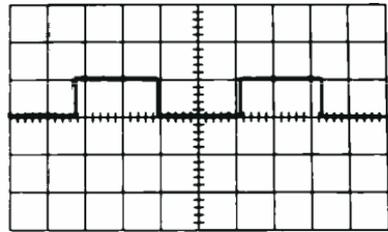
One of the main contributors to the low-frequency noise problem is drift, which results in dc "noise." For the lowest possible noise level, the base-emitter voltage ( $V_{BE}$ ) and  $h_{FE}$  would ideally remain absolutely constant. Thus anything we can do to approach this ideal condition will result in improved performance.

Fig. 6-49A is a schematic diagram of a typical differential amplifier, and Fig. 6-49B is the circuit equivalent to show functional analysis. Transistors Q1 and Q2 draw a very small current, so  $R_E$  can have a very high value. This means a constant-current source. A transistor might provide this source, as we will see shortly.

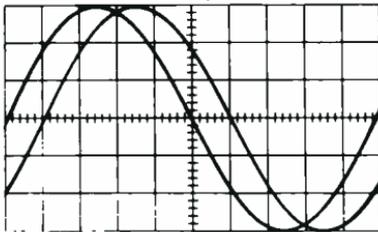
Assume a positive-going signal is applied to the base of Q1 only (the base of Q2 remains at a fixed dc potential). The low-resistance emitter output of Q1 is coupled directly to the low emitter input resistance of Q2. The positive-going signal applied to the emitter of Q2 decreases conduction, causing the collector to go positive (no phase inversion).



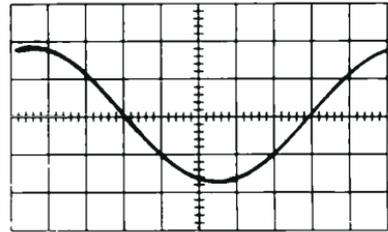
(A) Square waves of unequal amplitudes.



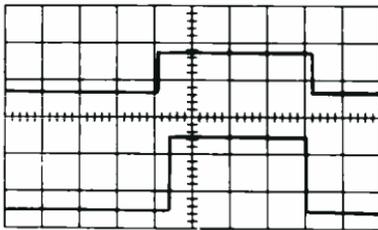
(B) Resultant output for input waveforms in A.



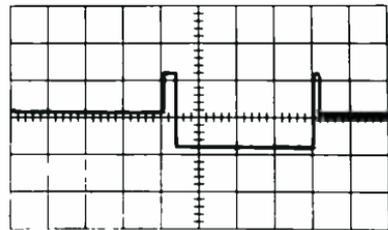
(C) Equal sine waves out of phase.



(D) Resultant output for C inputs.



(E) Rectangular waves, unequal height and width.



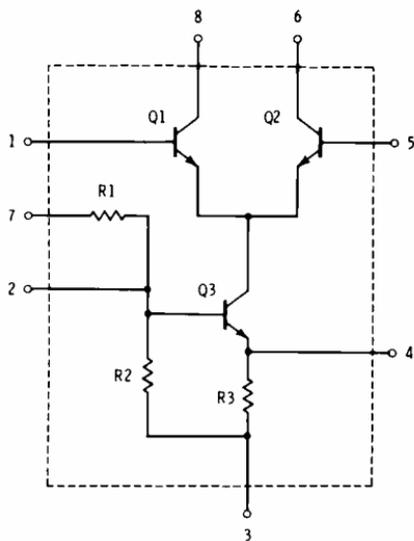
(F) Resultant output for input waveforms in E.

Courtesy Tektronix, Inc.

**Fig. 6-50. Waveforms for a differential amplifier.**

Now assume we make the dc potential applied to the Q2 base more positive. Then Q2 will draw more collector current. This would cancel the effect of the positive-going signal applied to the Q1 base. Here we see that if we applied the same signal to the base of both Q1 and Q2, there would be no change in collector voltage (assuming the  $V_{BE}$  and  $h_{FE}$  of the transistors were exactly matched).

Any difference in the voltage required at the base of Q2 to maintain a balanced condition is termed the *offset voltage*. For example, the Q2 base might require a dc voltage plus or minus 0.5 volt different from that applied to the Q1 base. The operation of the circuit is such that the Q2 collector responds only to the difference between the two base signal inputs.



**Fig. 6-51. Diagram of an integrated circuit (Type CA 3028).**

Courtesy RCA

Fig. 6-50 serves to illustrate the basic function of a differential amplifier. It remains now to explain why this type of circuit is becoming so popular in the audio and video fields as well as in many other applications.

First of all, if the current in one side of the differential amplifier increases, the current in the other decreases by a like amount. Thus drift, or low-frequency "noise," is drastically reduced.

Since a differential amplifier amplifies the difference between two applied signals, an in-phase feedback signal can be applied to one side of the amplifier for negative feedback. There are many other applications, such as agc or other forms of limiting audio amplifiers, in which this property of the differential amplifier may be used to advantage.

Base-emitter voltages tend to cancel each other in the differential amplifier. But any *difference* of base-emitter voltages cannot be distinguished from the signal. So it is important not only that  $V_{BE}$  of the two transistors be matched, but also that each transistor be at the same temperature, since both  $V_{BE}$  and  $h_{FE}$  are temperature-dependent.

The first method of maintaining a matched pair of transistors at the same temperature was by mounting the transistor pellets on isolated islands of a header in a multiple-lead TO-5 package. This method has been largely replaced by the *integrated circuit*, in which transistors, resistors, diodes, etc., are contained in one structure, which might also be enclosed in a TO-5 package. The IC is becoming quite common in new audio and video gear.

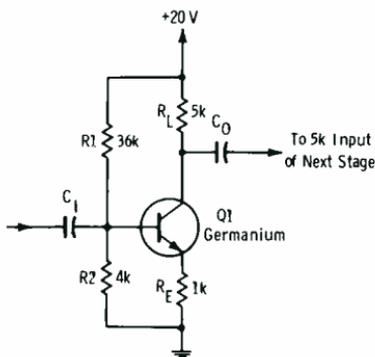
See Fig. 6-51. A *monolithic integrated circuit* is contained in one structure (silicon chip). This construction is different from other forms of integrated circuits that are made up of discrete components on a common substrate (thick-film, thin-film, hybrids, etc.). The circuit is contained in a TO-5 package with 8 pins. The particular circuit in Fig. 6-51 is an RCA CA 3028 IC.

In the circuit of Fig. 6-51, the collector current of Q3 is at very high impedance and is the constant-current source for Q1 and Q2. Note that Q3 is able to be controlled by connection of an external circuit.

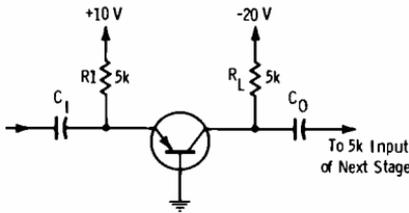
### EXERCISES

Q6-1. In Fig. 6-52, what is the voltage from base to ground?

**Fig. 6-52. Common-emitter circuit for practice problems.**

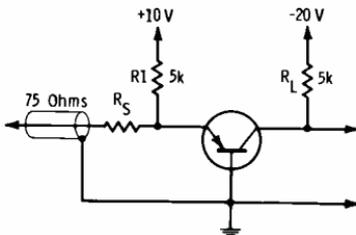


- Q6-2. In Fig. 6-52, what is the emitter current?
- Q6-3. In Fig. 6-52, assuming  $\alpha = 0.98$ , what is the collector current?
- Q6-4. In Fig. 6-52, what values (with respect to ground) of  $V_B$ ,  $V_E$ , and  $V_C$  would you expect to read on an external voltmeter?
- Q6-5. In Fig. 6-52, what is  $V_{CE}$ ?
- Q6-6. In Fig. 6-52, what is the collector power dissipation?
- Q6-7. If  $\alpha$  of Q1 in Fig. 6-52 is 0.98, what is  $\beta$ ?
- Q6-8. What is  $I_B$  in Fig. 6-52?
- Q6-9. What is the input impedance of the transistor in Fig. 6-52?
- Q6-10. What is the load impedance in Fig. 6-52?
- Q6-11. What is the voltage amplification of the circuit in Fig. 6-52?
- Q6-12. What is your short-cut visualization of the voltage-amplification value for Fig. 6-52?



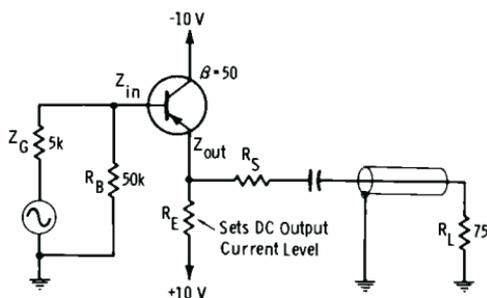
**Fig. 6-53. Common-base circuit for practice problems.**

- Q6-13. Now see Fig. 6-53. The same transistor is used as in Fig. 6-52. What are the emitter current, collector current, input impedance of the transistor, and load impedance?
- Q6-14. What dc  $V_E$  and  $V_C$  would you expect to measure in Fig. 6-53?
- Q6-15. What  $A_v$  should you expect in Fig. 6-53?
- Q6-16. How does the bandwidth of the circuit in Fig. 6-53 compare with that of the circuit in Fig. 6-52?
- Q6-17. What is the main disadvantage of the circuit of Fig. 6-53 compared to the circuit of Fig. 6-52?
- Q6-18. You have seen that the circuit of Fig. 6-53 has much greater voltage gain than the circuit of Fig. 6-52. What are the respective power gains?
- Q6-19. In the circuit of Fig. 6-52, if a large bypass capacitor is placed across  $R_E$ , what are the new values for voltage gain, transistor input impedance, and power gain?
- Q6-20. Does the type of transistor affect the frequency response?
- Q6-21. See Fig. 6-54. This is basically the same circuit as Fig. 6-53, and the same transistor is being used. Series resistor  $R_S$  has been added. If this input is to terminate a 75-ohm line feed, what should be the value of  $R_S$ ?



**Fig. 6-54. Common-base circuit with resistor added.**

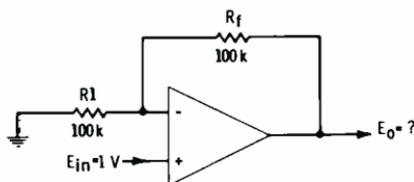
- Q6-22. What is the actual voltage gain of the circuit of Fig. 6-54?
- Q6-23. See Fig. 6-55. For the circuit shown, what should be the value of  $R_S$  if the emitter follower is to "see" a matched impedance?
- Q6-24. What peak-to-peak current would be required in  $R_L$  to develop a 1-volt peak-to-peak signal?



**Fig. 6-55. Emitter-follower circuit for practice problems.**

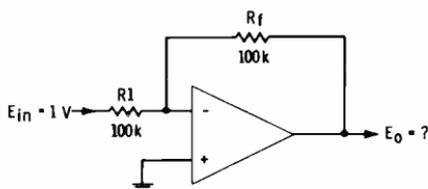
- Q6-25. What value of  $R_E$  should be provided to supply a 20-mA peak-to-peak signal current without clipping?
- Q6-26. If it is desirable to provide an output impedance of a fraction of an ohm, what method would be used?
- Q6-27. What is the most troublesome aspect of the conventional 75-ohm unbalanced distribution of tv systems?
- Q6-28. How do you isolate "glitching" sources?
- Q6-29. See Fig. 6-56. Note that resistors  $R_f$  and  $R_1$  are identical in value. What is  $E_O$  of this noninverting amplifier with  $E_{in} = 1$  volt?

**Fig. 6-56. Noninverting amplifier.**



- Q6-30. See Fig. 6-57. Resistors  $R_f$  and  $R_1$  are again identical in value. What is  $E_O$  of this inverting amplifier with  $E_{in} = 1$  volt?

**Fig. 6-57. Inverting amplifier.**

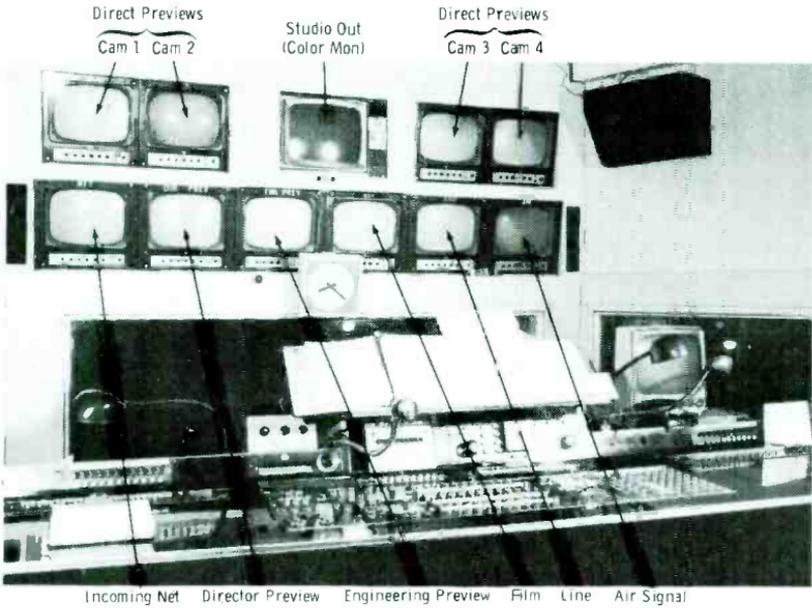


## The Picture and Waveform Monitoring System

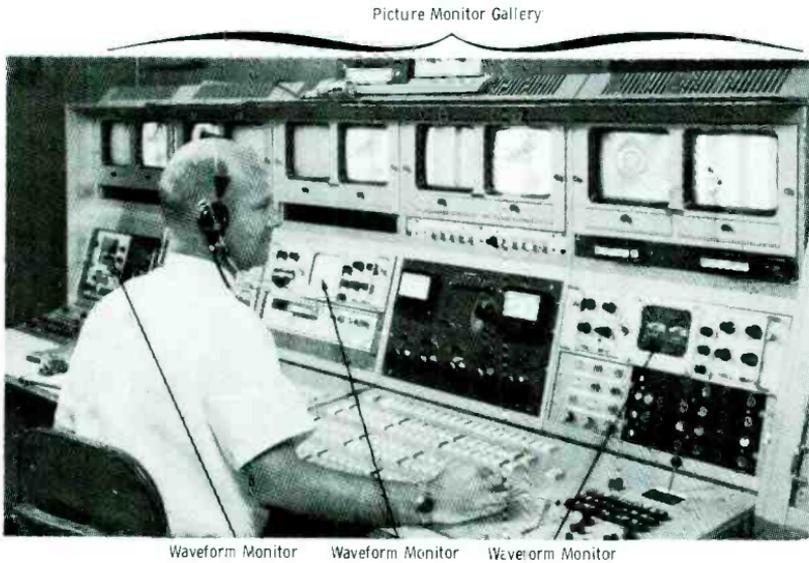
Picture and waveform monitors, and the monitoring system, constitute a sizeable part of the total studio installation. An appreciable allotment of maintenance time is normally required for this equipment.

Fig. 7-1 illustrates a typical monitor "gallery" as installed at a studio switcher position. The two switchable monitors, director preview and engineering preview, are handled as a part of the switcher system (Chapter 5). The camera-preview monitors in this particular installation are fed directly from the camera DAs. The incoming-network, line-output, and air-signal (off-air receiver) monitors are similarly fed directly from the respective DAs. An automatic preview crosspoint "fired" by the roll button causes the film monitor to display the film-camera output selected on the switcher as soon as the source is "rolled." The color monitor in the center of the top row is used to display the studio output for the switcher in use.

Fig. 7-2 shows another general monitoring arrangement, for a switcher position which also includes waveform monitors. Each camera-control position has an individual picture monitor and a waveform monitor, which includes a switching arrangement to permit observation of each color channel alone, all channels together, or the encoded output, as desired. Because of this, many switcher monitoring galleries do not include waveform monitors. Picture and waveform monitors also are normally included at some point to permit observing the line output, incoming network, and remote signals, and (when used) stl monitor returns. In addition to these general requirements in the control area, individual studio monitors must be placed in each studio, and a "house monitoring" system is



**Fig. 7-1. Typical monitor "gallery" at studio switcher position.**



Courtesy Cohu, Inc.  
**Fig. 7-2. Switcher position with both picture and waveform monitors.**

included for providing video and sound feeds to the offices and other areas around the building.

Monitor and receiver servicing is outside the scope of this text. We will cover only those features that are more common to broadcast applications than to home or industrial use. Basic picture-tube convergence procedures and specialized line-selection circuitry in waveform monitors will be covered.

## 7-1. THE COLOR PICTURE MONITOR

**NOTE:** The I-Q and color-difference forms of color demodulation are covered in a preceding volume in this series, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*. It remains to examine here the more advanced techniques and circuitry affecting monitor adjustments and maintenance.

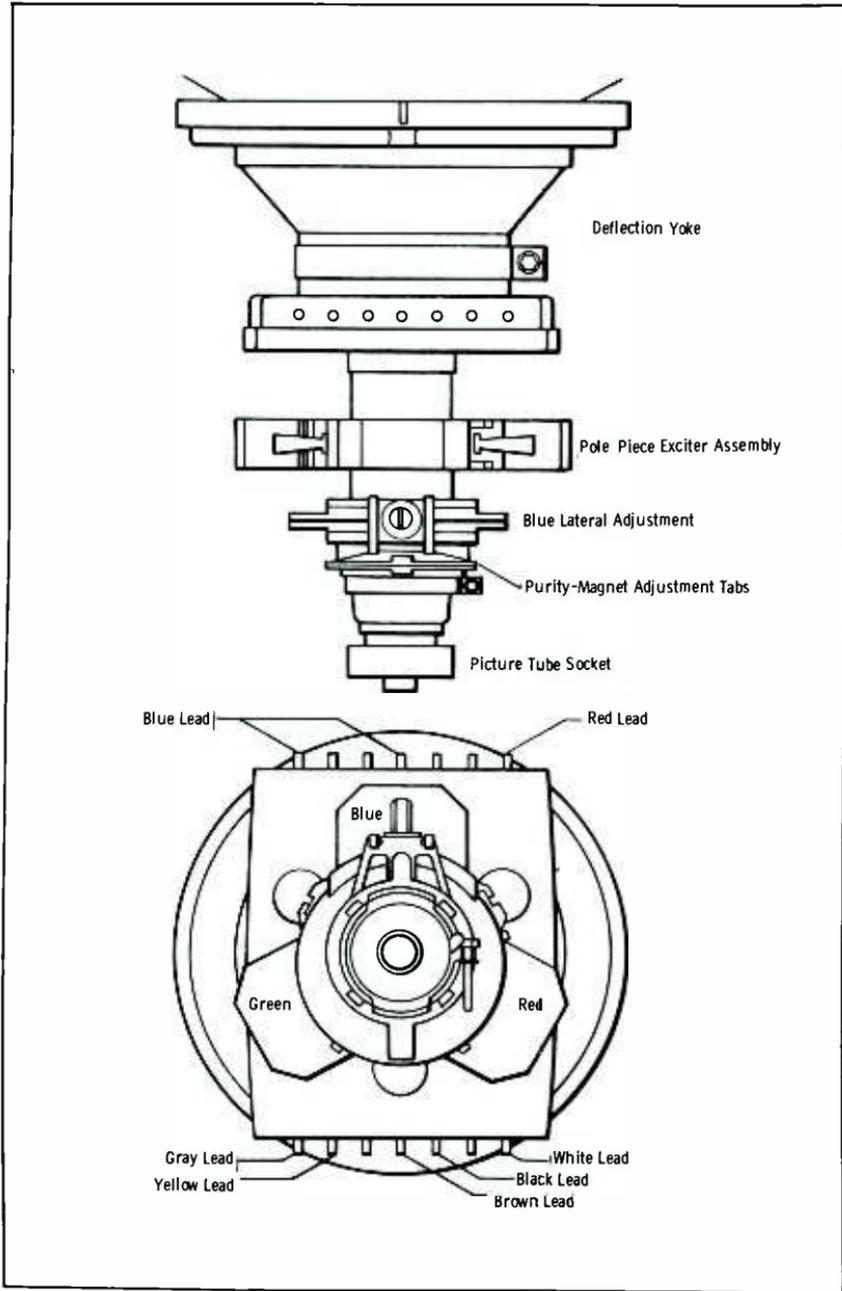
The techniques for aligning the color picture tube are fairly common to all makes of broadcast-type monitors, and a description of one monitor such as the Miratel MC-19 will suffice for our purpose.

Before actual color-picture alignment, it is always advisable to check and adjust (if necessary) the high voltage. **CAUTION:** Always remember to use extreme caution when the back of the color set is removed. The 25,000-volt regulated supply can be lethal! For high-voltage adjustment, follow the particular manufacturer's instructions, because many types of controls may be found here. In the simplest type, the high-voltage control is varied to obtain the specified ultor voltage (usually around 25 kV) with the brightness and kine-bias controls fully counterclockwise. A vvm with a high-voltage probe is required to check the ultor voltage on the kine-scope.

### Purity Adjustment

Purity-adjustment procedures for the Miratel MC-19 are as follows:

1. Verify that the BG1 (blue grid) and GG1 (green grid) switches are placed in the off position, since the red field is generally used to check purity.
2. Loosen the yoke clamp, and slide the yoke to the rear of the set.
3. Note the red field or fireball in the center of the screen.
4. If this field is not centered, adjust the purity-magnet adjustment tabs (Fig. 7-3) until it is. Slowly move the yoke forward, and simultaneously adjust the purity tabs in order to keep the red area in the center of the raster.
5. Continue this procedure until a pure red field completely covers the face of the picture tube. Be careful not to move the



Courtesy Ball Bros. Research Corp.

**Fig. 7-3. Convergence assembly, Miratel MC-19 color monitor.**

purity magnet from its correct position (in line with the gap between grids 3 and 4) on the neck of the tube.

### **Static-Convergence Adjustment**

Static convergence is accomplished by using the static controls to position the electron beams in the center of the raster. The static controls are located next to the yoke in the pole-piece exciter assembly. This assembly (Fig. 7-3) is shaped like an inverted Y. The blue static adjustment is located in the top leg, the red static adjustment is located in the lower right leg, and the green static adjustment is in the lower left leg. Located between the pole-piece exciter assembly and the purity-magnet adjustment tab is the blue lateral-adjustment control. Static convergence is adjusted as follows:

1. With a crosshatch video signal applied to the monitor and the BG1 switch turned to the off position, adjust the red and green static controls until the crossbars converge in the center of the raster.
2. Switch RG1 off, turn on BG1, and adjust the blue static control and the blue lateral magnet to converge the blue and green crossbars at the center of the raster.

In other areas of the raster, convergence becomes a dynamic process. Since dynamic convergence affects the static convergence, static convergence is necessarily touched up during the course of dynamic convergence.

### **Dynamic-Convergence Adjustment**

Dynamic convergence is adjusted as follows:

1. Refer to Figs. 7-4 through 7-9.
2. First adjust the red and green controls.
3. Continue with the blue adjustments.

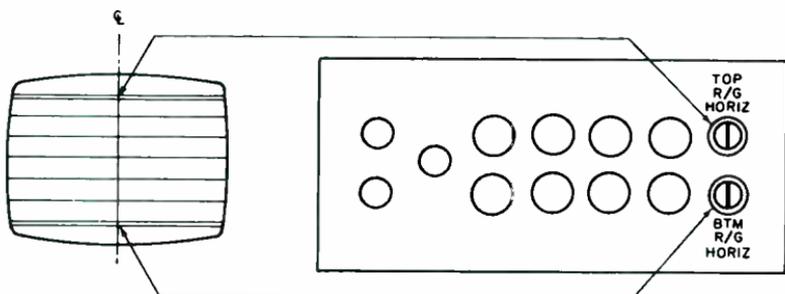
### **Pincushion Adjustment**

The pincushion-adjustment controls are located at the rear of the monitor alongside the high-voltage cage. They are operated as follows:

1. Use the phase control to adjust the tilt at the top and bottom horizontal lines.
2. Use the amplitude control to adjust the amount of pincushion correction applied to both the top and bottom lines.

### **Underscan Convergence**

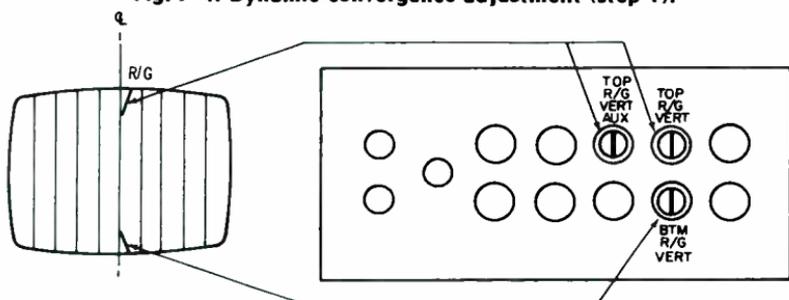
This adjustment procedure applies only to the deluxe series monitors. The procedure is as follows:



- A. Turn off blue beam.
- B. Converge top red and green horizontal bars at center line.
- C. Converge bottom red and green horizontal bars at center line.

Courtesy Ball Bros. Research Corp.

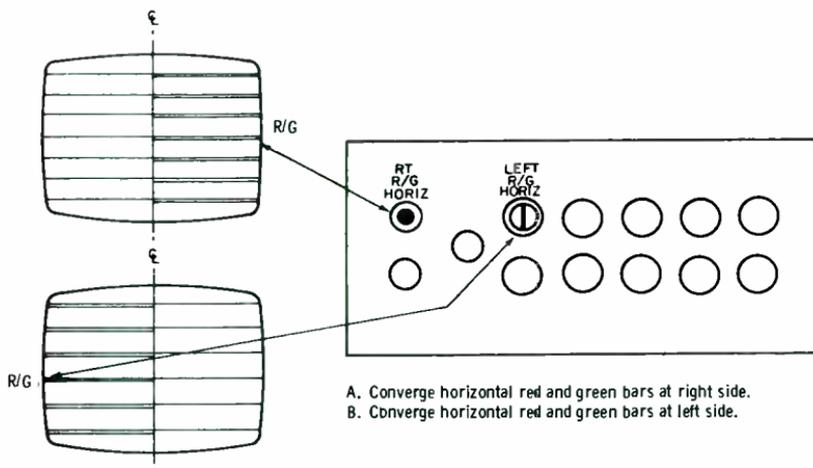
**Fig. 7-4. Dynamic-convergence adjustment (step 1).**



- A. Converge vertical red and green bars at top center.
- B. Converge vertical red and green bars at bottom center.

Courtesy Ball Bros. Research Corp.

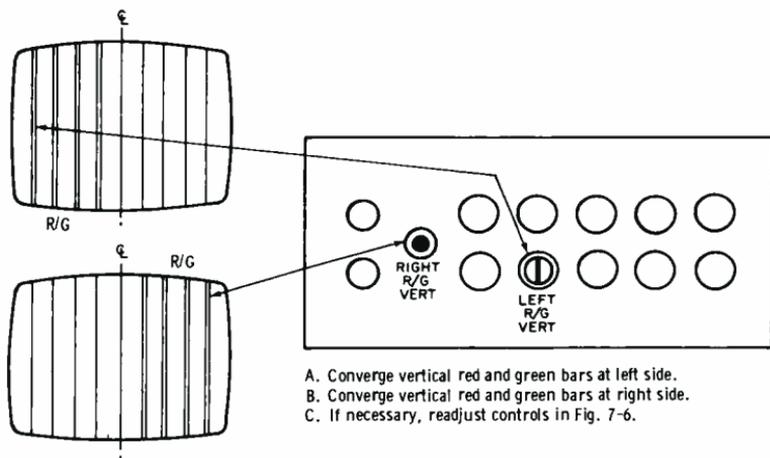
**Fig. 7-5. Dynamic-convergence adjustment (step 2).**



- A. Converge horizontal red and green bars at right side.
- B. Converge horizontal red and green bars at left side.

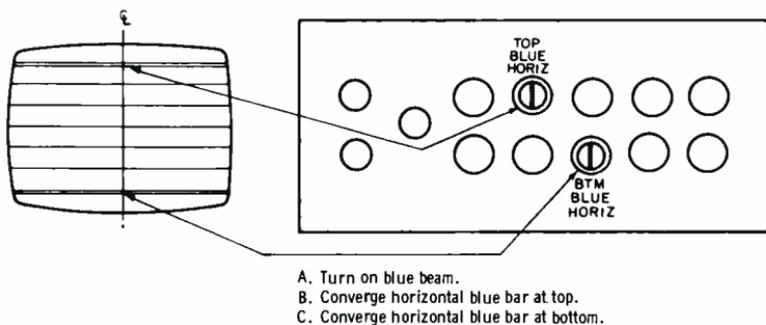
Courtesy Ball Bros. Research Corp.

**Fig. 7-6. Dynamic-convergence adjustment (step 3).**



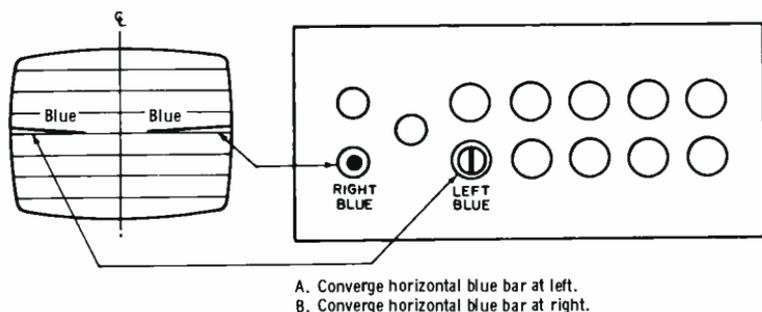
Courtesy Ball Bros. Research Corp.

**Fig. 7-7. Dynamic-convergence adjustment (step 4).**



Courtesy Ball Bros. Research Corp.

**Fig. 7-8. Dynamic-convergence adjustment (step 5).**



Courtesy Ball Bros. Research Corp.

**Fig. 7-9. Dynamic-convergence adjustment (step 6).**

1. Turn the underscan switch to the underscan position.
2. Adjust the width control for the correct aspect ratio, and adjust the auxiliary vertical-hold control to obtain vertical stabilization of the picture.
3. Turn off the blue beam, and apply a crossbar pattern.
4. Adjust the (red-green) vertical auxiliary control to converge the red and green bars at the top of the raster.
5. Turn on the blue gun, and adjust the left blue and right blue horizontal controls as shown in Fig. 7-9.

### **Color-Killer Adjustment**

The color-killer control is adjusted as follows:

1. With no signal applied to the monitor, turn the chroma control fully clockwise.
2. Adjust the color-killer control until the colored snow on the raster completely disappears.

### **Color-Temperature Adjustment**

The color-temperature adjustments are made as follows:

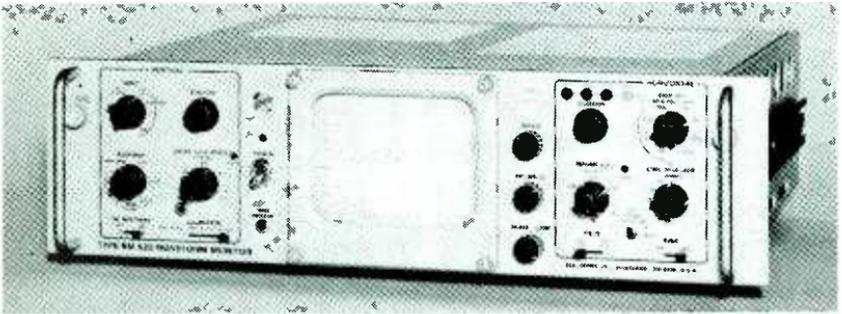
1. With no signal applied to the monitor, adjust the chroma, kine-bias, and red, green, and blue screen controls fully counterclockwise.
2. Turn the green and blue cathode-drive controls fully clockwise and place the service/normal switch in the service position.
3. Advance the red screen control fully clockwise and then back off approximately 20 percent. Adjust the kine-bias control until a red line is just visible.
4. Adjust the green and blue screen controls until the green and blue lines are also just visible.
5. Return the service/normal switch to the normal position.
6. To check the black-and-white tracking of the monitor, a video signal is applied to the monitor. A good black-and-white picture should be maintained within the normal range of both the contrast and brightness controls. If proper tracking is not obtained, alternately adjust the green and blue drive controls for the best black-and-white picture.
7. Turn the brightness control fully clockwise and adjust the kine-bias control to the point just before the raster starts to bloom.

## **7-2. THE WAVEFORM MONITOR**

The maintenance technician needs to be familiar with the specialized circuitry of modern waveform monitors. The most important

features include the "brightening pulse and video" output (see Section 3-5), the special type of dc restoration, and the method employed to select a given line of a total composite video signal. This specialized circuitry is in addition to the ultrastable video amplifier, with which the reader is assumed to be acquainted.

Due to the extensive use of the Tektronix Type 529 waveform monitor at the time of this writing, this monitor will be described. (The description is courtesy of Tektronix, Inc.) Any later model has circuitry that is similar to the circuitry of this unit; hence, it is important to become familiar with the theory of operation of the Type 529. The Type 529 is a hybrid model, employing some vacuum tubes in conjunction with solid-state circuitry. A front view of the unit is shown in Fig. 7-10.



Courtesy Tektronix, Inc.

Fig. 7-10. Tektronix Type 529 waveform monitor.

### Video Output Amplifier

The video output amplifier (Fig. 7-11) receives its signal from the emitter of Q114 (output of preamplifier). The input impedance

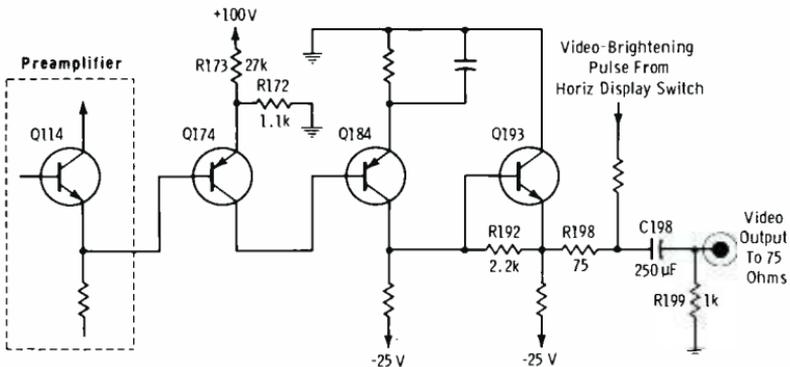


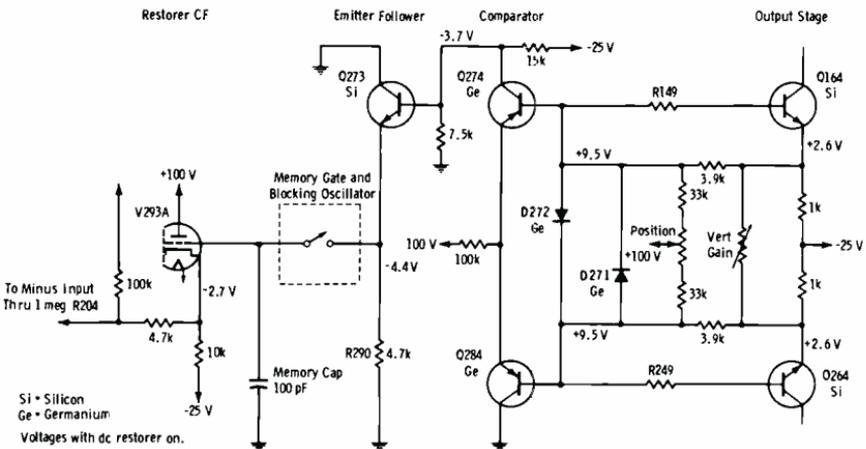
Fig. 7-11. Video amplifier for picture-monitor display.

of Q174 is greater than 50k due to the large amount of emitter degeneration, and Q114 is loaded lightly. Transistor Q174 produces an inverted current drive of essentially 1 mA per volt to feedback amplifier Q184-Q193. The emitter impedance of Q174 is R172 and R173 in parallel, or 1.06k. Since the collector current of Q174 changes linearly with the base-voltage drive, the signal-source resistance of the Q184 base is 1.06k. Then, the voltage gain of Q184 and Q193 (set by the feedback resistance divided by the input resistance) is 2.2k/1.06k, or 2.06. Assuming 3 percent signal loss in the preamplifier, the gain from the input to the Q193 emitter is 2. The feedback (R192) makes the video output amplifier have a very low output impedance. Thus, the amplifier is matched to the output coaxial line by inserting a series 75-ohm resistor (R198). This resistor in series with a 75-ohm load makes the overall gain 1. The reactance of C198 is so low at the frequencies passed by the circuit that it does not increase the output impedance over the resistance of R198. Resistor R199 assures that C198 is charged properly.

The sweep generator applies a positive pulse to the video output jack in the line-selector mode of operation. The intensifying-pulse amplitude is about 0.2 volt at the video output jack (0.1 volt when loaded with 75 ohms). The value of the pulse is set by current from a +100-volt pulse in the sweep-generator unblanking circuit.

**Keyed DC Restorer**

The keyed dc-restorer circuit includes the dc-restorer comparator, the blocking oscillator, and the restorer cathode follower. Voltages and waveforms on the dc-restorer diagrams were taken with the dc-restorer circuit on. The simplified diagrams of Figs. 7-12, 7-13,



**Fig. 7-12. Simplified keyed dc-restorer circuit.**

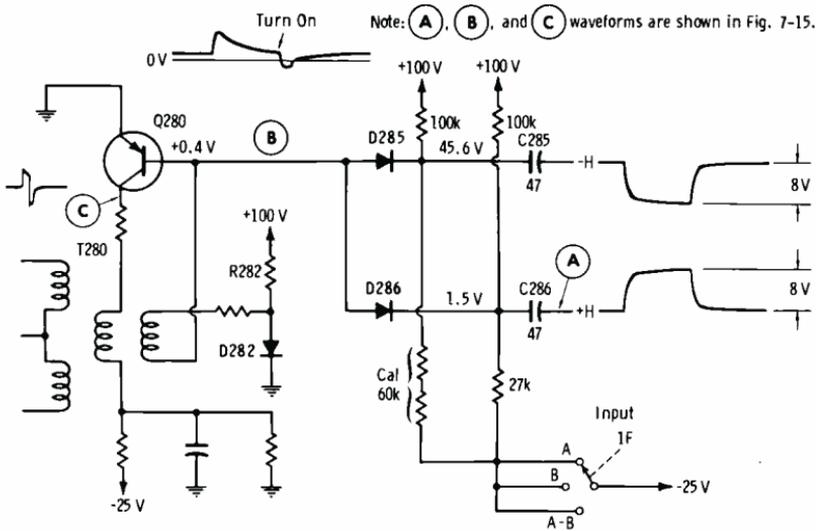


Fig. 7-13. Operation of diode switch feeding blocking oscillator.

and 7-14 and the waveforms of Fig. 7-15 will help during the following discussion of the dc-restorer circuit operation.

The dc-restorer comparator is a dual-input, single-output amplifier that amplifies a small part of the composite video which is normally near the crt screen center. The signal peaks are limited by parallel back-to-back germanium diodes D271 and D272 (Fig. 7-12). One of the two diodes conducts whenever the voltage across

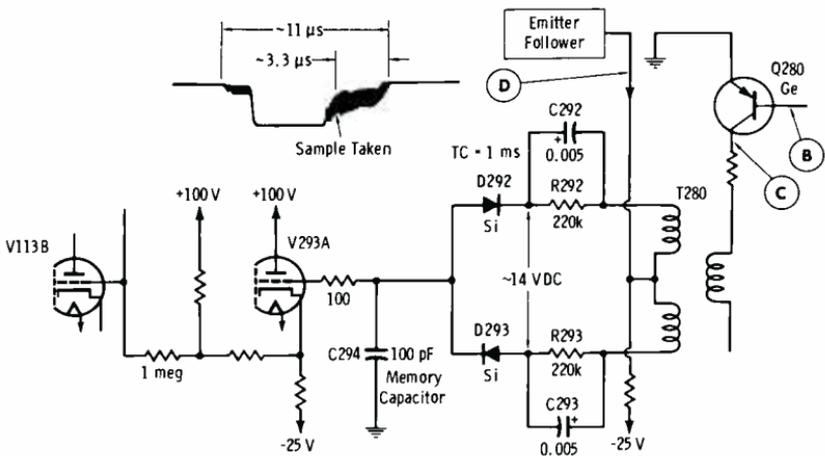
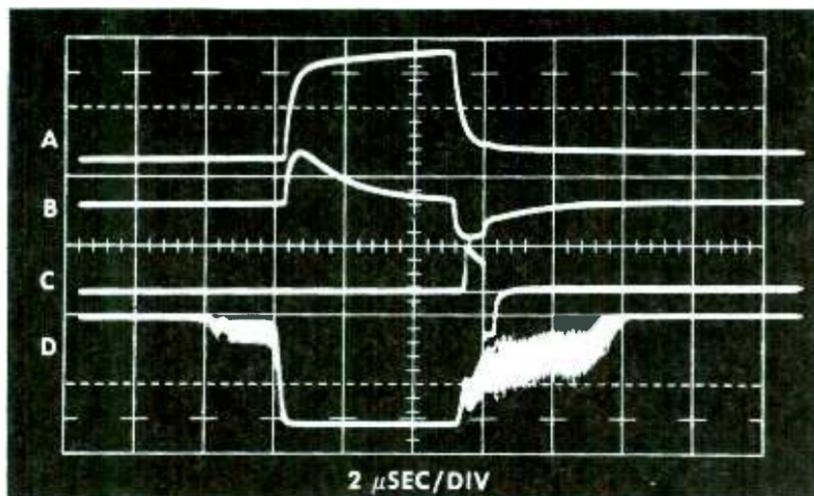


Fig. 7-14. Sampling and memory portions of dc-restorer circuit.



Courtesy Tektronix, Inc.

**Fig. 7-15. Time-coincident waveforms of keyed dc restorer.**

them exceeds 0.2 to 0.3 volt. When the diodes are not conducting, the two bases of the comparator amplifier are essentially shorted together, and there is no gain. When the signal at the base of Q274 goes negative (and the signal at the base of Q284 goes positive), D271 conducts. The opposite signal polarity causes D272 to conduct. The peak-to-peak signal input to the comparator is, therefore, limited to about 0.45 volt. The comparator output signal is typically 4.2 volts peak to peak at the emitter of the emitter follower (Q273). The comparator functions in the same manner as a differential stage. Waveform D in Fig. 7-15 shows the signal as it enters the memory gate. The other waveforms of Fig. 7-15 are discussed below in connection with the blocking oscillator.

The blocking oscillator (Fig. 7-13) normally is biased to cutoff by current through R282 and D282. A negative pulse applied to the base of Q280 causes it to conduct and go through one cycle of oscillation. The negative pulse arrives at the base of Q280 through a diode switching network from the sync regenerator.

Assume the vertical INPUT switch is at position A, as shown in Fig. 7-13. The diode switching circuit then reverse biases D285 about 45 volts and reverse biases D286 about 1 volt. Regenerated horizontal-sync pulses arriving at C285 and C286 have a peak-to-peak amplitude of about 8 volts. Thus, the negative pulse will not cause D285 to conduct, but the positive pulse will cause D286 to conduct at the end of the pulse as it goes negative. Capacitor C286 charges almost the full 8 volts of the rise of the positive pulse, and

then causes the cathode of D286 to fall about 7 volts. The base of Q280 does not fall as much as 8 volts because of the low base-to-emitter impedance at the time of turn-on.

The blocking oscillator turns on at the time the composite-video horizontal-sync pulse starts to change toward the back-porch level. The back-porch level occurs about  $0.25 \mu\text{s}$  later, so the blocking oscillator does not sample the video at the time of turn-on. As Q280 conducts, the L/R time constant of the collector-circuit inductance and resistance keeps Q280 conducting for about  $0.6 \mu\text{s}$  before the base drive decays. As the regenerative base turn-on drive to Q280 is stopped, the collector current drops rapidly. The dropping collector current is transformer-coupled back to the base of Q280 as a fast turn-off signal. The collector-voltage change is shown by waveform C in Fig. 7-15. The changing collector current is also transformer-coupled to the tapped secondary that supplies drive to the memory gate.

The memory gate (Fig. 7-14) consists of the center-tapped secondary of T280 (with its dc level set by the emitter follower), two parallel RC circuits, and two silicon diodes. As Q280 is turned on, the secondary voltage of T280 reverse-biases D292 and D293, and there is no current. As Q280 stops conducting, the secondary of T280 drives D292 and D293 into conduction and at the same time charges both C292 and C293 to about 7 volts. (The energy for charging the two capacitors comes from the core of T280.) Diodes D292 and D293 are computer diodes with equal forward drop at the time they conduct, making the voltage at their junction with C294 equal to the voltage at the center of the T280 secondary. Thus, during the backswing time of T280, the voltage at the center of the T280 secondary appears at C294. The backswing lasts for about  $0.4 \mu\text{s}$ , which is plenty of time to charge C294 to the voltage of the emitter follower. During each backswing, C292 and C293 are recharged to replace the small amount of voltage discharged by resistors R292 and R293. The time constant of the two RC circuits is 1 ms each, many times longer than the  $63.5 \mu\text{s}$  between sampling times.

The memory circuit consists of memory capacitor C294 and the grid of V293A. Since D292 and D293 are silicon diodes with very low reverse leakage, the charge on the memory capacitor is essentially constant between samples.

The four waveforms of Fig. 7-15 (taken with a four-channel oscilloscope) show the time relationships of the waveforms of the dc-restorer circuit. Note that some of the color burst is seen by the memory gate and memory capacitor. The reactance of the memory capacitor is about 440 ohms at 3.58 MHz, and the sampling diodes are turned on hard by the T280 secondary, so there is no color-burst rectification. Thus, the dc-restorer circuit acts on the average voltage

of the color burst, keeping the display at the same stable position with or without the color burst.

The restorer cathode follower acts as an impedance transformer that couples the sampled and stored voltage into the vertical-amplifier minus input grid.

The preceding information on the vertical amplifier and keyed dc restorer serves to emphasize the sophisticated type of circuitry found in modern waveform monitors. At this point, we have covered the "fast" type of keyed clamp as used in video-processing amplifiers to remove all hum components, and the "slow" clamp used in waveform monitors. The latter serves to hold the waveform at a fixed reference for black regardless of APL changes, while still displaying on the trace any hum component present in the signal. Also, we have examined a back-porch clamping circuit that functions without distorting the color burst, and without separation of luminance and chrominance components.

### Field and Line Selection

Of importance also are the "field-1 recognition" circuitry and the rather complex circuitry associated with the VIT signal display. A number of modifications have been made, depending on serial number, and it is vital for the maintenance engineer to acquaint himself with his particular equipment. The remote control of tv transmitters requires use of local test signals in the vertical interval (always check current FCC Rules), and the performance of the waveform monitor for this function is extremely important.

We will cover just the basic functions of the Type 529 monitor in this respect, and outline the Tektronix method of fundamental adjustment. Again, we must emphasize that much more detailed information (as contained in the instruction book for a particular instrument) is necessary for a thorough understanding of the selection of specific line displays in the vertical interval of the signal. Observe Fig. 7-16 during the following discussion.

*Vertical-Sync Separator*—The vertical-sync separator is primarily a differentiator and an amplifier biased off about 3 volts. The differentiator, C341-R341, shifts the signal-output dc level depending on the pulse duty cycle. If the negative peaks of the incoming rectangular-wave signal are of longer duration than the positive peaks, the output will be more positive than negative. Likewise, if the positive peaks of the incoming signal are of longer duration than the negative peaks (vertical sync), the output will be more negative than positive. During the time C341 receives only regenerated horizontal-sync pulses, the signal at the base of Q344 shifts between about 2 volts negative and 6.5 volts positive, and Q344 does not conduct. As the vertical-sync group occurs, the signal at the base



of Q344 shifts to between about 6 volts negative and 2.5 volts positive, turning Q344 on hard below about  $-3.5$  volts.

Both the field-1 recognition circuit and the field-trigger generator require a single pulse, the first of the vertical-sync pulses. The parallel combination of R346 and D346 and the capacitance of C347 allow only the first vertical pulse to pass through. As the collector of Q344 rises 20 volts in the positive direction, D346 passes the whole pulse to C347 (and C351). Capacitor C347 charges to more than half the peak voltage of the first pulse. The very high reverse resistance of D346 and the 22 megohms of R346 let C347 hold this charge. The second vertical-sync pulse is, thus, not passed since the cathode of D346 is already several volts more positive than it was before the arrival of the first pulse. (Some of the second pulse gets through the coupling capacitors, but does not affect the following circuits.) Resistor R346 discharges C347 (and C351) before the next vertical-pulse group arrives.

*Field-1 Recognition*—The field-1 recognition circuit is a one-shot multivibrator (monostable) with two input paths. The single vertical-sync pulse that arrives at the base of Q355 through C351 turns Q355 on and Q365 off. The switching action is regenerative due to emitter coupling, and coupling from the Q355 collector to the Q365 base. Capacitor C360 was initially charged to about 13 volts. As the voltage at the collector of Q355 falls, the base of Q365 is taken about 11 volts negative. The FIELD 1 SYNC control (R360), in series with R361, starts the base of Q365 back toward ground, discharging C360. As the voltage at the base of Q365 nears a value that would cause Q365 to turn on, a positive pulse coupled to the base through C361 will turn it on and reset the multivibrator. The time constant of C360 and R360-R361 is set such that the multivibrator changes state at the end of the last vertical-equalizing pulse. As the C365 collector goes negative at the end of the last equalizing pulse, C364 and R368 form a negative pulse that ramps up (changes linearly in the positive-going direction) for a period of 50 to 55  $\mu$ s. Capacitor C370 couples differentiated negative horizontal-sync pulses and adds them to the ramp at the junction of R369 and D370. If a horizontal-sync pulse occurs during the time the ramp is rising, the output through D370 is more negative than at any other time. A two-field interlace horizontal-sync pulse occurs in the middle of every other ramp.

All the other negative pulses at the junction of R369 and D370 charge C371 (through R372) to an essentially stable dc voltage (R371 does not appreciably discharge C371 between pulses). As field 2 occurs, the more negative pulse that coincides with the field-1 recognition ramp is coupled through D371 to the base of Q375, flipping the field-trigger generator so that Q375 conducts.

*Field-Trigger Generator*—The field-trigger generator is a bistable multivibrator that changes state each time a positive pulse arrives through C347 from the vertical-sync separator. The triggering pulse is coupled to the bases of Q375 and Q385 through diodes D374 and D384 and RC networks R375-C375 and R385-C385. The positive-polarity pulse turns off the conducting transistor regardless of which is conducting. The pulses arrive at a 60-hertz rate, causing the field-trigger generator to have a 30-hertz output rate at each collector.

If Q375 is off when the negative-going field-1 signal from the field-1 recognition circuit arrives at the base of Q375, then Q375 will be turned on. A positive signal from the sync separator to each collector at the start of each field, and a negative pulse to the Q375 base at each field 1 assures that the output of the field-trigger generator is always related to field 1 and field 2 of the composite video. The collector of Q375 always goes positive (toward ground) at the beginning of field 1. The collector of Q385 always goes positive at the beginning of field 2.

Field-shift switching is the selection of the correct output pulse from the field-trigger generator by a dual-input, single-output diode switch. Positive-going trigger pulses are needed by the sweep generator and by the delay generator. Thus, to trigger on field 1, the field-shift switch causes the collector signal from Q375 to be coupled on, and for field 2, it causes the collector signal from Q385 to be coupled on.

Assume a field-2 trigger is selected. The crt display will start the sweep on field 1 and show field 2 at center screen. The field switch (set at 2) applies a negative bias to the anode of D377, assuring that this diode cannot conduct the signal from the Q375 collector to the following circuits. Diode D387 will pass the positive portion of the differentiated signal from the Q385 collector. Differentiation of the Q385 collector signal is done by C384 and the parallel resistance of R388. As the voltage at the Q385 collector rises, C384 couples the first of the full step through to D387 and the rest of the circuit. Capacitor C384 soon charges, dropping the voltage at the cathode of D387 back to ground level. As the voltage at the Q385 collector falls, D387 disconnects the signal from the rest of the circuit, and R388 recharges C384 for the next positive pulse.

*Line Selector*—The line-selector circuit (Fig. 7-17) includes the delay generator and the line-pickoff circuit. The selected line-trigger pulses occur once each field, an adjustable amount of time after each vertical-sync pulse.

The delay generator is normally biased so that Q425 is conducting and Q415 is cut off. The collector of Q415 rests at +24 volts because of voltage divider R417, R418, R419, and Q434. (The LINE SELECTOR control current from -25 volts through R402 to the junc-

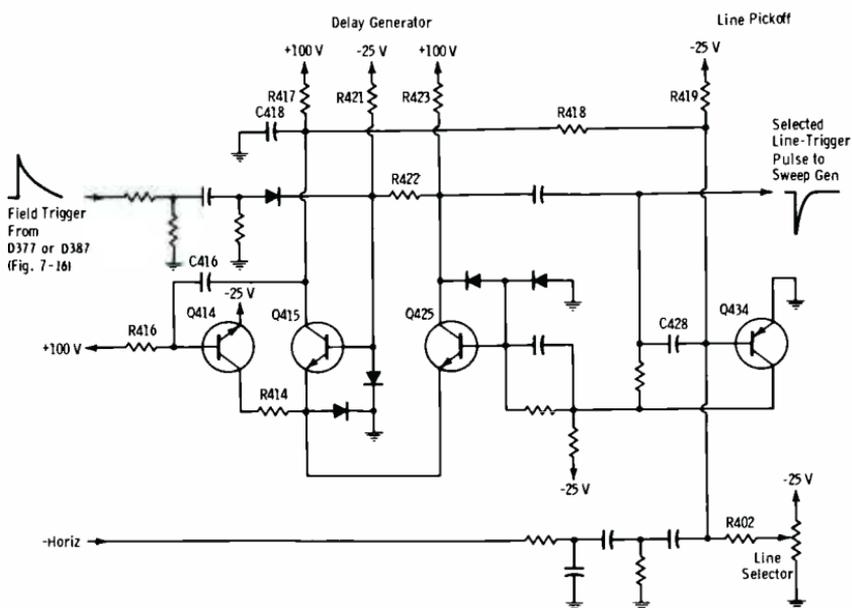


Fig. 7-17. Simplified diagram of line selector.

tion of R419 and C428 does not take the junction significantly below ground because there is 24 volts across R419 and there can be up to 25 volts across R402.) The collector of Q425 rests at  $-0.65$  volt, holding the Q415 base at about  $-6$  volts by the drop across R421 and R422. Transistor Q414 is saturated (collector voltage very near emitter voltage) because of the base current through R416. Thus, the current in the Q425 emitter is set by R414 and the  $-24$  volts at the Q414 collector, and the current in the Q425 collector is set by R423. The delay generator remains in this condition until a positive trigger pulse arrives at the base of Q415.

As a positive pulse turns on Q415, the current through Q414 and R414 shifts to Q415, and Q425 turns off. As the Q415 collector starts negative, C416 couples the voltage change into the base of Q414 in a direction to reduce its collector current. As the base of Q414 is taken far enough negative to almost turn off its collector current, the drop in Q415 collector voltage is nearly stopped until the current through R416 discharges C416. As C416 discharges slightly, the base voltage of Q414 turns on slightly more current. The current of Q414 is also the current in Q415, which again tends to discharge C416. The result of the feedback just described is that C416 is discharged in a linear manner by current through R416. The voltage at the junction of C416 and R416 remains essentially constant while

the collector of Q415 pulls the other side of C416 negative at a rate set by the current through R416.

When the collector voltage of Q415 reaches ground level, the current through R416 raises the base voltage of Q414 and increases its current. Increased current in Q414 pulls the emitter and base elements of Q415 negative, allowing the collector to go negative also. The common emitter-to-emitter lead of Q425 and Q415 drops negative until Q425 again turns on. The collector voltage of Q425 drops and quickly turns Q415 off, letting its collector voltage rise in the positive direction as R417 charges C416 to its original state. Capacitor C418 cancels the initial negative collector surge of Q415 caused by shifting the Q414 current from about 5 mA in Q425 to about 0.27 mA in Q415. Without C418, the collector voltage of Q415 would drop sharply negative at the time Q415 is triggered into conduction. The charging current of C416 is limited only by R417 since the negative end of the capacitor is tied to -25 volts through the Q414 base-emitter junction.

### Sweep Generator

The sweep generator is a triggered-sweep system for all positions of the DISPLAY switch except the 2-field position, for which a recurring synchronized sweep is produced. Positive field-trigger pulses stop the sweep for the 2-field and both line-selector positions (line selector set to a line from 16 through 21); negative line-trigger or selected line pulses start the sweep in all other modes of operation. The sweep rate for the 2-line position permits viewing the interlacing of the color burst. The sweep rate at 0.125H/cm lets you look at every other horizontal line and observe a noninterlaced color burst.

For the 2-line display (vertical INPUT switch at A, DISPLAY switch at 2-line), the sweep generator is held in one state by the sweep-gating multivibrator until the arrival of a negative trigger pulse. For the 2-field display, operation of the sweep-gating multivibrator is changed from a state in which a trigger is required to start the sweep, to a condition in which the sweep starts automatically and is stopped by a trigger pulse.

### Field-1 Sync Adjustment

The field-1 sync adjustment is made as follows:

1. Connect a 1-volt composite video signal to the right VIDEO INPUTS A connector.
2. Be sure that a 75-ohm termination is connected to the left VIDEO INPUTS A connector.

3. Set the controls of an external test oscilloscope as follows:
  - Sweep Rate:  $50 \mu\text{s}/\text{Cm}$
  - Triggering: +Ext, AC
  - Signal Input Coupling: AC
  - Vertical Deflection Factor (with  $10\times$  Attenuator Probe):  
 $5 \text{ Volts}/\text{Cm}$
4. Connect the  $10\times$  probe from the vertical-input connector of the external oscilloscope to the junction of D370 and C370 (Fig. 7-16).
5. Connect the  $1\times$  probe from the external-trigger input connector (external scope) to test point TP344 (Fig. 7-16).
6. Observe whether the ramp starts just after the sixth equalizing pulse (Fig. 7-18A).
7. Adjust by first rotating the FIELD 1 SYNC control (R360 in Fig. 7-16) fully counterclockwise. Then, rotate the control slowly in a clockwise direction until the ramp starts just after the sixth equalizing pulse. (The sixth equalizing pulse is the last pulse in the equalizing-pulse group following the vertical-sync pulses.) Figs. 7-18B and 7-18C show two possible displays that can be obtained when the control is adjusted incorrectly.

### Check Field Switch

Check the field switch as follows:

1. Use the test-equipment setup given above.
2. Set the FIELD switch to One.

NOTE: If the vertical-sync pulse is not completely displayed, position it to the horizontal center of the viewing area with the (HORIZ) POSITION control.

3. Check for a horizontal distance of  $1H$  ( $63.5 \mu\text{s}$ ) between the last horizontal-sync pulse and the first equalizing pulse (Fig. 7-19A). This indicates that field 1 is displayed at the center of the crt.
4. Set the SYNC switch to Ext and back to Int several times to check that the field-1 display is obtained each time the SYNC switch is returned to the Int position. This is a double check. A stable display of field 1 means that R360 is adjusted properly.
5. Set the FIELD switch to Two.
6. Check for  $0.5H$  ( $31.75 \mu\text{s}$ ) horizontal distance between the last horizontal-sync pulse and the first equalizing pulse (Fig. 7-19B). This indicates that field 2 is displayed at the center of the crt.

7. Repeat step 4, except check that a field-2 display is obtained each time the SYNC switch is returned to Int.

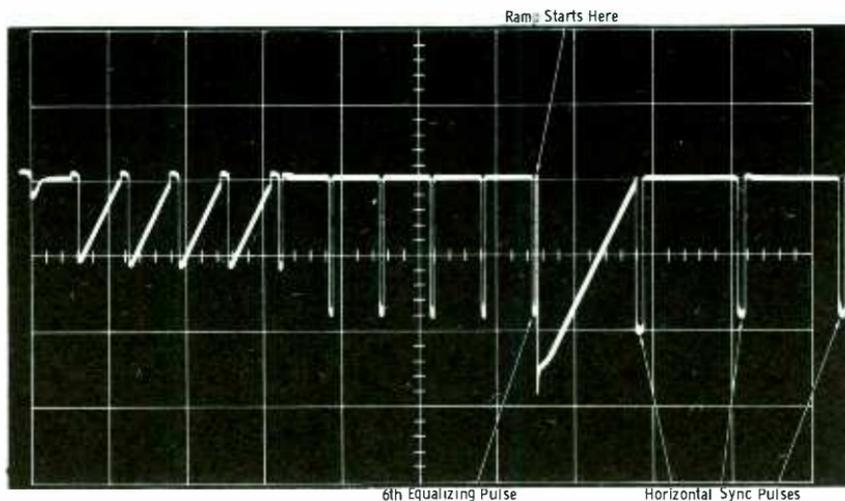
### Check Line-Selector Operation

The brightening-pulse amplitude must be 100 mV or more. The start of the brightening pulse corresponds to the start of the sweep, as observed on the Type 529 crt, and the width of the brightening pulse corresponds to the length of the sweep, as observed on the Type 529 crt.

With the LINE SELECTOR switch set to Variable and the LINE SELECTOR variable control set fully counterclockwise, the brightening pulse must start in the vertical-blanking time on or before the fifteenth line; with the LINE SELECTOR variable control set fully clockwise, the brightening pulse must start after the first 25 percent of the second field; the display on the Type 529 must remain stable throughout the rotation of the LINE SELECTOR variable control.

With the LINE SELECTOR switch set to 16, the brightening pulse must start at the start of line 16, and so on for each line through 21.

1. Connect a 75-ohm coaxial cable from the VIDEO OUTPUT connector of the Type 529 to the vertical-input connector of the external test oscilloscope.
2. Set the test-oscilloscope controls as follows:  
Input Selector: AC  
Volts/Div: 0.02



(A) Correct setting of control.

Fig. 7-18. Test-oscilloscope patterns for

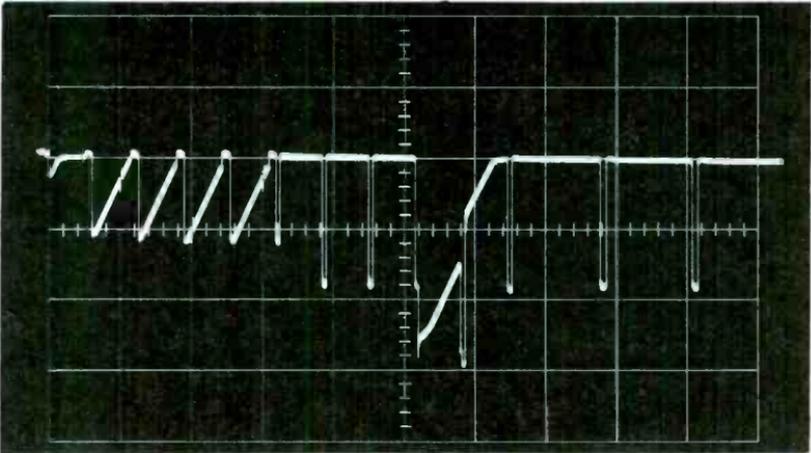
Time/Div: 2 ms

Triggering Mode: Adjusted for stable display

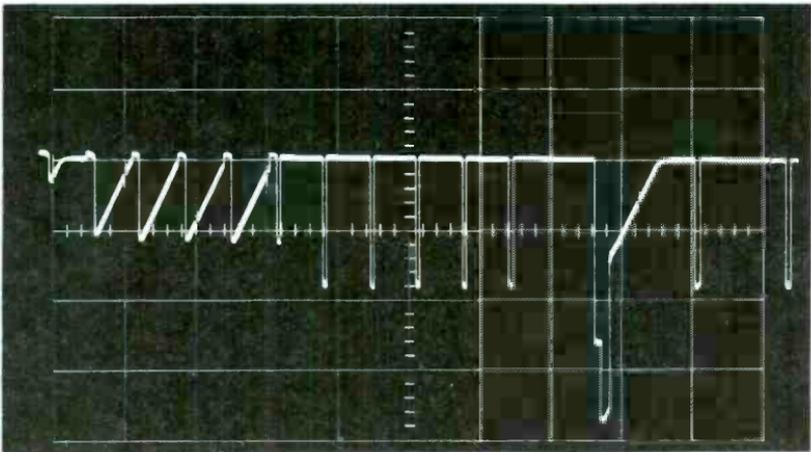
Trigger Slope: +Ext

NOTE: The external triggering signal must be at the video field rate.

3. Set the Type 529 DISPLAY switch to Line Selector .125H/Cm and the LINE SELECTOR switch to Variable.
4. Rotate the LINE SELECTOR variable control fully counterclockwise.

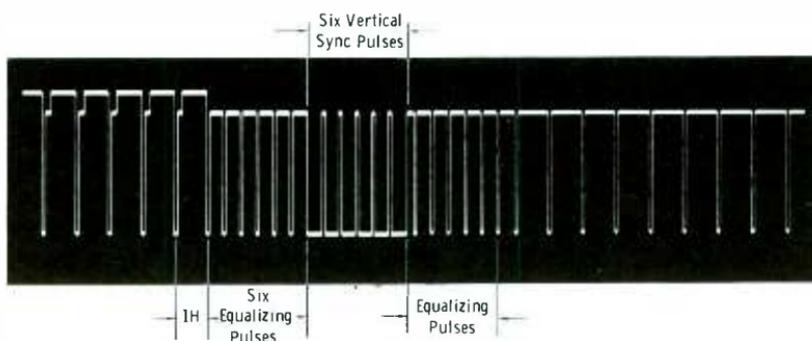


(B) Too far counterclockwise.

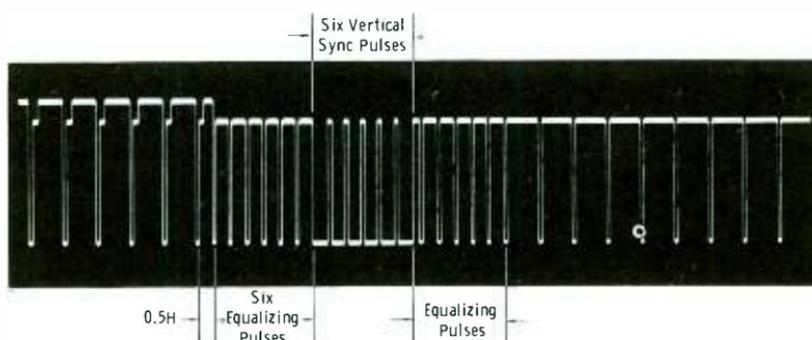


(C) Too far clockwise.

adjustment of field-1 sync control.



(A) Field switch set to one.



(B) Field switch set to two.

**Fig. 7-19. Checking for proper field display.**

5. Check the position (in time) and the amplitude of the brightening pulse on the test oscilloscope. The brightening pulse must be at least 100 mV high, and its start should be on or before the fifteenth line.
6. Rotate the LINE SELECTOR variable control fully clockwise.
7. Check the position (in time) and the amplitude of the brightening pulse on the test oscilloscope. The brightening pulse must be at least 100 mV high, and its start must occur after the first 25 percent of the second field.
8. Set the LINE SELECTOR switch to each numbered line position.
9. Check that the brightening pulse is at least 100 mV high and that it starts at the start of the line to which the LINE SELECTOR switch is set.
10. Set the FIELD switch to One and repeat steps 3 through 9.

NOTE: Line 10 starts at the first horizontal-sync pulse following the last equalizing pulse in both fields. Lines 16 through 21 can be found by counting forward from this point of reference.

### Adjust VIT Line-Selector Range

The VIT LINE SEL RANGE control (R458) is in the sweep generator.

1. Connect a dc voltmeter between the center arm of R458 and ground.

NOTE: Before proceeding, check the position of the LINE SELECTOR to ensure that the knob has not been misaligned.

2. Adjust the VIT LINE RANGE control so the brightening pulse just starts to jump between the 21st and 22nd lines as observed on the test oscilloscope.
3. Note the dc-voltmeter reading.
4. Adjust the VIT LINE SEL RANGE control so the brightening pulse just starts to jump between the 21st and 20th lines as observed on the test oscilloscope.
5. Note the dc-voltmeter reading.
6. Adjust the VIT LINE SEL RANGE control for a dc-voltmeter reading halfway between the two readings noted in steps 3 and 5.
7. Disconnect the dc voltmeter.

### Final Check of Line-Selector Operation

Use the same test-equipment setup as before.

1. Rotate the LINE SELECTOR variable control fully counterclockwise.
2. Check the position (in time) of the brightening pulse on the test oscilloscope. The brightening pulse must start on or before the fifteenth line.
3. Rotate the LINE SELECTOR variable control fully clockwise.
4. Check the position (in time) of the brightening pulse on the test oscilloscope. The brightening pulse must start after the first 25 percent of the second field.
5. Set the LINE SELECTOR switch to each numbered line position.
6. Observe whether the brightening pulse starts at the start of the line to which the LINE SELECTOR switch is set.
7. Set the FIELD switch to One and repeat steps 1 through 6.

## EXERCISES

- Q7-1. What should be checked prior to convergence adjustments for a color picture tube?
- Q7-2. What is the difference between "static" and "dynamic" convergence adjustments?

- Q7-3. On the Tektronix Type 529 waveform monitor, if you use the magnifier to observe the burst interval with the selector on the 2-line position, will the burst be interlaced or noninterlaced?
- Q7-4. How do you observe the noninterlaced burst on the Type 529 waveform monitor?
- Q7-5. If the input signal to a Tektronix Type 529 monitor contains a hum component, will this be displayed on the crt with the dc restorer on?

## Studio Measurements and Maintenance

### Part 1: Linear Distortions

It has become standard practice to consider video distortions as divided into three basic classes: linear distortion, nonlinear distortion, and interference (noise or cross talk). Linear distortion is any distortion independent of the signal amplitude, providing this amplitude is within the normal operating range of the equipment. Nonlinear distortion is a form of distortion which is amplitude-dependent, within the normal amplitude (and gain) range of the equipment. Interference cannot be classed in either of the above categories. For want of a better method, we will include the measurement of noise in Chapter 11 as a part of the proof of performance of the system.

**NOTE:** Linear-distortion measurements can be invalid if a significant amount of nonlinear distortion exists in the test path. If the test results vary with a change in amplitude of any test signal over the normal operating range, nonlinear distortion is a factor. In general, nonlinear distortion is more likely to occur in stl's, visual transmitters, and transmission terminal gear than it is in complete studio installations. However, it is obvious that such distortion can also occur at the studio, and tests for this condition must be made. Nonlinear distortions are covered in Part 2 of this chapter.

Linear video distortion can be divided further into two basic classes: (1) amplitude (gain) versus frequency, and (2) phase (delay) versus frequency. In either case, the frequency is varied at a given reference amplitude, and the corresponding amplitude or delay measurement is made.

To be both realistic and practical in meeting the needs of every telecaster, we will cover all the pertinent techniques associated with linear-distortion measurements: video-sweep, square-wave, multiburst-signal, and the preferred method employing the  $\sin^2$ -window generator.

**IMPORTANT NOTE:** Review Chapters 1 and 2 for basic back-to-back test-signal measurements (prerequisites for actual system tests), and basic test-signal characteristics.

## 8-1. THE SYSTEM MEASUREMENT PATH

Fig. 8-1 illustrates the basic principle of a studio-system measurement path. This allows measurement of the cumulative effects, involving all video DAs, interconnecting cabling with equalizers or equalizing amplifiers, switchers (including all the various switching paths involving faders, special-effects, etc.), and any stabilizing amplifiers or processing amplifiers in the path. Only the "air path" is measured in the example of Fig. 8-1; the "production path" (feeds to tape recorders for local productions, etc.) involves a separate measurement. The final jack field feeding the stl or telco terminal gear is terminated, and the oscilloscope is connected at that point. (Stl's involve a separate measurement except when overall performance is measured. See Chapter 9.) The test signal is fed to the input of each originating DA in turn.

## 8-2. SPECIAL TEST-EQUIPMENT ACCESSORIES

It is desirable to employ "keyed" test signals phased by the station sync generator to eliminate the test amplitude during horizontal and vertical-blanking intervals. This makes it possible to check the many types of amplifiers incorporating line-to-line clamps, which otherwise need to be modified if straight test signals are used.

When a multiburst or window generator is available, an external-signal input is often provided so that station sync and blanking may be inserted upon any desired test signal. When such provision is not made, the simple signal keyer of Fig. 8-2 can be constructed. The type 2N1143 transistor is excellent for this purpose. Application of sync and/or blanking pulses of negative polarity drives Q1 to cutoff for the input signal, and the amplitudes of setup and sync are adjusted by the respective controls shown.

The test-signal keyer functions as follows: The test signal is coupled to emitter follower Q1 through C1. A 75-ohm termination for the signal is provided by R1. Negative-going sync and blanking pulses appearing across common emitter resistor R4 serve to cut

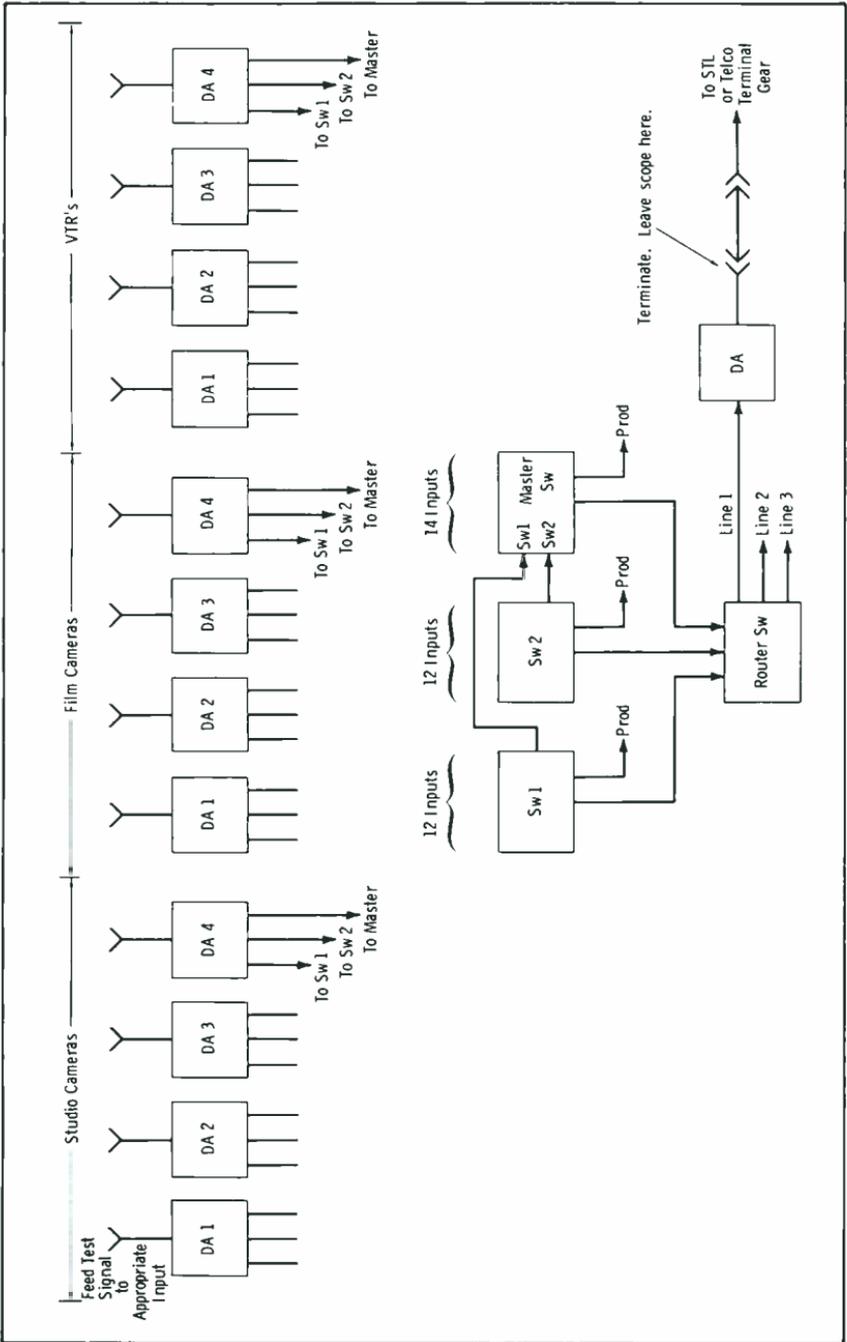
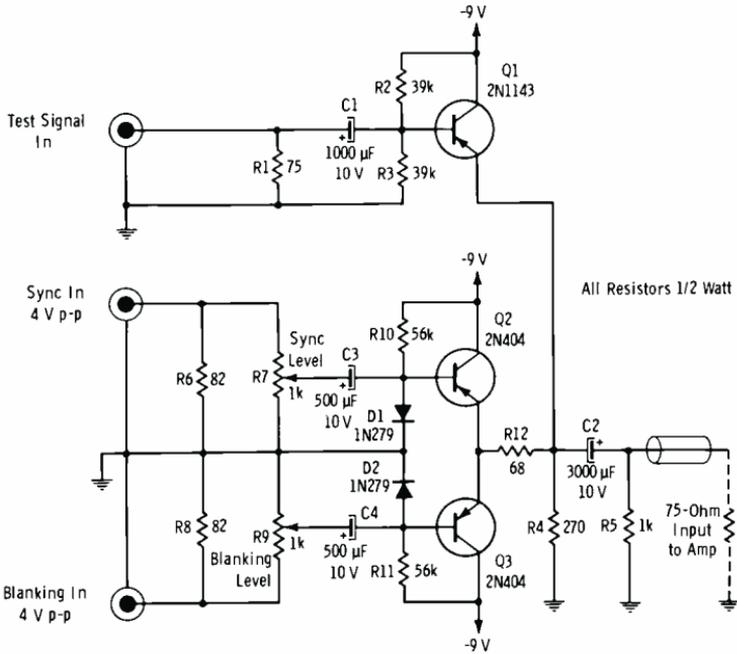


Fig. 8-1. Simplified block diagram of studio-system measurement path.



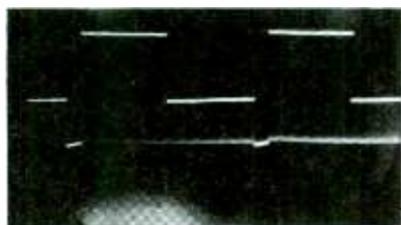
**Fig. 8-2. Keyer for test signals.**

Q1 off for their duration. The resultant composite signal is coupled through C2 to the input termination of the first system amplifier. The large capacitor values are necessary to pass 60-Hz waveforms without distortion.

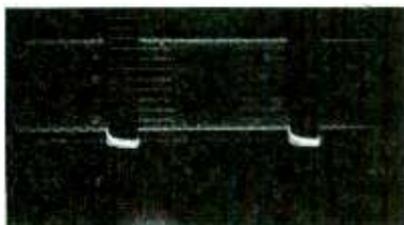
To use the keyer, adjust the amplitude of the input test signal from the generator to obtain the desired signal amplitude at the output. The maximum signal output level (to the system input) should be 1 volt peak to peak, including blanking and sync. The test-signal gain is somewhat less than unity.

Fig. 8-3A illustrates the keyed output when the test signal is a 60-Hz square wave. The setup (blanking) level is adjustable to the desired amount of pedestal. This type of signal results in a clean, composite blanking interval, and with the addition of sync no modification is necessary for units employing clamps.

The keyer may be used for sine waves, as shown in Fig. 8-3B with only blanking inserted. Fig. 8-3C illustrates the waveform after sync is inserted. This unit also enables the engineer to feed keyed video sweep to the system, with the same advantage of being able to leave all clamping circuits in an active condition, just as for any composite picture signal.



(A) 60-Hz square wave with station blanking.



(B) 1-MHz sine wave with station blanking.

(C) Same as B with station sync added.

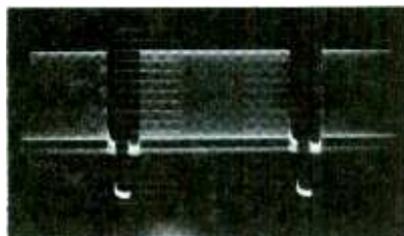


Fig. 8-3. Output of keyer in Fig. 8-2 for various test signals

Another useful accessory is the simple differentiating network of Fig. 8-4. The switch selects the proper time constant for horizontal or vertical drive, and a positive trigger from the trailing edge of the input pulse is delivered. Square-wave generators that accept external sync inputs are more stable with a positive trigger of short duration. This external trigger pulse allows synchronization of the square-wave generator to an integral harmonic of field or line frequency to obtain a stable pattern for measurements.

### 8-3. AMPLITUDE-VERSUS-FREQUENCY REQUIREMENTS

The amplitude-versus-frequency response of the television system must be such that the *overall* characteristic (through the transmit-

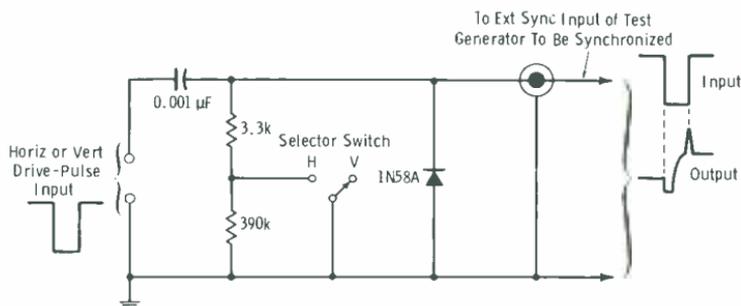


Fig. 8-4. Differentiating circuit for obtaining sync trigger.

ter) is reasonably flat over the required passband. It is realized, however, that the pickup device, whether it be an image orthicon, vidicon, or Plumbicon, utilizes a scanning beam that has a definite minimum spot size. Since this spot is not infinitely small, the waveform resulting from scanning across sharp vertical lines in the image will not be the ideal square wave, but more nearly a sine wave. This aperture distortion may be compared to passing the signal through a low-pass filter without phase distortion. Circuits used to compensate for this effect, in addition to pickup-tube output capacitance, do produce phase shifts which must be corrected by high-peaking or phase-correction circuitry that largely affects the gain at middle and low frequencies in the passband.

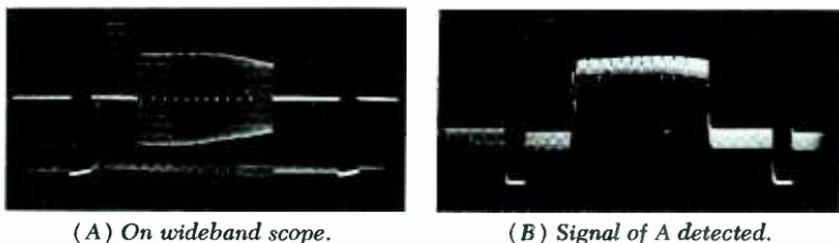
Since each pickup device varies over a limited range in characteristics, each camera chain incorporates the necessary correction circuitry for the pickup device. At this time, we are concerned primarily with distribution amplifiers, processing amplifiers, and all equipment which should exhibit flat frequency response with satisfactory amplitude and phase linearity. (NOTE: The stabilizing amplifier as operated at the transmitter location may be used to "pre-distort" the signal in an inverse relation to the nonlinearities of the transmitter. This is covered in Chapter 10.)

A word of caution is in order at this point. Technicians have been known to peak up video amplifiers to obtain a sharp, crisp reproduction on a master monitor. This seems to be particularly tempting with certain types of video amplifiers employing a single boost or peaking circuit which is adjustable from the front panel. Overpeaking is most likely to occur in stations where personnel are divided permanently between studio and transmitter, and the overall system function is not continually borne in mind. The practice of overpeaking to obtain a crisp picture on a studio monitor sometimes results in a deteriorated picture as viewed on a properly operating home receiver.

Video monitors themselves may be checked easily for resolution capabilities with a sine-wave generator. The most convenient method is to feed the generator output through a keyer, such as that shown in Fig. 8-2, so that pedestal and sync may be inserted for stable monitor operation. (For a monitor driven by external sync or drive, only the blanking pedestal is inserted.) As the frequency is increased, the vertical lines on the picture tube become thinner (and fainter) until nothing but the raster remains. By noting the maximum frequency at which the lines are just visible, the cutoff frequency of the monitor may be determined. The effect of brightness and contrast ratios on resolutions may also be observed, as well as the effect on the comparative resolution capability between various areas of the raster.

#### 8-4. KEYED VIDEO SWEEP

Complete systems may be checked with keyed video sweep without removal and modification of clamper circuits, as is required with straight video sweep. A schematic diagram of a simple keyer was shown previously in Fig. 8-2. Precautions in setup should be taken as illustrated in Fig. 8-5A, which shows the signal through the keyer as displayed by a Tektronix Type 524 scope in the wideband posi-



**Fig. 8-5. Keyed video sweep displays on Type 524 scope.**

tion. Adjust the video gain and blanking gain so that the sweep video is above the blanking level as shown. (Sync is not shown here, but it must be inserted prior to clamping stages.) This is necessary since the detector cannot discriminate between relative levels of video and blanking. Fig. 8-5B shows the signal of Fig. 8-5A after detection.

On a 30-MHz or 50-MHz scope (Tektronix Type 545 or 547), the wideband sweep will be perfectly flat provided the generator output is flat. A detector probe is not necessary when one of these more modern oscilloscopes is used. Always check the back-to-back characteristics of your sweep generator and scope to assure flatness.

**NOTE:** If, when the keyer is used, the blanking pulse should occur in the video-sweep trace on the scope, simply reverse the ac plug of the generator. In some cases, the sync-generator phase control may need adjusting with the generator on line lock. Since video-sweep techniques are normally carried out during off-the-air hours, color lock is not necessary.

The most efficient procedure in using the video sweep is to employ keyed sweep for overall system checks, and use normal video sweep (without blanking and sync) if it is necessary to service individual amplifiers not incorporating clamping circuits. For the keyed type of video sweep, insert only blanking when feeding inputs usually receiving noncomposite signals, with sync-insertion units incorporated. Insert both blanking and sync when feeding inputs normally receiving composite video. Use straight video sweep (un-

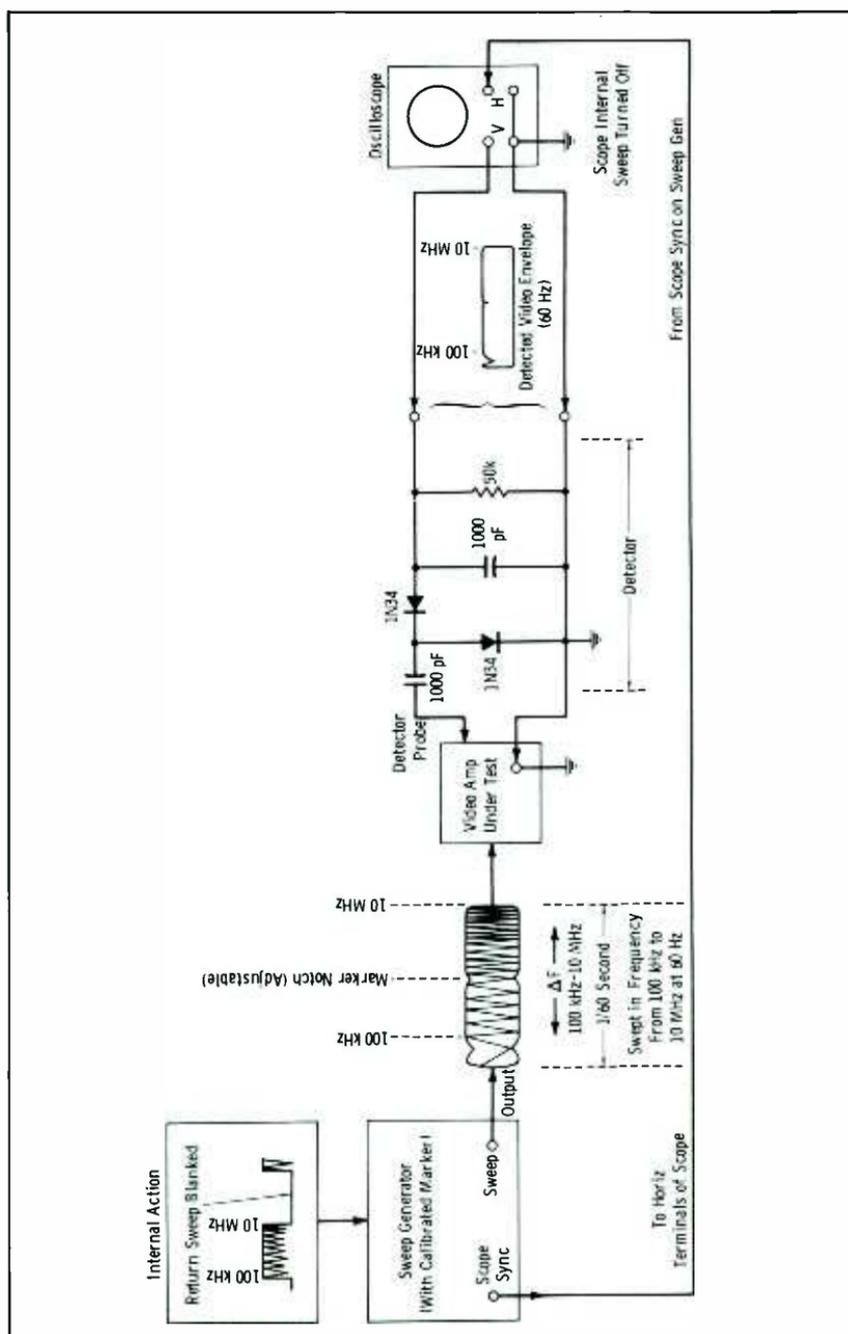


Fig. 8-6. Method of testing video amplifier for high-frequency characteristics.

keyed) for equipment not employing clamping circuits to avoid the unnecessary horizontal pulses on the vertical-rate sweep trace.

A sweep generator consists of a fixed-frequency oscillator the output of which is beat with the signal from a sweep oscillator frequency modulated at 60 Hz. The frequency modulation is such as to cause the beat frequency to swing over a usable range from about 100 kHz to 10 or 20 MHz. The frequency swing may be produced by a conventional reactance-tube circuit with 60-Hz excitation from the power line, or, in many cases, it may be produced by a motor-driven capacitor in the oscillator tank circuit. A 3600-rpm motor provides a 60-Hz sweep of the oscillator frequency. Such a sweep generator usually incorporates an absorption-marker generator that places a notch or series of notches at any reference frequency or frequencies over the usable range.

The fundamentals of checking the high-frequency characteristics of video amplifiers are illustrated in Fig. 8-6. The output frequency of the sweep generator is swept over a range of 100 kHz to 10 or 20 MHz, with a tunable frequency marker (notch) placed at any desired frequency. The sweep is repeated 60 times per second. This test signal is applied to the amplifier to be tested. A detector of the type shown in Fig. 8-6 is connected to the output of the amplifier when a 10-MHz scope such as the Type 524 is used. This detector rectifies the signal output as shown (in this case the amplifier is considered to be theoretically ideal: no distortion has occurred), and the output of the detector is fed to the vertical input of the oscilloscope, which traces a graph of output versus frequency over the passband above 100 kHz. For this test, the scope should have excellent low-frequency response so that no distortion of the 60-Hz square wave takes place. High-frequency response need extend no farther than 50 kHz, when a detector probe is used.

It is important not to overload the amplifier(s) when using video sweep. If the normal output is a 1-volt (peak to peak) signal, feed just enough input sweep level to result in a 0.5-volt (peak to peak) amplifier output. This minimizes any effect of nonlinear distortion. Remember to calibrate the detector probe so that you know how much loss occurs in the probe. For example, if the detector-probe gain is 50 percent, a 0.5-volt (peak to peak) actual output level reads 0.25 volt (peak to peak) through the probe.

Modern video-sweep generators for broadcast service have an internal output impedance (sending-end impedance) of 75 ohms. When the input of a system or individual amplifier is being fed, the proper 75-ohm termination should be used. In making response adjustments, which may require feeding an interstage circuit, the video-sweep generator feeds a high impedance unless the particular instruction manual for the unit specifies otherwise.

In checking individual units, the coaxial output cables should be disconnected and replaced with 75-ohm terminations. The detector probe is placed directly across the termination and retained in this position for response alignment.

Fig. 8-7 illustrates the general technique used in bench alignment procedures for a tube-type amplifier employing peaking circuits:

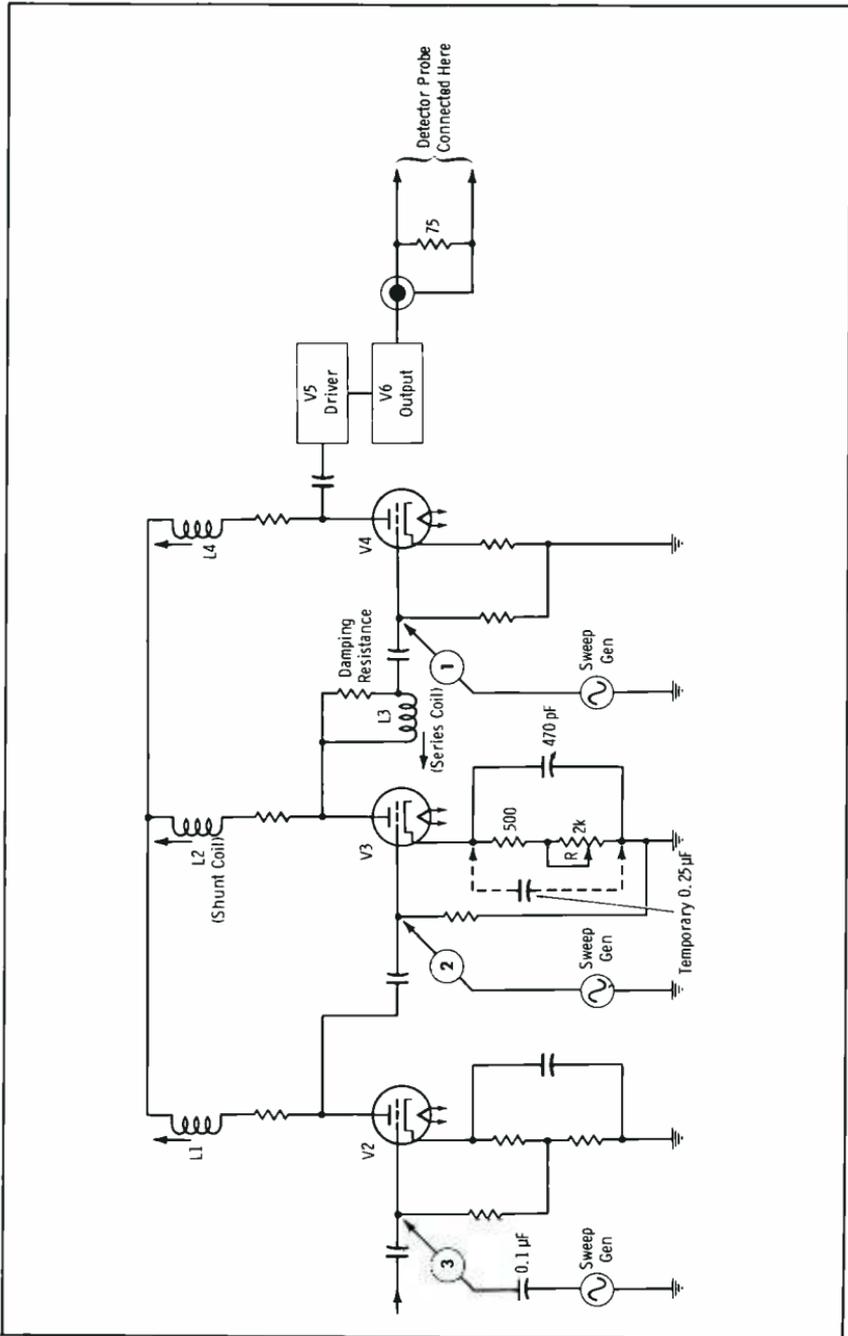
1. Connect the sweep generator in position 1. Adjust L4 for the flattest response.
2. Note that in stage V3 it is necessary to bypass temporarily the small capacitor with one of approximately  $0.25 \mu\text{F}$ . Either R or C may be variable in practice, but neither is adjusted in this step. Feed the sweep to point 2 and adjust L2 and L3 for flattest overall response.
3. Remove the temporary bypass. Note that the grid of V2 is at a dc potential because of the grid-return arrangement. Since the sweep generator does not normally employ a coupling capacitor at the output, it is necessary to feed point 3 through a  $0.1\text{-}\mu\text{F}$  capacitor to block dc from the generator output circuit. Adjust L1 of V2 and the R or C of V3 for the flattest response.

NOTE: The preceding is a general outline only to emphasize important precautions in technique. Always follow specific sweep procedures in equipment instruction manuals, when given and available.

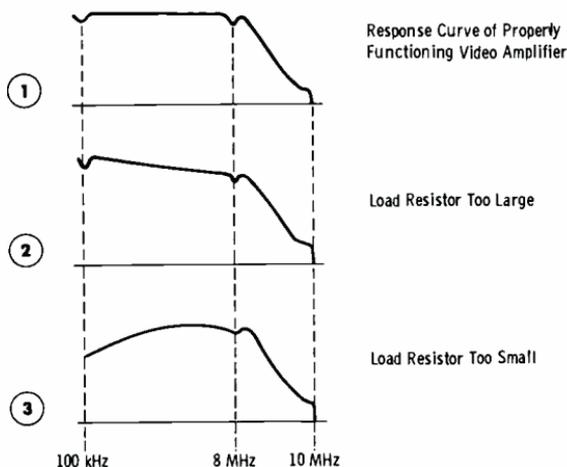
In all following examples of troubles, it is assumed that the back-to-back response of the sweep generator and scope (detected) is at least as good as curve 1 in Fig. 8-8.

In practice, the marker notch is set (by means of the calibrated marker dial) to occur at the point on the curve where the decided slope toward cutoff is noted. If the dial then reads 8 MHz, the response is flat from 100 kHz to 8 MHz. A minimum of 8 MHz is considered necessary in studio equipment to ensure minimum frequency and phase distortion. In many instances, modern commercial equipment tests well over 8 MHz.

If the plate load resistor should increase in value from the normal resistance, the trace obtained would appear similar to curve 2 in Fig. 8-8. We know that a higher value of coupling resistance causes a departure from flat response at both high and low frequencies. In this case, we are observing the high passband from 100 kHz to 8 MHz, and the droop toward the upper end of the band is noticeable. If the slope is very pronounced, phase distortion is bound to occur, and loss of resolution is apparent in the picture. Although this effect might be caused by an actual change in the value of the load resistor, this is not necessarily the sole cause. Anything that would affect the dynamic plate load so as to increase its effective impedance over the passband would have the same result. For example,



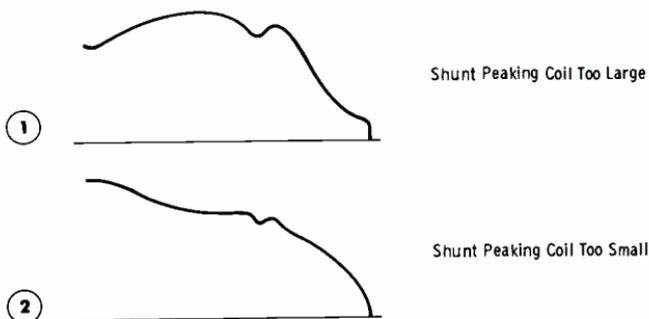
**Fig. 8-7. Alignment procedure for video amplifier.**



**Fig. 8-8. Effect of changes in value of load resistor on demodulated sweep-generator test signal.**

observe curve 2 in Fig. 8-9. This is essentially the same trace as curve 2 of Fig. 8-8, but it is caused by reduced inductance of the shunt peaking coil. Since this coil is part of the designed plate load, insufficient inductance causes an increase in effective plate load impedance. This may seem contradictory to the reader and will be clarified before we go further.

Peaking inductors in video circuitry compensate for circuit and tube capacitances which attenuate the higher frequencies. This compensation, in turn, allows a somewhat higher value of plate load resistance to be used than for an uncompensated circuit. Therefore, if a shunt peaking coil should fall to an insufficient value of inductance to produce the necessary compensation, the value of plate load resistance, in turn, becomes too high to maintain flat



**Fig. 8-9. Effect of shunt peaking coil on response curve.**

frequency response. This simply says that the dynamic load impedance has increased with a decrease in shunt inductive reactance.

Understanding of the basic circuit relationships materially aids the maintenance engineer in interpreting scope traces. Curve 3 of Fig. 8-8 is a typical trace when the load resistor has decreased from its normal value. Observation of curve 1 in Fig. 8-9 reveals that an increase in shunt peaking inductance has the effect of reducing the load resistance and therefore results in a similar trace.

The effect of the series peaking coil is shown in curves 1 and 2 of Fig. 8-10. When the series coil has more than the optimum value of inductance (curve 1), a gradual upward slope occurs from 100 kHz (low end of sweep) toward the middle of the sweep. The larger the value of inductance, the farther the hump is shifted to the left. Compare this with curve 3 in Fig. 8-8, which indicates the load

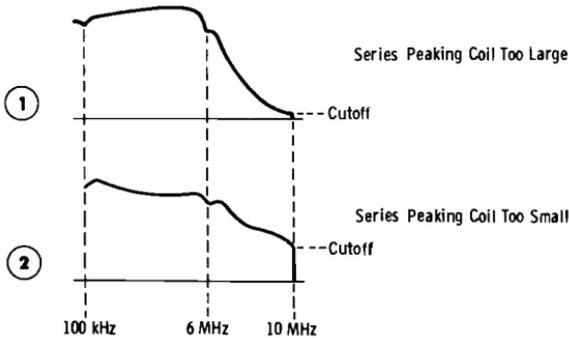
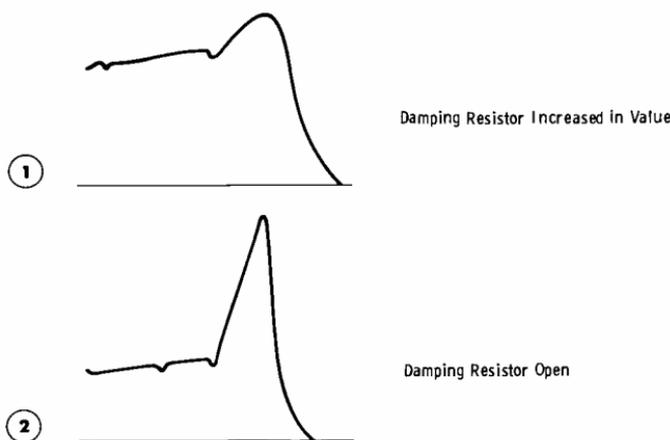


Fig. 8-10. Effect of series peaking coil on response curve.

resistor is too low in value. The major difference in the resulting traces is the extremely reduced cutoff level (at the start of the maximum downward slope of the curve) indicated in curve 1 of Fig. 8-10. This causes the notch to "slide down" on the sloping portion of the curve at the high end. From the slope of this curve, it may be inferred that the effective load resistance is reduced as the series peaking coil is increased in value, just as in the case of the shunt peaking coil. Increasing the value of the series coil lowers the resonant frequency (greater LC ratio) at which the series coil performs. This causes the hump in amplitude response to move to the left (lower in frequency), and the effective load impedance at the highest frequencies in the desired passband is reduced.

If the series peaking coil is too small in value (curve 2 in Fig. 8-10), the response from 100 kHz to the middle of the sweep is too large, and the amplitude at cutoff is increased. This condition also may be caused by a reduced value of damping resistor shunted across the series coil.



**Fig. 8-11. Effect of damping resistor on response curve.**

The effects of increased values of damping resistance are shown in Fig. 8-11. Curve 1 is displayed when the resistance value has increased to the point where inadequate damping of the resonance peak occurs. This is one possible cause of transient oscillation. Curve 2 indicates an open damping resistance that allows the resonant peak to appear.

The peaking circuits of most tube-type video amplifiers are adjustable, as indicated by the variable inductances in Fig. 8-7. The proper alignment of these stages constitutes an important function of the maintenance engineer both in initial setup of the amplifiers and in routine and priority checks of equipment. With the sweep generator connected at position 2 in Fig. 8-7, the effects of varying the adjustment of L2 and L3 may be noted. It will be observed that varying series coil L3 mostly affects the trace at the right of the pattern, and varying L2 mostly affects the trace through the center of the pattern. The marker-notch frequency should be set so that it appears at approximately the assumed limit of the flat portion of the curve, such as 7 or 8 MHz. If, on varying L3, the peak starts moving to the left, the adjustment should be made in the opposite direction to obtain as flat response as possible. Similarly, while the scope pattern is observed, L2 is adjusted to obtain the ideal response characteristic (curve 1 in Fig. 8-8). No more than approximately 2 percent variation should occur from 100 kHz to the limit of the flat portion of the curve. Always compare results with the manufacturer's specifications.

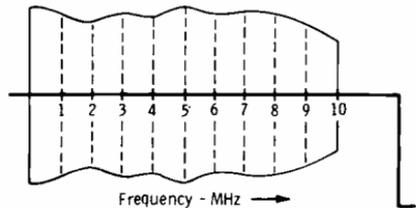
The typical traces shown in Figs. 8-8 through 8-11 assume only one component fault, as is usually the case in preventive maintenance or in trouble occurring during operation. If a number of am-

plitude variations appear in the pattern, several faults may exist simultaneously. In this case, the engineer familiar with the effect of any given adjustment on the corresponding pattern will establish a basis from which to proceed. It is important that the setup of test equipment and test leads produce no spurious response on the screen. Experience with any particular installation is necessary before the engineer can readily determine whether a trace is normal or abnormal.

If a great number of pronounced "humps" or "wiggles" occur on the scope screen, there probably is a poor ground connection. Always use the shortest possible ground leads in video-sweep measurements. If changing a ground connection to a different point changes the pattern, the grounding arrangement is faulty.

A typical overall video-sweep response of a modern studio installation is illustrated in Fig. 8-12. Inevitable "ripples" in response are caused by cabling, slight discontinuities of jack fields, nonexact terminations, etc. These hills and valleys should be no greater than  $\pm 0.5$  dB with no sharp changes across the passband. The response at 10 MHz should be down no more than 1 dB, or 10 percent on the voltage scale.

**Fig. 8-12. Typical overall video-sweep response of studio installation (wide-bandwidth oscilloscope).**



When overall video-sweep runs are made on studio equipment at periodic intervals, a record of the results obtained from each run should be kept as a "flag" to indicate deteriorating performance. Table 8-1 gives an example of how such a record may be put in tabular form.

## 8-5. THE MULTIBURST TEST SIGNAL

The multiburst test signal (developed in Chapter 2) provides a convenient "quickie" check of system performance up to 4.2 MHz (typically). Some multiburst generators designed for studio measurements provide bursts up to 8 or 10 MHz, but the standard signal as transmitted for VIT purposes and as normally used provides a white flag, 0.5 MHz, 1.5 MHz, 2 MHz, 3 MHz, 3.58 MHz, and 4.2 MHz (Fig. 2-19). Since the signal is constructed on standard station

**Table 8-1. Example of Tabular Form for Video-Sweep Performance**

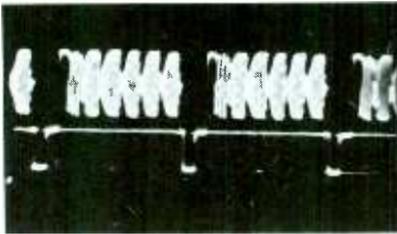
Reference Point	Frequency (MHz)	Response
First minimum	Approx 2 MHz	0.1 dB down
First maximum	Approx 4.5 MHz	0.2 dB up
Second minimum	Approx 6.5 MHz	0.3 dB down
Second maximum	Approx 9 MHz	0.3 dB up
10 MHz		0.6 dB down
Date: _____		
Engineer: _____		
Path (Include DA numbers, switcher(s), and switcher(s) path): _____		

sync and blanking, it is readily transmitted and measured and is easily interpreted by the average station operator.

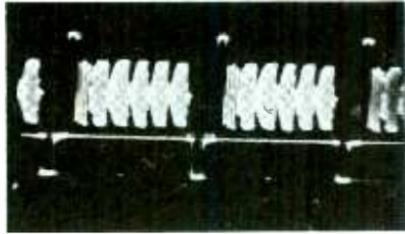
As indicated in Fig. 8-13, there are two basic proportionments of white-flag amplitude to burst amplitude. Fig. 8-13A shows the normal proportionment used for studios, AT&T Long Lines Division, and modern tape recorders and visual transmitters. Most older visual transmitters and video tape recorders exhibited considerable overload at the higher video frequencies when full-amplitude signals were used, causing the tests to be meaningless. (The high-frequency content of the average picture never approached reference white level for any considerable duration.) For this reason, the proportionment of Fig. 8-13B was used. With the white flag at 100 IEEE units, the burst peaks are at 70 IEEE units. With a 10-percent setup level, this results in a peak-to-peak burst amplitude of 60 IEEE units.

The keyed burst signal is the most convenient check for line or system frequency response from 0.5 MHz to the upper limit of the passband. The individual sine-wave bursts should be read peak-to-peak in voltage or IEEE units. The setup, of course, will change with attenuation of the burst frequency and should be disregarded in readings.

Fig. 8-14A illustrates a gradual rolloff with increasing frequency as would occur on a long unequalized line. For example, at 4 MHz the attenuation of RG-11/U cable is 0.4 dB per 100 feet. Fig. 8-14B shows the rising response that usually is the result of overpeaking. Fig. 8-15A is the hourglass display, which can be caused by faulty equalization for a rolled-off response. In this case, the higher-frequency end is overequalized, and the resulting picture actually is much inferior to that obtained from the gradual rolloff of Fig. 8-14A, since middle-frequency "holes" affect picture resolution drastically. This effect also can be produced by an open shield



(A) Proportionment of flag and burst for normal AT&T line checks and studio runs.

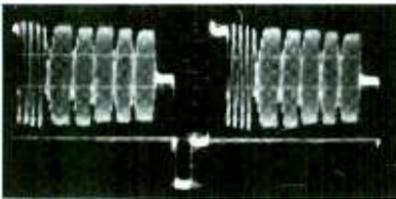


(B) Proportionment for response runs on older transmitters and video tape recorders.

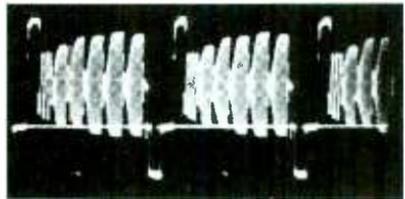
**Fig. 8-13. Keyed burst signals for overall transmission checks.**

ground on one end of the coaxial cable transmitting the signal, or by faulty terminations. Fig. 8-15B illustrates a shifted axis along the individual bursts as a result of frequency-selective harmonic distortion, which can be caused by overloads at certain frequencies or by overpeaking.

The axis-shift effect or an actual loss of high-frequency response is sometimes the result of intentional overpeaking in an attempt to obtain a sharp picture. But if an off-air monitor were placed side-by-side with a studio monitor displaying the overpeaked or overequalized signal, the modulation effects of the main transmitter

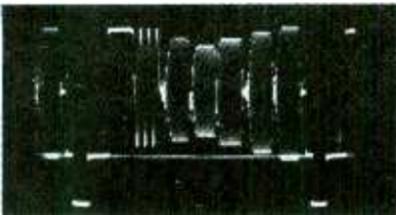


(A) Gradual rolloff with increasing frequency.

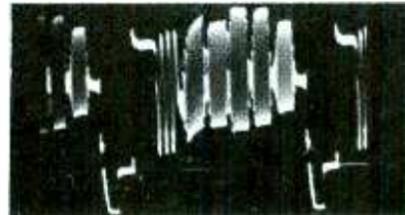


(B) Increase of gain with increasing frequency.

**Fig. 8-14. Keyed burst-signal displays.**



(A) Display resulting from poor mid-frequency response (hourglass effect).



(B) Harmonic distortion (shifted axis) caused by overload or overpeaking.

**Fig. 8-15. Keyed burst-signal displays indicating other types of distortion.**

and any studio-to-transmitter links (usually involving either microwave or equalized lines) would be most revealing. Most systems employing fm for video relay include low-frequency attenuation circuits to prevent excessive swings of the carrier frequency at the low video frequencies. This effectively limits the frequency excursion in the if strip of the microwave receiver so that differential phase at 3.58 MHz (color subcarrier) and any high-frequency sound subcarrier is within tolerable limits. As a specific example, the RCA TVM-1 stl transmitter uses an 8-dB attenuation at 60 Hz with gradually decreasing attenuation to 6 MHz. The video is restored in the receiver restoration network. With any such networks, an overpeaked signal with the higher-frequency components extending into the sync region will cause compression or actual clipping of the highs. Restoration of the lower frequencies does not remove the high-frequency compression that results in a harmonic distortion in direct ratio to the amount of overpeaking.

The amount of compression, of course, is also dependent on the peak-to-peak video level used at the modulator to obtain the 100-percent reference modulation. When this is held within the design limits of a particular system, the degree of compression from an overpeaked signal can be quite small. In this case, the major cause of severe edge effects is the ringing caused by the main-transmitter low-pass filter that employs a rapid cutoff above 4.18 MHz. It is also known that when high-frequency energy is appreciable (as is the case with sine waves or keyed video sweeps), some vacuum-tube circuits can exhibit considerable overloading at these frequencies while passing lower frequencies at normal gains.

## 8-6. THE SQUARE WAVE

The square-wave test signal should be used only when a standard window-signal generator is not available.

The square wave is a versatile test signal when the user follows the precautions with respect to the oscilloscope outlined in Chapter 1 of this book. When the system includes a unit incorporating clampers (such as a processing amplifier or the transmitter), the square wave should be keyed and sync inserted as discussed previously. Fig. 8-16 shows how the square wave can be utilized in a basic analysis of a television system or a single unit of such a system.

If the square-wave response indicates a loss of low-frequency gain with accompanying phase distortion, the raster will be gradually shaded from top to bottom. The video setup level will be reduced, resulting in excessive contrast and black compression on a monitor or receiver that was adjusted for a standard (distortion-

Test Signal	CRO Indication at System Output	Effect on Picture	Defect	Cause
		Normal picture.	No defects. Excellent low-frequency response. Negligible phase shift.	
		Shading top to bottom of picture. Loss of setup. (See text.)	Loss of low-frequency gain with leading low-frequency phase shift.	Coupling capacitors decreased in value; screen and cathode bypass capacitors; low-frequency compensation circuits out of adjustment; grid resistors decreased in value; defective screen resistors; clamping-circuit failure.
60 Hz		Shading top to bottom of picture. Increase of setup. (See text.)	Excessive low-frequency gain with lagging low-frequency phase shift.	Overcompensation of low-frequency correction circuits; coupling capacitors; screen and cathode circuits; grid resistors (open or high).
<p>p-p Amplitude With Tilt (x+y)</p>  <p>With Tilt</p> <p>Actual Pulse Amplitude</p>		$\% \text{ tilt} = \frac{x}{y} \times 100$ <p>Example: <math>x = 1 \text{ cm}</math> <math>y = 4 \text{ cm}</math> then: <math>\frac{1}{4} = 0.25 \times 100 = 25\% \text{ tilt}</math></p>		
		Normal Picture.	No defects. Excellent response at multiples of line-scanning frequency.	
		Very poor resolution.	Poor middle- and high-frequency response.	Improper adjustment of peaking coils; plate load resistors increased in value; peaking coils; or decreased value of peaking-coil shunt resistors.
15,750 Hz (Line Scanning Frequency)		Black-following-white streaking. (Negative streaking horizontally.)	Loss of low-frequency gain with leading low-frequency phase shift.	Coupling circuit time constants; improper peaking adjustments; defective peaking coils; bypass capacitors or bypass time constants; irregular gain at middle frequencies.
		White-following-white streaking. (Positive streaking horizontally.)	Middle- and low-frequency response too high with lagging low-frequency phase shift.	Misadjustment of phase-correction circuitry or low-frequency compensation circuit.
		Normal.	No defects. Good high-frequency and transient response.	
75 kHz		Fair to poor resolution.	Poor high-frequency response; poor rise time.	Improper adjustment of peaking coils; defective peaking coils or low value of peaking-coil shunt resistors; plate load resistors increased in value.
		Bad "edge effects"; "ringing" after fine vertical lines.	Excessive high-frequency response; nonlinear time delay; high-frequency cutoff too rapid.	Overpeaking from improper adjustment; plate loads decreased in value; peaking coils or open shunts on coils; poor lead dress.

Fig. 8-16. Use of the square wave.

free) picture. Overall tilt (studio to transmitter output) should be held under 2 percent for complete freedom from visually observable effects. In some cases, the processing amplifier will give a flat output with an input tilt of up to 10 percent.

When this type of overall distortion is noted, individual units should be checked separately. A 60-Hz square wave is useful for setting coupling-circuit time constants (where used), usually designated as *tilt controls*. For example, in the RCA TVM-1 series of microwave systems, such controls are employed in the modulator, monitor amplifier, and receiver amplifier. The square wave is fed to the transmitter and is observed at the klystron repeller with the modulator tilt control adjusted for flat response or the same tilt as indicated on the scope at the transmitter input. The monitor and receiver may then be adjusted for proper transmission of the square wave. Similarly, "lap-dissolve" amplifiers (mixing amplifiers) used with switching systems sometimes incorporate this type of control so that the tilt can be removed before distribution to the transmitter terminal gear.

It should be realized here that point-to-point, single-frequency sine-wave runs might indicate a response down to 60 Hz well within 1 or 2 dB of the response at the reference frequency, and yet the equipment can fail to pass a 60-Hz square wave with less than 20 to 30 percent or more tilt. Loss of effective coupling-circuit time constants, or clamping failure, will cause this type of distortion.

## 8-7. THE $\text{SIN}^2$ -WINDOW SIGNAL

Review Chapter 2 for the development of the  $\text{sin}^2$ -window test signal. It is sometimes termed the *pulse-and-bar* signal.

In an attempt to correlate test-signal measurement with an actual degree of picture impairment, the *K factor* is used. The K factor is basically defined in terms of a standard picture distortion which is a single echo spaced in time  $8T$  or more from the main transition. For example, if the peak amplitude of this single echo is 4 percent of the original transition amplitude, the K factor is defined as 4 percent.

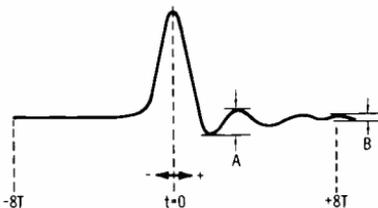


Fig. 8-17. Basic quantities involved in explaining K factor.

See Fig. 8-17. A signal transition with a sine-squared shape occurs at  $t=0$ . At a point spaced at  $+8T$ , a certain amplitude of "ring," or echo, exists. Let us arbitrarily assume that this amplitude is 4 percent of the original amplitude, so  $B = 4$  percent. A waveform distortion (A in Fig. 8-17) much closer to the transition is larger, but its effect as judged by an "average observer" is only equal to the picture impairment caused by echo B. Thus, although echo A may be 16 percent of the original amplitude, echo B, of only 4-percent amplitude, results in the same degree of picture impairment. We may construct a gratitude mask which defines limits within which a waveform must fit if it is to have a K factor equal to or less than the limits specified.

For the purpose of assigning a numerical value to a subjective assessment, we can say that for a K factor of 5 percent, picture impairment is noticeable to an experienced and critical observer, whereas a K factor of 3 percent is not noticed by the same individual.

See Fig. 8-18. Along the positive time base of the transition, for h.a.d. =  $0.250 \mu\text{s}$ , the time from  $t=0$  to  $t=T$  is  $0.125 \mu\text{s}$  (A in Fig. 8-18). Then the time to  $8T$  is  $8 \times 0.125 = 1 \mu\text{s}$ . When h.a.d. =  $0.125 \mu\text{s}$ ,  $T = 0.0625 \mu\text{s}$  (B in Fig. 8-18). Therefore  $8T = 8 \times 0.0625 = 0.5 \mu\text{s}$ . Obviously, the transition along the negative time base is the same, but progresses from  $t=0$  in the opposite direction.

Now see Fig. 8-19. At plus and minus  $8T$ , lines representing the limits of the K factor are spaced in reference to the amplitude at  $t=0$ . If the K-factor limit is to be 4 percent, then at the  $8T$  points

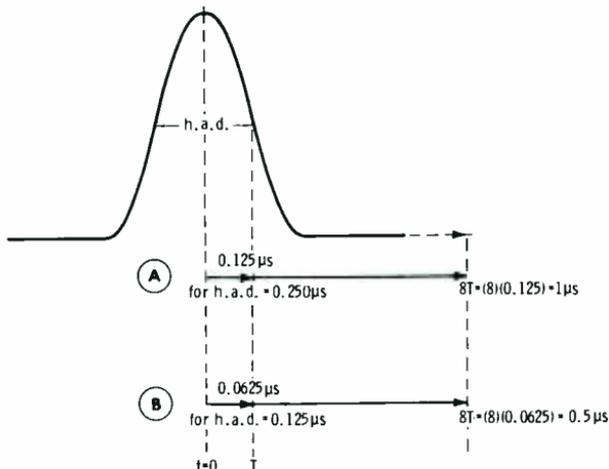


Fig. 8-18. Pulse transition along positive time base.

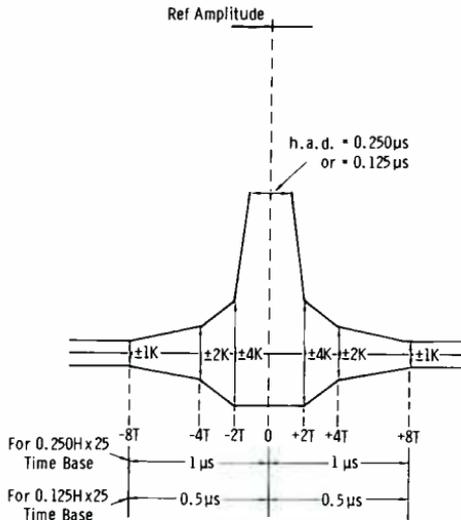


Fig. 8-19. Basic K-factor graticule.

the lines are spaced plus and minus 4 percent of the amplitude at  $t = 0$ . Echoes of larger amplitude may occur closer to the main transition with no increase in subjective picture impairment. Note, for example, that at  $2T$  the limit increases to 4 times that at  $8T$ . Thus, if the  $8T$  point has a K factor of 4 percent, the  $2T$  point is allowed an amplitude of  $4 \times 4 = 16$  percent to fit within the 4-percent K-factor mask.

Note also that the same mask can be used for pulses of either  $0.250 \mu\text{s}$  or  $0.125 \mu\text{s}$  h.a.d. by proper adjustment of the waveform-monitor time base (providing that proper precautions are used in interpretation for various systems, as described later). Table 8-2 lists time bases of the Tektronix Type 529 waveform monitor. The time base of  $0.250\text{H}/\text{cm} \times 25$  is the proper time base for the pulse with h.a.d. of  $0.250 \mu\text{s}$ . For the pulse with h.a.d. of  $0.125 \mu\text{s}$ ,  $0.125 \text{H}/\text{cm} \times 25$  is the proper time base.

Table 8-2. Waveform-Monitor Time Bases

Horizontal Magnification	Display Switch on 0.125H/cm		Display Switch on 0.250H/cm	
	H/cm	μs/cm	H/cm	μs/cm
×1	0.125	7.94	0.250	15.9
×5	0.025	1.59	0.05	3.18
×25	0.005	0.318	0.01	0.635



Fig. 8-20A is a photograph of the K-factor graticule on the Tektronix Type 529 waveform monitor. The solid lines outline a 4-percent K factor, and the dash lines outline a 2-percent K factor. The sweep rate for the pulse-and-bar signal (hardly distinguishable in Fig. 8-20A because the pulse is very faint and the bar is centered on the proper graticule lines) is  $0.125H/cm$ , thus displaying one horizontal line. Note that a 10-percent setup level is used.

Fig. 8-20B is a multiple exposure of the three basic sine-squared pulses from a standard generator. The time base used was  $0.125H/cm \times 25$ . Note that the pulse with  $0.125 \mu s$  h.a.d. (center) is well under the 2-percent line for the h.a.d., since we are looking directly at the generator output.

Also note in Fig. 8-20B that the T/2 pulse (right) is slightly below the reference amplitude of 100 IEEE units. This emphasizes that a sine-squared pulse generator *must* be adjusted with an external test oscilloscope that has satisfactory response up to 30 to 50 MHz. When the T/2 pulse is properly adjusted in amplitude with such an oscilloscope, The Type 529 specifications call for an amplitude of 94 to 100 IEEE units for the T/2 pulse display. The wideband response of the Type 529 waveform monitor is flat within 0.3 dB to 8 MHz.

The K-factor graticule includes the limits of flatness for the window signal, and the limits for the pulse-to-bar amplitude measurements for the K factor used. See Fig. 8-21, which indicates how

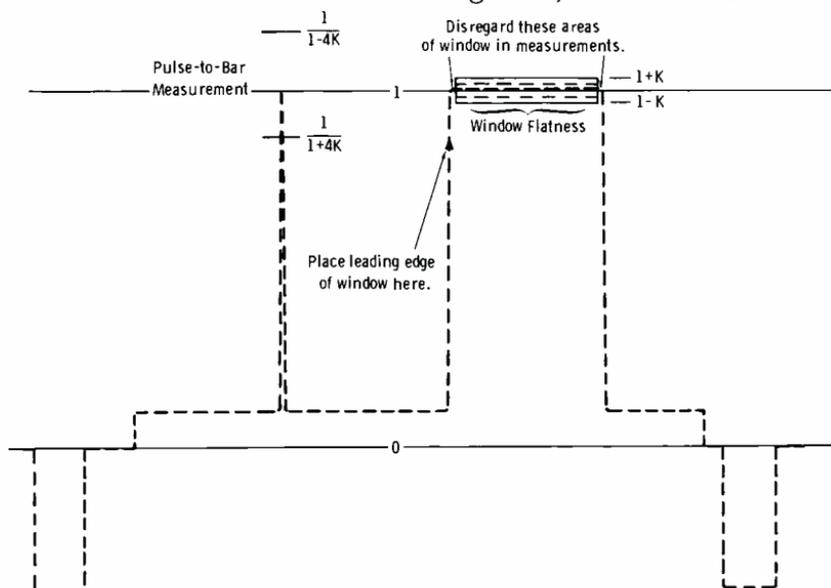
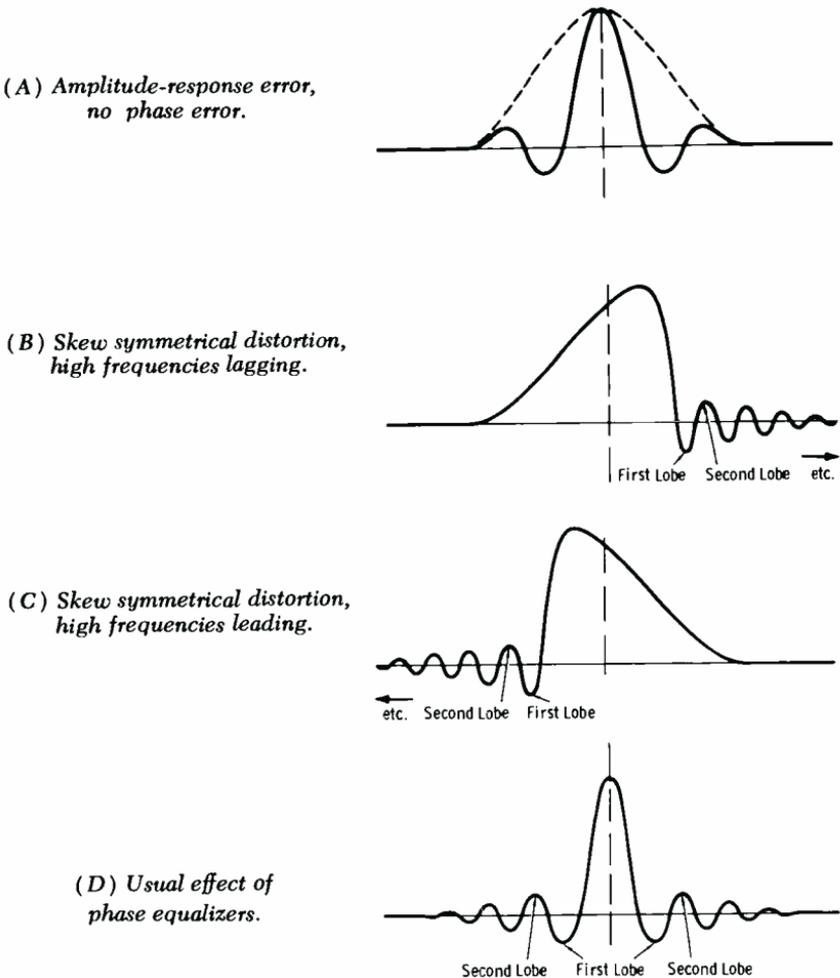


Fig. 8-21. Pulse and window K-factor limits, normal horizontal sweep.

these limit lines are established. Observe also in this drawing that an indicator is used to show the correct waveform centering to place the leading edge of the window signal. Note that an area on either side of the window is disregarded in this measurement, and that only the enclosed area along the top of the bar is used. This is true for either horizontal- or vertical-rate waveform presentations.

Fig. 8-22 illustrates the basic T-pulse responses encountered. In Fig. 8-22A, an amplitude-response error is apparent without phase-response error. Waveform errors this close to the transition do not impair the signal (unless excessive) as much as errors farther



**Fig. 8-22. Basic T-pulse responses.**

away. In fact, we should recognize this type of "error" as that obtained from "phaseless aperture correction" in camera chains. Thus we have a "crispening" effect of a single overshoot as compared with actual picture impairments such as would result from the remaining waveforms of Fig. 8-22.

Fig. 8-22B shows the "skew symmetrical" distortion caused when the delay increases with increasing frequency. Fig. 8-22C shows the opposite type of phase distortion, where the delay decreases with increasing frequency. In a system with a fairly rapid rolloff that employs phase equalizers to correct the resulting phase distortion, proper equalizer adjustment is indicated when ringing amplitudes are equally distributed preceding and following the pulse as shown in Fig. 8-22D.

The amplitude-frequency and amplitude-phase response at frequencies higher than about 100 kHz is most evident in the measurement of the  $\sin^2$  pulse. Amplitude-phase response at frequencies below 100 kHz is most evident in measurement of the window signal.

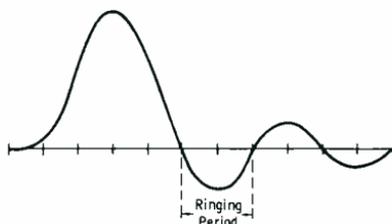
NOTE: Some generators place the  $\sin^2$  pulse following the window rather than preceding the window. This has no effect on basic understanding of measurement principles.

Distortions at low frequencies produce waveform distortion with a long time constant, as, for example, streaking. This is most evident in window measurement. Distortions at higher frequencies produce waveform distortions with shorter time constants as, for example, smearing, loss of resolution, or "edge effects" from bad transient response. This is most evident in  $\sin^2$ -pulse measurement or in window-signal transitions.

High-frequency rolloff results in loss of amplitude. Loss of amplitude results in a *widening* of the pulse, since the area of the pulse represents a constant dc component. A slow rolloff within the video band produces a large reduction in amplitude (and pulse-width increase) with little or no ringing. A rapid rolloff close to the top of the band but still within the desired video bandwidth produces both a reduction (perhaps slight) in amplitude, and ringing. A rapid rolloff (almost a cutoff) just above the video bandwidth concerned results in practically no effect on amplitude, but does produce ringing. The shape of the rolloff and whether the resulting phase shift is leading or lagging is revealed by the distribution of ringing before and after the pulse.

The window permits detecting low-frequency distortion, which has practically no effect on the  $\sin^2$  pulse. The window shows undershoot, overshoot, and horizontal tilt, depending on the time constant of the impairment. When used with the  $\sin^2$  pulse, the

Fig. 8-23. Ringing period.



window has the same rise time as the pulse so that no frequencies higher than the system test reference are introduced.

Ringing occurs at the frequency at which the gain dip occurs in the system being measured. The ringing amplitude depends on the sharpness of this gain-dip characteristic.

The ringing period (Fig. 8-23) is defined by the following relationship:

$$R_p = \frac{1}{f_c}$$

where,

$f_c$  is the cutoff frequency.

For example, if we have a 4MHz cutoff, the ringing period ( $R_p$ ) is:

$$R_p = \frac{1}{4(10^6)} = 0.250 \mu s$$

To find the cutoff frequency for a given measured ringing period:

$$f_c = \frac{1}{R_p}$$

where,

$f_c$  is the cutoff frequency in megahertz,

$R_p$  is the ringing period in microseconds.

In defining waveform distortions, certain terminology has become standard, and it is well to review this terminology here:

*Short-time waveform distortion (SD)* involves impairment of small picture detail in the horizontal direction. It is seen as blurring or smearing of a sharp brightness transition. It may or may not be accompanied by an overshoot or ringing to the right (or left) of the transition. Measurement of SD may be accomplished by observing the leading and/or trailing edge of the window signal displayed at the horizontal rate; the display may be expanded on the scope time base.

*Line-time waveform distortion (LD)* concerns a longer time constant than does SD, and it results in impairment of brightness

reproduction between the sides of a picture detail. When the detail is smaller than full picture height, the streaking is most noticeable to the right of the detail. Details extending all the way up and down the picture may result in streaking across the full raster horizontally. Measurement of LD is done across the top of the window signal viewed at the horizontal rate, and by the relationship of the leading and trailing edges to reference black.

*Field-time waveform distortion (FD)* results in impairment of brightness reproduction from top to bottom of the picture. Measurement of FD is done across the top of the window signal viewed at the vertical rate, and by the relationship of the leading and trailing edges to reference black.

*Relative chroma level (RCL)* is a measure of the faithfulness of reproduction of the saturation of all colors in a color picture. High RCL causes more vivid colors than intended; low RCL causes colors more pale than intended. Measurement of RCL is done most readily with the modulated 20-T pulse.

*Relative chroma time (RCT)* is a measure of relative chroma and luminance delay. The result of RCT errors is misregistration of all colors with their respective luminance components. Delayed RCT places chroma to the right of its luminance component; advanced RCT places chroma to the left. Measurement of RCT is done with the modulated 20T pulse.

Note that we have defined above only those waveform distortions associated with linear types of distortion. Nonlinear distortions are covered in Part 2 of this chapter.

Fig. 8-24 illustrates SD, which may become LD with a slightly lower rolloff point and more severe phase shift. The waveform of Fig. 8-24A indicates, in this example, a rolloff of high frequencies as indicated by the low amplitude of the pulse relative to the bar, and by the rounded window corners. Fig. 8-24B shows the resulting monitor display, which, in this example, is more appropriately termed "smearing" rather than "streaking." (Streaking is shown by Figs. 8-25 and 8-26.)

Note that in Fig. 8-24B, the pulse (point 1 on the waveform) is hardly distinguishable because of the low amplitude. The rounded transition from gray to white (point 2) results in a leading-edge smear, and the rounded fall-off toward black (point 3) results in white-after-white smearing. A truly "short-time smearing" would be less severe than this, affecting only the finer transitions (higher frequencies) of the picture. If the duration of the transition is up to about  $3 \mu\text{s}$ , it is termed "line-time smearing," as contrasted with SD, or "short-time smearing."

One type of LD is shown in Fig. 8-25. In this example, positive streaking is indicated preceding the window (black-after-black)

and following the window (white-after-white). Fig. 8-25A illustrates the window signal at the horizontal rate, and the resulting monitor presentation is shown in Fig. 8-25B. The reason for the appearance of Fig. 8-25B is indicated more clearly by Fig. 8-25C. Note the blacker-than-black tilt prior to the window, and the time

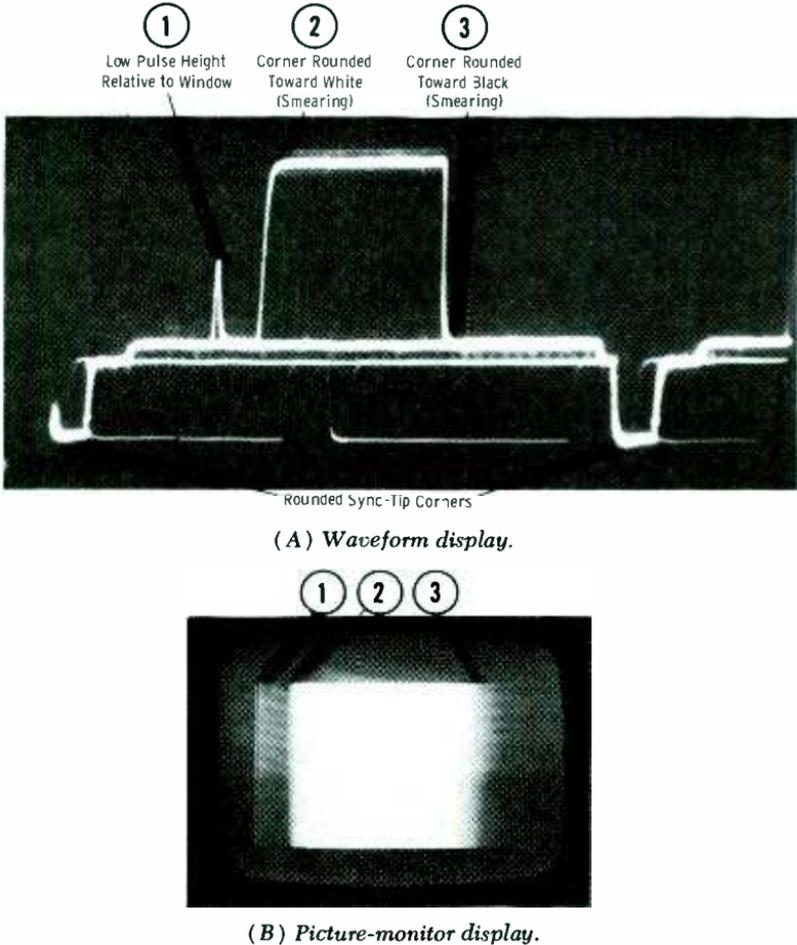
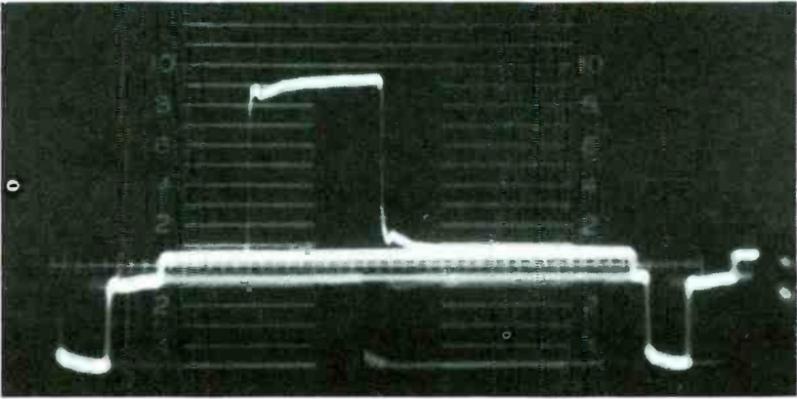


Fig. 8-24. Example of short-time waveform distortion.

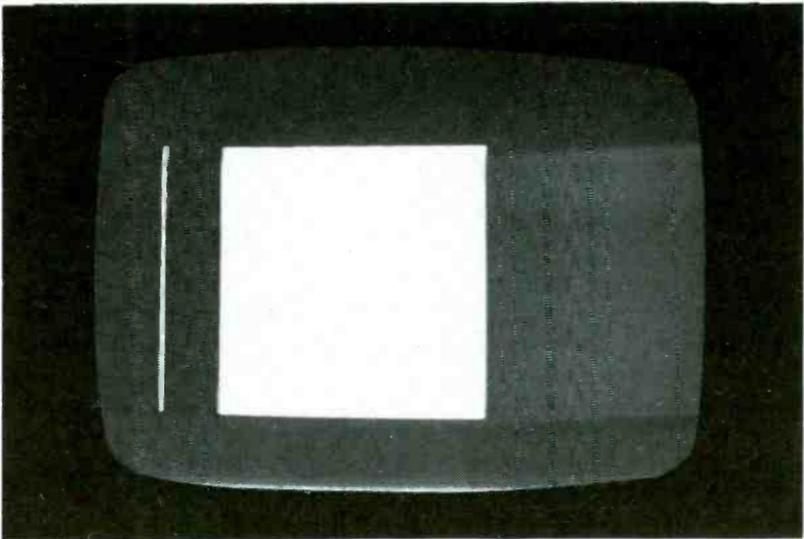
duration required to fall to black at the trailing edge of the window. In actual measurement, the  $1\text{-}\mu\text{s}$  intervals at the leading and trailing edges are not used, and the same durations for a and b relative to A are used. The window is approximately  $\frac{1}{2}H$  in duration, and the time from the trailing edge of the window to the lead-

ing edge of sync is about  $\frac{1}{4}H$ . So one-half of the window tilt is included in the  $a$  measurement, as indicated on the drawing.

If the type of distortion is strictly linear, dimensions  $a$  and  $b$  are equal. If nonlinear distortion is present, these dimensions may differ. If reducing the level of the test signal into the system changes the relative dimensions of  $a$  and  $b$ , nonlinear distortion is present, and a lower level of test-signal input should be used to check the actual linear distortion.



(A) Window waveform at horizontal rate.



(B) Picture-monitor display,  $\sin^2$  pulse added.

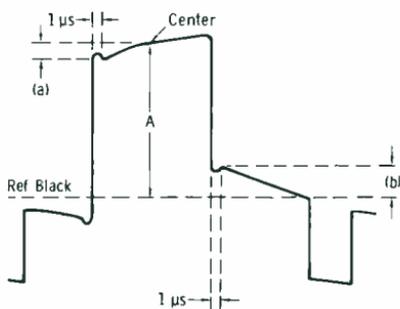
Fig. 8-25. Example of LD with leading

Fig. 8-25D shows the vertical-rate cro display of the same signal. The white-going setup between the bottom of the white signal and blanking serves as an accurate indicator of the percentage of distortion. This defect is the result of excessive gain at low frequencies and causes an increase in setup level, in addition to the streaking effect from the attendant low-frequency phase shift. Such distortion is usually the result of a defective equalizer on long lines, or overcompensation with low-frequency-compensation controls or tilt controls.

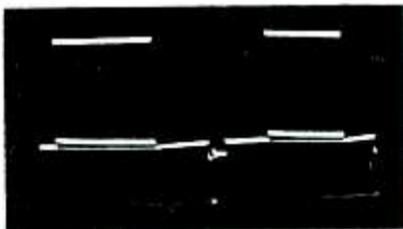
Fig. 8-26 illustrates LD resulting in negative streaking (black-after-white). Fig. 8-26A is the horizontal waveform. Fig. 8-26B shows the typical monitor presentation for this type of picture impairment. Fig. 8-26C illustrates how this form of LD impairs the display of lettering. The vertical-rate display (Fig. 8-26D) indicates clearly the loss of setup, which occurs because this type of phase distortion is the result of insufficient gain at low frequencies, up to about the tenth harmonic of the nominal line-scanning frequency of 15,750 Hz. It will usually be found in practice that the loss of gain occurs below the first few harmonics, or approximately 50 kHz.

The modulated 20T pulse is the most convenient method of displaying RCL and RCT. Fig. 8-27 typifies the display when pure amplitude distortion exists (no phase distortion). A change in am-

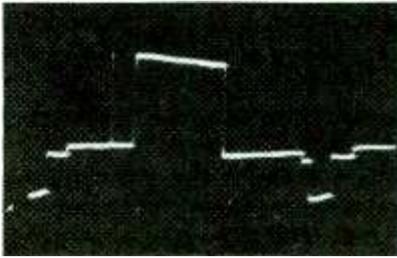
(C) Details of window signal of A.



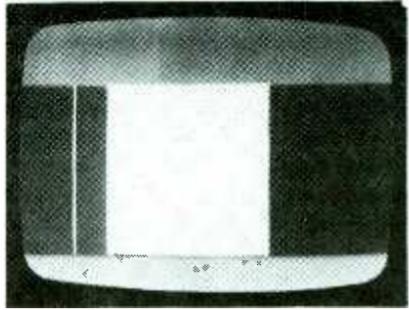
(D) Vertical-rate waveform display.



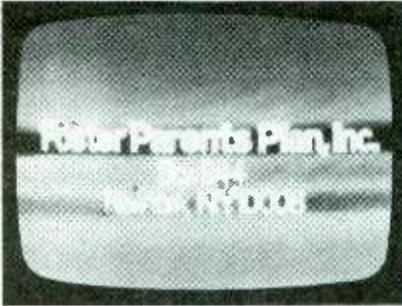
and trailing positive streaking.



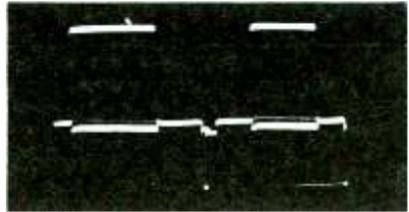
(A) Horizontal-rate waveform display.



(B) Picture-monitor display corresponding to A.



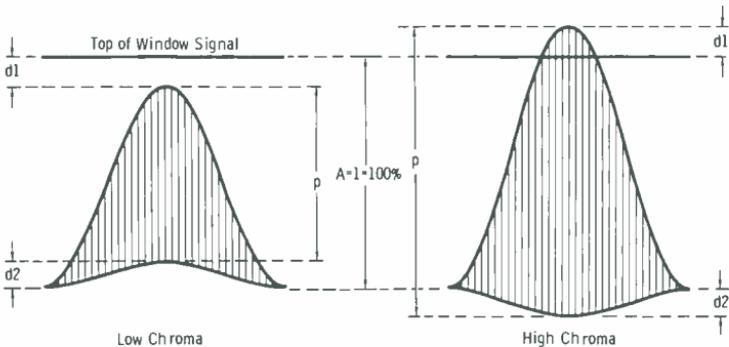
(C) Effect on transmission of lettering.



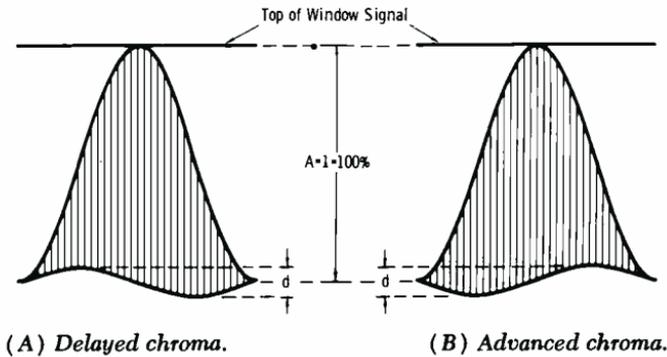
(D) Vertical-rate waveform as displayed on monitor.

**Fig. 8-26. Line-time waveform distortion with negative streaking.**

plitude of the 3.58-MHz subcarrier results in a cosine-shaped distortion of the base line, and a departure from reference peak level (top of window signal). When the distortion is linear, dimensions  $d_1$  and  $d_2$  are equal. If these dimensions are unequal, nonlinear dis-



**Fig. 8-27. RCL only, no phase distortion.**



**Fig. 8-28. RCT only, no amplitude distortion.**

tortion (differential gain) is present. In this case, linear distortion normally can be measured by reducing the test-signal input to one-half the normal input level, or about 0.5 volt peak to peak.

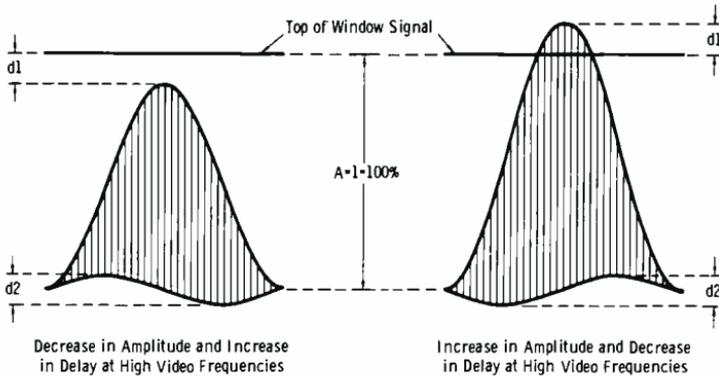
Dimension  $p$  in Fig. 8-27 represents the peak-to-peak level of the 3.58-MHz signal. Therefore (assuming  $d_1$  and  $d_2$  are equal),  $RCL = p$ . Thus assuming  $p$  is 80 percent, RCL is 80 percent, or simply 0.8.

Fig. 8-28 typifies RCT (without amplitude distortion). In Fig. 8-28A, the envelope of the 3.58-MHz subcarrier has a sinusoidal base-line distortion indicating a delay. In Fig. 8-28B, the sinusoidal base-line distortion indicates an advance. Although dimension  $d$  can be expressed as a percentage of  $A$ , the scope display does not provide a very convenient method of specifying the actual group delay in nanoseconds. The maintenance technician normally is interested only in the fact that he has a delayed-chroma or advanced-chroma problem, not in the measurement of actual delay. When this measurement must be determined, a calibrated variable chroma delay or advance is used at the generator output and is adjusted for a flat baseline. The degree of RCT (delay or advance) can then be read directly from the calibrated dial in nanoseconds. Otherwise, RCT may be specified in terms of whether chroma is delayed or advanced, and the percentage of  $d$  to  $A$ .

It is often the case that RCL and RCT distortion occur simultaneously. Fig. 8-29 represents typical displays. The figure is self-explanatory if the preceding two figures are understood.

## 8-8. CHECKING THE VIT SIGNAL

Review Fig. 1-24 and related text for a method of obtaining the VIT display on the oscilloscope. Review Figs. 2-36 and 2-37 and related text for details of the VIT signals.



**Fig. 8-29. RCL and RCT simultaneously.**

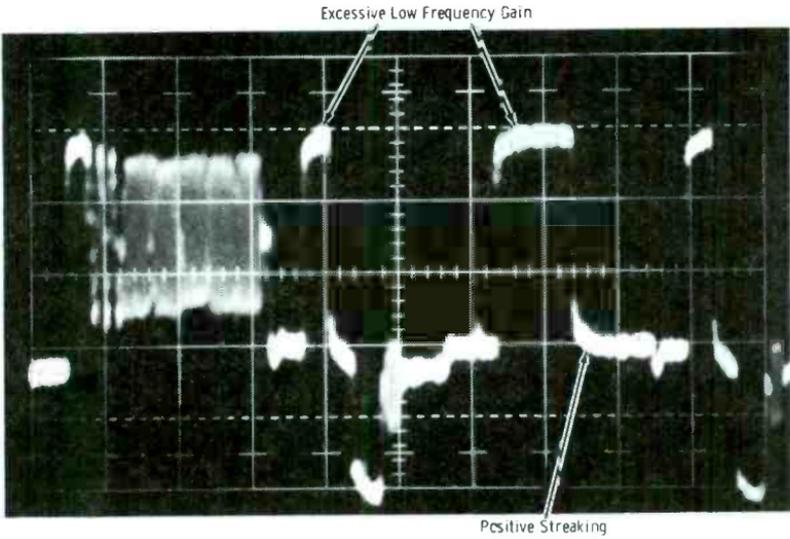
Fig. 8-30 illustrates the portion of the VIT signal normally used for an indication of linear distortion. Fig. 8-30A shows the typical trace of a signal exhibiting positive streaking. Fig. 8-30B shows the same signal after corrective equalization on the faulty leg of the complete route.

NOTE: VIT waveforms appear “noisier” and therefore not as “clean” as full-frame test signals. This is normal.

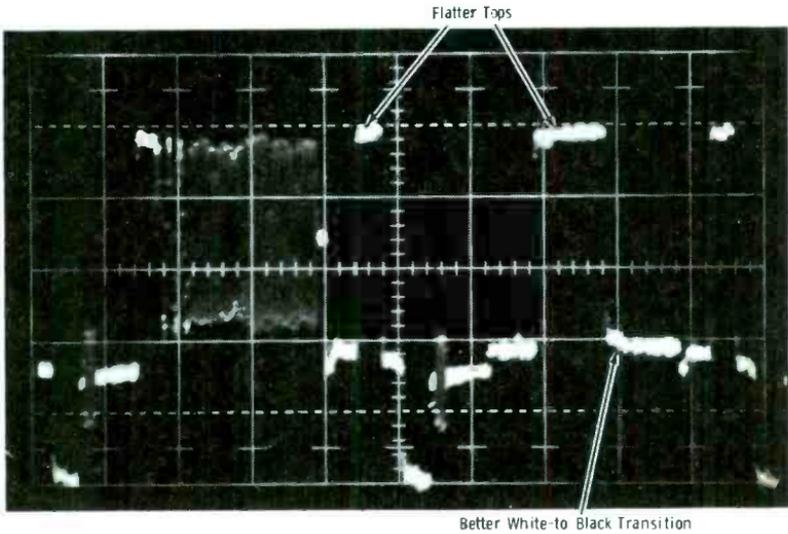
The channels for the intercity transmission of network television program signals are one of the links in the overall television system. Most of these channels are furnished and maintained by the Bell System under Tariff FCC 260. (In the case of some local channels, an independent telephone company may provide the facilities. These cases are included.) The tariff does not define the technical specifications of these channels beyond the broad definition that they are suitable for the transmission, in one direction only, of U.S.-standard, 525-line NTSC color television signals which are to be broadcast by a television station.

The following information is based on an agreement between the technical representatives of the major television networks and those of the Bell System, acting through the Network Transmission Committee, on the performance objectives for these facilities from a source to any receiving point as determined by observation of the prescribed test signals. The prescribed test signals and the methods of applying and measuring them are described below. The signals may be applied either full-frame or during the vertical-blanking intervals as vertical-interval test signals (VITS).

In order to establish a uniform reference objective, it is required that between the source and any receiving point at which observa-



(A) Excessive low-frequency gain and positive streaking.



(B) Waveform of A after corrective equalization.

Fig. 8-30. Portion of VIT signal normally used for linear-distortion indication.

tons are made there shall be only one transmitting and one receiving local channel and that the Bell System shall have the responsibility for and the control of the alignment of all the facilities included.

It is also intended that the performance objectives will apply on a relatively long-term basis. There are no present plans to establish a program of statistical analysis of network performance, so judgment must be used in determining when a report of trouble should be made to the Bell System. In this connection, it should be noted that tolerances are set safely below the point at which a picture impairment would be noticeable. Minor deviations may therefore be tolerated until it is evident that they are a symptom of a chronic condition which should be corrected. In general, such a condition should be expected to be cleared by routine maintenance procedures within a 24-hour period. At the other extreme, if a visible picture impairment or a significant deviation of a test signal is observed, the condition should be reported and corrective action initiated immediately.

In addition to their usefulness in evaluating normal network transmission quality, the performance standards may be used as a guide in evaluating transmission performance when special network configurations are established, as for example, an in-and-out feed through the equipment and facilities of an intermediate station. If it is judged from observation of test signals that such an arrangement is causing undue transmission impairment, cooperative tests by Bell System and network personnel should be initiated to isolate the cause of the irregularity.

We can now go into the procedures for actually measuring the performance characteristics of television network facilities, and state the current objectives. It is inevitable that these objectives will be upgraded continually, and the maintenance technician should keep abreast of developments.

It is essential that oscilloscope adjustments be standard so that uniform displays are obtained at all observing points. This uniformity will facilitate the interchange of information between technicians and, when photographs are mailed to a central location, will make comparisons between photographs more effective.

We will first describe those measurements normally used for linear distortions. Then we will give the current objectives of those signals used for nonlinear distortion, even though the actual measurement procedures are covered in Part 2 of this chapter.

### **Multiburst Signal**

1. When the scope used employs a frequency-response switch, place it in the flat position.

2. Select line 17 of either field. Center the display horizontally.
3. For the IEEE scale, adjust the vertical positioning and gain so that with the blanking line on zero, the middle of the top of the white flag is exactly on 100.
4. The peak-to-peak amplitude of each burst must be between 80 and 95 IEEE divisions. This is 0.57 volt to 0.68 volt peak to peak where 100 IEEE units = 0.714 volt.
5. The peak-to-peak amplitude of the color-sync burst must be between 36 and 44 IEEE units. This is 0.26 volt to 0.31 volt peak to peak.
6. For photographing, adjust the trace intensity and/or scale lighting so that the scale and the displayed trace both appear with good contrast.
7. For recording, record both the top and bottom peak scale position of each burst as well as the peak-to-peak amplitude of the burst.

### **T Sine-Squared Pulse**

1. Use flat response of the scope.
2. Select line 18 of field 1.
3. Center the pulse horizontally with maximum horizontal magnification ( $\times 25$ ).
4. Adjust the vertical positioning and gain so that with the pedestal (actual base of the pulse) at zero, the top of the pulse is exactly at 100 on the IEEE scale.
5. The peak-to-peak amplitude of the overshoots, either leading or following the pulse, whichever is larger, must not be greater than 25 percent.
6. For photographing, since the pulse trace is rather dim, the scale lighting should be reduced accordingly and a longer exposure made. Experiment is called for here.
7. For recording, record the peak-to-peak amplitude from the first negative to the first positive peak of the overshoot, either leading or lagging, whichever is greater.

### **Window Signal**

1. Use IRE rolloff if available (easier to read).
2. Select line 18 of field 1.
3. Center the window portion of the signal horizontally using appropriate horizontal magnification to the extent that it improves the display.
4. Adjust the vertical positioning and gain so that with the pedestal (actual base of the window) at zero, the middle of the top of the window is exactly at 100 on the IEEE scale.

5. The slope across the top of the window must not exceed 4 percent. In order to eliminate distortions at the transistions, take the reading on the middle half of the window, and double this reading to represent the overall slope.
6. For recording, record in IEEE scale divisions the slope across the middle half of the top of the window and multiply by 2.

### **Stairstep Signal (Differential Gain and Phase at 3.58 MHz)**

1. Use the differential-gain characteristic on oscilloscopes equipped with an internal high-pass filter.
2. Select line 18 of field 2.
3. Adjust the vertical positioning and gain so that the largest step displayed is exactly 100 IEEE divisions peak to peak centered between the 0 and 100 scale markings.
4. The difference in peak-to-peak amplitude between the smallest-step burst and 100 must not be more than 15 IEEE units, or 15 percent.
5. Differential phase must not exceed 5 degrees.
6. For recording, record the difference between the peak-to-peak amplitude of the smallest step displayed and the segment at 100. Also note whether the envelope of the bursts, progressing from left to right, slopes upward (positive differential gain) or downward (negative differential gain).

**IMPORTANT NOTE:** Techniques for measurement of differential gain and differential phase are presented, along with waveform photographs, in Part 2 of this chapter.

## **Part 2: Nonlinear Distortions**

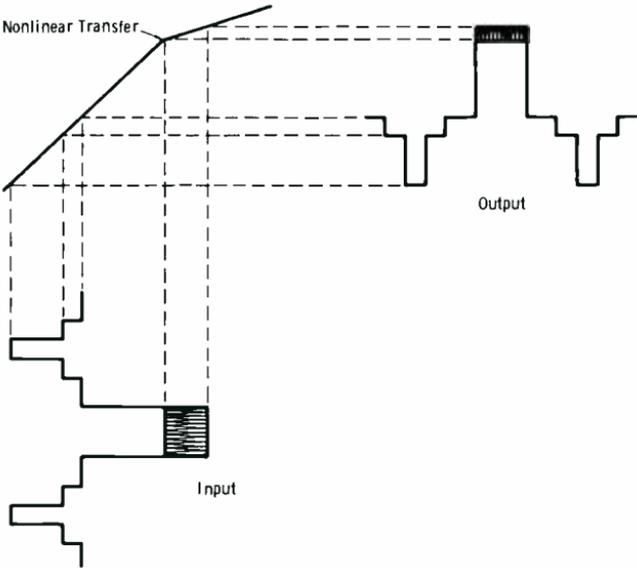
In this part of the chapter, we will consider the measurement of distortions which are amplitude-dependent. It must always be kept in mind that this assumes the proper operating amplitude range of the equipment is not exceeded.

### **8-9. DEFINITIONS OF TERMS**

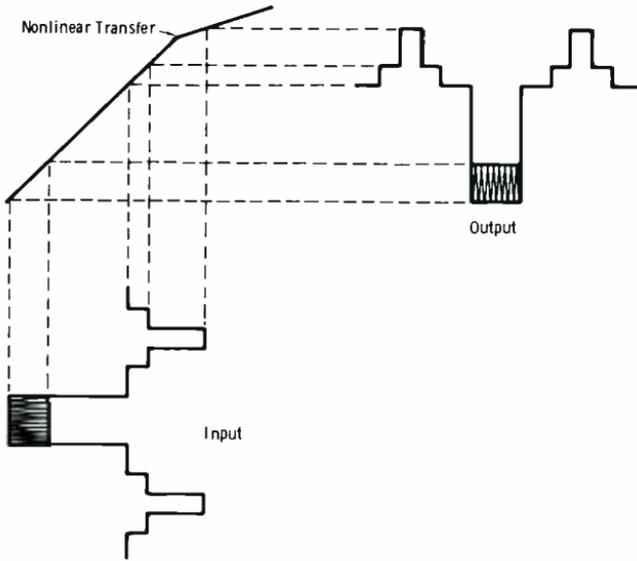
Before we proceed, it is important to have a thorough understanding of the meaning of certain terms. These basic terms are defined in this section.

#### **Incremental Gain Distortion**

Incremental gain distortion is the basic cause of amplitude and phase nonlinearity (Fig. 8-31). When the transfer curve is not linear, clipping or compression (depending on the severity of departure from the linear) occurs. Considering a single stage, the com-



(A) Compression of rf component in white direction.



(B) Compression of sync pulse.

Fig. 8-31. Effect of polarity on portion of signal compressed.

pression may occur in either the white or sync direction, depending on the polarity of the input signal to the offending stage.

The shaded area of the signals in Fig. 8-31 represents an rf component, such as the color subcarrier in the region of 3.58 MHz. If the tip of the white pulse in Fig. 8-31A represents peak white, no white compression would occur, and no effect at all would be evident on a monochrome receiver. However, the color component, such as could occur in this region on saturated yellows or cyan, would be seriously compressed. This results in the following three major errors:

1. The hue represented by the color signal will be washed out in highlights and completely lost if the degree of nonlinearity is at a clipping level.
2. Differential gain (defined below) will exist.
3. Differential phase (defined below) will exist and will actually shift the intended hue to a different color toward the highlight areas. For example, yellow could be faithfully reproduced in the lowlight areas, but could change toward a desaturated green or red in the highlight areas.

In practice, the shape of the transfer curve can result in any degree of stretch or compression anywhere from black to white. If the transfer should take an "S" shape (such as occurs in most uncompensated visual transmitters), both white and black are compressed, and grays are stretched.

In Fig. 8-31B the sync region is compressed. If the departure from the linear region is at a lower point, even the video setup level can be lost. Thus there exist two major causes of loss of setup in a transmission path: loss of low-frequency response, as previously described, and incremental gain distortion resulting in a nonlinear transfer.

### Differential Gain

Differential gain is the difference between (a) the ratio of the output amplitudes of a small high-frequency sine-wave signal at two stated levels of a low-frequency signal on which it is superimposed, and (b) unity. Differential gain may be expressed as a percentage by multiplying this difference by 100, or it may be expressed in dB by multiplying the common logarithm of ratio (a) by 20. In this definition, level means a specified position on an amplitude scale applied to a signal waveform. The low- and high-frequency signal must be specified. This definition is based on that established by IRE subcommittee 23.4 and approved by the American Standards Association (ASA).

Note that this measurement technique is similar in every respect to intermodulation-distortion methods in audio. In practice, differential gain is normally expressed in percent rather than decibels. The low-frequency component for television measurement is the line frequency of 15,750 Hz; the high-frequency superimposed component is a sine wave at the color-subcarrier frequency of 3.579545 MHz (which, for simplicity, we refer to as 3.58 MHz).

Notice also that the word "distortion" is absent in the definition. Any amount of differential gain is distortion. Therefore the addition of this term would be redundant.

### **Differential Phase**

Differential phase is the difference in output phase of a small, high-frequency sine-wave signal at two stated levels of a low-frequency signal on which it is superimposed. In this definition, level means a specified position on an amplitude scale applied to a signal waveform. The low- and high-frequency signals must be specified. This definition is based on that established by IRE subcommittee 23.4 and approved by the ASA.

Differential phase is measured in degrees with the same type of test signal as that used for differential gain. Again, in this case the term "distortion" is redundant.

### **Summary of Definitions**

Incremental gain distortion results in both differential gain and differential phase. When the color signal (the high-frequency component) is clipped or compressed, saturation errors in the colors result. Even though the gain at 3.58 MHz is brought back to normal after compression, the axis is shifted, resulting in a different phase relative to the reference. Incremental gain distortion can result in luminance errors in monochrome transmission and/or sync compression or loss of setup level.

Differential gain means that the chrominance signal level changes as the luminance signal varies between black and white. As stated before, this produces saturation error in the color information. Provided there is no compression or clipping of the luminance level, no effect harmful to monochrome transmission will be produced.

Differential phase means that the phase of the chrominance signal (which represents hue) changes as the luminance signal varies from black to white. This results in shifting of actual hues as the brightness varies. For a monochrome picture, differential phase at 3.58 MHz is not important, except that the measurement is an excellent criterion of overall monochrome system performance.

NOTE: Before proceeding, it is suggested that the reader review Section 2-5 of Chapter 2.

### 8-10. FILTER FOR DIFFERENTIAL-GAIN MEASUREMENTS

Modern waveform monitors incorporate suitable (and selectable) internal filters for various purposes. When such a capability is not available, or when an external test scope is to be used, an external filter is necessary.

Equipment necessary to measure amplitude linearity (low-frequency) and differential gain at 3.58 MHz is quite simple. All that is required is a wideband scope and a crossover network of the type shown in Fig. 8-32. This filter incorporates a switch for select-

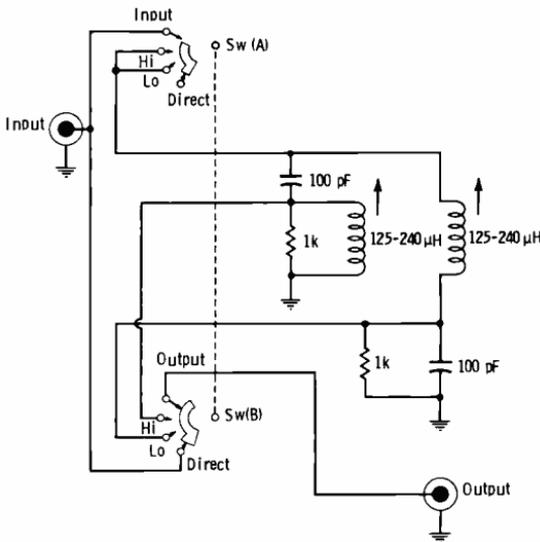
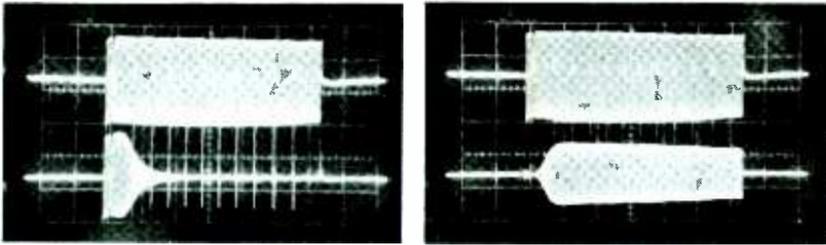


Fig. 8-32. Crossover filter for 3.58-MHz checks on staircase.

ing direct, low-pass, and high-pass positions. The standard response of the various positions should be as shown in Fig. 8-33. The markers are at 1-MHz intervals to 10 MHz. Note that in the low-pass position the response is negligible at 3.58 MHz, and only the low-frequency 15,750-Hz component of the signal is passed (Fig. 8-33A). In the high-pass position, the response is negligible at frequencies below about 0.5 MHz, while 3.58 MHz is passed (Fig. 8-33B). Note that any loss in amplitude of the 3.58-MHz component is not of interest, since only the *difference* in levels at the steps between black and white is important.

The measurement of differential phase requires special equipment, such as the Tektronix vectorscope, the RCA color-signal analyzer, or the Telechrome video-transmission test-signal receiver.

(A) *Low-pass position.*(B) *High-pass position.***Fig. 8-33. Characteristics of crossover filter.**

A burst regenerator is required when the measuring equipment is remote from the sending equipment. For example, if measurement of an stl is required, the sending equipment with the reference 3.58-MHz oscillator is located at the studio, but the measuring gear must be at the transmitter. In this case, it is necessary to frequency-phase lock a stable reference oscillator in the measuring equipment with the signal received from the studio. The Tektronix vectorscope is capable of this mode of operation.

The basic action is as follows. The test signal is applied to two paths: (A) a crystal filter circuit that converts the 3.58-MHz signal component into a constant-amplitude, constant-phase reference, and (B) a bandpass amplifier that amplifies only the 3.58-MHz signal (and its associated sidebands when a color signal is used) so that the amplitude and phase of each of the 3.58-MHz steps is preserved as received. (The test signal is the same as that used for differential gain, as described previously.)

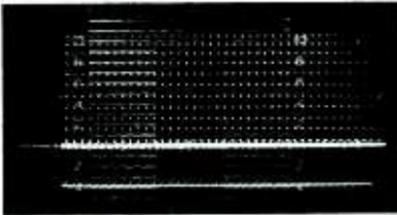
The reference sine wave and the amplified signal are applied to a phase demodulator the output of which is amplified and viewed on the cro. To measure differential phase, the reference-signal phase is adjusted so that it is nominally  $90^\circ$  out of phase with the signal. Under this condition, zero output from the demodulator is obtained. Slight differences from  $90^\circ$  result in an output that is proportional to the amount of phase difference. The horizontal-sync pulse is used as a reference line, since no 3.58-MHz signal exists during this interval. Differential phase of the steps is measured by varying the phase of the reference with a calibrated dial and successively bringing the sync step in line with the other steps. Phase difference between the steps is read from the dial.

## 8-11. AMPLITUDE LINEARITY AND DIFFERENTIAL GAIN

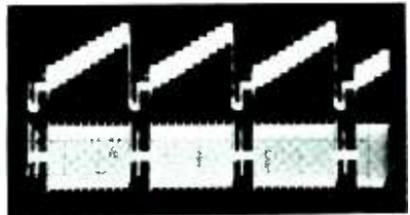
As in every type of testing, the first step is to know exactly the back-to-back characteristics of the test generator and measuring

equipment. This is to emphasize that you must first provide a standard for system comparison. Check the linearity of the steps on the oscilloscope, and use the internal adjustments of the stairstep generator to obtain a step on every 10 IEEE scale units. It is most convenient to employ a long time base on the scope and observe the horizontal-rate pulses at a vertical rate (see Fig. 8-34). This makes possible a most precise adjustment of linearity of the steps. Note that for an uncluttered view the pattern in Fig. 8-34 has been vertically displaced slightly from the graticule lines.

Next, be certain the generator itself does not produce differential gain on superimposed 3.58-MHz signals. The upper trace of Fig. 8-35 shows the direct generator output, whereas the lower trace is the 3.58-MHz component obtained through the high-pass filter. (The scope gain was increased for the lower trace.) Differential gain from the generator should be zero.



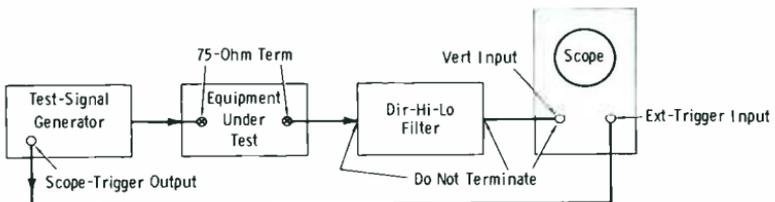
**Fig. 8-34.** Stairstep-generator output waveform displayed at vertical rate with long time base.



**Fig. 8-35.** Displays used to check output of stairstep generator for differential gain.

Fig. 8-36 shows the proper setup for amplitude-linearity and differential-gain measurements. The external trigger supplied by the test-signal generator is important for trace stability, particularly when the one-in-five-lines (variable-APL) signal is used.

Amplitude linearity normally refers to the luminance scale represented by the low-frequency step component. The upper trace of Fig. 8-37 shows the stairstep input signal, either without the 3.58-MHz signal superimposed or as viewed through the low-pass



**Fig. 8-36.** Test setup for checking amplitude linearity and differential gain.

filter when the 3.58-MHz signal is present. The lower trace shows almost complete clipping of the last (white) step.

The upper trace of Fig. 8-38 shows the output of a system under test, viewed with the filter in the direct position; compression in the white region is quite obvious. The actual amount of differential gain is measured by viewing this signal through the high-pass filter as in the lower trace. (The scope gain was increased for the lower trace.) Note that the transients (spikes), which are due to the short rise times of the steps, provide convenient markers to indicate the different steps.

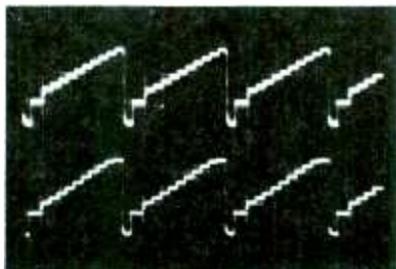


Fig. 8-37. Stairstep displays through low-pass filter.

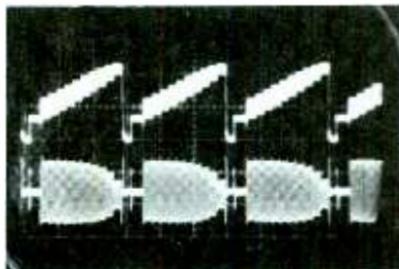


Fig. 8-38. Differential gain in white region.

Differential gain normally is expressed as a percentage. On a wideband scope, the 3.58-MHz component through the high-pass filter may be “blown up” so that the highest block is 100 IEEE units, or 100 percent, and the amount of compression (smallest block) can be read in percent response. The peak-to-peak block amplitude is considered in the measurement. However, it is useful information to state whether the measurement involves compression or expansion, and there are certain ambiguities that must be avoided.

In practice, it is most convenient to use the following relationships in the measurement of differential gain: *Compression*:

$$\text{Differential Gain} = (1 - b/a) \times 100$$

where,

a is equal to the height of the uniform (linear) portion of the signal,

b is equal to the height of the *smallest* block of the signal.

*Expansion*:

$$\text{Differential Gain} = (b/a - 1) \times 100$$

where,

a is equal to the height of the uniform (linear) portion of the signal,

b is equal to the height of the *largest* block of the signal.

An example to illustrate the precaution required in the computation of compression and expansion follows. Assume a is 20 IEEE units and b is 15 IEEE units. Then, for compression, we have:

$$\begin{aligned} \text{Differential Gain} &= (1 - 15/20) 100 \\ &= (1 - 0.75) 100 \\ &= (0.25) 100 \\ &= 25 \text{ percent (compression, or } \textit{negative} \\ &\quad \text{differential gain)} \end{aligned}$$

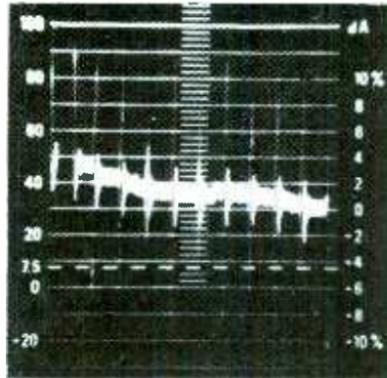
Now assume a is 15 IEEE units and b is 20 IEEE units. Then, for expansion, we have:

$$\begin{aligned} \text{Differential Gain} &= (20/15 - 1) 100 \\ &= (1.33 - 1) 100 \\ &= (0.33) 100 \\ &= 33 \text{ percent (expansion, or } \textit{positive} \\ &\quad \text{differential gain)} \end{aligned}$$

Tests should be run at all three standard APL levels. (Review Figs. 2-27 and 2-28 of Chapter 2.) It is obvious that the 50-percent APL condition most nearly simulates average transmission. Where time or facilities are limited, the 50-percent APL condition should be selected. However, a complete story of system performance can be obtained only from tests at 10, 50, and 90 percent APL.

The Tektronix Type 520 vectorscope has a provision for measuring differential gain, and the display is different from those just discussed. See Fig. 8-39. The differential-gain mode of operation is selected by pushing a button, and the 100%-75%-MAX GAIN switch is set to the Max Gain position. In this mode, a 5-percent change deflects the trace 5 percent, or 25 IEEE units. If there is no differential gain, the 3.58-MHz components of all steps of the 10-step staircase signal will lie on a straight line. The photo of Fig. 8-39 indicates that the first step is 3.6 percent higher than the tenth step. Thus 3.6-percent negative differential gain is present. (The 3.58-MHz component of the tenth, or whitest, step is 3.6 percent lower in amplitude than the 3.58-MHz component of the darkest step.) The IEEE scale on the left and the differential-amplitude scale on the right are automatically illuminated when the DIFF GAIN button is depressed. In the normal mode of operation, the conventional vector graticule is displayed.

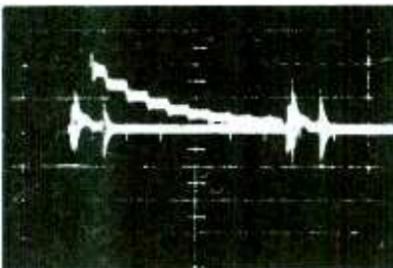
**Fig. 8-39. Differential-gain display on Tektronix Type 520 vectorscope.**



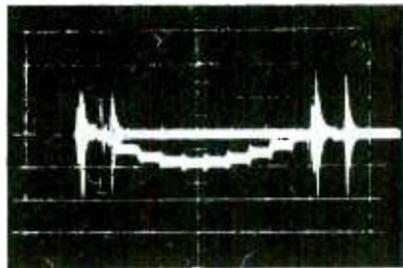
## 8-12. DIFFERENTIAL PHASE

Differential phase is measured with the same type of test signal as that discussed in previous sections, with special measuring equipment basically described in Sections 8-10 and 8-11.

The output of the system is fed to the measuring device to determine the phase shift in degrees. Fig. 8-40 shows typical traces obtained from the measuring equipment with a stairstep signal. The differential phase can occur (show major departure from a straight line) at the end of the trace (toward white or black), or it can occur in mid trace (gray area). In the case of Fig. 8-40A, it occurs at the end of the trace, so one end is brought to the reference level by the position control with the calibrated phase knob set on zero. Then the calibrated knob is adjusted to bring the other end of the trace to the reference line, and the phase difference between the ends of the trace is read on the calibrated knob. The same procedure is used for Fig. 8-40B, except the ends are first nulled with the position control (calibrated knob on zero), and



*(A) Right-hand end adjusted to reference.*



*(B) Left and right ends of trace adjusted to reference.*

**Fig. 8-40. Traces obtained in differential-phase measurement.**

then the calibrated knob is adjusted to null the maximum center amplitude. Then the phase difference is read from the calibrated knob.

In general, a system which has little differential gain will also have minimum differential phase. Compression or clipping of the 3.58-MHz component that results in differential gain will also result in differential phase. However, in rare cases (more rare at the studio than at the transmitter) differential phase can exist even though differential gain is very small. Parallel paths in chrominance and luminance channels, and impedance elements which may have constant impedance at line scanning frequencies but are variable in the 3.58-MHz region, can result in differential phase. The latter condition is more likely to occur in the transmitter than at the studio.

**Table 8-3. Differential Phase Data**

APL	$\Delta\theta$ in Degrees	Amplitude Region
10%	0	Black
	+0.5	Center
	-3	White
50%	0	Black
	+1	Center
	-2	White
90%	0	Black
	+1	Center
	-1	White

Note that 0° reference is at blanking level.

Differential phase is normally tabulated in either of two ways, as follows:

1. The values of differential phase with respect to the value at the blanking level. "Plus" implies leading phase; "minus" implies lagging phase (see Table 8-3).
2. The maximum range of the differential phase (difference of extreme values). Note that for the tabulated data of method 1, method 2 would simply list differential phase as 3.5 degrees.

In general, the studio equipment connected to the transmitter input terminals should show a maximum of 1.5° differential phase at 3.58 MHz. The transmitter should show a maximum of 3.5° for an overall allowance of 5°. The maximum overall differential phase (from origin to transmitter output) allowed by the FCC is 10°. This overall tolerance may be tightened at any time; always check current FCC Rules.

The Tektronix Type 520 vectorscope is also equipped to measure differential phase. The vertical deflection of the display is greatly magnified and is inverted on alternate lines. This allows the use of a trace-overlay technique for measuring small phase changes between black and white; the standard linearity test signal is used. Fig. 8-41 illustrates the resulting display. In Fig. 8-41A, the first step of the staircase has been nulled, by means of the variable phase control with the calibrated phase control on 0. Note that in this particular instance, the greatest difference occurs at step 5. The calibrated phase control is turned to null step 5, as shown in Fig. 8-41B. Differential phase is then read directly from the calibrated phase control, which also indicates whether plus or minus values are involved.

### 8-13. THE BASIC INTERNATIONAL WAVEFORM GRATICULE

The Satellite Technical and Operational Committee-Television (STOC-TV) was established in North America in 1967. This committee was formed in recognition of the need to develop unified operating practices, technical quality criteria, and standards for international television connections using communications satellites. Its membership comprises companies operating intercontinental satellite links and domestic satellite systems, and also major terrestrial common carriers and broadcasting networks in the United States and Canada.

In September of 1974, an ad hoc subcommittee was formed to recommend a standard operational graticule for waveform monitors used to check the technical quality of long-distance connections containing links by satellite. The subcommittee completed its work early in 1976, and the main committee approved the guidelines in July of 1976. Fig. 8-42 illustrates the basic graticule (for NTSC color) which resulted from this study. The Tektronix 1480-series waveform monitors use this graticule, and it is recommended for network-affiliated stations in the interest of a "common language" for spelling out transmission characteristics.

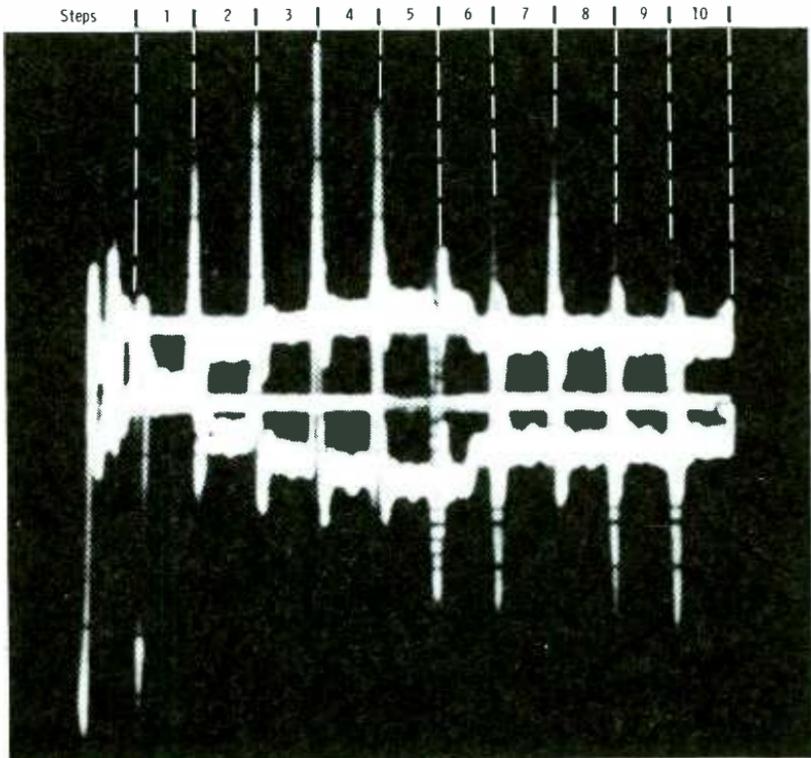
The basic graticule is inscribed on the face of the cathode-ray tube to minimize parallax error. An additional graticule may be mounted externally when required. To reduce the possible masking of fine waveform details as a result of an overbusy or cluttered graticule view, graticule lighting on the waveform monitor is arranged so that only one graticule is visible and the other graticule is subdued so as not to interfere with the measurement.

The video level scale of Fig. 8-42 is the IEEE scale standardized for operational level measurements (140 IEEE units = 1 volt). The IEEE scale division markings are shown on the left, and car-

rier percentage markings are shown on the right. The nominal black-level setup (7.5 IEEE units) is shown as a dashed line. In addition, three fine-division scale segments are included to enhance precision of sync-level, blanking-level, and white-level measurements.

The numbered guidelines of Fig. 8-42 are as follows:

1. The limit outline for the measurement of line-time linear distortion. The box (labelled L.D.) is for a 5-percent tolerance limit; the short lines inside indicate the 2-percent line-time distortion level. Obviously, with  $\times 2$  vertical amplification, the outline and the interior marks would indicate  $2\frac{1}{2}$ - and 1-percent tilt, respectively.
2. The pulse-to-bar ( $K_{pb}$ ) ratio scale is graduated in terms of the quality rating factor and has divisions from +4 percent to -5 percent in 1-percent increments. These limits apply

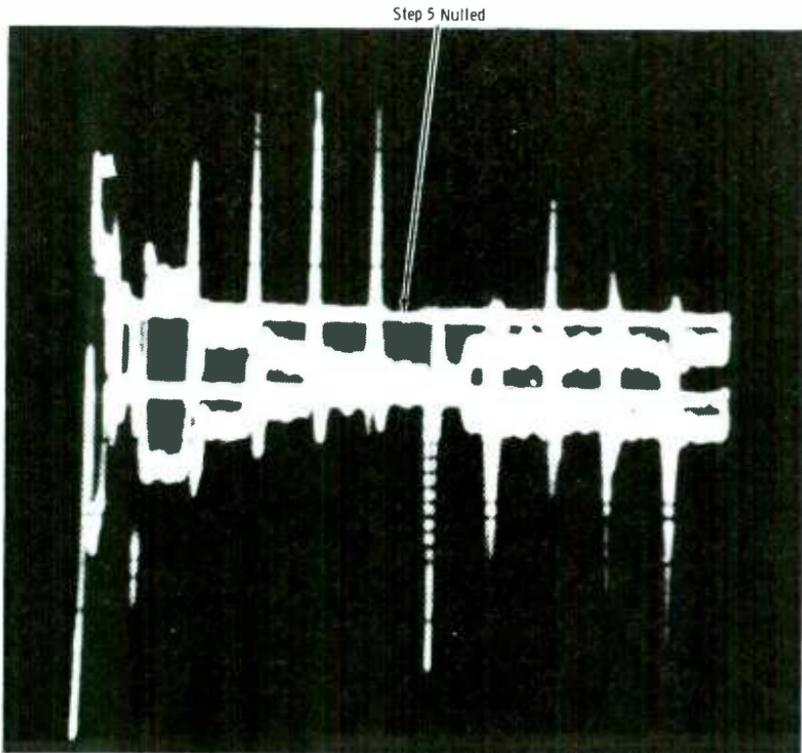


(A) Bottom step nulled.

Fig. 8-41. Differential-phase measurement

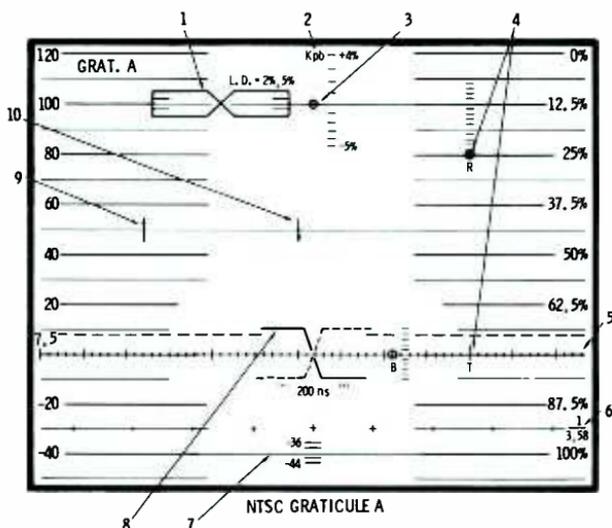
to  $\times 1$  vertical amplification, and obviously would be halved for  $\times 2$  amplification.

3. An unlabelled target point used for alignment of the external graticule, which has a corresponding target point.
4. Note the accentuated division labelled  $\tau$  on the 0 IEEE unit line, aligned with the auxiliary target point  $\kappa$  above it on the 80 IEEE unit (25-percent modulation) line (RT stands for "rise time"). A time scale is provided on the 0 IEEE unit line (5) for the measurement of waveform durations. The value of the divisions along this line depends on the sweep time setting. By adjusting a waveform transition (pulse edge) for a deflection of 100 IEEE units, aligning it vertically between the  $-10$  IEEE and  $+90$  IEEE levels, and aligning it horizontally so that it passes through point R, it is possible to measure the rise or fall time in terms of divi-



(B) Greatest error nulled.

with Tektronix Type 520 vectorscope.

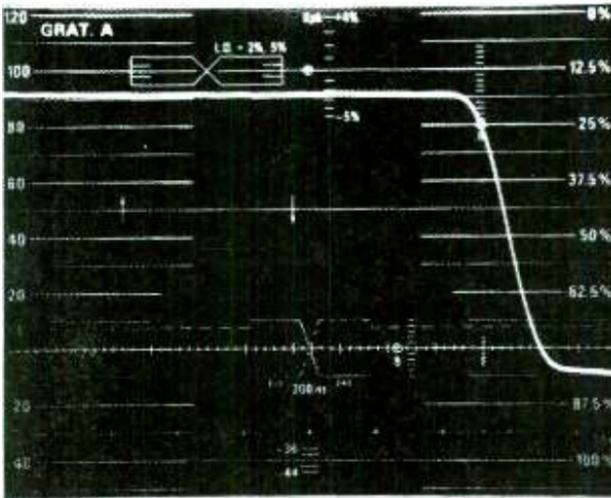


**Fig. 8-42. Basic international standard waveform-monitor graticule for use with 525-line NTSC color television system.**

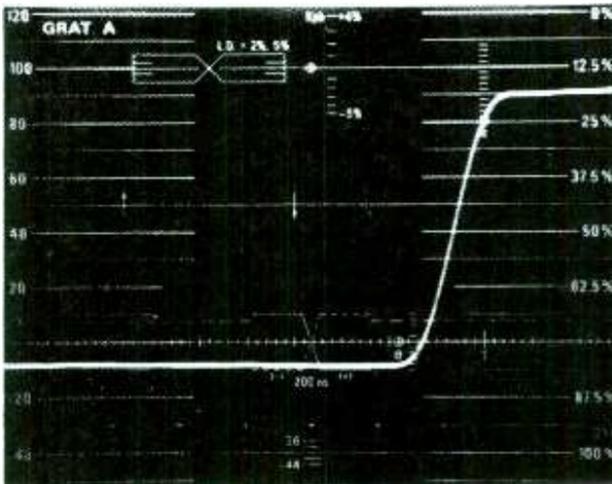
sions to the right or left of division T. This procedure measures the rise or fall time between the 10- and 90-percent points of the waveform, as required.

Fig. 8-43 is an example of measuring the rise and fall times of horizontal sync. Fig. 8-43A shows the leading edge of sync at  $0.1 \mu\text{s}$  per major division, indicating a rise time of  $0.12 \mu\text{s}$ . The same time base is used in Fig. 8-43B, indicating a trailing-edge fall time of  $0.125 \mu\text{s}$ . Maximum rise and fall time of all pulses is  $0.003H$  ( $0.19 \mu\text{s}$ , or 190 nanoseconds) leaving the studio, and  $0.004H$  ( $0.25 \mu\text{s}$ , or 250 nanoseconds) from the transmitter.

5. The 0 IEEE unit line basically described under (4) above. The target point labelled B on this line is for the correct placement of the baseline of a test waveform such as a  $\sin^2$ -window signal. With standard VITS, baseline point B will fall between the  $12\frac{1}{2}T$  modulated pulse and the bottom step of the modulated staircase.
6. The  $-30$  IEEE unit line where a special scale labelled  $1/3.58$  is provided. The divisions on this line correspond to a half wavelength of the color subcarrier when the prescribed horizontal sweep rate is used, or to a full wavelength when a sweep rate half as fast is used. This scale serves for quick checks of timing accuracy by using the color burst to see if its zero crossings coincide with the scale divisions.



(A) Leading edge of sync.



(B) Trailing edge of sync.

Courtesy Tektronix, Inc.

Fig. 8-43. Measuring rise and fall times of horizontal sync.

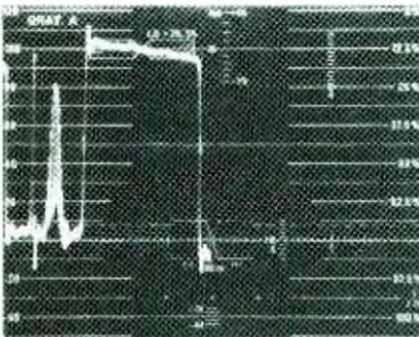
For example, on the Tektronix 1480-series waveform monitors, with the time base set to  $0.1 \mu\text{s}/\text{division}$ , you can position the zero-axis crossings of the color burst to check that there is one-half cycle of burst per mark. This verifies the  $0.1\text{-}\mu\text{s}/\text{division}$  timing of the monitor. Then with the time base set at  $0.2 \mu\text{s}/\text{division}$ , position the color burst so

that it passes through the 3.58-MHz marks on the graticule. Check that there is 1 cycle of burst per mark. This verifies the  $0.2\text{-}\mu\text{s}/\text{division}$  timing accuracy.

7. Fine-division scale segments to enhance the precision of adjustments and measurements of sync-tip level.
8. These limit outlines on the 0 IEEE (baseline) scale are for the measurement of relative chroma time when the relative chroma level is zero. The solid outline is for positive and the dashed outline for negative relative chroma time. Both indicate a 200-nanosecond limit with  $\times 1$  vertical amplification.
9. The target point with the ascending arrow (9) is for correct placement of the leading edge of the bar waveform. A bar waveform of the duration required for correct use of the L.D. box mentioned under (1) above would then have its trailing edge go through the target point which has a descending arrow (10).

As a practical example, see Fig. 8-44. This shows an off-the-air measurement made through a standard demodulator. The display indicates 5-percent line-time distortion.

10. See (9) above for the use of this target point.



**Fig. 8-44. Display indicating 5% line-time distortion.**

Courtesy Tektronix, Inc.

### Part 3: The Audio

In the early days of television, we were concerned with the transition of audio (radio) technicians to the art of picture construction and transmission. Now, a "new breed" of technicians has appeared, whose training has concentrated on the visual medium, to the possible detriment of adequate training in the aural portion of broadcasting. In this part, we will provide the necessary background in audio to make meaningful the measurements and main-

tenance techniques involved in the all-important sound portion of the studio installation.

Understanding the terminology and correct usage of signal generators, meters, and measuring devices used in broadcasting is vitally important to testing and maintenance procedures. Therefore, considerable attention will be given to the units of measurement and their application.

#### 8-14. BASIC DEFINITIONS IN AUDIO

When sound is increased in magnitude, the *loudness* is said to increase, the impression to the brain being roughly proportional to the logarithm of the ratio of the acoustical power of the two sound levels. Loudness is a complex function dependent on many variables and is covered more fully in other texts.

For example, suppose a speaker driven with 1 watt has its driving power increased to 2 watts. It is meaningless to say that the power was increased by 1 watt unless it is also stated that the original power was 1 watt. What is important is that the power was doubled. The ear interprets this as a certain change in loudness; but the same degree of change is perceived with an increase of only  $\frac{1}{2}$  watt if the original power was  $\frac{1}{2}$  watt, or with an increase of 2 watts if the original power was 2 watts.

The common logarithm of the ratio of two powers is an expression of their relationship in bels:

$$\text{bels} = \log_{10} \left( \frac{P_2}{P_1} \right)$$

where,

$P_1$  is the reference power,

$P_2$  is the power being compared to  $P_1$ .

The bel is too large a unit for practical use in broadcasting work, so a unit equal to one-tenth of a bel, the decibel (dB), is commonly used. Therefore the difference in level between  $P_1$  watts and  $P_2$  watts is given by:

$$\text{dB} = 10 \log_{10} \left( \frac{P_2}{P_1} \right)$$

To avoid cumbersome computations, tables and graphs are normally used. (See the decibel table in Appendix A for the conversion of ratios to decibels.) Note that a power ratio of 2/1 is 3.01 dB; this is normally stated as 3 dB.

Zero dB may designate any convenient reference level. Although it is normally based on the ratio between two powers, the dB can also indicate absolute power, provided the reference level (zero

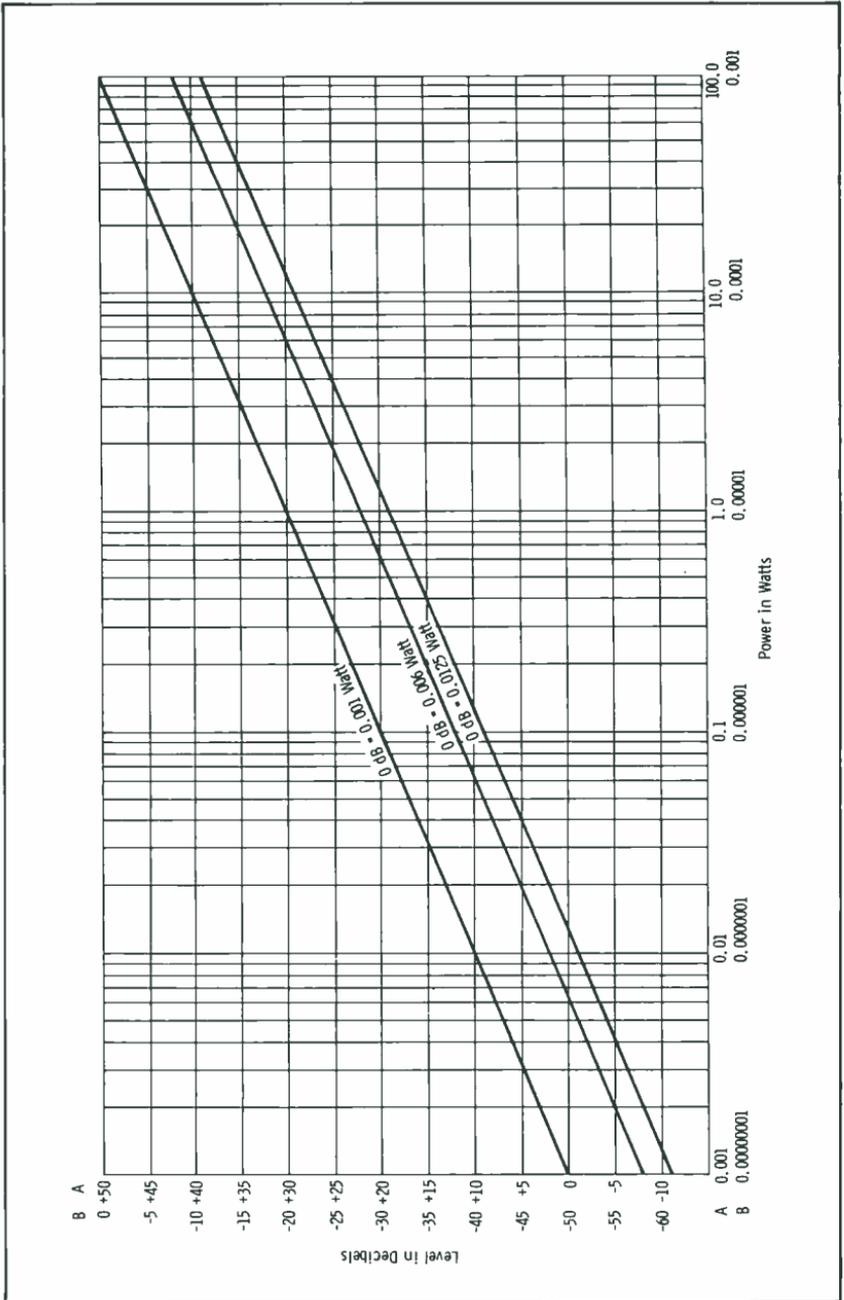


Fig. 8-45. Relationship of decibels and power level for three reference levels.

level) is specified. In the past, so-called standard reference levels have been variously specified as 1, 6, 10, 12.5, and 50 milliwatts (mW). The 1-mW reference level is most widely used today, but the practicing engineer will occasionally find 6 and 12.5 mW referred to as 0 dB. The term *dBm* is used to indicate that zero level is 1 mW. Note that power levels expressed in dB are independent of impedance values. Fig. 8-45 shows a graph of dB versus power for three reference levels. Note from Fig. 8-45 that:

1. To convert from a 1-mW reference to a 6-mW reference, add  $-7.78$  dB.
2. To convert from a 1-mW reference to a 12.5-mW reference, add  $-10.97$  dB.
3. To convert from a 6-mW reference to a 1-mW reference, add  $+7.78$  dB.
4. To convert from a 12.5-mW reference to a 1-mW reference, add  $+10.97$  dB.

With any reference level, a plus sign indicates so many "decibels up." A minus sign indicates so many "decibels down." The statement of a *power ratio* in dB is independent of reference level or impedance. The statement of *absolute power* in dB is meaningless unless the reference level is stated.

When decibels are related to voltage or current, the value of impedance must be taken into account since the voltage across or the current through an impedance depends on the impedance as well as the power level:

$$E = \sqrt{WR}$$

where,

E is the voltage across the impedance,

W is the power in watts,

R is the impedance (purely resistive) in ohms.

Power is proportional to the square of the voltage or current:

$$W = \frac{E_2^2}{R} = I^2R$$

When a number is squared, the logarithm of that number is doubled; therefore, when considering dB relative to voltage ratios, the following relation applies:

$$\text{dB} = 20 \log_{10} \left( \frac{E_2}{E_1} \right)$$

When considering current ratios, the relation is:

$$\text{dB} = 20 \log_{10} \left( \frac{I_2}{I_1} \right)$$

The decibel table in Appendix A lists dB relative to voltage or current ratios as well as power ratios. Note that for a given ratio of power, the number of decibels is one-half the value for the same ratio of voltage or current.

The vtvm is a convenient tool that is often used in gain or loss measurements. Fig. 8-46 shows the dBm-to-volts relationship for the impedances given.

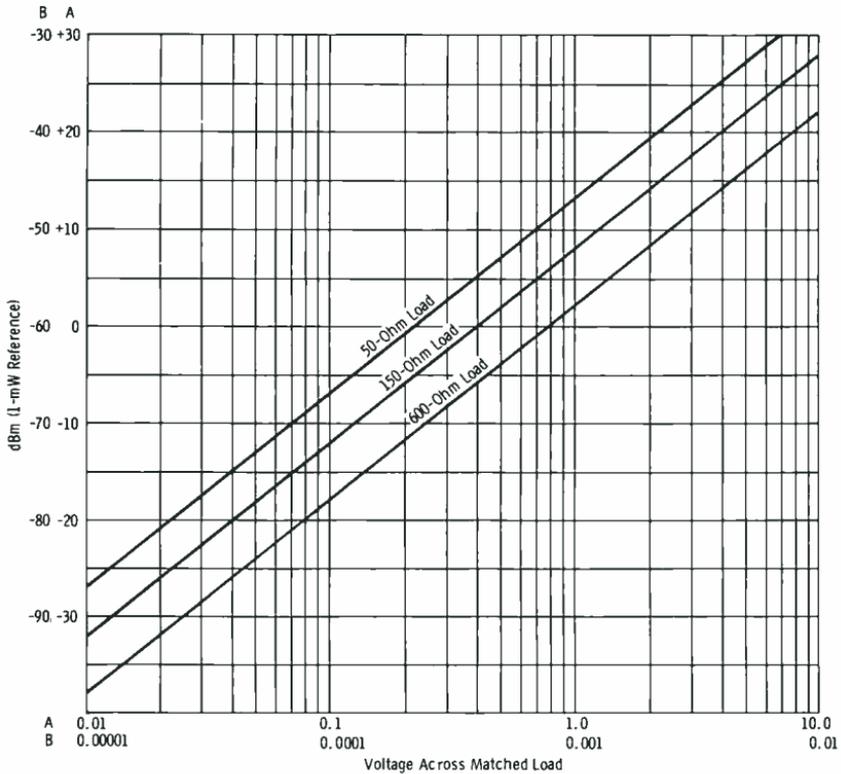


Fig. 8-46. Relationship of dBm and load voltage.

## 8-15. VOLUME UNITS AND VU METERS

The VU meter is a standardized instrument intended for the monitoring of audio program content. Since the power in program signals is constantly fluctuating, the meter reading must be standardized as to whether it is a peak, rms, or average reading, and

the meter must have specified ballistic characteristics, such as speed of response and damping.

The standardized VU meter employs a full-wave, dry-disc rectifier. The dynamic characteristics of the meter movement are such that if a sinusoidal voltage in the frequency range concerned and of such amplitude as to give reference deflection (under steady-state conditions) is suddenly applied, the pointer will reach 99 percent of reference deflection in 0.3 second (within 10 percent). The pointer will then overshoot reference deflection by a minimum of 1 percent and a maximum of 1.5 percent.

Fig. 8-47 illustrates the conventional external circuitry involved with the use of the standard VU meter for monitoring transmission

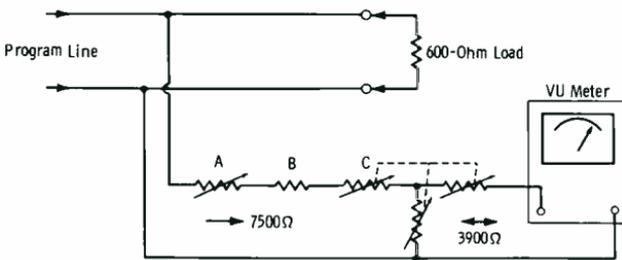


Fig. 8-47. External circuit of standard VU meter.

levels. In practice, the signal level is such that the range of the meter must be extended. The impedance of the meter itself varies with the voltage across the meter terminals and, therefore, must be isolated by a resistance network. The ballistic characteristics described are dependent on the meter source impedance, which must be 3900 ohms. The dB reading of the calibrated, variable multiplier (C in Fig. 8-47) plus the scale reading of the meter yield a measurement of the transmitted level. The complete network contains the following components:

- A. Zero adjuster, approximately 800 to 1000 ohms.
- B. Fixed resistor, approximately 3200 ohms, selected so that with A at mid-position,  $A + B = 3600$  ohms.
- C. Meter multiplier, "T" attenuator, 3900 ohms input and output.

The meter input impedance, as seen by the program line, is 7500 ohms, except when a 1-mW position is provided. This is a test position only and is not used in program monitoring.

The volume unit implies a complex wave, a waveform with a higher peak-to-rms ratio than a sine wave. When the VU meter is used to measure steady, single-frequency sine-wave signals, the

reading should be referred to as so many dBm. Volume units should never be used to indicate the level of a sine-wave signal.

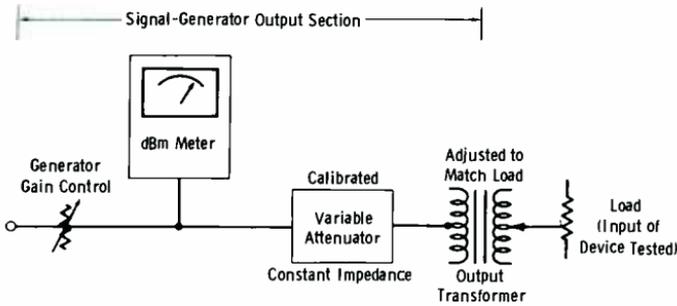
The volume unit has as its reference a steady-state condition, however. When the instrument is connected in the standard circuit configuration (Fig. 8-47), the maximum sensitivity is such that a +4-dBm, 1-kHz sine wave in the 600-ohm load causes the meter to deflect to the 0 VU point. Most program lines are fed with a +8-dBm level. In this case, the external multiplier is set to +8, and the meter indicates 0 VU at +8 dBm.

The VU meter is, by definition, properly calibrated only when connected across 600 ohms. When it is connected across any impedance other than 600 ohms, the reading must be corrected by adding  $10 \log_{10} (600/Z)$ , where Z is the actual impedance in ohms. EIA and FCC specifications on noise and distortion measurement require that a meter with standard VU characteristics be used. In this case, the meter on the measuring equipment reads an actual 1 mW of sine-wave power in 600 ohms for zero reference.

Always remember the basic differences in terminology between program and test signals, which may be summarized as follows: *Reference volume* is that strength of program signals that causes 0 VU (or 100 percent) deflection under the conditions described. This definition is arbitrary because the complex nature of the waveform makes a definition in fundamental terms impossible. *Reference level* is that steady-state condition in which there is 1 mW of a 1000-Hz sine wave in a 600-ohm impedance across which the meter indicates 0 reference level. This should be termed 0 dBm. However, when the meter is connected per standard practice according to Fig. 8-47, the maximum sensitivity is +4 dBm for 0-VU deflection. The actual level is the setting of the multiplier in dB plus the meter reading. Some multiplier arrangements allow a 1-mW position for test purposes only; this position taps down on the attenuator for a lower multiplier resistance.

## 8-16. DECIBELS IN PRACTICE

The first consideration is feeding the output of the sine-wave signal generator to the input of the system or device being tested. (Actual techniques are covered later. Only the proper reading of decibels is of concern here.) Fig. 8-48 is a typical arrangement. The generator dBm meter is always loaded by the constant impedance of the variable attenuator and the primary of the output transformer. The actual output is the reading of the meter minus the setting of the calibrated attenuator. For example, to feed a microphone preamplifier, the generator gain might be adjusted to give a meter reading of +15 dBm and the attenuator set to 65 dB. The



**Fig. 8-48. Typical method of connecting signal-generator output to input of audio system on which measurements are being made.**

actual output is then  $15 - 65 = -50$  dBm, a typical value for the input circuit of a microphone preamplifier.

The generator output-transformer secondary is then adjusted (usually by means of a switch connected to transformer taps) to match the load, which is the input of the preamplifier. This is normally 600, 150, or 50 ohms. The actual dBm input to the device is independent of the value of the load. Remember that 0 dBm sets the reference level at 1 mW in any load.

The voltage across the load and the current through the load are dependent on the value of the load in ohms. For example, 0 dBm (1 mW) results in 0.774 volt across 600 ohms, 0.387 volt across 150 ohms, and 0.224 volt across 50 ohms (Fig. 8-46).

Provided the generator output is matched to the load impedance, no conversion in meter reading (in dBm) is necessary when feeding the device or system, regardless of the input impedance. This is simply a practical application of Ohm's law, but it has resulted in some confusion in practice. The explanation of why the power remains the same regardless of impedance should be reviewed, as follows:

When the impedance is reduced from 600 to 150 ohms, the ratio is 4/1. Since the turns ratio of the output transformer is the square root of the impedance ratio:

$$\text{Turns Ratio} = \sqrt{600/150} = \sqrt{4} = 2 \text{ to } 1$$

Since the voltage developed is in direct proportion to the turns ratio, just one-half the voltage is developed across 150 ohms as for 600 ohms at any reference level. However, calculation shows that the same power exists in the load for either case. Therefore the level indicated on the generator dBm meter (minus the attenuator setting) is the actual level at the system input, provided the impedances are matched.

When the load does not match the output impedance of the generator, it is convenient to measure the voltage gain and then convert to decibels. The generator is adjusted to the closest match available, and the following calculation performed:

$$E_L = E_{OC} \frac{R_L}{R_L + R_O}$$

where,

$E_L$  is the voltage across the external load,

$E_{OC}$  is the open-circuit voltage (twice the voltage that would appear across a load equal to  $R_O$ ),

$R_L$  is the external resistance load,

$R_O$  is the output impedance of the generator.

For example, assume the external load is 250 ohms, the generator output impedance is set at 150 ohms, and the generator output is  $-10$  dBm. From Fig. 8-46, the voltage across a 150-ohm load would be 0.125 volt, so  $E_{OC}$  in this example is 0.25. Substituting in the formula gives:

$$\begin{aligned} E_L &= 0.25 \frac{250}{250 + 150} \\ &= 0.25 \frac{250}{400} = 0.156 \text{ volt} \end{aligned}$$

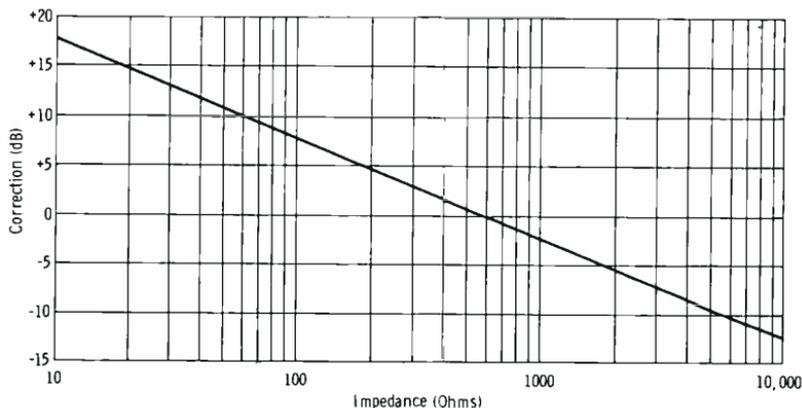
Note that  $-10$  dBm represents a power of 0.1 mW (Fig. 8-45). The power ( $E^2/R$ ) in the 250-ohm load is:

$$\frac{0.156^2}{250} = \frac{0.0243}{250} = 0.097 \text{ mW}$$

This is less than 0.5 dB from the  $-10$ -dBm reference of 0.1 mW (Fig. 8-45), an error that normally may be disregarded. Unless the mismatch is 2 to 1 or greater, any correction factor usually can be ignored, except for the most precise measurements.

As mentioned previously, noise and distortion measurements are made with a meter with standard VU characteristics to meet EIA and FCC specifications. This VU meter (or any VU meter) is properly calibrated only when connected across 600 ohms. For example, assume it is necessary to measure the gain of an amplifier with 150 ohms input and output impedance. Further assume that the gain of the amplifier is 40 dB. If the input is to be  $-20$  dBm, the output should be  $+20$  dBm. As described, the generator should feed  $-20$  dBm into the amplifier, and no conversion factor is involved. However, the output must be measured with a meter calibrated in dBm for 600 ohms. The impedance ratio between 600 and 150 ohms is 4 to 1, which, for a given power, corresponds to a voltage ratio of

2 to 1. A 2-to-1 voltage ratio is equal to 6 dB (Fig. 8-46 or dB Table in Appendix A). Therefore, the meter reading is 6 dB low across 150 ohms, and the correction factor is +6 dB. Then the meter at the output of the amplifier should read  $20 - 6$ , or +14 dBm. The actual output is then  $14 + 6 = 20$  dBm. Fig. 8-49 gives the correction factor for impedances from 10 ohms to 10,000 ohms, for the reference 0 dBm = 1 mW in 600 ohms.



**Fig. 8-49. Correction factor for VU meter connected across various impedances.**

Note that with the usual broadcast-type signal generator and measuring equipment no correction factor is involved at the system input. But if the measurement is made across other than 600 ohms at the output, the proper correction factor must be applied to the reading.

There are many applications in practice in which the standard VU meter is used across impedances other than 600 ohms. It is important to understand the proper interpretation for maintenance and level setting. For example, it may be desirable to monitor all inputs to a crossbar switcher where all the inputs are 150 ohms.

Assume the proper level at this point is +10 dBm. If the standard minimum insertion loss multiplier is used, the level across 600 ohms would be +4 dBm for 0-VU deflection. However, the correction factor for 150 ohms is +6 dB. Therefore, 0-VU deflection now indicates 0 dBm +4 +6 dB, or +10 dBm in 150 ohms.

Now further assume that an external test dB meter which does not employ the program-line bridging network is used, and the reference level of 1 mW in 600 ohms is stated on the scale of the meter. In this case, the 150-ohm crossbar input level should indicate +4 dBm (0 dBm with the external multiplier set on +4). Then with the +6-dB correction factor, the actual level is +10 dBm.

The preceding information is important to the installation engineer and to the maintenance department. Once the VU meters are installed in a system, the operating engineer is not interested in absolute levels; he needs only to see that the program level is maintained at the zero reference level on peaks.

Always bear in mind that when a VU meter peaks at 0 VU or 100 percent on program material, actual instantaneous peaks will occur at well over 1 mW in 600 ohms. The peak factor of the average program wave is generally taken as 10 dB over the peak-to-rms value of a sine wave. For this reason, a unit or system is sometimes tested and measured with a sine-wave power of 10 dB above the program operating reference. The Bell Telephone test board commonly feeds tones at 10 dB over the program operating level when measurements for distortion and cross talk are being made. This is important to the operator who may be monitoring the incoming network line for purposes of setting level on network circuits. When there is any doubt, the local test board concerned should be contacted to ascertain the level being transmitted.

## 8-17. MICROPHONE OUTPUT IN dB

The output of a microphone may be expressed in terms of either voltage or power. Since the output is obviously dependent on the magnitude of excitation, the reference is normally made to either 1 or 10 dynes per square centimeter. (A level of 0.0002 dyne/cm<sup>2</sup> is considered to be the threshold of audibility.) All microphone output ratings are expressed at a single stated frequency.

When the output rating is given in terms of voltage, the reference is 1 volt (open circuit), usually at 1 dyne per square centimeter. The expression is abbreviated dBV, which indicates decibels with a reference of 1 volt as 0 dB. Thus, if a microphone specification sheet states the output as -60 dBV, this indicates that an open-circuit voltage of 60 dB below 1 volt is generated with a sound pressure of 1 dyne/cm<sup>2</sup>. When the microphone is connected to a matched load, the voltage is decreased to one-half the open-circuit voltage, or 6 dB less. The effective output is then -66 dBV. Note from the decibel table (Appendix A) that -60 dB is 0.001 of the reference voltage, or 1 millivolt (open circuit). An additional 6 dB cuts this value in half, giving an effective 0.5-mV signal.

Most broadcast-type microphones are rated in terms of power output (dBm) at a stated sound pressure. Typical ratings are from -50 to -65 dBm at a sound pressure of 10 dynes/cm<sup>2</sup>. Note that 10 dynes/cm<sup>2</sup> is 50,000 times the sound pressure at the threshold of hearing, 0.0002 dyne×cm<sup>2</sup>. A voltage ratio of 10,000 to 1 is 80 dB. The ratio of 50,000 to 10,000 is 5, which gives an additional 14 dB

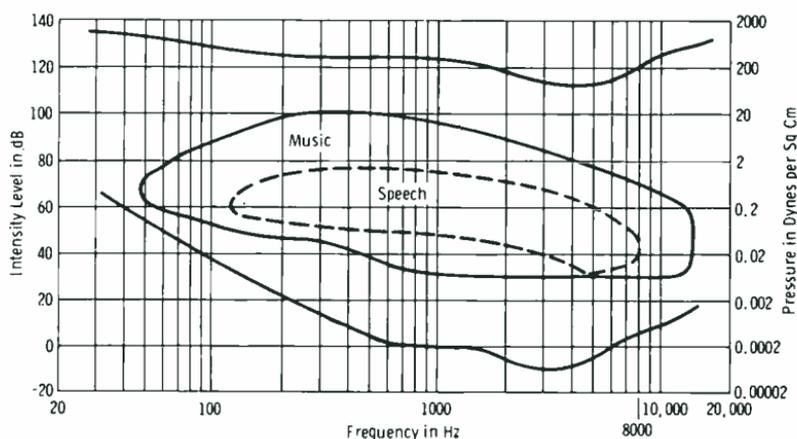


Fig. 8-50. Sound-level ranges.

(approximately) and a final level of 94 dB above threshold level. This is in the upper region of the average program sound pressure encountered in practice. (See Fig. 8-50, which is a graph of sound-level ranges in dynes/cm<sup>2</sup> and decibels, where 0.0002 dyne/cm<sup>2</sup> is 0 dB at 1000 Hz.)

The EIA (formerly RETMA) microphone system rating is also a power rating, but it is a ratio in dB, relative to 1 mW and 0.0002 dyne/cm<sup>2</sup>, of the power available from the microphone to the square of the undisturbed sound field pressure in a plane progressive wave at the microphone position. (This simply specifies the axis of the microphone relative to the sound front.)

The EIA system rating is given by:

$$G_M = \left( 20 \log_{10} \frac{E}{P} - 10 \log_{10} R_{MR} \right) - 50 \text{ dB}$$

where,

- $G_M$  is the microphone system rating (sensitivity) in dBm,
- $E$  is the open-circuit voltage generated by the microphone,
- $P$  is the sound pressure in dynes per square centimeter,
- $R_{MR}$  is the microphone rating impedance.

Microphone rating impedances ( $R_{MR}$ ) for broadcast-type units are given in Table 8-4.

Table 8-4. Microphone Impedance Ratings

Nominal Impedance	Rating Impedance ( $R_{MR}$ )
19-75 ohms	38 ohms
75-300 ohms	150 ohms
300-1200 ohms	600 ohms

Note that you can readily convert from the EIA microphone system rating to the effective output rating in dBm. It is only necessary to consider the difference in acoustical pressure between 0.0002 dyne/cm<sup>2</sup> and 1 or 10 dynes/cm<sup>2</sup>. The ratio of 1 to 0.0002 is 5000 to 1, or 74 dB. The ratio of 10 to 0.0002 is 50,000 to 1, or 94 dB (as discussed previously). Thus, to convert from a  $G_M$  rating to effective output, for 1 dyne/cm<sup>2</sup> add 74 dBm and for 10 dynes/cm<sup>2</sup> add 94 dBm. For example, if a certain microphone has a  $G_M$  rating of -150 dBm, the effective output level at 1 dyne/cm<sup>2</sup> is  $-150 + 74 = -76$  dBm. At 10 dynes/cm<sup>2</sup> the effective output level is  $-150 + 94 = -56$  dBm.

A natural question that may occur at this point is why the voltage ratio is used in expressing ratios in terms of decibels and then also used in adding to a power level in dBm. Actually, when a microphone is connected to an unloaded input transformer, its output cannot be expressed in power since no appreciable power is delivered. The effective output level of the microphone is simply that level which, when added to the amplifier power gain in dB, gives the correct output level from the amplifier in dBm.

The effective output level of a microphone connected to a matching impedance is given by:

$$P_0 = 1000 \frac{E_G^2}{4R_M}$$

where,

$P_0$  is the output level in milliwatts,

$E_G$  is the open-circuit output in volts,

$R_M$  is the nominal microphone impedance in ohms.

The power in milliwatts can then be converted into dBm.

## 8-18. AUDIO MAINTENANCE

Part 3 of this chapter has covered the basic theory and application of decibels and volume units, primarily to enable the reader to handle properly the aural proof of performance as discussed in Chapter 11. The techniques of microphone phasing and the actual maintenance of microphones, turntables, audio amplifiers, reel and cartridge tape machines, etc., are fully covered in a companion volume of this series, *AM-FM Broadcasting: Equipment, Operations, and Maintenance*. The reader is referred to that volume for troubleshooting techniques for all aural equipment.

### EXERCISES

- Q8-1. Give the basic definition of (A) linear distortion, (B) nonlinear distortion.

- Q8-2. How can you determine the effect of nonlinear distortion on a linear distortion?
- Q8-3. If the amplitude of the 3.58-MHz burst package is 50 percent of the amplitude of the 500-kHz burst package (peak-to-peak values) in the multiburst signal, what is the loss in dB at 3.58 MHz?
- Q8-4. For the multiburst signal as normally proportioned for AT&T or studio checks, what is the peak-to-peak amplitude of the multiburst cycles in IEEE units when the white flag is 100 IEEE units?
- Q8-5. When the K-factor graticule is used with the Tektronix Type 529 waveform monitor, what is the proper time base to use for a sine-squared pulse of (A)  $0.250\text{-}\mu\text{s}$  h.a.d., or (B)  $0.125\text{-}\mu\text{s}$  h.a.d.
- Q8-6. If the ringing period of Fig. 8-23 measures  $0.15\ \mu\text{s}$ , at what frequency is the measured gain dip?
- Q8-7. What determines the *amplitude* of the ringing in Q8-6?
- Q8-8. Assume the condition of Q8-6 for Fig. 8-23. Will a pulse of  $0.0625\text{-}\mu\text{s}$  h.a.d. show a different ringing period than a pulse of  $0.125\text{-}\mu\text{s}$  h.a.d.?
- Q8-9. What is the difference between smearing and streaking?
- Q8-10. Can differential gain at 3.58 MHz occur without luminance distortion?
- Q8-11. In stating power levels in decibels, must impedance values be stated?
- Q8-12. In relating voltage or current to decibels, must impedance values be stated?
- Q8-13. You have 0 dBm in 600 ohms. What is the rms signal voltage?
- Q8-14. You have 0 dBm in 150 ohms. What is the rms signal voltage?

## Microwave Systems

Television relay microwave systems operate in the 2000-MHz (1990–2110 MHz), 7000-MHz (6875–7125 MHz), and 13,000-MHz (12,700–13,250 MHz) bands. These are normally referred to as the 2-GHz, 7-GHz, and 13-GHz bands, respectively (GHz is the abbreviation for gigahertz, or thousands of megahertz). The 7-GHz region is the most popular band of frequencies for studio-to-transmitter links (stl's). The 13-GHz region is most commonly used for miniature (horn-type) microwave units that link cameras to mobile trucks, and the 2-GHz band is commonly used for relay from the mobile truck to the studio.

This chapter is concerned with the proper testing and maintenance of microwave systems, primarily as applied to permanent installations such as stl's. It must be assumed in this treatment that the proper planning and initial installation have been completed to the satisfaction of the station and that initial proof of performance confirms standards of fading margins in signal-to-noise ratio measurements. The complex fields of site selection and path surveys are beyond the scope of this treatment. The manufacturer of the particular equipment used normally furnishes either the complete service or the data necessary to plan the initial installation.

NOTE: Site selection and mobile applications are covered in a companion volume in this series, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals*.

### 9-1. VIDEO MEASUREMENTS

Measurements on a microwave system that has been properly installed are affected directly by transmitter and receiver tuning; therefore, this subject is of initial importance. It is important that

the reader be familiar with basic microwave circuitry as covered in the reference cited in the note above.

1. Set the wavemeter in the transmitter head on the exact operating frequency desired, noting any calibration-correction data the manufacturer may have supplied for the frequency setting.
2. Apply 60-Hz sine-wave modulation. This is normally provided by a test position on the video selector switch at the control unit.

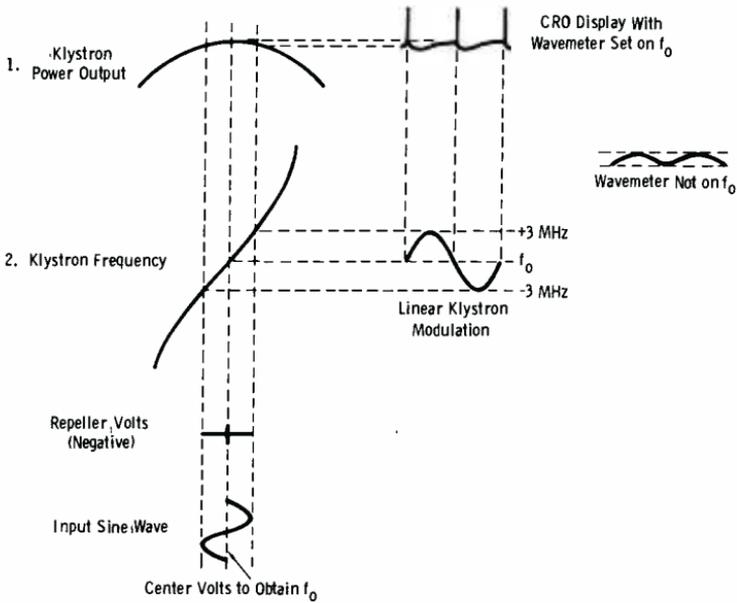
The wavemeter feeds a meter and/or test point for the purpose of relative power indication and frequency adjustment. In the Raytheon KTR-1000K/R system, the wavemeter absorbs part of the signal fed to the frequency-detection crystal, thereby giving an indication of the klystron frequency. The frequency notch is indicated on the control-unit tuning meter and rf-head test meter so that system tuning can be monitored readily from either unit.

The RCA TVM-1 system monitors crystal current on a meter and employs additional amplification for a test point to which a cro is connected for visual observation of the notches. The RCA TVM-6 is essentially the same.

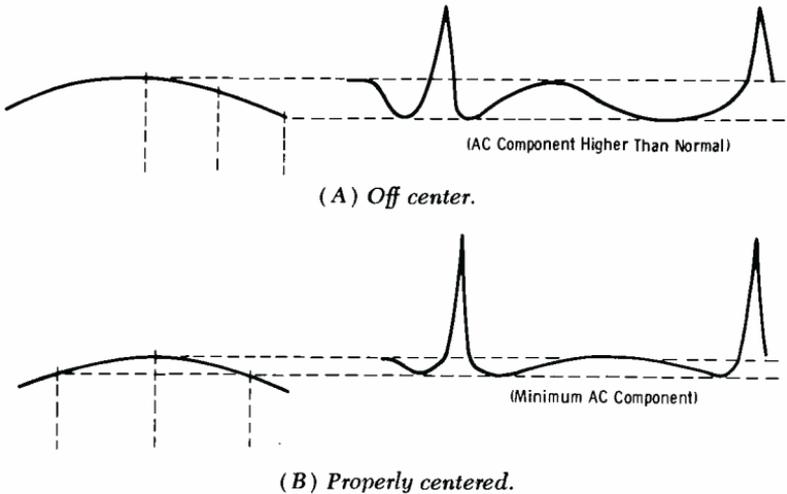
3. In Fig. 9-1, the applied sine wave is of such amplitude as to cause a 6-MHz deviation of frequency ( $\pm 3$  MHz from the operating frequency,  $f_0$ ). The repeller voltage and the mechanical screw of the klystron are varied alternately to obtain maximum crystal current and minimum ac component, as observed on the external cro. Note from curve 1 in Fig. 9-1 that the power output varies at twice the rate of the modulating voltage, which results in a 120-Hz pattern. (The wavemeter absorbs power each time the frequency swings through  $f_0$ ; thus the notches occur at a 120-Hz rate for a 60-Hz modulating voltage.) The actual power deviation is quite low for normal modulation level and could not be observed except for the gain of the amplifier and the oscilloscope.

NOTE: The RCA TVM-6 employs an 8-MHz deviation.

The most sensitive indication of klystron centering is the ac component. Fig. 9-2A illustrates how the ac component is increased with the klystron off center. When the klystron is not properly centered, either white compression or sync compression will occur, particularly at high modulation levels. When modern klystrons are properly centered (Fig. 9-2B) (maximum crystal current and minimum ac component at the frequency test jack), no compression will occur at levels up to almost twice the normal modulation level. The limiting factor usually is the modulator amplifier.



**Fig. 9-1. Correct tuning of a transmitter klystron.**



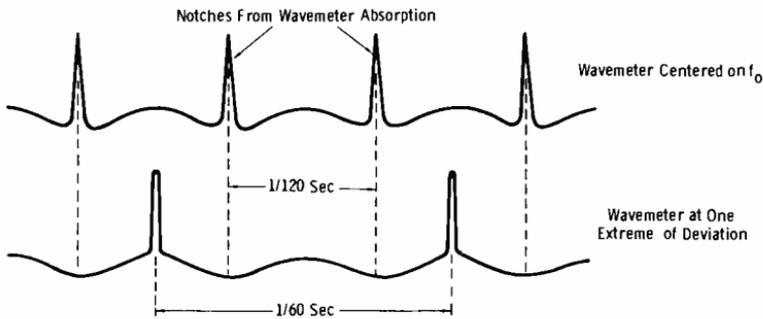
**Fig. 9-2. Waveforms indicating klystron centering.**

The electrical and mechanical tuning of the receiver klystron should be adjusted for maximum receiver crystal current and maximum signal level or agc voltage developed. A very sensitive indication of proper receiver tuning is available when a sound subcarrier

is employed. This subcarrier causes a certain degree of sync buzz in the sound if the receiver is detuned. With normal video but no sound applied, observe the audio output of the sound demodulator with an oscilloscope set for vertical-rate display. This permits precise tuning of the receiver klystron for minimum cross talk into the audio channel. With high scope gain, the 60-Hz pulse trains are clearly visible if the receiver is detuned. The scope may also be connected to the output of the audio noise and distortion meter for higher gain.

Proper frequency deviation under modulation must also be obtained before video measurements are made. The normal procedure using a wavemeter is as follows:

1. With the test 60-Hz signal applied, proper tuning is indicated by the 120-Hz signal shown in the upper trace of Fig. 9-3.
2. Unlock the calibrated dial of the wavemeter, and rotate the dial until the notches just merge into a 60-Hz signal (lower trace of Fig. 9-3). Read the frequency at which this occurs.
3. Now rotate the dial in the opposite direction until the notches again just merge. Read this frequency. If the normal 100-percent modulation is to be 6 MHz (peak to peak), the deviation on each side of the center frequency should be 3 MHz.



**Fig. 9-3. Waveforms indicating deviation from center frequency.**

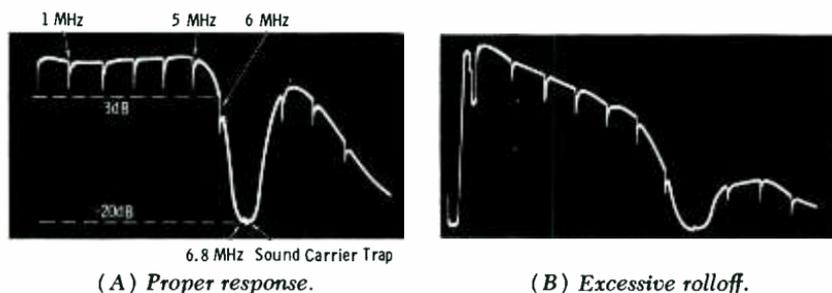
4. If the total peak-to-peak swing is not correct, adjust the amplitude of the 60-Hz test signal until the required deviation is obtained. Measure the peak-to-peak value of this sine wave at the klystron repeller terminal with a low-capacitance scope probe. For example, the type 220 klystron will require approximately 15 volts (peak-to-peak) of signal (modulation sensitivity about 400 kHz per volt) to obtain a peak-to-peak deviation of 6 MHz.
5. Now bypass the predistortion network (if used), and feed a normal video level to the modulator. Adjust the signal-modula-

tion gain to obtain the peak-to-peak value required, as determined in step 4. Reinsertion of the predistortion network will reduce the swing at 60 Hz to 2.4 MHz if the usual 8-dB network is employed. (A loss of 8 dB means a reduction ratio of 0.4.)

### Microwave Frequency Response

Initially, the transmitter monitor amplifier should be measured for frequency response and amplitude linearity so that overall measurements can be conveniently related to individual characteristics of the transmitter and receiver. When the engineer is certain of the monitor characteristics, the transmitter modulator and klystron response will be revealed by the normal test signals applied to the system input.

The video sweep is most convenient to use for wideband response measurement. (NOTE: If the system employs line-to-line clampers, keyed sweep should be used, as described previously.) Fig. 9-4A shows the proper overall response when a sound diplexer is used and the subcarrier frequency is 6.8 MHz. The sound notch should be centered on the subcarrier frequency with a width (normally) of about 300 kHz. This adjustment is usually provided by trimmers in the passive mixing network for video and audio at the transmitter.



**Fig. 9-4. Detected video sweep of stl system.**

Note from Fig. 9-4A that for a 6.8-MHz sound subcarrier, video rolloff starts at approximately 5 MHz, is down 3 dB at 6 MHz, and is down 20 dB at 6.8 MHz. Although this rolloff is not as sharp as that employed at the visual transmitter, it does result in slight ringing at the trailing edge of a sharp vertical transition in the picture. This is normal in microwave systems employing diplexed sound.

Fig. 9-4B shows the detected sweep on a system with excessive rolloff. If this occurs on the overall system measurement at the receiver video output, and the transmitter monitor output indicates

the waveform of Fig. 9-4A, the rolloff is actually at the receiver. In stl installations, two units at both the transmitting and receiving positions are normally employed. During the test period, a common receiver should be used to measure the relative response of the main and standby transmitter units (when operated on the same frequency). Thus if the response through (for example) receiver 1 and transmitter 1 is as shown in Fig. 9-4A, but the response through receiver 1 and transmitter 2 is as shown in Fig. 9-4B, then transmitter 2 needs servicing. The same response should be evident from the transmitter-2 monitor video.

Poor frequency response is usually the result of sagging transconductance in modulator tubes or transistors, faulty coaxial lines, or bad terminations. Complete alignment of the modulator circuits may be necessary at intervals of a year or two, for tube-type circuitry.

Similarly, microwave-receiver if circuitry seldom needs complete alignment, since modern units employ if bandwidths of 18 to 30 MHz, which is far more than the normal 4- to 8-MHz deviation of the carrier. Therefore, poor frequency response is almost always the result of faulty tubes or plate loads, as previously discussed. Solid-state systems (such as the RCA TVM-6) are relatively free of service requirements over long periods of time.

### **Amplitude and Phase Linearity**

Amplitude and phase linearity is checked in the conventional manner with the stairstep signal, as described in Chapter 8. Always check the maximum allowable peak-to-peak value of signal input at which compression just starts. This provides a warning flag as to the need for further checks. For example, if the normal peak-to-peak video is 1 volt, raise the input level of the stairstep until compression just starts. Record this value, and use it for future reference.

For installations which must meet color standards, differential gain and phase at 3.58 MHz must also be measured. (The techniques are the same as those in Chapter 8.) Always adjust the transmitter monitor cavity for zero differential gain, as shown in Fig. 9-5. This will normally result in minimum differential phase. Remember that for overall differential-phase measurements, a burst-controlled oscillator unit must be employed at the receiver unless units are measured back-to-back on the bench.

Causes of poor amplitude and phase linearity are the same as those outlined in Chapter 8. In rare cases, it is possible that the klystron is at fault here, although the normal indication of a poor klystron is lowered signal strength and poor signal-to-noise ratio. If replacement is indicated, the "lighthouse" unit must be replaced with an entire assembly.



**Fig. 9-5. Display permitting measurement of differential gain.**

### Measuring Video Signal-to-Noise Ratio

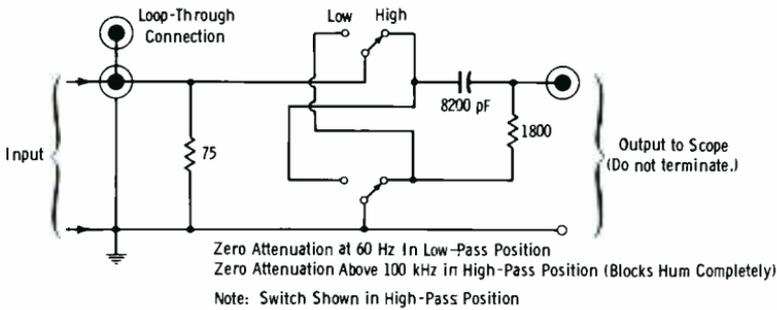
The video signal-to-noise ratio is measured in two parts: (1) video signal to random noise, and (2) video signal to hum level. The measurement of the ratio of video to random noise is simply termed signal-to-noise measurement, and it is understood that this measurement must not include hum level. Video-to-hum content is a separate measurement.

In addition, the signal-to-noise ratio may be measured in terms of peak-to-peak video and peak-to-peak noise, or peak-to-peak video and rms noise. The former is most often used in practice, since it requires a wideband (10 MHz) scope, which is readily available at the station. The noise may be specified in terms of rms when a wideband (4 MHz) vtmv, calibrated in rms values of a sine wave, is available. If desired, a 20-dB conversion factor may be employed between the two methods of measurement. For example, if the p-p video to p-p noise measures 28 dB, the p-p video to rms noise is  $28 + 20 = 48$  dB.

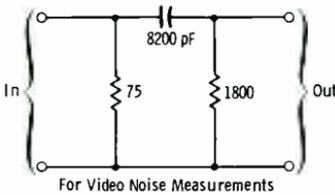
**NOTE:** The 20-dB conversion factor takes into account the conversion from peak-to-peak to rms, plus the fact that rms noise measurement should be limited to a 4-MHz bandwidth rather than the 10-MHz or more bandwidth of the scope amplifier.

EIA standards for stl's call for a minimum ratio of p-p video to rms noise of 58 dB. Thus to meet this standard, the ratio of p-p video to p-p noise should be at least 38 dB. In any event, the measurement should indicate a signal-to-noise ratio of at least the minimum required for the conditions outlined in Section 9-3. The random noise content of the signal is usually a reliable indication of the path effectiveness.

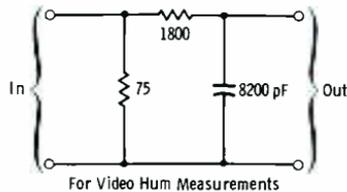
To measure both signal-to-noise and signal-to-hum ratios, a filter similar to that illustrated in Fig. 9-6 should be constructed and used. The input circuit provides a termination for the receiver video



(A) Filter circuit.



(B) Equivalent circuit for high position.

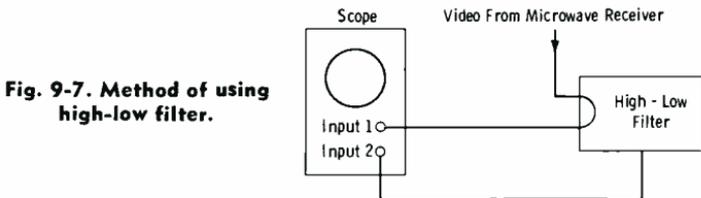


(C) Equivalent circuit for low position.

**Fig. 9-6. Video filter for stl signal-to-noise measurements.**

output. The filter output should go directly to the scope vertical-amplifier input without termination.

A convenient method of hookup is shown in Fig. 9-7. The reference video level fed from the microwave transmitter is measured by scope input 1. The filter output at input 2 is then available at the flick of a switch. This assumes a dual input or switchable input as available on the Tektronix Type 524 scope. The filter provides zero attenuation at 60 Hz in the low position, and zero attenuation above 100 kHz in the high position. No insertion loss is involved, and



**Fig. 9-7. Method of using high-low filter.**

the readings are direct. Since the filter is frequency-selective, the video waveform is distorted. Measure only the characteristic (hum or noise) desired, which is measured with no modulation according to the following procedure.

1. If sound diplexing is used, turn off the sound modulator and demodulator.
2. Leaving any predistortion and restoration networks in place, feed the 60-Hz test sine wave through the transmitter at the normal deviation (100-percent modulation).
3. Set the receiver video output level to the station standard (normally 1 volt peak to peak). If the hookup of Fig. 9-7 is used, the signal should be applied at scope input 1. Use vertical-rate sweep. All gains have now been set at reference level.
4. Remove the test signal at the transmitter, and remove the coaxial line feeding the video input. Substitute a termination.
5. At the receiving position, observe scope input 2 with the filter in the high-pass position. Increase the gain of the scope to maximum and read the peak-to-peak voltage of the random noise. The resulting voltage ratio is converted to dB by using a dB table or by computing 20 times the logarithm of the ratio.

NOTE: Some engineers prefer to use the scope internal 60-Hz sweep with the sweep attenuator adjusted to collapse the trace to about a quarter-inch width. This gives a more readily measured noise indication at very low noise levels.

The measurement of the ratio of video signal to hum is made in the same way, except that the filter is placed in the low-pass position for the hum-content trace. Hum content should be at least 40 dB down, or at least the amount listed in the particular manufacturer's specifications. Hum is minimized by reversing one power lead at a time, by adequate shielding, and by good power-supply filtering and regulation. Most modern modulators and klystrons have dc filament voltage supplied for further suppression of hum. Check these supplies occasionally for ripple content.

## 9-2. DIPLEXED-SOUND MEASUREMENTS

It is necessary that the technician be familiar with the practical use of the audio noise and distortion meter. The basics of the audio signal generator were covered in Chapter 8. The audio noise and distortion meter and its use are covered in Chapter 11.

The microwave sound-diplexing equipment employs the same pre-emphasis/de-emphasis curve as the aural transmitter. This calls for a modification of techniques when the stl is involved, either in an overall measurement from the studio microphone input to the transmitter output or in measuring the stl alone.

In this section we are concerned with measurements on the stl only. Overall measurements through the main aural transmitter are covered in Chapter 11.

### Frequency Response

A suggested procedure for frequency-response measurements is as follows:

1. Feed a 1-kHz tone to the modulator at the specified program input level. Adjust the modulator gain to obtain 100-percent modulation (0 VU on the meter).
2. At the receiver, adjust the sound-demodulator gain for reference output (0 VU). Calibrate the measuring equipment, and run response measurements at 50, 100, and 400 Hz. The measuring equipment is connected across the proper termination resistance (usually 600 ohms) required for the demodulator output.
3. Reduce the audio input to the modulator 20 dB at 1 kHz. This is the new input reference level. Increase the gain of the measuring equipment at the demodulator output by 20 dB. Using the new reference, run response measurements at the necessary spot frequencies up to 15 kHz.

Since the overall measurement includes de-emphasis, the response curve should be relatively flat. In any event, the response must be adequate to meet the requirements for FCC proof-of-performance runs on the overall system.

### Distortion Measurement

To measure distortion, proceed as follows:

1. With a 1-kHz tone applied to the modulator and the gain adjusted for 100-percent modulation, apply termination to the demodulator output and calibrate the noise-distortion meter connected across this termination.
2. Make the distortion measurement at 1 kHz. Then reduce the tone-generator frequency to 50 Hz and *use the same generator output level as for the 1-kHz tone.* (The modulation meter will fall considerably because of the pre-emphasis curve.)
3. At the receiver, adjust (if necessary) the gain to hold 0 VU output. Make the distortion measurement at this frequency. Repeat Steps 2 and 3 at all spot frequencies back up to 1 kHz (hold the demodulator output constant; hold the signal-generator output constant).
4. At all frequencies above 1 kHz, *hold the modulator VU meter at 0 VU by reducing the output level of the signal generator at each new frequency.* At each new frequency, the gain of the distortion meter will need to be increased to compensate for the de-emphasis curve. *Do not touch the demodulator gain or modulator gain.* Thus it is necessary to calibrate the noise-dis-

tortion meter at each new frequency. Make distortion measurements at the required spot frequencies up to 15 kHz.

Distortion in tube-type equipment is normally produced by the reactance tube and the tubes in the modulator used for audio amplification. In the demodulator, distortion is caused by the audio amplifier tubes following the discriminator. This assumes that the stl transmitter and receiver circuits are properly tuned, and that the diplexing tuned circuits are centered.

Distortion is excessive when the overall measurements through the studio equipment to the transmitter output indicate noncompliance with FCC proof-of-performance specifications (covered in Chapter 11), and a spot check of the stl alone indicates sound diplexing to be the limiting factor. When this occurs, it is best to remove the stl units completely for bench checkout by back-to-back hookup through an attenuating coupler. If it is not considered wise to do this with the standby equipment during broadcast hours, the entire alignment normally can be completed in about four hours during nonbroadcast time. The manufacturer's instructions regarding adjustments that affect distortion should be studied ahead of time and the proper preliminary planning done to minimize down time.

These adjustments normally include modulator and demodulator discriminator primary adjustments for minimum distortion, which also affect the discriminator secondary adjustment for balance (zero output at no modulation). When such adjustments are made, a frequency standard set at the subcarrier frequency should be available to ensure that the subcarrier traps hold adjustment, or to provide a visual method of readjusting the frequency or traps. Because of the wide variance of circuitry, it is necessary to follow exactly the manufacturer's instructions for the particular model of diplexing equipment involved. In general, the sound notch must be centered on the subcarrier frequency with a *flat-bottomed* width of around 300 kHz (Fig. 9-4). If the sweep is observed on a wideband display (undetected), the flatness of the notch is more evident.

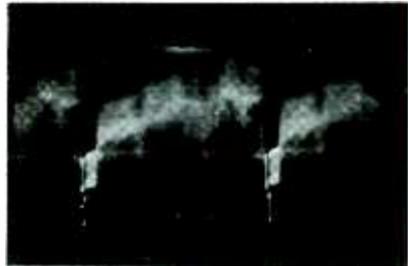
### **Aural Signal-to-Noise Ratio**

The following procedure for measuring aural signal-to-noise ratio may be used:

1. Feed a 1-kHz tone to the modulator at normal program level. Adjust the modulator gain control for 100-percent modulation (0 VU).
2. Adjust the demodulator gain for 0 VU output, with the proper termination applied and the noise meter connected across this termination. Calibrate the noise meter.

3. Remove the tone modulation and substitute a resistor equal to the input impedance (normally 150 to 600 ohms). Measure the residual noise.

The noise level of modern diplexing equipment should be between  $-70$  and  $-85$  dBm with the units back-to-back on the bench. In service, this level depends on the effectiveness of the path. To expect a sound signal-to-noise ratio of 70 dB, it is necessary to obtain at least a 25-dB video signal-to-noise ratio on a p-p video to p-p noise basis (Section 9-3).



**Fig. 9-8. Wideband cro display of video signal with diplexed sound.**

The rf level of the aural subcarrier added to the video should be sufficient to maintain an adequate signal-to-noise ratio, but not high enough to result in sound bars in the picture output. Fig. 9-8 illustrates the sound-subcarrier fill in the sync region. In practice, the subcarrier level is adjusted to approximately 50 percent of the sync level.

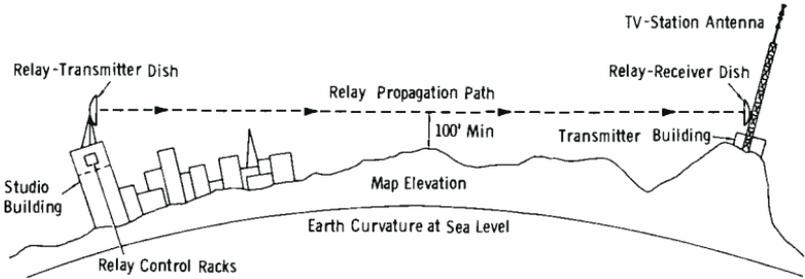
### 9-3. WHEN TO SUSPECT A FAULTY INSTALLATION (SYSTEM EVALUATION)

An engineer who may have inherited a faulty installation must have some means of determining such faults. The primary measurement of a satisfactory path is the signal-to-noise ratio consistently obtained from a series of tests in various weather conditions.

When unsatisfactory signal-to-noise measurements are obtained, the final criterion is a back-to-back measurement on the bench. Connect the transmitter head to the receiver head through a variable attenuator or fixed attenuator that gives an attenuation equal to that of the path. These waveguide couplers are normally supplied with the initial installation. If not, one may be purchased or rented.

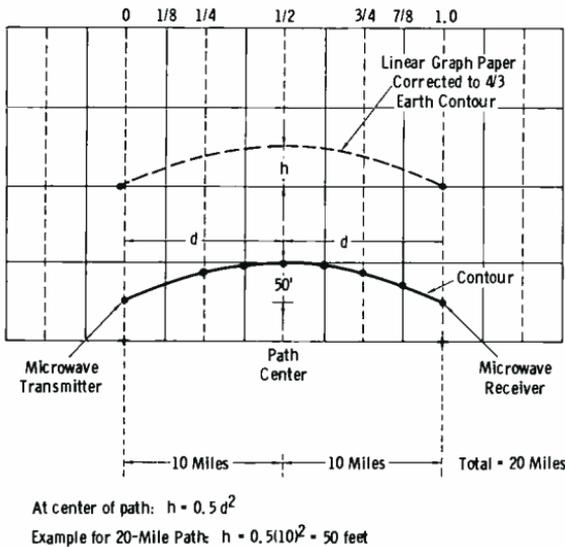
Provided the back-to-back measurement reveals nothing faulty in the units themselves, you may proceed with the notion that either antenna misalignment has occurred (as might happen with loose mountings) or the path is at fault. Antenna alignment should, of

course, be tried first. If nothing is gained, investigate the path profile (see Fig. 9-9 for an example). If a 100-foot clearance exists at the nearest point to the center of the beam, the path is not at fault, since this clearance provides an adequate safety margin.



**Fig. 9-9. Basic principle of stl.**

1. Plot a profile of the transmission path. Graph paper which presents the curvature of the earth on a radius  $4/3$  times its true value may be employed. For limited use, it is more convenient to use ordinary linear graph paper and the data of Fig. 9-10. Paper with ten squares to the inch is ideal for this purpose.
2. The path profile and obstructions on the path may be charted from topographic maps. The topographic map gives the height



**Fig. 9-10. Method of plotting profile of transmission path.**

**Table 9-1. Minimum Transmission-Path Clearance  
(in Feet) Above 4/3 Earth**

Path Length (Miles)	1/8 & 7/8 Distance	1/4 & 3/4 Distance	1/2 Distance
5	13	16	19
10	21	27	32
15	29	38	45
20	38	49	57
25	46	59	69
30	54	71	82

above sea level of the surface of the earth, to which are added the heights of major obstructions. Maps for specific areas east of the Mississippi River may be obtained from the U.S. Geological Survey, Washington, D.C. 20242. For maps of areas west of the Mississippi River, write to the U.S. Geological Survey, Denver, Colorado 80225.

3. The clearance from the tallest obstruction in the path should be at least that shown in Table 9-1.

The above technique will enable you to determine roughly if the initial path survey was adequate. When you are in doubt, obtain the services of a competent and experienced microwave organization or the manufacturer of the equipment used.

**Table 9-2. Approximate Gain (Parabolic Reflector Only)**

Dish	Gain (dB)		
	2000 MHz	7000 MHz	13,000 MHz
4 ft	25	37	42
6 ft	28	40	45

You can also use the antenna-system gains and free-space attenuation to determine approximately what signal-to-noise ratio should be expected. Table 9-2 shows the gain of a parabolic reflector as a function of size and operating frequency. Fig. 9-11 shows system gain for the designated dish and reflector sizes at 7000 MHz. (Similar information is obtainable from the manufacturer of a given microwave unit.) Table 9-3 is a tabulation of the free-space loss for the indicated microwave bands up to a 30-mile path length. This is the maximum distance normally employed for a single hop.

Assume the following data:

Power output of transmitter: 1 watt (Let this be 0 dBW.)

Frequency: 7000 MHz

Path length: 20 miles

Table 9-3. Approximate Free-Space Loss

Path Length (Miles)	Loss (dB)		
	2000 MHz	7000 MHz	13,000 MHz
5	117	127	133
10	123	133	140
15	126	136.5	143.5
20	129	139	146
25	131	141	147
30	132	142.5	148

Where: Free-space loss in dB =  $37 + 20 \log f + 20 \log D$ ;  $f$  = operating frequency in MHz;  
 $D$  = distance in miles.

Antenna-system gain at each end: 35 dB (This is a total gain of 70 dB.)

The net path loss is:

$$A = A' - G_t - G_r$$

where,

$A$  is the net path loss,

$A'$  is the free-space loss (from Table 9-3),

$G_t$  is the transmitter antenna-system gain (Table 9-2 or Fig. 9-11),

$G_r$  is the receiver antenna-system gain (Table 9-2 or Fig. 9-11).

For the example above

$$\begin{aligned} A &= 139 - (+35) - (+35) \\ &= 139 - 70 \\ &= 69 \text{ dB} \end{aligned}$$

The receiver power input is:

$$P_r = P_t - A$$

where,

$P_r$  is the receiver power input,

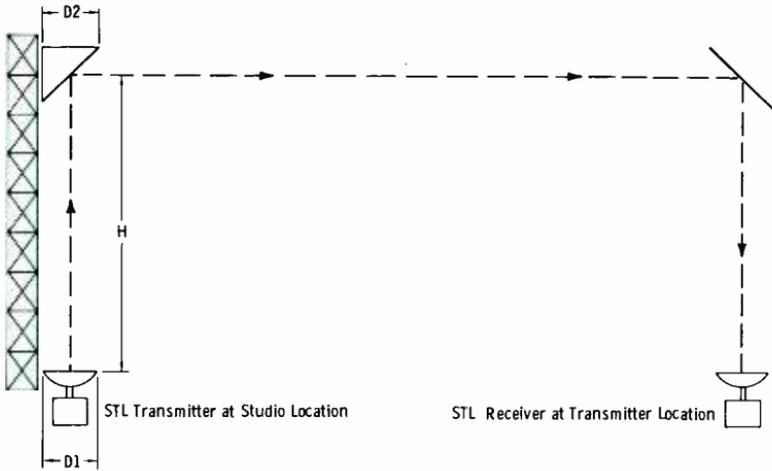
$P_t$  is the transmitter power output,

$A$  is net path attenuation.

This formula states that subtracting the net path loss (in dB) from the transmitter power output (for 1 watt this is expressed as 0 dBW, or 0 dB above 1 watt) gives the power input to the receiver. The path loss calculated above is 69 dB, so:

$$P_r = 0 - 69 = -69 \text{ dBW}$$

Fig. 9-12 is a graph of the expected signal-to-noise ratio versus power input to the receiver (dBW) for the RCA microwave-relay system. Note that for a power input of  $-69$  dBW, the video signal-



(A) Antenna arrangement.

H Dimension (Feet)	Gain (dB)		
	D2 = 4 ft	D2 = 6 ft	
80	35	39	D1 = 4-ft Parab
100	34	38.5	
120	33	38	
140	32	37.5	
160	31	37	
180	30	36.5	
200	29	36	
80	36	40.5	D1 = 6-ft Parab
100	35	40	
120	34	39.5	
140	33	39	
160	32	38.5	
180	31	38	
200	30	37.5	

(B) Approximate antenna-system gain (parabola and reflector), 7 GHz.

Fig. 9-11. Computation of gain when passive reflectors are used.

to-noise ratio should be 38 dB on a peak-to-peak video to peak-to-peak noise basis. This is approximately 58 dB on a peak-to-peak video to rms noise basis, as covered previously under techniques of measurement.

Note also from Fig. 9-12 that this receiver input power should result in better than a 72-dB signal-to-noise ratio for duplexed sound.

In evaluating microwave-relay performance, the following fundamentals serve as a basic guide:

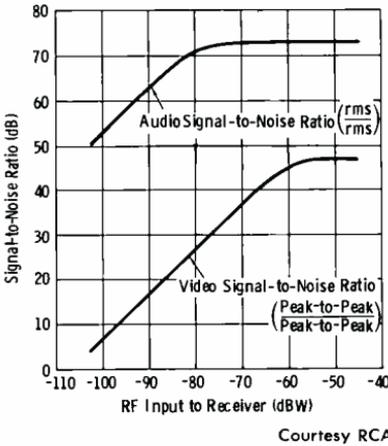


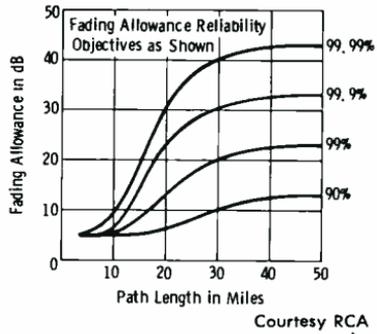
Fig. 9-12. Graph showing signal-to-noise ratio versus receiver input.

1. In general, noise will be visible in the picture when the signal-to-noise ratio deteriorates to less than 24 dB.
2. Since a microwave beam sometimes is bent and scattered by atmospheric conditions, not only adequate clearance, but also adequate fading margin must be provided. The picture becomes unusable at a signal-to-noise ratio of 8 dB. This 8-dB figure establishes a basis for computing the fading margin.
3. For example, if a 24-dB signal-to-noise ratio exists, the fading margin is only 16 dB ( $24 - 8 = 16$ ). From Fig. 9-13, a 16-dB fading margin indicates only about 98-percent reliability on a 25-mile path. This would mean an outage of about 117 hours in an average broadcast year, entirely unsuitable for applications of a continuous nature such as an stl. Note that for a reliability of 99.99 percent, the fading margin for a 25-mile path should be 37 dB. This requires a signal-to-noise ratio of  $37 + 8 = 45$  dB.
4. From Fig. 9-13, find the allowable fading margin for 99.99-percent reliability, especially if the service is stl. Assume the path length is 20 miles; this requires a fading margin of 31 dB. Then a minimum signal-to-noise ratio of  $31 + 8 = 39$  dB should be indicated by an average of measurements made under average weather conditions. Normally, measurements made daily or nightly over a period of a week or two give a reliable indication of the practical signal-to-noise ratio of a system.

## EXERCISES

- Q9-1. What determines the frequency of a klystron?
- Q9-2. What are the most usual indications of a deteriorating klystron?

**Fig. 9-13. Graph showing fading allowance.**



- Q9-3. What measurements are most critical in indicating proper tuning of the stl receiver?
- Q9-4. Is the video modulation for microwave systems normally am, or is it fm?
- Q9-5. Is the audio modulation for sound diplexing on an stl normally am, or is it fm?

## Transmitter Terminal Gear

Transmitter terminal gear includes all equipment at the transmitter location prior to the actual transmitter input, including test and monitoring facilities for both the aural and visual transmitters.

Figs. 10-1 and 10-2 show typical block diagrams of video and audio terminal gear, respectively. The drawings have been simplified to eliminate the jack fields often used at the inputs and outputs of various units for greater flexibility. These diagrams may be termed "typical," although considerable variation of program monitoring facilities exists from station to station. The use of either remote control or an automatic transmission system (ATS) is optional. (NOTE: At the time of this writing, ATS had not been approved for television stations, but approval was expected at an early date.)

### 10-1. VIDEO TERMINAL GEAR

In Fig. 10-1, the use of stl's is assumed. Some stations employ land facilities of the telephone company, and the stl receivers are replaced with Telco terminal equipment. Signal sources are stl 1 or 2, a slide chain, and various test-signal generators.

#### Video Air Path

The air and monitor switcher receives all signal-source outputs and provides two outputs: (1) a monitoring bus, and (2) the air bus. Any source may be selected for either bus at a given time.

The air-bus output is isolated by distribution amplifier (DA) 2, which feeds processing amplifiers 1 and 2. The processing amplifier (usually shortened to "proc amp") provides individual controls for sync, blanking, and video amplitudes so that the necessary amount of sync stretch may be applied to compensate for transmitter non-

linearity. In some cases, amplitude and even phase controls for the color burst are provided.

The output of either proc amp may be selected for the air path by the proc-amp switch; the output of this switch is fed to the color equalizing terminal gear. It is the purpose of this unit to predistort the signal so that the transmitter output is correct. Such correction is used not only to compensate for the transmitter characteristics, but also for certain "average receiver" characteristics as discussed below. It should be understood that the equalization mentioned for the color-equalization group is mostly that of *phase* as distinguished from amplitude provided in the video proc amp.

*Low-Pass Filter*—The low-pass filter provides rapid rolloff of video response above 4.2 MHz. The filter response is within 1 dB to 4.18 MHz, and falls to -20 dB at 4.75 MHz.

*Distribution Amplifier 3*—Distribution amplifier 3 provides precise termination for the low-pass filter, isolation of the low-pass filter from the low-frequency phase equalizer, sending-end termination for the low-frequency phase equalizer, and gain to offset losses in this control system.

*Low-Frequency Phase Equalizer*—The low-frequency phase equalizer has in its input circuit an equalizer to reduce the *amplitude* response at lower frequencies to result in a flat response out to 4 MHz. This is required to compensate for a slight deficiency in amplitude response of the low-pass filter and phase equalizers. The remaining circuitry in this unit compensates for delay distortion produced by the transmitter and vsb filter. NOTE: No compensation is provided at the transmitter for low-frequency delay distortion in receivers.

*Receiver Equalizer*—Compensation is provided for the high-frequency delay distortion that results from the sharp cutoff in the receiver from full response at 4 MHz to almost no response at 4.5 MHz (sound-carrier separation from picture carrier).

*Distribution Amplifier 4*—Distribution amplifier 4 serves the same purposes as DA 3, except that it is placed between the receiver equalizer and the variable equalizer.

*Variable Phase Equalizer*—The variable phase equalizer provides variable high-frequency phase compensation for the delay distortion in the transmitter.

*Distribution Amplifier 5*—Distribution amplifier 5 is the final DA for the selected video signal. One output goes to the monitoring system (transmitter input designation), and the other output goes to the transmitter video input.

### Video Monitor Path

Distribution amplifier 1 provides isolation for the monitor bus, and feeds the Mon Bus position of the master monitor switcher. Thus

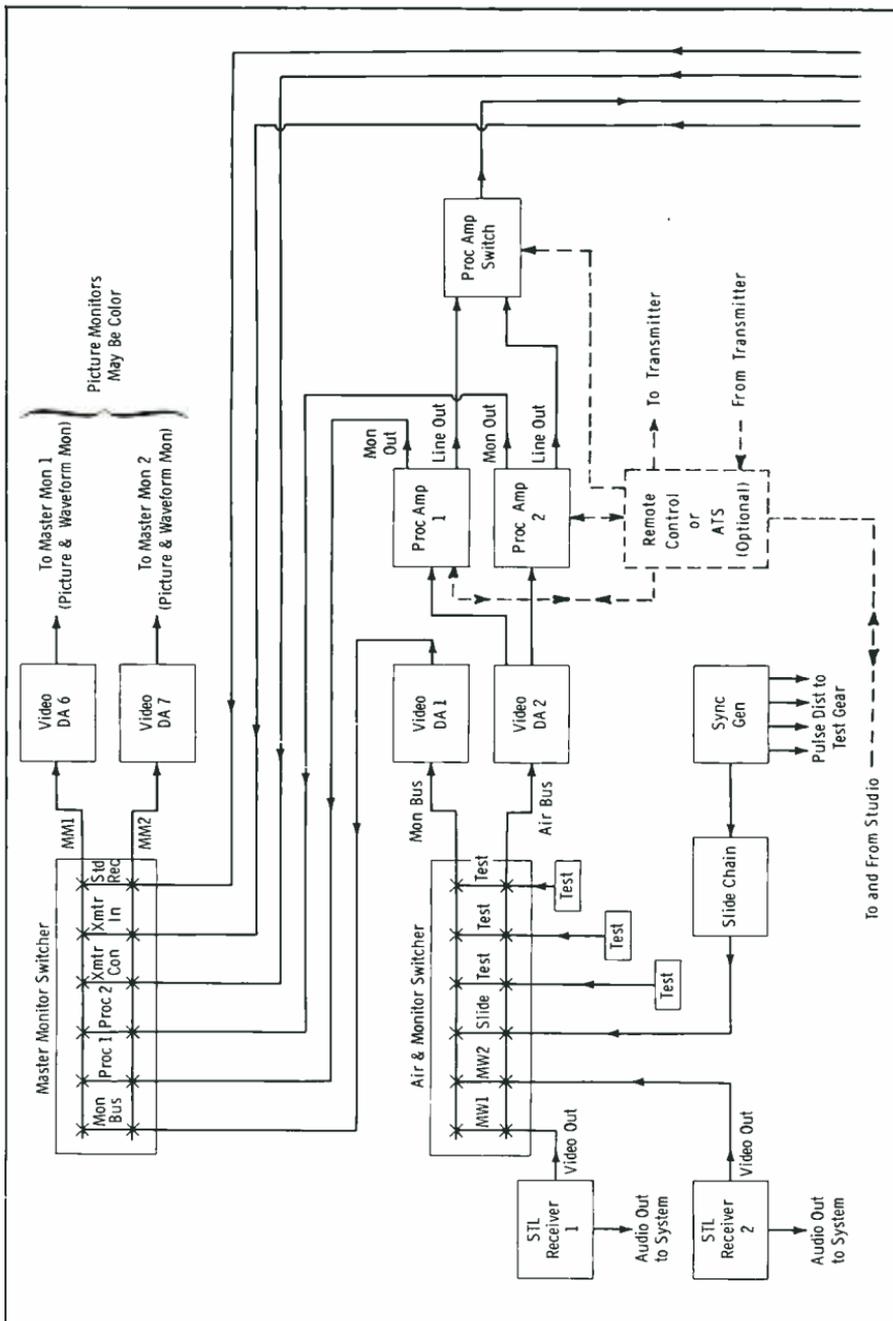
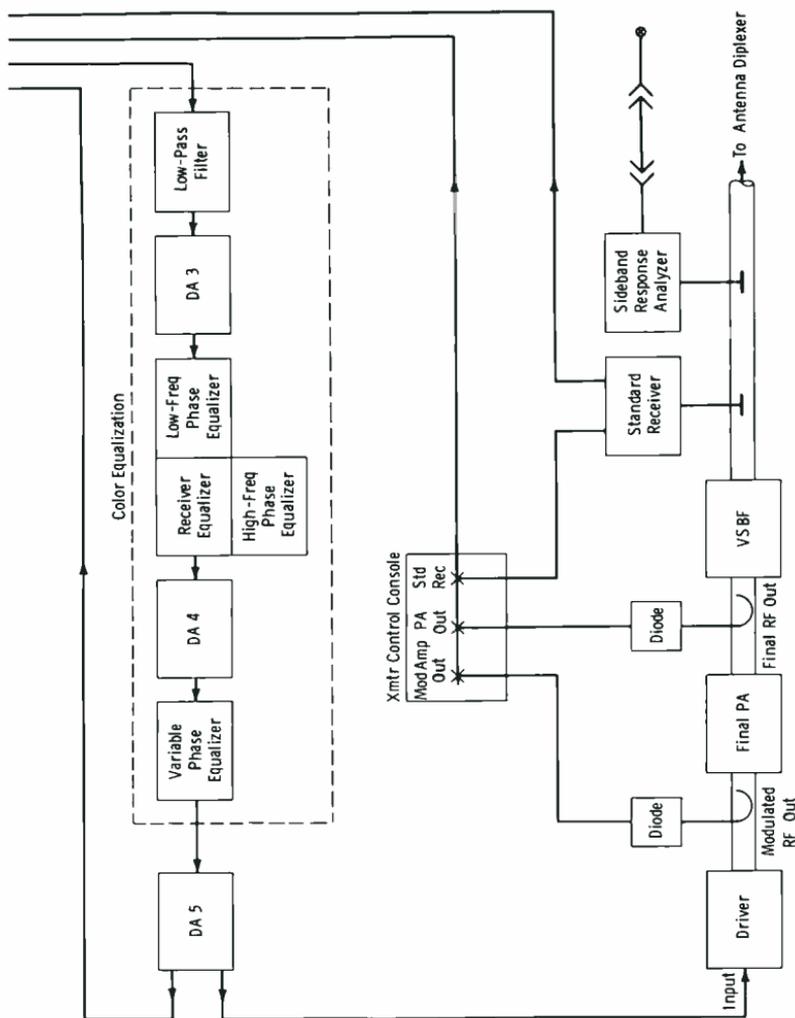


Fig. 10-1. Block diagram of video



transmitter terminal gear.

the monitor row of the air and monitor switcher preselects what is to be observed on a master monitor when its associated MON BUS button is depressed.

Another preselect position is provided at the input marked XMTR CONSOLE. This input receives the source selected from the monitor buttons on the transmitter control console.

With the monitoring arrangement shown, either master monitoring position may be displaying the video signal from almost any point between the stl output and the output of the vestigial-sideband filter. This permits level setting or any other form of maintenance

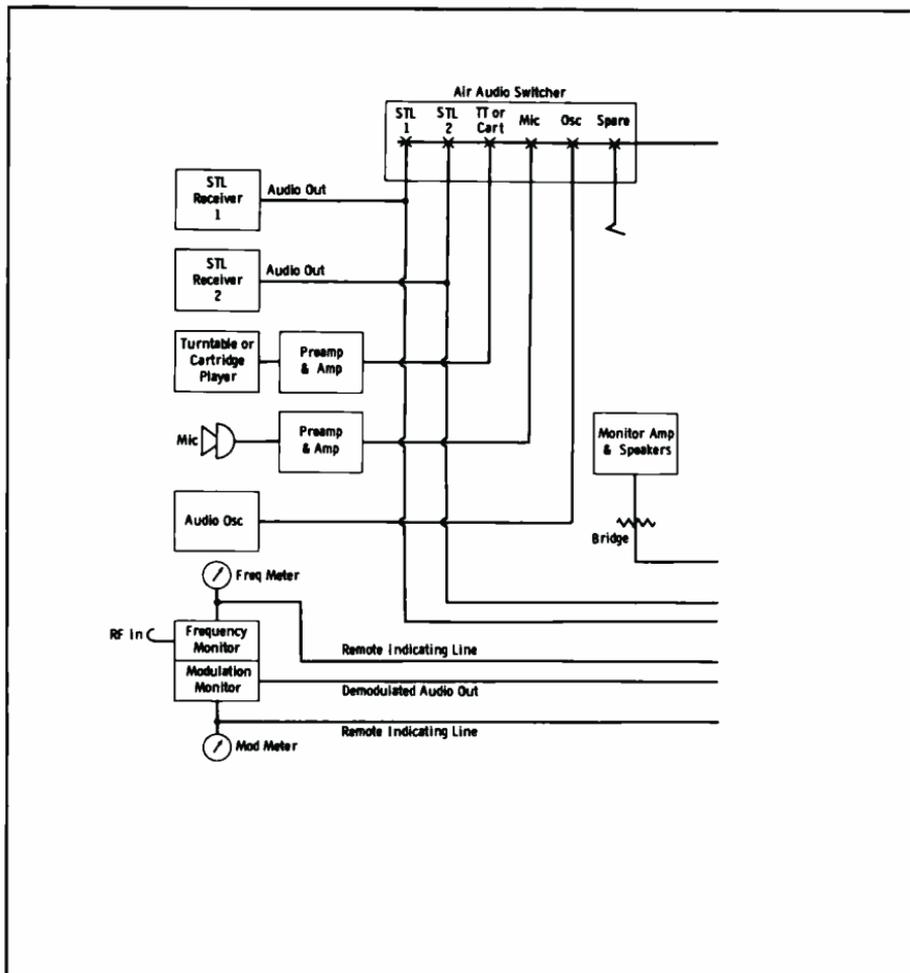
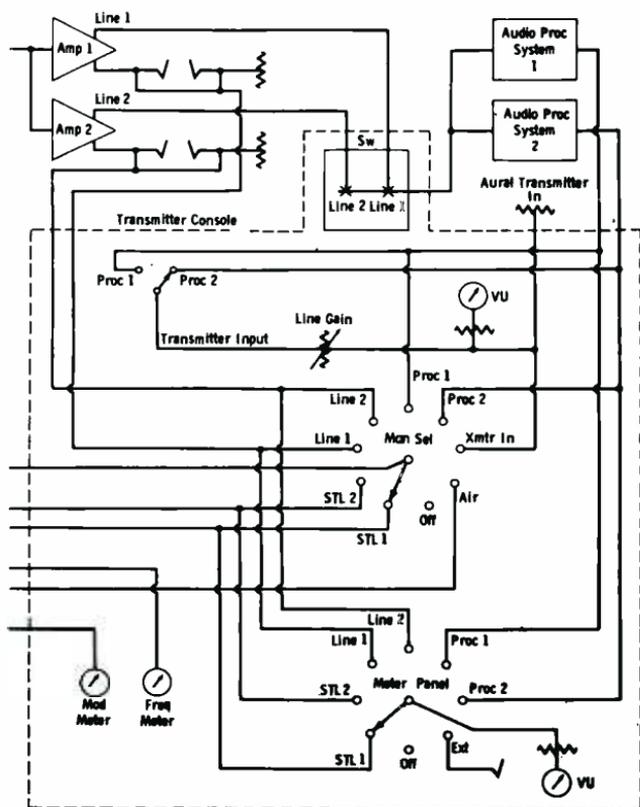


Fig. 10-2. Block diagram of

to be carried out on duplicate paths while the air path is monitored simultaneously.

The standard receiver is variously termed "vsbf demodulator," "sideband demodulator," etc. It normally includes an internal or external chopper which may be switched on when desirable for measurement of depth of modulation.

The sideband response analyzer provides a quick check on modulation sideband characteristics, during test periods. It allows immediate display and evaluation of the effectiveness of transmitter adjustments as they are made.



audio transmitter terminal gear.

## 10-2. AUDIO TERMINAL GEAR

As may be noted from Fig. 10-2, the same amount of planning goes into facilities for quickly checking continuity of the audio path as is true for the video path. By means of the monitor selector switch on the transmitter control console, the direct audio output of stl 1 or 2, the output of either switcher amplifier (line 1 or 2), the output of audio processing system 1 or 2, the transmitter input, or the air signal from the modulation monitor may be selected. In addition, another external air receiver is sometimes used, as well as an aural output from the tv demodulator, when provided.

Note also that the meter panel permits the VU monitoring of any of these separate points independent of the position of the monitor selector switch. Thus levels may be set or servicing performed with maximum efficiency.

The audio processing system may consist of age, limiting, and/or clipping amplifiers. Much of this equipment must be disabled or bypassed when proof-of-performance runs are made, as outlined in the next chapter.

## 10-3. TERMINAL-GEAR LAYOUT

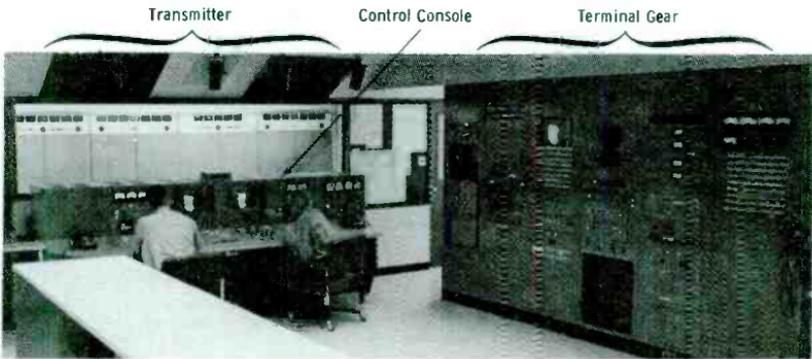
As in the studio, the transmitter equipment is largely rack-mounted, with controls and monitoring facilities centrally located at a control-console position. Fig. 10-3A is a view looking from the rear of the WBBM-TV control room past the control desk through a window into the transmitter room. Terminal gear is housed in the racks at the right in this view. Stl receivers and a precise frequency-offset control for the transmitter are housed in the first cabinet to the left.

Fig. 10-3B is a view of the WBBM-TV Transmitter Room, with the transmitter on the left. On the right, a window gives a view into the transmitter control room. The patch facility below the window permits coupling of test gear in any desired configuration.

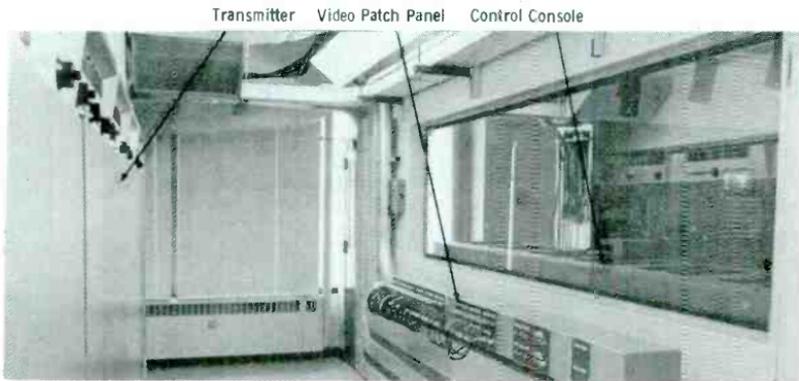
Fig. 10-3C is a closeup view of the control console with the transmitter in the background through the window. The functions of the panels are self-explanatory, except perhaps for the power panel (far right). This panel contains meters for measuring the ac line voltage (phase-to-phase or single phase to ground), ac current for each phase, emergency-generator voltage (also phase-to-phase or any one phase to ground), and emergency-generator frequency.

## 10-4. VIDEO LEVEL SETS

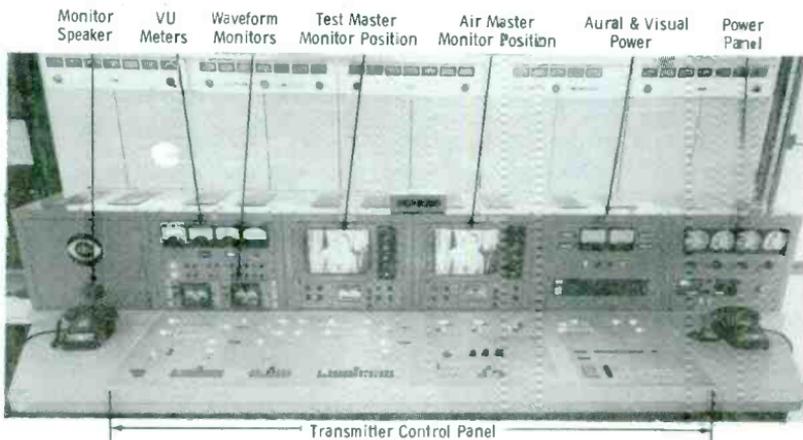
The standard nominal amplitude of the composite video signal for distribution is 1 volt peak to peak. The gain control on each stl



(A) Transmitter control room.



(B) Video patch facility.

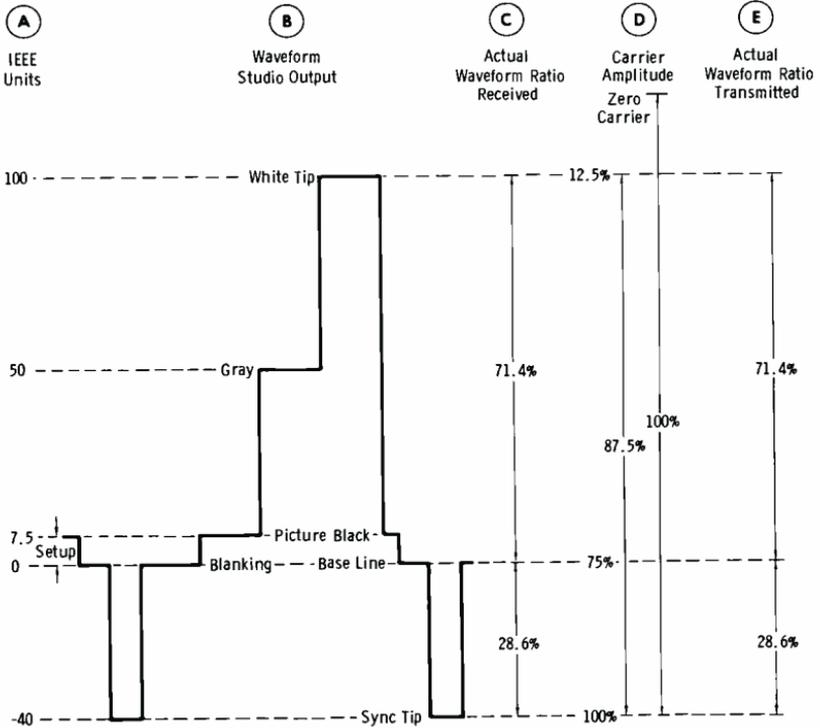


(C) Transmitter control console.

Courtesy WBBM-TV

Fig. 10-3. Transmitter facilities at WBBM-TV.

receiver should be adjusted for the standard value at the output terminal. Similarly, the gain of the switcher DA, proc amps, and other DAs should be adjusted in turn for the 1-volt output. The DAs in the color-correction path normally are used to obtain slightly greater than unity gain to compensate for the inevitable loss in the phase equalizers so that the output of each is again 1 volt. The signal input to the transmitter is normally 1 volt peak to peak.



**Fig. 10-4. Relationship of waveform to carrier modulation.**

See Fig. 10-4. A three-step waveform is shown with various levels called out in IEEE units. Recall that the ratio is usually termed 70 percent video to 30 percent sync, although the actual ratio is 71.4 percent video to 28.6 percent sync.

When this signal is applied to modulate the visual carrier, peak white level must be adjusted so that a nominal value of 12.5 percent of maximum carrier exists. The carrier must never be modulated to zero value (carrier cutoff) for the following reasons:

1. Carrier cutoff causes severe buzz in intercarrier-type receivers as a result of loss of the visual-carrier frequency reference.

2. Transfer linearity is particularly hard to maintain in the region between 10 percent and zero carrier, and is likely to vary with many operating parameters, including APL.

The FCC tolerance on reference white level is 10 to 15 percent of maximum carrier, for a nominal value of 12.5 percent.

Note from Fig. 10-4D that with peak white adjusted for 12.5 percent of maximum carrier, the blanking base line represents 75 percent of maximum carrier, and sync tips represent maximum carrier (100 percent). Thus, full visual modulation represents use of 87.5 percent of the carrier, and sync is 25 percent of the carrier and video is 75 percent of the carrier. In speaking of the rf carrier, the ratio is rightly termed 25 percent sync to 75 percent video.

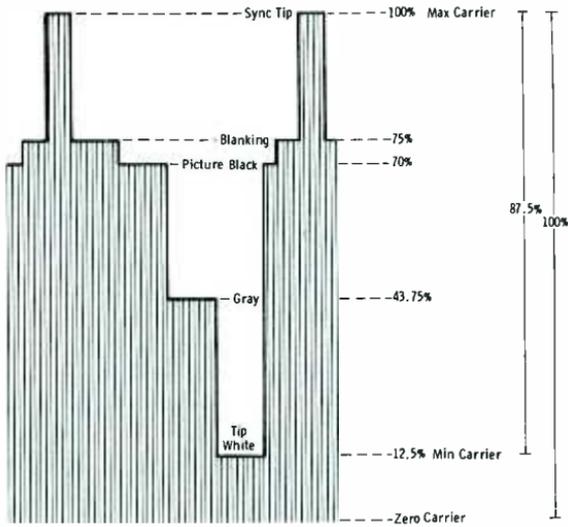
Note carefully that the transmitted signal (Fig. 10-4E) has exactly the same video-to-sync ratio as the studio output (transmitter input) of Fig. 10-4B. This simply emphasizes an important concept in video levels: a nominal ratio of 70 percent video to 30 percent sync is required in order to obtain a modulated-carrier ratio of 75 percent video to 25 percent sync. *This is true only when everything in the video path, including the modulator, is perfectly linear.*

To clarify the concept still further, Fig. 10-5A represents an rf carrier modulated with the signal of Fig. 10-4B. This is standard negative modulation (decrease in carrier power with increase in signal white), as used in the United States. Representative values (nominal) of different gray-scale levels are given in percent modulation. Fig. 10-5B shows a representative test-signal display on a cro driven from a tv demodulator incorporating an electronic chopper to establish a zero-carrier reference. Fig. 10-5C typifies the same type of display during normal programming; the signal is chopped at a 120-Hz rate and displayed with a 30-Hz sweep rate.

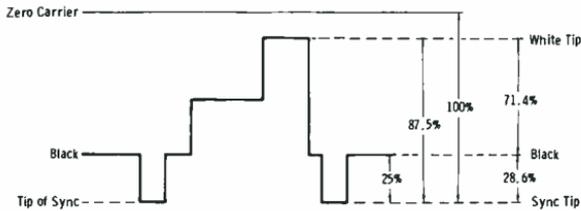
Fig. 10-6 gives the FCC tolerances for all the reference levels. *Caution:* Always keep abreast of current FCC rules; they do change as the state of the art advances.

The video-to-sync ratio required to obtain the proper carrier modulation is normally adjusted in the video stabilizing or processing amplifier. In some instances, the necessary processing is built into the transmitter prior to the point at which modulation occurs. Because of the usual transmitter nonlinearity, this ratio may be something entirely different from that of the transmitted signal.

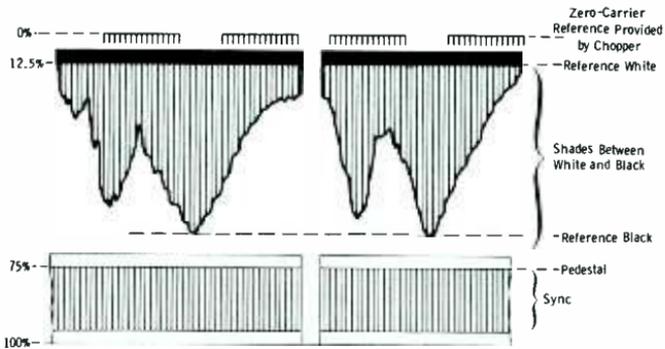
In practice, the sync and video levels are adjusted in the processing amplifier so that the resultant transmitter input gives the required video-to-sync ratio on a properly adjusted tv demodulator operated with a chopper to indicate depth of modulation. The usual overall transfer characteristic of the transmitter results in some compression at both the white and sync extremities of the carrier.



(A) Modulated carrier.

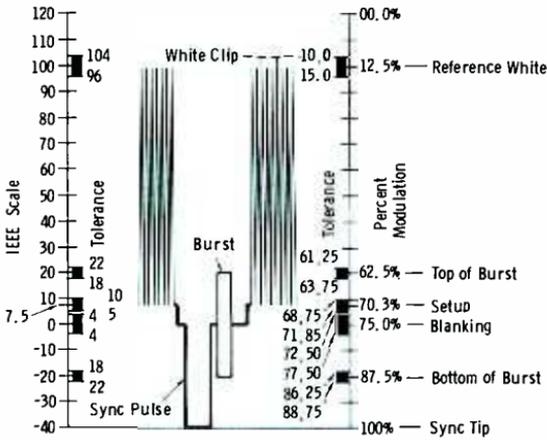


(B) Display on cro.



(C) Display during program.

Fig. 10-5. Representations of waveforms for visual modulation.



Note: Tolerance value specified at blanking level applies to sync amplitude only. The variation of blanking with respect to the tolerance of setup shown is assumed to be zero.

Fig. 10-6. Tolerances for video levels.

Thus it is not unusual to have a sync-to-video ratio of about 50-50 (as compared to the normal 30-70) into the transmitter to give the proper signal ratio at the modulated carrier output. In addition, some amount of white stretch is usually required as observed on a modulated staircase signal. This is more fully treated in the next chapter.

### 10-5. COLOR PRECORRECTION

Because of the nature of vsb transmission, the transmitter and "standard receiver" must be considered as a system in relation to proper amplitude balance and phase characteristics of the upper and lower sidebands. At best (as you are probably aware), the nature of this modulation-demodulation process is such that it causes slight leading white and trailing smear on a white-to-black transition. In practice, you also note an "offset of porches" from an uncompensated transmitter.

The transmitter exhibits low- and high-frequency distortions due to the cutoff characteristics. The receiver exhibits low-frequency distortion due to the vestigial-sideband tuning. Also, due to the aural traps in the receiver, it exhibits considerable high-frequency distortion in the frequency-phase relationship.

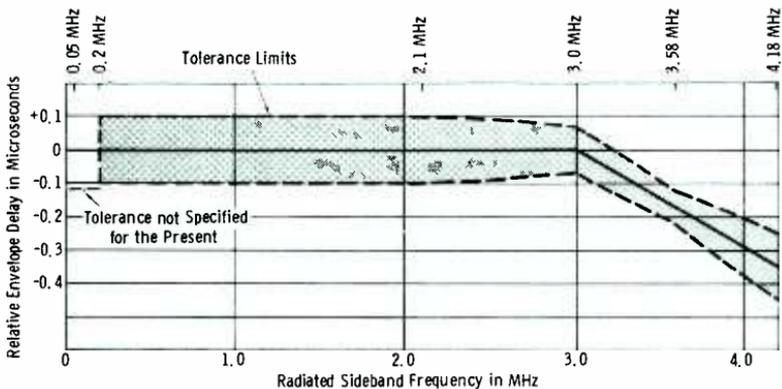
As pointed out in Chapter 2, if skew symmetry can be achieved, color precorrection may be used to practically eliminate quadrature distortion in the overall transmission-reception process.

Time-delay distortion in color signals must be held to a minimum to avoid the "funny-paper" effect of misregistration. (This term is

a carryover from the early days of printing color comics, when color registration was quite imperfect compared with modern printing.)

The delay tolerance for color is based on the average delay in the region from 0.05 to 0.2 MHz, since this is a region within which the phase properties of the vestigial-sideband filter are not a factor. In the sideband-filter cutoff region, corresponding to video frequencies from 0.75 to 1.25 MHz, the corresponding phase distortion must be brought within tolerable limits for good color transmission. At higher frequencies up to 4.18 MHz, the transmitter characteristic is intended to compensate for the average receiver characteristic. The FCC rules provide for envelope-delay compensation at the transmitter for normal receiver errors in this respect.

Fig. 10-1 illustrates a typical series of units employed to meet the preceding requirement (color equalization group). As a result of the general lack of suitable measuring equipment, most engineers have relied on square-wave response to judge the transmitter radiation phase characteristics. Fig. 10-7 illustrates the FCC specification for the transmitter envelope-delay curve for color tv transmissions. The specifications themselves are as follows.



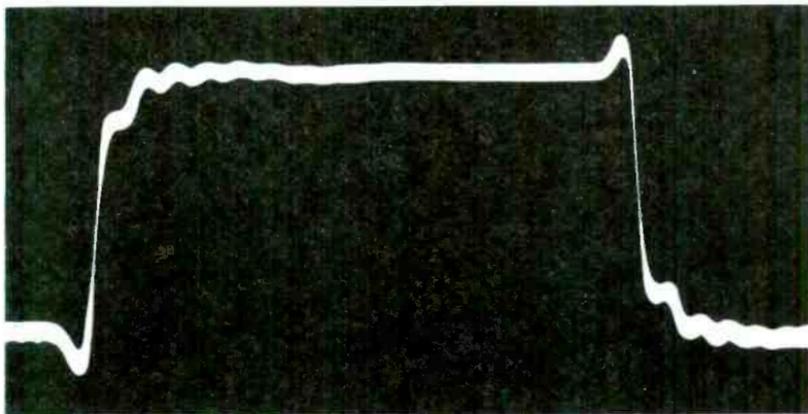
**Fig. 10-7. Envelope-delay tolerances.**

A sine wave, introduced at those terminals of the transmitter which are normally fed the composite color picture signal, shall produce a radiated signal having an envelope delay, relative to the average envelope delay between 0.05 and 0.20 MHz, of zero microseconds up to a frequency of 3.0 MHz; and then linearly decreasing to 4.18 MHz so as to be equal to  $-0.17$  microsecond at 3.58 MHz. The tolerance on the envelope delay shall be  $\pm 0.05$  microsecond at 3.58 MHz. The tolerance shall increase linearly to  $\pm 0.1$  microsecond down to 2.1 MHz, and remain at  $\pm 0.1$  microsecond down to 0.2 MHz. (Tolerances for the interval of 0 to 0.2 MHz are not specified at the present

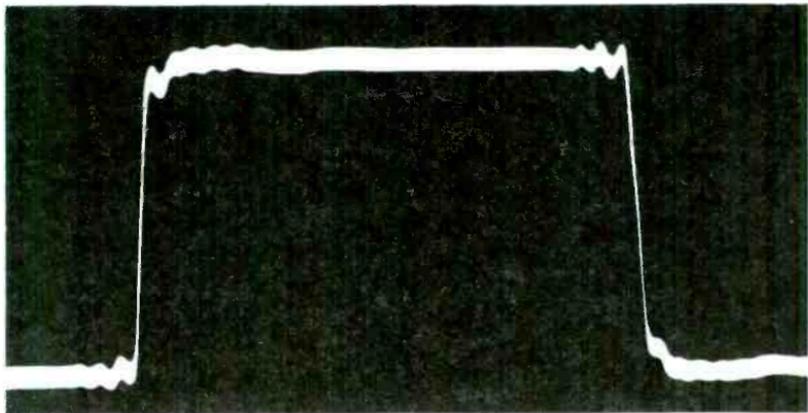
time.) The tolerance shall also increase linearly to  $\pm 0.1$  microsecond at 4.18 MHz.

Fig. 10-8A shows a typical uncorrected transmitter response to a 100-kHz square wave. The transients preceding the transitions result from low-frequency phase distortion inherent in the attenuation of the lower sideband. Ringing after the transitions is caused by phase distortion that results from attenuation of the upper sideband. These defects indicate delay distortion outside the tolerance for color transmission.

When properly adjusted, the precorrection networks can minimize but not completely eliminate the phase distortion. Fig. 10-8B



(A) *Uncorrected response.*

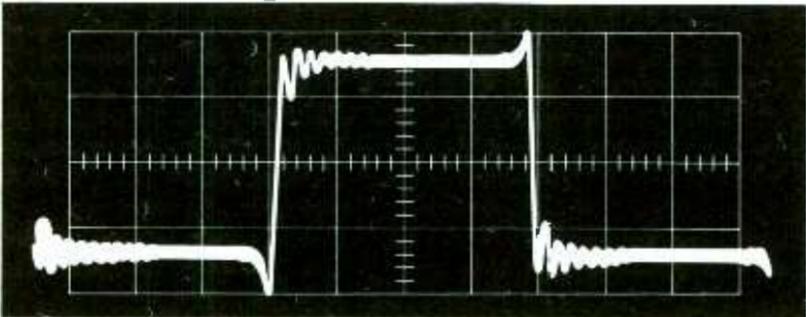


(B) *With phase correction.*

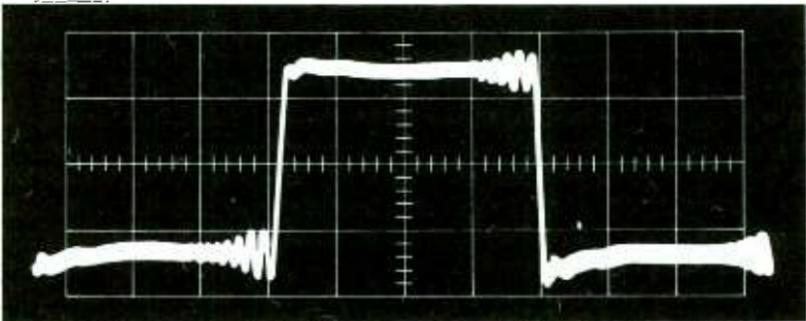
Fig. 10-8. Transmitter square-wave response.

shows the response to the 100-kHz square wave in a phase-corrected installation. The preceding (anticipatory) transients have been eliminated, and the corners following the transitions have been squared. These improvements are made by proper adjustment of the low-frequency response. The ringing has been distributed before and after the transition and reduced in magnitude by the high-frequency phase correction.

The precise FCC specifications can be checked only by using special equipment, such as the recently developed RCA BW-8 envelope-delay measuring set, or the Rohde & Schwarz Type LFM. It is very likely that in the near future (check the current FCC rules) it will be mandatory to use such special equipment for envelope-delay measurements. Up to the time of this writing, the square-wave test along with differential-gain and -phase measurements at 3.58 MHz have been accepted by the FCC. We will discuss in Chapter 11 the adjustments of color pre-correction circuitry using the sine-squared pulse.



(A) *Uncompensated modulator-demodulator response.*



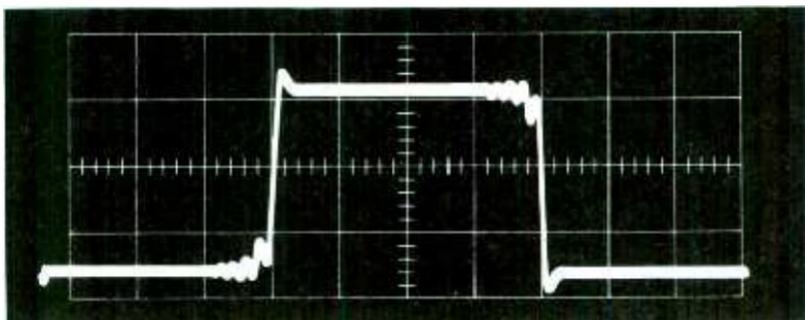
(B) *Predistortion by fixed equalizer.*

**Fig. 10-9. Color pre-correction**

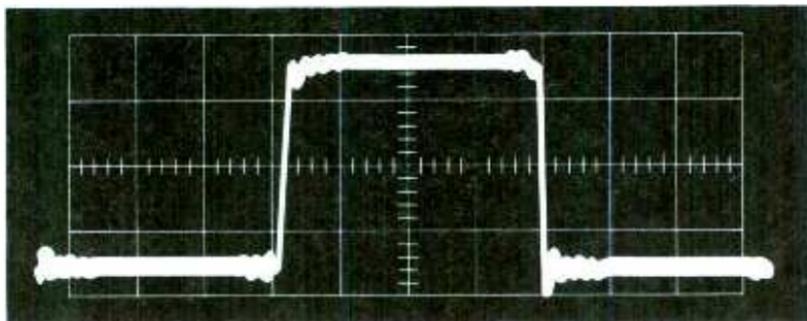
We will discuss briefly here the actual step-by-step procedure in using the square wave to set color pre-correction adjustments. Note first (Fig. 10-1) that color equalization (pre-correction) consists of the following units:

1. A low-pass filter which rapidly attenuates the video response above 4.2 MHz while minimizing attendant phase distortion.
2. A variable low-frequency phase equalizer.
3. A switchable (in-out) fixed receiver equalizer.
4. A switchable (in-out) fixed high-frequency delay.
5. A variable high-frequency equalizer.

You must have a standard demodulator that has been accurately adjusted at the factory. This should be checked every few years. Very few stations find it economical to invest in the specialized test equipment required for proper setting of "wing-trap" and "sound-trap" adjustments to complement the individual demodulator delay equalizer.



(C) *Low- and high-frequency phase equalization.*



(D) *Resultant properly equalized reproduction.*

**adjustments with square wave.**

Use a 100-kHz square wave to modulate the transmitter only about 20 percent. The faster the rise time, the more will be the ringing, which facilitates the adjustments. In any event, the rise time should be  $0.5 \mu\text{s}$  or better. The low-pass filter increases the rise time. Apply the square wave to the DA that feeds the chain of phase equalizers.

The waveform in Fig. 10-9A is the type of presentation that occurs when the transmitter phase is not equalized. Fig. 10-9B shows the type of predistortion accomplished by the fixed receiver equalizer; this ringing is opposite to that caused by the typical receiver sound notch. Fig. 10-9C shows the combined effect of low- and high-frequency phase equalization to obtain the properly equalized reproduction of Fig. 10-9D. The important characteristic to look for here is that the ringing is symmetrically distributed about the transitions.

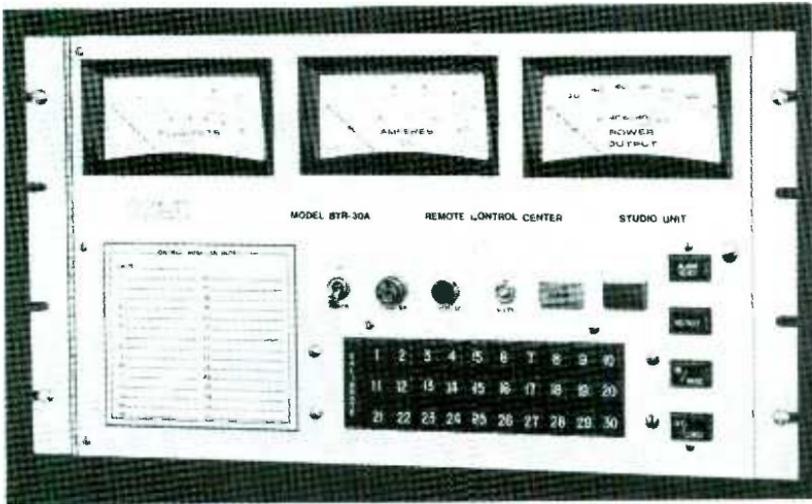
For the initial adjustments (visual carrier only, with sound carrier off), remove the demodulator sound notch and also the transmitter receiver delay equalizer. This gives the best modulator-demodulator combination and allows the greatest accuracy in setting low- and high-frequency delay adjustments. The waveform of Fig. 10-9C is the normal appearance at the transmitter input and the modulator output if the modulator itself needs no correction. The demodulator is fed from an rf pickup in the antenna line following the vsb filter and harmonic filters; a directional coupler is used so that only the forward wave is observed. The variable equalizers should be adjusted in conjunction with switching the fixed equalizers in and out to observe the effect. When all frequency components of the pulse are delayed an equal time, the ringing about the transitions will be symmetrical and of minimum amplitude. The ringing will increase and become nonsymmetrical when you reinsert the demodulator sound notch, but this effect should be nullified by restoring the receiver delay equalizer to the predistortion circuits.

In the next chapter, we will discuss a somewhat more detailed procedure for adjusting envelope distortion precorrection circuitry using the sine-squared-pulse technique. This pulse is being used more and more in visual-transmitter measurement and adjustment, and we have therefore deferred this study so that it is included in the general group of sine-squared-pulse tests on visual transmitters.

## 10-6. REMOTE CONTROL OF TRANSMITTERS

Remote control of both vhf and uhf television transmitters has been authorized by the FCC. This type of operation will eventually become as commonplace in television transmission as it has been for some time in radio broadcasting.

A remote-control system consists of one transmitter control unit containing most of the setup adjustments, and a studio control unit



Courtesy RCA

**Fig. 10-10.** Studio unit of RCA Model BTR-30A remote-control system.

with switchable metering and control adjustments required to meet FCC specifications. The units are interconnected by either a land line or a subcarrier on the station stl.

The RCA Model BTR-30A remote-control system (Fig. 10-10) is an all-solid-state system of modular construction. It has 30 metering channels and 60 individual control functions. An indication of the stepper-relay position is provided on the front panel of the transmitter unit. This is especially useful during weekly calibrations. Color-coded, push-button switches on the transmitter unit are used for local control. The local/remote buttons are illuminated red and green for quick indication of system status. Swing-away front panels on both units provide access to all printed-circuit modules, and all initial and routine adjustments are made from the front. An extension board is provided for testing each module.

Included with the unit is a five-input alarm system. A contact closure is used to activate any one of the five inputs. This can be utilized for continuous surveillance, sensing such things as illegal entry, temperature, flooding, and the like. The alarm signals are returned to the studio unit as part of the telemetry information. When an alarm condition exists, a visual indication is given at the control point by the amber ALARM lamp.

The Model BTR-30A requires one two-way, communications-grade signal circuit between the control point and the transmitter site. Fail-safe provisions meet FCC requirements, and they function with the loss of primary power or control-information reception at

the transmitter unit or with a malfunction of the equipment itself. The fail-safe tone generated in the studio unit also actuates the stepper relay at the transmitter unit. The tone is momentarily interrupted, creating, in effect, short-duration pulses which control the stepper. An interruption of approximately 15 seconds trips the fail-safe circuitry. Two functions, designated ON/RAISE and OFF/LOWER, can be performed on each channel selected. A metered indication of the parameter being controlled can be observed simultaneously. The frequencies used are: fail-safe, 920 Hz; on/raise, 790 Hz; and off/lower, 670 Hz. High-Q, temperature-stabilized capacitors and toroidal inductors are used in all oscillator and tone-detector circuits to assure drift-free operation. Two types of control outputs are available: a preselected external control voltage, or contact closures.

Telemetry is accomplished by converting the dc sample voltage from the transmitter to a frequency in the range 22 to 36 Hz by means of a linear voltage-controlled oscillator. This signal is relayed to the studio unit and converted back to a dc voltage proportional to the input sample for display on any of the 4-inch taut-band, panel-mounted meters. Multiturn calibration controls are provided on the transmitter unit.

For wire service, a single voice-grade telephone line (full duplex) is required for interconnection; dc continuity is not required. The telemetry information (22 to 36 Hz) is returned to the studio unit by means of an amplitude-modulated 1280-Hz signal. Thus all audio signals appearing on the telephone line are confined to the range from 670 Hz to 1280 Hz.

A block diagram of the system is shown in Fig. 10-11. For radio service, the system is designed to mate with RCA aural stl equipment. Control information is transmitted to the transmitter unit on subcarriers multiplexed on the stl. Included in the system are a control-subcarrier generator and detector. A 26-kHz control-subcarrier frequency is used for monophonic stl systems. Telemetry information is sinusoidal and subaudible (22 to 36 Hz). The return path for the telemetry information can be on an sca channel of an fm transmitter, the main channel of an am transmitter, or some other radio circuit capable of handling frequencies from 22 to 36 Hz.

NOTE: Digital-type remote-control systems are covered in a companion volume in this series, *Digitals in Broadcasting*.

## 10-7. PREVENTIVE MAINTENANCE FOR TERMINAL EQUIPMENT

Following is an effective routine of preventive maintenance on transmitter terminal gear. This routine is adaptable for use in any station installation.

*Daily*

1. Check the tuning and afc of stl receiver 1.
  - A. Turn the afc switch off. Check the manual tuning for minimum sync buzz in the sound (when diplexing is used) or for minimum differential gain or phase (using the stairstep signal with 3.58-MHz modulation, Chapter 9).
  - B. Restore the afc. Check to see that the afc holds the receiver tuning at the optimum point found in Step A. If it does not, recenter the afc pull-in point according to the manufacturer's instructions. (Normally, this simply involves adjusting the afc discriminator balance for zero volts at the frequency it is desired to hold, but the instructions for the specific receiver must be followed.)
  - C. Check and record the rf signal level on the meter. Note the weather conditions (clear, fog, rain, snow, etc.).
  - D. Take panel-selector voltage readings and a crystal-current reading (when provided). Make note of any deviation from previous or normal readings.
  - E. Check the level and quality of the video-signal output, through other terminal gear to the transmitter input. Adjust the video processing amplifier for the proper ratio of sync to video at the transmitter output.
  - F. Check the level and quality of the sound output from the stl receiver.
2. Repeat schedule 1 for stl receiver 2.
3. Check (and adjust if necessary) the audio level and sound quality through the entire audio system to the transmitter input. Include a check of the number of decibels of compression or limiting; this will normally be 3 to 6 dB, depending on station use and operating philosophy.
4. Check the operation of the Emergency Broadcast System receiver.

*Weekly*

1. Brush and vacuum the wiring and component sides of the terminal equipment. In cleaner installations, this may become a monthly operation.
2. Check the color pre-equalization adjustments.
3. Check the video processing amplifier(s) for proper white-stretch and/or differential-phase adjustments for proper pre-correction of the transmitter output.
4. Clean all picture-monitor kinescopes and filter plates. Check for proper geometric linearity by means of a grating signal. Adjust the size and linearity if necessary.

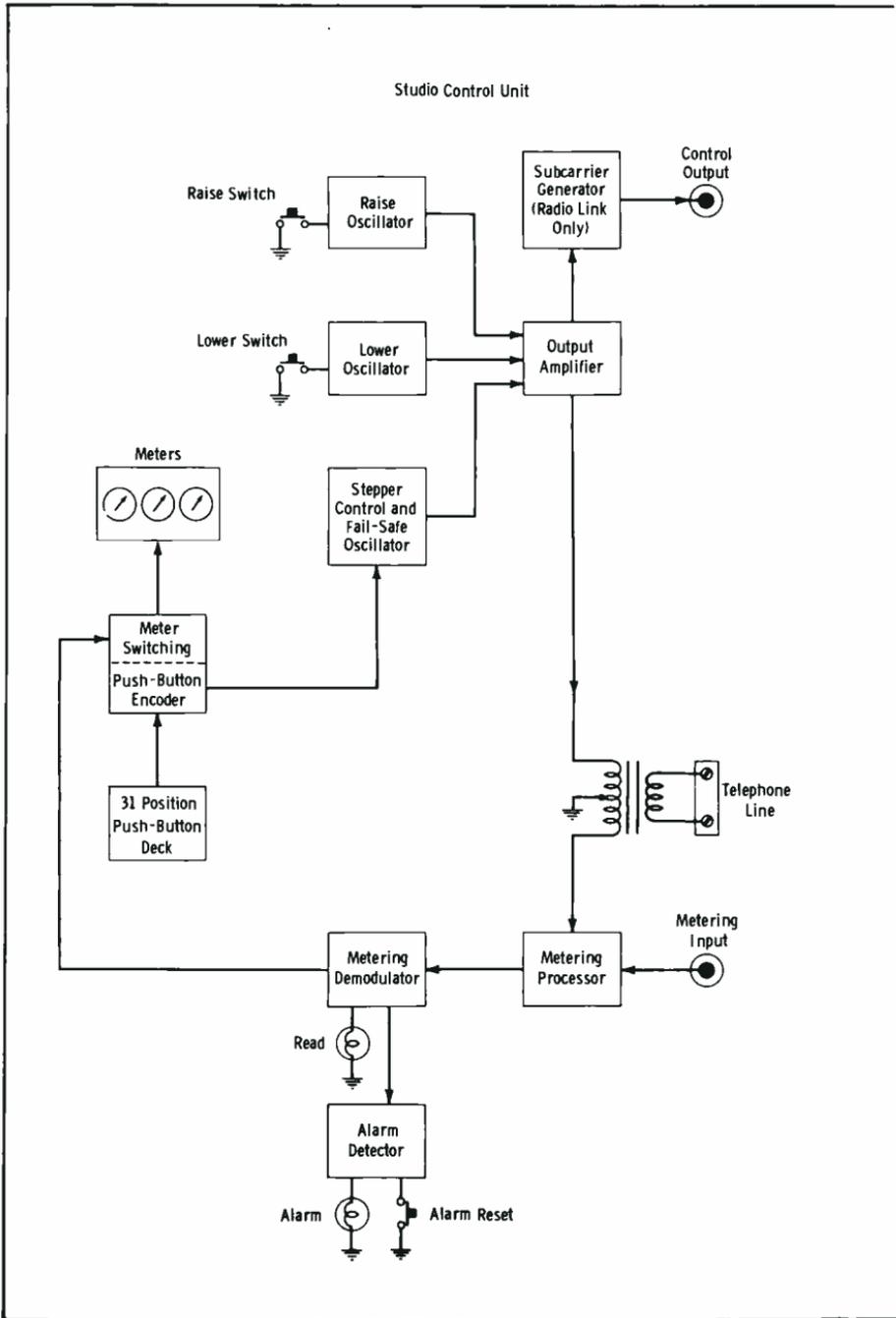
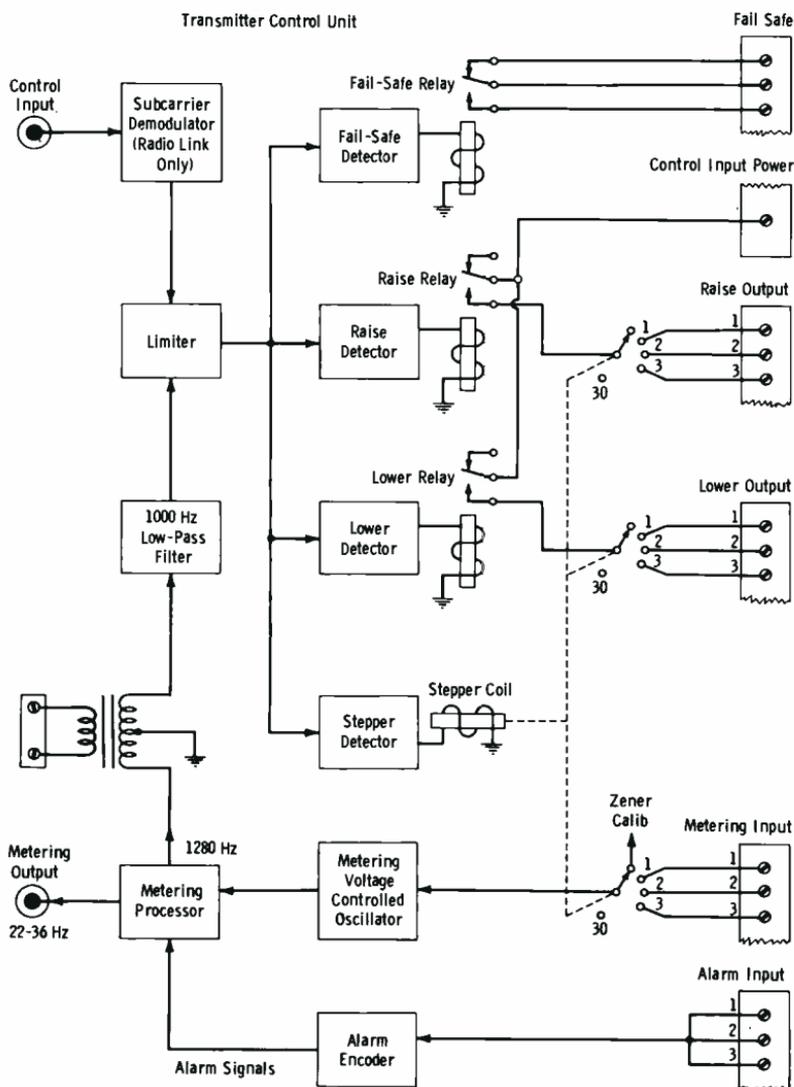


Fig. 10-11. Block diagram of



RCA Model BTR-30A remote-control system.

5. Check the sync generator for frequency stability and conformity with EIA standards. Adjust the pulse widths when necessary.
6. Optimize the slide chain for resolution and geometric accuracy, using a grating signal.
7. Check the operation of the tv demodulator and chopper, side-band analyzer, and all test gear in the terminal-equipment racks.

### *Monthly*

1. Run video measurements, equivalent to those used on proof-of-performance runs, from the studio to the output of the stl receivers. Some stations prefer also to add measurements through the transmitter on this monthly basis.
2. Carry out schedule 1 for the audio system from the studio microphone input to the stl output, or on through the entire transmitter.

### EXERCISES

- Q10-1. Should a cro monitor at the visual-transmitter output show 25 percent sync and 75 percent video, considering only the video-to-sync ratio?
- Q10-2. Should the ratio of sync to video at the transmitter input be exactly 30 percent to 70 percent for full modulation?
- Q10-3. Assuming reference video white is set to result in a carrier level of 12.5 percent of maximum carrier, full visual modulation represents use of what percent of the carrier?
- Q10-4. What test signal is normally used to set the video-to-sync ratio and the degree of white stretch (if any)?
- Q10-5. What unit normally incorporates color-signal precorrection for differential gain and differential phase?
- Q10-6. What units are used to precorrect for the envelope-delay characteristics of the transmitter?

## The Television Transmitter and Proof of Performance

It is assumed here that the reader is familiar with the fundamentals of visual and aural transmitters, tuning procedures, and antenna principles.<sup>1</sup>

### 11-1. PARALLEL TRANSMITTER OPERATION

In earlier years, the primary purpose of operating final power amplifiers in parallel was to increase the power output into the antenna system. More recently, two complete transmitters operating with the finals in parallel (through a special combining network) have been used for the purpose of being able to stay on the air (at reduced power) in the event of failure of one of the transmitters. Fig. 11-1 illustrates such a combined transmitter.

One of the basic requirements for transmitters operating in parallel is that they be mutually independent. That is, if one transmitter should fail, there must be no effect whatsoever on the other, which will continue to operate without distortion. The term "mutual independence" simply means an effective lack of coupling of one transmitter to the other, even though the outputs are normally combined for full-power operation.

Adding the outputs of two separate transmitters in a combining circuit while maintaining mutual independence is normally achieved by means of a *hybrid* constructed from coaxial elements. The basic

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<sup>1</sup> See, for example, Harold E. Ennes, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals* (Indianapolis: Howard W. Sams & Co., Inc., 1971) Chapters 13 and 14.

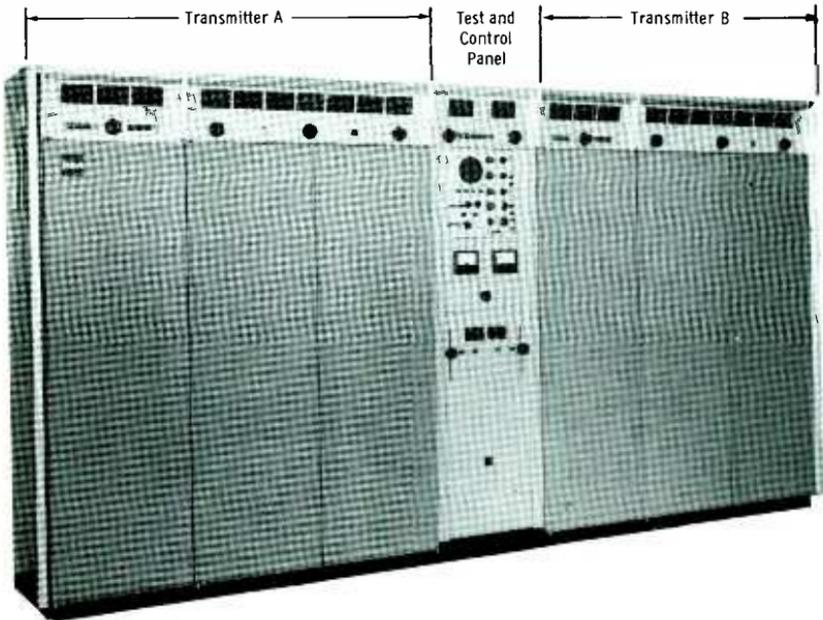


Fig. 11-1. Parallel 12.5-kW vhf television transmitter.

Courtesy RCA

principle of a symmetrical coaxial hybrid is represented in Fig. 11-2. A direct connection (for this particular hybrid) exists between arms 1 and 2, and between arms 3 and 4. There is no direct connection between arms 1 and 4, and arms 2 and 3. Power can flow, however, between arms 1 and 4 if arms 2 and 3 are not terminated, or are connected to a reactive load. Conversely, power can flow between arms 2 and 3 if arms 1 and 4 are not terminated or are connected to reactive loads. This is to say that power can flow between, for example, arms 1 and 4 only by virtue of reflections at arms 2 and 3. The development of hybrid theory is beyond the scope of this

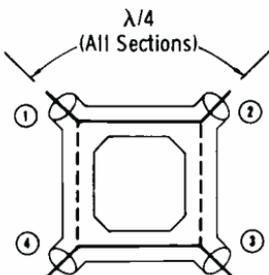


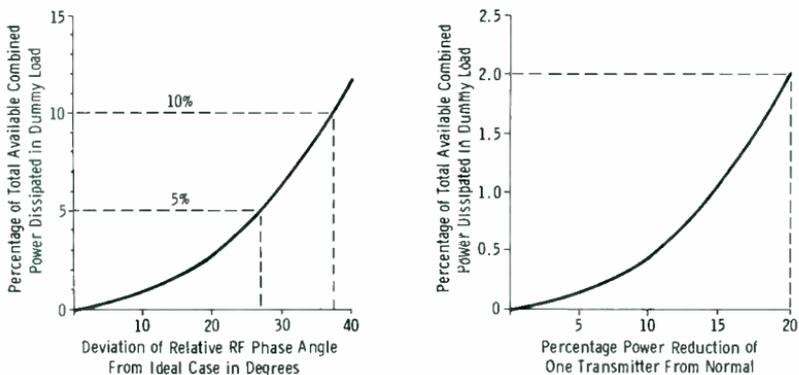
Fig. 11-2. Basic symmetrical coaxial hybrid.

book, but the maintenance engineer or technician should be familiar with certain basic characteristics, as follows:

1. If a single rf drive is applied to, say, terminal 1, with terminal 4 terminated and terminals 2 and 3 working into matching loads, the power at terminal 1 is equally divided at arms 2 and 3. This is an example of a hybrid used as a *power divider*.
2. If two separate rf signals are connected at 1 and 4, and if they are of the same frequency and amplitude with controllable (variable) phase, the *sum* of the signals at 1 and 4 can be made to appear at either arm 2 or 3 (alone), dependent on the phase angle between the rf signals applied at 1 and 4. The condition of mutual independence of transmitters A and B remains valid. This is an example of a *hybrid combiner*. In practice, any desired division of power between 2 and 3 may be obtained simply by adjustment of the rf phase angle between the transmitters feeding terminals 1 and 4.

When the hybrid is operated as a combiner, one output port feeds the diplexer or antenna system, and the opposite output port is terminated in a dummy, or reject, load.

Assume we have two transmitters in a combined operation with identical amplitudes fed into the hybrid combiner, but with relative rf phase of the two carriers which deviates from the ideal case. Fig. 11-3A shows the ratio of power dissipated in the dummy load to the total available combined power, as a function of relative rf phase deviation from the ideal. Note that only 5 percent of the total power is dissipated in the dummy load for a phase deviation of approximately  $26^\circ$  from the ideal relationship. For a deviation of approximately  $38^\circ$  from ideal, about 10 percent of the total combined

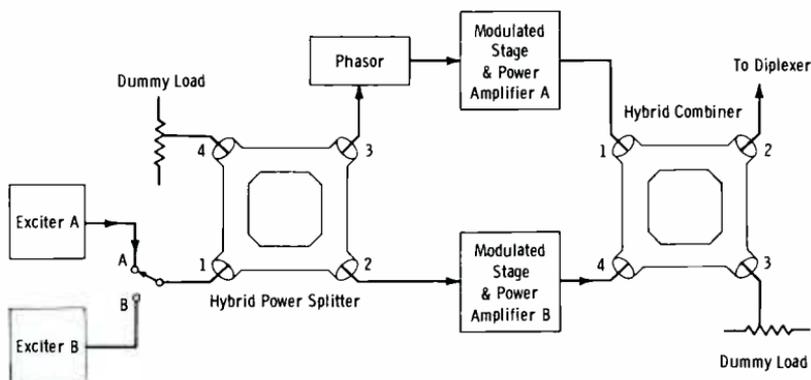


(A) As a function of phase difference. (B) As a function of power reduction.

**Fig. 11-3. Ratio of power dissipated in combiner dummy load to total available combined power of two sources.**

power is dissipated in the dummy load. It can be seen, therefore, that the hybrid combiner has considerable tolerance to phase shift, since the relative phase between carriers can be made quite stable.

Now assume that we have two transmitters in a combined operation with an ideal rf phase relationship at the input ports of the hybrid, but with the relative amplitude of the power output of one transmitter reduced. Fig. 11-3B shows the result. For a relative power reduction of 20 percent, only about 2 percent of the total normal combined power is dissipated in the reject load. Thus we note that in addition to a considerable tolerance to phase shift, the hybrid is relatively insensitive to power unbalances which might be expected in normal operation. In practice, the power in the reject load usually runs less than one to two percent of the total over extended periods of time.



**Fig. 11-4. Basic principle of parallel-operated visual transmitters.**

The simplified block diagram of Fig. 11-4 shows how two transmitters are paralleled. (This illustration represents the visual sections; combining of the aural sections is essentially the same in principle.) A common rf drive source (the exciter unit) is selected either automatically or manually to drive both transmitters through a power divider, which may be a hybrid network as shown. The power at arm 1 is divided equally between arms 2 and 3. The relative rf phase of the transmitter outputs required for proper combining is controlled by means of a variable lumped-constant phasor or motor-driven continuously variable coaxial "line stretcher" in series with one of the output terminals of the exciter power-dividing hybrid.

It should be evident that the amount of power in the combining dummy load (output side of Fig. 11-4) is a sensitive criterion for adjustment of the relative rf phase between transmitters. Thus, in

the example of Fig. 11-4, if equal rf amplitudes of the two transmitters are achieved at arms 1 and 4, the phasor may be adjusted for minimum power at dummy-load arm 3 as indicated on a wattmeter at that location. When this is achieved, the sum of the rf signals appearing at arms 1 and 4 appears at arm 2 of the hybrid, with theoretically zero power in the reject load at arm 3.

Although the visual-transmitter rf phase may be different at terminals 1 and 4 for maximum (sum) rf at terminal 2 of Fig. 11-4, since both visual-transmitter modulated stages operate in parallel, equal depth of modulation and *video* phase coincidence on each carrier must be obtained. Video-modulation phase coincidence is readily obtained by making the electrical lengths of the video input cables from the common source equal, and by maintaining reasonably normal adjustments in the transmitter video circuitry.

A convenient measure of video phase coincidence is obtained by comparing the video outputs of identical demodulators coupled to like points on the output transmission lines of each transmitter. The demodulators are used with a differential-type waveform monitor, or test oscilloscope. A null indicates video phase and amplitude coincidence.

Since the visual modulation processes are paralleled, the combined result is an average of the two. Thus if there is a variation in amplitude-versus-frequency response or modulation linearity from one transmitter to the other, the combined performance is never worse than that of the transmitter with the greater variation (due to the averaging effect), and it can be better than the performance of either when a cancelling of the variation occurs.

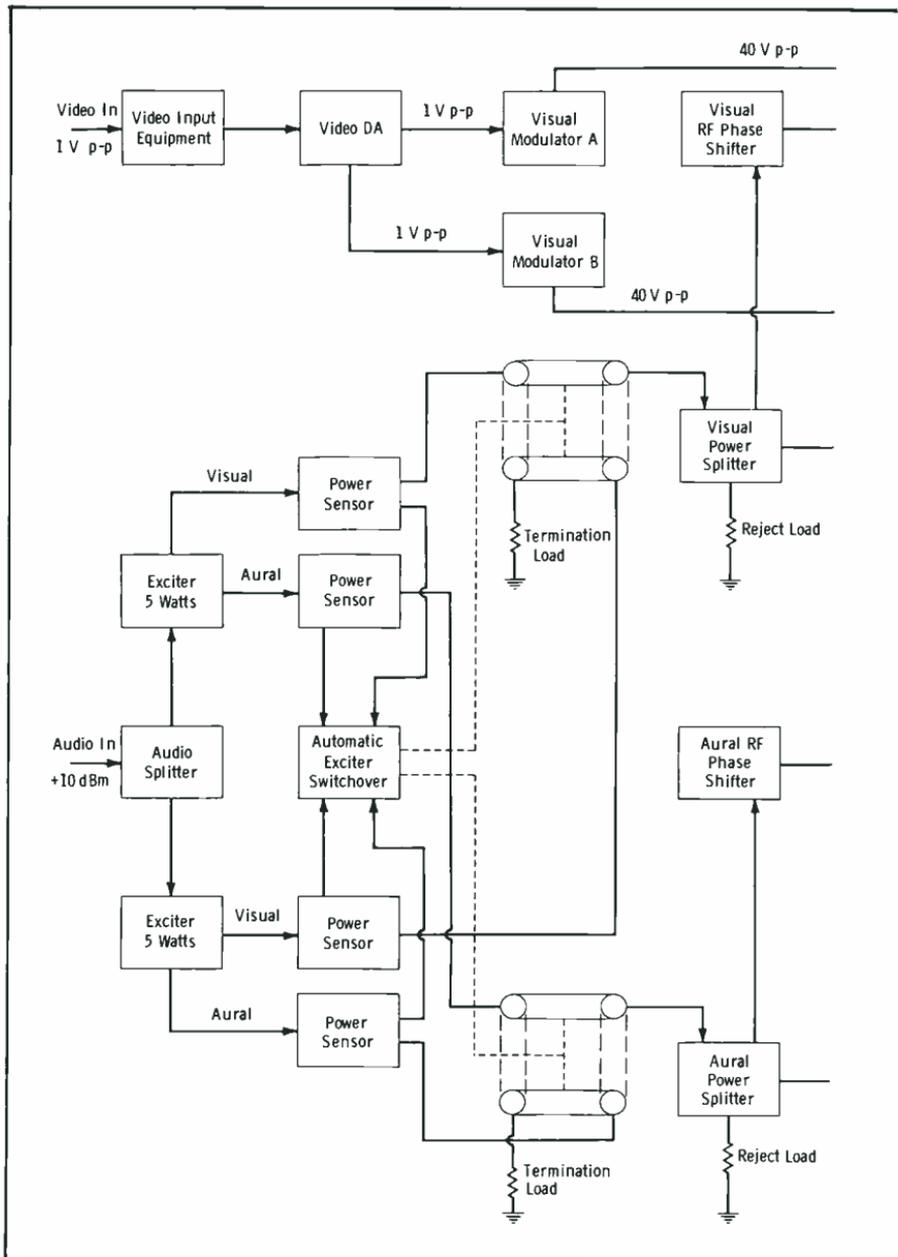
NOTE: Since the aural transmitters are frequency modulated and the power-combining process depends on the maintenance of a reasonable rf phase relationship, aural modulation is usually performed prior to the parallel circuits. We will examine this in more detail shortly.

## 11-2. THE RCA TYPE TT-30FL PARALLELED TRANSMITTER

The RCA Type TT-30FL is a vhf-tv (channels 2-6) transmitter that provides a peak visual power of 30 kW and an aural power of 7.5 kW. These power levels are attained by operating two 15-kW units in parallel.

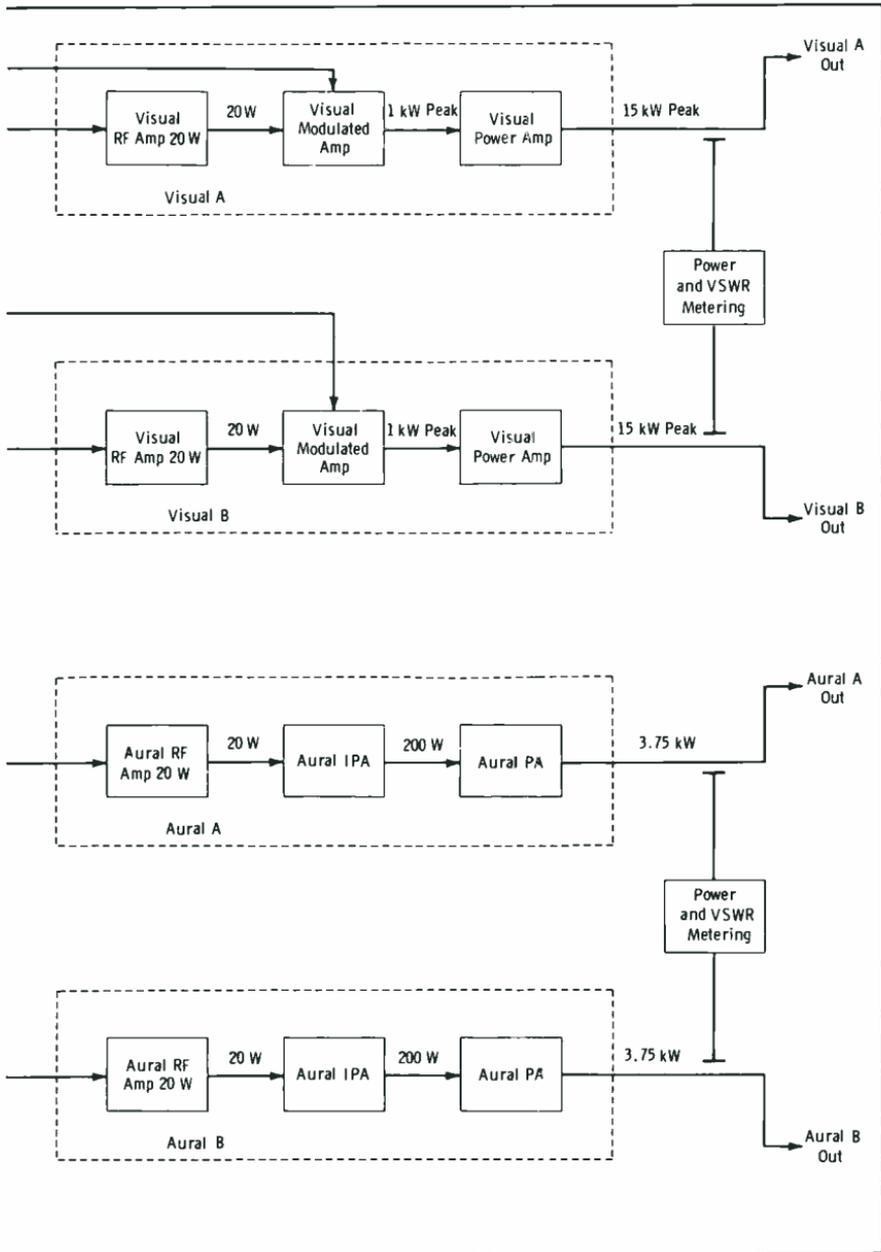
### General Description

Fig. 11-5 is a simplified block diagram of the RCA Type TT-30FL transmitter. Two identical 15-kW transmitter units are operated in a parallel system. Their outputs add up to 30 kW peak visual power and 7.5 kW (rms) aural power. The operation of each transmitter unit is completely independent of the other. Should a failure occur



(A) Transmitter

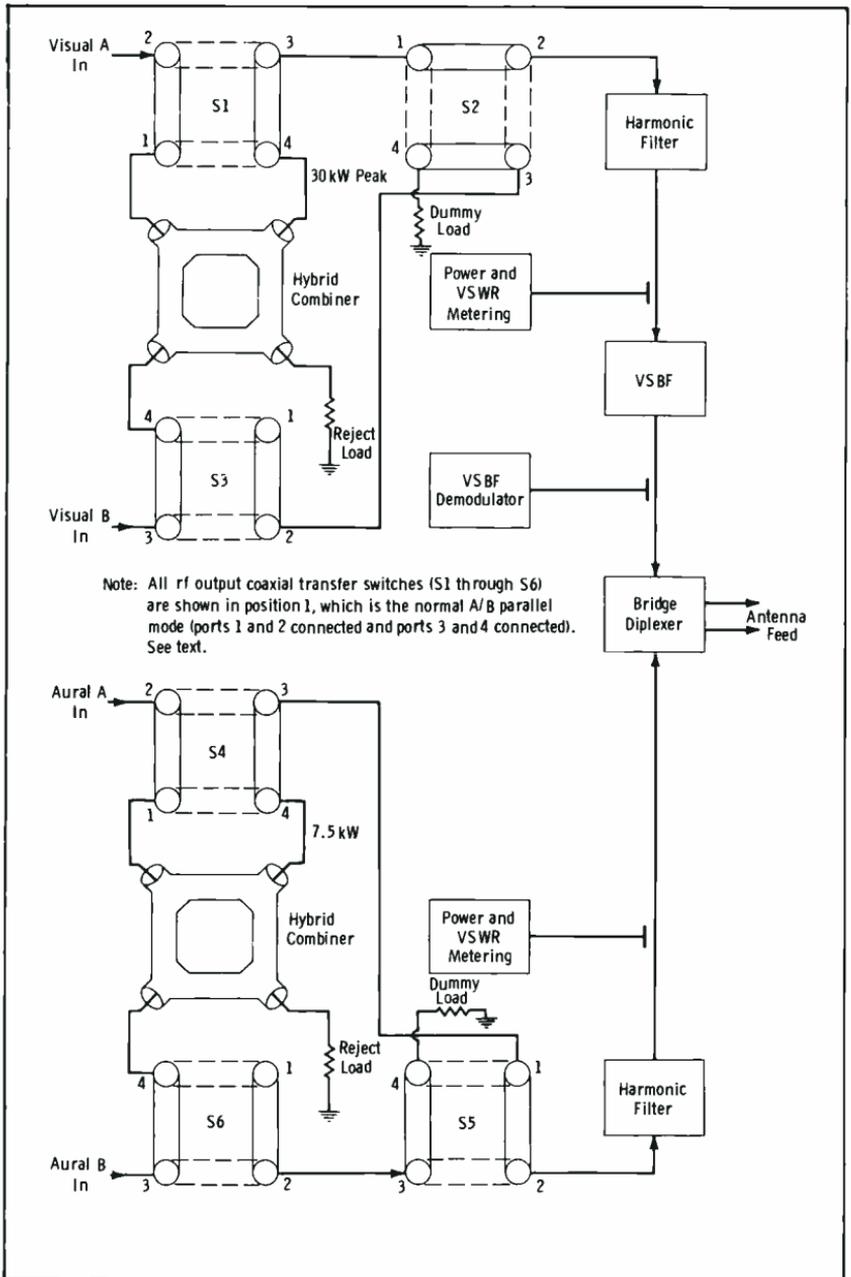
Fig. 11-5. Simplified block diagram of



NOTE: RF output switching on next page.

stages.

RCA Type TT-30FL low-band vhf transmitter.



(B) RF output switching.

Fig. 11-5. Simplified block diagram of RCA Type TT-30FL low-band vhf transmitter—cont.

in one unit, the other continues to provide a nondistorted signal into the antenna system.

Two identical exciters feed into an automatic exciter-switchover circuit. This circuit terminates the output of the standby exciter while the output of the operating exciter feeds the visual and aural stages. The visual output is cw, and the aural output is frequency modulated by the program audio. If the output of the operating exciter degrades or fails, the standby exciter is automatically put in service. The switcher circuit operates so quickly that no noticeable break in video or audio occurs.

The outputs of the two 15-kW transmitter units are combined to provide antenna input powers of 30 kW visual and 7.5 kW aural. If a failure occurs in a transmitter unit, only one input port of the hybrid is fed, and the remaining half power is divided between the other two ports; one-fourth power output in the combining network results, or a reduction of 6 dB. This reduction has little effect within the primary service area of the station. The full power of the one operating transmitter can then be switched directly into the antenna to reduce the power loss to only 3 dB. This final switching operation is initiated from a single push-button switch at an appropriate program time, and it requires about two to three seconds. The reason for this time lapse is that the carrier voltage is temporarily removed during the coaxial contact switching and then is automatically restored after the operation is completed and stabilized. Removal of the carrier power prevents arcing and damage to the switch-gear contacts.

## RF Output Switching

When the transmitter is in the normal A/B parallel mode of operation, all output coaxial transfer switches are in the positions in Fig. 11-5B; ports 1 and 2 are connected, and ports 3 and 4 are connected. This is designated position 1. Now assume a malfunction occurs in transmitter B or that transmitter B is to be serviced for some reason. The A-AIR/B-TEST push button is pressed to energize a relay. This causes coaxial transfer switches S1, S3, S4, and S6 to switch to position 2 (ports 1 and 4 connected, and ports 2 and 3 connected). Transmitter A is thus connected into the antenna, and transmitter B is connected into the test load. Note that S2 and S5 remain in position 1 (ports 1 and 2 connected, and ports 3 and 4 connected) for this mode of operation.

If the reverse of the above condition exists, the A-TEST/B-AIR push button is pressed. In this case, *all* coaxial transfer switches are placed in position 2 (ports 1 and 4 are connected, and ports 2 and 3 are connected).

### **Exciter Unit**

The 5-watt aural/visual exciter, its power supply, and its metering circuits consist of seven plug-in modules in one standard frame. See Fig. 11-6 for a simplified block diagram.

The aural exciter circuitry uses an fm oscillator in an afc feedback loop. The oscillator is followed by a buffer amplifier, a doubler, and two amplifiers to reach the 5-watt output level at carrier frequency. The center frequency of the oscillator is accurately maintained by a reference frequency that is 150 kHz above the fm oscillator frequency. A dc error voltage that represents the difference between the center frequency and the reference corrects for any oscillator drift.

The visual exciter circuitry consists of a temperature-controlled visual oscillator followed by a doubler and two amplifiers to reach the 5-watt output level at carrier frequency.

### **Aural and Visual 20-Watt Amplifiers**

Identical 20-watt amplifier modules are used in the aural and visual rf chains following the 5-watt exciter. These amplifiers and their power-supply and metering circuits consist of five plug-in modules in one standard frame.

The aural and visual amplifiers provide the power necessary to drive the visual modulated amplifier and the aural intermediate power amplifier. As an example of conservative design, the 20-watt amplifiers are required to deliver only about 10 watts.

### **Solid-State Visual Modulator**

The visual modulator and its power supply consist of four standard plug-in modules. A total of 38 transistors of 8 types and 25 diodes of 6 types are used. Motorized controls are furnished for video-gain, sync-gain, and pedestal adjustments, for convenience in adapting to remote control. Fig. 11-7 is a simplified block diagram of the solid-state visual modulator section.

The output of the visual modulator is a 70-volt peak-to-peak video signal that is applied to the visual modulated amplifier. As another example of conservative design, only about 40 volts peak-to-peak is needed for 100-percent modulation.

Differential-phase and -gain correction are accomplished in separate circuits with negligible interaction between the two functions. A feedback clamp circuit that operates from the modulator output or the detected rf output of the transmitter provides dc restoration. The clamp circuit tends to hold the transmitter output constant, even when there are power-line variations.

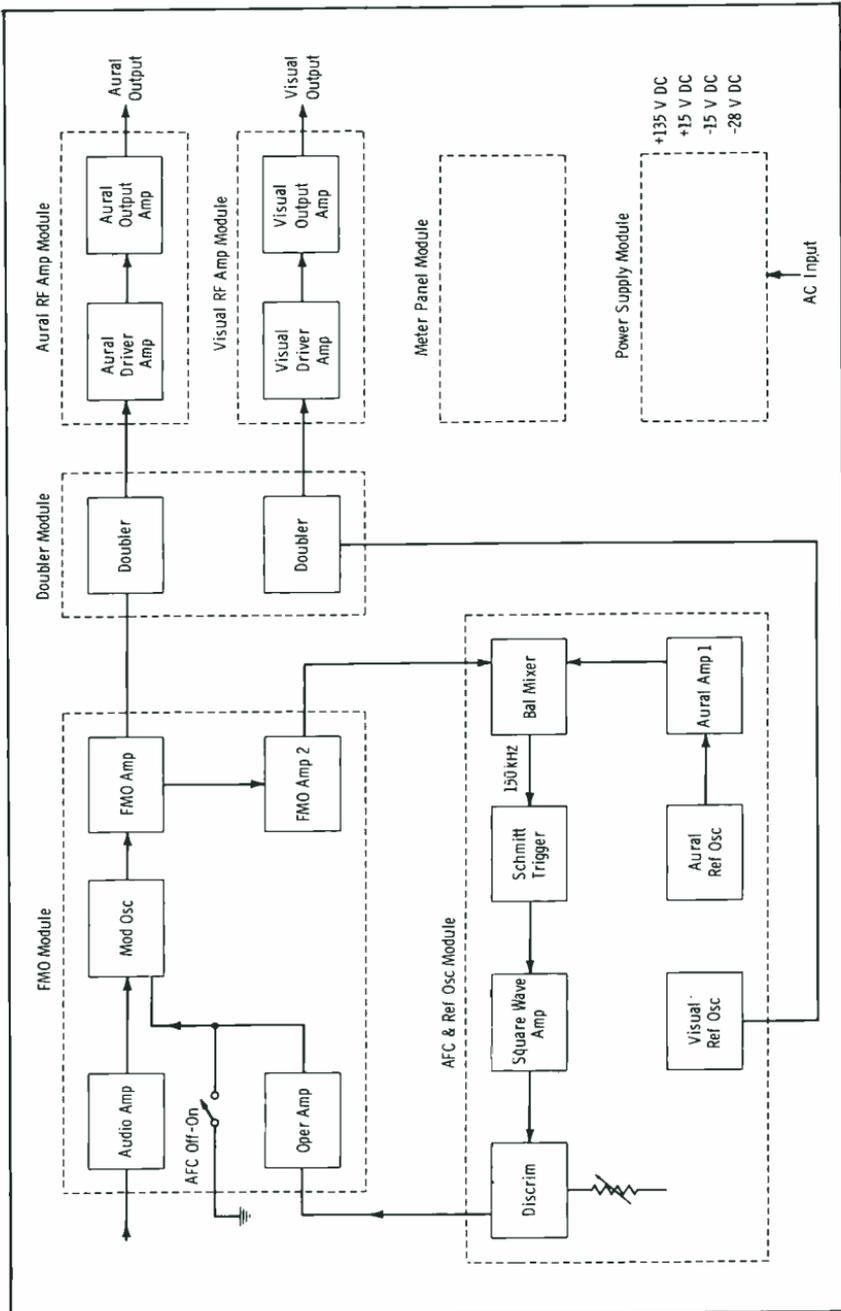


Fig. 11-6. Block diagram of solid-state 5-watt exciter.

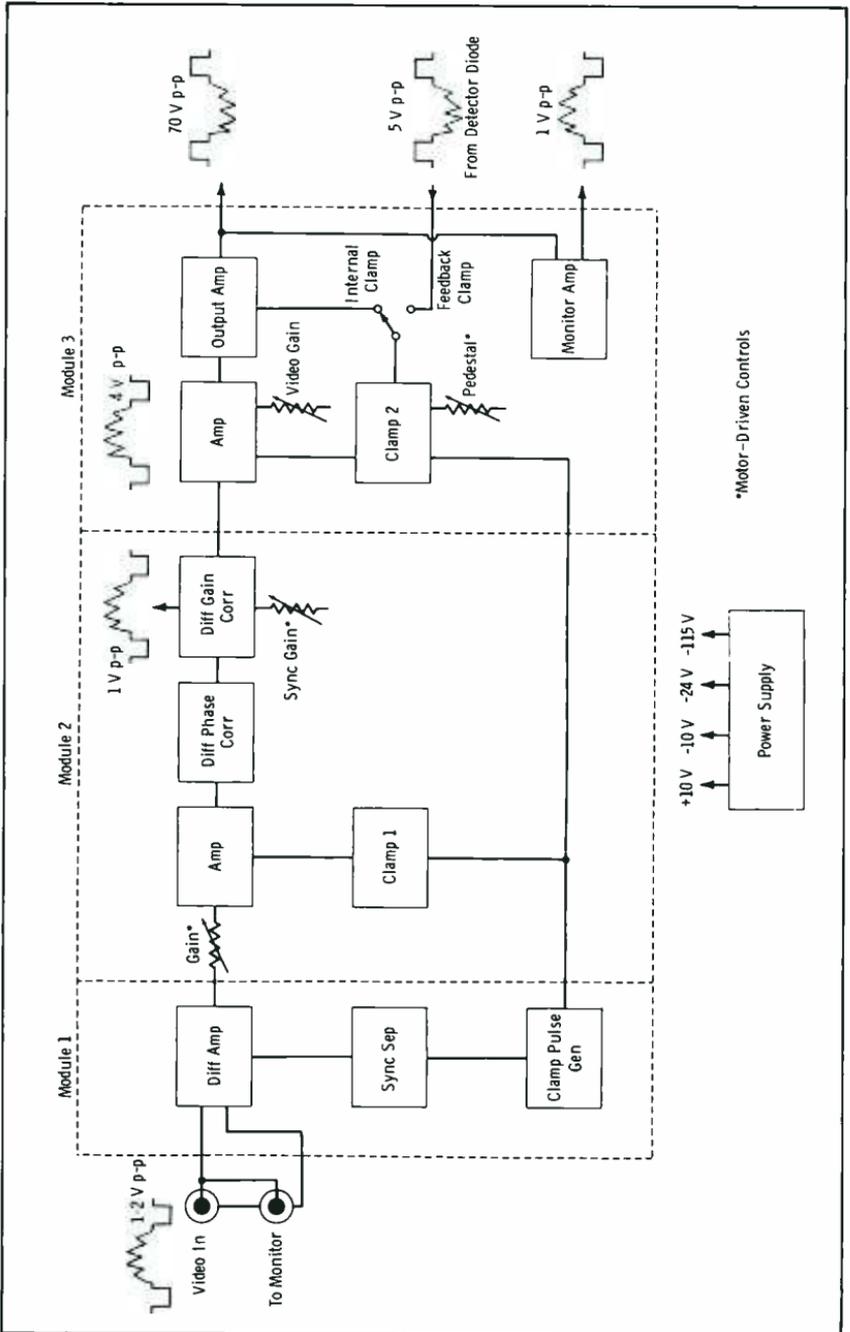


Fig. 11-7. Block diagram of solid-state visual modulator.

### Visual Modulated Amplifier and PA

The visual modulated amplifier uses two Type 8791 Cermolox tubes in a push-pull grid-bias-modulated circuit. The input circuit is heavily loaded to provide good stability and drive regulation, and the Type 8791 tube has extremely linear transfer characteristics.

The visual power amplifier uses a zero-biased high-mu 3CX-10,000A7 triode in a cathode drive circuit. The cathode circuit is a part of the double-tuned overcoupled output circuit of the visual modulated amplifier, providing a circuit with an extremely wide bandwidth. Therefore, tuning of the overcoupled output circuit determines bandwidth. This simplified pa tuning also reduces phase distortion.

The use of a zero-biased triode makes the pa circuitry simple, dependable, and easy to service. The grid is at dc ground, eliminating bypass capacitors and a bias supply, and no neutralization is required.

### Aural IPA and PA

The aural intermediate power amplifier uses a single Type 8791 Cermolox tube operating as a class-C amplifier. The input circuit is heavily damped to present an excellent match to the aural 20-watt amplifier. Because of the broad bandwidth of the circuitry, it is not necessary to retune the input when changing tubes. The plate circuit of the aural ipa is tuned and matched to the pa cathode by using a pi network.

The aural pa uses a 3CX3000A7 zero-biased high-mu triode operating as a ground-grid class-C amplifier. It requires no fixed bias and does not need neutralization.

### Motorized Transmitter Tuning

Motor-driven adjustments are provided on a tuning control panel located in each 15-kW transmitter unit. To select a function to be adjusted, the push button associated with the function is depressed. The push button lights, giving an indication of the selection. Associated with each selection push button is an INCREASE/DECREASE switch which, when depressed, operates a 24-volt tuning-control motor. Metering is accomplished by rotating a 20-position selector switch to the appropriate position. All tuning adjustments are read on the meter provided.

The transmitter control circuits and metering can be located at a console or at a remote location. The circuits are ready for full remote control and automatic logging, and eventual computer control of the transmitter.

### Power Distribution and Control Circuits

The power supplies for each 15-kW transmitter use silicon rectifiers which are well protected against surges, transients, and overloads. The control circuits provide a choice of single-button sequential starting or a step-by-step startup procedure. Automatic or manual reset following an overload is provided. Fig. 11-8 is a simplified power-distribution diagram.

The high-voltage and intermediate high-voltage supplies both employ three-phase full-wave rectifier circuits and furnish plate potentials for all tubes in the transmitter. Screen potentials for the Type 8791 tubes are obtained from a single-phase full-wave regulated supply. No bias supplies are required. The three-phase, 60-hertz main power line to each 15-kW transmitter should be capable of delivering a minimum of 50 kVA and a maximum of 300 kVA at 208/240 volts  $\pm 5$  percent. Two high-voltage transformers are provided with primary taps and are operated from the power-line voltage. Power to the remaining components of the transmitter is supplied through a distribution transformer equipped with primary taps so that the output voltage is always 230 volts. No taps are necessary on any other transformers. Constant-voltage transformers are employed to maintain all filament voltages constant to within one percent.

### 11-3. BROADBANDING

Broadbanding the transmitter concerns the proper tuning to obtain the required amplitude-versus-frequency response, while at the same time maintaining proper power output and proper upper and lower sideband shape factors, along with color characteristics within EIA or (at least) FCC specifications.<sup>2</sup>

Remember that bandwidth (separation between humps in the response of an overcoupled circuit) is largely affected by the degree of coupling, whereas flatness across the top of the response curve is largely affected by loading. In general, if the amplifier plate current is higher than specified for a specific power output, the bandwidth (degree of coupling) is too wide, and the long-term power (plate) dissipation of the final tubes might be exceeded. If the power output is being obtained with much less plate current than

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<sup>2</sup> The proper tuning of conventional rf circuits and of overcoupled circuits as applied to television transmitters is covered in Chapter 14 of Harold E. Ennes, *Television Broadcasting: Equipment, Systems, and Operating Fundamentals* (Indianapolis: Howard W. Sams & Co., Inc., 1971).

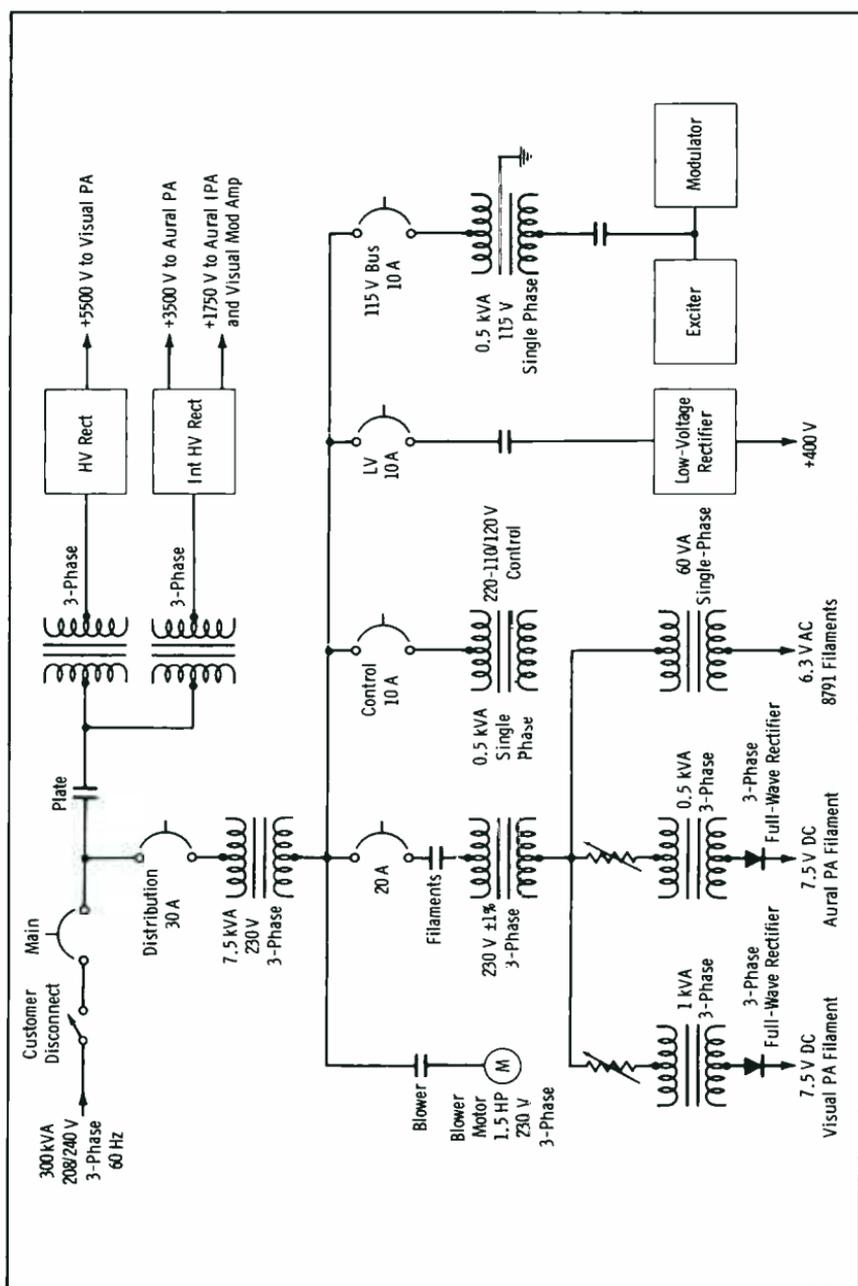


Fig. 11-8. Simplified diagram of power distribution in RCA Type TT-30FL transmitter.

specified, chances are that the bandwidth is too narrow. Either of these conditions alerts you to the fact that it is time to run complete frequency-response checks. This need normally also is indicated by the daily checks with the sideband response analyzer. The use and interpretation of the response curves for this application are covered elsewhere.<sup>3</sup> (We will touch briefly on tuning procedures for modern uhf klystrons later in this chapter, since the cited reference largely describes vhf tuning procedures.)

The FCC proof-of-performance data are normally compiled from single-frequency sine-wave runs with 200 kHz (0.2 MHz) as the reference frequency. The condition for this measurement is specified by the FCC as follows (always check current FCC Rules):

The attenuation characteristics of a visual transmitter shall be measured by application of a modulating signal to the transmitter input terminals in place of the normal composite television video signal. The signal applied shall be a composite signal composed of a synchronizing signal to establish peak output voltage plus a variable-frequency sine-wave voltage occupying the interval between synchronizing pulses. (The synchronizing signal referred to . . . means either a standard synchronizing waveform or any pulse that will properly set the peak.) The axis of the sine wave in the composite signal observed in the output monitor shall be maintained at an amplitude 0.5 of the voltage at synchronizing peaks. The amplitude of the sine-wave input shall be held at a constant value. This constant value should be such that at no modulating frequency does the maximum excursion of the sine wave, observed in the composite output-signal monitor, exceed the value 0.75 of peak output voltage. The amplitude of the 200-kHz sideband shall be measured and designated 0 dB as a basis for comparison. The modulation signal frequency shall then be varied over the desired range and the field strength or signal voltage of the corresponding sidebands measured. As an alternate method of measuring, in those cases in which the automatic dc insertion can be replaced by manual control, the above characteristic may be taken by the use of a video-sweep generator and without the use of pedestal synchronizing pulses. The dc level shall be set for mid-characteristic operation.

A sync signal and sine wave of the FCC-specified type are shown in Fig. 11-9. The sine wave occupies the region of 25 to 75 units; therefore, the axis is at 50 units of a composite signal that places sync peaks at 100 percent of carrier (100 units). The transmitter modulator gain is left adjusted for the original normal input of 10 to 100 IEEE units of video. Note, therefore, that full video modulation of the transmitter does not occur on single-frequency sine-wave runs.

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<sup>3</sup> *Loc. cit.*

**Fig. 11-9. Proportionment of sine wave and sync pulses for transmitter frequency-response runs.**

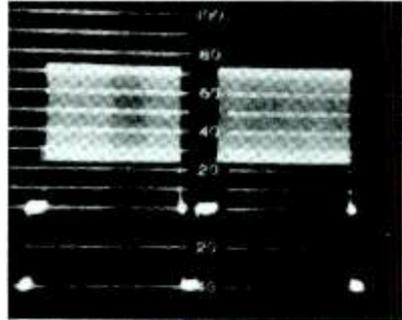
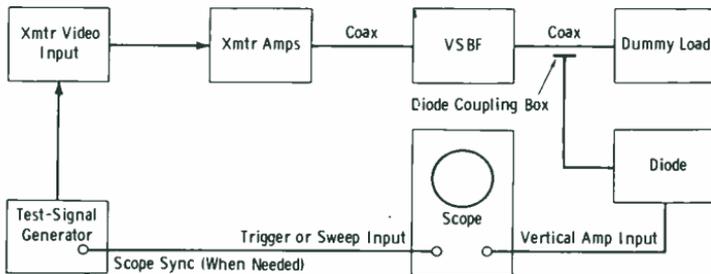


Fig. 11-10 illustrates a typical test setup for transmitter frequency-response (attenuation-characteristic) runs. Although this meets the FCC requirements (video test signal to transmitter input terminals), it is good operating practice to send the test signal through the stl (when used). This then results in an overall test of stl, transmitter terminal gear, and transmitter. The transmitter should be operated into a dummy load.



**Fig. 11-10. Setup for measuring transmitter frequency response.**

The 0.2-MHz sine wave (Fig. 11-9) is the reference level on the scope and is designated 0 dB. The frequency is then increased in steps, with the input amplitude to the transmitter held constant, and the response is tabulated as in Table 11-1. These data may then be transferred to the diode-demodulator curve of Fig. 11-11. The FCC visual-transmitter frequency-response requirements for monochrome and color are tabulated in Chart 11-1.

Fig. 11-12 illustrates how you can plot your own curve for the particular application involved. This can be drawn on a master and copies made on the station duplicating equipment (when available).

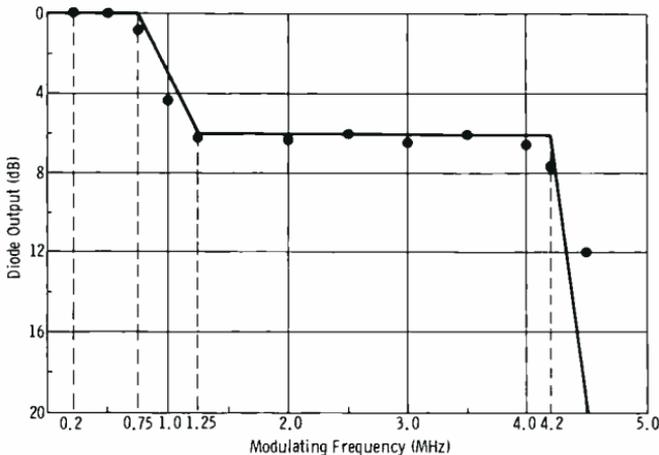
Note that with vestigial-sideband transmission (see Fig. 11-13) the characteristic of the sideband attenuation (upper and lower) must be considered. The FCC specifications (at the time of this

**Table 11-1. Video Frequency-Response Data**  
(Example of typical readings taken on WTAE-TV main visual runs)

Frequency (MHz)	Diode Response	
	Percent	Decibels
0.2	100	0
0.5	100	0
0.75	90	-0.9
1.0	60	-4.4
1.25	49	-6.2
2.0	49	-6.2
2.5	50	-6.0
3.0	48	-6.3
3.58	49	-6.2
4.0	48	-6.3
4.2	42	-7.5
4.5	25	-12.0
4.75	08	-21.9
5.0	05	-26.0

writing—always check current FCC Rules) in this respect are as follows:

The field strength or voltage of the lower sideband . . . shall not be greater than -20 dB for a modulating frequency of 1.25 MHz or greater and in addition, for color, shall not be greater than -42 dB for a modulating frequency of 3.579545 MHz (the color subcarrier frequency). For both monochrome and color, the field strength or voltage of the upper sideband as radiated or dissipated and measured



**Fig. 11-11. Graph plotted from data in Table 11-1.**

### Chart 11-1. Visual-Transmitter Frequency-Response Requirements

NOTE: Always check current FCC Rules.

Monochrome	Color
<p>The overall attenuation characteristics of the transmitter, measured in the antenna transmission line after the vestigial sideband filter (if used), shall not be greater than the following amounts below the ideal demodulated curve.</p> <p>2 dB at 0.5 MHz                  2 dB at 1.25 MHz                  3 dB at 2.0 MHz                  6 dB at 3.0 MHz                  12 dB at 3.5 MHz</p> <p>The curve shall be substantially smooth between these specified points, exclusive of the region from 0.75 to 1.25 MHz. Output measurement shall be made with the transmitter operating into a dummy load of pure resistance and the demodulated voltage measured across this load.</p>	<p>A sine wave of 3.58 MHz introduced at those terminals of the transmitter which are normally fed the composite color picture signal shall produce a radiated signal having an amplitude (as measured with a diode on the rf transmission line supplying power to the antenna) which is down <math>6 \pm 2</math> dB with respect to a signal produced by a sine wave of 200 kHz. In addition, between the modulating frequencies of 2.1 and 4.1 MHz, the amplitude of the radiated signal shall not vary by more than <math>\pm 2</math> dB from its value at 3.58 MHz. At the modulating frequency of 4.18 MHz, the amplitude of the radiated signal shall not be down more than 4 dB below its value at 3.58 MHz.</p>

... shall not be greater than -20 dB for a modulating frequency of 4.75 MHz or greater.

When field-strength readings of the upper and lower sidebands are made, the test signal of Fig. 11-9 is used. The field strength

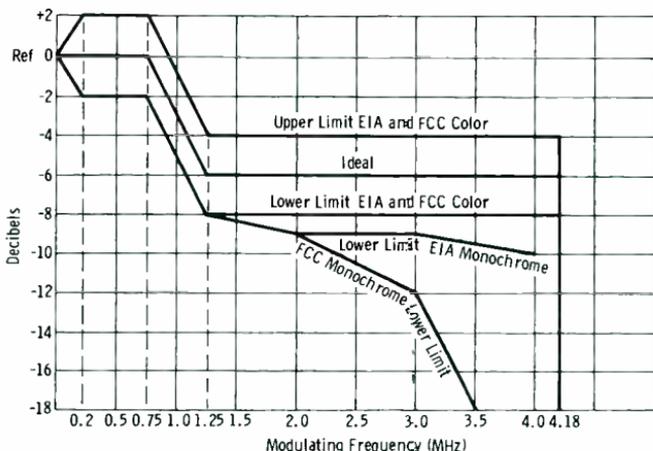
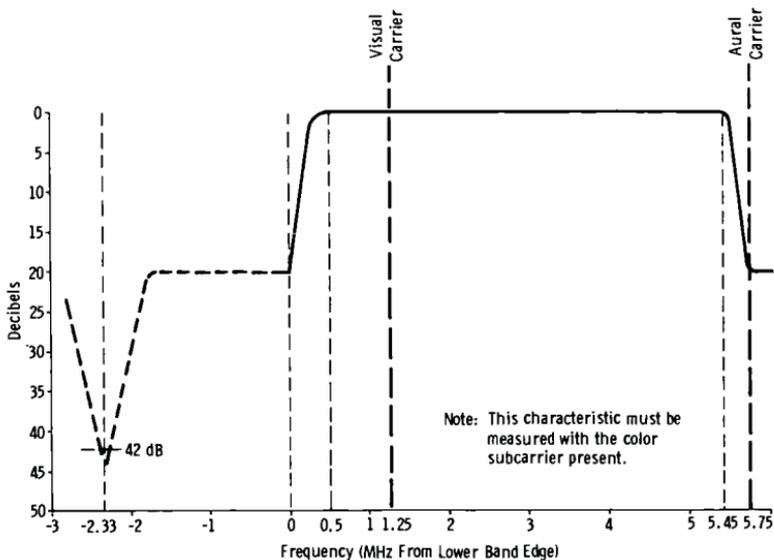


Fig. 11-12. Example of frequency-response graphs.

(for example) with a modulating signal of 2 MHz is measured at the carrier frequency plus 2 MHz for the upper sideband, and at the carrier frequency minus 2 MHz for the lower sideband. The field strength of the carrier frequency itself is the reference 0 dB.

In practice, the upper-sideband response is measured as previously described for Fig. 11-11. A video sweep, or video sweep incorporated with the sideband analyzer, is employed for the lower-sideband at-



**Fig. 11-13. Visual-sideband attenuation characteristic.**

tenuation characteristic. The transmitter is fully modulated (87.5 percent), and the marker is moved to 0.2 MHz for the reference 0 dB. Maximum scope gain must normally be used to measure the lower-sideband response at frequencies above 1.25 MHz. Note that when color standards must be met, the lower-sideband energy at 3.58 MHz must be at least 42 dB down.

Notice also that with vestigial-sideband transmission, the response as measured by the diode may be incorrect due to faulty transmitter tuning for the proper sideband attenuation. Therefore, the sideband attenuation characteristic *must* be ascertained and recorded as part of any proof-of-performance run.

Field-strength measurements for these characteristics or for harmonic radiation are not normally required, except upon demand by the FCC. This may be required if numerous complaints involving your transmitter frequency have been filed with the Commission.

#### 11-4. UHF KLYSTRON TUNING

The klystron is normally supplied from the factory pretuned for the specific station operating frequency. However, it should be checked (and tuned if necessary) to assure proper transmission of color information from the visual amplifier. Tuning procedures may differ with the type of klystron and the specific transmitter in which it is used. The information presented here is largely as used for the RCA Type TTU-30A uhf television transmitter.

**NOTE:** The subject of how to change a Klystron is covered in Section 11-5.

The cavities contained in the visual klystron are tuned as follows (Fig. 11-14 illustrates the approximate tuning points and a simplified diagram of the visual klystron):

1. Cavities 1 and 4 (bottom and top cavities, respectively) are tuned for maximum power at the carrier frequency. (The first cavity is tuned slightly above the carrier frequency on low channels to flatten the response.)
2. Cavity 2 is tuned about 3 MHz below the lower sideband for suitable klystron gain and optimum response at the lower end of the passband.
3. Cavity 3 is tuned about 3 MHz above the upper sideband for suitable klystron gain and optimum response at the upper end of the passband.

In addition to obtaining maximum power output, a flat sideband response from  $-1.0$  to  $+4.5$  MHz should be maintained. To determine the proper tuning points, it is necessary to use the sideband response analyzer to observe the transmitter response from the visual klystron.

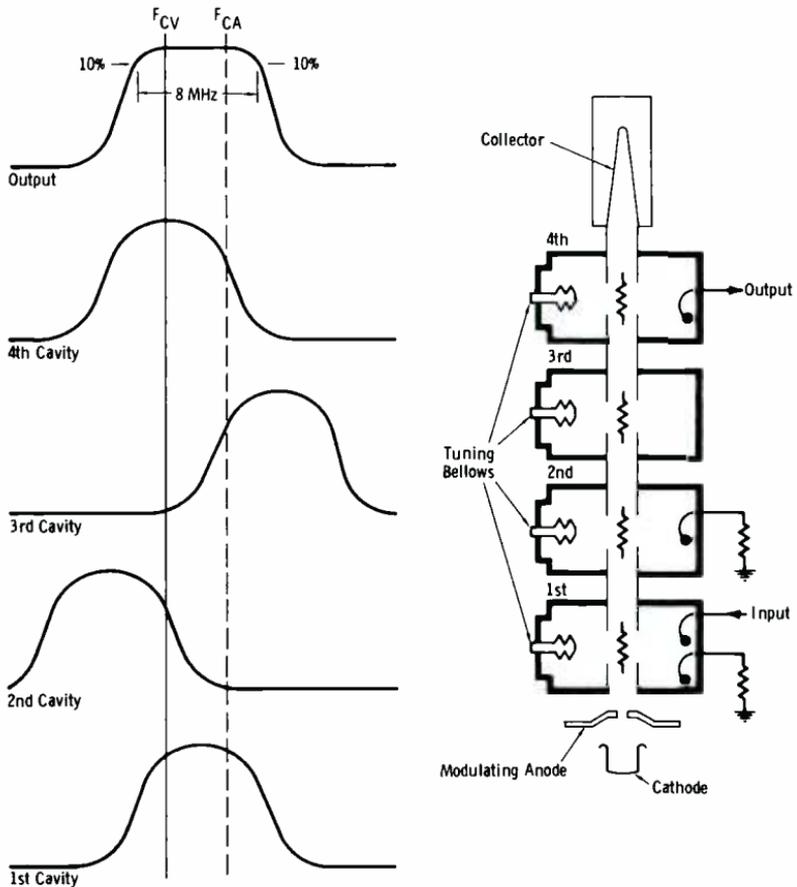
The following description applies specifically to the RCA Type TTU-30A transmitter, and is presented here courtesy of RCA.

Prior to testing the visual klystron, remove all power from the transmitter and then disconnect the rf load from the aural transmission line. Reconnect the rf load to the visual output transmission line following the reflectometer unit. In addition, a directional coupler should be inserted between the reflectometer unit and the rf load to provide a monitoring point for observing sideband response on the sideband analyzer.

To tune the visual klystron, proceed as follows:

1. Insert a Thruline wattmeter between the visual-driver output cable and rf input connector on the visual klystron.

2. Connect a coaxial cable between the monitoring directional coupler mounted in the visual transmission line and the rf converter unit associated with the analyzer. In addition, ensure that the video output of the analyzer is connected to video input jack J201 on the exciter/modulator.



**Fig. 11-14. Tuning of visual klystron.**

Courtesy RCA

3. Insert the bar in switch 4S6 and remove the bar from switch 4S7. Remove the three plugs in the front of the magnet assembly and the plug in the plate immediately below the magnet assembly, to gain access to the cavity tuning sockets.
4. Operate the following circuit breakers and switches to the on position:

<i>Item</i>	<i>Symbol</i>
Main circuit breaker	5S1
Knife switch	4S1
PA COOLING switch	1S4
MAGNET switch	1S5
CONTROL switch	1S6
EXCITER switch	1S7
LOW VOLTAGE switch	1S8

5. In addition, operate HIGH VOLTAGE HIGH switch 4S2A to off and HIGH VOLTAGE LOW switch 4S2B to on. Then depress the TRANSMITTER ON push button. Set visual attenuator 1A1 to  $\frac{2}{3}$  of maximum attenuation. In addition, make sure that the fm exciter and the exciter/modulator are operating.

CAUTION: Prior to applying high voltage to the visual klystron, ensure that the vapor cooling system is operating properly. In addition, make sure that the rf (dummy) load is correctly installed, and that a sufficient amount of cooling water is being circulated through the rf load.

6. Set the limit needle (red line) on BODY meter 2M3 at 150 milliamperes. When the TRANSMITTER READY indicator lights, depress the HIGH VOLTAGE ON push button. Maintaining the magnet current at the center of the current range, adjust the lower and upper limit needles (red lines) on MAGNET meter 2M4 for minimum current indication on the BODY meter. Minimum current should occur between 25 and 30 amperes of magnet current.
7. Observe the output response on the sideband response analyzer. If the response is nearly correct and a substantial power output is apparent, depress the HIGH VOLTAGE OFF button. Operate the HIGH VOLTAGE LOW switch to off and the HIGH VOLTAGE HIGH switch to on. Then reapply high voltage to the transmitter.
8. Perform tests for response, gain, and frequency response versus brightness level on the visual klystron as follows:
  - A. To test response, operate the AC/DC switch on the exciter/modulator to the ac position. Adjust the SWEEP WIDTH control on the sideband analyzer for +6.0 MHz and the VIDEO SWEEP OUTPUT control for 20-percent modulation. Observe the output of the visual klystron as monitored by the sideband analyzer. The output response should be flat between -1.0 and +4.5 MHz at the mid-characteristic level (7.5-kilowatt output).
  - B. To test the gain, set the VIDEO SWEEP OUTPUT control to zero. Adjust the PEDESTAL LEVEL control until the rf wattmeter

connected to the rf load indicates 17 kilowatts. The power input to the klystron should be nominally 2.5 watts, but no more than 5.0 watts as indicated on the ThruLine wattmeter.

- C. To test frequency response versus brightness level, readjust the VIDEO SWEEP OUTPUT control for 20-percent sweep-modulation input. Observe the variation of frequency response at the 22.5-percent and the 67.5-percent brightness levels referenced to mid-characteristic. At each brightness level, readjust the VIDEO GAIN control to maintain the 20-percent modulation level. These levels should not exceed  $\pm 1.5$  dB from visual carrier minus 0.75 MHz to visual carrier plus 4.18 MHz.

If acceptable results were not obtained for any one of the three tests, recheck the output of the exciter/modulator and visual driver, and then perform the above tests. If acceptable results still cannot be obtained, tune the klystron as follows:

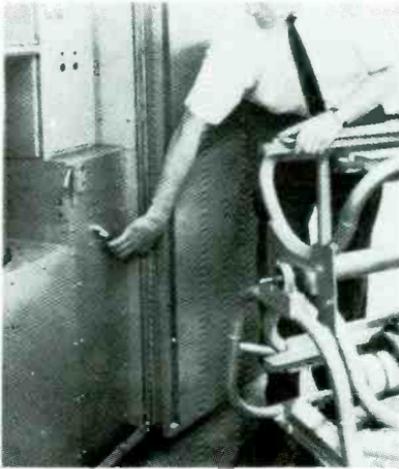
**NOTE:** When tuning the klystron cavities, rotate the tuning tool clockwise to increase frequency and counterclockwise to decrease frequency.

Set the PEDESTAL LEVEL controls for a 2.5-watt indication on the ThruLine wattmeter. Using the klystron tuning tool, tune the fourth cavity at the carrier frequency and the first cavity at approximately 2.0 MHz above the carrier frequency. Note that when tuning the second and third cavities, care should be exercised not to exceed 20 kilowatts output with a 2.5-watt input to the klystron. Normally the second cavity should be tuned at 4 to 6 MHz below the carrier frequency, and the third cavity should be tuned at 7 to 9 MHz above the carrier frequency. Proper tuning should be approached from outside of these tuning points (e.g., start cavity 2 lower than minus 6 MHz and cavity 3 higher than plus 9 MHz). The second and third cavities should be alternately tuned in small increments, until 2.5 watts input produces 17 kilowatts of output power. Then check to ensure that 20-percent sweep modulation remains flat within  $\pm 1.5$  dB when the PEDESTAL LEVEL control is set to produce power-output levels of 13.7, 7.5, and 1.5 kilowatts. It may be necessary to retune the first cavity near the upper sideband and the third cavity higher to reduce the resulting rise in high-frequency response, in order to achieve optimum results in frequency response versus brightness levels. If the frequency response deteriorates near the 1.5-kilowatt output level, recheck the exciter/modulator setup with respect to the drive applied to the visual mixer.

To set up the transmitter for the correct visual output power, proceed as follows:

1. Disconnect the video output of the sideband analyzer from video input jack J201 on the exciter/modulator; then connect station sync to this jack. Insert the monitoring diode into the 20-dB connector of directional coupler 1Z7. Connect the output of the monitoring diode to the dc-coupled input of an oscilloscope, terminated in 75 ohms at the oscilloscope.
2. Operate HIGH VOLTAGE HIGH switch 4S2A and HIGH VOLTAGE LOW switch 4S2B to off. Depress the TRANSMITTER ON push button. Then operate the AC/DC switch to the dc position and set the linearity-corrector switch to on.
3. When the TRANSMITTER READY indicator lights, depress the HIGH VOLTAGE ON push button. Adjust the level of the sync input to 40 percent as indicated on the oscilloscope connected to the monitoring diode. After the sync level has been adjusted, depress the HIGH VOLTAGE OFF push button.
4. Connect the monitoring diode to the directional coupler preceding the rf (dummy) load. Ensure that excessive rf will not be applied to the monitoring diode, since permanent damage to the diode could result. Operate the HIGH VOLTAGE HIGH switch to on, and reapply high voltage to the transmitter. Connect the oscilloscope to the monitoring diode and monitor the demodulated signal in the transmission line.
5. Increase the visual excitation by operating the VISUAL EXCITATION RAISE push button until the rf wattmeter indicates 19 kilowatts or the blanking level is at 75-percent modulation. If 19 kilowatts of power is obtained before 75-percent blanking-level modulation occurs, reduce the sync stretch in the exciter/modulator by varying the extreme left-hand LINEARITY potentiometer, to establish the proper peak of sync level.
6. If 40-percent sync is being applied to the klystron and the 75-percent blanking level is achieved before 19 kilowatts of power is indicated on the rf wattmeter, it will be necessary to increase the klystron beam current and/or voltage. Slight increases can be made by raising the output of the induction voltage regulators by a small amount. If it appears necessary to exceed 90 kilowatts of input power, the power metering devices should be recalibrated and the klystron tuning procedure carefully repeated.

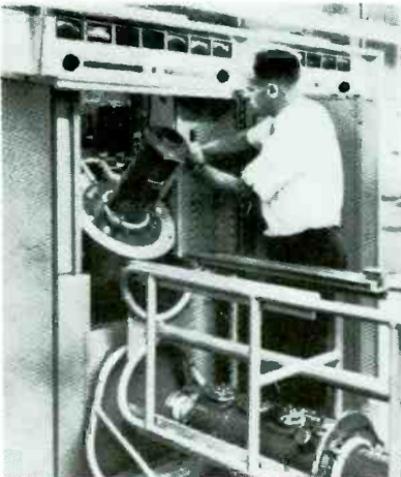
This completes the tuning procedure for the visual klystron. Depress the HIGH VOLTAGE OFF push button, and remove power from the transmitter by depressing the TRANSMITTER OFF push button.



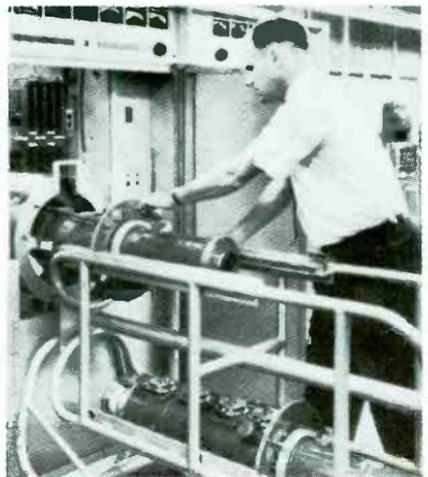
(A) *Open water drain.*



(B) *Disconnect inlet.*

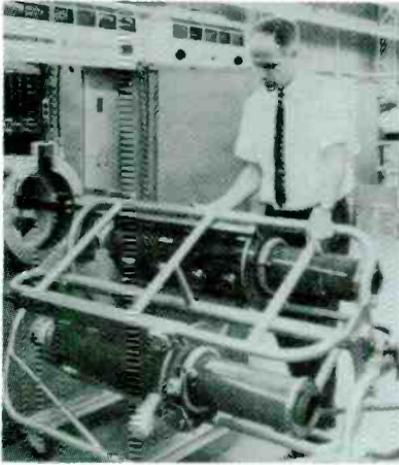


(C) *Tilt klystron out.*

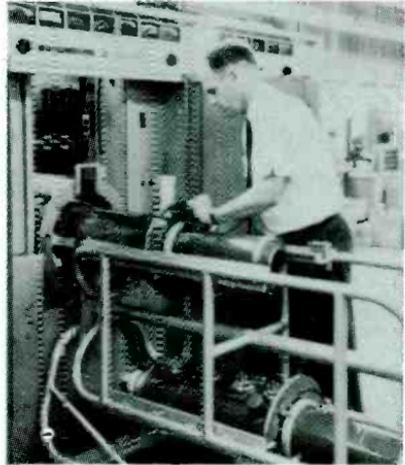


(D) *Roll klystron out.*

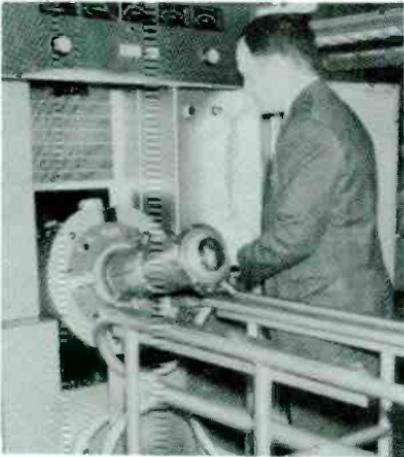
**Fig. 11-15. Procedure for**



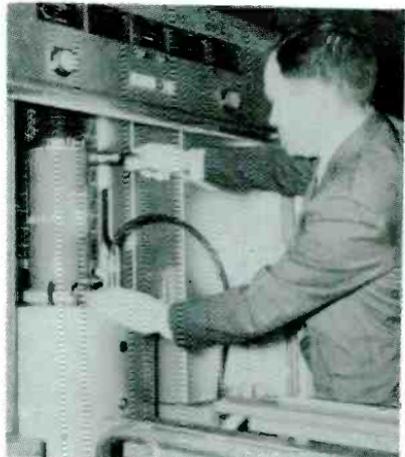
(E) Tumble carriage.



(F) Roll klystron in.



(G) Tilt klystron into transmitter.



(H) Reconnect water, close drain.

changing a klystron.

Courtesy RCA

NOTE: If the licensed station power is less than 30 kilowatts, the absolute values specified in the preceding instructions for measuring peak of sync and test power levels must be reduced accordingly.

## 11-5. HOW TO CHANGE A KLYSTRON

We should not leave the subject of the uhf klystron without describing the procedure of klystron replacement. The subject is both interesting and informative for the student and the many television technicians who have been concerned primarily with the more conventional tubes used in vhf transmitters.

The aural and visual amplifiers of the RCA Type TTU-30A transmitter each use a vapor-cooled integral-cavity klystron of the Varian Type VA-890 series. The high-power klystrons may be easily installed by one operator. The factory-tuned klystron is transferred in a horizontal position directly from the shipping carriage. By means of a built-in loading device, the klystron can then be easily installed in the transmitter from the klystron carriage. No unusual ceiling height is required since the klystron remains in a horizontal position until it has been completely installed in the transmitter. It is then tilted into a vertical position by a device which is an integral part of the transmitter. Further, factory pretuning eliminates the station-site preparation required by external-cavity designs. Fig. 11-15 illustrates the step-by-step procedure for changing a klystron.

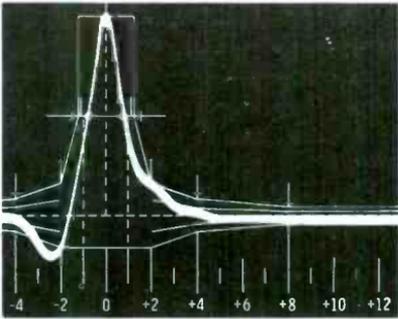
## 11-6. ENVELOPE-DELAY ADJUSTMENT WITH THE SIN<sup>2</sup> PULSE

Adjustment of color precorrection circuitry by means of a square wave was described in Chapter 10. We will cover here how to use the sine-squared pulse for these adjustments. In either case, a standard tv demodulator is required.

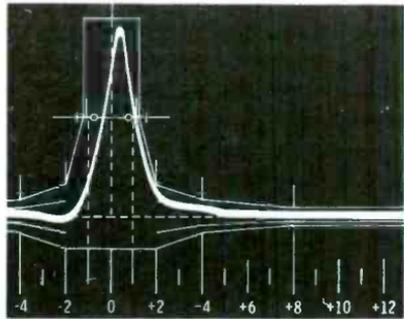
1. See Fig. 11-16A. This is a 2T pulse (most convenient to adjust low-frequency delay) with no transmitter compensation. The demodulator sound notch is out, and the receiver-delay equalizer is out. Adjust the low-frequency delay unit to minimize the anticipatory overshoot at the leading edge of the pulse (Fig. 11-16B).

NOTE: The markings along the bottom of the graticule are in unit intervals of T.

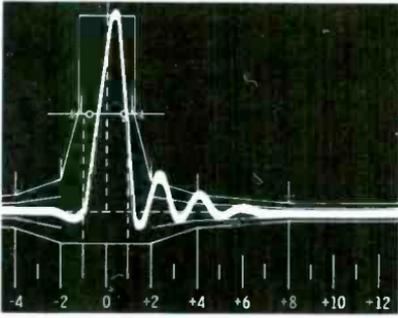
2. The waveform in Fig. 11-16C is the T pulse (twice the frequency spectrum of the 2T pulse) without high-frequency phase equalization. Adjust the high-frequency delay (variable)



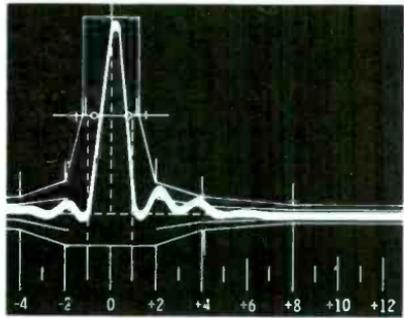
(A) 2T pulse, no phase compensation.



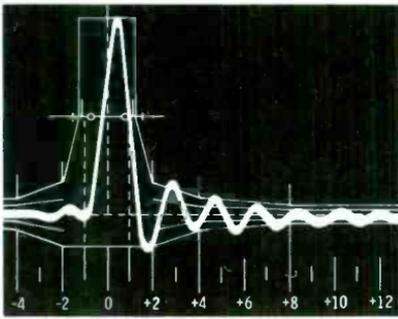
(B) 2T pulse with phase compensation.



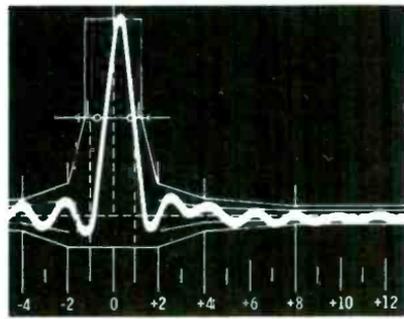
(C) T pulse, no phase compensation.



(D) T pulse with phase compensation.



(E) T pulse, demodulator aural trap.



(F) T pulse, receiver equalizer.

Fig. 11-16. Use of  $\sin^2$  pulse in envelope-delay adjustment.

and switch the fixed unit in and out to obtain minimum amplitudes and best possible distribution of ringing. Fig. 11-16D illustrates satisfactory phase compensation.

3. Switching the demodulator aural notch back in will result in the waveform of Fig. 11-16E. Now switch the receiver-delay equalizer (fixed) in. Fig. 11-16F shows the result.
4. Repeat the above procedures until satisfactory results are obtained. The transmitter should be properly broadbanded (Section 11-3) prior to color-equalization (envelope-delay) adjustments.

NOTE: The modulated 20T pulse is a sensitive indicator of envelope delay. Review Figs. 8-27, 8-28, and 8-29 and related text.

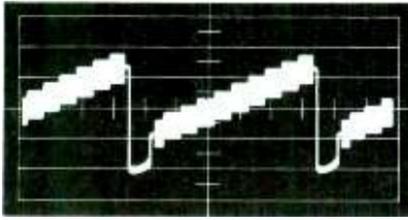
## 11-7. AMPLITUDE AND PHASE LINEARITY

Current FCC rules simply state that for color transmission the amplitude linearity "shall be substantially linear." The industry has generally established a maximum allowable nonlinearity of 10 percent. It must be realized that this degree of nonlinearity is an overall amount and cannot all occur, for example, in either the white or sync region.

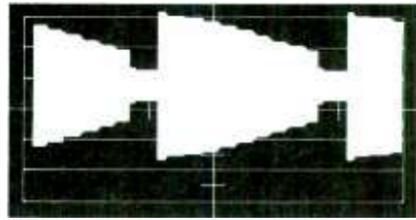
Amplitude linearity of the transmitter is measured with the stair-step signal, as previously described (Chapter 8). However, a special problem which calls for a modified procedure exists at the transmitter.

Visual-modulator transfer characteristics inherently compress in the white direction and sometimes (simultaneously) in the sync direction. Adjustments in the transmitter itself, or in an external processing amplifier, are provided to predistort the signal so that the overall transfer curve is linear.

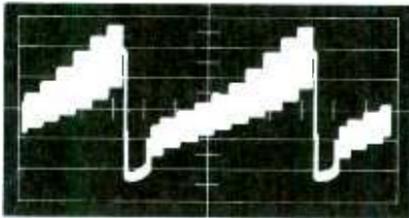
Transmitter nonlinearities are compensated in the video before modulation onto the carrier. Fig. 11-17 serves as a review of the basic principles. Linear video as in Fig. 11-17A normally results in white compression, which is particularly apparent on the superimposed 3.58-MHz (color subcarrier) component (Fig. 11-17B). A waveform similar to that in Fig. 11-17C is necessary to result in a minimum of differential gain at 3.58 MHz (Fig. 11-17D) and a linear step response. NOTE: Always use full modulation in this test. This means the top of the 3.58-MHz component on the whitest step should be adjusted, as monitored with the chopper, to an amplitude where 10 percent of the carrier remains. (Some stations use 15-percent minimum carrier, which is acceptable.) *Caution:* When adjusting the degree of white stretch, be sure to maintain reference full modulation.



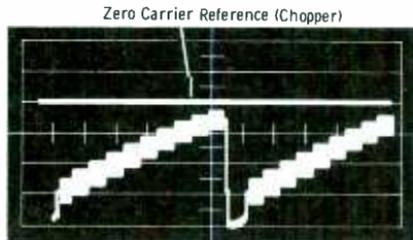
(A) *Stabilizing-amplifier output, white stretch off.*



(B) *Demodulator output (3.58 MHz), no white stretch.*



(C) *Stabilizing-amplifier output, white stretch on.*



(D) *Demodulator output with white stretch on.*

**Fig. 11-17. Compensation for transmitter nonlinearity.**

Then check the transmitter differential phase at 3.58 MHz. It is usually necessary to employ predistortion for this effect also. Differential phase results from parallel impedance paths, which simply means that the loading of circuits (even in some distribution amplifiers) can vary somewhat depending on amplitudes, particularly at the higher frequencies. Predistortion for differential phase is accomplished by employing nonlinear impedances prior to modulation; the nonlinearity is opposite to that of the transmitter characteristics, and normally is adjusted in the same stabilizing or processing amplifier that provides white stretch.

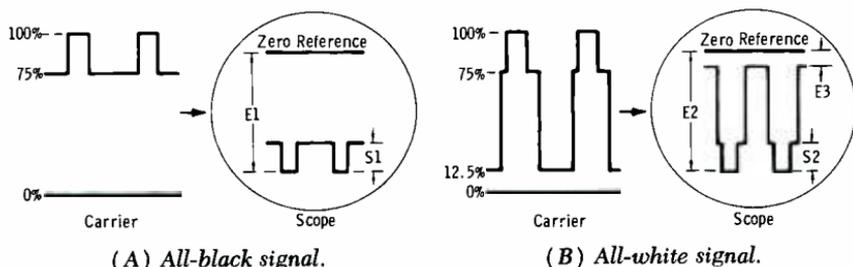
The techniques of measurement are the same as those already covered in Chapter 8, with the added function of the precorrecting circuits. Modern transmitter installations are capable of being adjusted with little difficulty to well within 3-percent differential gain and  $3^\circ$  of differential phase at 3.58 MHz.

## 11-8. MEASURING WHITE-TO-BLACK REGULATION

For waveform monitoring, the output of the demodulator is fed to a keyer circuit and then to an oscilloscope. The purpose of the keyer circuit (also termed vibrator or chopper) is to short-circuit the output of the detector intermittently, providing an additional

line on the scope screen to represent zero output. The dc axis must be constant; the ac video-signal axis is variable, depending on light or shade in the original scene. Periodic shorting of the demodulator produces a zero reference level representing no signal.

Fig. 11-18 illustrates the application of such monitoring to the measurement of modulation characteristics. If an all-black video signal with a sync-pulse height  $S_1$  above pedestal level is fed into



**Fig. 11-18. Measurement of modulation characteristics and transmitter regulation, white to black.**

the transmitter, the resultant scope pattern is as shown in Fig. 11-18A. The ratio of  $S_1$  to  $E_1$  is an expression of the *modulation capability* of the transmitter for an all-black signal, with respect to the sync pulses. If the transmitter is left adjusted as before and an all-white signal is fed to the transmitter input, the scope pattern appears as in Fig. 11-18B. The ratio of  $E_3$  to  $E_2$  is an expression of transmitter modulation capability for an all-white signal, with respect to the sync pulses. For a properly adjusted transmitter, these ratios should be practically equal. In other words, the variations of blanking and sync levels with changes in picture brightness from black to white must be held to an absolute minimum. The FCC standards limit this variation to 10 percent of the amplitude of an all-black picture. When functioning properly, modern transmitters hold well within 3 percent in going from black to white.

The percent of variations under such conditions may be determined as follows:

$$\text{Blanking-level variations} = \frac{(E_2 - S_2) - (E_1 - S_1)}{E_1 - S_1} \times 100\%$$

$$\text{Sync-level variations} = \frac{E_2 - E_1}{E_1} \times 100\%$$

The preceding arrangement also enables the operator to set the maximum white level of the video signal to 12.5 percent ( $\pm 2.5$  percent) of the peak sync amplitude.

### 11-9. ASSEMBLING DATA ON PROOF OF PERFORMANCE (VISUAL)

All exhibits should be dated and signed by the engineer making the measurements. The chief engineer then completes the proper engineering reports on the FCC form for license renewal from the data, and attaches the exhibits to the forms.

*Exhibit 1.* Draw a block diagram of the test-signal path. Record the type, manufacturer, and serial number of each unit employed in the measurement.

*Exhibit 2.* Tabulate the frequency response (upper and lower sidebands).

*Exhibit 3.* Plot the data of Exhibit 2 on suitable graphs. This should include a plot on the ideal diode-response curve for the upper sideband. The lower-sideband response can be plotted on suitable linear graph paper.

*Exhibit 4.* Tabulate the pertinent data as in Fig. 11-19. Regulation measurements were described in Section 11-8. The video signal-to-noise ratio is measured as described in Chapter 9 for the stl. NOTE: The signal-to-noise ratio at the transmitter output can normally be considered to better than that obtained on the stl alone. This

<b>EXHIBIT 4</b>	
(With typical measurements at WTAE-TV transmitter)	
REGULATION: Black:	100
White:	103
DIFF. GAIN (3.58 MHz):	3%
DIFF. PHASE (3.58 MHz), 50% APL:	2°
VIDEO SIGNAL/NOISE RATIO:	40 dB
FINAL PLATE VOLTS:	3.65 kV
FINAL PLATE CURRENT:	8.2 A, Black Level
INDICATED POWER IN DUMMY LOAD:	13.2 kW, Black Level
FREQUENCY METER READING:	+100 Hz
Latest Calibration Date:	_____
By:	_____
Date:	_____
Engineer:	_____

Fig. 11-19. Example of tabulation of proof-of-performance data.

is because the stl is measured with a wideband scope response. While this is also true of the main transmitter, the attenuation characteristic (sharp rolloff above 4.18 MHz) removes the higher-frequency noise measured directly at the stl output.

*Exhibit 5.* Take a photograph of the transmitted test-pattern signal from the face of a station monitor. This is normally required on the initial proof only, but it is a good practice to include this and the following photos on each proof.

*Exhibit 6.* Photograph the 100-kHz square-wave response or 20T-pulse response (color installations only). Or, tabulate envelope delay versus frequency if direct measurements can be taken.

*Exhibit 7.* Photograph the video waveform (after vsbf) at normal line rate, showing 7.5-percent setup and full modulation.

*Exhibit 8.* Photograph (on an expanded scale) the horizontal-sync interval with 0.005H markers.

## 11-10. FCC RULES FOR AURAL-TRANSMITTER PROOF OF PERFORMANCE

Always keep abreast of changes in the FCC Rules. The following Rules were applicable at the time of writing:

1. The transmitter shall operate satisfactorily with a frequency swing of  $\pm 25$  kHz, which is considered 100-percent modulation. It is recommended, however, that the transmitter be designed to operate satisfactorily with a frequency swing of at least  $\pm 40$  kHz.
2. The transmitting system (from input terminals of microphone preamplifier, through audio facilities at the studio, through telephone lines or other circuits between studio and transmitter, through audio facilities at the transmitter, and through the transmitter, but excluding equalizers for the correction of deficiencies in microphone response) shall be capable of transmitting a band of frequencies from 50 to 15,000 Hz. Pre-emphasis shall be employed in accordance with the impedance-frequency characteristic of a series inductance-resistance network having a time constant of 75 microseconds. The deviation of the system response from the standard pre-emphasis curve shall lie between two limits. The upper of these limits shall be uniform (no deviation) from 50 to 15,000 Hz. The lower limit shall be uniform from 100 to 7500 Hz, and 3 dB below the upper limit; from 100 to 50 Hz the lower limit shall fall from the 3-dB limit at a uniform rate of 1 dB per octave (4 dB at 50 Hz); from 7500 to 15,000 Hz the lower limit shall

fall from the 3-dB limit at a uniform rate of 2 dB per octave (5 dB at 15,000 Hz).

3. At any modulating frequency between 50 and 15,000 Hz and at modulation percentages of 25, 50, and 100 percent, the combined audio-frequency harmonics measured in the output of the system shall not exceed the rms values given in Table 11-2.

**Table 11-2. Modulation Frequency Versus Distortion Percentage**

Modulation Frequency	Distortion Percent
50 to 100 Hz	3.5
100 to 7500 Hz	2.5
7500 to 15,000 Hz	3.0

Measurement shall be made employing 75-microsecond de-emphasis in the measuring equipment and 75-microsecond pre-emphasis in the transmitting equipment, and without compression if a compression amplifier is employed. Harmonics shall be included to 30 kHz.

NOTE: Measurements of distortion using de-emphasis in the measuring equipment are not practical at the present time for the range 7500 to 15,000 Hz for 25- and 50-percent modulation. Therefore, measurements should be made at 100-percent modulation and on at least the following modulating frequencies: 50, 100, 400, 1000, 5000, 10,000, and 15,000 Hz. At 25- and 50-percent modulation, measurements should be made on at least the following modulating frequencies: 50, 100, 400, 1000, and 5000 Hz.

It is recommended that none of the three main divisions of the system (transmitter, studio-to-transmitter circuit, and studio facilities) contribute over one-half of these percentages, since at some frequencies the total distortion may become the arithmetic sum of the distortions of the divisions.

4. The transmitting-system output noise level (frequency modulation) in the band of 50 to 15,000 Hz shall be at least 55 dB below the audio-frequency level representing a frequency swing of  $\pm 25$  kHz.

NOTE: For the purpose of these measurements, the visual transmitter should be inoperative, since the exact amount of noise permissible from that source is not known at this time.

5. The transmitting-system output noise level (amplitude modulation) in the band of 50 to 15,000 Hz shall be at least 50 dB below the level representing 100-percent amplitude modulation.

NOTE: For the purpose of these measurements, the visual transmitter should be inoperative, since the exact amount of noise permissible from that source is not known at this time.

6. If a limiting or compression amplifier is employed, precaution should be maintained in its connection in the circuit due to the use of pre-emphasis in the transmitting system.

When an stl is included in the audio measurements by sound duplexing, follow the procedure described in Chapter 9 for checking the stl. Note that in measuring the overall system, which includes the main tv aural transmitter, the gain must be dropped at the studio by 20 dB when the 1-kHz frequency is reached. In order to restore reference modulation at the transmitter, the gain must be increased at the transmitting location by this amount. This is the new 0-dB reference level.

The aural frequency response should be tabulated and plotted within the pre-emphasis curve, as outlined later. (Note from the FCC Rules that the visual transmitter should be off for tv aural-transmitter measurements.) Record the aural transmitter final-stage plate voltage and plate current. Tabulate the noise and distortion measurements (also as outlined later in this chapter). Date and sign the forms. Chart 11-2 summarizes the visual and aural proof-of-performance methods.

## 11-11. AURAL-TRANSMITTER PROOF-OF-PERFORMANCE TECHNIQUES

Review Part 3 of Chapter 8 before proceeding. Be sure you understand the description of the audio signal generator illustrated in Fig. 8-48. Fig. 8-48 shows a typical audio-oscillator feed used for a broadcast proof of performance. Remember that this proof requires the signal-generator output to feed a studio microphone-preamplifier input.

Fig. 11-20 shows a block diagram of a typical noise and distortion meter, which is also used to check frequency response. In measuring harmonic distortion, the following takes place:

- A. The amplitude of a single-frequency sine wave is measured.
- B. A tuning circuit is adjusted to suppress the fundamental frequency. This is done by a sharply tuned circuit and bridge control to achieve at least 80 dB of suppression at the fundamental frequency.
- C. The remaining measured amplitude is the total harmonic distortion.

### Chart 11-2. Summary of Station Proof-of-Performance Measurements

(After vsbf with transmitter operating into dummy load)

Visual	Aural
<p>1. CARRIER FREQUENCY (By station monitor; with notation of latest calibration, date, and by whom made.)</p> <p>2. RF OUTPUT POWER (PEAK) WITH PEDESTAL ADJUSTED TO EXACTLY 75% OF PEAK CARRIER (Reflectometer should be calibrated at least once each 6 months. Make appropriate notation on log.)</p> <p>3. TRANSMITTER REGULATION, BLACK-TO-WHITE</p> <p>4. SIDEBAND RESPONSE AND ATTENUATION CHARACTERISTIC (If calibrated receiver is available, run measurements on point-to-point basis with single-frequency sine waves and sync combined. Or use sideband analyzer. Photos of cro display should be filed.)</p> <p>5. SIGNAL/NOISE RATIO AND SIGNAL/HUM RATIO</p> <p>6. LINEARITY AT FULL MODULATION</p> <p>7. DIFFERENTIAL GAIN (3.58 MHz) AT FULL MODULATION</p> <p>8. DIFFERENTIAL PHASE (3.58 MHz) AT FULL MODULATION</p> <p>9. PHOTOS OF: Test Pattern from Monitor 1/2 line-rate sweep of above on cro 1/2 field-rate sweep of above on cro Expanded horizontal-blanking interval with 0.005H markers</p>	<p>1. CARRIER FREQUENCY (By station monitor; with notation of latest calibration, date, and by whom made.)</p> <p>2. RF OUTPUT POWER (rms) <math>E_p \times I_p \times F</math> where: <math>E_p</math> is equal to plate volts of final stage, <math>I_p</math> is equal to plate current of final stage, <math>F</math> is equal to efficiency factor supplied by manufacturer.</p> <p>THE FOLLOWING TO BE MADE WITH VISUAL TRANSMITTER OFF:</p> <p>3. SIGNAL/NOISE RATIO (A) FM (B) AM</p> <p>4. AUDIO FREQUENCY RESPONSE (Measuring input level required to hold reference modulation percentage.)</p> <p>5. HARMONIC DISTORTION (50-15,000 Hz. Distortion meter must measure harmonics to 30 kHz.)</p>

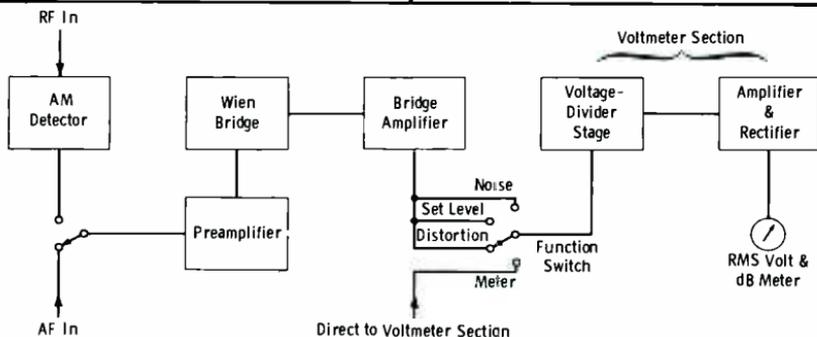


Fig. 11-20. Block diagram of Hewlett-Packard noise-distortion meter.

Normally the transmitter output measurement is taken from a special output of the modulation monitor designed for this purpose. However, an am detector may also be provided as shown in Fig. 11-20. The use of this in fm measurements is described later.

Always be sure of the back-to-back characteristics of your audio oscillator and noise-distortion (N-D) meter before taking complete proofs or after tube or transistor changes or other servicing of the measuring equipment. The direct combination of these instruments should indicate a frequency response flat within 1 dB from 30 to 15,000 Hz. The distortion should measure less than 0.2 percent from 60 Hz to 20 kHz and not more than 0.35 percent from 20 to 50 Hz. The noise should measure at least 70 dB below the output level of the audio oscillator. If it does not, substitute new tubes one at a time (allowing good warmup time) or follow the manufacturer's instructions for the adjustments necessary to bring the equipment within the above specifications or the specifications for the particular type used.

For frequency-response runs, always consider the back-to-back response of the measuring equipment in calibration of the readings for the input level at the studio. For example, if the back-to-back response is down 2 dB at 5000 Hz relative to 1000 Hz (reference frequency), then feed an input two decibels higher to the studio microphone input at this frequency.

Fig. 11-21 shows a typical setup for a complete aural proof-of-performance run. Remember that the terminal gear could be part of an stl rather than Telco line equipment. We must again emphasize that when stl's with diplexed sound are used, follow the procedure for aural runs described in Chapter 9. This procedure is modified

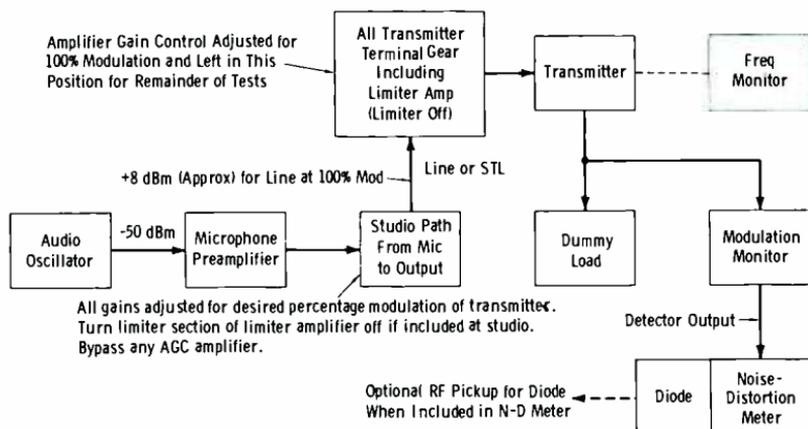


Fig. 11-21. Setup for aural proof of performance.

somewhat from that for normal landline runs due to stl transmitter sound pre-emphasis, which is in addition to that of the regular tv aural transmitter.

### Frequency-Response Runs

A frequency-response run includes the following steps:

1. With the audio oscillator set on 1000 Hz, feed  $-50$  dBm to a studio microphone input (Fig. 11-21). If the oscillator meter indicates zero dBm at the adjusted gain, the calibrated attenuators will total 50 dB. Usually the oscillator meter has a full-scale reading of  $+15$  to  $+17$  dBm. It is advisable to adjust the oscillator gain to obtain, say,  $+10$  dBm (to obtain a midscale reading) and set the attenuators for a total of 60 dB to result in a  $-50$ -dBm output. Usually the back-to-back measurement previously described gives the best results with this type of operation, particularly when distortion and noise measurements are taken. Usually there are three sets of calibrated attenuators—tens, units, and tenths—so that precise readings may be made.
2. Adjust the associated faders and master gain control for reference line output (an indication of zero on the studio-line VU meter). This is actually  $+4$  to  $+12$  VU. All faders and the master gain control should be in approximately the normal operating positions. The term “approximately” must be used, since the microphone-fader setting will in most cases actually be somewhat higher because of the high peak factor of speech waves compared to a sine wave.
3. Be sure that any agc amplifier is bypassed (patched around). Also be sure that if a limiting amplifier is employed, the limiter section is turned off.
4. At the transmitter, adjust the line or limiter amplifier to obtain 100-percent modulation. The limiter section of the limiter amplifier must be off.
5. Record the oscillator attenuator reading. Refer to Fig. 11-22 for an example of tabulated response data. Note that the attenuator reading for 100-percent modulation at 1000 Hz is simply recorded along the entire top row. Assume this to be 60 for illustrative purposes. Copy the figure 60 also in row 2 for 1000 Hz.
6. Tune the oscillator to 50 Hz and readjust the attenuators (if necessary) to again obtain 100-percent modulation. Record the new attenuation figure in row 2 under 50 Hz. Repeat this procedure for the other frequencies listed.
7. Fill in row 3 by subtracting the readings in row 2 from those in row 1. This is a record of the response variation. Note that, for

## ..... 100% MODULATION

Hz	50	100	400	1000	5000	10,000	15,000
(1)	60	60	60	60	60	60	60
(2)	57	58	59	60	69	74	76
(3)	+3	+2	+1	0	-9	-14	-16

- (1) Record attenuator reading for the 1000-Hz reference signal in this row.  
 (2) Record the attenuator readings for the specified frequencies in this row.  
 (3) Record the af response variations, which are obtained by subtracting row (2) from row (1), in this row. These final figures are to be used in plotting the graphs.  
 NOTE: The figures inserted in this example indicate pre-emphasis measurement.

**Fig. 11-22. Example of data from frequency-response check of television aural transmitter.**

example, if it was necessary to reduce the attenuation 3 dB at 50 Hz, this indicates that a level 3 dB *higher* was required at 50 Hz relative to the level at 1000 Hz to obtain 100-percent modulation. Therefore the response of the system is *down* 3 dB at this frequency relative to the response at the reference frequency, and is so plotted on the graph (Fig. 11-23).

8. The entire process is repeated at the lower percentages of modulation required. Two acceptable methods may be used:
- The audio-oscillator attenuators may be raised in value to obtain the new (lower) percentage of modulation, or
  - The faders on the console may be adjusted to obtain the new percentage, the -50-dBm input to the microphone preamplifier at the reference frequency of 1000 Hz being retained. This method is preferred, to avoid extremely low-level inputs from the audio signal generator to the microphone input.

The frequency-response measurement may be taken either with or without de-emphasis. It is a good engineering practice to run this check both ways. This proves the proper transmitter pre-emphasis first and then proves that the station de-emphasis network in the monitor circuit is essentially complementary to the 75-microsecond pre-emphasis.

Fig. 11-22 is an example of tabulated data obtained from measuring the transmitter output without de-emphasis. These figures are shown plotted on the standard 75-microsecond pre-emphasis curve in Fig. 11-23. Note in this example that the original run, where the 1000-Hz reference is 0 dB, results in values for frequencies of 5000 Hz and 10,000 Hz that fall outside the tolerable limits. Note, however, that at 5000 Hz (farthest out of the range) a shift of -0.8 dB brings the measurement within the limits. Therefore, a new curve

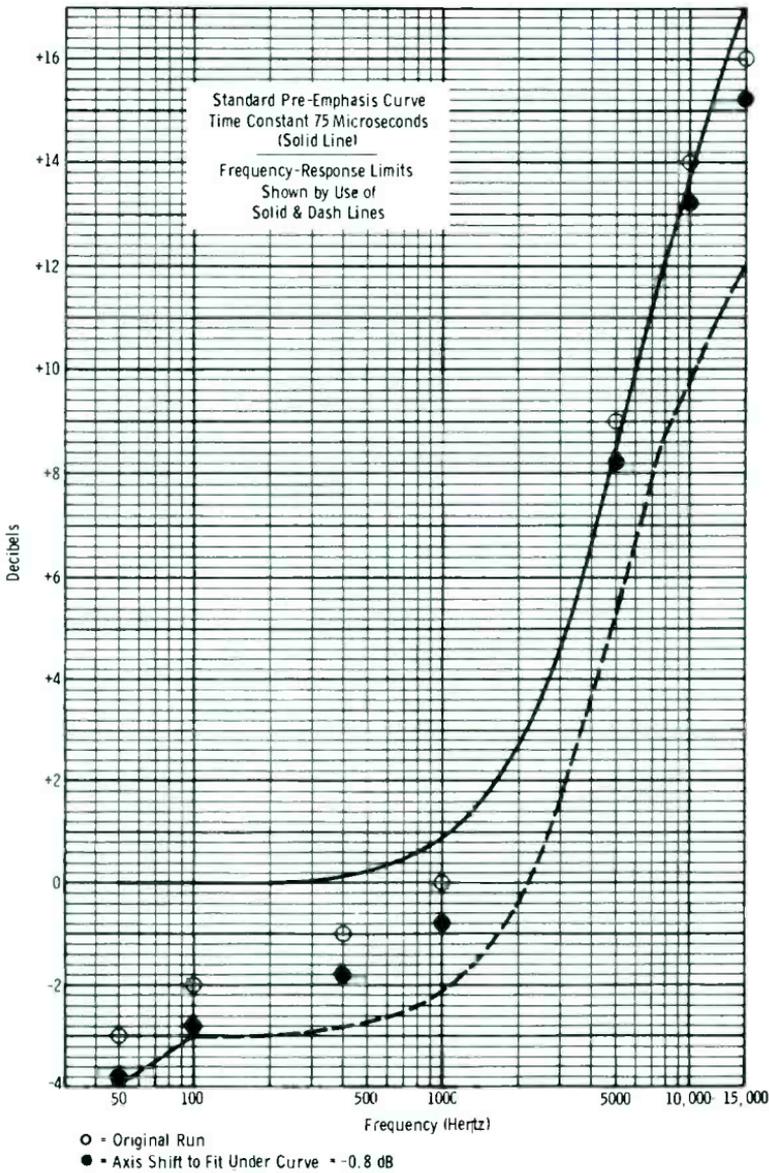


Fig. 11-23. Sample plot of frequency-response data for fm transmitter.

may be drawn (as shown by solid dots) with an axis shift of  $-0.8$  dB, and the proof is satisfactorily completed. If this shift had caused other plotted values to fall outside the limits, remedial measures would have been needed to correct the fault. Measurements must be taken at least at the frequencies specified in the top row of the

table of Fig. 11-22 and for modulation percentages of approximately 25, 50, and 100 percent.

The response is then run with standard de-emphasis, and the overall curve is drawn on standard response-curve paper.

Remember that the basic idea of a proof run is to prove that the system can be brought within specifications by the proper adjustments or servicing. A proof is a record of the performance as it exists at the time of measurement. This is no guarantee that a single stage in the system may not deteriorate the next hour, day, or week to result in different performance. However, the more often such tests are made and proper remedial steps taken to correct deficiencies, the better the overall results will be.

When a system fails to meet response specifications, first ascertain whether the trouble is at the studio, at the transmitter, or in the line. When the transmitter is at a separate location from the studio, feed the audio oscillator directly into the line. If the trouble persists, run a characteristic curve on the line (or stl) itself to isolate or determine if the result is cumulative. If the trouble is at the studio, it is a simple matter to run studio-only frequency-response checks.

### Noise Level

Between 50 and 15,000 Hz, fm hum and noise must be at least 55 dB below the level representing 100-percent modulation ( $\pm 25$  kHz deviation), and am hum and noise must be at least 50 dB below the level representing 100-percent amplitude modulation. The reference frequency is 1000 Hz, and this measurement may actually be performed in step 5 of the preceding procedure by removing the oscillator (after the reference level has been established) and terminating the microphone input with a resistor equal to the input impedance. Leave all the gains as originally set, but be sure that no other source faders are open. The noise and distortion meter is then placed in the noise position and its sensitivity increased to obtain the noise level in the unmodulated carrier.

The input stages of preamplifiers are the most common source of noise. If an oscilloscope is available, connect it to the scope terminals of the N-D meter and determine if the noise is caused by a hum component. Many amplifier or preamplifier power supplies employ "hum" pots which should be adjusted while the noise measurement is observed. Use the same trace-down procedure for hum and noise as that previously described for locating the cause of an improper frequency response.

The output noise level for fm is measured in two categories, fm noise and am noise. The fm noise includes any noise in the entire system that would result in frequency modulation of the carrier. This noise is measured in dB below the level corresponding to 100-

percent modulation, which for tv broadcasting is a frequency swing of  $\pm 25$  kHz. This measurement must be made with standard 75-microsecond de-emphasis. The indicating instrument must have ballistic characteristics similar to those of a standard VU meter.

In measuring the am noise level, it is necessary to obtain the noise level in terms of what would correspond to 100-percent amplitude modulation of the transmitter, but obviously it is not possible to amplitude modulate the fm transmitter to 100 percent. Some other means must be used to calibrate the noise meter.

The arrangement shown in Fig. 11-24 may be used for this purpose. A diode rectifier rectifies a small amount of rf energy from the output of the transmitter. A dc voltage proportional to the carrier output of the transmitter appears across the 600-ohm resistor with the switch in the rf position. Should the carrier output be amplitude modulated, there would also appear an ac voltage that would be proportional to the percentage of modulation. If the carrier were amplitude modulated 100 percent with a sine wave, this ac voltage

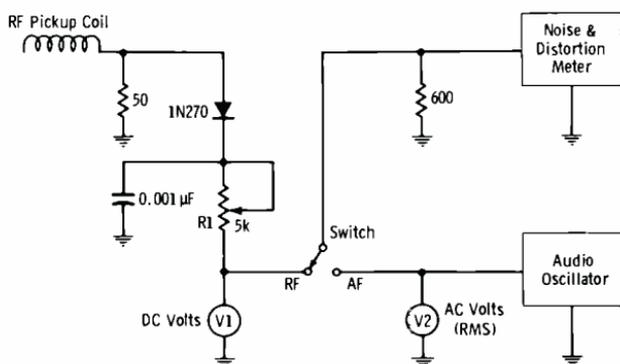


Fig. 11-24. One method of measuring am noise level of fm transmitter.

(rms) would be equal to 0.707 times the dc voltage. (The rectifier is a peak-voltage rectifier.) Thus an external calibration of the noise meter can be set up. This is done with an audio oscillator having a 600-ohm output adjusted so that there appears across the output an ac voltage equal to 0.707 times the rectified dc voltage. This voltage is fed into the noise meter, and the latter is adjusted for full-scale deflection. Thus calibrated, it is ready for use in measuring the am modulation level that appears across the output of the diode rectifier. The actual steps follow.

1. A diode rectifier is coupled to the output of the transmitter (Fig. 11-24). In some transmitters, one-half of the audio-

monitor coupling links may be used. Adjust R1 to obtain a convenient reading, such as 1 volt, on V1.

2. With the switch in the rf position, measure the dc output voltage of the diode rectifier by means of voltmeter V1.
3. Throw the switch to the af position, and adjust the output of the audio oscillator so that voltmeter V2 indicates an ac voltage equal to 0.707 times the dc voltage just measured. The noise meter is then adjusted to zero dB. (The reference is now set.)
4. Return the switch to the rf position, and read the noise level indicated by the noise meter. This is the measurement of the am noise level.

Fig. 11-25 illustrates an alternate method of measuring am noise without the conversion factor of 0.707 described previously. Only one voltmeter is required. The procedure is as follows:

1. Open S2 and close S1. Adjust the trimmer capacitor (if used) and R1 to obtain a convenient voltmeter reading, for example 2 volts.
2. Open S1 and close S2. Adjust R2 to obtain exactly the same voltage.
3. Calibrate the noise meter for 0 dB reference.
4. Open S2 and close S1. Take the noise reading according to the instructions for using the noise and distortion meter.

### Audio-Frequency Harmonic Distortion

Distortion must be measured for fundamental frequencies of 50, 100, 400, 1000, and 5000 Hz at 25, 50, and 100-percent modulation. In addition, at 100-percent modulation, distortion at the frequencies of 10,000 and 15,000 Hz must also be measured. The measurements must include harmonics to 30,000 Hz. The maximum allowable dis-

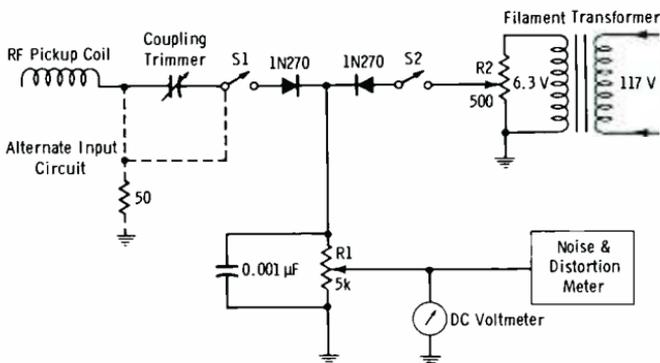


Fig. 11-25. Alternate method of measuring am noise of fm transmitter.

tortion (measured through a standard 75- $\mu$ s de-emphasis circuit) is given in Table 11-2. A typical procedure for the measurement of audio-frequency harmonic distortion is as follows:

1. The reference modulation at the reference frequency (1000 Hz) is obtained as in the frequency-response procedure. The noise and distortion meter is then placed in the set level mode, and the meter is adjusted to 100-percent calibration.
2. The noise and distortion meter is then placed in the distortion mode, and the tuning and bridge-balance controls are adjusted for minimum reading. The sensitivity is next increased in steps of 10 dB, and these controls are readjusted for minimum each time.
3. The percentage is read directly on the most sensitive scale on which it is possible to obtain a minimum reading by use of the tuning and balance controls.

Excessive harmonic distortion can be caused by faulty tubes or transistors, low power-supply voltages, or overdriving of any amplifier or chain. Again, this is a case of tracing down the source of distortion first to the transmitter, line or stl, or studio. Also, if excessive noise is present, the distortion measurement is likely to be high because of the noise level.

## 11-12. PREVENTIVE-MAINTENANCE PROCEDURES

The importance of a rigid preventive-maintenance schedule at a tv transmitter should be obvious. It remains to examine in detail the methods and procedures involved. General schedules may be outlined as follows:

### Daily Procedures

1. Throughout the operating day, in addition to recording meter readings, make daily reports on any peculiarities in meter readings, time and duration of any abnormal waveform observations, any unusual frequency-deviation readings, and any unusual temperature readings in water-cooled or vapor-cooled tubes. Record the time and indicated circuits of overloads.
2. After shutdown, investigate any of the peculiarities listed in (1).
3. Immediately after shutdown, *feel* all components such as capacitors, inductors, transformers, relays, insulators, and grid and anode connections of high-power tubes for excessive heating. Feel blower motors. Get the "feel" habit to become familiar with normal operating temperatures.

4. Should general abnormally high temperatures be revealed, check for correct cabinet temperatures, and check air filters for cleanliness. Check cabinet temperature of the air around all high-voltage rectifier tubes.
5. Check the pressure in gas or dry-air filled transmission lines.
6. Observe all components such as resistors, meter hands (for zero set), insulators, etc. Watch for blistering or discoloration on resistors. Watch all electrolytic capacitors for bulging sides or leaking insulation. Form the habit of observing along with feeling for normal appearance and operation. Cultivate the sense of smell to analyze any unusual odors.

#### **Weekly (In Addition to Daily)**

1. Carry out overall alignment procedures. This serves two purposes: to keep the operator familiar with the procedure, and to aid in observing any slight changes in stage-by-stage tuning. Realignment is absolutely necessary about every two months if optimum transmitter results are to be obtained.
2. Clean and polish all safety gaps.
3. *Dust off all surfaces.* Use a small forced-air stream in spaces not readily accessible with a rag and cleaning fluid. Thoroughly clean and polish all insulators with a rag and carbon tetrachloride and inspect closely for cracks. Clean all terminal boards, and inspect them for tightness of connections. After dusting, clean the entire transmitter with a vacuum cleaner.
4. Check blower motors, and check blower belts for proper tension. Inspect air filters and clean or replace them, if necessary. Check blower interlock switches for freedom of operation and cleanliness of contacts. See that the oil level in blower motors is correct.
5. Check all door interlocks and safety switches to be sure they operate properly.
6. Check spare crystals to be sure they will operate properly in an emergency. At the same time, check neutralization of stages by removing the crystal.
7. Calibrate the reflectometer against the dummy-load reading.

#### **Monthly (In Addition to Weekly)**

1. Remove all receiving-type tubes and test them with a good dynamic tester. Any tubes falling below 10 percent of their normal transconductance value should be replaced. Be sure to check a new tube before installing it. While the tubes are removed, thoroughly vacuum all sockets, and check for tightness of socket wiring. Examine all grid or plate caps and connections.

2. Clean all relay and contactor contacts. Watch for badly worn contacts and replace them if necessary. Clean the pole faces on contactors.
3. Clean and polish all tuned line-circuit elements and connections.
4. Clean all audio equipment, including attenuators and switching contacts.
5. Calibrate the visual and aural transmitter frequency monitors with a primary-standard frequency-measuring service.
6. Clean all monitoring equipment, including switches.
7. Clean, inspect, and check for proper operation of all automatic control equipment, such as time-delay relays, overload relays, etc.
8. If a water-cooling or vapor-cooling system is employed, check the entire system for any visible signs of leaking and for electrical leakage.

#### **Quarterly (In Addition to Monthly)**

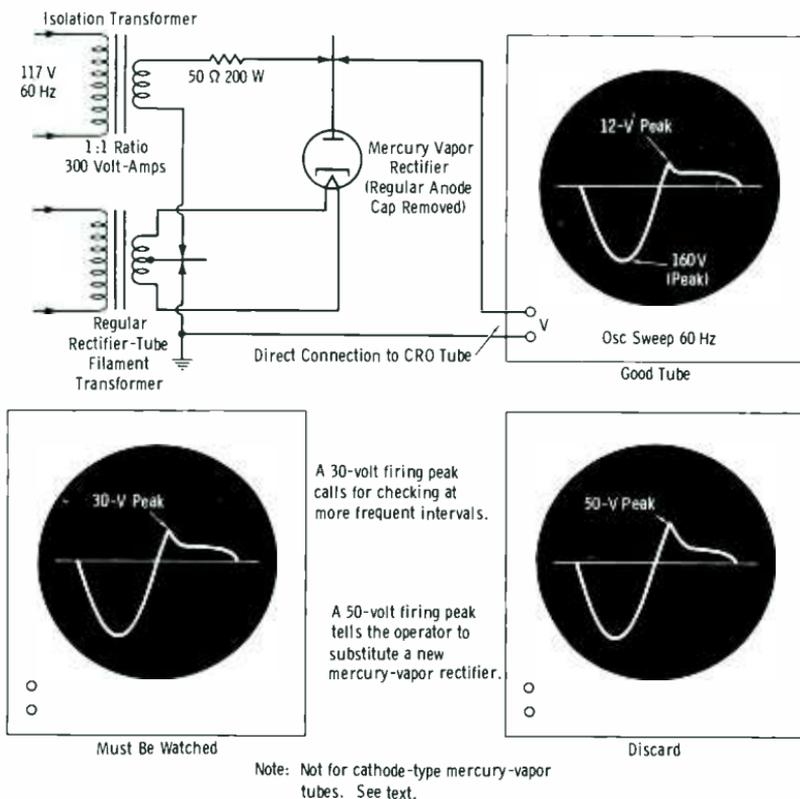
1. Check all filament voltages with an accurate voltmeter.
2. Operate all spare mercury-vapor rectifiers. Even though they have been preheated and stored upright as they should be, operate them for a 15-minute period with filament voltage only. Check for high-voltage operation and run them for several hours before storing them again in their upright containers.
3. Operate spare high-level tubes for several hours at their normal ratings to prevent formation of gas within the envelope. This also serves to double-check their operation in case of emergency.
4. Check all filter-bank surge resistors (where used) with an ohmmeter.
5. Check overall system performance as to picture resolution, waveform, and aural noise and distortion, and keep accurate records of tests. The cause of any undue deviation from normal should be located by stage isolation.

#### **Mercury-Vapor Tubes**

Mercury-vapor rectifier tubes (when used) should not be neglected in maintenance schedules. Unless proper precautions are taken, excessive lost air time will be caused by faulty rectifiers. These tubes should be observed whenever possible during each operating day. A good mercury-vapor rectifier is characterized by a healthy, clear-blue glow. A greenish-yellow color usually indicates a faulty tube or one which will soon cause trouble.

Because of the importance of foreseeing such trouble and because of the lack of familiarity of the average operator with methods for testing this type of tube, the reader should become familiar with the

maintenance procedure illustrated in Fig. 11-26. Since cathode-ray oscilloscopes are common at tv transmitter installations, the operator may conveniently use this most accurate check. An isolation transformer that has a rating of at least 300 volt-amperes should be used with a series current-limiting resistor of 50 ohms as shown. The mercury-vapor rectifier tube is left in its regular socket with its regular plate-cap connection removed. The secondary of the isolation transformer is then connected in series with the resistor to the rectifier plate, and the other lead is connected to the filament center tap. The vertical-deflection plates of the oscilloscope are connected directly across the tube in the same manner. With the scope self-synchronized with the 60-Hz power line and with power applied to the filament of the tube being checked, the scope pattern will show both the nonconducting half-cycle and the conducting half-cycle, as in Fig. 11-26. The sharp peak at the start of conduction reveals the tube condition in operation. A good tube will fire at between 10 and



**Fig. 11-26. Method of checking condition of mercury-vapor rectifier tube.**

20 volts, as indicated by the amplitude of this peak on a calibrated screen. A tube approaching the end of its useful life will require a higher firing voltage and will break into conduction later in the conducting interval. When this breakdown peak reaches 30 to 40 volts, the tube must be tested at more frequent intervals, preferably once a week. When the firing peak reaches close to 50 volts, the tube must be replaced with a new rectifier.

Always remember that mercury-vapor rectifiers must have their filaments operated at normal voltage for a minimum of 30 minutes, and then be stored *upright* to prevent the mercury from splashing back on the envelope and elements. Tubes which have been accidentally jarred must again be preheated before application of the anode potential.

*Caution:* This type of simple check is possible only for gas tubes (such as the 866 and 8008) in which the filament directly faces the plate. Certain types of tubes (such as the 673) employ an indirect (cathode) heater which provides an extra element between the filament and plate. This projection is similar in action to a grid, since it is connected to one side of the filament and is either positive or negative with respect to the plate, depending on which way the filament transformer is connected. The best way to field-test this type of tube is described in an RCA bulletin, *Pulse Method of Testing Hot-Cathode Gas Tubes*, Application Note AN-157. If this bulletin is not included in your instruction book, write to RCA and request a copy. The circuitry is more involved, but the test is worth your time.

### 11-13. PRECISE FREQUENCY CONTROL

Frequency control, frequency monitoring, and frequency checking are highly important items for the television operator and maintenance technician. It is *extremely* important to keep abreast of all current FCC Rules and Regulations, and it is the responsibility of every licensed person on duty at the transmitter to be aware of all changes that occur.

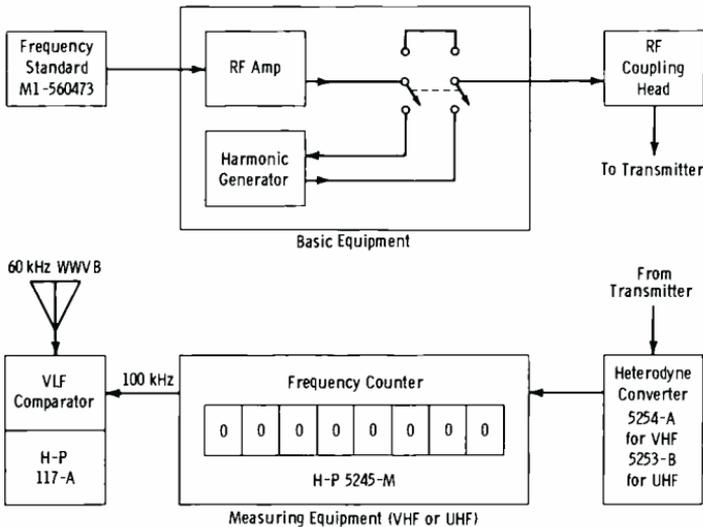
It is generally known in the television industry that interference between two or more cochannel television stations is reduced if their picture carrier frequencies are offset by a fixed amount and if the difference frequency is held constant within a very small tolerance. For optimum results, the offset frequency must be an even multiple of the frame frequency, and this offset must be held constant within plus or minus five hertz per station.

The RCA TFC-1B precise frequency control system supplies the equipment necessary to control the carrier frequency of one station with a typical stability of 4 parts in  $10^9$  over a 30-day period. The

guaranteed stability is 5 parts in  $10^{10}$  over a 24-hour period after 30 days of continuous operation. This stability adequately meets requirements for carrier stability to gain the benefits of reduction in cochannel interference.

It is desirable to be able to determine the exact frequency offset between any two participating stations at least once per month. The method would be to measure precisely the carrier frequency of each participating station with reference to National Bureau of Standards station WWVB, which operates on 60 kHz. This allows each station to make an independent and continuous check of its carrier frequency with reference to WWVB.

The block diagram of Fig. 11-27 shows the method of measurement. The basic equipment provides a precise carrier control system to replace the crystal of any RCA vhf or uhf transmitter, and can be adapted to most other makes of tv transmitters. For measuring equipment, a vhf or uhf option is available. The vhf option provides the equipment necessary to measure the carrier frequency with reference to WWVB to an accuracy of one hertz or better. It also provides a highly stable frequency counter which can be used for other rf measurements from 0.1 Hz to 500 Hz with reference to WWVB. The uhf option provides the equipment necessary to measure carrier frequency with reference to WWVB to an accuracy of one hertz or better. It also provides a highly stable frequency counter which can be used for other rf measurements from 500 MHz to 3000 MHz with reference to WWVB.



**Fig. 11-27. Precise frequency control and measurement.**

## EXERCISES

- Q11-1. What is the procedure for determining the operating power of the tv aural transmitter?
- Q11-2. If the aural transmitter must feed 1000 watts of power into the transmission line, the final-stage plate voltage used is 2700 volts, and the efficiency factor is 0.78, what plate current would you "load" into the final aural amplifier?
- Q11-3. Where do you obtain the efficiency factor (F) used in determining the aural-transmitter output power?
- Q11-4. What is the approximate value of the efficiency factor for a standard tv aural transmitter?
- Q11-5. Define the terms (A) frequency swing and (B) frequency deviation.
- Q11-6. Define the terms (A) percentage modulation and (B) 100-percent modulation.
- Q11-7. Give the procedure for determining (A) average and (B) peak power of the visual transmitter.
- Q11-8. If you find the transmitter frequency response varies with APL, what possible trouble could exist?
- Q11-9. What important precaution must you observe when adjusting white-stretch predistortion to obtain a linear transmitter video output?
- Q11-10. What should you do first when you find you cannot meet differential-gain or differential-phase specifications?



## Decibel Table

dB	Current and Voltage Ratio		Power Ratio		dB	Current and Voltage Ratio		Power Ratio	
	Gain	Loss	Gain	Loss		Gain	Loss	Gain	Loss
0.1	1.01	0.989	1.02	0.977	8.0	2.51	0.398	6.31	0.158
0.2	1.02	0.977	1.05	0.955	8.5	2.66	0.376	7.08	0.141
0.3	1.03	0.966	1.07	0.933	9.0	2.82	0.355	7.94	0.126
0.4	1.05	0.955	1.10	0.912	9.5	2.98	0.335	8.91	0.112
0.5	1.06	0.944	1.12	0.891	10.0	3.16	0.316	10.00	0.100
0.6	1.07	0.933	1.15	0.871	11.0	3.55	0.282	12.6	0.079
0.7	1.08	0.923	1.17	0.851	12.0	3.98	0.251	15.8	0.063
0.8	1.10	0.912	1.20	0.832	13.0	4.47	0.224	19.9	0.050
0.9	1.11	0.902	1.23	0.813	14.0	5.01	0.199	25.1	0.040
1.0	1.12	0.891	1.26	0.794	15.0	5.62	0.178	31.6	0.032
1.1	1.13	0.881	1.29	0.776	16.0	6.31	0.158	39.8	0.025
1.2	1.15	0.871	1.32	0.759	17.0	7.08	0.141	50.1	0.020
1.3	1.16	0.861	1.35	0.741	18.0	7.94	0.126	63.1	0.016
1.4	1.17	0.851	1.38	0.724	19.0	8.91	0.112	79.4	0.013
1.5	1.19	0.841	1.41	0.708	20.0	10.00	0.100	100.0	0.010
1.6	1.20	0.832	1.44	0.692	25.0	17.7	0.056	$3.16 \times 10^2$	$3.16 \times 10^{-2}$
1.7	1.22	0.822	1.48	0.676	30.0	31.6	0.032	$10^2$	$10^{-2}$
1.8	1.23	0.813	1.51	0.661	35.0	56.0	0.018	$3.16 \times 10^2$	$3.16 \times 10^{-3}$
1.9	1.24	0.803	1.55	0.646	40.0	100.0	0.010	$10^3$	$10^{-3}$
2.0	1.26	0.794	1.58	0.631	45.0	177.0	0.006	$3.16 \times 10^3$	$3.16 \times 10^{-4}$
2.2	1.29	0.776	1.66	0.603	50.0	316	0.003	$10^4$	$10^{-4}$
2.4	1.32	0.759	1.74	0.575	55.0	560	0.002	$3.16 \times 10^4$	$3.16 \times 10^{-5}$
2.6	1.35	0.741	1.82	0.550	60.0	1,000	0.001	$10^5$	$10^{-5}$
2.8	1.38	0.724	1.90	0.525	65.0	1,770	0.0006	$3.16 \times 10^5$	$3.16 \times 10^{-6}$
3.0	1.41	0.708	1.99	0.501	70.0	3,160	0.0003	$10^6$	$10^{-6}$
3.2	1.44	0.692	2.09	0.479	75.0	5,600	0.0002	$3.16 \times 10^6$	$3.16 \times 10^{-7}$
3.4	1.48	0.676	2.19	0.457	80.0	10,000	0.0001	$10^7$	$10^{-7}$
3.6	1.51	0.661	2.29	0.436	85.0	17,700	0.00006	$3.16 \times 10^7$	$3.16 \times 10^{-8}$
3.8	1.55	0.646	2.40	0.417	90.0	31,600	0.00003	$10^8$	$10^{-8}$
4.0	1.58	0.631	2.51	0.398	95.0	56,000	0.00002	$3.16 \times 10^8$	$3.16 \times 10^{-9}$
4.2	1.62	0.617	2.63	0.380	100.0	100,000	0.00001	$10^9$	$10^{-9}$
4.4	1.66	0.603	2.75	0.363	105.0	177,000	0.000006	$3.16 \times 10^9$	$3.16 \times 10^{-10}$
4.6	1.70	0.589	2.88	0.347	110.0	316,000	0.000003	$10^{10}$	$10^{-10}$
4.8	1.74	0.575	3.02	0.331	115.0	560,000	0.000002	$3.16 \times 10^{10}$	$3.16 \times 10^{-11}$
5.0	1.78	0.562	3.16	0.316	120.0	1,000,000	0.000001	$10^{11}$	$10^{-11}$
5.5	1.88	0.531	3.55	0.282	130.0	$3.16 \times 10^6$	$3.16 \times 10^{-7}$	$10^{12}$	$10^{-12}$
6.0	1.99	0.501	3.98	0.251	140.0	$10^7$	$10^{-7}$	$10^{13}$	$10^{-13}$
6.5	2.11	0.473	4.47	0.224	150.0	$3.16 \times 10^7$	$3.16 \times 10^{-8}$	$10^{14}$	$10^{-14}$
7.0	2.24	0.447	5.01	0.199	160.0	$10^8$	$10^{-8}$	$10^{15}$	$10^{-15}$
7.5	2.37	0.422	5.62	0.178	170.0	$3.16 \times 10^8$	$3.16 \times 10^{-9}$	$10^{16}$	$10^{-16}$
								$10^{17}$	$10^{-17}$

## Answers to Exercises

### CHAPTER 1

- A1-1.* One nanosecond (formerly called one millimicrosecond) is one thousandth of a microsecond. Therefore, to convert nanoseconds to microseconds, divide by 1000. Thus  $35 \text{ ns} = 0.035 \mu\text{s}$ .
- A1-2.* (A) 10 MHz. (B) Since one gigahertz is one thousand megahertz, to convert from megahertz to gigahertz, divide by 1000. So  $10 \text{ MHz} = 0.01 \text{ GHz}$ .
- A1-3.* Picofarad is now the accepted term for micromicrofarad, or one millionth of a microfarad. Therefore, (A)  $470 \text{ pF} = 0.00047 \mu\text{F}$ , and (B)  $470 \text{ pF} = 470 \mu\mu\text{F}$ .
- A1-4.* You don't really care about the circuitry. The important thing to know is the attenuation of the detector probe. Observe a wide-band oscilloscope display of the output of the video sweep generator (undetected), and adjust the generator output to give a convenient peak-to-peak amplitude, such as 1 volt. Use the amplitude at the 1-MHz marker. Then install the detector probe on the scope without changing either the amplitude of the signal or the scope calibration, and measure the peak-to-peak detected signal at the 1-MHz point. The amplitude will now normally be between 0.5 and 0.75 volt p-p with a 1-volt signal.
- A1-5.* You should adjust the amplitude control of the video sweep generator to obtain  $0.75 \times 0.5 = 0.375$  volt peak-to-peak.
- A1-6.* Suppose a signal is being measured at the transmitter output. If the test-oscillator frequency dial indicates 4.18 MHz, but the actual frequency is 4.25 MHz, the sharp rolloff above 4.18 MHz will indicate that the proper FCC color specifications are not met. Thus, faulty calibration of the test oscillator can cause the transmitter operator to attempt retuning, etc., which actually is not required.

- A1-7.  $4.6 \times 0.02 \times 10 = 0.92$  volt
- A1-8. No. Review Time-Difference Measurements in Section 1-4.
- A1-9. Time difference =  $(1 \mu\text{s}/\text{cm}) (3 \text{ cm}) = 3 \mu\text{s}$ .
- A1-10. (A) Duration =  $(10 \mu\text{s}/\text{cm}) (3 \text{ cm}) = 30 \mu\text{s}$ . (B) The frequency is the reciprocal of the duration;  $\frac{1}{30 \mu\text{s}} = 33.333 \text{ kHz}$ .

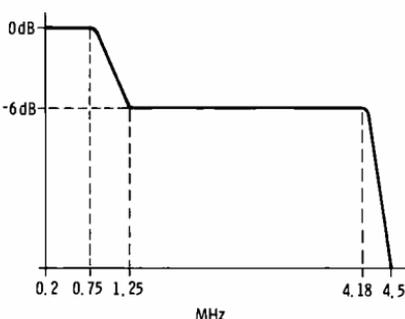
## CHAPTER 2

- A2-1. At the point of 50-percent amplitude response on a linear slope extending from minimum response at 0.75 MHz below the visual carrier to maximum response at 0.75 MHz above the visual carrier.
- A2-2. Exactly the same as for vsb, since frequencies in the lower sideband more than 0.75 MHz from the carrier frequency are severely attenuated by the receiver tuning.
- A2-3. Approximately 340 tv lines, both horizontal and vertical.
- A2-4. Yes, because the "average picture tube" in the receiver does not have a linear transfer characteristic.
- A2-5. The stairstep or sawtooth signal for luminance, and the same signal with 3.58 MHz superimposed for color.
- A2-6. The window (bar) observed at the field rate.
- A2-7. The window (bar) observed at the line rate.
- A2-8. The T or 2T pulse, or the T/2 pulse for wide-bandwidth systems.
- A2-9. The pulse and bar, through comparison of their heights.
- A2-10. The modulated 20T or 12.5T pulse and bar signal.

## CHAPTER 3

- A3-1. The purpose is to cause the rectifier output to change, in coordination with the output voltage, in such a direction that minimum voltage drop across the series regulator(s) occurs. This reduces power dissipation in all series regulator elements.
- A3-2. Remote sensing leads connect from the supply to the load end of the camera cable, and therefore permit sensing the voltage at the load rather than at the supply terminals. Thus the load can be held constant regardless of the length of the camera cable.
- A3-3. So that any rfi components can be made to occur during blanking and not appear as beat patterns during active picture scanning.

- A3-4. The power supply would operate normally, but it would not shut down if excessive current occurred in the load. (The over-current safety feature will not work.)
- A3-5. The color-sync burst normally is maintained at the same peak-to-peak amplitude as the sync pulse. If the sync is adjusted to 0.3 volt (above blanking), then the burst amplitude is 0.3 volt peak-to-peak. The FCC specification is 0.9 of sync amplitude to 1.1 of sync amplitude.
- A3-6. No. It is eliminated during the 9H interval of equalizing and vertical-sync pulses.
- A3-7. 8 to 10 cycles.
- A3-8. Note from the statement concerning this subject in the FCC rules that the color-transmitter response is tightly controlled at the color subcarrier frequency of 3.58 MHz. Note also that the applicable specifications are tied to the "ideal detector curve" shown in Fig. B-1. The first 0.75 MHz of the modulation frequency range is double-sideband; hence the diode response is 100 percent in this region. Due to vestigial-sideband transmission, the diode response at 1.25 MHz will be down 6 dB *if the transmitter has 100-percent response to this frequency*. In essence, then, the actual attenuation-vs-frequency response of the color transmitter must be within plus or minus 2 dB for frequencies up to 4.18 MHz (using response at 200 kHz as a reference) to meet FCC requirements.



**Fig. B-1. Ideal detector-diode curve.**

- A3-9. No. The camera head normally employs a gamma of less than unity (black stretch).
- A3-10. Yes. If not, luminance distortion (which also affects colors) will occur.
- A3-11. Phase sensitivity.
- A3-12. Multipath reflections with attendant phase shift at various frequencies in the radiated signal. Also, of course, the receiver circuits can affect color reproduction.

- A3-13. 1. Yellow would go greenish.  
2. Cyan would go bluish.  
3. Green would go bluish (cyan).  
4. Magenta would go reddish.  
5. Red would go yellowish.  
6. Blue would go toward purple (magenta).  
7. White would remain white.
- A3-14. Errors in the input-signal amplitudes at the encoder (all inputs must be equal for white). Improper I or Q white balance adjustments. Carrier unbalance. I or Q coefficient errors in the receiver.
- A3-15. Low overall chroma gain.
- A3-16. Good human "flesh tones."
- A3-17. A red with reduced luminance value. This could be caused by lack of proper gamma correction (black stretch) in the camera chain, or by video unbalance in the encoder.
- A3-18. Differential gain, in which the chroma gain decreases with an increase in brightness.

#### CHAPTER 4

- A4-1. 31.468 kHz, counted down from the subcarrier frequency standard.
- A4-2. Six leading and six trailing.
- A4-3. Field 1 has a full-line interval prior to the 9H vertical time. Field 2 has a half-line interval prior to the 9H vertical time.
- A4-4. Approximately  $1.5 \mu\text{s}$  per 1000 feet, or  $0.0015 \mu\text{s}$  per foot.
- A4-5. Approximately  $1.9^\circ$ .
- A4-6. No. The phase of the color-subcarrier burst is adjusted in the encoder. The flag pulse influences only the timing (position) of the burst.
- A4-7. One way is to use a time base of  $0.1 \mu\text{s}/\text{cm}$  and use  $0.1\text{-}\mu\text{s}$  markers (with the Tektronix Type 524 scope). Trigger the scope with horizontal-drive pulses. You can use a time base of  $1 \mu\text{s}/\text{cm}$  if scope triggering instability occurs on the extremely short time base. Or, you can use  $10 \mu\text{s}/\text{cm}$  with  $\times 10$  sweep magnification. There should be five of the markers between the trailing edge of horizontal sync and the first (full) cycle of burst. On more modern scopes, markers are not necessary.
- A4-8. If bursts are occurring during the entire 9H interval, the 9H eliminate pulse obviously has been lost completely. When some of the bursts are eliminated, but not for the full nine lines, the burst-eliminate pulse has excessive slope so that the eliminate

keyer does not hold off for the entire interval. The most likely cause of this trouble is insufficient clipping at a clipper stage. This could result from low output level of a stage, or a leaky coupling or bypass capacitor (either of which would reduce the amplitude). Your first check should be to scope the pulse at the multivibrator output to observe the capability of obtaining a 9H pulse width. If no trouble is apparent here, you will most likely find a badly sloping pulse from the clipper, as noted above.

- A4-9. 18 peaks.
- A4-10. The color-sync burst is normally maintained at exactly the same peak-to-peak amplitude as the sync pulse. If the sync is adjusted for 0.3 volt above blanking, then the burst is 0.3 volt peak to peak. The FCC tolerance range is 0.9 of sync amplitude to 1.1 of sync amplitude.
- A4-11. No. It is eliminated during the 9H interval of equalizing and vertical-sync pulses.
- A4-12. Eight to nine complete cycles.
- A4-13. The voltage-controlled oscillator.
- A4-14. The subcarrier frequency is too high, the sync generator is on crystal control (instead of subcarrier-frequency control), or the sync generator is in free-run operation. Normal dot travel is from bottom to top.
- A4-15. No. The subcarrier-frequency oscillator is a temperature-controlled crystal circuit. It is improbable (after normal operating temperature is reached) that an error greater than  $\pm 50$  Hz will exist (the tolerance is  $\pm 10$  Hz). But even assuming a  $\pm 100$ -Hz error, since the total division is 113.75, the final error would be  $100/113.75 = 0.87$  Hz. So once the counters are properly centered, the frequency adjustment can be made at any time, as the count is quite broad.
- A4-16. The oscillator frequency of 3.579545 MHz is reduced to 31.46852 kHz by the counter chain. This is usually done with three divider stages ( $5 \times 7 \times 13 = 455$ ) and a  $\times 4$  multiplier, to make the total division  $455/4 = 113.75$  times. Double check this by multiplying:  $31.46852 \text{ kHz} \times 113.75 = 3.579545 \text{ MHz}$ .
- A4-17. Not if you switch directly from one source to another. The receiver simply uses the burst phase it receives for synchronous demodulation. But if you *mix* two separate sources, they must have the same system phase.

## CHAPTER 5

- A5-1. To compensate for the large shunt-capacitive effects on the bridging input line.

- A5-2. The machine control panel and the video-switcher panel.
- A5-3. A film island employs a multiplexer that allows one camera to handle at least two film projectors and one slide projector. Therefore all interconnections are normally made through the multiplexer unit.
- A5-4. Equalization and extra gain.
- A5-5. With a chrominance signal input of 1 volt peak-to-peak, the measureable cross talk should be 0.001 volt, or 1 millivolt, peak-to-peak. (See decibel table, Appendix A.) Review the discussion of cross-talk measurement in Section 5-3.

## CHAPTER 6

- A6-1. The total current through R1 and R2 is  $20/40k = 0.5$  mA (base current assumed negligible). The 0.5 mA through R2 produces  $0.5 \times 4k = 2$  volts from base to ground.
- A6-2. Since Q1 is germanium, the emitter voltage to ground is about +1.8 volts. The emitter current is  $1.8/1000 = 0.0018$  A = 1.8 mA.
- A6-3. The collector current is  $0.98 \times 1.8 = 1.76$  mA.
- A6-4.  $V_B = +2$  V;  $V_E = +1.8$  V;  $V_C = +11.2$  V.
- A6-5. Since  $V_E$  is 1.8 V and  $V_C$  is 11.2 V, then  $V_{CE} = 11.2 - 1.8 = 9.4$  V.
- A6-6.  $P_C = (9.4 \text{ V}) (1.76 \text{ mA}) = 16.5$  mW.
- A6-7.  $\beta = \alpha / (1 - \alpha) = 0.98 / 0.02 = 49$ . (For quick analysis, the value of  $\beta$  can be assumed to be 50.)
- A6-8.  $I_B = I_C / \beta = 1.76 \text{ mA} / 50 = 35 \mu\text{A}$  (approx).
- A6-9. Input  $Z = (r_{tr}) (h_{FE}) = (26/1.8 + 4 + 1000) (50) = 1018 (50) = 50,900$  ohms.
- A6-10. Load  $Z = 5k$  in parallel with  $5k = 2.5k$  (approx).
- A6-11.  $A_v = R_L / r_{tr} = 2500 / 1018 = 2.5$  (approx).
- A6-12. With a large  $R_E$ , you know that  $A_v$  is essentially  $R_L / R_E$ , or  $2500 / 1000 = 2.5$  times.
- A6-13.  $I_E = 10/5k = 2$  mA.  $I_C = (0.98) (2) = 1.96$  mA (can be assumed to be 2 mA—same as emitter current—for rapid analysis). Input  $Z = r_{tr} = 26/2 + 4 = 17$  ohms. Load  $Z = 5k$  in parallel with  $5k = 2.5k$ .
- A6-14.  $V_E$  should be about +0.2 to +0.3 V.  $V_C$  should be about -10 V.
- A6-15.  $A_v = \frac{\text{Load } Z}{\text{Input } Z} = 2500 / 17 = 147$  (approx).

- A6-16. There will be very little actual difference because the degenerative emitter resistance in Fig. 6-52 greatly reduces the input capacitance. But note the extreme difference in gain to get approximately the same bandwidth.
- A6-17. The extremely low input impedance. To obtain an input impedance equivalent to that of Fig. 6-52, a large series resistor would be necessary in the signal path to the emitter, greatly reducing the voltage gain.
- A6-18. For Fig. 6-52:

$$A_p = \beta \frac{R_L}{r_{tr}} = 50 (2.5) = 125$$

For Fig. 6-53:

$$\begin{aligned} A_p &= \alpha^2 \frac{R_L}{r_{tr}} = (0.98)^2 (147) \\ &= (0.96) (147) = 141 \end{aligned}$$

- A6-19. The new  $r_{tr} = 26/1.8 + 4 = 18$  ohms (approx). Then:  
 $A_v = 2500/18 = 139$   
 Input  $Z = (18) (50) = 900$  ohms  
 $A_n = (50) \frac{2500}{18} = 6950$

Although a higher impedance than in Fig. 6-53 has been obtained at approximately the same voltage gain, the bandwidth is less.

- A6-20. Yes, very much. High-frequency types are manufactured which offer much improvement in bandwidth over lower frequency types.
- A6-21. The input impedance of the transistor is 17 ohms (see A6-13). So,  $R_s = 75 - 17 = 58$  ohms. The 75-ohm line is "transformed" to an impedance of 5000 ohms in parallel with the following load. Both voltage and power gain are provided in this transformation.
- A6-22. The input signal now appearing at the emitter is  $17/75 = 0.23$  (approx) of the signal at the end of the 75-ohm line. Therefore, the actual voltage gain is now  $0.23 \times 147 = 34$  (approx). Compare this result with A6-15.
- A6-23.  $Z_{out} = 5000/(50 + 1) = 100$  ohms (approx) looking back into the emitter itself. Then  $R_s = 100 - 75 = 25$  ohms.
- A6-24.  $1/75 = 13.3$  mA (approx) peak to peak.
- A6-25. The total supply voltage is 20 volts (collector and emitter work between minus 10 and plus 10 volts, respectively). So  $R_E = V/I = 20/0.02 = 1000$  ohms. A signal swing of more than 20 mA peak to peak will then cause clipping.

- A6-26. Use of a stage with low output impedance (such as another emitter follower) to drive an emitter follower.
- A6-27. Cross talk and voltage-gradient ("glitching") problems. In cases of definite faulty grounding, picture "hum" also can result.
- A6-28. A glitch results from a voltage built up across a common impedance of the system. It normally occurs only upon operation of such mechanisms as slide changers or in certain types of video-switching systems. Small transients occur in any system with long unbalanced coaxial lines; the problem is to minimize such transients (white flashes in the picture when the switch or slide change occurs) to a negligible amount. Some cases are actually so severe as to cause picture "rolls" in home receivers upon actuation of the troublesome device.

The first thing to do is to provide as much electrical suppression of the offending equipment as possible by shielding of the control box and by coil and contact surge suppressors (diodes or RC suppressor arrangements). If the manufacturer of the particular equipment is out of ideas for any further help, you must go further into your own system.

One factor is the tally-light system. First, disable the tally-lamp power. Is the "glitching" reduced? Then disable the tally-relay control voltage. Be sure *any* relays in the system are properly "arc suppressed" as called for by the manufacturer. Glitching when a certain camera chain is switched on the air has been known to be caused by the tally-light relay in the camera head. Try different suppressor combinations across these coils and/or contacts.

It may be that ground loops exist only when certain pieces of equipment are connected into the system. Even a faulty ground on a viewfinder was found in one case to cause glitching when a slide changer operated. When the viewfinder was removed from the camera, the glitch practically disappeared. (The particular camera-chain routing happened to be nearest to the offending slide-change control box.) Even though it "doesn't make sense," try disconnecting any equipment that can be spared during tests for this condition. All coax must have solid shield connections at both ends. When you isolate any line or sections of units which seem to emphasize the glitch problem, examine these for "clean" grounds, shielding, solid interconnecting harness, etc.

When the trouble is in a video switcher, even tube conditions can cause excessive switching transients. Cathode interface of tubes will do this. In the case of relay-type switchers (as well as straight mechanical contacts in push buttons), be sure an overlap interval occurs rather break-before-make operation. The usual overlap interval is 200 to 300 milliseconds.

Problems of this nature can be minimized by going to 124-ohm balanced video-distribution systems. Many of the latest solid-state distribution amplifiers provide such an arrangement.

$$\begin{aligned}
 \text{A6-29. } E_o &= E_{in}(1 + R_f/R_1) \\
 &= 1(1 + 1) \\
 &= 1(2) = 2 \text{ volts}
 \end{aligned}$$

Thus we have a voltage gain of 2 for this noninverted signal.

$$\begin{aligned}
 \text{A6-30. } E_o &= -E_{in}(R_f/R_1) \\
 &= -1(1) \\
 &= -1 \text{ volt}
 \end{aligned}$$

Thus we have an inverted signal with no gain.

## CHAPTER 7

- A7-1. (1) Check the high voltage, and adjust it if necessary. (2) Check the purity, and adjust it if necessary.
- A7-2. The static controls involve dc currents and affect convergence mainly around the center of the raster. The dynamic controls involve in-step ac waveforms that compensate for changes in the deflection-beam paths as the beams depart from the central area of the raster.
- A7-3. The burst will be interlaced.
- A7-4. With the sweep rate set at 0.125H/cm, you look at every other horizontal line and observe the noninterlaced color burst.
- A7-5. Yes. The time constant is long relative to the line duration of 63.5  $\mu$ s. The dc restorer does, however, function effectively in holding the blanking line at reference level for varying APL.

## CHAPTER 8

- A8-1. (A) Distortion which is independent of amplitude, providing the normal operating range is not exceeded. (B) Distortion the degree of which varies with signal amplitude, within the normal operating range.
- A8-2. By reducing the test-signal input amplitude and noting whether the degree of distortion changes.
- A8-3. This is a voltage ratio of 0.5, or 6 dB. (Refer to the dB table in Appendix A.)
- A8-4. 90 IEEE units for 10-percent setup, or 92.5 IEEE units for 7.5-percent setup.
- A8-5. (A) 0.25H/cm  $\times$  25 (0.01H/cm, or 0.635  $\mu$ s/cm).  
(B) 0.125H/cm  $\times$  25 (0.005H/cm, or 0.318  $\mu$ s/cm).
- A8-6.  $f_c = \frac{1}{R_p} = \frac{1}{0.15} = 6.7 \text{ MHz}$

- A8-7. The severity (rapidity) of the rolloff at 6.7 MHz, and whether the pulse h.a.d. is  $0.0625 \mu\text{s}$  or  $0.125 \mu\text{s}$ .
- A8-8. No, the ringing period will be the same. The amplitude of the ringing will be larger for the  $0.0625\text{-}\mu\text{s}$  pulse than for the  $0.125\text{-}\mu\text{s}$  pulse.
- A8-9. Smearing is an impairment of shorter duration than streaking. Fine detail in the picture can be smeared with no evidence of streaking. Fig. 8-24 is an example of smearing (exaggerated almost to the point of streaking). Figs. 8-25 and 8-26 are examples of streaking.
- A8-10. Yes; see Fig. 8-31A.
- A8-11. No. Power levels are independent of impedance values.
- A8-12. Yes. The voltage across or the current through an impedance depends on the impedance value as well as the power level.
- A8-13. Zero dBm is 1 milliwatt, and  $E = \sqrt{WR}$ . Therefore:  

$$E = \sqrt{(0.001)(600)} = \sqrt{0.6} = 0.774 \text{ volt}$$
- A8-14.  $E = \sqrt{(0.001)(150)} = \sqrt{0.15} = 0.387 \text{ volt}$

## CHAPTER 9

- A9-1. The designed physical dimensions of the cavity, a mechanical tuning screw in the cavity, and a variable negative potential on the reflector electrode.
- A9-2. Lowered rf signal strength (usually indicated on a meter at the receiver), and deteriorating signal-to-noise ratio.
- A9-3. (A) Minimum sync-pulse amplitude at the output of the sound demodulator (video applied but no audio), and (B) minimum differential phase at 3.58 MHz.
- A9-4. Fm.
- A9-5. Fm.

## CHAPTER 10

- A10-1. If it does, you have compression; see Figs. 10-4, 10-5, and 10-6. This assumes, of course, that the signal includes a reference white level and that the modulation is full modulation.
- A10-2. Not necessarily. The ratio may require precorrection to obtain the proper modulated output.
- A10-3. 87.5 percent.
- A10-4. The stairstep signal with a superimposed 3.58-MHz component.
- A10-5. The video stabilizing or video processing amplifier.

- A10-6. The color precorrection units indicated in Fig. 10-1 and described in the text.

## CHAPTER 11

- A11-1. The *indirect method* is used. The power is taken to be the product of the plate voltage ( $E_p$ ) and plate current ( $I_p$ ) of the aural final stage, multiplied by an efficiency factor ( $F$ ):

$$\text{Operating power} = E_p \times I_p \times F$$

- A11-2. Since:

$$\text{Power output} = E_p \times I_p \times F$$

then:

$$I_p = \frac{P_o}{E_p F} = \frac{1000}{(2700)(0.78)} = 0.475 \text{ A} = 475 \text{ mA}$$

- A11-3. This factor is established by the manufacturer for each type of transmitter approved by the FCC. The numerical value of the efficiency factor is supplied to the customer with each transmitter.

- A11-4. The efficiency factor has a rather wide variation. It can have a value as low as 0.5 to 0.8 in a transmitter with a power output in the range from 250 to 1000 watts. In higher-power transmitters, the efficiency factor may run as high as 0.98.

- A11-5. (A) The term *frequency swing* is used to define the instantaneous departure of a frequency-modulated carrier from the center frequency as a result of the audio modulating signal applied to the transmitter.

(B) The term *frequency deviation* means the amount by which the aural center frequency or visual carrier frequency has drifted from the assigned frequency.

NOTE: The term *deviation* is often used interchangeably with *swing* when referring to a frequency-modulated carrier. This can be confusing if not properly related.

- A11-6. (A) the term *percentage modulation* (as applied to fm) means the ratio of the actual frequency swing to a set maximum frequency swing defined as 100-percent modulation.

(B) For the tv aural transmitter, a frequency swing of  $\pm 25$  kHz is defined as 100-percent modulation.

- A11-7. (A) Average power is:

$$\text{Average power} = I_{\text{rms}}^2 R$$

or:

$$\text{Average power} = \frac{E_{\text{rms}}^2}{R}$$

when the visual transmitter is operating into a dummy load of essentially zero reactance and a resistance equal to the transmis-

sion-line surge impedance, while transmitting a standard black picture. (B) Peak power is the average power obtained in (A) multiplied by the factor 1.68. This multiplying factor is obtained from the fact that the ratio of the rms voltage of the carrier during horizontal-sync time to the rms voltage averaged over the entire line time is equal to 1.295, which when squared gives 1.68.

- A11-8.** Assuming you have eliminated vacuum tubes as possible contributing factors, run the same check only to the modulator output. The modulator circuit must see a constant load impedance over the entire video band down to and including dc. Some modulator power supplies have the internal resistance designed into a matching network consisting of both high-frequency and low-frequency sections to make the load impedance independent of frequency. This check will tell you whether the trouble is at the point of modulation (and the trouble *can* be transmission-line regulation between the rf driver and modulated stage, as well as in the modulator circuit), or after this stage in the final rf amplifiers.

If you discover you have a problem here, you may have inherited a less-than-ideal installation. Variations in frequency response with power level (assuming proper tuning) are caused by a change in effective loads with power change. The electrical load must remain constant (within narrow limits) at all power levels (APLs) encountered under picture operation. The transmission line from the modulated amplifier to the power amplifier sometimes must be electrically a half wavelength (or a multiple) at the operating frequency. The more it varies from this length, the greater is the change in loading with power-level changes. We are referring now to "frequency-selective" loading as contrasted to transmitter regulation from black to white. It is possible for a transmitter to meet specifications for regulation from black to white, but still exhibit variations in frequency response with power level due to poor transmission-line regulation. Of course, it is possible that loading adjustments and "line-stretchers" are not optimized. Also, the transmission line between the rf driver and the modulated stage cannot be ignored, since its length is usually critical. If you have this problem, contact the transmitter manufacturer for advice. If it is necessary to start cutting new lines, do it. Many of us have "inherited" problems. The "sin" is not doing anything about them.

- A11-9.** Be sure to recheck the depth of modulation when you increase or reduce white stretch. The most obvious chance for trouble occurs when you increase white stretch and fail to recheck modulation, since you can be modulating to carrier cutoff. Always be sure to maintain a minimum of 10 percent carrier on white, since very few transmitters can be made linear beyond this point even though carrier cutoff is not reached.

*A11-10.* Again assuming you have eliminated vacuum tubes (and power supplies) as possible contributing factors, run the frequency-response versus power-level check as discussed in A11-8. Both of these troubles are caused by the same general condition: a load that is not constant at all frequencies within the intended pass-band. This can occur in straight video amplifiers ahead of the transmitter modulation. If differential-gain and -phase specifications are good up to the transmitter input, but the performance of the transmitter itself varies with APL, the most likely sources of trouble lie in the modulator, tuning procedures, or poor transmission-line power regulation.

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