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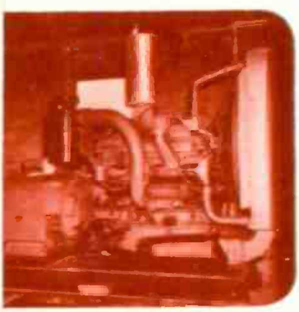

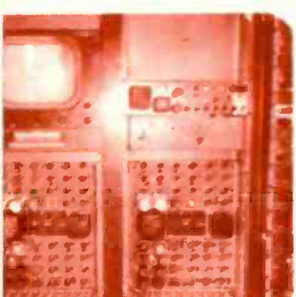

Television Systems Maintenance

by Harold E. Ennes

TELEVISION SYSTEMS MAINTENANCE

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Television Systems Maintenance combines the latest techniques in the standardization of television systems maintenance procedures with a complete analysis of system malfunctions. A certain amount of fundamental systems theory has been included where important in maintenance practice.

The book treats the television system 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.

Television Systems Maintenance includes the latest troubleshooting and maintenance techniques, together with discussions of standards for picture signal analysis, amplitude calibration and maintenance of levels, sync generators, pulse distribution systems, video switchers, microwave systems, transmitter maintenance and proof of performance, and many other topics pertinent to television systems and equipment.

Numerous easy-to-follow diagrams, line drawings and actual photos enhance the author's practical explanations. The book is designed to serve the need for a ready reference for broadcast station personnel, as well as a basic text for home study or classroom use. Every broadcast engineer, technician, or student, and everyone engaged in manufacturing and other activities which require a knowledge of television broadcasting, should have this volume.

ABOUT THE AUTHOR

Harold Ennes has been associated with various phases of radio engineering since 1930. He entered the broadcast field in 1936 as a staff engineer with station WIRE, Indianapolis. Later he installed the first FM broadcast station in Indianapolis—noncommercial WAJC for Jordan College of Butler University—and was the station's chief engineer for four years. In addition, he taught radio and television at Butler University for five years. Since 1958 Mr. Ennes has been maintenance supervisor for Television City, Inc. (WTAE-TV Pittsburgh). He has written numerous articles and books on the various aspects of radio and television broadcasting. Other SAMS broadcast references by Mr. Ennes include: *AM-FM Broadcast Operations*, *AM-FM Broadcast Maintenance*, and *Television Tape Fundamentals*.

HOWARD W. SAMS & CO., INC.
THE BOBBS-MERRILL CO., INC.

20069
$5.95 (In Canada $7.15)

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Formerly

BROADCAST ENGINEERING NOTEBOOKS: TELEVISION SYSTEMS MAINTENANCE

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HOWARD W. SAMS & CO., INC.
THE BOBBS-MERRILL CO., INC.
INDIANAPOLIS • KANSAS CITY • NEW YORK

FIRST EDITION

FOURTH PRINTING — 1967

Library of Congress Catalog Card Number: 62-21970

PREFACE

Television systems maintenance procedures have only recently started to become more standardized. While there are still many variations in techniques, these are steadily approaching systemized 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 considered important in maintenance practice.

The TV 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 other volumes of this series.)

In particular this book covers the visual units of the telecast system. Major differences in the sound portions, as contrasted with standard FM radio systems (diplexed audio, STL's, and special considerations of measurement of the aural transmitter), are treated accordingly. Standard maintenance techniques for FM audio systems are thoroughly covered in another book (*AM-FM Broadcast Maintenance*) by the author, and are not duplicated here. *AM-FM Broadcast Maintenance* also includes details concerning component parts that are just as applicable in television system maintenance; therefore, it contains data of value to the television engineer and should be consulted for details on general maintenance techniques.

HAROLD E. ENNES

ACKNOWLEDGMENTS

The author hereby extends his appreciation to the stations and manufacturers whose contributions assisted in making this book possible.

Data from Stations WISH-TV, WFBM-TV, WTAE, and from General Electric, RCA, and AT&T are acknowledged where applicable on charts and photographs.

Acknowledgment is also due to RCA for the basic description of transistorized power supplies in Section 2, and for the photos and description of the TS-40 Vertical Switcher in Section 4.

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SECTION 1

STANDARDS FOR PICTURE SIGNAL ANALYSIS

A "standard" is that which is established for calibration of an instrument to indicate when (for example) a "volt is a volt" or a "microsecond is a microsecond." This book is not to be considered as any revelation of "standard practice" in the field. It is realized that some readers may have honest differences of opinion on maintenance techniques. However, whatever the technique used, a "standard" starting point must be established for comparison.

To emphasize this point, consider the measurement of corner-to-corner resolution of a camera, looking at a studio test chart. It is perfectly valid for the operator to observe this on his monitor and make the necessary adjustments required 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 preventing performance deterioration beyond a limit set at the lower end by FCC Standards and at the higher end by the Chief Engineer of the particular station. This higher end (as it should be) is usually limited 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 dictated by previous experience for a given pickup tube and camera, the maintenance engineer must first know the characteristics of the monitor he is using before a valid measurement can be made. It is entirely possible for a monitor to exhibit 50, 100, or more lines difference in resolving power (either plus or minus) between the corner and center of the raster. How to measure this monitor characteristic is covered in Section 5 of this book.

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

1-1. THE OSCILLOSCOPE PERSONALITY

Every scope exhibits its own "personality" as observed on the CRT. Getting acquainted with the individual scope characteristic is the initial step in calibration of the instrument.

The two basic types of information displayed by the CRT are *amplitude* and *time*. *Waveshape* is not really a third basic type of information; it is simply amplitude versus 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, such as rise and decay times of pulses, cutoff (ringing) frequency, etc. The amplifier within the scope itself becomes the standard which must be considered in measurements.

Fig. 1-1 shows four response curves which presently concern the telecasting industry. The so-called Gaussian curve (A) has a roll-off suited for best transient response. The approximate relationship of this curve is such that the 0 to 3 db point is equal to the 3 db to 12 db point. 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.)

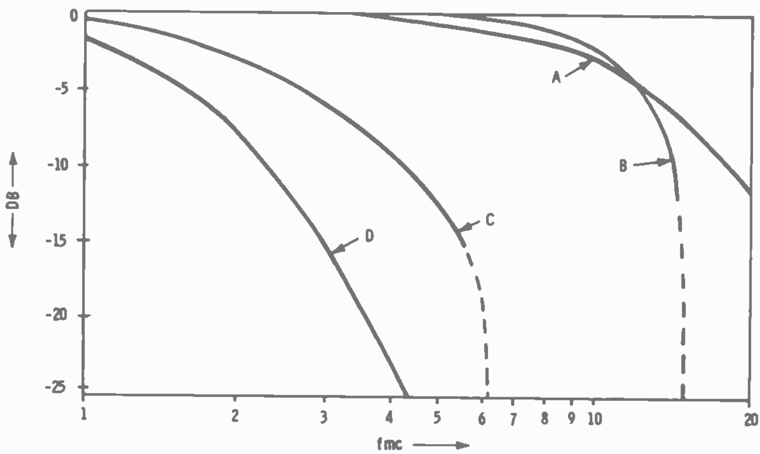


Fig. 1-1. The four major response curves with which the telecasting industry is concerned. (A) "Gaussian" or normal response—gradual roll-off (9 db/octave), rise time approximately 0.035 microsecond. (B) Flat response to 5 or 6 mc. More rapid roll-off than curve A. Rise time approximately 0.035 microsecond. (C) Old IRE curve. Rise time 0.175 microsecond. (D) New IRE curve. Rise time 0.3 microsecond.

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

The old IRE response curve (C) was originally adopted for comparative level checks. We are currently in the process of a changeover to the new IRE curve (D) adopted in 1958 as being more indicative of true luminance levels. NOTE: the exact specifications of this curve are given in Section 2.

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

1. NORMAL (Gaussian roll-off)—most suitable for waveform analysis, particularly where transient response becomes a major factor.
2. FLAT (to 5 or 6 mc)—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.

1-2. 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 megacycles (to the 3 db down point),
RT is the risetime in microseconds (measured between 10% and 90% of peak value),
k is the factor lying 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 must be less than 3%. In fact, a system has an equivalent bandwidth and rise time only within the limits of 3% 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 shows the tabulation of rise time for bandwidths from 1 to 10 mc within the preceding limitations. Going back to Fig. 1-1, it may be noted that the 3 db down point of either the Normal or Flat curve falls in an area which safely indicates a bandwidth of

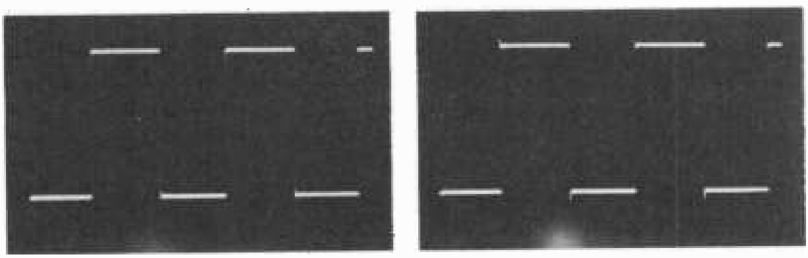
Table 1-1. Bandwidth Rise Time for K = 0.35 Overshoot Under 3%

BW (megacycles)	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

10 mc. 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. 1-2A and 1-2B illustrate the difference in overshoot of a 75-kc square wave with the scope set for Normal and Flat response, respectively.

It is the shape of the curve that is actually being changed when video peaking coils are adjusted. Leading and trailing transients of a rapid transition in picture content must be adequately controlled by the maintenance personnel. Hence complete familiarity with the scope amplifier characteristic is necessary.

A good square-wave generator with reasonably fast rise time and a flat top response curve completely free of wrinkles is one Tektronix Type 105 generator with a rise time of 0.02 microsecond and a flat top response curve completely free of wringles 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 at least one fifth that of the pulse to be measured. Now the rise time of a scope with a 10-mc 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 better. A fast rise time pulse is necessary for transient response checks in terms of overshoot or undershoot. Since the square wave generator must be terminated in 75 ohms to preserve rise time, it becomes impractical to check the pulse directly on the CRT since sensitivity is not sufficient for accurate measurement. Hence, it is necessary to determine the "standard" rise time (and overshoot) for any particular combination of generator and scope before a valid check can be made on external amplifiers or systems.



(A) Scope set for Normal response. (B) Scope set for Flat response.

Fig. 1-2. Tektronix 524 scope response to a 75-kc square wave (rise time 0.02 microsecond).

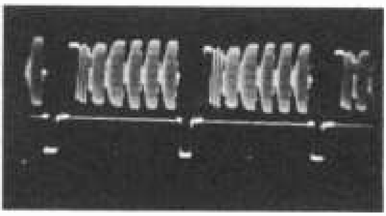
The total rise time of a pulse through a series of cascaded stages is equal to the square root of the sum of the squares of individual stage rise times (assuming overshoots less than 3%). When, for example, an amplifier with a rise time of 0.02 microsecond is feeding an amplifier with 0.04 microsecond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

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

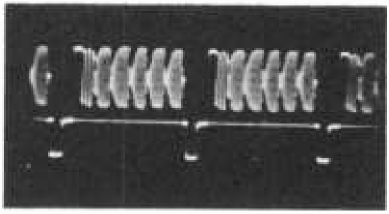
Understanding this relationship will enable the maintenance engineer to closely estimate the condition of his test equipment, even though an extremely wideband scope is not available or necessary—provided he is certain of the scope characteristics. It also emphasizes the better-known premise that an amplifier output must be directly compared 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, capacities, etc., it is important to check the input display at the point of connection so that the output can be properly interpreted.

1-3. SCOPE PROBES AND INITIAL CALIBRATION

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. 1-3A shows the display of a keyed burst signal taken with a direct probe across a 75-ohm termination. Actually the display obtained will depend on the length, type, and condition of the cable used; one probe could show a decided roll-off of higher frequencies, while another



(A) Using direct probe across 75-ohm termination.



(B) Using cathode-follower probe across 75-ohm termination.

Fig. 1-3. Scope response to keyed sine-wave burst.

could indicate a roll-off at low frequencies. Similarly, pulses would have varying rise times and overshoots, depending on the duration and repetition rates. Fig. 1-3B is the display obtained with a cathode-follower probe across the same termination. Use the direct probe only where IRE response is used. No probe loss can be tolerated when checking for the presence of extremely low-level signals, and certainly the direct probe has no place in scope calibration.

For most applications the 10:1 capacity divider probe should be used. For a scope with 1-megohm input shunted by a 40-mmf capacity, the simplest 10:1 probe consists of a series-connected 9-meg resistor shunted by a trimmer capacitor of 3 to 12 mmf. When connected to the scope, the input impedance from the probe tip becomes 10 megohms shunted by approximately 12 mmf. The trimmer capacitor is adjusted so that the RC product 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 kc and adjusting the trimmer so that the leading edge is not rounded on the top (undercompensated) or does not overshoot (overcompensated). This adjustment should be checked often and must always be checked when used with a different scope, even though it is one of the same make and model. Frequency response and transient response of the scope itself should be checked with this probe so that all variables are calibrated.

An important point to remember when using the 10:1 probe is that the preamplifier will normally be used ahead of the main vertical amplifier in such scopes as the Tektronix 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 requires the extra gain. It is, therefore, imperative that single-frequency 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 attenautor settings are the same as when equipment checks are made.

The cathode-follower (CF) probe overcomes the above limitation. A typical CF probe has an input impedance (connected to the scope) of 40 megohms with a 4-mmf shunt capacity, and it provides a gain slightly less than unity. Thus, the preamplifier need not be used across normal 1-volt terminations. The cathode-follower probe does, however, have limitations of its own.

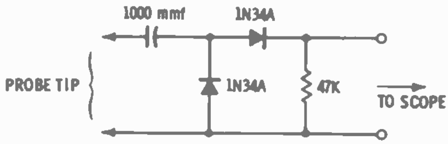
A. Depending on the design and voltage used, a single amplitude of about 5 volts unidirectional (10 volts p-p) is the maximum that can be handled without compression. Thus, the probe is not normally used in servicing equipment where higher signal levels occur, unless an additional voltage-divider probe is attached to the CF input.
B. Due to design limitations on the input time-constant, the low-frequency, square-wave response is poor (about 20% tilt on a 60-cycle square wave).
C. Since a DC voltage appears at the cathode output, the DC input of the scope cannot be used.

A logical step-by-step initial calibration of the scope can be outlined as follows:

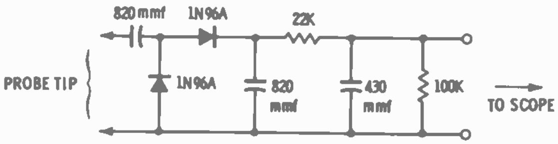
1. *Video Sweep (Detected)*. Terminate the video sweep generator directly at the generator output connector in 75 ohms. Use a video detector probe (Fig. 1-4). The probe in Fig. 1-4A will read approximately 75% of actual p-p output signal, while the higher isolation probe in Fig. 1-4B will read about 50%. Adjust the output amplitude for 1 volt, which will read approximately 0.75 volt with probe A or 0.5 volt with probe B. Also, adjust the scope gain to provide a convenient display (Fig. 1-5). This enables a check of the flatness of the sweep generator itself, since the detected sweep envelope does 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 ob-

served on the scope, the deviations must be plotted as a correction factor for equipment checks and scope calibration.

2. *Video Sweep-Wideband.* (This should only be observed after determining the flatness of the sweep generator as in Step 1.) Although used only in very special cases (and with extreme care), the RF envelope may be observed directly without



(A) Simplest type with a detected output that is approximately 75% of peak-to-peak value.



(B) Probe with additional isolation and filtering providing detected output of approximately 50% of peak-to-peak value.

Fig. 1-4. Two types of peak-to-peak video sweep detector probes.

detection as a "quickie" check on scope amplifier response (Fig. 1-6). This check, however, is valid only if the probe to be used for 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

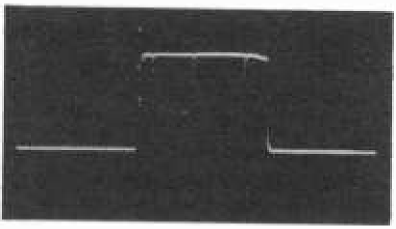
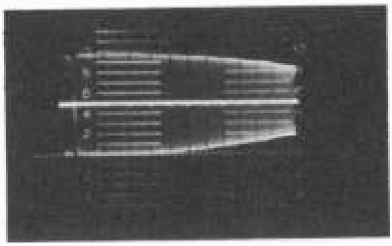
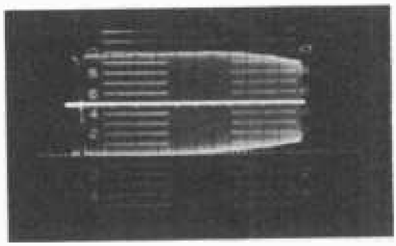


Fig. 1-5. Detected video sweep markers at 1, 5, and 10 mc.

with all probes in stock, and through the scope preamp as well as to the vertical amplifier input. Use varying levels from the sweep generator to enable 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 in-accurate in system measurements. Normally there will be some correction factor when using the preamp as when feeding the vertical amplifier direct. Plot these responses either on a graph or by tabulation in peak-to-peak values. Normally the detector probe is employed when using video sweep. The wideband display provides a quick check of scope response to single frequency sine waves or similar applications, such as keyed sine-wave bursts (Fig. 1-3).



(A) Scope set on Normal response.

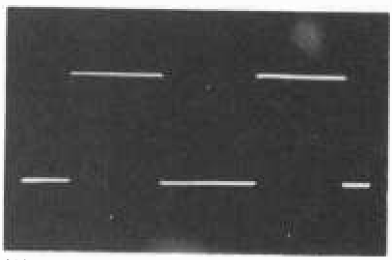


(B) Scope set on Flat response.

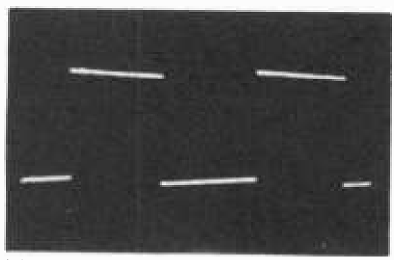
Fig. 1-6. Undetected video sweep display up to 10 mc.

3. *Single-Frequency Sine-Wave Checks*—The most accurate method of checking scope amplifier frequency response is to run single-frequency, sine-wave checks over the range of 100 kc to 10 mc. The same generator and probes should be used for scope calibration as will be used for system checks. Commercial sine-wave generators, such as the Hewlett-Packard 650-A incorporate a frequency-compensated metering circuit at the output to maintain a constant input to the scope or equipment at all frequencies. If a generator of this type is not available, a VTVM with good response to 10 mc can be used across the terminated generator output. As in Step 2, it is good practice to check all probes and all attenuator settings that are likely to be used in system checks. When the calibration is posted on the scope, the particular generator, meter, and probe should be identified, unless all such items have been found to be directly interchangeable. Such checks should normally be made about twice a year, or at any time that considerable maintenance (tube or component changes, etc.) has been required on the scope or signal generators.
4. *Low Frequency and Transient Response*—Determine the rise time and per cent of overshoot of the square wave as read on the scope, both through the preamp and main amplifier, at the frequencies normally used. Unless a fast rise-time generator is available, higher-frequency square waves (above 75 kc) are not particularly useful because for response checks

at the higher frequencies the rise time of the pulse must be faster than the rise time of the amplifiers to be checked. Keep in mind the discussion associated with Fig. 1-2. A 60-cycle square wave fed to the Tektronix 524AD (DC position) should have an absolutely flat top, as shown by Fig. 1-7A. Fig. 1-7B shows the normal amount of tilt introduced by the input coupling capacitor when the scope is on the AC position. Remember that the last two (highest gain) positions of the above scope are AC only, since the preamp is used on these positions. An adjustable grid time constant (Low-Frequency Compensation control) 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.



(A) 60-cycle square wave with scope on DC position.



(B) 60-cycle square wave with scope on AC position.

Fig. 1-7. Square-wave response patterns.

It is possible for a wideband scope amplifier to exhibit a leading edge overshoot from a vacuum-tube defect known as cathode interface. This low-frequency phase shift results from series resistance and capacitive bypassing effects of a chemical interface layer forming between the cathode sleeve and the oxide cathode coating. 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.

A. Adjust the frequency of the square-wave generator to 500 kc. The waveform should have a rise time of 0.2 microseconds or less.
B. Adjust the time base so that several cycles of the square wave are displayed. If an overshoot appears with a duration of 0.2 to 0.6 microseconds, (see Fig. 1-8), chances are good that one or more tubes in the vertical amplifier 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-kc square wave completes one cycle in 2 microseconds, thus a pulse width of 1 microsecond as shown by Fig. 1-8. The overshoot duration will normally be between 20% and 60% of the total pulse width when cathode interface is present. As a double check, plug the scope into a *Variac* and increase the line voltage to the upper limit allowed. If cathode interface is present, the increased tube heater voltage will reduce overshoot, and a decrease of line voltage will increase 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 *Variac* adjusted to deliver the lower limit of line voltage (105-108 volts) to emphasize the effect of cathode interface.

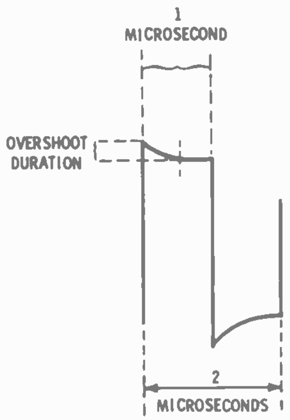


Fig. 1-8. Vacuum-tube cathode interference distortion of a 500-kc square wave.

1-4. THE SWEEP TIME BASE

Recognizing 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 (one at the studio and one at the transmitter) 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 the Tektronix 180A 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 650A, 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 Section 1-5).

Most heterodyne frequency meters employ a stable crystal oscillator which is used for calibrating the frequency of the 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. These 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. 1-9 shows a typical setup for calibrating the frequency meter. Assume that the calibration book shows a crystal check point at a frequency of 2,000 kc. The dial setting of the meter is adjusted to correspond to the number given in the calibration book for 2,000 kc. 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 if sufficiently close) is produced. In Fig. 1-9 the frequency of the beat is 300 cps, indicating that the variable oscillator is 300 cycles off (high or low) from 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. 1-10 shows 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 situated between 3,900 and 4,000, the exact dial reading must be between these numbers. The reading on the UNITS dial is read directly below the arrow 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 its scale which coincides most closely with a line on the UNITS dial. The value of .7 coincides with 33 on the Units dial; therefore, the exact reading of the dial setting shown in Fig. 1-10 is 3,900 + 27 + .7, or 3,927.7. The frequency reading corresponding to this number must be obtained from the calibration book which is included with each frequency meter.

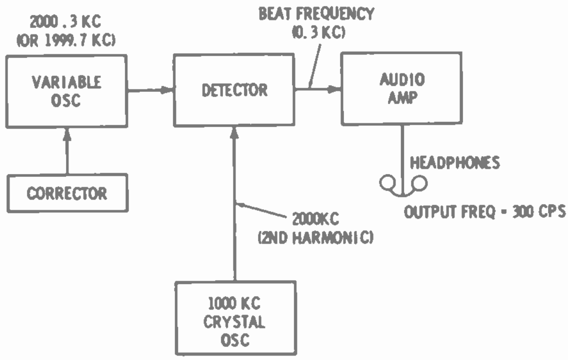


Fig. 1-9. Typical setup for calibrating a heterodyne frequency meter.

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. 1-10 lies betwen the numbers 3,925.5 and 3,927.9, as taken from the calibration book, and the frequencies corresponding to these dial settings are 3,669 kc and 3,670 kc. The difference between the listed dial settings is the difference between the dial reading and the next higher listed dial settings, as the corresponding difference between the listed frequencies is the frequency difference between the unknown frequency and the next higher listed frequency. This is shown in the formula:

$$\frac{2.4}{.2} = \frac{1\ \text{kc}}{x\ \text{kc}}$$

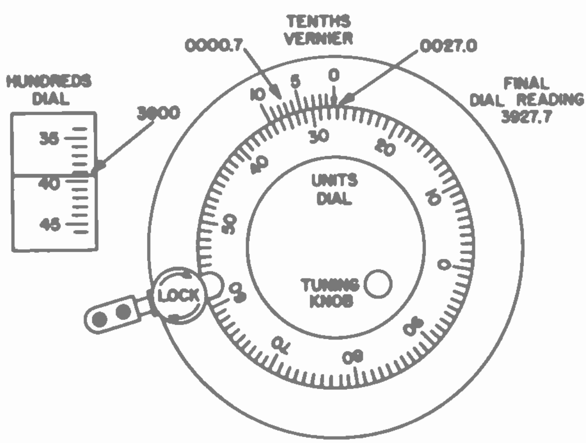


Fig. 1-10. Dial of a typical heterodyne frequency meter.

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

$$\begin{array}{ll}
 & \textbf{FREQUENCIES} \\
\text{diff. 1 kc}\left\{\begin{array}{l}3{,}669\text{ kc}\\ \text{unknown kc}\\ 3{,}670\text{ kc}\end{array}\right. & \left.\begin{array}{l}\\ \\ \end{array}\right\}\text{diff.}\times\text{kc}
\end{array}
\qquad
\begin{array}{ll}
 & \textbf{DIAL SETTINGS} \\
 & \left.\begin{array}{ll}3{,}925.5 & \text{(listed)}\\ \text{diff. .2}\left\{\begin{array}{l}3{,}927.7\\ 3{,}927.9\end{array}\right. & \begin{array}{l}\text{(dial reading)}\\ \text{(listed)}\end{array}\end{array}\right\}\text{diff. 2.4}
\end{array}$$

Solving for x in this formula gives a frequency difference of .083 kc. This figure is then subtracted from the higher listed frequency, producing 3,669.917 kc as the frequency corresponding to the dial reading. The last two significant figures can be discarded for all practical purposes.

When a suitable frequency meter is not available, crystals at 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 crsytal 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 mc, which is the roll-off 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 or WWVH. The most common WWV or WWVH frequencies used are 2.5, 5, and 10 megacycles (see Section 1-5).

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 will be:

$$\text{Sweep time/cm} = \frac{\text{cycles/cm}}{\text{osc. freq.}}$$

For example, on the Tektronix 524AD scope, a 1-mc 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 become completely familiar with the sweep linearity of the scope. Usually a slight amount of nonlinearity will be indicated. Fig. 1-11 illustrates a scope 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. Sweep linearity can usually be improved in case of an excessive amount of nonlinearity by selecting horizontal amplifier tubes for balance while observing markers on the trace, as in Fig. 1-11. It is most important, however, to determine what portion of the sweep the linear region is over.

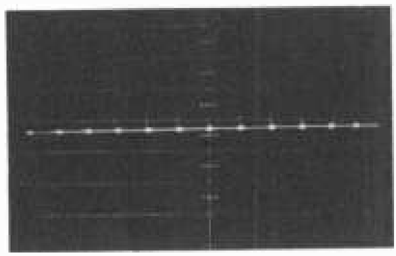


Fig. 1-11. Scope sweep with 1-microsecond markers to observe sweep linearity.

1-5. USING WWV OR WWVH FOR CALIBRATION

Mention has been made of checking frequency standard calibration against WWV or WWVH transmissions. (WWV is located in Ft. Collins, Colorado and WWVH is in Maui, Hawaii). 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 (due to 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. 1-12 illustrates one satisfactory method of calibrating the secondary frequency-standard crystal against WWV or WWVH transmissions. Tune the receiver to 2.5 or 5 mc (10 mc, if necessary for good reception), using the receiver bfo if it is required for accurate dial setting. Wrap a wire around the frequency meter crystal (with the crystal switch on) and 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 instructions with 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.

1-6. 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 AT&T. 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 and more when used as secondary-voltage standards. Eight of these cells in series will provide a standard of 10.8 volts

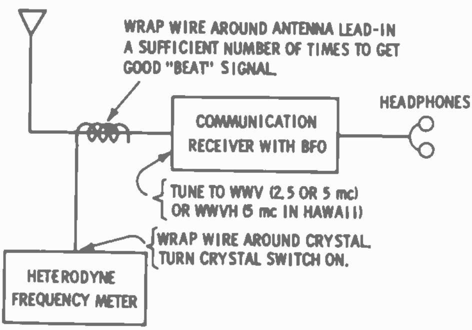


Fig. 1-12. Test setup for zero-beating frequency-standard crystal with National Bureau of Standards transmission.

within 1%, which can be used to calibrate DC meters. The pulse calibration circuit of most scopes employs a DC check point for proper adjustment which requires a voltage scale that can normally 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.

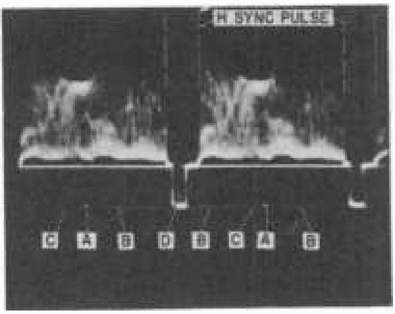
The properly calibrated scope can then be used to check the AC meter scales. The peak-to-peak value displayed multiplied by 0.3535 gives the rms meter value.

1-7. COMPOSITE VIDEO WAVEFORM INTERPRETATION

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

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

Scopes most commonly used in television broadcasting employ a time base calibrated in microseconds rather than frequency. Thus, the "horizontal rate" of 15,750 cps is 63.5 microseconds (the reciprocal of 15,750). To display what is commonly termed "two lines" at the horizontal rate, the scope-sweep time is adjusted to 7,875 cps (one-half the line rate) or 127 microseconds.



(A) Equalizers coincident with vertical sync serrations.

(B) Vertical sync pulses.

(C) Vertical sync serrations.

(D) Equalizers coincident with horizontal sync.

Fig. 1-13. CRO horizontal-rate display of two lines.

With a scope such as the Tektronix 524AD it is desirable to spread these "two lines" over the 10-cm full-width scale. The time base is therefore adjusted to 12.7 microseconds/cm by placing the sweep time to 10 microseconds/cm, and the multiplier to 1.27, resulting in two lines at a horizontal rate across the graticule within the 10-cm scale.

The video picture waveform, contrary to what is expected, does not display a single line of varying 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 impossible *unless* the camera were continually 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 of the human eye, as well as the fact that all the lines of the picture are traced out on top of each other during a single sweep of the scope beam, results in a pattern similar to that shown in Fig. 1-13.

The line appearing at the sync-tip level is also 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 V 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 V pedestal is some 13 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 lines at the sync tip level are the V sync pulses placed atop the V pedestal. They are of longer duration than the H sync, and they are serrated in form.

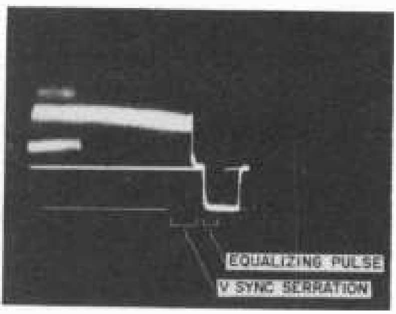


Fig. 1-14. Expanded horizontal sweep showing details of vertical sync serrations and equalizing pulse coincident with horizontal sync.

Notice that the equalizing pulse is also visible, although the trace produced in very faint. This is due to the fact that the pulse is at a high frequency (31.5 kc) relative to the time base, and exists only at 60-cps intervals. Fig. 1-14 shows the equalizer pulse more clearly due to the expanded sweep. The leading edge of *alternate* equalizing pulses coincides with the leading edge of a horizontal sync pulse. Since the width is one half that of H sync (Section 3 details this waveform), it appears to split the H sync pulse in half. The leading edge of alternate equalizers *do not* coincide with H sync, but they are coincident with serrations of the vertical sync pulse, as shown in Fig. 1-13. Note, then, that the equalizer *coincident* with H sync exists only at 30 cps (every other pulse) which explains the dim trace.

When the monitor sweep is set at 30 cps (33,333 microseconds), the *field frequency* waveform is displayed as in Fig. 1-15. A single sweep now takes place in 1/30 second, the lines of one field are traced out to the left of the V pedestal, and the following field is traced out to the right. The line at the extreme bottom is composed of the *horizontal* sync pulses for the lines comprising 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 V sync pulses are also on this line immediately under the V pedestal, but they are hardly distinguishable as such due to their short time duration in ratio to the scope sweep of 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 above fields.

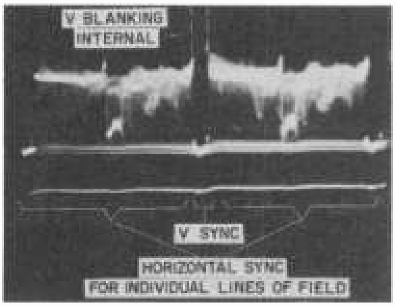


Fig. 1-15. CRO vertical-rate display of two fields.

Note from the preceding description and Fig. 1-13 that conventional horizontal-rate scope sweep results in a pattern which contains all the lines (at a horizontal rate) of the field. It is, however, possible to observe only a single line of the field on the Tektronix 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 (see Fig. 1-16). This sweep is obtained internally from the composite television signal by establishing

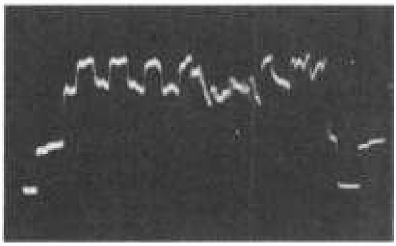


Fig. 1-16. CRO display of single line by line-selected delayed sweep with time base of 6.35 microseconds per centimeter.

a coarse time delay from a vertical sync pulse (off 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/cm, a single line will occur in the full-scale 10 centimeters of the scope graticule. (Fig. 1-16 illustrates this feature.)

When using the Sweep Delay control in this manner, the sweep is triggered only 30 times per second. The resulting display is correspondingly dim and should be viewed through the viewing hood provided for this purpose when much ambient light exists.

The photo of Fig. 1-16 required 1/2 second exposure of 3,000 ASA polaroid film (at f:5.6) compared to 1/30 second exposure for the other photos.

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 scope. The picture on the monitor is brightened during the time of the sweep gate as determined by the selected time base. The Sweep Delay control is rotated while watching the monitor until the desired line of the picture signal is selected. Thus, the amplitude of test-chart bandwidth wedges may be measured relative to grey (100%) areas.

To emphasize the fact that verticals are apparent in horizontal-rate CRO displays (and vice versa), the upper trace in Fig. 1-17 shows the appearance of 60-cycle hum in the video signal viewed

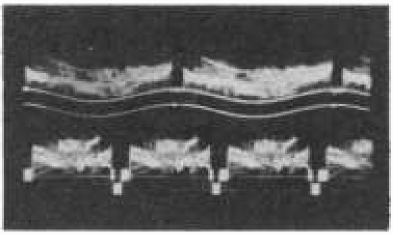


Fig. 1-17. Upper trace is vertical-rate display of signal with large 60-cycle hum component. Lower trace is horizontal-rate display of same signal.

at the vertical rate. The lower trace in Fig. 1-17 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 vertical "tilt' 'or shading; the horizontal sync tip and porches will be thickened an amount proportional to the vertical defect.

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

SECTION 2

AMPLITUDE CALIBRATION AND MAINTENANCE OF LEVELS

After several years of television broadcasting, the standards of video level measurement are still somewhat uncertain. This situation is analogous to that which existed in aural broadcasting before the standardization of the VU meter. Video level measurement is more complex than measurement of the audio portion since the ratio of several different signal levels is involved. Therefore waveform presentation on an oscilloscope is required.

2-1. REQUIRED SYNC-TO-VIDEO RATIO

Fig. 2-1 represents the transmitter carrier wave modulated with a standard window signal. Sync tips are 100% of carrier, blanking level 75%, and peak white 12.5% of maximum carrier. This "white setup" is fixed by FCC standards at 10 to 15% of carrier (12.5%

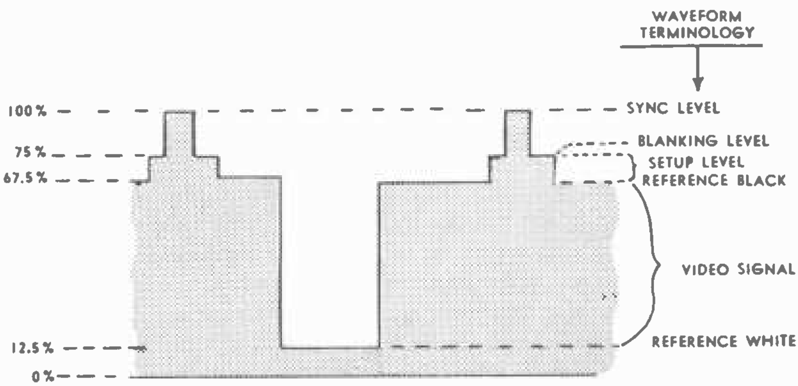


Fig. 2-1. Transmitter carrier representation when modulated with standard window signal.

nominal) to avoid carrier cutoff. The latter results in audio buzz in intercarrier-type receivers due to loss of the carrier frequency reference.

It is evident that the signal amplitude from the studio line is represented on the modulated carrier as that portion between 12.5 and 100% of maximum carrier values. Since full modulation of the transmitter occurs at 87.5% of carrier, the necessary sync-to-video ratio of the input signal to result in 25% sync is 25/87.5 = 0.286, or 28.6% of the total composite signal. Thus, disregarding special circuits such as sync stretching stages of transmitters or stabilizing amplifiers, the input signal (to a linear transmitter) must be 71.4% video and 28.6% sync to obtain the FCC requirement of 25% sync in the radiated signal.

Notice also that, on recovery of the video signal by detection of the transmitted carrier, *exactly the same ratio of video to sync* should exist; i.e., 71.4% video and 28.6% sync. This represents a modulated carrier ratio of 75% video and 25% sync. (Discussed more fully in Section 8 on transmitter proof-of-performance.)

2-2. TERMINOLOGY AND STANDARDS OF VIDEO LEVELS

The original standards for studio line output (transmitter input) set in 1946 established a 2-volt composite signal level comprised of 0.5-volt sync and 1.5-volt video, or 25% sync to 75% video. All early transmitters employed sync stretching circuits to properly adjust the respective amplitudes and to compensate for the inherent sync compression.

In 1950 this standard was changed to a composite level of 1.4 volts; 1-volt video to 0.4-volt sync. This change was largely because of difficulty in obtaining good amplitude linearity in existing equipment over the 2-volt range (particularly in a large number of cascaded amplifiers, as in network transmission). This new standard not only established better amplitude linearity characteristics, but it also provided a compatible 28.6% sync to 71.4% video ratio. Some transmitter manufacturers deleted the usual sync stretching circuits from transmitters and obtained proper sync/video ratios in external stabilizing amplifiers.

Since the advent of color television, many stations (even those operating monochrome only) have established a 1.0-volt composite signal as a standard for the line output level. Because of the overshoot of the color subcarrier on color-bar transmissions, amplitude linearity problems again manifested themselves for the 1.4-volt standard. Numerous tests indicated the desirability of reducing the transmission level to 1.0 volt composite, and this is now standard practice for AT&T television operating centers and the majority of commercial broadcast stations.

The new 1.0-volt standard maintains the same sync/video ratio as the older 1.4-volt standard. The voltage ratios of the old and new standards, as correlated with the IRE scale (adopted for use as an industry-wide standard), is shown in Fig. 2-2. Although the 1.0-volt signal is normally spoken of as 0.7-volt video to 0.3-volt sync), the actual voltage values are 0.714 video to 0.286 sync. By calling all values in IRE units, a standard is established which

For 1.4 volt signal: volts per unit = 1.4/140 = 0.01v/unit
For 1.0 volt signal: volts/unit = 1.0/140 = 0.00714/unit

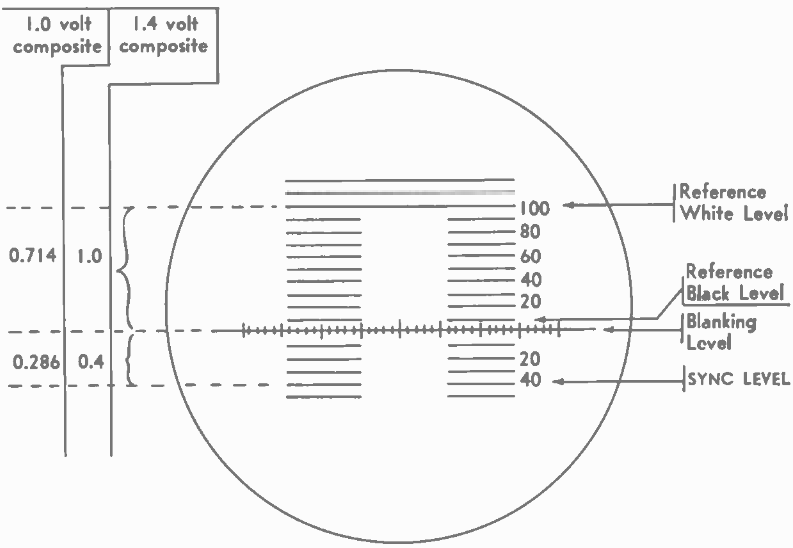


Fig. 2-2. Standard IRE scale showing correlation of 1.0-volt and 1.4-volt signals.

eliminates any confusion. It is only necessary to calibrate the scope gain so that 140 IRE units is 1 volt peak-to-peak. For the old standard, the scope is calibrated for 1.4 volts peak-to-peak on 140 IRE units.

Fig. 2-3 illustrates the three scales recommended for the various points to be monitored. Operating scale No. 1 is for use at the camera control unit when sync *is not* added at this point. Reference white is at 100, reference black is at 10, and the blanking level is at 0. It is noted that the reference black level at 10 is a continuous line as is the blanking level at 0. This *set-up* level is very important. In a theoretically perfect transmission system, black level and blanking level could be maintained the same, thus utilizing to the fullest extent the video amplifier gains. In practice, however, some amplitude distortion exists, resulting in at least slight amounts of overshoots in the black region. When this occurs, some retrace lines will be visible on the picture, unless the receiver controls are adjusted to clip black peaks. This results in *compression of blacks*. By raising the set-up value to about 10%, optimum operating conditions are realized. The setup must be constant and of the same value between cameras (studio or film), studios, networks, etc., to maintain constant background brightness in the home receiver. (NOTE: Some stations employ a 5% set-up level.)

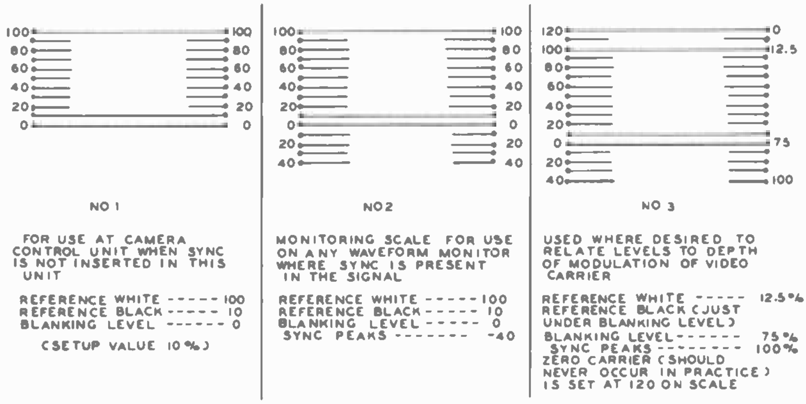


Fig. 2-3. Video level measuring scales recommended by IRE Standards (black negative polarity).

Operating scale No. 2 in Fig. 2-3 is recommended where the composite signal level (sync added) is to be measured. Reference white is at 100, reference black is at 10, the blanking level is at 0, and the sync peaks are at −40. This type of scale is common at the line monitoring position.

Operating scale No. 3 is recommended at transmitter locations where it is desirable to relate the arbitrary IRE units to depth of modulation of the video transmitter carrier wave. Reference white of 100 represents 12.5% carrier, zero carrier being represented by 120 on the scale. Reference black is at 10, and blanking level at 0 represents 75% carrier. Sync peaks at −40 represents 100% carrier modulation. The relationship to the FCC specifications of carrier levels may now be observed. Zero carrier level (which should never occur in practice) is set opposite 120 on the IRE scale, and maximum carrier opposite −40. Blanking level (zero on scale) then occurs at 75% of maximum carrier, and reference white (100 on scale) occurs at 12.5% of maximum carrier.

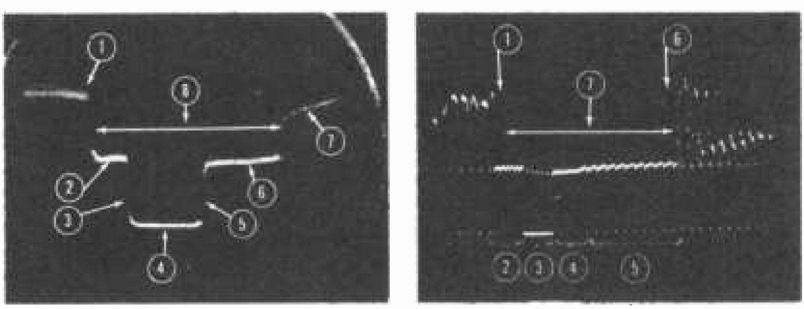
It is becoming increasingly important that all TV engineers "talk the same language." Fig. 2-4 illustrates two scope displays and the descriptive terminology associated with their analysis.

This terminology, as well as that included in the Glossary of Terms Concerning TV Waveform Levels at the rear of this section, is approved by AT&T and local telephone companies. Terminology used in describing specific troubles is presented where appropriate in future sections.

It is common practice to call out levels from blanking toward white and blanking toward sync as "100 over 40," "85 over 30," etc., which refers to IRE units in each direction. 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 IRE units on the scope and read each burst frequency in IRE units occupied by the individual burst. Reference white is established by a pulse immediately following blanking for use in this adjustment.

The set-up 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 IRE 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% setup at the line output stabilizing amplifier. (NOTE: the stabilizing amplifier is discussed in Sections 3, 5, and 6.)

All level checks not involving frequency response (such as keyed burst signals) should be made with the scope response on the IRE position. If the scope is on wideband response, the small-



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) Definition of terms, expanded 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) Definition of terms, expanded vertical.

Fig. 2-4. Analysis of expanded sweep displays.

energy overshoots of the higher frequency components will be apparent. If these overshoots are held below 100 IRE 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 AT&T.

Since the waveform monitor of the master unit is often used to indicate the final video level as it leaves the studio for the transmitter, the operator will find a choice of two pattern characteristics associated with this circuit. For example, the RCA TM-6 master monitor incorporates a variable vertical amplifier response for the CRO waveform oscilloscope. In the wideband position, the response is essentially flat to 5 megacycles. When used as level measurement only, the response is "rolled-off" at the high end, as illustrated in the IRE curves of Fig. 2-5. The solid line indicates the recommended roll-off, with the dashed lines showing allowable tolerances in the amplifier response for level measurements.

NOTE: This is the old IRE curve adopted in 1950, which is still used in many instances. It should be modified to the new curve of Fig. 2-6, which was adopted in 1958 and amended in 1961.

Referring to Fig. 2-5, 1 is the normal appearance of the waveform displayed on a wideband scope when the signal input is from a monoscope camera. Notice that this pattern can indicate the quality of a signal as well as the level in amplitude, showing distribution of lights and darks in the pattern and good shading characteristics. Display 2 is the same monoscope signal with artificially produced overshoots (transients) as viewed on the same wideband scope. It is recalled that these characteristics lead to major differences in handling overall gain by different operators. As a result of a study from which the IRE standard was compiled, it was found that such roll-off characteristics reduced disagreements among operating personnel concerning levels, and still provided sufficient indication of overshoot so that an undue amount would be apparent. Display 3 in Fig. 2-5 is the pattern produced from signal 1 on the narrow bandwidth CRO. Note that, since highs are reduced in response, the entire pattern appears to be of the same intensity. From the same characteristics the overshoots of 2, when displayed as in 4, are not as apparent (except in the "black" region), and a more uniform level indication results.

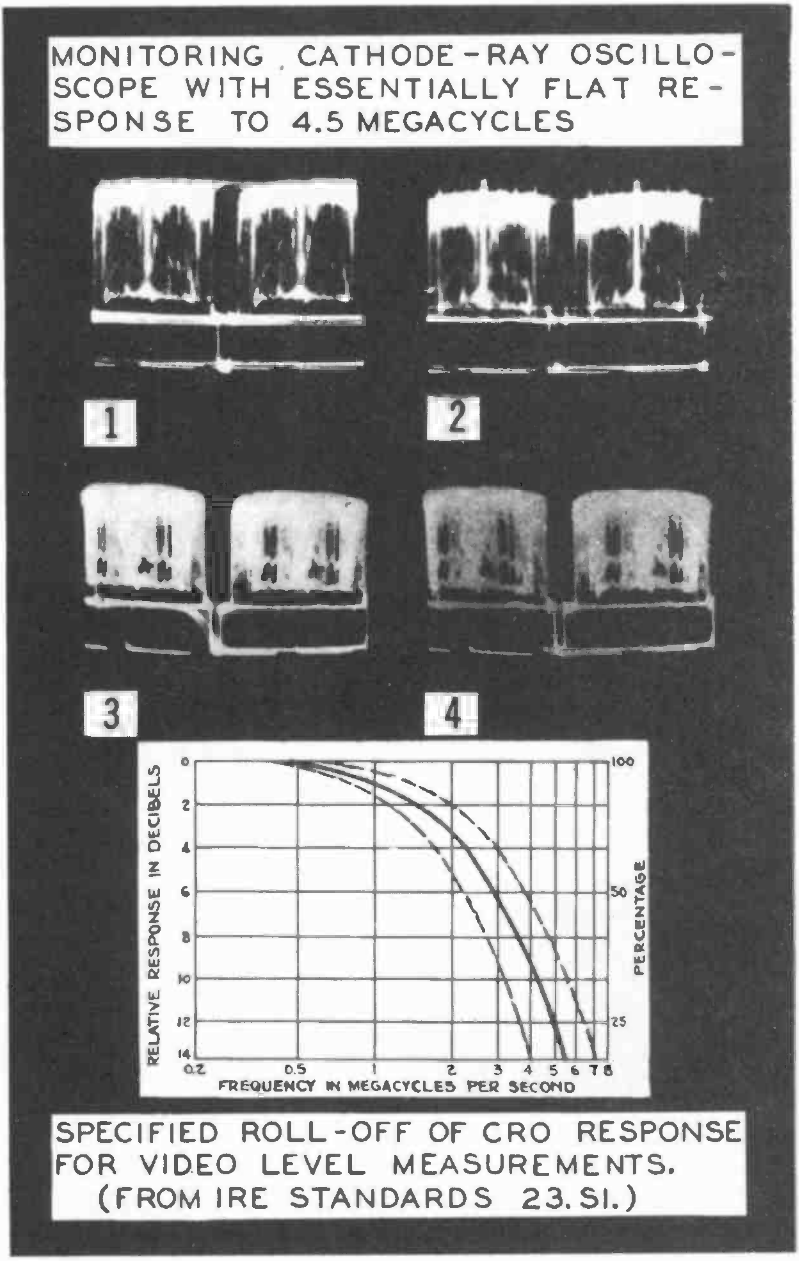


Fig. 2-5. Monoscope camera signal waveform on CRO with different response characteristics. Response curve at bottom is old IRE standard of 1950, still used by many stations.

In installations where such facilities are provided, the maintenance department should be charged with the responsibility of periodically checking this roll-off characteristic of the monitor. From the new curve in Fig. 2-6 the amount of roll-off that should occur at 3.58 mc is approximately 20 db. This was approximately 8 db on the old curve (Fig. 2-5). The new curve is more suitable

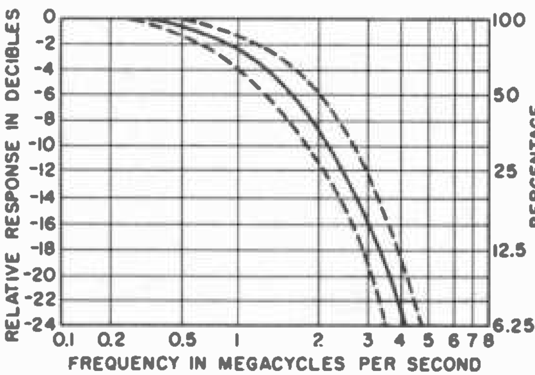


Fig. 2-6. Frequency response for standard oscilloscope (IRE roll-off).

for luminance control of composite color signals. This response characteristic may be conveniently measured by means of the sweep generator method, using the scope on the monitor to display the response, as shown by Fig. 2-7. In actual practice a slight rise may occur in this response around the "crossover" frequency (approximately 4 mc); it should remain at least 20 db down above 5 mc. The IRE response circuit should be adjusted to meet

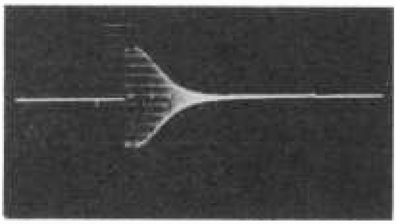


Fig. 2-7. CRO display of the new IRE response characteristic.

the specifications of Table 2-1, which are the exact values spelled out in the 1961 revision of the curve of Fig. 2-6, narrowing somewhat the tolerance in response. There should be no sudden break in the response curve above 3.58 mc, and the maximum response above 5 mc should be at least 20 db down with respect to that of 100 kc, the reference frequency. The scope should indicate less than 2% tilt in a 60-cps symmetrical square wave.

2-3. FACTORS AFFECTING LEVEL VARIATIONS

Level variations are classified into two major groups—short-term and long-term variations. The short-term grouping may include anything from the rapid "bounce" observed on scenic content changes between radical differences in duty-cycle (or unclamped points of observation, or unregulated power supplies) to variations occurring in a single program from laxity in "gain riding" or variations in distribution amplifiers operating at fixed gain (usually unity). Long-term variations (usually considered a week or more) are not troublesome, provided the maintenance department is aware of the required frequency of level checks for the particular installation.

Table 2-1. Frequency Response for Standard Oscilloscope (IRE Roll-Off)

Frequency (mc)	RELATIVE RESPONSE (db) Center	Lower Limit	Upper Limit
0.2	—0.1	—0.2	0.0
0.5	—0.7	—0.9	—0.5
0.7	—1.2	—1.6	—0.8
1.0	—2.5	—3.1	—1.9
1.5	—5.3	—6.4	—4.3
2.0	—8.8	—10.1	—7.5
2.5	—12.4	—13.8	—11.0
3.0	—15.9	—17.9	—14.4
3.6	—20.0	—22.5	—18.5
4.0	—22.7	—25.3	—21.2
Rise time approximately 0.3 microsecond.			

Proper "gain riding" of video levels on network feeds (where the operator may be uncertain as to whether the scene contains "reference white") is materially aided by a calibration pulse keyed in for a one- or two-line duration during the vertical blanking interval. The upper trace in Fig. 2-8 is the appearance of the ABC network transmission at normal vertical-rate sweep. The lower trace in Fig. 2-8 is expanded to show more clearly the one-line interval which occurs on the nineteenth line of vertical blanking.

Rapid variations in level that occur when switching between sources are often due to "hitting" the input coupling capacitor of a distribution amplifier with DC, as would occur with a leaky output capacitor from the source being switched. This effect is most pronounced with amplifiers employing heavy DC negative feedback circuits to obtain low line driving impedances over a wide bandwidth.

Most video amplifiers using several negative feedback loops in cascade, or overall feedback, give no indication of a drop in gain when weak tubes are present. Low emission tubes, or tubes with high-resistance internal element shorts, can, however, aggravate very short-term level changes. It is desirable to establish routine tube checks at 60- to 90-day intervals on such equipment or to run video sweep checks which normally provide a clearer indication of tube condition than tube testers can reveal.

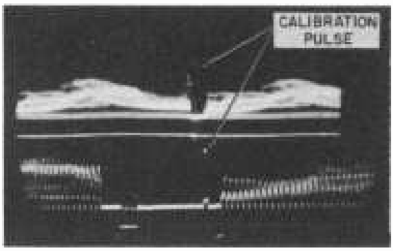


Fig. 2-8. Upper trace shows calibration pulse representing 1 volt peak-to-peak (normal vertical sweep). Lower trace is expanded sweep showing pulse on 19th line of vertical blanking.

When video AGC amplifiers are used, a check for proper operation should be made at least on a monthy basis. A satisfactory procedure can be outlined as follows:

1. Check any gas voltage-regulator tubes for regulation on a reliable tube tester. If any check outside regulation limits, replace them with new tubes that do check within limits. Gas regulators are usually used to supply screen voltage regulation for variable-gain stages where AGC action occurs.
2. Check all set-up controls as outlined by the manufacturer's instructions for the amplifier used. When the range of any gain controls needs to be increased to obtain proper operation, check for bad tubes. The Good-Bad scale is sufficient for transconductance checks, since this allows the normal manufacturer's tolerance for the particular tube being tested. In addition, check for shorts and gas.
3. AGC amplifiers usually operate with either a 3-, 6-, or 12-db control range. Assuming a 6-db control range, check as follows:
   (a) Reduce the input 6-db (50% on voltage scale). The output should remain constant within 0.5 db and fall off as the input is further reduced.
   (b) Increase the input 6 db over normal. The output should again remain constant within 0.5 db.
   (c) Try rapid variations of level within this range. No more than ±0.5-db variation should occur. Most critical to proper operation are any adjustments affecting bridge-type balancing networks, reserve gain for AGC control,

and the conduction threshold centering of the AGC rectifier.

Notice that for Step 3a, a 3-db control range AGC would require the input to be reduced to 70% on the voltage scale; the 12-db range would require a reduction to 25% of normal on the voltage scale.

2-4. CHECKING REGULATED POWER SUPPLIES

The principle of a vacuum-tube regulator is shown in Fig. 2-9. To review this function:

1. If the load draws more current or if the AC input to the rectifier section falls, the result would normally be lower terminal voltage.
2. Resistor R1, tube V2 and gas regulator tube V3 are in series across the rectifier filter output. V3 holds the cathode of V2 at a constant positive potential with respect to ground. The setting of R2 (voltage adjust control) determines the bias on the DC amplifier (V2).

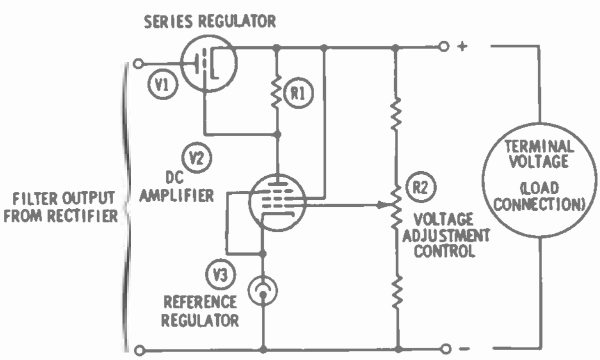
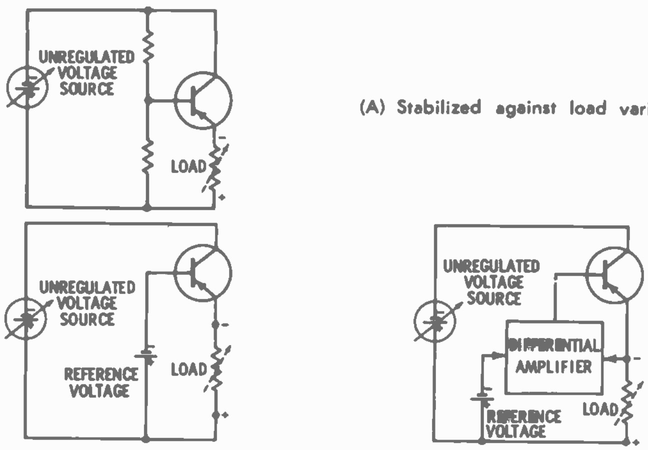


Fig. 2-9. Basic vacuum-tube regulator circuit.

3. A reduction in terminal voltage results in a more negative bias on V2, less current through V2, and hence less current through R1.
4. The decreased IR drop across R1 results in less negative bias on the series tube (V1). This, in turn, results in lowered series resistance, hence it increases terminal voltage by the amount decreased in the preceding Step 1.

Most regulated power supplies of recent vintage employ transistors and (in some cases) zener diodes. A review of transistorized regulated power supplies is assisted by the drawings in

Fig. 2-10. The circuit shown in Fig. 2-10A will probably be recognized immediately as 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 whose impedance is automatically controlled in such a way that it tends to compensate for the impedance changes (or current changes) in the load, thus maintaining an essentially constant voltage across the load. This action may be explained by noting that the voltage drop across the emitter-to-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



(A) Stabilized against load variations.

(B) Stabilized against both load and source variations.

(C) Further stabilized by error gain (amplified correction).

Fig. 2-10. Principle of transistorized voltage regulators.

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 significantly altered 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 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 in passing that a regulating transistor of the PNP type 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 is frequently 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. 2-10A is reasonably effective in stabilizing the output voltage against load variations, it does not remove variations due to 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. 2-10B 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 normally be a *reference diode*, (zener diode) which is a semiconductor diode with enough reverse bias to operate in the so-called "breakdown region." A diode operated in this manner behaves very much like the familiar 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. Reference diodes are preferable to gaseous voltage-regulator tubes for most transistorized power supplies, because they operate at lower voltages (usually of the order of 5 to 6 volts on up to 60 volts or more) and because they are generally superior in stability and inherent regulation.

The degree of regulation attainable with the circuit in Fig. 2-10B 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. 2-10C. 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.

A graph of the current through and the voltage across a reverse-bias junction diode (zener diode) is shown in Fig. 2-11. At a certain value of reverse-bias voltage the current rapidly increases while the voltage across the diode remains essentially constant. This "breakdown voltage," which may be anything between 2 and 60 volts or more, depends on the construction of the diode. This characteristic is similar to that of the gas tube regulator which begins conduction at a certain voltage and continues to conduct varying amounts of current while maintaining constant voltage across the elements. The zener diode is used in transistorized regulated power supplies to hold an element of the transistor at a given reference voltage.

The maintenance of regulated power supplies is extremely important to overall video-level stability. The following four tests enable the maintenance engineer to keep a running check on the condition of his regulated supplies.

Test 1 To determine the voltage output range at fixed load. Use a fixed load that will draw at least two-thirds of the maximum rated load current. Rotate the Voltage Adjust control to its extremes and record the minimum

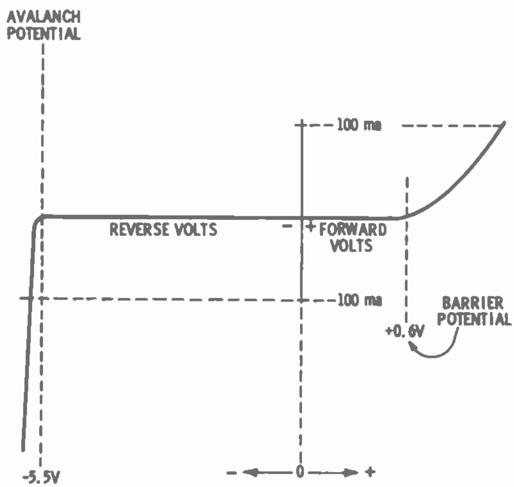


Fig. 2-11. Basic curve of 5.5-volt zener diode.

and maximum voltages. For example, the normal available range of a 280-volt regulated supply might be from 270 to 300 volts, at a given load current. Failure to reach the normal maximum voltage is usually the result of a weak DC amplifier tube or voltage adjust tube (or transistor).

Test 2 In vacuum-tube regulators check series regulator tube currents for balance. Most regulated supplies incorporate a Meter Selector switch on the panel for measuring individual regulator-tube currents. Chart 2-1 shows the application of such readings. Notice that the total load in this example is 1014 milliamperes; therefore, the ideal average for each of the six tube sections is 1014/6 = 169 ma. Since maximum tube life and stability can be expected when these currents are balanced within ±10% (20% total variation), a record is kept of individual currents as shown, and it is compared to the minimum and maximum values which should occur for the given load. This indicates the need for a tube change before trouble occurs, barring any sudden failure.

Chart 2-1. Tabulated Data for Test No. 2

SERIES TUBE CURRENTS (ma)		*OPERATING DATA*
V1A	168	
V1B	160	Total = 1014 ma
V2A	180	Average/Section = 169 ma
V2B	182	Lowest Desirable = 152.1 ma
V3A	164	Highest Desirable = 185.9 ma
V3B	160	

Test 3 Input voltage regulation (voltage output with fixed load and varying input AC line voltage). The setup for this and the following test is shown in Fig. 2-12. Adjust the power supply to be tested to 0.5 volt above the reference supply, and connect a voltmeter between the two outputs to measure this voltage difference. By means of the

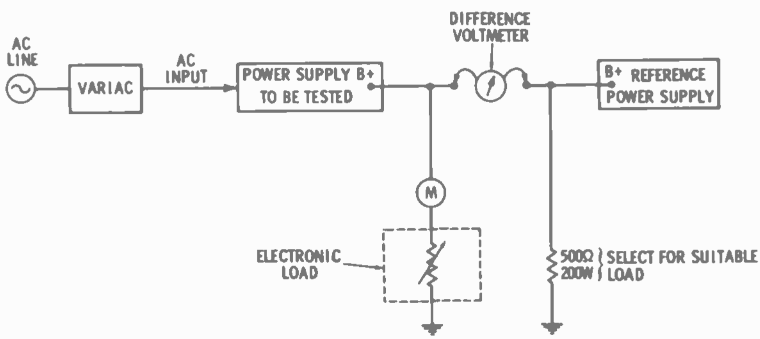


Fig. 2-12. Test setup for checking power supply regulation.

Variac, make measurements at the reference line voltage (usually 117 volts), then record the measurements over a specified range, such as 100 to 130 volts AC input. Chart 2-2 is the data recorded at station WTAE for an RCA WP-15B supply, showing the excellent input voltage regulation of this supply with a fixed load.

Test 4 Output voltage regulation under variable loads (fixed AC line voltage with varying load current). In this test the *Variac* is fixed for an AC input of 117 volts, and the

Chart 2-2. Tabulated Data for Test No. 3

LINE VOLTS	*VO*	*ACTUAL VOLTS* VARIATION
117(ref.)	280.5	0(ref.)
100	280.94	+ 0.44
105	280.75	+ 0.25
110	280.62	+ 0.12
115	280.5	0
120	280.42	− 0.08
125	280.37	− 0.13
130	280.31	− 0.19

electronic load on the supply under test is varied over a specified range. Chart 2-3 is the corresponding data for the WP-15B, again indicating extremely good regulation under varying loads, and very low internal power-supply resistance.

Chart 2-3. Tabulated Data for Test No. 4

LOAD CURRENT (ma)	*VO*	*ACTUAL VOLTS VARIATION*
1000(ref.)	280.54	0(ref.)
400	280.52	− 0.02
600	280.5	− 0.04
800	280.52	− 0.02
1200	280.54	0
1400	280.56	+ 0.02
1500	280.51	− 0.03

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,
ΔV_o is the change in output voltage,
ΔI_L is the change in output current.

For example, a voltage output changes 0.1 volt with a load current change of 1,000 ma (1 amp):

$$R_o = \frac{0.1}{1} = 0.1 \text{ ohm}$$

The condition of power-supply filters and general regulation efficiency should be checked at least several times yearly by observing the ripple content of the output voltage on an oscilloscope. An increase in ripple content will indicate the need for filter replacement or better regulation efficiency before reaching troublesome proportions. Typical commercial regulated power supplies have a maximum of 2 millivolts (p-p) ripple content on a 280-volt regulated output. The ripple on an unregulated voltage may be as high as 2 volts (p-p) on a 400-volt DC output.

2-5. HOW TO CHECK DIODES

The operating condition of zener diodes (or any other type of diode) is most reliably checked with an osciloscope and the simple associated circuitry shown in Fig. 2-13. The upper trace in Fig. 2-

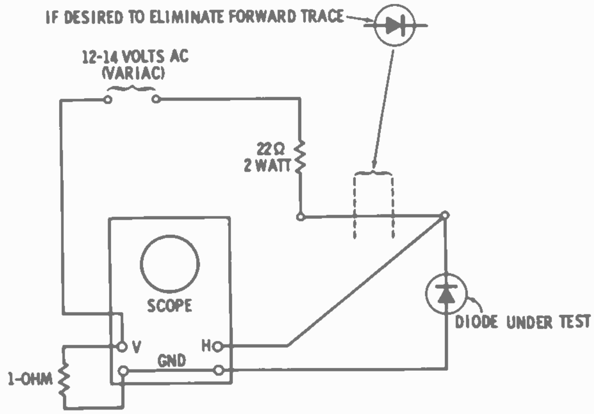


Fig. 2-13. Test setup for checking zener (or any other type) diode.

14 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 dotted lines in Fig. 2-13, which results in the lower trace shown in Fig. 2-14.

Fig. 2-15 is an interpretation of the trace in Fig. 2-14. As the *Variac* voltage output is increased from zero, the voltage is traced horizontally (A-B) along the scope graticule which may be calibrated in volts/cm. When the zener breakdown voltage is reached, the horizontal trace should remain the same length as the current curve increases. If desired, the vertical scale (B-C) may be calibrated in milliamperes/cm. Care should be taken not to exceed the maximum current specification (wattage) of the zener diode being tested (voltage applied across the diode times the diode current).

The same test circuit should be employed to check regular diodes 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. 2-16. This loop should remain stable without jitter or erratic "looping" as the voltage is varied around the normal operating level. Other

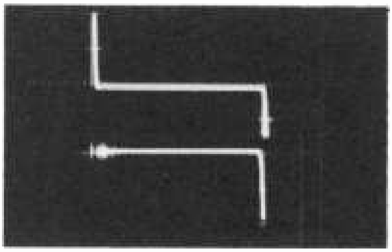


Fig. 2-14. Upper trace shows CRO display of zener diode response; in the lower display the forward trace is eliminated.

diodes (such as the type 1N279) do not reveal a loop, as shown by the lower trace in Fig. 2-16. There should be no instability of trace as the voltage is varied around the normal operating level.

2-6. POWER DISTRIBUTION

To prevent emergency maintenance, and for maximum flexibility, power-supply distribution is highly recommended. A close-up view of one typical installation is shown in Fig. 2-17, with the schematic illustrated in Fig. 2-18. Load receptacles are at the top,

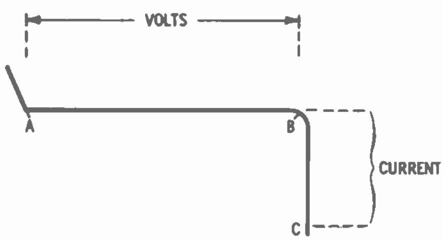


Fig. 2-15. Interpretation of the trace in Fig. 2-14.

and power-supply receptacles are at the bottom. A receptacle for a spare power supply is in the center of the bottom row (Fig. 2-17). Two sets of barrier terminals intercept each circuit, one in the power-supply rack and the other in the load-equipment rack. Each power supply terminates at a Jones receptacle on one of the three power-supply patch panels; one panel is used for each type of supply (i.e., WP-15, WP-33, and 580-D). Each receptacle marked Power in the bottom row is connected to a 580-D supply;

nine of these are in regular use, and the tenth is a spare. There is likewise one spare WP-15 and one spare WP-33 on each of the other panels. In all panels the spare is located in the middle of the row to keep the length of the patch-cords to a minimum.

The upper row of receptacles shown in Fig. 2-17 is connected to an amplifier or group normally fed by a 580-D supply. Each normal supply is connected to its normal load by the patch-cord.

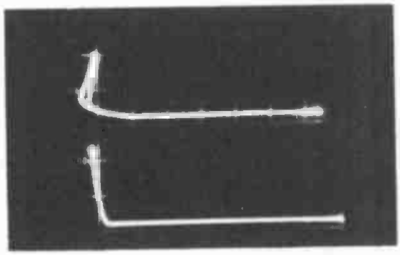


Fig. 2-16. Upper trace shows characteristics of 1N34 diode; the lower display is for a 1N279 diode.

Deep-bracket receptacles and plugs are used to prevent accidental removal. The pins are male on the power-supply receptacles and female on the load to prevent hazard of shock or accidental grounding while patching. Primary AC is wired through the

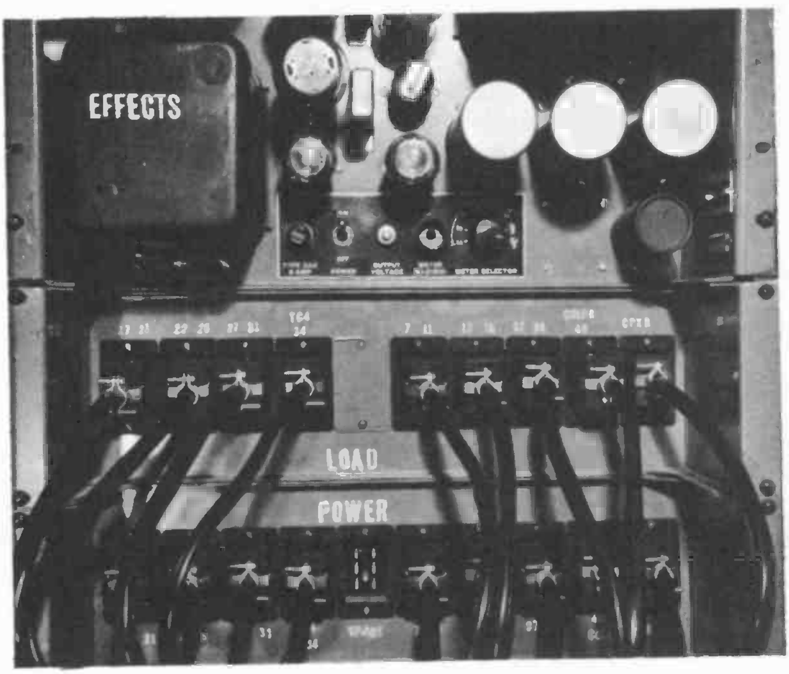


Fig. 2-17. Typical power-supply patch panel as installed at WISH-TV.

receptacles to each power supply so that the supply can be energized immediately.

A spare power supply may be put in service in a matter of seconds with this system, and the patch-panels have literally paid for themselves many times over.

The barrier terminals in this particular arrangement are integrated in the system to permit easy reassignment of equipment when required by changes in the system or relocation of equipment in the racks. In addition to this feature, they are advantageous as test points when troubleshooting. The presence or absence of voltage at a barrier terminal immediately determines whether trouble is in the amplifier or in the power supply.

Fig. 2-18 shows a typical circuit from power supply to equipment—a WP-33 power supply which normally feeds TA-5D stabilizing amplifier No. 6 with a remote-control position at the master control desk. All wiring in the WISH-TV plant is color-coded, which further assists in isolating circuits. Fig. 2-18 further illustrates the manner of handling centering voltage, which is supplied by a WP-33 power supply, however not used with a load as shown. A strap is inserted on the load receptacle that grounds the centering voltage. On pieces of equipment requiring centering voltage (master monitors or cameras) no strap is employed. This method of wiring permits the use of any power supply for any load of comparable capacity. This jumper strap performs the same function as the jumper across C3 in the WP-33 power supply.

Fig. 2-19 is a schematic of the power-supply distribution panels employed at WTAE in Pittsburgh. All power supplies are RCA type WP-15's which supply a total of 1.5 amps of regulated B+ voltage. In this arrangement a spare power-supply output may be plugged into the receptacle on the distribution panel in place of the normal supply in case of failure. Each circuit is separately fused, and neon indicating lamps show failure of any fuse.

2-7. GLOSSARY OF TERMS CONCERNING TV WAVEFORM LEVELS

The following terms are currently used in operation and maintenance, and they are all approved by the Long Lines Division of the American Telephone & Telegraph Company for use in television customer relations. Only those terms concerned with normal or abnormal levels are defined herewith. Terminology of specific types of troubles in waveforms will be covered where applicable in future sections.

NOTE: See Fig. 2-20 for specific application of terminology applied to video levels.

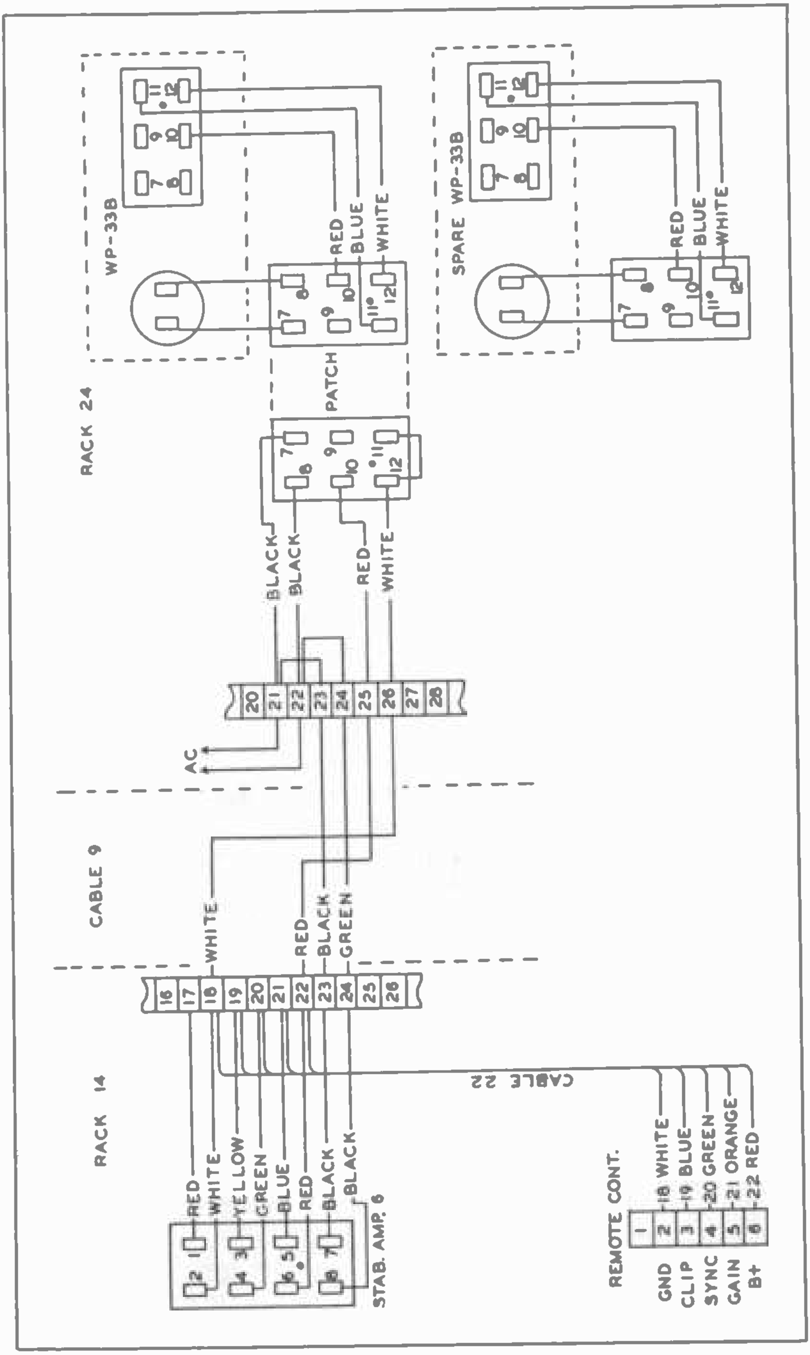


Fig. 2-18. Wiring diagram of the WISH-TV power supply patch system.

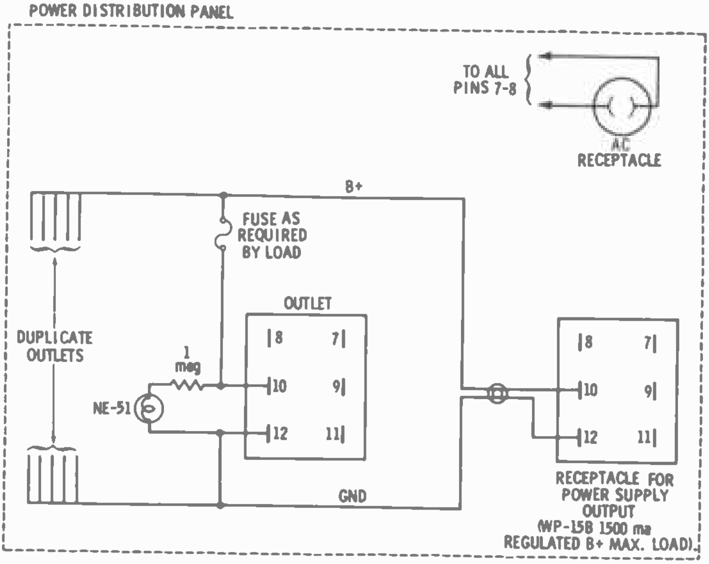


Fig. 2-19. Example of power supply distribution at station WTAE.

black peak—The point of maximum excursion of the picture signal in the black direction at the time of observation.

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.

bounce—Sudden variation of level (hence brightness) of the picture signal.

breathing—Amplitude variations similar to bounce but at a slow regular rate.

clipping—Sharp cutoff of the video signal peaks. May affect either the white (positive) peaks or the black (negative) peaks. The sync amplitude may be affected on a composite signal.

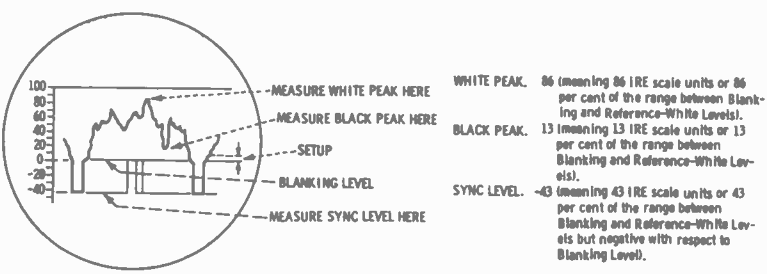


Fig. 2-20. Use and interpretation of IRE standard scale.

compression—An undesired decrease in 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," which is 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 the standard polarity for the transmiter input to produce a positive image to the viewer. 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 in level between blanking and reference black levels.

sync level—The level of the tips of the sync pulses.

video-in-black—Condition where 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.

SECTION 3

SYNCHRONIZING GENERATOR AND PULSE DISTRIBUTION SYSTEMS

The sync generator consists of two basic sections—timing and pulse shaping. The simplest timing relationship which can be used is a 15,750-cps oscillator for the "line" frequency and a 60-cycle pulse derived directly from the power line for the field frequency. This form of "timing generator" is employed in a number of closed-circuit industrial systems since the signal is not intended for regular broadcast. However, in order to produce the field interlace required for standard broadcast transmission, it is necessary that the field frequency be derived from the master oscillator frequency. When two independent generators are used (as in the industrial system), a form of "random interlace" results, since the frequency relationships are not locked. Also, the actual number of lines constituting a single frame will vary with the magnitude of drift between the line and field frequencies.

3-1. WHAT THE SYNC GENERATOR MUST DO

Fig. 3-1 serves as a "capsule" review of what the sync generator must do, and it should be referred to during the following discussion.

From the familiar waveforms of (N) and (P) note that field 1 has a full line (H) to the first equalizer pulse while field 2 has a half-line interval. This is the requirement for odd-line, interlaced (two-to-one) scanning, and the master oscillator must, therefore, be twice the line frequency so that either H or ½-H pulse are available. (NOTE: field 1 and field 2 are arbitrarily numbered. In this description, field 1 refers to the full-line interval before the 9-H vertical time, and field 2 refers to the ½ line as shown. Also remember that field 1, therefore, has ½-H spacing from the last equalizer pulse to the first H sync pulse, which is now field 2, while field 2 has H spacing from the same point in time, new field 1.) This characteristic is useful in setting the vertical blanking duration described later.

A in Fig. 3-1 represents the 31.5-kc triggers from the master oscillator, initiating the leading edge for B, C, D, and F. The width of pulse B determines the front porch interval since the trailing edge of this pulse "times" the H sync information (waveform E). Some generators use a tapped delay line for this purpose.

Although a considerable difference exists in methods used by various manufacturers of sync generators, the timing will be as shown in Fig. 3-1. The master-oscillator frequency, which is divided by 525, triggers the leading edge of vertical blanking, vertical drive, and the various gates shown. Delayed triggers (B-prime) still provide the coincidence timing of H sync during the long 9-H vertical interval.

The importance of timing accuracy and allowable tolerance may be realized from the following examples. Consider the observation of a thin vertical line representing a frequency of 4 mc in the picture. If the timing on alternate lines of the raster is shifted by as much as one-half cycle, fine detail is lost. Since a complete cycle of 4 mc occurs in 0.25 microsecond, a half cycle occurs in 0.125 microsecond. This means that the *overall* tolerance (sync generator and monitor or home receiver) in variation of timing between successive horizontal pulses is 0.125 microsecond for a 4-mc detail. This is roughly an accuracy of 1 part in 600 for overall H timing. For the sync generator alone EIA standards allow approximately 0.008-H in an averaging process of 20 to 100 lines, or about 0.05 microsecond (less than one-half of 0.125 microsecond).

The tolerance of vertical timing is even more severe. Since only one-half line difference exists for alternate fields, complete loss of interlace occurs with a vertical timing error of ½ H (32 microseconds). In practice the overall vertical difference in line spacing must be less than 10% to preserve the illusion of perfect interlace. This is approximately 3 microseconds out of the vertical interval of 16,667 microseconds, or less than 1 part in 5,000.

Before leaving the basics of interlace, it is pertinent to very briefly review the real function of equalizing pulses. They do not, as the name implies, equalize the charge on the vertical sync integrating capacitor between the H and ½-H fields. Fig. 3-2 will make this clear. Time t_1 to t_3 represents H (field 1), and t_2 to t_3 represents ½ H, or field 2. If the first V sync pulse occurred at t_3, the integrating capacitor would start charging with a 2/1 voltage difference between fields, resulting in a much earlier firing of the oscillator for field 2 than for field 1. However, if a

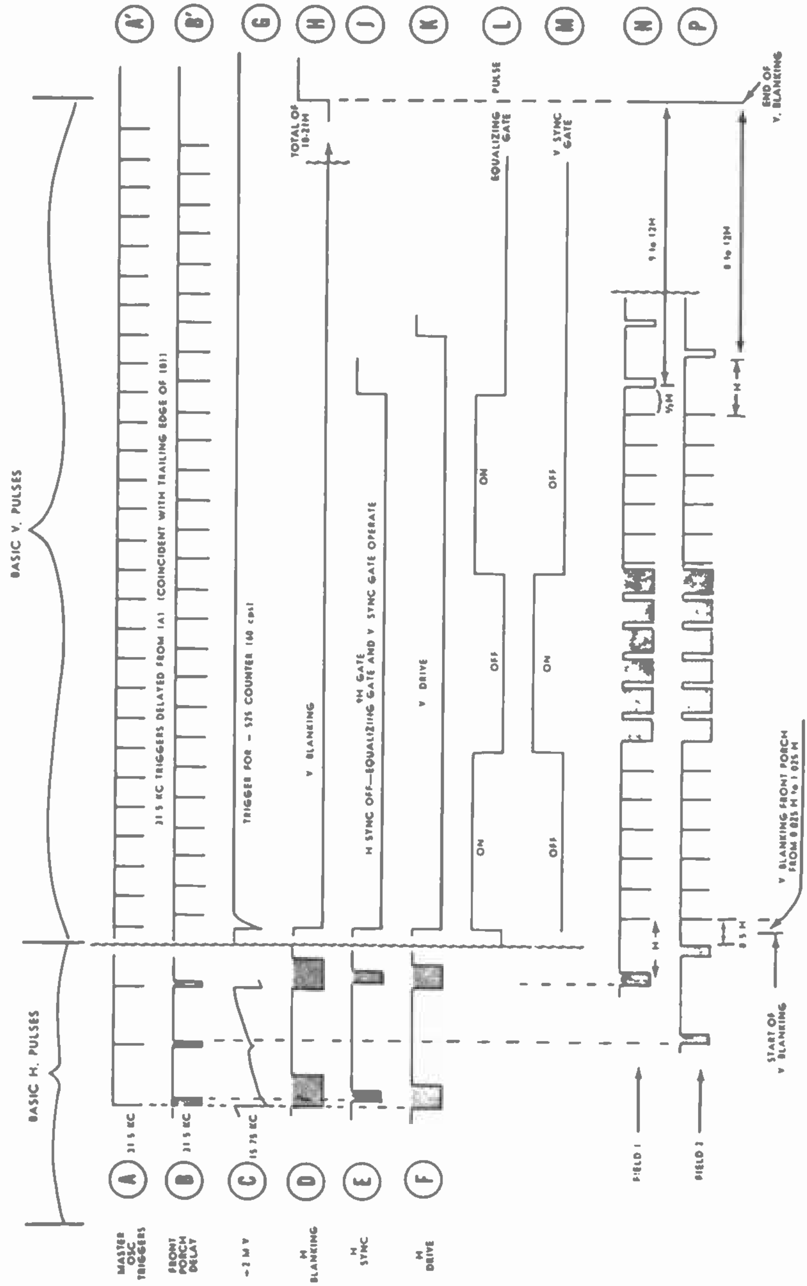


Fig. 3-1. Basic pulse timing of the sync generator. Because of space limitations the horizontal time scale of the vertical interval is shortened compared to the horizontal pulse scale.

3-H waiting, or equalizing, interval (t_4) is allowed, the following condition prevails:

Field 1: $H + 3H = 4H$
Field 2: $0.5H + 3H = 3.5H$
Ratio: $4H/3.5H = 1.14/1$

Thus, by the time the first V sync pulse occurs at t_4 the charge on the integrating capacitor is sufficiently equalized between the even and odd fields to fall within the requirements of interlace, assuming proper adjustment of the receiver vertical hold control.

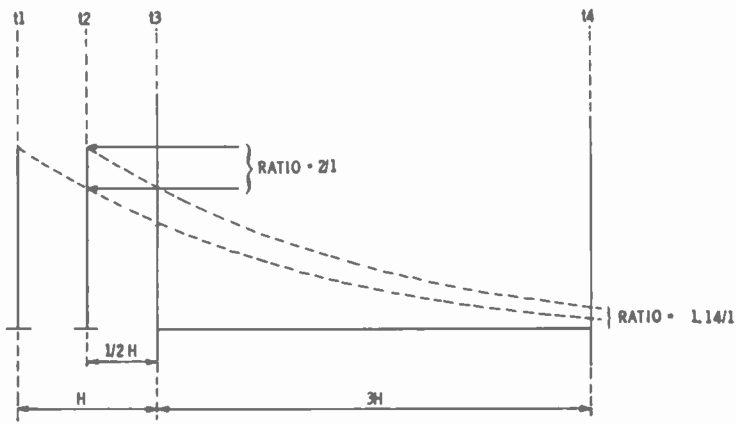


Fig. 3-2. Purpose of the 3H "equalizing interval" before vertical sync.

Obviously, pulses must be present during this interval to maintain H sync. These pulses must be twice the H frequency so that equalizers 1, 3, and 5 supply triggers for field 1, and equalizers 2, 4 and 6 supply them for field 2. The trailing edge of alternate V sync pulses (serrations) then step the H sync oscillator during that interval. Equalizing pulses are made one half of the width of H sync pulses so that no shift in the AC axis will occur at the transition between the line frequency and the double-frequency equalizers. Preventing this shift in axis is important to attenuate the inherent 30-cycle component of horizontal sync introduced as a result of alternate fields being displaced by one half of a line (30-cps).

3-2. TYPICAL SYNC-GENERATOR CIRCUITRY

In all modern sync generators certain circuit configurations are typical, regardless of manufacture. Due to the exceptionally wide use of the RCA Type TG-2 generator, this unit will be reviewed with more detail than provided in the Instruction Book as an aid to the maintenance engineer. More recent solid-state circuitry will then be reviewed.

NOTE: It is assumed that the reader has a basic understanding of electronic circuitry as applied to multivibrators, counters, clippers, etc. However, fundamentals of such action are included with special emphasis on points important to maintenance techniques.

The RCA studio (rack mount) sync generator (TG-2A), and the field (portable) generator (TG-12A) employ identical chassis with the same electrical characteristics. Fig. 3-3 illustrates the TG-2A and associated components as installed at Television City (WTAE) in Pittsburgh, Pa.

All multivibrators employed in this generator are based on the basic monostable (one-shot) circuit shown in Fig. 3-4. Notice that in the absence of triggering the A-section cathode is held near ground potential by CR1 conduction while the B-section cathode receives the full −70 volts. The resultant heavy conduction of the B section causes maximum plate-current flow through R4, placing a negative bias on the A-section grid, thus holding that tube below cutoff. The stable mode of this circuit is with the A section cut off and the B section conducting.

A negative trigger is applied through C2 and C1 to the grid of the B section driving this section to cutoff. The voltage across

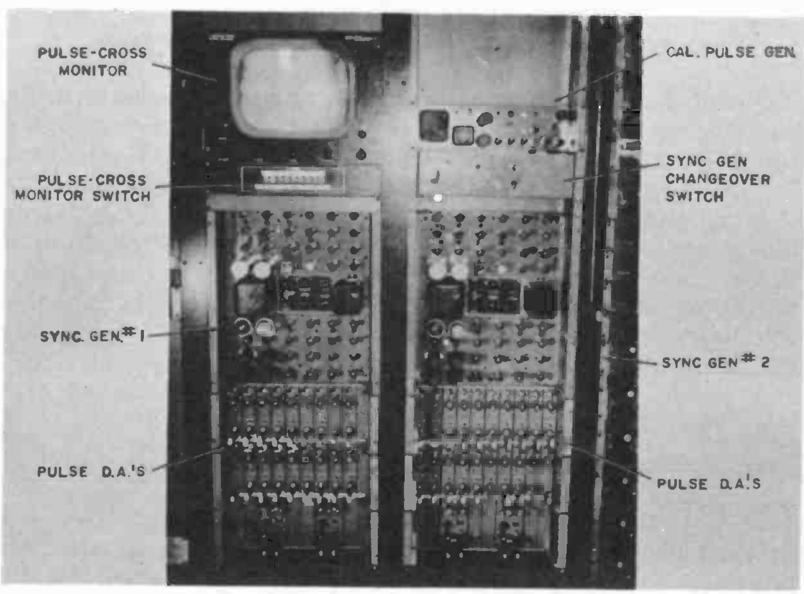


Fig. 3-3. Pulse racks at station WTAE.

R4 rises from a negative value to exactly ground potential. Zero grid bias allows the A section to conduct sufficiently so that the rise in cathode voltage cuts off diode CR1, removing that cathode from ground. Since this current also flows through plate resistor R3, a negative-going pulse which drives the B section well beyond cutoff is developed. The resultant charge on C1 now starts to leak off through R1. The duration of the B-section cutoff is determined by the C1-R1 time constant. As soon as the grid voltage intercepts the cut-off value, the B section again conducts, returning the A section to the initial cut-off mode. The multivibrator will remain in this stable mode until the arrival of the next trigger pulse.

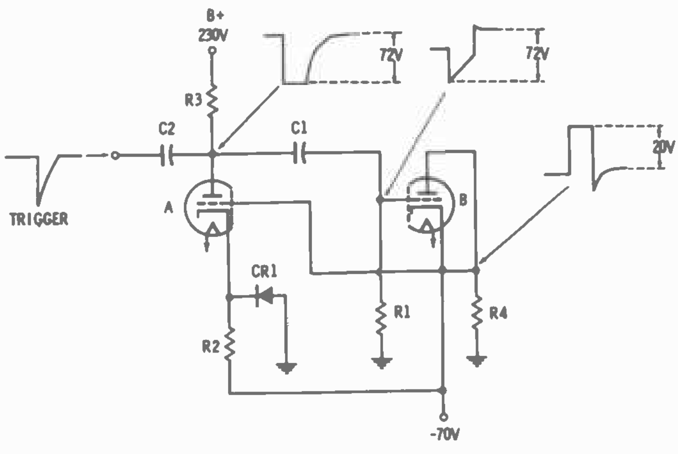


Fig. 3-4. Basic multivibrator circuit.

Keep in mind that the timing-cycle duration of this basic multivibrator (hereafter designated MV) depends on the R1-C1 time constant plus the value of the A-section cathode resistor which predetermines the conduction level. It should also be realized that this MV may be used as a counter by designing its natural time constant so that it runs for slightly less than the number of input triggers to be counted. Thus, when the desired count is reached, the next input trigger starts a new timing cycle.

The complete block diagram of the RCA TG-2A sync generator is presented in Fig. 3-5. The 31,500-cycle master frequency is generated by MV V13, which is a modification of the basic circuit to allow a free-running mode of operation. Essentially this modification consists of an added LC ringing circuit, an added trimmer capacitance across C1 (variable), and a variable effective resistance for R1. The trimmer capacitance is used to set the MV frequency at precisely 31.5 kc with the effective grid resist-

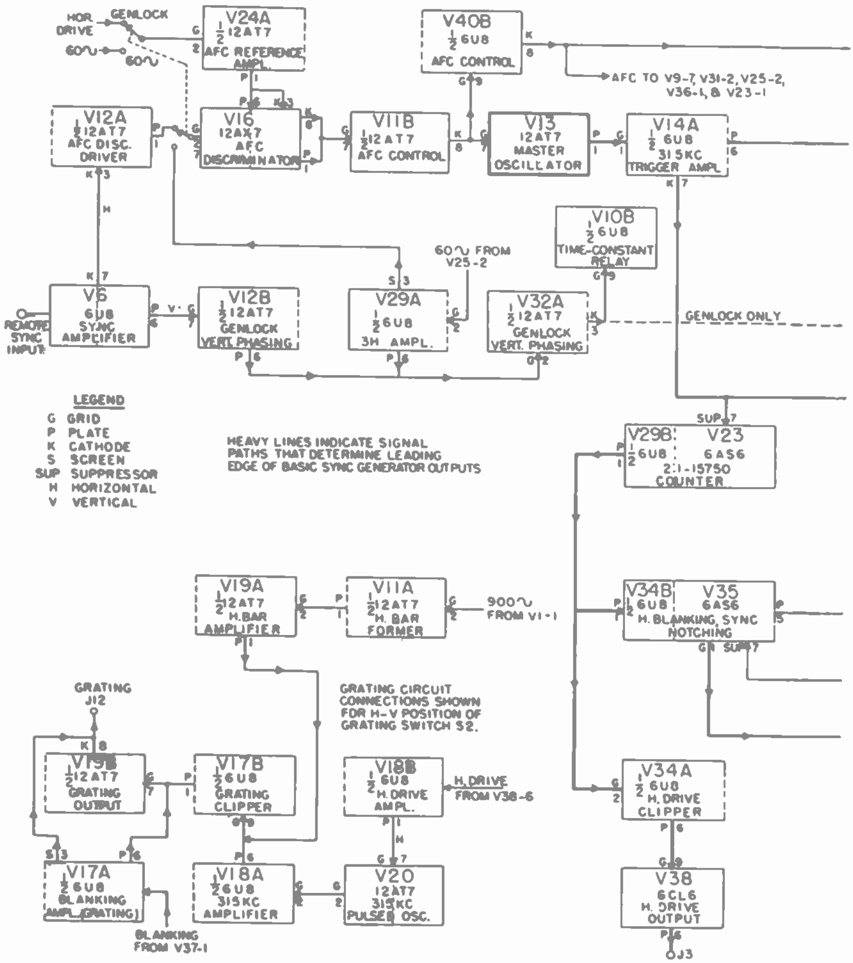
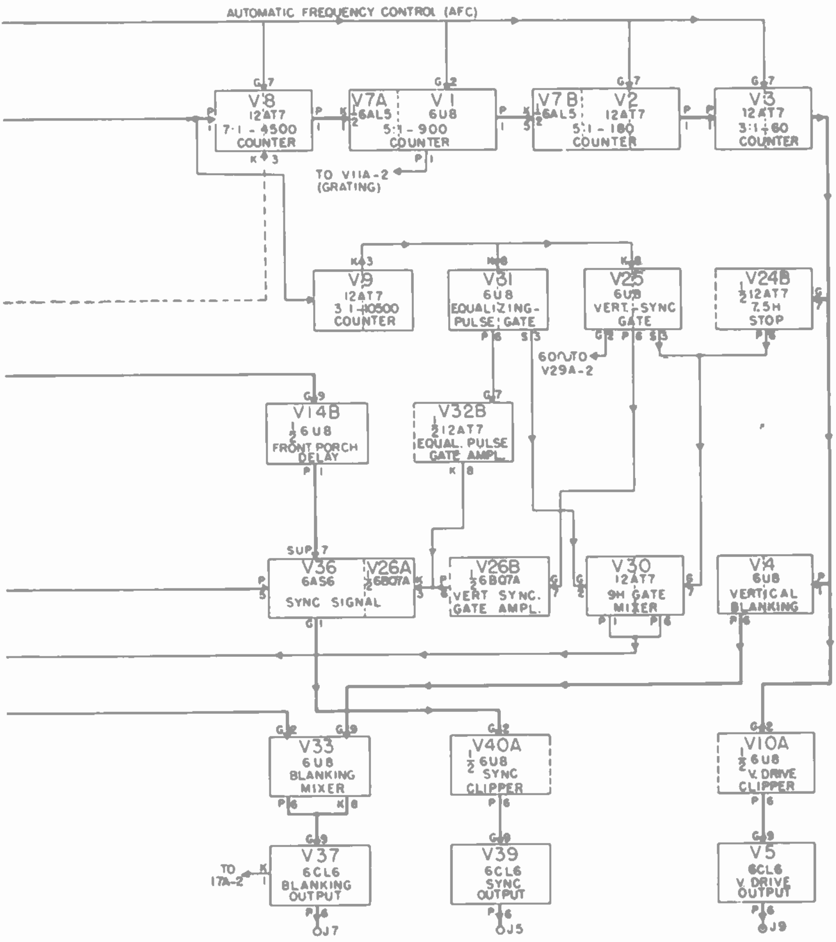


Fig. 3-5. Block diagram of



the RCA TG-2A sync generator.

ance of R1 set in the electrical center of its value. Variation of R1 is then capable of shifting this frequency approximately 5% each side, and it is used for the stability tests described later.

The master-oscillator frequency may be controlled by setting the front panel frequency-control switch to any one of the following positions:

1. Genlock
2. 60-cycle (line lock)
3. Off (free running)
4. Crystal
5. External

For monochrome telecasting the generator is normally run with 60-cycle AFC. This function may be understood by following the block diagram of Fig. 3-5 and the simplified schematic of Fig. 3-6. The AFC discriminator (V16) compares the countdown existing at the vertical sync gate (V25) with the 60-cycle wave derived from the power line and develops a DC voltage which controls the master oscillator by means of AFC control tube V11B. Tube V16 conducts only during a relatively short interval (3 lines or 190 microseconds compared to a 60-cycle alternation interval of 8333 microseconds) of the V sync gate, charging capacitor C1 to the value of the line voltage pulse existing at that instant. As you can see from Fig. 3-6, the AFC voltage that is developed will depend on the position of the V pulse relative to the 60-cycle power line and will tend to hold this position to occur at the time when the power line excursion is crossing its zero axis. In this way the master oscillator is stabilized at a frequency exactly 525 times the local power-line frequency. This generator also employs a variable phasing control to shift the phase of the 60-cycle line reference voltage so that proper synchronization of motion picture projectors with short-application shutters (as used with iconoscope camera chains) may be obtained.

When the selector switch is placed in the XTAL position, V11B (Fig. 3-5) is connected as a crystal oscillator at 94.5 kc and the master oscillator acts as a 3:1 counter.

When it is necessary to superimpose local material on network or remote shows, the Genlock position is used. In this position the incoming net or remote signal is brought into a stabilizing amplifier where the supersync region is stripped from the composite signal and used as triggers for the local sync-generator control. The video-only portion of the signal is then fed to the local switcher input where it can be handled as any local camera signal (noncomposite input).

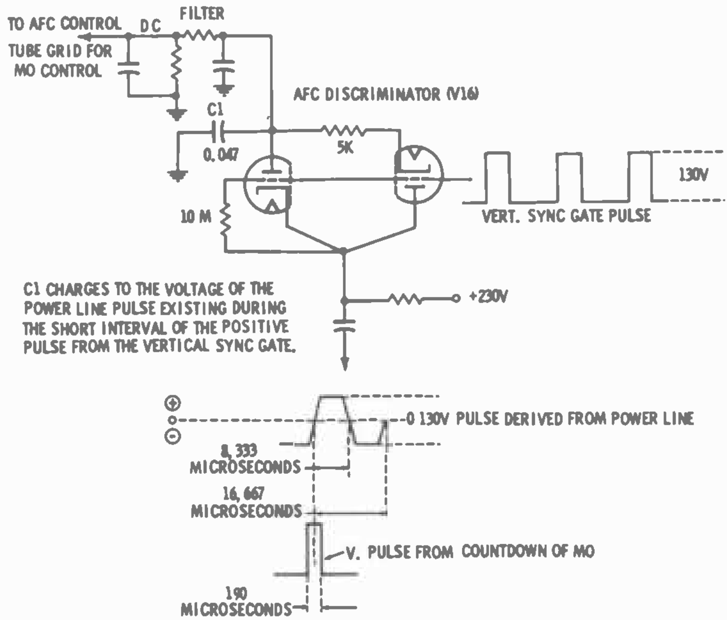


Fig. 3-6. AFC action to maintain constant master-oscillator frequency.

Two conditions must be met to lock in the local sync generator with that at a remote point.

(1) The frequency of the local master oscillator must be made exactly equal to that at the remote generator.
(2) The vertical pulses of the local sync generator must be brought precisely into phase with those at the remote point.

The Genlock operation of the TG-2 and TG-12 generator functions as follows: the stripped-off remote sync is amplified by V6A (Fig. 3-5). The horizontal component is separated by a differentiation network and is then amplified and clipped by V6B and V12A. Positive pulses from V12A feed AFC discriminator V16, in place of the 3-H vertical sync gate pulses as described for the line-lock function. When in Genlock operation, the charging capacitor (C1 in Fig. 3-6) is removed from ground return. At the same time the grid of V24A receives a horizontal drive pulse from the local sync generator and produces a sawtooth voltage which is applied to V16 in place of the 60-cycle, power-line wave. This time the DC control voltage fed to AFC control tube V11B serves to lock the frequency of the local sync generator to that of the remote generator.

The Genlock AFC must hold in the presence of noise, thus requiring a long time constant for the AFC control voltage. At the same time, for quick and reliable initial lockin, the DC control circuit should have a relatively short time constant. This problem is mastered in the RCA generator by incorporating a relay with the coil placed in the cathode circuit of Genlock relay tube V10B. The Genlock vertical phasing pulse from the cathode of V32A biases V10B so that the current is very small in the presence of this pulse. The relay is therefore open, and the total capacitance for the AFC time-constant is 220 mmf. When vertical phasing has been completed (as described below), the phasing pulse disappears, causing V10B to conduct heavily. This current operates the relay whose contacts shunt an additional 0.01 mfd across the 220 mmf capacitance.

Vertical phasing is accomplished by causing the 7:1 counter (V8) and all following counters to miscount until the local 60-cycle field rate is lined up exactly with the remote signal. Remote vertical sync is separated by an integration circuit, then amplified by V12B. The negative pulse from the plate of V12B is mixed with the positive-pulse local vertical sync gate appearing at the plate of V29A, and it is applied to the grid of Genlock vertical phasing tube V32A. This tube is biased so that it will conduct during the positive local vertical pulse until the negative remote vertical pulse is brought into position to cancel it. As long as V32A is conducting (vertical pulses not in coincidence), the 7:1 counter (4,500 cycles) will miscount, causing the fields to slip until exact coincidence is achieved.

Positive 31,500-cycle pulses at master oscillator V13 plate appear as positive pulses of stabilized amplitude (waveform A in Fig. 3-7) at the cathode of buffer amplifier V14A. These pulses drive the front-porch delay generator (V14B). Remember that blanking is inserted in the camera control units, whereas sync pulses are inserted at some point after final switching. In order to fix the horizontal front porch at approximately 1.6 microsecond, the width of pulse B of Fig. 3-7 is made variable by the Front Porch control so that this front-porch value is obtained when measured at the program line output. This delay MV is triggered from the trailing edge of the master-oscillator pulse by obtaining negative triggers from differentiated waveform A′. Note, therefore, that the EIA synchronizing signal (E in Fig. 3-7) is delayed with respect to the trailing edge of the master-oscillator pulse. The trailing edge of the undelayed pulse is used to trigger H blanking (D in Fig. 3-7) through the 2:1 counter V23 and V29B, thus causing blanking to start ahead of sync by the front-porch interval. The horizontal blanking MV (V34B and V35) incorporates a variable capacitor to adjust the time constant for

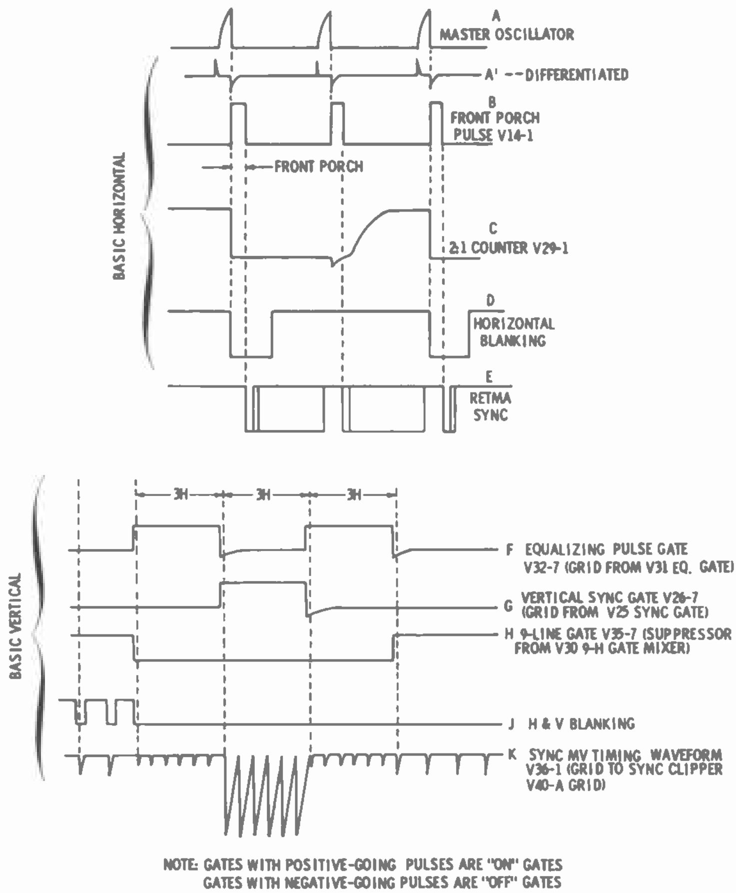


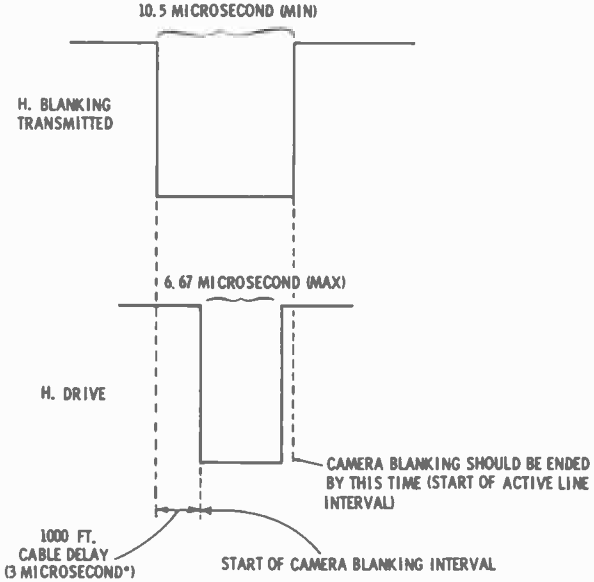
Fig. 3-7. Derivation of pulses with the RCA TG-2A sync generator.

the desired width of H blanking. The signal from V35 is fed to V33 where it is mixed with V blanking from V4 and is clipped by a crystal diode to produce composite blanking to drive output stage V37. All outputs allow amplitude adjustment (usually 4 volts p-p) without affecting the 75-ohm sending-end termination impedance.

The plate waveform of H blanking MV (V34B) is differentiated and then clipped in V34A to produce H driving pulses for output stage V38. The width of this pulse is fixed and stabilized at a nominal value of 0.1 H, or 6.3 microseconds.

In a similar manner, the V blanking waveform at the plate of V4B is differentiated and then clipped by V10A to produce V drive pulses for output stage V5. Vertical drive width is fixed at approximately 0.04V, or about 670 microseconds.

It is important to understand why the driving signals to the camera chains are normally about one half of the width of their respective blanking widths. This is particularly important at the horizontal frequency, as is evident in Fig. 3-8. Remember that the transmitted composite blanking is inserted at the camera control unit, as well as the driving signals for the camera. However, the camera cable may be as long as 1,000 feet in some instances,



*SINCE CABLE DELAY IS APPROX 1.5 MICROSECONDS PER 1000 FT, TOTAL DELAY (TO AND FROM CAMERA) IS 3 MICROSECONDS.

Fig. 3-8. Why horizontal drive is approximately one-half the width of horizontal blanking.

and allowance must be made for the cable delay, which is roughly 1.5 microsecond per 1,000 feet of cable. Since the total path is to and from the head (2,000 ft), total allowance must be made for a 3-microsecond delay. It may be observed from Fig. 3-8 that if the H drive pulses were the same width as the H blanking pulses, camera blanking would not be ended at the start of the active line interval in cases where long camera cables are employed (unless drive is regenerated and narrowed). This effect is not so important at the vertical frequency since 3 microseconds is negligible as compared with a total of about 1,250 microseconds.

Notice from the previous description that a "pulse narrowing" effect is used. To meet the requirements of waveforms, square- and rectangular-shaped pulses must not only be of the exact required width, but they must also have steep leading and trailing edges and flat tops. The fundamentals of controlling these characteristics are illustrated in Fig. 3-9. When the time relationship of the square wave T is approximately equal to the time constant of RC, the output of the waveform voltage across the resistance is shown at the immediate right. When T is much less than RC, as one quarter of the time constant or less, the output voltage is more nearly equal to the applied waveform. When T is much greater than RC, as four time constants or more, the waveform becomes that shown at the bottom drawing. In circuits used, for example, to narrow an applied pulse, a relatively small capacity and low resistance (making T greater than RC) serves

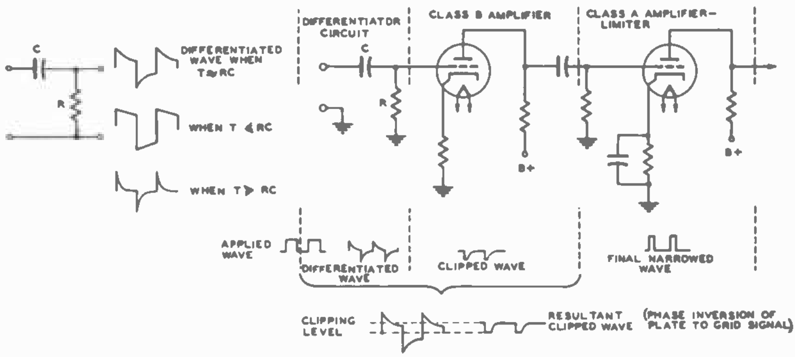


Fig 3-9. Method of obtaining the desired pulse waveshape.

to attenuate the low frequencies and easily pass the highs. Thus, the condition of T being much greater than RC results in the waveform shown at the lower left of Fig. 3-9. The pulse is initially narrowed here, but it is not of the required waveform. The right-hand portion of this figure illustrates how the differentiated wave is clipped and squared to produce the final waveform. The class-B amplifier clips and squares the wave, the resulting narrowed wave being reinverted in a class-A amplifier. The more the value of C in the differentiator is reduced, the greater the narrowing of the wave as applied to the class-B stage. This stage, aside from being biased to cutoff, operates at low plate voltage (or is over-driven), hence squaring off the applied peaks due to the overdriven condition of the plate current.

The derivation of composite sync is somewhat more complex than for the other pulses. It will be helpful here to note the type of timing waveform (K in Fig. 3-7) at the output of the sync MV (V36-V26A). When this waveform is amplified and clipped, the standard composite sync is obtained. The sync MV (V36-V26A) fundamentals of operation are as follows:

(1) Conduction of the equalizing pulse gate amplifier V32B reduces the amplitude of the MV timing waveform during the equalizing pulse interval.
(2) Conduction of vertical sync gate amplifier V26B greatly increases the amplitude of the MV timing waveform during the vertical sync interval.
(3) The suppressor of V36 receives negative triggers (31,500 cps) from the output of the front-porch delay circuit.

Complete derivation of composite sync is as follows: the vertical sync gate (V25) and the equalizing gate (V31) are multivibrators which receive triggers from the 60-cycle counter (V3) and then trigger each other to produce the gates shown by F and G in Fig. 3-7. With the equalizing pulse gate on, the vertical sync gate is held off, and the grid of V32B receives a positive trigger which causes it to conduct, reducing the amplitude of the sync MV output.

At the end of three lines (six leading equalizing pulse intervals) the equalizing gate MV under the influence of its own natural time constant and by triggering from the 3:1 counter (V9) reverses its mode of operation. This triggers the vertical sync gate on, causing vertical sync gate amplifier V26B to conduct, greatly increasing the output amplitude of the timing waveform. At the start of this interval the grid of V32B received a negative trigger to drive this stage to cutoff.

Again at the end of 3H, the vertical sync gate MV, by the discharge of its own natural time-constant and by triggering from V9, reverses its mode of operation, returning the sync MV to the initial vertical interval state. This is the trailing equalizing pulse interval which completes the vertical sync interval constituting 9H (9 lines).

At the beginning of this 9H interval the 7.5H Stop Tube (V24B) received the same trigger from the 60-cycle countdown that initiated operation of the vertical interval multivibrators. This negative pulse causes conduction in V24B to cease, thereby raising the screen voltage on the V sync gate and allowing it to be on when triggered. When the 7.5 H stop tube is conducting, the screen of the V sync gate is returned to a high negative potential which prevents conduction. The grid-circuit time constant of the 7.5H stop tube is such that the capacitor will discharge to allow V24B to conduct somewhere between 6 and 9 lines, hence the term "7.5H Stop" tube. This will prevent the vertical timing function from recurring until the next field interval trigger from the 60-cycle countdown is received.

Note also from the block diagram in Fig. 3-5 that the two vertical interval gates are mixed by the 9H gate mixer stage

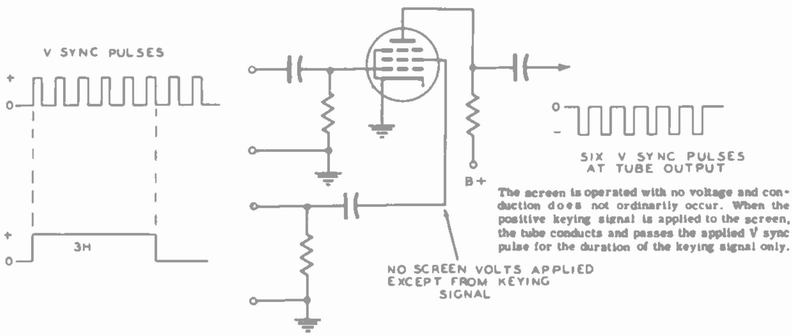
(V30) to produce the waveform shown by H in Fig. 3-7. The 9H gate (negative) is connected to the suppressor of V35 so that this tube is held at cutoff during this interval. The reason for this is that V35 is the so-called "notching" tube, and the control grid receives horizontal blanking pulses. During the regular horizontal intervals (all times except during the 9H interval described above) the negative-going H blanking pulse drives the tube to cutoff, whereas in the intervals between blanking pulses (positive-going pulse) the tube conducts and shunts the sync MV so that it can be triggered only during H blanking pulses. This is to say that pulses are recurring only at the line frequency of 15,750 cps. This notching action must not occur during the 9H interval where the pulses must recur at 31,500 cps, hence V35 is held at cutoff during this time to switch the pulse rate from 15,750 to 31,500 cps.

It will be noted from the block diagram that this generator also incorporates a "grating signal" generator with outputs of 900 cps and 315 kc. The grating pattern is used to check monitor linearity.

As noted, a gate circuit is common to all sync generators. The gate is said to be "closed" when pulses continuously applied to the grid are not passed to the output plate circuit. The gate is "open" when the grid pulses are passed to the output. A gate will take the form of either a "keying-in" or a "keying-out" circuit, as shown by Fig. 3-10.

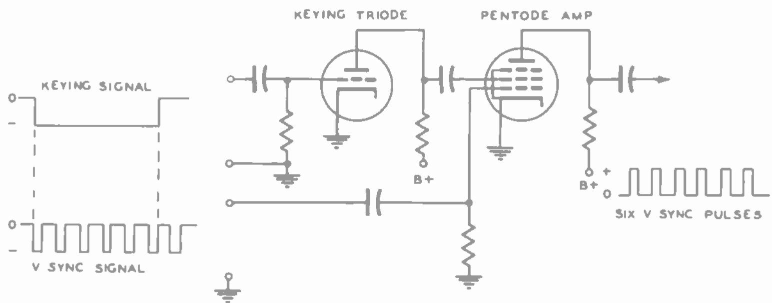
Fig. 3-10A shows the basic action of a keying-in circuit. The screen of the pentode tube is operated at 0 potential, and circuit constants are such that the tube is normally non-conducting. Thus, although the V sync pulses may be continuously applied to the grid, no corresponding signal voltage appears at the plate output. However, a keying signal of 3H duration and of positive polarity applied to the screen will cause the tube to conduct during this time interval. Thus, six V sync pulses of inverted polarity appear at the plate output. Fig. 3-10B illustrates a commercial type of keying-in circuit. Here the keying signal must be of negative polarity so that the amplified pulse applied to the screen of the pentode tube is a positive pulse, causing conduction during this time interval.

In some instances it is necessary to utilize a keying-out circuit. One such application is in eliminating the 15,750-pps H sync pulses during the V sync interval every 1/60 of a second. A commercial type of keying-out circuit is shown in Fig. 3-10C. In this circuit the screen of the pentode is operated with a low screen voltage of approximately 25 volts. The continuously applied H sync pulses at the grid circuit are passed during the time the screen voltage is +25 volts. When the positive 60-pps keying



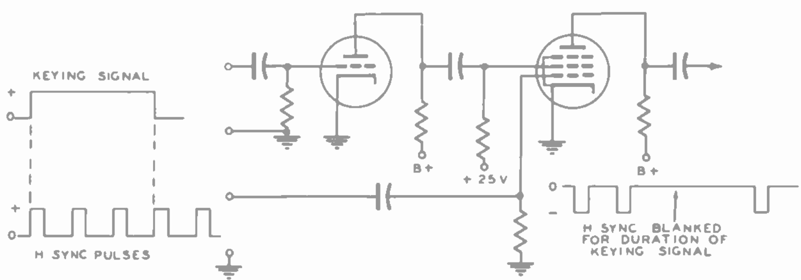
The screen is operated with no voltage and conduction does not ordinarily occur. When the positive keying signal is applied to the screen, the tube conducts and passes the applied V sync pulse for the duration of the keying signal only.

(A) Basic function of keying-in circuit.



As in (A) above, the pentode screen has no B+ voltage. A negative keying pulse of the desired width applied to the triode results in a positive pulse of the same duration applied to the screen of the pentode. The pentode then conducts, and the V sync pulses continously applied to the grid of the pentode are passed for the duration of the keying pulse.

(B) Commercial-type keying-in circuit.



The pentode screen-grid is operated at approximately 25 volts at which value the tube conducts. A positive keying signal applied to the triode grid results in a negative keying pulse of over 25 volts applied to the pentode screen. This drives the pantode to a non-conducting period over the duration of the keying signal.

(C) Commercial-type keying-out circuit.

Fig. 3-10. Keying circuits.

signal is applied at the grid of the triode, a negative pulse of over 25 volts is applied to the screen of the pentode, cutting off conduction during this interval. Thus, the H pulses are keyed-out 60 times per second for the required time as determined by the duration of the keying signal.

The pulse duration of a multivibrator output is determined by the time constant of the "off" tube relative to the "on" tube. Fig.

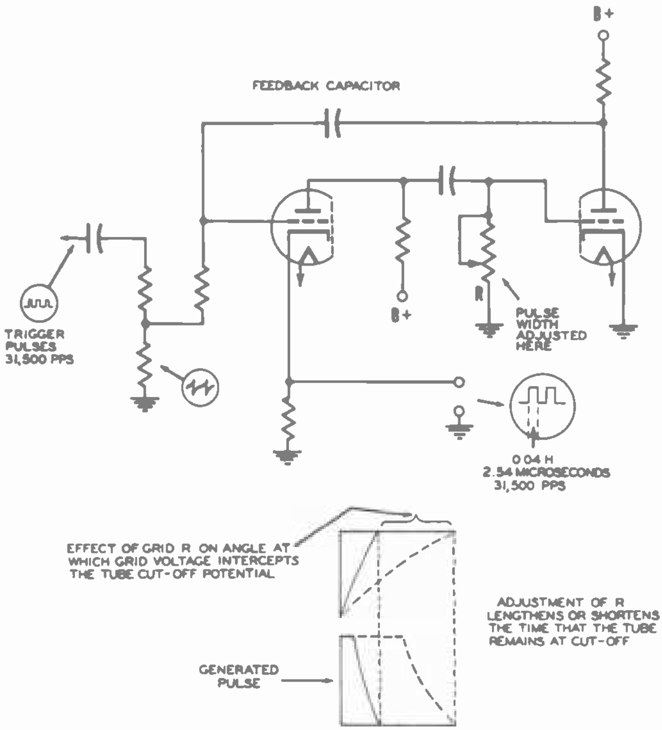


Fig. 3-11. Basic MV circuit employing width control. The timing shown here is for equalizing pulses.

3-11 shows the basic action of an equalizing pulse multivibrator. This circuit is triggered from squared pulses derived from the 31,500-cps, sine-wave oscillator and clipped to form the square wave for triggering. The grid resistor adjusts the time over which the oscillator is active for each trigger pulse and hence determines the width of the output pulses. The action of this circuit is basic to any multivibrator incorporating pulse-width adjustment.

Clipping circuits vary in methods and applications. Such circuits may be used to square-off extremities of a sinusoidal wave, or they may have pulsed waveforms applied in which case the clipper removes either positive or negative tips of the waveform. Any of the usual methods (series or parallel diode clipping, class-C amplifier, plate saturation, or grid clipping) are found in commercial equipment. For convenience, fundamental clipper circuits are reviewed in Fig. 3-12.

The timing section of some vacuum-tube types of sync generators and almost all transistorized sync generators employ a series of binary scalers to derive the 60-cps countdown from the 31.5-kc master-oscillator frequency. The output of each binary is a wave having one positive and one negative-going pulse per cycle of operation. However, only the negative-going or positive-going pulse (depending on design) will trigger the next binary, resulting in a division of 2. Thus, a series of binaries in cascade result in total divisons that are integral powers of 2. Stated another way:

$$\text{Total division} = 2^n$$

where,

n is the number of binary scalers.

To obtain an odd number of divisions as required in the sync generator, feedback loops are provided, as shown by one type of binary counter in Fig. 3-13. In this case:

$$\text{Total division} = 2^n - 2^p$$

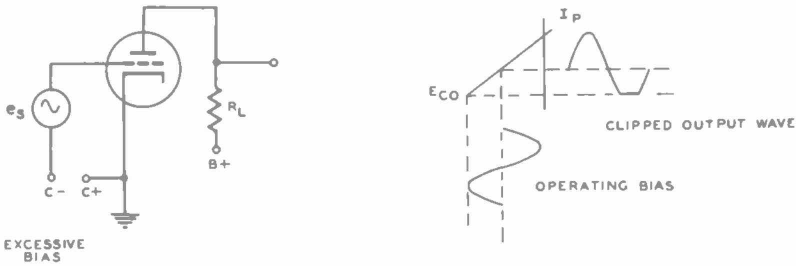
where,

p is the order number of scaler to which feedback from the output scaler is supplied.

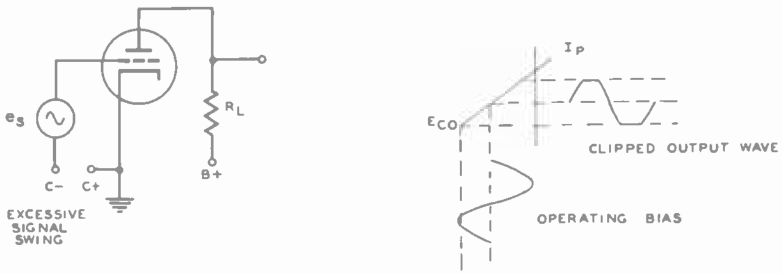
Normally, if this is at the input of the first scaler, $p = 0$; if at the output of the first scaler, $p = 1$; if at the output of the second scaler, $p = 2$; etc. In the case of a magnetic-core binary scaler, the power by which the numeral 2 is raised may be either plus or minus depending upon which polarity of current is induced through the windings by the feedback pulse.

Observe Fig. 3-13. The ten identical binary stages have an inherent count of 2^{10}, or 1024. To obtain a count of 525, the third and the fifth to ninth stages are reset by the output of the 1024th pulse. The reset pulse also enters the number 1 binary in such a way that it is set to the −1 state instead of the zero state. Note from the table in Fig. 3-13 that the sum of the feedback is 499. Thus a count of 499 is preset into the counter before the first pulse arrives at the input. The total 1024 count minus the preset 499 results in the desired 525.

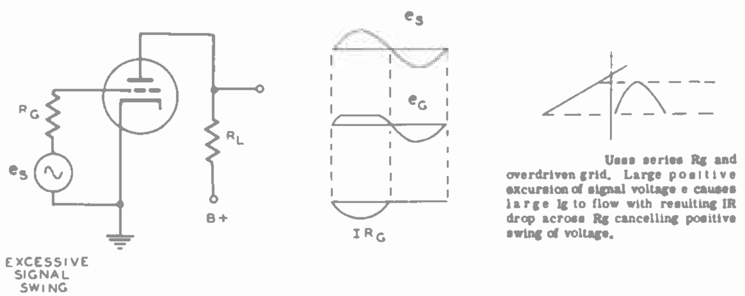
Some sync generators employ completely tube-free (solid-state) circuitry. A binary scaler similar (but not exactly the same) to that described is a common factor in all such units.



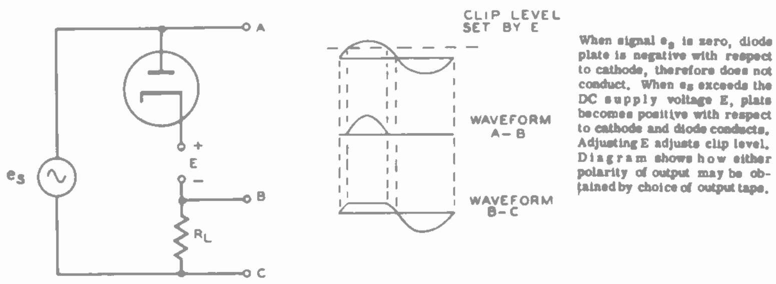
(A) Plate-current cutoff clipper.



(B) Plate-current cutoff and saturation clipper.



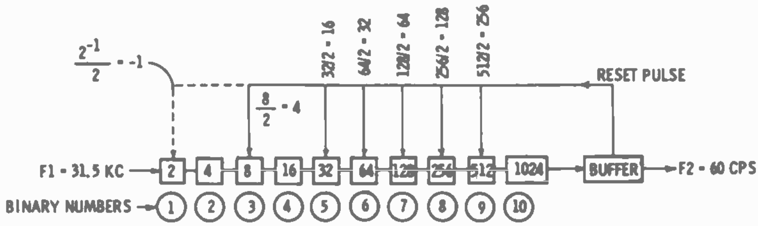
(C) Grid clipper.



(D) Diode clipper.

Fig. 3-12. Fundamental clipper circuits.

Fig. 3-14 shows a basic bistable multivibrator (MV) circuit using transistors. NOTE: a bistable circuit has two stable states. When triggered by an input pulse, the MV switches to the second stable state where it remains until triggered again. Some sync generators set pulse widths by means of a delay line which feeds a pulse for initial triggering and a delayed pulse to turn the MV off.



BINARY NO.	SUM OF RESET	
1	-1	3
3	4	
5	16	
6	32	
7	64	
8	128	
9	256	
TOTAL RESET = 499		

TOTAL DIVISION OF BINARIES = 2^n
WHERE n = NUMBER OF BINARY SCALERS
TOTAL = $2^n = 2^{10} = 1024$

TOTAL DIVISION (WITH RESET) = 1024 - 499 = 525

Fig. 3-13. One type of binary scaler.

The circuit in Fig. 3-14 has two separate trigger inputs. A trigger pulse at input A will change the MV state. A trigger of the same polarity at input B, or of opposite polarity at input A will again trigger the state of the multivibrator. The output is taken between the collector and ground of either transistor (or both), the output pulse being opposite in polarity between transistors.

When the transistor is cut off, collector current is essentially zero and the output resistance is high. Fig. 3-15A shows a typical voltage divider for use with the circuit of Fig. 3-14. The load resistor (R_L) voltage drop is only 2 volts, leaving almost the full collector supply voltage (V_{CC}) at the collector element (point C). The voltage drops across the coupling resistor (R_c) and base resistor (R_B) are 21 and 7 volts, respectively, providing there is a negative voltage (−5 volts) at point B. This forward-biases transistor X2 (Fig. 3-14), thus causing it to conduct. The "on" transistor is driven to saturation by the design of the resistor values and bias voltages.

Referring now to the voltage divider in Fig. 3-15B, the saturation current of the "on" transistor flowing through R_L results in the full V_{CC} drop across this resistor, hence essentially zero voltage at point C. The base supply voltage (V_{BB}) is divided by resistors R_c and R_B so that +1.5 volts results at point B. The "on" transistor is held at cutoff by the reverse bias on the base; (polarities are reversed for NPN transistors). Capacitors C1 and C2 in shunt with the coupling resistors (Fig. 3-14) sharpen the pulse wavefronts for more rapid switching of the transistors, hence faster rise and decay times of the pulse outputs. Some circuits employ diodes in place of the RC combination.

Basic forms of transistorized "logic" or gating circuits are shown in Fig. 3-16. In Fig. 3-16A the positive pulses at both inputs are required to change the state of operation from conduction to cutoff. With positive pulses applied only to the A

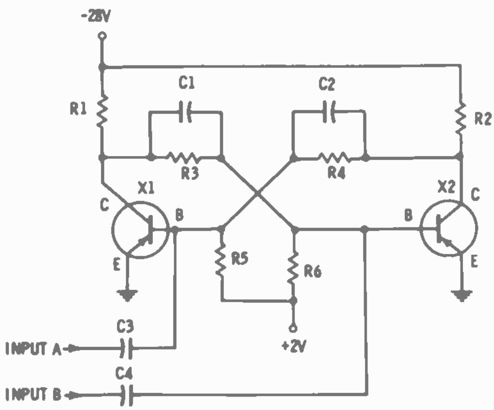


Fig. 3-14. A basic bistable multivibrator.

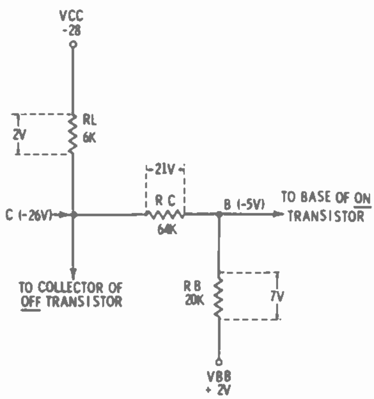
input, transistor X2 continues to conduct from the forward base bias. The resulting current through R_L continues to supply a negative potential at the output. Application of a positive pulse simultaneously to both inputs overcomes the forward bias and drives both transistors to cutoff. Thus the voltage drop becomes essentially zero through R_L and the output rises to ground (zero) potential. The example illustrates vertical sync pulses continuously applied to input A, gated "on" by a six-line gate pulse at input B.

Fig. 3-16B illustrates a reverse bias shunt gate, or "inhibition" gate. The transistor acts as a switch in shunt with the signal path. With no control pulse applied to the base, the transistor is nonconducting (open switch) and the signal pulse applied to the circuit is passed to the output. A negative-going control pulse applied to the base overcomes the reverse bias and the transistor conducts (closed switch), shorting the output terminals to zero. Thus this gate circuit is similar to the keying-out circuit previously described for vacuum tubes.

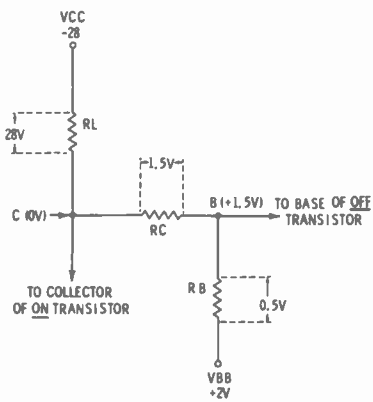
NOTE: This circuit can be designed as a keying-in circuit by employing forward bias on the base, while using a positive-going control pulse.

3-3. SYNC-GENERATOR ADJUSTMENTS

Proper adjustment of the station synchronizing generator depends entirely on the technician's familiarity with his oscillo-



(A) With negative voltage applied to base of "on" transistor.



(B) With positive voltage applied to base of "off" transistor.

Fig. 3-15. Voltage division of multivibrator circuit for "off" and "on" transistors.

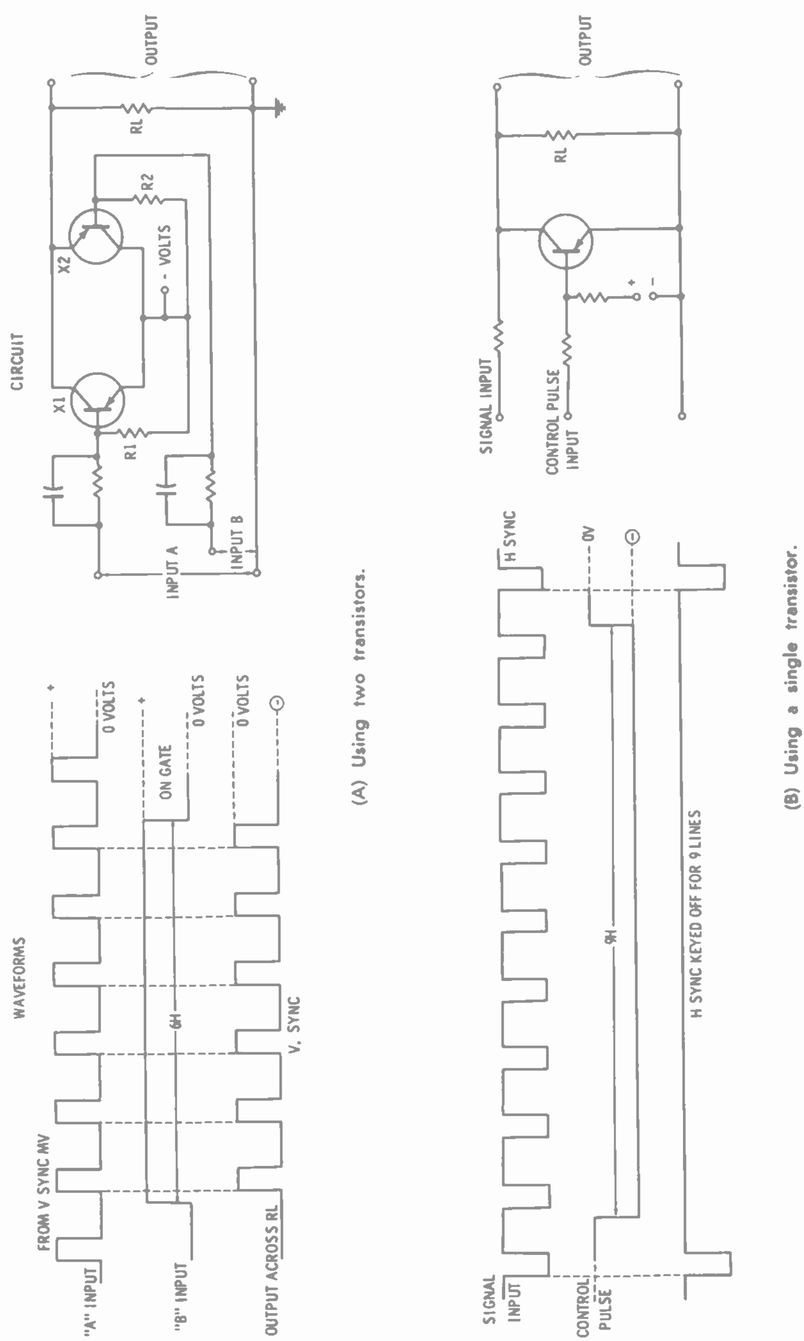
scope, the trouble taken to accurately maintain marker and time base standards, and the technique of adjustments. The reader should be thoroughly familiar with Section 1 of this book.

Frequency Adjustment

Proper operation of the counters and pulse-gating circuits will usually result if the master oscillator is within 5% of the 31.5 kc nominal frequency. It is entirely practical to adjust this frequency within 1% or less, greatly extending the stability factor over a period of time with changing tube and component characteristics.

Most sync generators provide some means of controlling the master oscillator frequency in any one of the following conditions:

(A) Free running for test or emergency use.
(B) Crystal control. Usually used for test setups. May also be used when power-line frequency is unreliable, as in some



(A) Using two transistors.

(B) Using a single transistor.

Fig. 3-16. Basic transistorized gating circuits.

remote locations, or when using standby power sources, such as emergency power generators.

(C) Line-lock to local 60-cycle power line. This is the most common operating condition for monochrome pictures.

(D) 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 net or remote signal.

(E) Any selected external signal, such as "color-lock" for color programs.

When a crystal circuit is employed, the crystal frequency is usually 94.5 kc, or three times the master oscillator (hereafter termed 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 scope probe (low-capacity type) at the 31.5-kc test point. With the switch in the XTAL position, the waveform should look like the upper trace

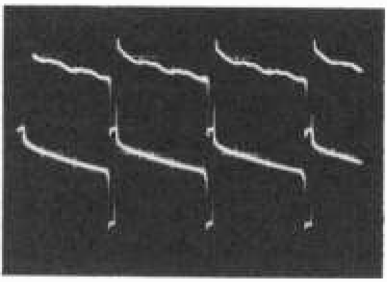


Fig. 3-17. Upper trace shows 31.5-kc waveform with MO on crystal control; lower trace is the same waveform with MO free running.

in Fig. 3-17. Since the crystal is 94.5 kc, a division of 3 is 31.5 kc and is indicated by three sine-wave cycles. If a different indication occurs, the MO frequency adjustment control should be rotated for proper frequency. In the RCA type TG-2A generator, a capacitor is used as a coarse adjustment and a time-constant potentiometer as a fine (about 3%) adjustment. Use the capacitor for setting the exact frequency with the time-constant control in midposition. This same procedure is followed on similar types of MO circuitry in other generators.

Step 2. The procedure in Step 1 assures proper frequency under crystal control. This step sets the free-running frequency to 31.5 kc within close limits. This step assumes use of a triggered scope, such as the Tektronix 524. Step 3 is an alternate method when triggered sweep is not available. Step 4 describes a vernier adjustment for precise setting providing counters are properly functioning. Expand the scope sweep to display one cycle at 31.5-kc in the crystal position and position the leading edge of the pulse on a verticle graticule line. Switch the MO control from XTAL to the Off (free-running) condition; if the pulse shifts from the reference position, adjust the MO frequency until no shift occurs between crystal control and free-running modes. The lower trace in Fig. 3-17 should be observed when the MO is free running.

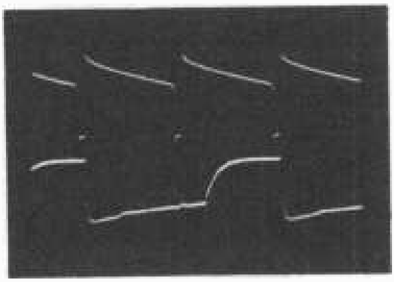


Fig. 3-18. Upper trace is 31.5 kc MO waveform; lower trace is 15.75 kc (÷2 stage) wave showing "division pip."

Step 3. Fig. 3-18 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. 3-18A is the waveform of the 31.5 kc MO and the lower trace is the output of the "divide-by-two" (15.75 kc) circuit showing the characteristic "division pip" of the alternate pulse. Expand the waveform in the lower trace as in Fig. 3-19 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 horizontally. If this width changes in the "free-running" position compared to the width when on crystal control, adjust the MO frequency so that no change occurs between crystal and free-run modes.

Step 4. With the MO in the free-running mode, observe any 60-cycle signal (such as at the counter output, vertical drive, or blanking) with the scope trigger selector on Line position. This shows any "slip" with 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%, 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.

Step 4 is normally the only step necessary for routine checks of the MO frequency. However, the previous steps should be exercised occasionally to check the crystal control and, of course, are necessary if trouble is experienced in counter stages.

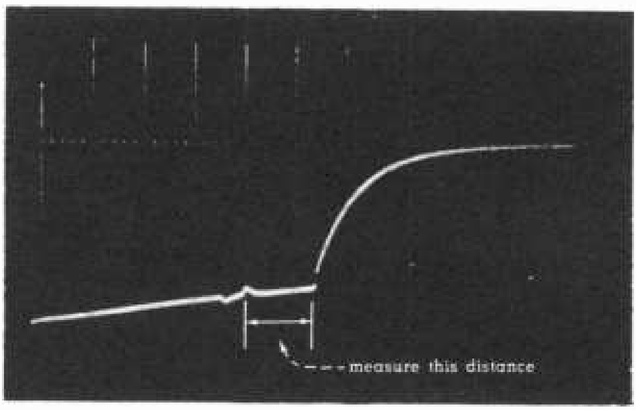


Fig. 3-19. Same as lower display in Fig. 3-18 except on expanded scale.

Some generators that do not include a crystal for test have a 94.5-kc ringing circuit in the MO to sharpen wavefronts and provide a reasonable degree of short-term stability. (Long-term stability is provided by the AFC circuit.) Step 4 is normally used to adjust the frequency for this type of MO, but any indication of faulty 60-cycle pulses always poses the question as to whether the trouble is in the counters or master oscillator. By use of the low-capacity 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 cycle of sine waves.

An alternate method is the old standby of employing a 31.5-kc frequency standard and some form of comparison. The "beat" method is subject to error in pulse circuits due to the rich harmonic content. The frequency standard signal may be displayed on the scope, being triggered externally by the same signal. Without disturbing the external trigger, observe the MO output and adjust the circuit for the same frequency and maximum stability of trace.

Whenever a sync generator becomes erratic, the first step should be to change the MO control from the usual line-lock position to the free-run or crystal control. If the pulses stabilize, the MO frequency may be out of range of AFC control, or the AFC circuitry is erratic.

Checking the Countdown

Some of the older-type sync generators employing step-by-step counters featured continuous display of counter circuits action by means of a small scope on the front panel. More recent generators have deleted this feature 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 TG-2A the first counter (7/1) contains a variable adjustment while the remaining counters use fixed constants with no adjustable controls. Test points are generally provided at the various counter outputs on all generators for the purpose of testing and servicing.

Because of the nature of binary counters employing feedback pulses to obtain the odd-number countdown (525-1), a direct counting check is usually not practical. Trouble in the preceding binary stage is indicated by the first grid check point where random or unstable pulses exist.

Generators using multivibrator-type counters (such as the RCA-TG2) normally employ four stages as follows:

STAGE NO.	INPUT FREQUENCY	DIVISION	OUTPUT FREQUENCY
1	31,500	7	4,500
2	4,500	5	900
3	900	5	180
4	180	3	60

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 7/1 divider circuit by adjusting the sweep to obtain 1 cycle of the waveform shown by the upper trace in Fig. 3-20. The counter should show a division of 7.
2. Apply the same signal to the external sync input of the scope and adjust the sweep for five complete cycles.
3. Without changing the external sync connection or the sweep,

place the scope vertical amplifier probe in the following (5/1) counter. Exactly one cycle should now occur in the same interval, as shown by the lower trace in Fig. 3-20.

Repeat the same procedure for the remaining stages by triggering the sweep with the counter *preceding* the one to be checked, and adjusting the sweep for the number of cycles that equals the division of the counter to be checked.

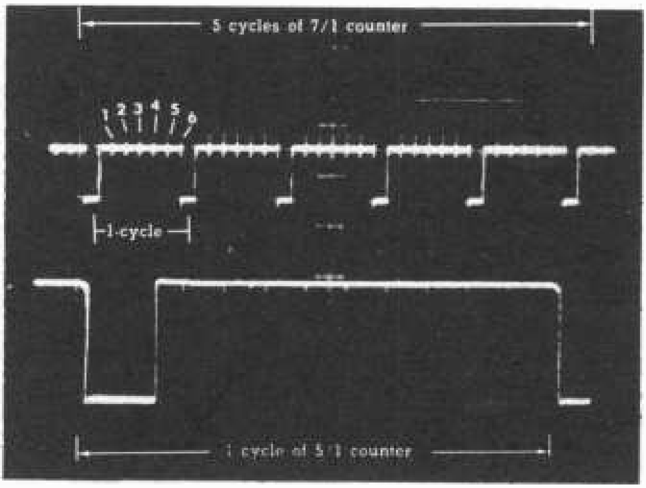


Fig. 3-20. Upper trace is five cycles of 7/1 counter; lower trace shows one cycle of 5/1 counter.

An alternate method (less accurate) is to measure the frequency of the pulse by using the time base of the scope in microseconds (see Table 3-1). In this case the engineer must assure himself that the time base is both accurate and linear (Section 1).

Setting Pulse Widths

Pulse widths are set by either of two general methods:

(A) Variable time constant of triggered multivibrators.
(B) 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 between various manufacturers. It is usually important to follow the recommended sequence of adjustments in the instruction manual for the particular generator involved.

For convenience, pulse widths in terms of both H and microseconds are listed in Table 3-2. Fig. 3-21 is an exaggerated draw-

Table 3-1. Time Base of Counter Circuits

STAGE	TOTAL DIVISION	FREQ	TERMS OF H	WIDTH OF PULSE* IN MICROSECONDS
MO	1	31,500	½	31.75
÷ 7	7	4,500	3.5	222.0
÷ 5	35	900	17.5	1,111.0
÷ 5	175	180	87.5	5,556.0
÷ 3	525	60	262.5	16,668

* From leading edge of pulse to leading edge of following pulse.

ing illustrating the standards of pulse measurements. The width is normally measured at 0.9 peak amplitude where "peak" is taken as the positive-going direction. (Horizontal blanking is an exception where, for the purpose of setting aspect ratio, width is measured at the 50% amplitude point.) Also note that pulse rise and decay times are measured as the time interval between the 10% and 90% amplitude points.

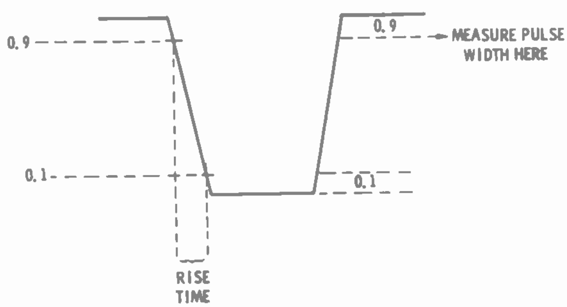


Fig. 3-21. Standard for measuring pulse width and rise time.

By cross-mixing sync and blanking outputs and adjusting the scope time base to obtain the waveform shown by Fig. 3-22, horizontal sync, equalizing, horizontal blanking, and vertical serration widths may be adjusted from this one display. The markers are 0.005H (½ of 1% of H) for maximum accuracy. Table 3-3 lists the proper number of 0.005H markers for the appropriate intervals.

From Table 3-2, H sync should be adjusted to 4.8 microseconds, or 0.075H. For a width of 7.5% of H there will be 15 of the 0.005H markers present as shown. The 1 microsecond markers could be used here if desired, as well as for setting total width of H blanking to 11.1 microseconds, with slightly less accuracy. The vertical sync serration should be 4.5 microseconds, or 7%H, which is

Table 3-2. Pulse Table

PULSE	MINIMUM		NOMINAL		MAXIMUM	
	H	Micro-seconds	H	Micro-seconds	H	Micro-seconds
H Blanking	0.165	10.5*	0.175	11.1*	0.178	11.3*
H Sync	0.07	4.445	0.075	4.8	0.08	5.08
Equalizing	0.45 of H. Sync	2.0	0.5 of H. Sync	2.4	0.04	2.54
V. Serration	0.06	3.8	0.07	4.5	0.08	5.08
V. Blanking	18.375	1167*	19.7	1250*	21	1333*

* H and V blanking must be of proper ratio to establish 4/3 aspect ratio.

NOTE: EIA resolution and ball charts are based on following values:
H. Blanking = 11.1 Microseconds
V. Blanking = 1,250 Microseconds

indicated by 14 of the 0.005H markers. Where the time base and sweep linearity are known to be accurate, the graticule lines only may be used. For example, the vertical serration could be set by adjusting the scope for 2.25 microseconds/cm, and the serration adjusted to cover exactly 2 cm.

Vertical blanking can be accurately set by using horizontal drive pulses as markers. By using a 1,000-ohm resistor from the H 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 3-2,

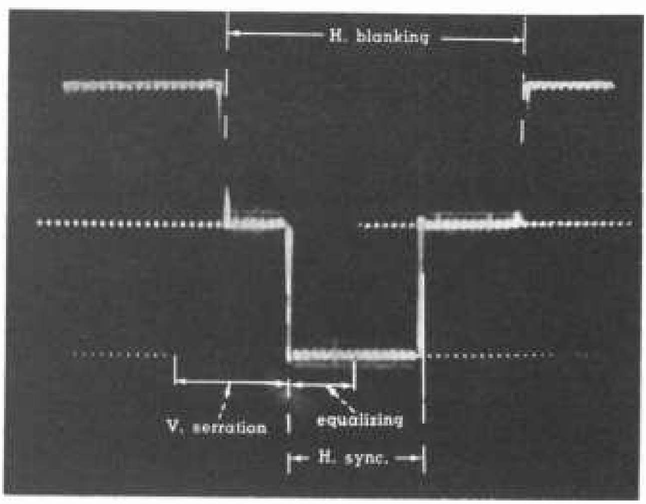


Fig. 3-22. Cross-mixed sync and blanking with 0.005H markers.

Table 3-3. 0.005 H Table

PULSE	MIN.	NOMINAL	MAX.
H. Blanking	33	35	35.6
H Sync	14	15	16
Equalizing	6.3	7.5	8
V. Serration	12	14	16

vertical blanking (nominal value) is 19.7H; thus, 19 markers plus 0.7 indicates the proper width. The 0.7 value can be interpolated quite accurately, provided the operator is familiar with the linearity of the scope sweep.

Actually the use of markers as described is not necessary if the scope incorporates variable delayed sweep, as provided in the Tektronix 524-AD. In this method simply leave sync and blanking cross-mixed, as in H pulse measurements, use the 200-

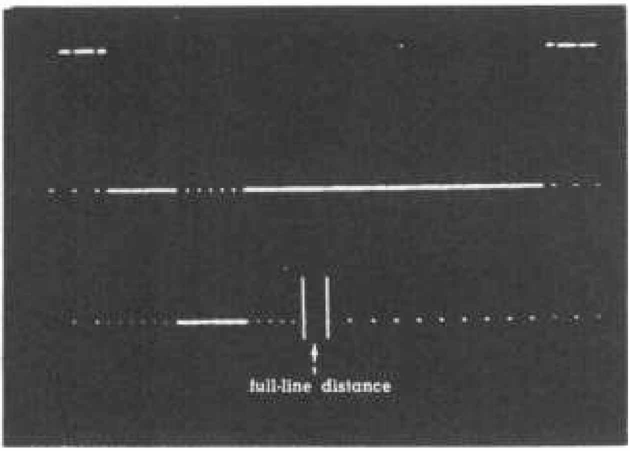


Fig. 3-23. Vertical blanking interval (waveform P of Fig. 3-1) adjusted to 1.9H.

microsecond time base and delayed sweep, and select the field with the full line (H) interval following the last equalizing pulse (Fig. 3-23). The last equalizing pulse is the end of the first 9H interval of vertical blanking, plus the 0.25H to 1.025H vertical front porch which should be counted. The remaining H sync pulses may then be counted to arrive at the total vertical blanking interval as shown.

3-4. 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 H 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 V blanking and sync in the normal active line area of the monitor. This displays both fields (interlaced) 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 cps), a single field is displayed with half the number of pulses, (noninterlaced presentation). In either case, expansion of the vertical sweep is normally used to allow more critical observation of the pulses.

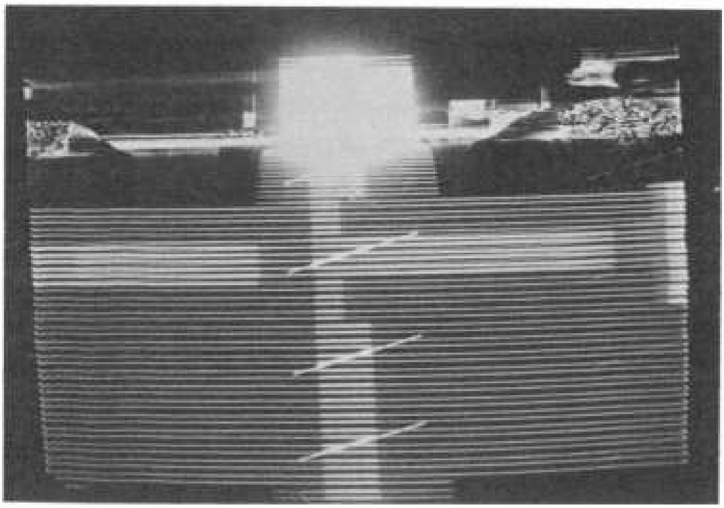


Fig. 3-24. Interlaced pulse-cross display on composite signal.

Fig. 3-24 is the pulse-cross presentation of a line-output signal with 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 horizontal front-porch, sync, blanking, equalizing, and V sync widths. Vertical blanking is conveniently checked by counting the number of blanking lines. Some stations construct graticules with normal pulse widths indicated after an accurate check of the generator with the oscilloscope.

The pulse-cross is extremely useful both as a continuous monitor and as a servicing tool. A 9 × 1 switch panel is used at station WTAE to allow selection of signal from a number of points to feed the monitor, but this switch panel is normally left on the "stand-by generator" position. This enables continuous monitoring of whichever generator is in the stand-by position (composite sync only), as shown in Fig. 3-25. Fig. 3-26 is an exploded view of this presentation with identification of pulses.

The "cross" is formed at the in-line-with-H sync period. The reader can readily understand the sequential presentation of the

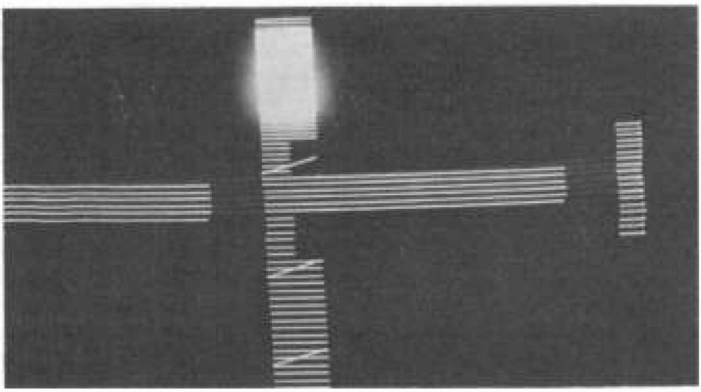


Fig. 3-25. Pulse-cross display, sync only.

monitor (Fig. 3-26) if he will visually move field 2 (waveform P of Fig. 3-1) to the right one-half line so that the H sync pulses of both fields are in vertical alignment and the last H pulse of field 2 is in line with the first equalizer of field 1. Now, observing Fig. 3-26, note that the in-line pulses (those occuring at H intervals)

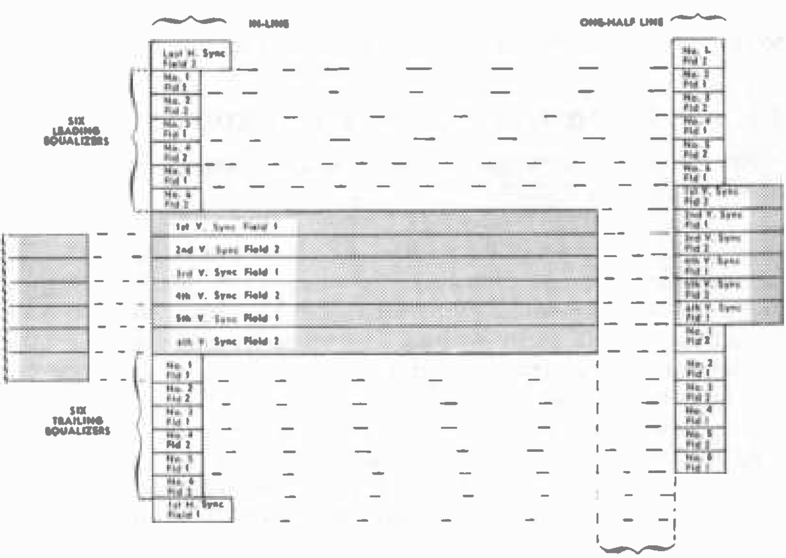


Fig. 3-26. Exploded view of pulse cross with pulse identification.

are equalizers 1, 3, and 5 of field 1, and 2, 4, and 6 of field 2, spaced on alternate lines due to interlace of fields. The half-line intervals and the remainder of the presentation is obvious in pulse identification by following the preceding procedure.

With an interlaced type of pulse-cross monitor loss of interlace, such as could be caused by a vertical countdown error of 0.5H,

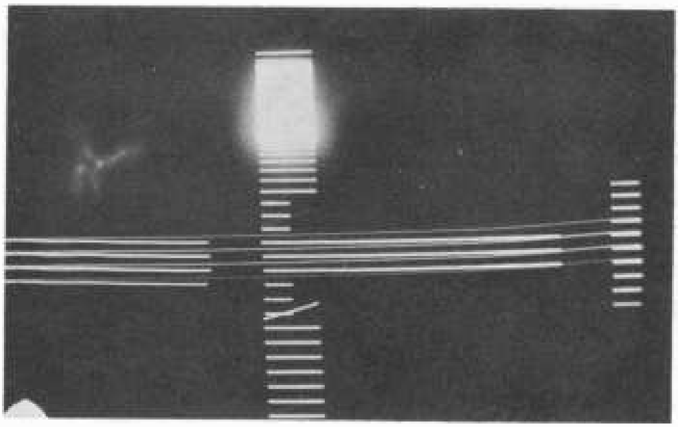


Fig. 3-27. Interlaced pulse-cross display when sync generator has lost interlace. Note the apparent loss of one field.

is readily apparent, as illustrated by Fig. 3-27. Brightness of the display will be greater than normal for a given adjustment due to the double tracing of identical raster lines.

3-5. SUGGESTED MAINTENANCE PROCEDURES

An adequate preventive maintenance schedule can be outlined as follows:

30 Days: Stability, pulse widths and EIA Standards check. This includes an overall stability check where possible, a check (and adjustment if necessary) of MO frequency, counter chain, and all pulse widths, as well as measurement of pulse amplitude, overshoot, and rise times both at the generator output and all pulse distribution amplifier outputs with cables connected as in normal operations.

90 Days: Tube check and a repeat of the 30-day schedule. All tubes should be checked for transconductance, shorts, and gas. Normally the "good-bad" scale which allows manufacturers tolerance of performance is sufficient for the transconductance check. When tubes are replaced, it is always important to recheck pulse widths in about 3 days and again after 7 days. When experience dictates that pulse widths do not hold within minimum to maximum tolerances for 30 days, this schedule should be shortened accordingly.

The overall stability test is a good indication of any deterioration of tubes and/or components. The test is possible if the MO has a vernier frequency adjustment which will allow a 3 to 5% frequency variation from the exact operating frequency. In this test it is assumed that the MO has been adjusted to 31.5 kc in the free-running mode with the vernier control at midrange by one of the methods previously described.

Cross-mix sync and blanking at the test points and observe this on the scope or pulse-cross monitor. When using the Tektronix 524 scope, the best display is obtained by using delayed sweep on a 200 microsecond time base and rotating the delay control to observe an entire vertical blanking field plus several lines. The field-shift key should then be used to observe the opposite field. This procedure is unnecessary if an interlaced pulse-cross is available.

Rotate the vernier frequency control through the extremes of rotation. Maximum stability is indicated if no erratic behavior of pulses exists between extremes. If H pulses become erratic or change frequency, check the divide-by-two circuit (15.75 kc) with the scope. If this output remains stable, check each following H 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 and V sync instability occurs, check any 9H gating circuit and/or 3H stop or delay circuits and particularly the counter chain.

Keeping tabs on the rise and decay times of pulses is also a good procedure from the standpoint of troubles casting their shadows before them. The time interval of leading and trailing edges of H and V sync, equalizing pulses, H blanking, and H drive 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, as shown by Fig. 3-21. 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% as measured by a scope with good transient response (Normal position on the Tektronix 524-AD) and with interconnecting cables in place. When long coax runs are involved, an overshoot with "ringing" sometimes occurs. If the pulse waveshape is important (not used simply as triggers), the termination should be varied slightly around the nominal value for best transmission. It is sometimes necessary to provide complex terminations on extremely long cable runs.

3-6. SYNC CROSSTALK

This term can be applied either to crosstalk within the sync generator itself, or to the so-called "windshield wiper" effect similar to the horizontal motion caused by co-channel interference on a home receiver.

Crosstalk within the generator itself is usually caused by a very small amount of any one of the counter frequencies "leaking" into the MO waveform. This type of crosstalk 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 crosstalk can be determined by considering the number of such horizontal line displacements occurring from top to bottom of the raster as follows:

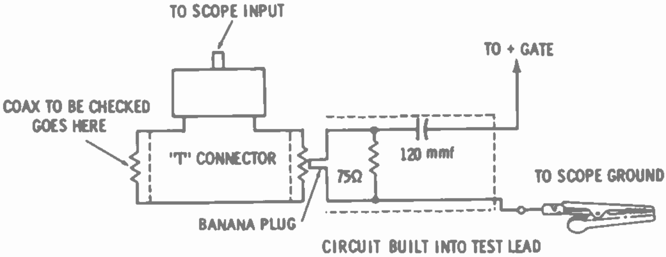
CROSSTALK FREQUENCY	NUMBER OF DISPLACEMENTS TOP TO BOTTOM
4,500	70
900	14
180	3

If this effect should be evident on all driven monitors, the counter indicated should be additionally shielded or wiring rerouted until the interference is eliminated.

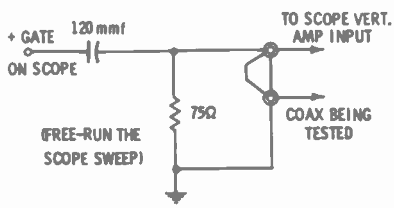
A much more prevalent type of sync crosstalk is that which occurs between two nonsynchronous sources, such as local and network or local and video-tape signal. This trouble is evident as a vertical bar or line moving back and forth horizontally through the signal when a remote source not tied to the local sync generator is observed. When this condition exists, the local sync will also crosstalk with the signal from the video-tape recorder in the 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 coax cable. Each cable should be disconnected from the sending end and checked with an ohmmeter from center conductor to shield to determine if the termination resistance is obtained. If the shield is open, no continuity will exist. Always twist both 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 only one person is possible with the construction of the simple tester shown in Fig. 3-28. Fig. 3-28A illustrates the physical construction and Fig. 3-28B shows the schematic view. The two components (resistor and capacitor) can be part of the cable running from the plus gate binding post to the UHF connector. The line to be checked is connected to the opposite UHF "T" connector. The arrangement shown in Fig. 3-28 is suitable for use with the Tektronix wide-bandwidth scopes which have sweeps adjusted to cover only the 10-centimeter graticule, such as the 551, 535, etc. Procedure is as follows:



(A) Physical construction.



(B) Schematic diagram.

Fig. 3-28. Coaxial line tester for Tektronix 551 and similar scopes.

1. With the connection shown in Fig. 3-28A, 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. If the cable is properly terminated, no return pulse will appear and the base line will be perfectly smooth. A slight bump may appear at junctions, such as jack fields, splices, etc.
3. If the cable is open, a reflected pulse of positive polarity will appear (Fig. 3-29A). If the cable is shorted, a negative reflected pulse occurs as in Fig. 3-29B.
4. The distance to the discontinuity can be very closely obtained by the following analysis.
   (a) The speed of light in feet per microseconds = 983.5 ft/microseconds.

(b) Multiply this figure by the propagation factor of the cable used (See Table 3-4).
(c) Divide by two. (The pulse must travel to the discontinuity, then return). This is the multiplying factor for the indicated space in microseconds.

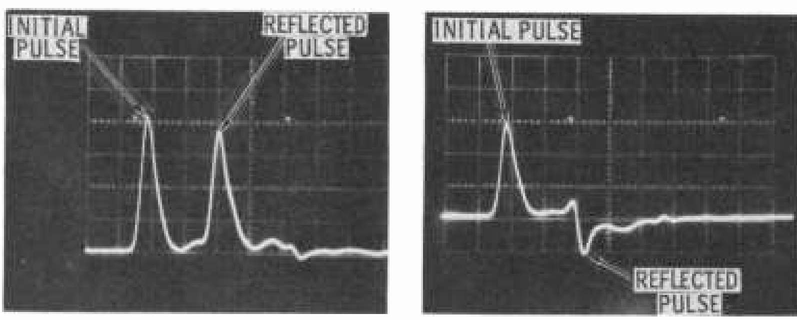
Table 3-4. Coaxial Cable Chart for Tester

CABLE TYPE	PROPAGATION FACTOR	MULTIPLYING FACTOR
Solid poly	0.66	325
Foam poly	0.82	404
½" styro	0.89	439
¾" styro	0.90	443

The graphs in Figs. 3-30 and 3-31 are presented for convenience where the cable has a propagation factor of 0.66, as is common for most video runs. In this case the multiplying factor is 325. Thus, if the indicated delay of the reflected pulse is 0.5 microsecond, the length to the discontinuity is:

$$325 \times 0.5 = 162.5 \text{ feet}$$

The photos of Fig. 3-29 were made with a scope sweep time base set on 0.1 microsecond/cm, indicating approximately 72 feet to



(A) Open circuit indication. (B) Shorted cable indication.

Fig. 3-29. Scope displays indicating coaxial cable defects.

the discontinuity, or 0.22 microseconds. The Tektronix Model 551 scope was used.

NOTE: When the discontinuity is at a distance of less than approximately 40 feet (causing the reflected pulse to ride on the initial pulse), use a 0.1-microsecond delay line in series with the cable tested and subtract this amount from the final computation. A 70-foot roll of RG 11/U or 59/U wil provide this amount

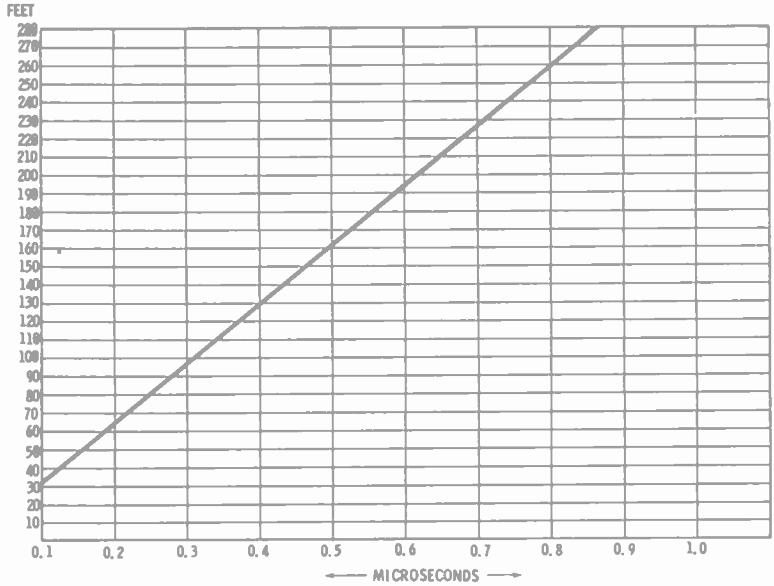


Fig. 3-30. Delay in microseconds versus cable lengths (10 to 280 feet) for propagation factor of 0.66.

of delay in lieu of a delay cable. (These cables have a delay of approximately 1.5 microseconds per 1,000 feet.)

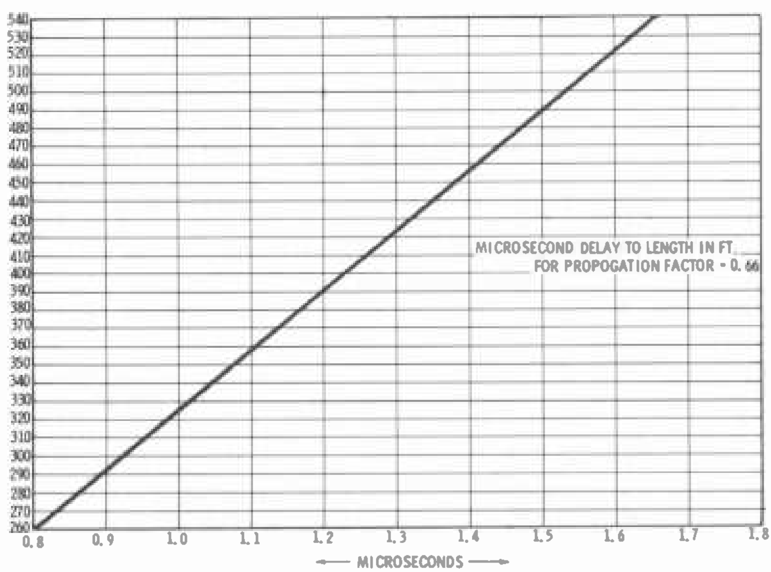


Fig. 3-31. Delay in microseconds versus cable lengths (260 to 540 feet) for propagation factor of 0.66.

Due to the much longer horizontal sweep of the Tektronix 524 scope, the simple circuit of Fig. 3-28 is not suitable for this unit; however, the arrangement shown in Fig. 3-32A can be used. Although obviously more complex than the previously described tester, the transistor and all other components are included within the cable as before. Due to initial pulse inversion in the transistor circuit, the minus gate output of the scope is used. The magnifier positioning voltage divider within the scope was modified as shown in Fig. 3-32B to obtain the very small amount of current required at −20 volts, which is approximately .5 to .75 ma. Resistor R327 in the scope (Model 524-AD) was replaced with the two resistors shown, and the junction was brought out to a jack (or binding post) on the front panel. The pulse output is approximately 0.125 microsecond wide at the base.

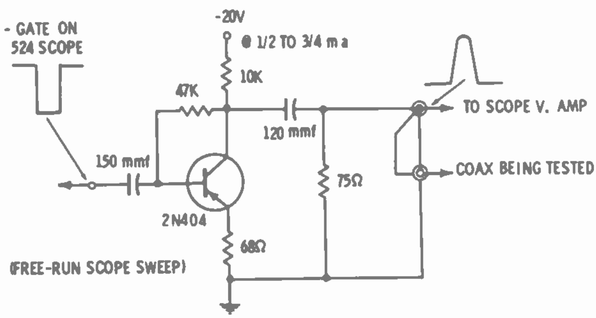
3-7. PULSE DISTRIBUTION

Pulse-distribution amplifiers are used to provide isolated runs of the required pulses to various points in the plant. Fig. 3-33 shows the basic pulse distribution of station WTAE in Pittsburgh. Note that since monochrome and color must be integrated in this case, all monochrome pulse runs occur on a delayed basis to that of the color system. This is necessary to maintain the same front-porch width on the composite transmitted signal between color and monochrome signals. A shift in front porch would cause a shift in the raster (picture) area at the receiver. Since the normal delay through a color system is about 1.2 microseconds, horizontal drive, blanking and sync pulses must be delayed accordingly for monochrome. It may be observed that with sync pulses inserted after the final switching point in the system, if set for the normal 1.6 microseconds front porch on monochrome, only about 0.4 microsecond of front porch would exist for the color system.

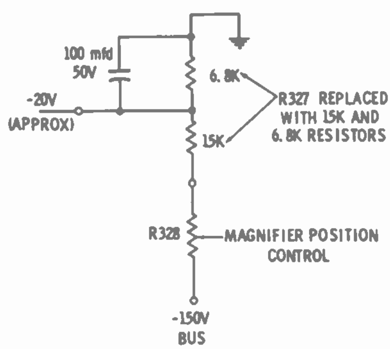
Pulse distribution amplifiers are normally unity gain devices used strictly for isolation of feed lines to the many points requiring pulses. Without this isolation, no interconnected unit could be disconnected while the station is on the air. In some applications the amplifiers are used to regenerate the pulse to "clean up" rise times, overshoots, ripple, etc. as caused by some of the older types of delay lines.

Some amplifiers provide two to six outputs for a single input. The type shown in Fig. 3-33 are single-in, single-out units of the plug-in variety for flexibility in meeting emergencies or servicing and maintenance. Actually, only a single coax provides the input for a complete group; the inputs are carried by straps on the plug receptacle to adjacent units in a group.

Pulse distribution outputs should be checked on a regular basis for rise times, overshoots, ringing, proper amplitude, and any indication of hum or noise. When any deterioration is evident, the external coax feed should be removed during maintenance time and a termination provided directly at the amplifier output to determine whether the trouble is internal, or caused by the external feed. The pulse amplifier is usually either a "go-no-go" device using regenerative clippers or multivibrators, or it employs



(A) Schematic of the coaxial-line tester.



(B) Scope modification to supply —20 volts for the transistor.

Fig. 3-32. A transistorized coaxial-line tester for use with the Tektronix 524 scope.

heavy negative-feedback circuits. In either case, the best tube check that can be performed is a performance test using an oscilloscope. Transistorized pulse distribution amplifiers are being used more frequently in modern installations and the performance test is mandatory.

3-8. COMPOSITE PULSE CHARACTERISTICS

The preceding paragraphs of this section have considered only the horizontal and vertical drive signals and composite blank-

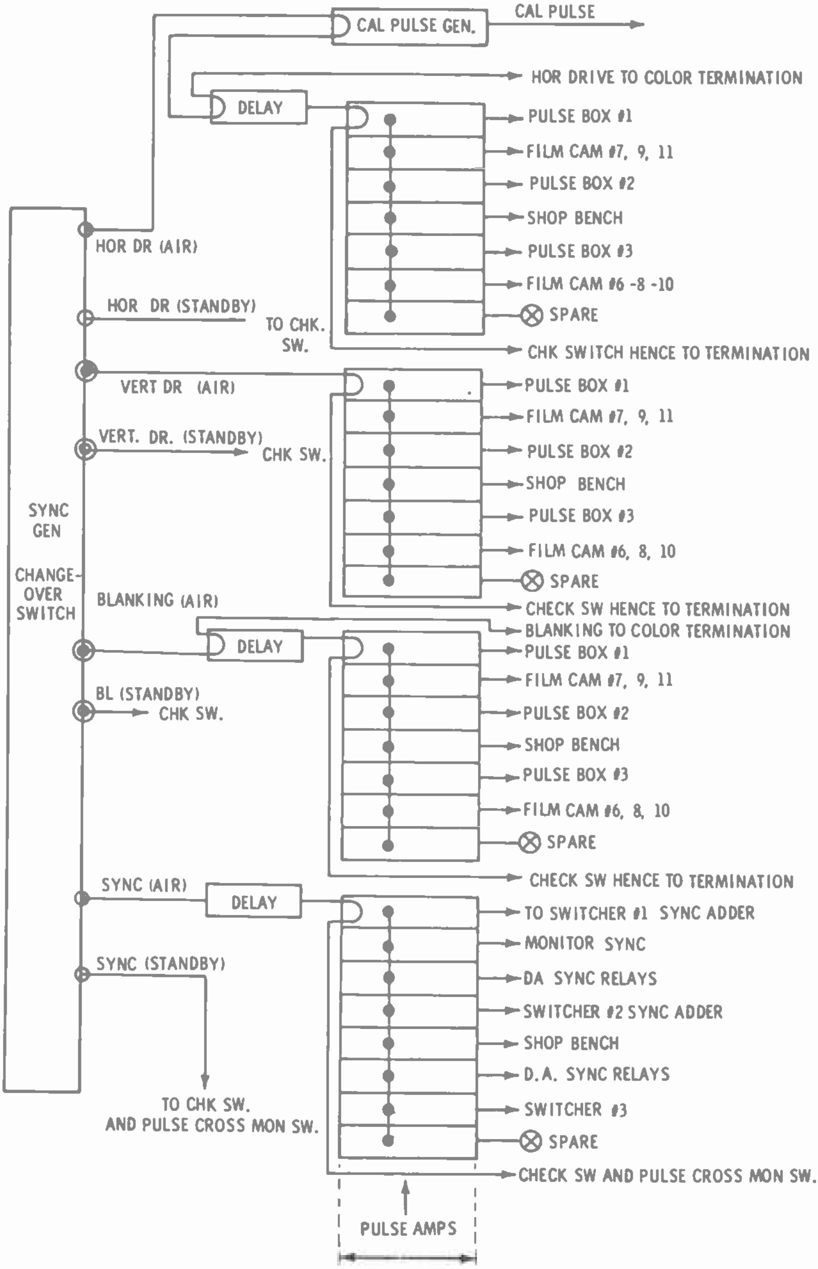


Fig. 3-33. Pulse distribution system of station WTAE.

ing and composite sync. The actual composite signal as fed to the transmitter for broadcasting involves the proper combination of composite blanking and composite sync to establish authorized front and back porches, as well as the effect of combining the video signal (over the entire gamut of duty cycles or average picture levels) on the pulses from the sync generator.

Setting the Front-Porch Width

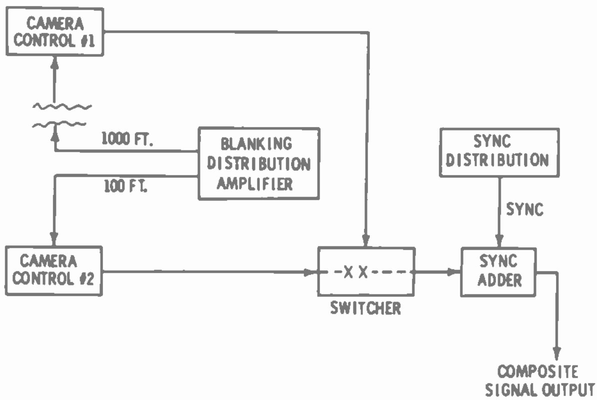
When centralized camera controls are employed with approximately the same length camera cables (within 100 feet), no special problem exists in adjusting the front-porch width. Keep in mind that this width is determined by the adjustable delay in formation of the horizontal sync pulse from the leading edge of horizontal blanking.

Blanking is normally inserted in the camera control unit, and sync is inserted either in the final studio switching system or in a stabilizing or sync-adder amplifier following the studio switcher. In a few cases blanking is also inserted in the stabilizing or adder amplifier at a fixed amplitude of between 5% and 10% of the composite signal level. The procedure in this case is quite simple.

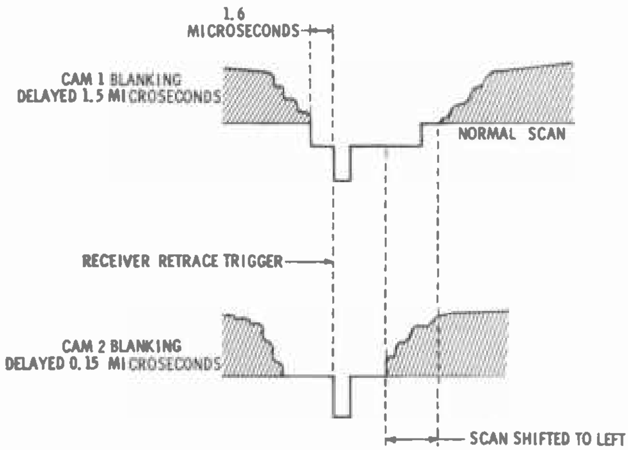
1. Transmit a test pattern or any slide or chart with reference white around the outer extremities. (The video level at the picture edge must be sufficient to be readily measured on the CRO.)
2. Observe the composite signal after the final switching point, which includes the sync pulse.
3. Adjust the front-porch control (this is either a multivibrator pulse-width control or a delay-line adjustment) until the interval between the end of the picture information and the leading edge of horizontal sync is between 1.27 microseconds (minimum) and 1.59 microseconds (maximum). This meets the FCC specification for front-porch width of 0.02H to 0.025H. Users of the Tektronix 524 scope normally use the 0.025H markers in setting this adjustment.

When camera control units are not centralized, or where camera cable lengths may vary from more than 100 up to 1,000 feet, the problem becomes more complex. Fig. 3-34 illustrates an example where a 900-foot difference exists between two control units and the system blanking distribution. In this case the blanking pulse is delayed approximately 1.5 microseconds to control unit 1, and only 0.15 microsecond to control unit 2. If the sync generator, front-porch width control is adjusted to obtain a normal front porch when observing the camera 1 signal after sync insertion, then a switch to camera 2 signal will result in a lengthened front porch (Fig. 3-34B). Since the receiver retrace is triggered by the leading edge of horizontal sync and the picture is unblanked by the end of horizontal blanking, a lengthened front porch causes the picture area to shift to the left. Similarly, if the front porch is adjusted for normal on the camera 2 signal, a switch of the camera 1 signal will result in a narrowed front porch, and the picture would shift to the right on receivers.

Thus, where camera control units are more than 100 feet (0.15 microsecond) from each other with respect to the system blanking distribution, it is necessary to add delay lines to the nearest control units to equal the delays of the farthest units. This is most conveniently accomplished by simply using the same length of



(A) Wide variation in cable length causing delay.



(B) Effect of cable length on scan waveform.

Fig. 3-34. The effect of a blanking delay on the front-porch width.

feed lines to every control unit; excess cable can be coiled up when necessary.

A similar problem can exist even when centralized control units are used but the length of camera cable is greatly different between units. Camera blanking is normally formed from horizontal and vertical drive pulses. (Vertical-rate pulses are of no consequence in this discussion since the few microseconds of delay encountered has no bearing on the long vertical blanking interval of 1,250 microseconds.) Camera blanking must "fit under" the composite signal blanking inserted in the control unit. For this reason the pickup tube (camera) blanking is normally about 7 to 9 microseconds compared to the 11-microsecond horizontal blanking transmitted.

Now if the camera cable should be as much as 1,000 feet in length, the total camera blanking delay is 3 microseconds (1.5 microseconds per 1,000 feet, and the pulse experiences delay both to and from the camera). It may be observed here that if camera blanking duration is longer than 8 microseconds, the interval is not completed by the end of receiver blanking. When this occurs, the front porch width is not determined by the delay in the sync generator between the leading edge of blanking and the leading edge of sync, but rather is actually determined by the end of camera blanking. Again, if one camera cable length is only 100 feet, and another camera cable length is 1,000 feet, switching from one to the other will cause a picture area shift on the receiver. In this case *horizontal drive* pulses must be delayed to the camera units with the *shortest* cables to equal those of much longer lengths.

In any event, the maintenance engineer is charged with the responsibility of assuring that the front-porch width remains within the FCC specifications from all signal sources. Preferably this difference should be maintained within no more than 0.1 microsecond between sources.

Allowable Pulse Degradation

It has been the impression of some engineers that if the studio sync generator is maintained within EIA standards (which all commercial sync generators equal or exceed), the transmitted waveform, as specified by the FCC, can readily be obtained.

This is a dangerously false impression, as the experienced engineer well knows. Actually, very little degradation of pulses is allowed between EIA sync generator waveform standards and the FCC radiated waveform specifications. Since this aspect of television system maintenance has been scantily covered, Figs. 3-35 through 3-37 are presented to trace allowable degradation. Fig. 3-35 is the EIA studio sync generator standards; Fig. 3-36 is the

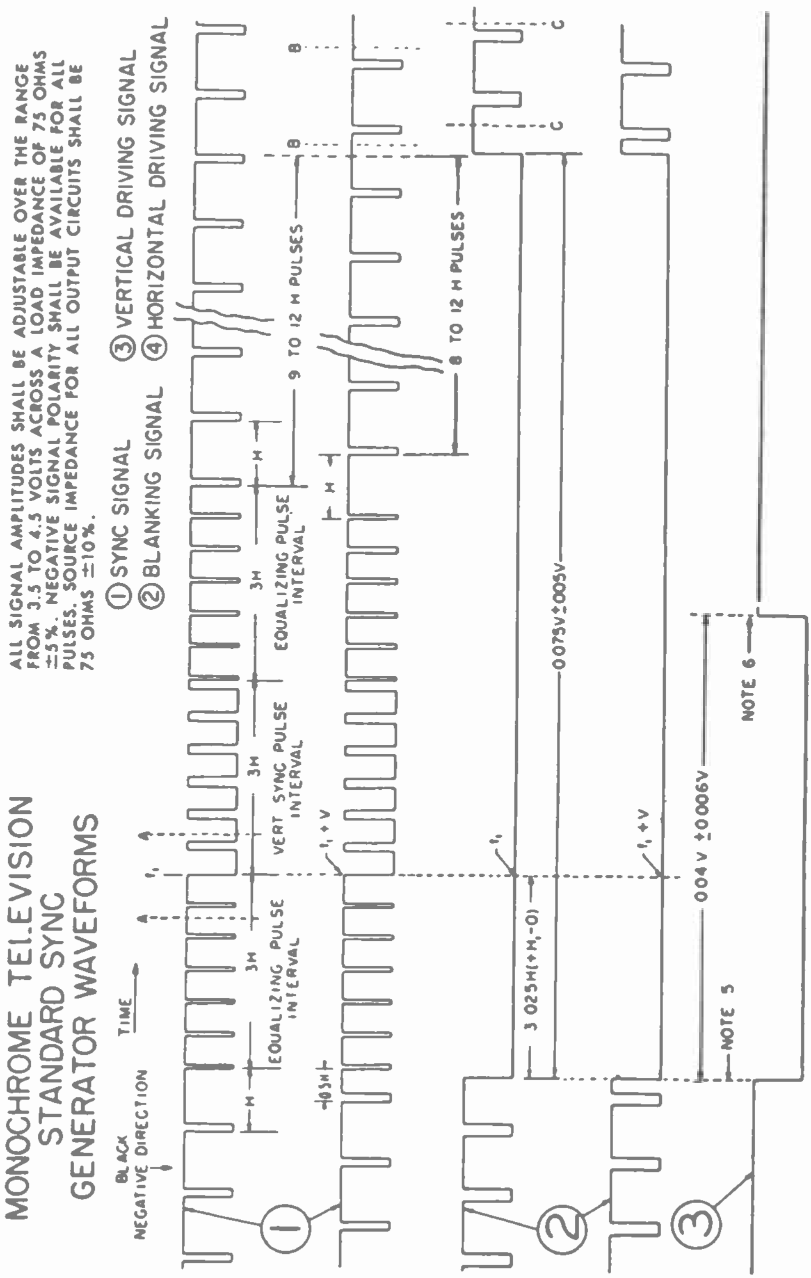
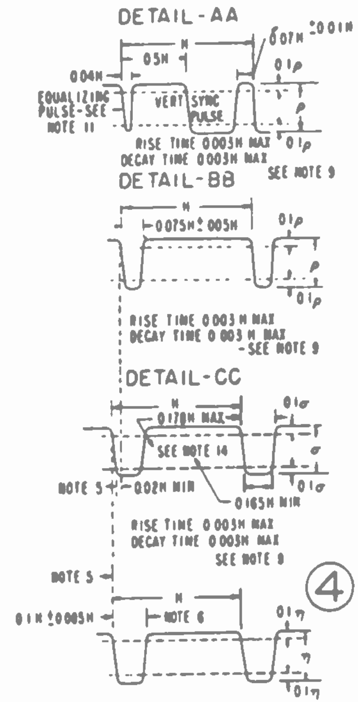


Fig. 3-35. Recommended

sync generator waveforms.

NOTE:

1. H = time from start of one line to start of next line.
2. V = time from start of one field to start of next field.
3. Leading and trailing edges of vertical driving and vertical blanking signals should be complete in less than 0.1H.
4. All tolerances and limits shown in this drawing are permissible only for long time variations.
5. Timing adjustments, if any, shall include this condition.
6. The vertical driving pulse duration shall be 0.04V, ±0.006V. The horizontal driving pulse duration shall be 0.1H, ±0.005H.
7. The time relationship and waveform of the blanking and sync signals shall be such that their addition will result in a standard RETMA signal. The time relationship shall be adjustable in order to satisfy this relationship for the condition where the blanking signal is delayed with respect to the sync signal over the range from 0.0H to 0.05H.
8. The standard RETMA values of frequency and rate of change of frequency for the horizontal components of the sync signal at the output of the picture line amplified shall also apply to the horizontal components of the output signal from the recommended sync generator.
9. All rise and decay times shall be measured between 0.1 and 0.9 amplitude reference lines.
10. The time of occurrence of the leading edge of any horizontal pulse "N" of any group of twenty horizontal pulses appearing on any of the output signals from a standard sync generator shall not differ from "NH" by more than 0.0008H where H is the average interval between the leading edges of the pulses as determined by an averaging process carried out over a period of not less than 20 nor more than 100 lines.
11. Equalizing pulse area shall be between 0.45 and 0.5 of the area of a horizontal sync pulse.
12. The overshoot on any of the pulses shall not exceed 5%.
13. The output level of the blanking signal and the sync signal shall not vary more than ±3% under the following conditions:
    A. The A.C. voltage supplying the sync generator shall be in the range between 110V and 120V and must not vary more than ±5V during test.
    B. A period of 5 hours' continuous operation shall be considered adequate for this measurement, after suitable warm-up.
    C. The room ambient shall be in the range between 20 and 40 degrees C. and shall not change more than 10 degrees C. during this test.
14. Adjustment shall be possible between minimum and maximum limits so that the aspect ratio can be set to the normal value.



Rise and decay times shall be equal to or less than rise and decay times of horizontal blanking—See note 9

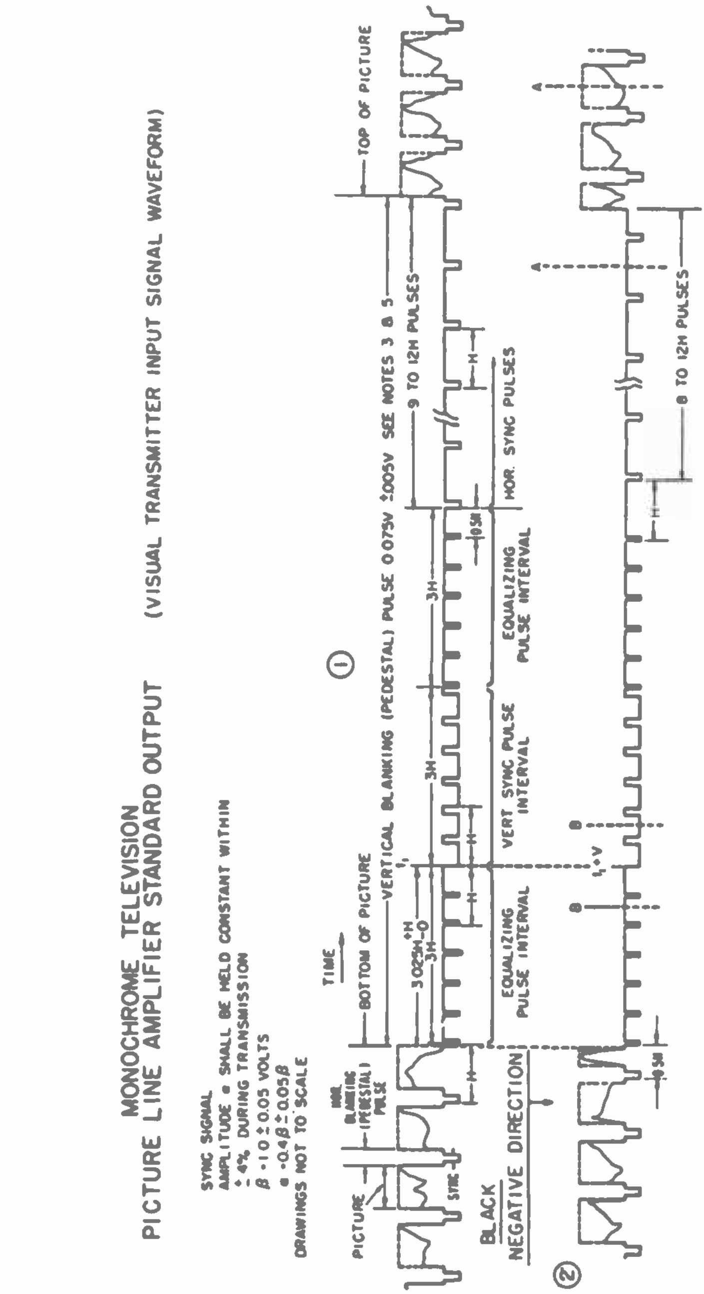
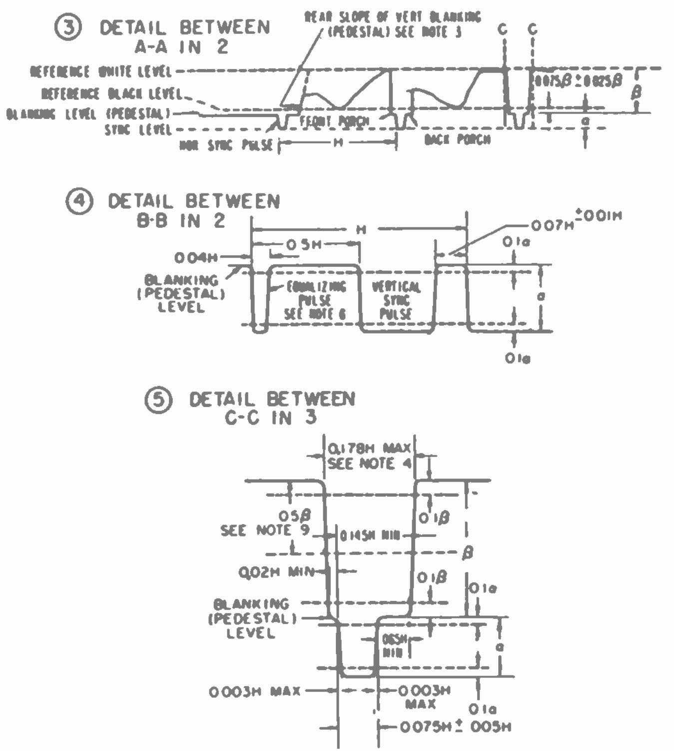


Fig. 3-36. Picture



RISE TIME & DECAY TIME OF ALL PULSES 0.003H MAX.
(SEE NOTE 7)

line amplifier output.

NOTE:

1. H = time from start of one line to start of next line.
2. V = time from start of one field to start of next field.
3. Leading and trailing edges of vertical blanking (pedestal) should be complete in less than 0.1H.
4. Leading and trailing edges of horizontal blanking (pedestal) shall be steep enough to preserve min. and max. values of durations under all conditions of picture content.
5. All tolerances and limits shown in this drawing are permissible only for long-time variations.
6. Equalizing pulse area shall be between 0.45 and 0.5 of the area of a horizontal sync pulse.
7. All rise and decay times shall measure between 0.1 and 0.9 amplitude reference lines.
8. The overshoot on blanking (pedestal) signal shall not exceed 0.02β at the beginning of the front porch and shall not exceed 0.05β at the end of the back porch. The overshoot on sync signals shall not exceed 0.05a. Any other extraneous signals on the blanking (pedestal) signal shall not exceed .02β.
9. For setting aspect ratio, the horizontal blanking (pedestal) duration should be set to 0.175H at the 0.5β point.
10. Rise time and decay time of horizontal blanking shall not exceed 0.003H.

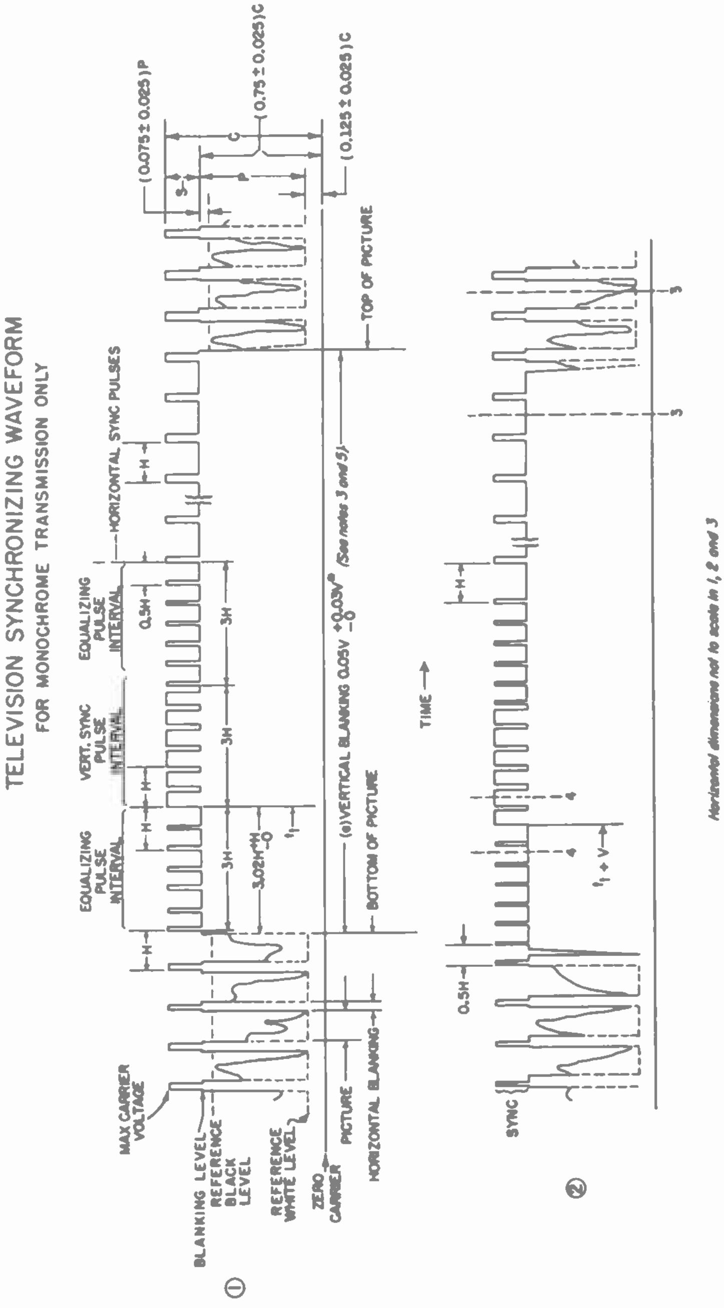
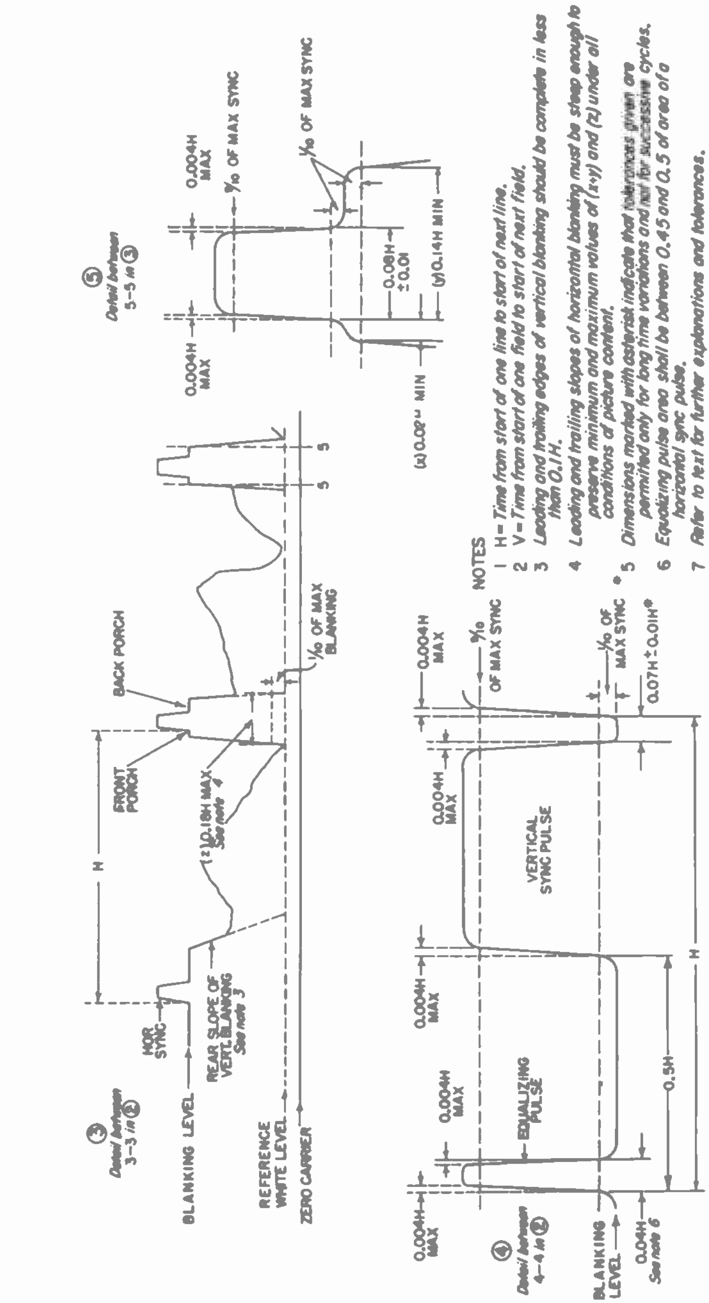


Fig. 3-37. Transmitter



output waveform.

EIA picture line amplifier output (transmitter input) standards; and Fig. 3-37 presents the FCC radiated waveform specifications.

Note that in Fig. 3-36 there is no allowance for degradation of pulses at all from the sync generator output. In addition, all amplifiers handling the composite signal to be transmitted must be capable of rise times (now carrying both pulse and picture components) to preserve minimum and maximum values of:

(1) front porch,
(2) leading edge of sync to end of back porch,
(3) base of white level interval (0.18H maximum width).

In addition, overshoot on the blanking signal must not exceed 2% of the sync pulse amplitude at the beginning of the front porch, or 5% at end of the back porch. Overshoot on the sync signal must be no greater than 5% of sync pulse amplitude. Extraneous signals (hum, etc.) must not exceed 2% of the composite signal level. The use of a stabilizing amplifier (which reforms pulses) before the transmitter input may be necessary and is usual.

From Fig. 3-27 it can be observed that only the rise time of horizontal sync, equalizing, and vertical sync pulses have allowable degradation from the EIA studio sync generator output standards (from 0.003H, or 0.19 microsecond, to 0.004H, or 0.25 microsecond). This is due to the restricted bandwidth of transmission, which must be no more than 4.18 mc, and down at least 20 db at 4.5 mc.

While the complete testing of the transmission path for bandwidth and transient response requires special test equipment and techniques, the engineer has a continuous display of these characteristics if he is well acquainted with such restrictions on the composite pulse shapes. After some experience in the field, an interpretation of the CRO display of the composite signal will indicate the general path conditions to the engineer. Following is a brief outline of the pulse interpretation as to system condition as discussed in detail in Sections 5 and 6:

1. Poor low frequency response is indicated by a downward tilt on the base of vertical interval pulses (upward tilt at the tips) and a loss of "setup" from the originally adjusted value.
2. Poor high-frequency response is indicated by a rounding of the horizontal sync pulses compared to those leaving the sync generator.
3. Poor transient response is indicated by overshoots or undershoots on the pulses, and on the beginning and ending of the porches.

SECTION 4

THE VIDEO SWITCHER

Since most TV stations (even the smallest ones) must use more than one camera source at some time during their daily schedule, some means for switching signal sources must be included. A means of fading out or in on a given signal and the momentary mixing of two separate video signals for purposes of lap-dissolving or deliberate superimposition of pictures are also needed.

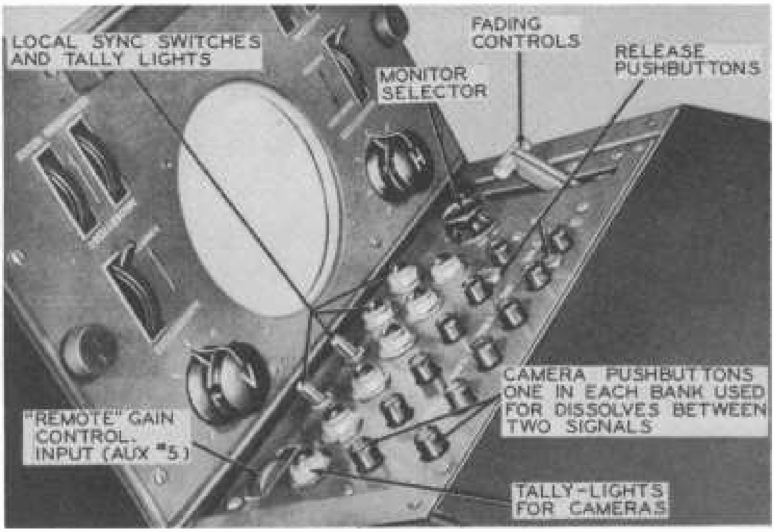
There are three general types of video switchers:

1. Mechanical push-button switching with video on the actual switch contacts. The bank of switches is interlocked to prevent more than one source being "punched up" at a time. This type of switcher may or may not employ a means of fading or lap dissolving between channels. It is used primarily (in installations of the past 10 years) in portable field units.
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 has been the most commonly employed for studio installations of all except the smallest.
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 vertical blanking time following vertical sync. This type of switching system is rapidly replacing the relay type in older stations, and is used almost universally in new stations.

4-1. FUNDAMENTALS OF SWITCHING SYSTEMS

Fig. 4-1 illustrates the RCA type TS-10A switching panel. Although an older type of unit, it is still used in many station installations across the country. Fig. 4-2 is a block diagram of this unit.

The video amplifiers are located in the lower compartment of the console. There are two two-stage amplifiers for each picture channel. Another two-stage amplifier feeds the monitor from a



Courtesy RCA

Fig. 4-1. RCA type TS-10 switching panel.

selector switch, allowing preview or on-air monitoring. Observation of the block diagram also shows the sync-interlock system which automatically adds local sync when the remote sync fails (when broadcasting from a remote point), or it may be used when local sync is otherwise required. In this system sync signals for local telecasts are usually added in the following stabilizing amplifier position.

Controls for the switching amplifier project through the inclined top panel of the desk as shown in the photo. These controls consist of two banks of push buttons that select the on-the-air signal, two toggle switches for adding local sync to incoming remote signals, gain controls for the remote signals, fading controls (also used for lap-dissolving), a three-position switch for selecting either of the two remote or on-the-air signals for display on the monitor, and tally lights which indicate the inputs being used.

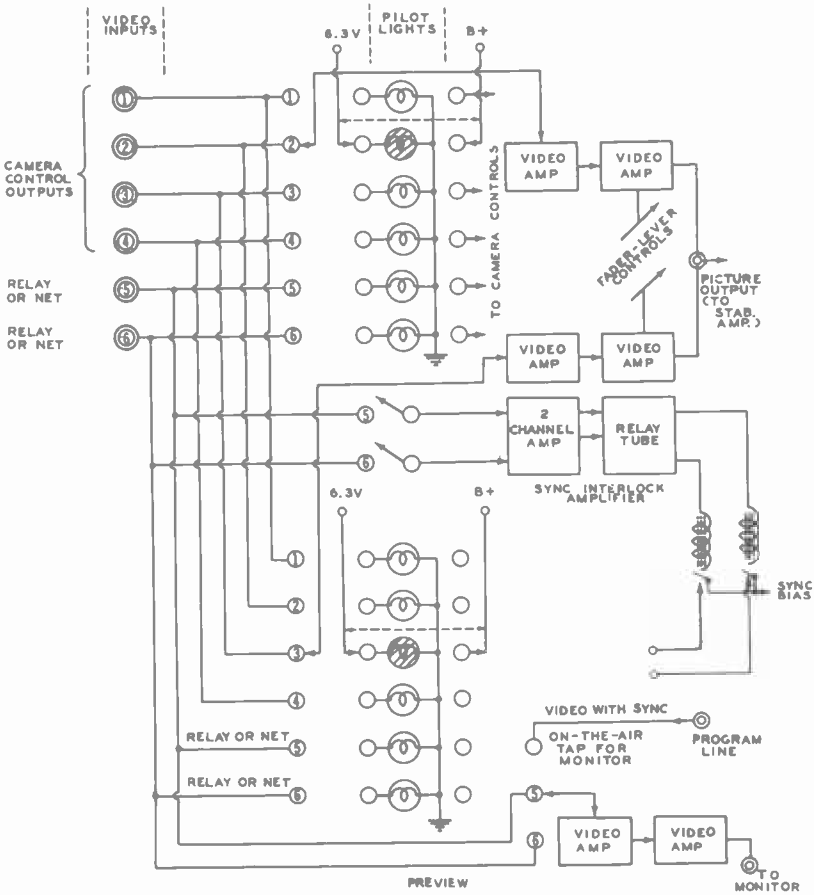
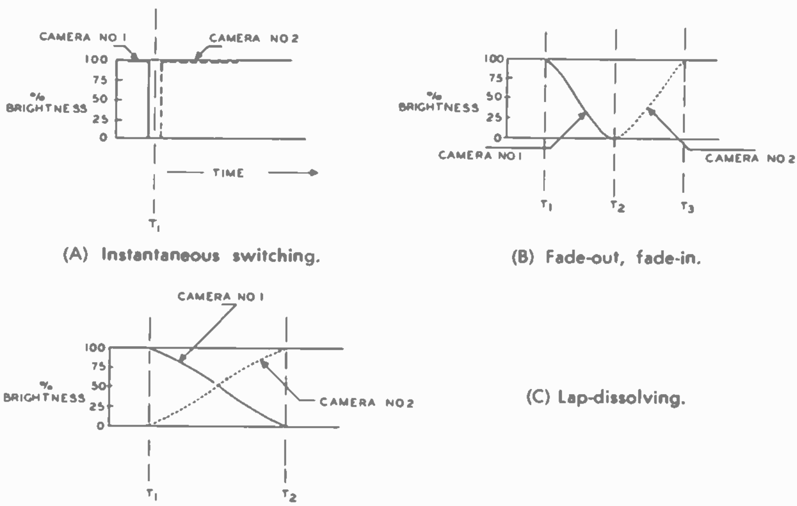


Fig. 4-2. Block diagram of the RCA TS-10 switching system.

The camera control units are connected to the switcher by means of standard 75-ohm flexible coaxial cables. The particular unit under discussion accommodates the outputs from four camera controls and two remote or network sources. The remote or network video signals are accompanied by sync from that source, whereas local camera signals fed to the mixer unit are only video and blanking without sync. The two banks of push-button switches feed separate amplifiers which have their outputs connected together to feed the line connecting the stabilizing amplifier. This amplifier is usually rack-mounted in the control room. When it is desired to switch instantaneously from one camera to another, one bank of push buttons is used. Depressing any push button releases any other push button on the same bank by an interlocking feature. This prevents two sources from being fed to a common amplifier at the same time. When it is desired to lap-dissolve or superimpose two pictures, both banks of push buttons are used. Assume, for example, that four cameras are being used; number 1 is on the air at this time. Also assume that this is being accomplished by having the camera No. 1 push button on bank No. 1 depressed. Now, by depressing camera No. 3 push button on bank No. 2, both camera output signals are fed through respective channels and combined in the output. Each picture may be adjusted in relative brightness by adjusting the respective fader control (lever-type control) shown at the right of the control panel. The fader controls can either be operated together or separately. When adjusting the relative brightness of the two different sources, as in superimpositions, the handles are operated independently. Thus either picture may be faded out entirely at any time ratio.



(A) Instantaneous switching.

(B) Fade-out, fade-in.

(C) Lap-dissolving.

Fig. 4-3. Three of the most common types of video switching techniques.

Fig. 4-3 shows the technical definition of the three common types of switching in use during telecasts. In Fig. 4-3A, camera 1 and camera 2 are switched instantaneously. In other words the camera 2 push button on the same bank of switches on which the camera 1 push button was depressed, was operated, immediately releasing camera 1 and feeding the camera 2 signal. Fig. 4-3B shows the fade-out, fade-in technique. The camera 1 push button is depressed on one of the banks, and it is adjusted at reference brightness level up to time t_1. At this instant the camera 2 push button on the other bank is depressed, but its fader control is adjusted at zero brightnes. Simultaneously, the fader for camera

1 is adjusted from reference brightness (100% on scale) toward zero. By time t_2 camera 1 is faded out and the fader for camera 2 is adjusted from zero brightness toward reference brightness. By time t_3 the camera 2 signal fully occupies the screen, and it is entirely faded in. In Fig. 4-3C, the camera 1 push button is depressed on one of the banks and its fader is adjusted at reference brightness. The camera 2 push button on the other bank is also depressed but its fader is at cutoff and only the camera 1 signal is transmitted up to time t_1. At this time the fader for camera 1 is adjusted toward cutoff, and, simultaneously the camera 2 fader is adjusted toward reference brightness. Thus, between t_1 and t_2 both signals appear on the screen while signal 1 is decreasing and signal 2 is increasing. At time t_2 the camera 2 signal fully occupies the screen and camera 1 is completely faded out.

In installations where a large number of video sources are involved, remotely controlled relays are used to switch the video signals. In this type of switching the relays may be rack mounted with all the various coaxial cables centralized and with individual relay switching panels installed at points most convenient for any particular installation. The push buttons themselves are nonlocking in this type of switching, and the interlocking action takes place in the relay system. That is, the push keys operate the relay coils, which themselves are connected in a lock-up type of circuit that drops out any relay when another relay is operated.

Fig. 4-4 shows the basic input-output bus arrangement of the two fader banks in an AB-type relay switcher. Fig. 4-5 is a view of one studio switcher panel at station WBBM (CBS) in Chicago. In this circuit each bus feeds a cathode-follower, which in turn contacts the transfer relay contacts. This is a D-type relay (make before break) spring combination which momentarily places two signals on the air to prevent picture disturbance. The "overlap" time may be 200 to 300 milliseconds.

Fig. 4-6 is a simplified block diagram of the RCA TS-20A video relay switching system. This type of relay and method of mounting is illustrated in Fig. 4-7. The contacts extending through the chassis are the inputs to the video contacts. The output or "operated" side stands above the relay frame. The basic chassis provides the circuits and relays for switching six inputs to two outputs. When additional inputs or outputs are required, auxiliary relay panels are employed. Two general types of switching are possible with relays, namely gap or overlap. The instantaneous switching method illustrated in Fig. 4-3A is known as *gap* switching, or break-before-make contacts. *Overlap* switching is make-before-break contact; thus a "black gap" between signal switching is avoided. Either type of switching is obtainable in the RCA relay system. (Overlap switching is used for on-the-air signals;

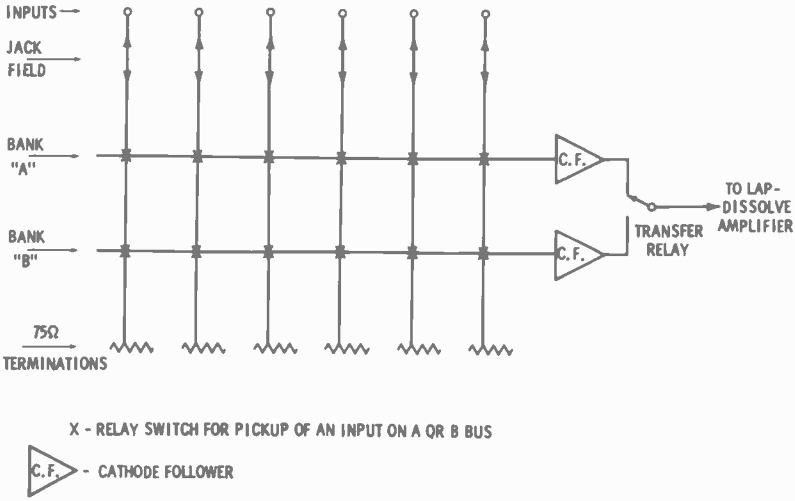
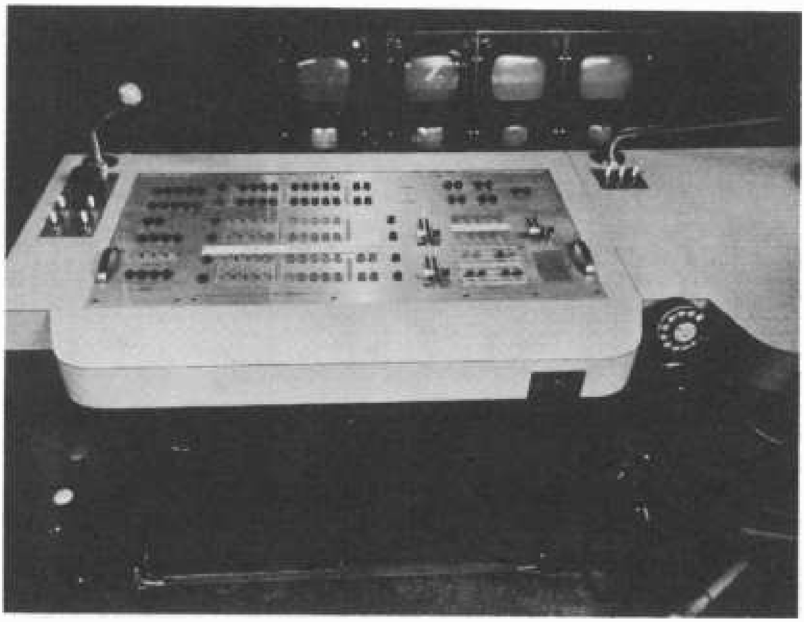


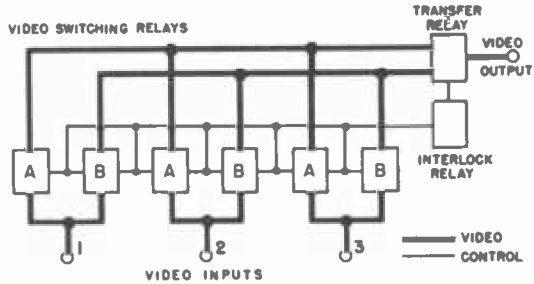
Fig. 4-4. Basic input and output system in an AB-type switcher.

preview buses may be optionally made gap switching.) Lap-dissolving simply involves the use of a lap-dissolve (mixing) amplifier in conjunction with a relay system.



Courtesy CBS

Fig. 4-5. Typical studio switching console with master monitor and three camera control monitors in background.



Courtesy RCA

Fig. 4-6. Block diagram illustrating basic switching arrangement of RCA switching system. Relays A and B are operated electrically by push buttons.

The basic action of a video relay switching system is illustrated in Fig. 4-8. Each incoming video signal is connected to *make con-*

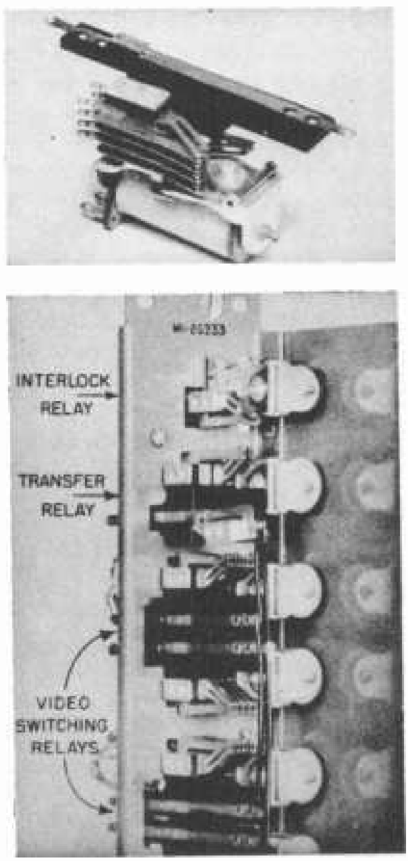


Fig. 4-7. Video switching relays.

Courtesy RCA and Broadcast News

tacts on a *pair* of switching relays. Two video bus wires which run to a transfer relay are connected to the other side of these contacts. When the associated Channel 1 push button is operated, relay A will close, connecting the incoming video signal to one of the transfer bus wires. Other contacts on this same switching relay operate the transfer relay connecting the output line to the video bus wire and hence to the video signal. When another push button (not shown) is operated, separate contacts on the transfer relay "inform" the new video switching pair that bus A is in use, thereby causing switching relay B to operate. If the description had started out at a time when A was already in use, operation of the push button shown would have caused relay B to operate instead of A. When the video contacts on relay B close, the new incoming video line is connected to the B video transfer bus wire. At this same instant other contacts on relay B close, operating an interlock relay which, in turn, releases the transfer relay. In this way the outgoing line is transferred from the original input to the new input. The circuits of relay A are now released from the "busy" hold of the transfer relay contacts and the preceding process is repeated on subsequent operation of the push buttons.

In this system the time of the signal transfer is determined entirely by the transfer relay rather than being governed by the operate and release time of the individual relays. By choice of appropriate connections to the video transfer bus wires, either gap (break-before-make) or overlap (make-before-break) switching is available.

Another 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 relay system. Or, an audio console with its associated relays may provide for operation of the video relay bays. Such systems are usually made variable so that either separate or tied-in operation may be used.

When several channels must be switched simultaneously between several outgoing lines, *preset switching* methods are used.

4-2. THE VERTICAL-INTERVAL SWITCHER

At the time of this writing, at least three major broadcast equipment manufacturers are supplying solid-state (transistor-diode combinations) switching systems tied to the local sync generator for switch timing. These manufacturers are EMI/US (General Communications Division), RCA, and Sarkes-Tarzian. Due to the superior performance and minimum maintenance requirements, it is very likely that all systems of the future, including portable switchers, will employ some form of solid-state vertical interval switch circuitry.

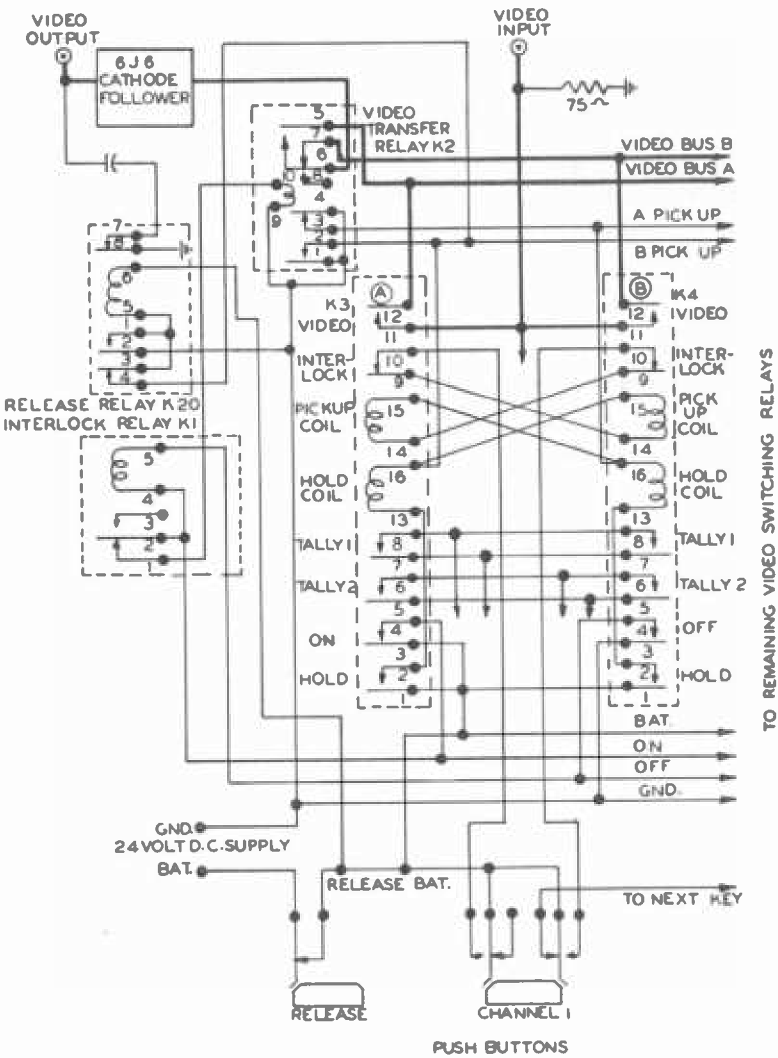


Fig. 4-8. Circuit details of RCA relay switching system.

Aside from a significant reduction in power and space required, transistor circuitry permits extremely rapid switching transitions—on the order of 1 microsecond. Hence, picture-to-picture transfer occurs between successive television fields, thereby eliminating the chance of visible disturbance to the picture. The life expectancy of transistors is virtually unlimited; they are compact, extremely reliable, and generate very little heat in operation.

By virtue of the short lead lengths involved, the compact switching assemblies made possible by transistors have excellent video performance in both monochrome and color applications.

The basic components of the RCA TS-40 Transistorized Switching System (with the exception of power supplies) are illustrated in the simplified block diagram in Fig. 4-9. These components consist of cross points, cross-point groups, latch circuit plates, coupling circuit plates, output amplifiers, mixing amplifiers, distribution amplifiers, and sync or blanking adders. Power required

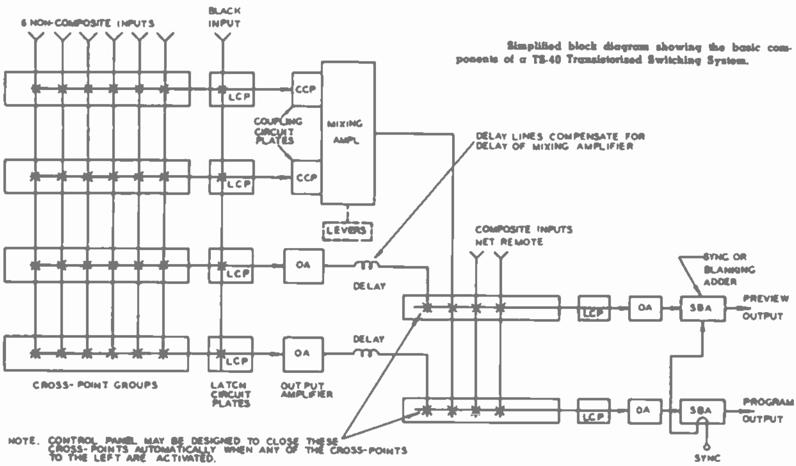


Fig. 4-9. Simplified block diagram of the RCA TS-40 transistorized switching system.

by the system is obtained from a transistor power supply, DC plate supplies, and a filament and bias supply. A typical rack layout is shown in Fig. 4-10.

Six cross points are mounted on an etched circuit board to form a cross-point group (Fig. 4-11). This assembly, fitted with a plug-in connector along one edge, has an output bus that joins all six cross points to form a six-input, single-output switching element. For convenience in planning and installing systems, the cross-point group is supplied as the basic plug-in switching module.

Each cross-point is functionally equivalent to a high-quality relay having two sets of contacts, one for a video signal and one for a tally circuit. It consists, in essence, of a semiconductor diode switch which is turned on and off by a transistorized flip-flop circuit. The circuit is bistable; that is, it will remain indefinitely in either the Off or the On position until activated externally.

The special voltages required for the transistorized circuits are all supplied by the WP-40 power supply. In addition, standard 280-

volt supplies (such as the RCA 580-D or WP-15 or 16) are required for the amplifier complement, and a 24-volt supply is required to operate tally lamps and auxiliary relays. (The coils of the tally relays operated directly by the cross points are powered by the WP-40, but the circuits controlled by their contacts require external power.)

The coaxial fittings at the rear of the WP-40 power supply are for the composite sync input and the output of the trigger pulse

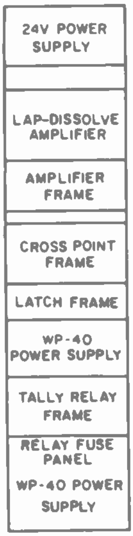
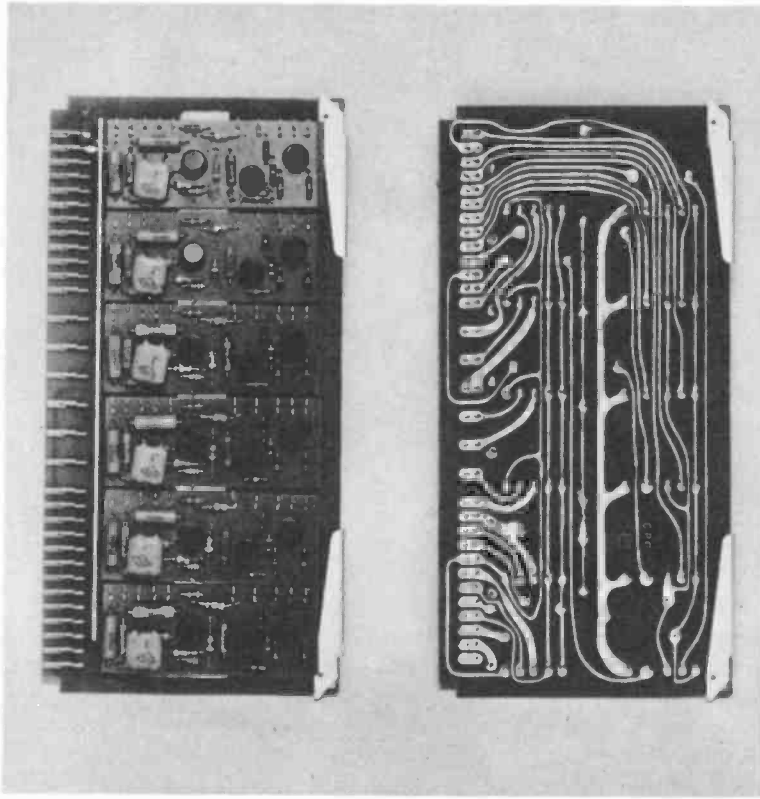


Fig. 4-10. Typical rack layout of the RCA TS-40 switching system.

generator incorporated in the supply. This generator consists of a transistorized circuit for deriving pulses suitable for triggering TS-40 cross points (through the push-button switches on the control panel). The pulses are derived during the vertical sync interval so that the switch always occurs shortly after a vertical retrace period, thus minimizing the probability of a vertical roll-over when switching between pictures of widely different duty cycles. The output pulses are at a level of about 30 volts peak-to-peak, and they are conducted by a coaxial cable from the trigger pulse generator to the trigger circuit plates mounted under the control panel. The pulse rise time is deliberately made quite long so as to limit the high-frequency energy in the pulses. This permits them to be conducted along ordinary wires from the control panel to the cross points without significant cross talk between leads.

In addition to push buttons and fader mechanisms, there are two types of etched wiring circuit plates that are mounted beneath the control panels to serve important functions in TS-40 systems. The first of these is the trigger circuit plate. This circuit is a single-input, six-output, transistorized amplifier which serves to distribute the trigger pulses generated by the WP-40 transistor power supply to as many as six rows of push buttons. (Additional trigger circuit plates may be used for panels with more than six rows of buttons.) Each time a push button is pressed, it connects the corresponding cross point to the source of pulses derived from

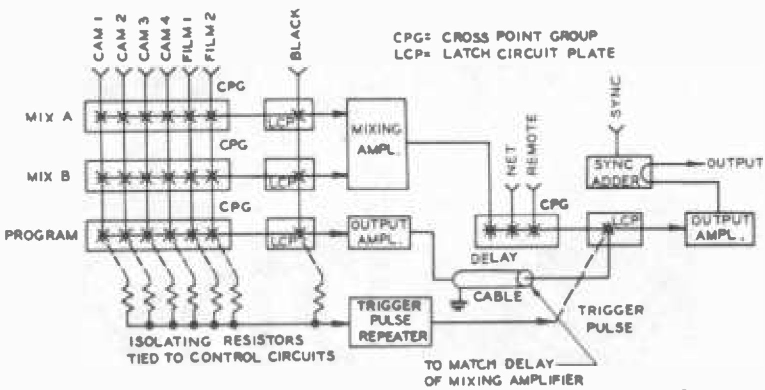


Courtesy RCA

Fig. 4-11. Front and rear views of a cross-point group which consists of six transistorized cross points mounted on an etched circuit board.

vertical sync. The very first pulse that passes through activates the cross point, and the complete switching action occurs near the end of vertical blanking.

The second special control-panel component is the trigger pulse repeater. This device may be used to make any cross point a "slave" of one or more other cross points so that the "slave" will always be activated when any one of the "masters" is in use. This feature is very useful in switching systems that employ delay compensation to keep the total time delay through the system constant no matter which signal path is punched up. Fig. 4-12 is a simplified sketch illustrating the function of the trigger pulse repeater in a system with delay compensation. The push buttons for the secondary switch (shown at the right) may be mounted in the same row as the others in the program bus; this is done so that, functionally, the operator may treat them as part of the same switching bus. No button is required for the cross point operated by the trigger pulse repeater. When any of the cross points to the left of the delay compensation line (actually a length of coaxial cable) are operated, the trigger pulse repeater produces a pulse to close the second cross point automatically. Thus, the circuit is completed through to the output.

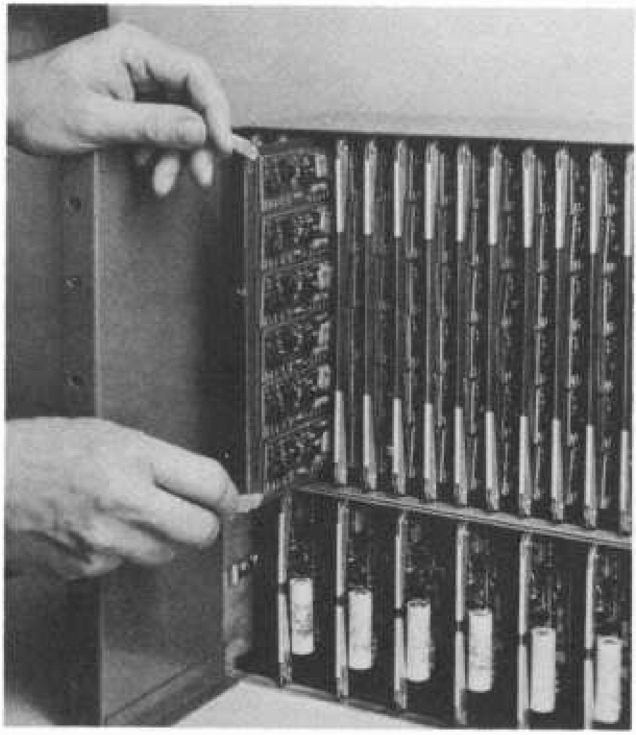


Courtesy RCA

Fig. 4-12. Simplified diagram illustrating the function of the trigger pulse repeater in a system with delay compensation.

The trigger pulse repeater is actually an amplifier followed by a clipper. The isolating resistors, required at its input to prevent cross talk between the several push-button circuits, cause a substantial reduction in the level of the trigger pulses applied to the repeater. However, its gain is sufficient to produce trigger pulses of normal amplitude at its output.

Another basic building block for the TS-40 systems is the latch-circuit plate. This circuit is mounted on an etched-circuit board which plugs into a frame normally mounted below the cross-point frame (Fig. 4-13). It performs the same function as the mechanical latching bar in a "direct" push-button switcher; that is, it trips off the circuit previously turned on each time a push button is operated, thus assuring that each output bus carries only one signal at a time. One latch-circuit plate is required for each independently latched output bus (consisting of up to four cross-point groups) and is connected to the cross points through two busses, designated *Latch Trigger* (LT) and *Latch Operate* (LO). The

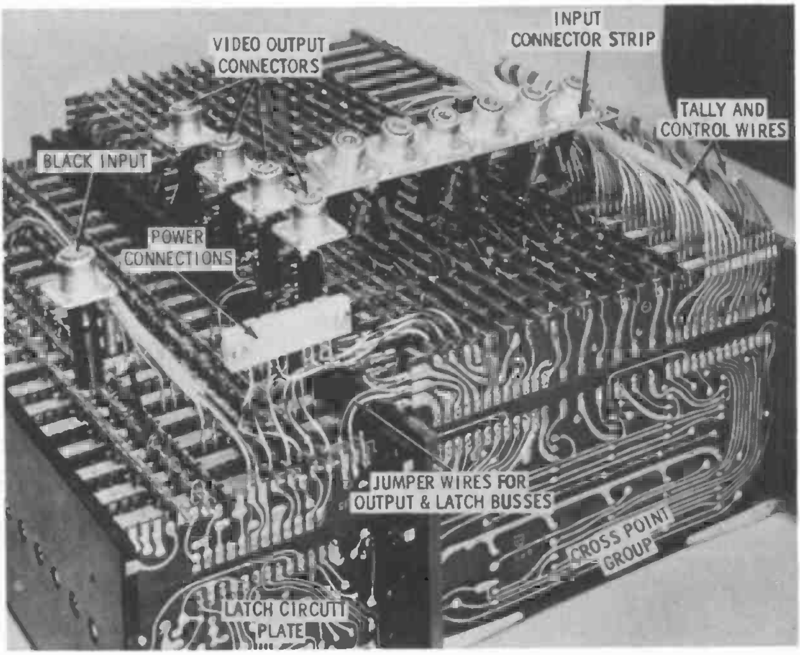


Courtesy RCA

Fig. 4-13. A cross-point frame that will accommodate up to 20 plug-in cross-point groups (120 cross points). Immediately below is the frame that accepts the latch-circuit plates.

latching operation is automatic, requiring no extra connections to the push-button control panel. Each time a cross point is actuated by its individual control button, it produces a low-level signal on the latch-trigger bus. This signal is amplified and clipped by the latch-circuit plate, and is fed back along the latch-operate bus to all of the cross points connected to the same output. The amplified latch-operate signal triggers off whichever cross point was previously on. The entire sequence of operation is extremely fast, on the order of 1 microsecond.

Fig. 4-14 shows the internal components of a cross-point frame and latch frame removed from their housings to illustrate, in greater detail, the manner in which interconnections are made. Note that both the crosspoint groups and the latch-circuit plates "mate" with narrow etched-wiring interconnection strips which serve to bring the conductors to the appropriate positions at the back of the frame. The tally and control leads are conveniently grouped at one end of the cross-point interconnection strips separate from the video bus structure. External wires are attached to the etched-wiring pieces by means of simple, slip-on edge connectors that require no solder for installation. The output, latch-trigger, and latch operate busses appear at one end of the cross-point interconnection strips where double-width slots are provided for installing the jumper wires used to connect adjacent



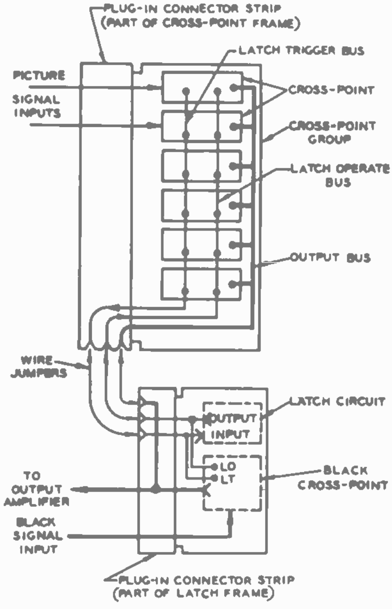
Courtesy RCA

Fig. 4-14. Cross-point and latch frames as they appear with the steel housing members removed and all internal components in their normal relative locations.

cross-point groups in parallel and to the latch-circuit plates. The video output connections are also made at the same point through coaxial cable fittings mounted on the latch frame.

The coaxial fittings used for both input and output connections are of an unusual design, combining features of standard coaxial jacks and cartridge fuse holders. The video signals are actually brought in and out through small fuses which protect the transistors and diodes from damage if excessive voltages are accidentally applied to the interconnecting cables. The fuses are of sufficiently low impedance so that they do not degrade the performance of the system.

The interconnections between cross-point groups and latch-circuit plates are shown in simplified form in Fig. 4-15. The latch-circuit plate includes a cross-point circuit identical to all the others, but intended for use by the black signal. (In monochrome systems black may consist of no signal at all, but in color systems it is desirable to provide a black signal containing the color synchronizing burst, possibly supplemented by a fixed pedestal.) The black cross-point is connected to the same busses as all the other cross-points in the same output chain, but its control circuit is interconnected with one of the power supply busses in such a way that the switcher always come up in a black condition when power



Courtesy RCA

Fig. 4-15. Interconnections between cross-point group and latch-circuit plate.

is first applied. If it were not for this feature, the bistable cross-point circuits might be activated in random fashion—some off and some on—when power is applied.

The output signal from a cross point is at a relatively low level, because each cross point handles a little less than one-tenth of the total signal current applied to the input of the switcher. This current division is necessary to permit up to ten cross points to be connected to each input. There are also minor losses involved in the cables required for delay equalization within the system. In order to restore normal signal voltage level for system distribution, a coupling circuit plate is used. The coupling circuit consists of a transistorized amplifier with an input impedance of 75 ohms and an output impedance of about 1,200 ohms.

In a TS-40 system provision is made to mount the coupling-circuit plate within the output amplifier. This amplifier utilizes etched wiring and is identical to the TA-12 distribution amplifier, except that the coupling-circuit plate has been substituted for a conventional input coupling capacitor. The TA-12 amplifiers are also useful in many other applications wherever a unity-gain, single-input, single-output isolation amplifier is required. The same basic design is satisfactory for picture signals at a one-volt level, CW subcarrier at two volts peak-to-peak, and pulse signals at four volts peak-to-peak.

Up to ten output or distribution amplifiers may be mounted in the amplifier frame. This frame also serves as a housing for other such items in the TA-12 family of equipment as the sync or blanking adder and the heater and bias supply. The sync or blanking adder may be connected in series with the output of a distribution or output amplifier whenever there is a need for adding sync pulses to noncomposite signals. An interlock relay is included in the unit, making the adder suitable for use at the output of switching systems handling both composite and noncomposite signals. This same device may be used for adding a small amount of blanking for so-called fixed set-up operation. If both sync and blanking addition are required at the same location, two sync or blanking adders may be connected in series.

Since many maintenance engineers are still relatively unfamiliar with transistorized circuits, the following description of TS-40 operation is particularly important. The following points are in effect, the basic design decisions made early in the TS-40 development program which established the framework for most of the specific circuit functions.

1. So that the cleanest possible transitions from picture to picture can be provided, the TS-40 is designed to switch on the microsecond time scale in a manner that permits control during the vertical blanking interval. This requirement is a basic factor leading to the selection of transistors and diodes in lieu of relays as the basic switching elements. (Reliability, size, and power requirements are also important factors in this basic choice.)
2. The switching elements are good enough with respect to frequency response and linearity to pass color signals with negligible distortion even in large systems where several "passes" through the switcher may be involved.
3. To minimize the total B+ power requirements, the TS-40 employs *passive* (or resistive) isolation to minimize the ef-

fects of capacity stacking on input busses and to prevent transitions on one output bus from causing detectable transients on another. (This approach eliminates the need for input amplifiers to provide *active* isolation.) Special attention has been given to the preservation of a uniform input impedance for each cross point in both the off and on conditions.

4. To eliminate the need for clamped amplifiers and to permit full freedom in the handling of composite and nonsynchronous signals, the TS-40 is designed to introduce negligible DC shift (or bounce) when switching from one signal to another. The input signals are introduced at a point where the DC voltage is zero, and the cross points are designed for uniform DC conditions at their outputs.
5. To permit the use of momentary-contact push buttons and to eliminate the need for relays or other devices in special holding circuits, the cross points themselves are made self-latching; that is, they remain indefinitely in either the off or on condition until a deliberate change is made in an external control circuit. This requirement has led to the selection of a bistable flip-flop circuit as an important part of each crosspoint.
6. As in any remotely-controlled video switcher, there is provision for interlocking the latching action of all cross-points connected to a given output, so that only one signal appears at a time. There is added provision for completing the entire latching cycle in only a microsecond or two.
7. In addition to the video output from each cross point, a second output of sufficient power to operate a tally relay is available. The tally system is a slave to the video cross point, and not vice versa, so the tally relay contacts have no influence on video switching reliability. The relay is employed only to provide low-cost isolation for the multiplicity of tally and auxiliary circuits (such as sync interlocks and audio ties) normally required.
8. When a cross point is turned off, its video isolation at 3.58 mc is on the order of 76 db, thus permitting large numbers of cross points to be combined in complex systems with reasonable cross-talk performance.

Brief Review of Overall System

The block diagram of a simple TS-40 system shown in Fig. 4-16 should serve to illustrate the functional relationships of the various circuits to be described. The actual switching elements are known as cross points, and they are mounted in groups of six, as shown near the center of the diagram. Video input signals are

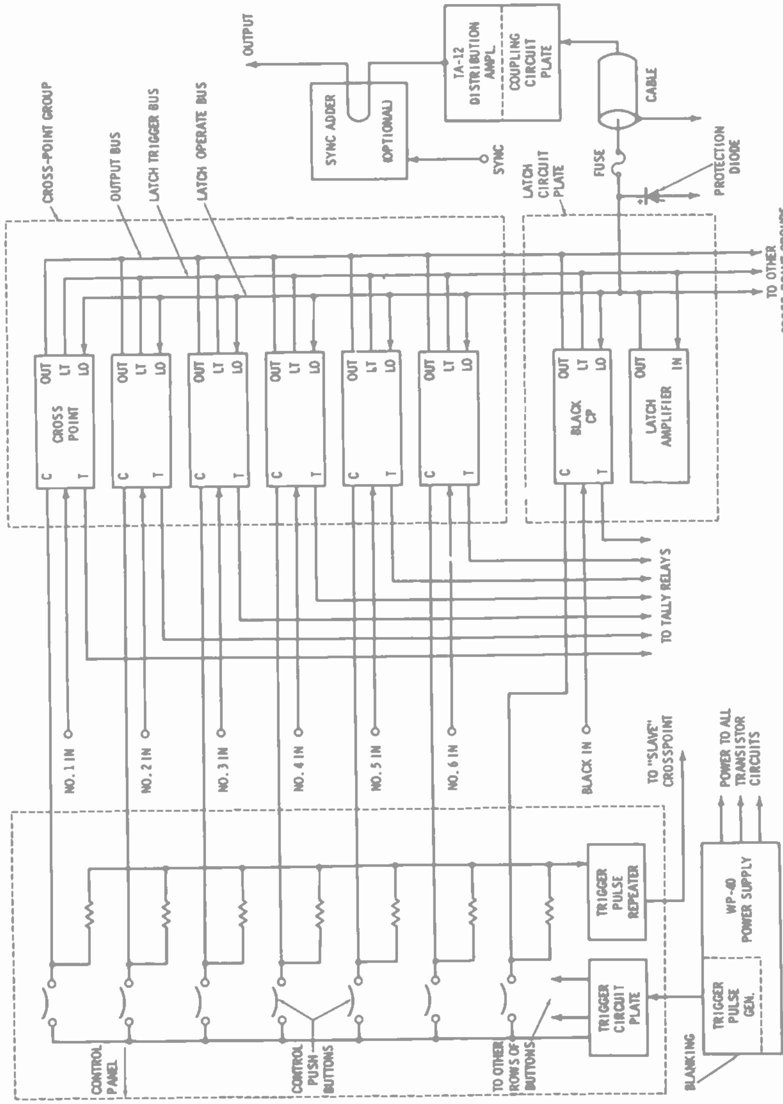


Fig. 4-16. Block diagram of a simple TS-40 system.

brought in through small fuses and horizontal copper buses which join all the cross points to be connected to a given input. When the number of cross points connected to a bus is less than eleven, impedance trimming resistors are employed to adjust the input impedance to 75 ohms. The cross points are turned on by pulses generated by a special circuit in the WP-40 power supply and distributed by way of the push-button switches in the control panel. A trigger-circuit plate mounted beneath the control panel provides a separate output for each independent row of push buttons. Where a "slave" cross point is required to close a secondary switch in systems where delay compensation is employed, an appropriate trigger pulse is developed by using a trigger pulse repeater whose input is tied to a group of push-button circuits through isolating resistors. Each cross point has an independent tally output, which is normally used to operate a multiple-contact relay to control the several tally lamps and auxiliary circuits associated with each cross point.

All cross points associated with a given output are joined by three buses, designated Output, Latch Trigger, and Latch Operate. Up to four cross-point groups may be joined by these buses. The buses are also connected to a *latch-circuit plate,* which contains an additional cross point for the black video signal plus a latch amplifier. The function of the video output bus is obvious (note that the connection to the output amplifier is brought out through a fuse and a short length of coaxial cable). A protection diode serves to carry the fuse-blowing current in one direction in the event that an excessive voltage is applied to the cable. The cross-point circuit itself will safely carry the fuse-blowing current in the other direction. The output amplifier is equivalent to a TA-12 distribution amplifier with a transistorized *coupling-circuit plate* installed at its input. A TA-12 sync adder may be placed in series with the output amplifier if sync addition is required. The latch-trigger bus conveys a relatively low-level input signal to the latch amplifier; each time a cross point is turned on, it signals this fact by placing a small current on this latch-trigger bus. The latch amplifier then generates from the low-level trigger signal an output pulse of sufficient amplitude to serve as an off trigger for whatever cross point was previously on. The latch-operate pulse is conducted to all cross points simultaneously, but its action is automatically overridden in the case of the cross point to which the control signal is being applied.

Basic Switching Circuit

The basic switching action is illustrated by the simplified schematic diagram in Fig. 4-17. The actual video switching element is a crystal diode, which is connected in series with an 1,100-ohm isolation resistor. The anode side of this diode is held at about +3 volts (through the transistor in the following coupling circuit) so that it is kept in the conductive state except when its cathode is connected to a still higher positive voltage (about 5 volts) through a transistorized flip-flop. Very good video isolation (of the order of 76 db) is provided when the diode is cut off, because the video signal is simultaneously shorted to AC ground at the output end of the isolation resistor through the flip-flop and the

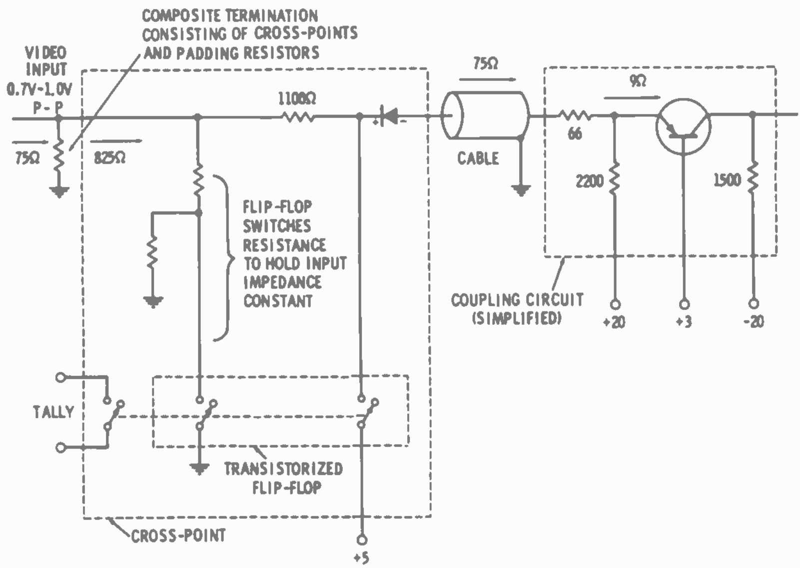


Fig. 4-17. Basic circuit of the RCA TS-40 switcher.

+5-volt power supply. In addition to the 1,100-ohm series resistor, a shunt impedance is provided to bring the net impedance of the cross-point circuit down to 825 ohms, and this shunt impedance is automatically adjusted by the flip-flop to maintain the input impedance constant for both the off and on conditions. Up to eleven cross points may be connected across an input bus to form a normal 75-ohm termination for a coaxial cable; if less than eleven output buses are required, padding resistors may be employed to adjust the input impedance to the proper value. A switching transistor capable of delivering enough power to operate a tally relay is also controlled by the same flip-flop which controls the video switching action.

Because the signal current injected into the input of the switcher is divided into as many as eleven different branches, each cross point handles only a fraction of the normal signal current. If this signal current were passed through an ordinary

75-ohm load resistor at the end of the 75-ohm cable needed to link the switcher output with the rest of a studio system, the voltage developed would be of the order of 60 millivolts. To provide the necessary voltage gain to restore the normal 0.7 to 1.0-volt signal level, a transistorized circuit known as a "coupling circuit plate" is employed. The key feature of this circuit, shown in considerably simplified form in Fig. 4-17, is a common-base amplifier. Almost all of the signal current injected into the emitter of this amplifier passes through to the collector, where it can be applied to a load resistor of sufficient value (approximately 1,500 ohms) to produce the normal voltage level. A unity-gain, tube-type amplifier may then be employed as an impedance transformer to develop the same voltage swing across a 75-ohm load. Since the inherent input impedance of the common-base amplifier is only about 9 ohms, a 66-ohm resistor is placed in series with its input to form a proper termination for the 75-ohm line linking the actual output bus to the output amplifier. A 2,200-ohm resistor provides bias current for both the common base amplifier and the switching diode.

Transistorized Flip-Flop Circuits

Because a transistorized flip-flop circuit plays a vital role in every cross point, it is desirable to comment briefly on the general properties of such circuits before describing in detail the specific circuit used.

From the circuit diagram of a very simple transistorized flip-flop shown in Fig. 4-18, the reader can see that a transistorized flip-flop is closely related to a tube-type, flip-flop, or bistable multivibrator. The flip-flop consists of a pair of transistors (X1 and X2) in nominally identical circuits with the collector of one direct-coupled to the base of the other, and vice-versa. When either side conducts, it holds the opposite side at cutoff by applying a reverse bias to its emitter-base junction. The conduction state can be reversed by applying an appropriate trigger signal, usually at one of the base connections.

With the voltage and current values shown in Fig. 4-18, transistor X2 will conduct. Transistor X2 operates in a state of saturation, so its impedance from collector to emitter is very low, and all three leads are at essentially ground potential. This means that there must be about 10 milliamperes following in its collector load resistor, and about 1 milliampere in its base bias resistor. The voltage at the base of transistor X1 is determined by noting that it is connected to a voltage divider consisting of a 20K bias resistor and the 5K cross-coupling resistor. Since the total voltage drop from the bias supply to the grounded collector of transistor X2 is 20 volts, the current flowing in this voltage divider is 0.8 ma, and the voltage at the base of X1 is +4 volts. Because this is positive with respect to the grounded emitter, the transistor is kept cut off, and it appears like a very high impedance (of the order of 100K) from collector to ground. The only significant source of current for the X1 collector load resistor is, therefore, the 5K cross-coupling resistor. Since this current path has a total resistance of 7K between the −20-volt collector supply battery and the grounded base of transistor X2, the current flow is about 3 ma. The voltage at the collector of X1 (−14V) is determined by subtracting the drop across the load resistor from the supply voltage.

The conduction state of the flip-flop circuit may be reversed by applying, to the point marked control, a *negative* voltage of

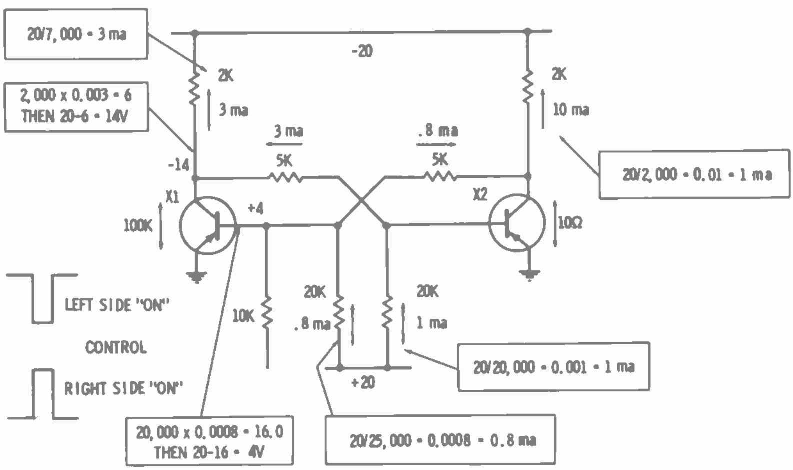


Fig. 4-18. Basic transistorized flip-flop circuit.

sufficient level to overcome the +4 volts bias on the base of X1. Once transistor X1 begins to conduct, its collector voltage begins to *rise* toward ground potential. This rising voltage is coupled through a 5K resistor to the base of transistor X2, where it begins to cut off the current flow. As the current in transistor X2 decreases, its collector voltage *falls* (in the direction of the collector supply voltage), and this falling voltage is cross-coupled to the base of transistor X1, where it reinforces the action of the trigger signal which initiated all the changes. This positive feedback causes the current and voltage changes to continue very rapidly until transistor X1 is fully saturated and transistor X2 is cut off, at which time the changes will cease and the circuit will remain indefinitely in this state. The original conditions could be restored either by applying a *negative* trigger signal to the base of transistor X2 or applying a *positive* trigger signal to the base of transistor X1.

TS-40 Cross-Point Circuit

The complete schematic of the cross-point circuit is shown in Fig. 4-19. The cross point utilizes the basic principles illustrated by previous figures, but it also includes a number of refinements which will now be described.

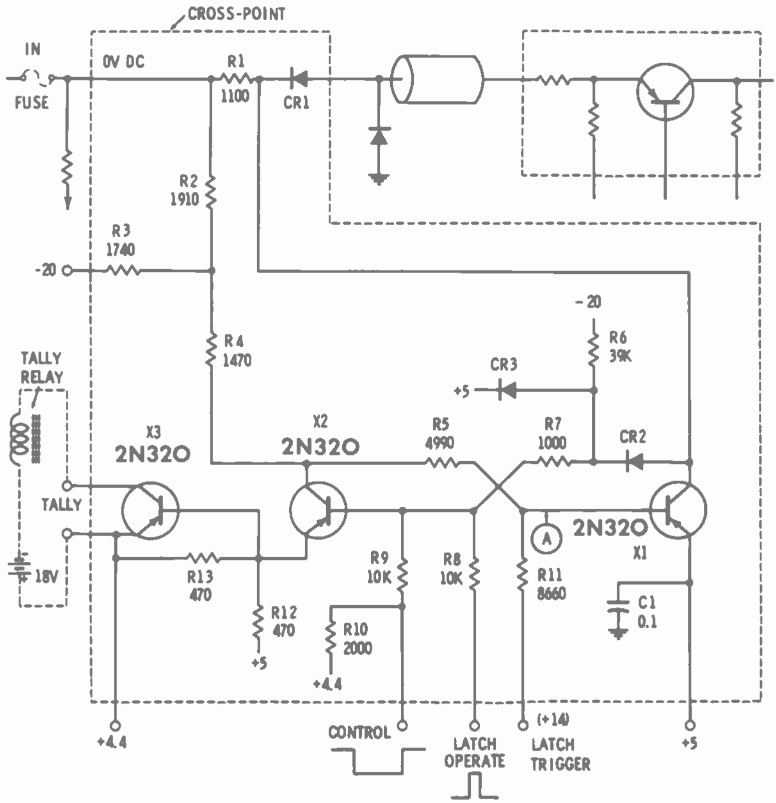


Fig. 4-19. Complete cross-point circuit.

In the complete circuit, transistors X1 and X2 form a flip-flop basically similar to that shown in Fig. 4-18, while X3 is a tally switching transistor connected in series with one side of the flip-flop. The 1,100-ohm isolation resistor in series with the actual video switching diode (CR1) forms part of the collector load for X1. The operation of the circuit can best be explained by describing the specific reason for each detail of the circuit that represents a departure from the very simple circuit of Fig. 4-18.

Note, first of all, that the emitters of the flip-flop are not connected directly to ground, but instead are connected to slightly positive supply voltages. The reason for this is that transistor X1 is utilized as the switching element which applies a source of +5 volts to the cathode of switching diode CR1 when the cross point is turned off. The diode itself is returned to a source of +3 volts rather than ground so that the input signal, at the opposite end of the isolation resistor, can be introduced at a point where DC potential is zero (the junction of R1 and R2). Instead of being returned directly to the source of +5 volts, the emitter of X2 is returned to a slightly lower voltage (4.4 volts) through the base-to-emitter junction of a third transistor (X3). When X2 conducts, indicating that the cross point is on, X3 also conducts and closes a circuit with enough current-carrying ability to operate a 1-watt tally relay placed in series with its collector power supply. The major function of the +4.4-volt supply is to provide the bias to keep both X2 and X3 fully cut off when X1 is in the conducting state. When X1 is conducting, the emitter of X2 and the base of X3 are both held at about 4.7 volts by means of the voltage divider consisting of R12 and R13, and the base of X2 is nominally at +5 potential, since it is connected to the collector of X1 through R7 and the forward-biased diode CR2. Thus, both X2 and X3 have reverse biases of about 0.3 volts on their emitter-to-base junctions.

The collector load resistors are not returned independently to the −20-volt collector supply. but they are connected through a common 1,740-ohm resistor (R3). This arrangement provides automatic adjustment of the DC conditions so that the junction of R1 and R2 remains at ground potential (DC-wise) in both the off and on conditions. Resistors R2 through R5 are all of 1% tolerance, because they are involved in the impedance-balancing arrangement which holds the AC input impedance constant for both the on and off conditions. The equivalent AC input impedances for both the conditions are shown in Fig. 4-20. When the cross point is on, isolation resistor R1 is in series with the switching diode (which has an impedance of about 25 ohms in its forward-biasing condition) and the 75-ohm input impedance of the coupling circuit plate. R2 leads to another circuit branch attached to the input, and R3 is, in effect, a shunt to AC ground. R4, a collector load resistor, is effectively in series with the very low equivalent impedance of X2 (in a saturated condition) and X3 (with a strong forward bias). When the cross point is off, the 1,100-ohm isolation resistor is in series only with the 8-ohm equivalent impedance of X1 (in a saturated condition), and the only significant path between R4 and ground is through R5, the cross-coupling resistor which ties the collector of X2 to the base of X1. Simplification of the equivalent resistances shown in Fig. 4-20 by the step-by-step combination of the several series and parallel components would show that both networks are equivalent to 825 ohms within a fraction of 1%.

Diode CR2 (in Fig. 4-19) is used as part of the coupling link between the collector of X1 and the base of X2 primarily in order to disconnect this cross-coupling network from the video path when the cross point is on, thus reducing the number of components affecting the video transmission. When the cross point is on, X2 is held in a conducting state by the forward-bias current drawn through the emitters of both X3 and X2 and the bias network consisting of R7 in series with R6. Ignoring the small voltage

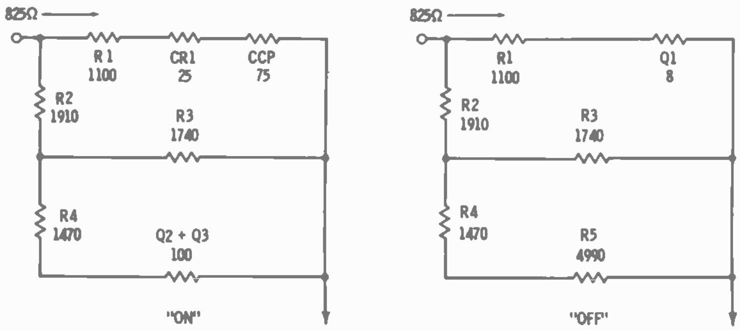


Fig. 4-20. Input equivalent circuits for a cross point.

drops across the emitter junctions, we find that the current through the bias network is approximately equal to the total supply voltage (24.4 volts) divided by the total resistance (40K) or about 0.6 ma. This current flow holds the cathode of CR2 at about +3.8 volts, which is more positive than its anode; hence, the diode is nonconducting. When the cross point is "off," however, the anode of CR2 is connected to +5 through saturated transistor X1, and it conducts freely through R6. Its cathode, therefore, remains close to +5 volts and provides (through R7) the positive bias needed to keep X2 cut off. The only function of R7 in this network is to provide a reasonable impedance across which the triggering signals needed to reverse the state of the flip-flop are built up.

Two types of trigger signals are applied to the base of X2, as shown in Fig. 4-19. The on trigger signals are relatively wide, negative-going 60-cycle pulses applied through a push-button control switch. Off trigger signals are much narrower pulses of positive polarity generated automatically by the latching circuit connected to every output bus. R8 and R9 are isolation resistors which make the impedances of the latch or control circuits negligible as far as the operation of the cross point itself is concerned. R10 is a shunting resistor which lowers the input impedance of the control circuit down to about 1,670 ohms to minimize the susceptibility of the control lead to impulse noise or cross talk. The control signal pulses are generated in the WP-40 power supply and are made much wider than the latch pulses so that they will always override the latch pulses when both occur simultaneously.

The bias required to hold X1 at cutoff when the cross point is on is obtained from a +14-volt supply source connected through R11. The current through R11 is about 0.3 ma less for the on condition than the off condition, and this current change serves as the latch trigger signal which initiates the action of the latch-circuit plate.

The only capacitor in the entire cross-point circuit is a bypass for the emitter of X1. In a normal switching system, only one cross point in each output bus is on at any one time, and a large number remain in the off condition with their video signals shorted to the +5-volt supply through their X1 transistors. Thorough bypassing of the +5-volt supply (to avoid cross talk) is assured by attaching a separate bypass capacitor to each cross point and carrying an individual ground lead for each capacitor back to a common, low-impedance ground.

Diode CR3 plays no part in the normal operation of the cross-point circuit, but it serves as a protection device to prevent damage to the transistors or other components in the event that excessive voltages or currents are applied to the video input or output lines. The input and output cables are connected through fuses with current ratings approximately 10 times greater than the normal signal currents. If for some reason the collector of X1 tends to go excessively positive, the conduction path through CR2, CR3, and the +5-volt power supply will safely carry enough current to blow the fuse before other damage occurs. If the accidental occurrence tends to make the same point go excessively negative, a safe conduction path for the fuse-blowing current is provided by CR1 and a protection diode associated with the output fuse-holder (located on the latch frame).

Latch Circuit

The latch-circuit plate, one of the basic building blocks for TS-40 switching systems, contains a cross-point circuit identical to the one previously discussed as well as the latch circuit shown in Fig. 4-21. The cross point on the latch-circuit plate is the recommended point for introducing a black signal to make sure that it comes up in the on condition when power is first applied to the system or when power is momentarily interrupted.

The terminal marked LT (for latch trigger) not only serves as an input terminal for the latch circuit, but also serves as a source of +14 volts for all cross-points connected to the latch-circuit plate. Transistor X1 is an emitter follower, which serves as a voltage regulator to maintain the +14-volt source constant, independent of the number of cross points that may be attached to the latch-circuit plate. This regulator also serves to isolate the latching action from the +20-volt bus so that many latch circuits may be employed in a single system without interaction between them. Resistors R3 and R4 establish the operating bias for X1, and R2 drops the effective collector supply voltage to minimize the collector dissipation.

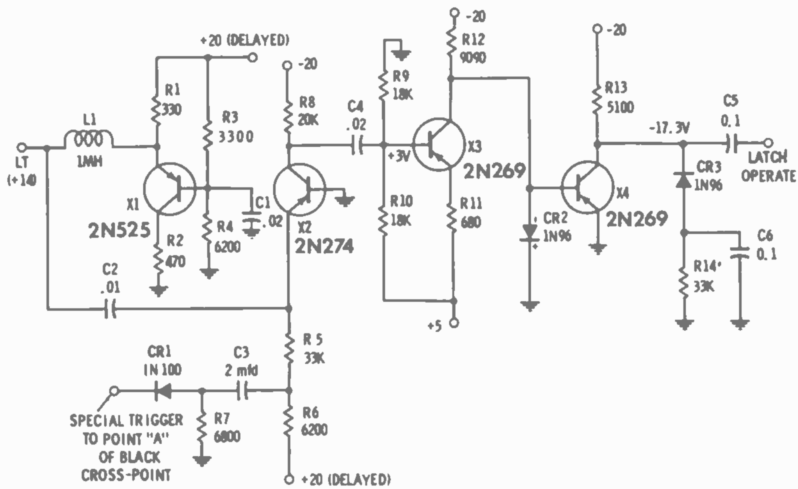


Fig. 4-21. Latch circuit.

The latching action is initiated by a slight decrease in current in the latch-trigger lead of an individual cross point which results when the cross point is triggered on. The current change is quite rapid (of the order of a fraction of a microsecond) and develops a small positive-going voltage step across L1; this voltage step is then coupled by C2 to the input of X2, connected as a straightforward common-base amplifier. Resistors R5 and R6 provide an operating bias of approximately 0.51 ma (20 volts divided by 39.2K); most of this bias current flows through the load resistor, R8, giving a steady-state voltage of about −10 volts on the collector. The input impedance of this amplifier is very low (about 20 ohms), so its performance is quite independent of the number of cross points connected to its input. Each cross point has a source impedance on the order of 8,600 ohms; 25 of them in parallel (the system maximum) offer a source of about 350 ohms, which is still large in comparison with 20 ohms. The gain of this stage is sufficient to produce a voltage swing of about 2 volts (positive-going) when the small current step is applied to its input.

The +20-volt source which supplies bias to both X1 and X2 is delayed slightly by a special circuit in the WP-40 power supply so that it always appears last when the power supply is turned on. This delay is employed to assure that all cross points except one will remain off when beginning operations. Until the delayed +20 volts appears, the +14-volt source applied through the latch trigger bus is also missing, and all cross points must be off, because the video shorting transistor (X1 in Fig. 4-19) can be cut off only when the +14-volt supply is present. When the delayed +20 volts appears, it not only activates transistors X1 and X2 of the latch circuit, but it also produces a special trigger signal across R7. This trigger signal is actually formed by the charging surge of capacitor C3, and its peak amplitude is nominally 10 volts, because R6 and R7 form an effective voltage divider during the charging period. The special trigger signal (positive-going) is coupled by diode CR1 to the black cross point at the point marked A on Fig. 4-19 (the base of X1). This special trigger turns the black cross point on, while all others remain off.

Diode CR2 plays an important role in the operation of the last two stages of the latch circuit. The stage using X3 is so heavily biased by the combination of R9 and R10 that it will operate in a saturated condition (i.e., with the collector at virtually the same potential as the emitter) were it not for the presence of CR2, which provides an alternate path for the collector current and holds the collector at nominally ground potential. Actually, the anode of CR2 is positive with respect to ground by a few tenths of a volt, which is enough reverse bias to keep X4 normally cut off. When the 2-volt rise appears across R8, signalling the beginning of a latching cycle, stage X3 is driven to cutoff. The emitter resistor, R11, not only stabilizes the stage by degeneration, but it also adjusts the stage gain so that nearly all of the input signal is required to achieve cutoff, thus assuring good noise immunity. The voltage at the base of X4 does not change immediately when the current through X3 begins to decrease, because diode CR2 holds this point at nearly ground potential until the current flow through X3 is reduced enough to cut off the diode. When X3 finally reaches complete cutoff, X4 is driven into full saturation, by virtue of the direct connection of the emitter-to-base junction to the −20-volt supply through a relatively small resistor (R12). In its saturated state the collector of X4 rises to nominally ground potential. The voltage change resulting from the latch signal is, therefore, equal to the steady-state voltage appearing on the collector in the absence of a signal. This voltage is about −17.3 volts, established by the voltage-dividing effect of R13 and R14. Diode CR3 opens when the latch signal appears (because capacitor C6 holds the voltage on its anode essentially constant during the brief latching period). The chief function of CR3 is to permit the connection of a bypass capacitor (C6) to the latch-operate bus during the normal operating periods when no switching is taking place, so as to reduce any possibility of cross talk on this bus. The clipping action of X4 assures that the latch-operate signal will always have the same amplitude, regardless of the number of cross points connected.

All operations associated with the latching function are actually initiated by *step* changes in voltage or current, but if an oscilloscope is used to observe the waveforms at any of the significant signal points, it would be noted that the latch signal *appears* to be a pulse. The reason for this is that when the cross point previously on finally goes off as a result of the latch-operate step signal, its latch-trigger lead puts out a signal of opposite polarity from that which initiated the latching signal. This opposite polarity signal passes completely through the latching chain, effectively shutting off the latching circuits in preparation for the next switching operation. The entire latching cycle requires from about 1 to 3 microseconds, depending primarily upon the number of cross-points connected to the latching buses.

Coupling-Circuit Plate

The major function of the coupling circuit shown in Fig. 4-22 is to recover the voltage lost as a result of the 1,100-ohm isolation resistors in the cross points. The circuit consists of a common-base amplifier (X1) in series with an emitter follower (X2).

The input of the coupling circuit plate is designed to serve as a proper terminating impedance for the short length of coaxial cable required to conduct the signal from the switcher proper to the output amplifier. The nominal input impedance of X1 is about 8.5 ohms, so a 66.5-ohm resistor (R2) is placed in series to adjust the mid-band impedance to the required 75 ohms. R1 and C1 cause the terminating impedance to rise at low frequencies (below a few hundred kilocycles) to compensate for the nonuniform frequency response of the video switching diode at the sending end of the cable. R3 and C2 trim the impedance downward slightly in the high-frequency region (above about 4 megacycles) to compensate for an opposite variation in the inherent input impedance of the transistor circuit.

R5 is a bias resistor that supplies bias current both to X1 and to the video switching diode in whichever cross point is in the on condition. The DC voltage drop across R5 is a little less than 17 volts, so the current through it is about 7.5 milliamperes. The video switching diode takes 2.5 milliamperes of this current, while the remaining 5 milliamperes flow in X1. R4 and C3 serve as a decoupling filter to prevent video cross talk on the +20-volt bus.

The base of X3 is connected to the +3-volt power supply through a second decoupling filter, consisting of R6 and C4.

The load impedance of X1, which was represented by a simple 1,500-ohm resistor in Fig. 4-17, actually consists of resistors R8, R10, and R11, supplemented by a diode clipper and several components for trimming the frequency response. Resistors R8, R10, and R11 are effectively in parallel as far as the signal is concerned, yielding an equivalent impedance of 1,540 ohms. R10 and R11 also form a voltage divider which effectively lowers the DC voltage

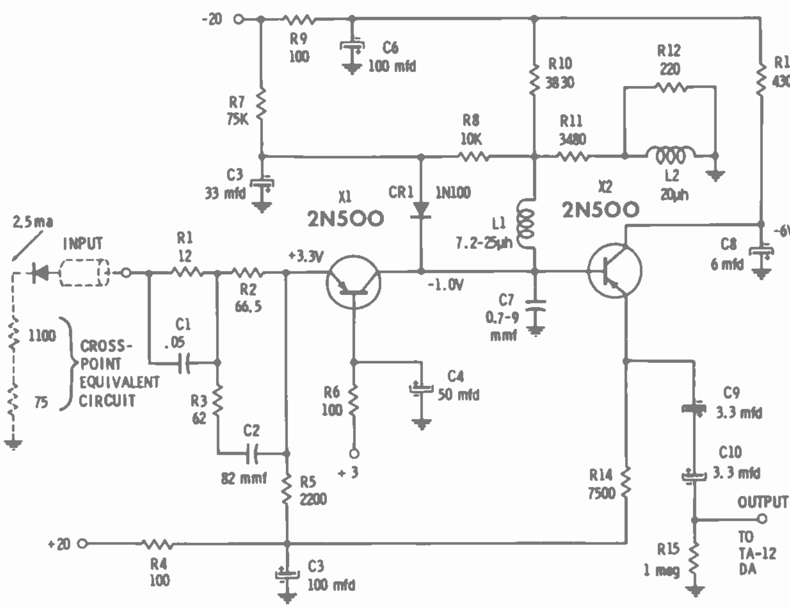


Fig. 4-22. Coupling circuit.

on the collector to about −1.0 volt relative to ground (or −4 volts relative to the emitter), thus reducing the collector dissipation. R10 and R11 also form the bias network for the base of X2. L1 serves as a conventional shunt peaking coil for trimming the frequency response. Variable capacitor C7 is employed to trim the value of capacitance to be balanced out by the shunt peaking coil, and compensates for the capacitance tolerances in the other components. L2 and R12 form a fixed peaking network to compensate for a frequency-response variation in X1 which would, if uncorrected, cause a slight dip in the response in the vicinity of 1 megacycle.

Diode CR1 is a clipper that limits the maximum excursion of the narrow, negative-going pulse which marks each switching interval. The TS-40 operates on an overlap basis, that is, a cross point comes on a microsecond or so before the previous circuit is latched off. During this very brief overlap period, bias resistor R5 at the input to X1 in the coupling circuit plate must momentarily supply bias current for *two* cross points. This leaves only 2.5 milliamperes flowing in X1, resulting in the generation of a negative-going pulse of the same width as the latching interval. Diode CR1 is biased by R7 and R8 in such a way that it is nonconducting for all normal signal excursions, but it conducts during the switching transitions and clips the pulses to a level only slightly more negative than the sync tips of a maximum-white signal. The residual pulse has insufficient energy content to cause difficulty in later portions of the system. Capacitor C5 serves primarily as a bypass for the back side of the clipper, but it also forms a low-boost network in conjunction with R7 to compensate for the low-frequency roll-off of the output coupling network. This low-boost action is also augmented slightly by R9 and C6, although the major function of these components is the decoupling of the collector supply leads.

The emitter-follower circuit employing transistor X2 does not significantly affect the gain of the coupling circuit plate, but it effectively isolates the input capacitance of the tube-type output amplifier from the load impedance of X1, and it also provides a lower output impedance for the coupling circuit plate. As noted earlier, resistors R10 and R11 form the DC bias network for the base of X2. Resistor R14 in the emitter circuit is also involved in establishing a reasonable emitter-current bias of about 3 milliamperes. The actual output impedance of the circuit is equal to this resistance shunted by the emitter impedance of X2, which is much lower (of the order of about 100 ohms). Resistor R13 reduces the voltage on the collector to minimize dissipation, and capacitor C8 bypasses R13 to prevent any effect on the frequency response. The output capacitors are of a polarized type, and it is necessary to connect two of them back-to-back to form a non-polarized capacitor because the polarizing voltage can actually be reversed if the circuit is operated with nothing connected to its input.

As mentioned previously, delay equalization is used in various paths of the TS-40 switching system. In order that the consequent frequency response through extra cable lengths be the same as the short-run cables, video equalization circuits are employed, as shown by Fig. 4-23 for the cable lengths indicated.

4-3. SWITCHER MAINTENANCE

The only difference between maintenance of the switcher system and other units is in the type of control switches. This 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 switching systems is minute amounts of dirt or foreign matter on the contacting surfaces. In the case of field equipment, mountings and connections

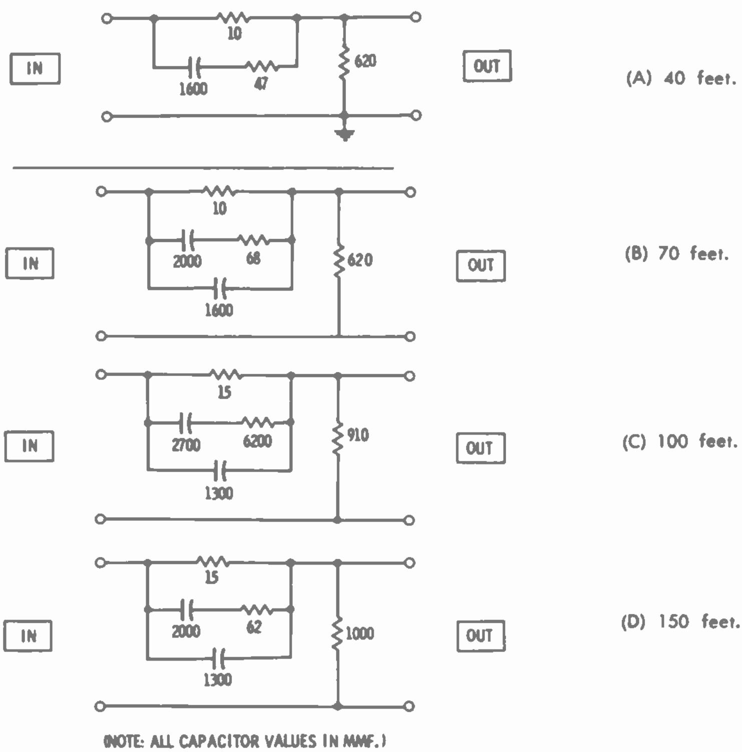


Fig. 4-23. Typical video equalizer circuits for various lengths of cable.

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 in inspecting and servicing certain types of switches.

Relays are normally enclosed in a dust-tight cover over the panel mountings, and should never be serviced except when trouble is indicated. When necessary, relays should be inspected for correct spacing and proper line-up 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 a sufficient tension. At longer intervals, crocus cloth dipped in carbon tet should be used for this process and then followed by a dry linen cloth.

A trouble commonly encountered in relay-type switchers employing 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 can only be remedied in stubborn cases by careful selection of new tubes. When this DC bounce is coupled to a following distribution amplifier employing DC-coupled feedback circuits, severe fading for two or three cycles may occur until the feedback is stabilized.

A good way to check for tube selection in such instances is to employ an adjustable APL (average picture level) stairstep generator (as described in Section 6) or use two slides from a film chain with widely different densities. The effectiveness of each selected tube can be critically evaluated by means of an oscilloscope connected to the distribution amplifier output by noting the deflection in centimeters as the duty cycle is varied or the two slides are switched alternately. A video monitor in conjunction with the scope allows correlation between 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.

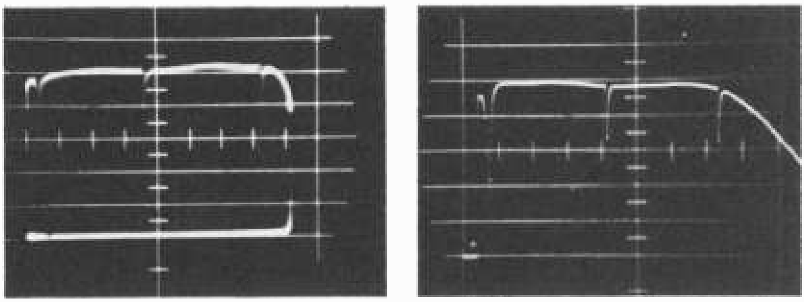
Switching systems should be checked for frequency response, amplitude, and phase linearity at least every 30 days for optimum results. The best tube check that can be run on modern systems is the one mentioned previously. System checks are described further in Sections 5 and 6.

With the video sweep generator connected to a given input of the switcher (such as No. 1 input) be sure to check all paths of the signal:

Fader A bank.
Fader B bank.
Fader A bank through Engineering Preview.
Fader B Bank through Engineering Preview.
Engineering Preview Normal.
Engineering Preview on air.
All Directors Preview paths.

Obviously, the extent of such procedures depends on the switching system facilities, which range from very simple to extremely complex layouts.

Fig. 4-24 is the video response curve of the TS-40 switcher installed at station WTAE in Pittsburgh. Markers are at 4 and 8 mc. Fig. 4-24A is the appearance with internal 60-cycle sweep used on the scope. Many technicians prefer this type of sweep because of the well defined base line produced. In Fig. 4-24B is the same response as observed with a normal 60-cycle time base. This type of observation is often preferred because it allows you to see the actual shape of the roll-off beyond the systems upper response.



(A) With internal 60-cycle sweep used on scope.

(B) With normal 60-cycle sweep time base on scope.

Fig. 4-24. Video response curves of the TS-40 switcher.

The shape of the roll-off is quite important for good transient response.

It is very important to properly maintain levels through the various switcher paths to avoid overloading any one amplifier in the path. Use the following general procedure.

1. Feed a calibration pulse, or any fixed picture slide to a given switcher input. (Do not use a single-frequency sine wave, since the level then depends on the frequency response which may vary slightly at a fixed frequency).
2. Check each amplifier for each destination (air, preview, etc.) and adjust the level for the level which should exist. (Normally 0.7 volt p-p for a noncomposite signal).
3. A final check should be made at the output bus by switching the signal input through every possible path including the two fader banks.

SECTION 5

FREQUENCY AND TRANSIENT RESPONSE

In properly maintaining a television broadcast system, the maintenance department is concerned with the following general characteristics:

1. *Amplitude versus frequency response*—includes frequency- and phase-response correction circuits to obtain good transient response over the available system bandwidth.
2. *Amplitude linearity response*—includes so-called gamma correction circuits to compensate for certain film characteristics, transmitter modulation, and (in color) for picture tube characteristics.
3. *Differential gain*—concerns any change in gain of a given single frequency with change in level from black reference to white reference. Although it is most important in color systems, it is also a sensitive indicator of general system performance in monochrome.
4. *Differential phase*—involves change of phase of a given frequency with change in level from black to white. Again, it is of prime importance to color systems, and also is an extremely sensitive indicator of amplifier performance for monochrome systems.

Since the system bandwidth must be as great as practicable and also maintain the best possible transient response, these two characteristics must be simultaneously considered in adjustment. This section deals entirely with this problem, whereas amplitude and phase linearity, which require somewhat different test gear and techniques, are discussed in Section 6.

5-1. VIDEO AMPLIFIER THEORY FOR THE MAINTENANCE ENGINEER

The following review of video amplifier theory along with the details of Section 1 will enable the maintenance engineer to properly analyze and correct deficiencies that are revealed in television system measurements.

An important aspect of a video amplifier is its gain over a sufficiently broad bandwidth that is expressed in the familiar "gain-bandwidth" product as follows:

$$\text{Gain} \times \text{Bandwidth} = \text{Upper Frequency Limit}$$

The upper frequency of any tube at which gain is reduced to unity is:

$$f_u = \frac{g_m}{(2\pi)(C_t)}$$

This is simply the ratio of tube transconductance to total shunt capacity. The upper frequency limit (f_u for which gain equals unity) is expressed in megacycles when the total tube and wiring capacitance (C_t) is given in micromicrofarads and the transconductance (g_m) is given in micromhos.

For example, the input capacity of a single triode section of the type 12AT7 is 2.2 mmf, and the output capacity is 0.5 mmf. To this it is necessary to add a typical value of stray circuit capacitance of 15 mmf. C_t is therefore 17.7 mmf, which we may round off as 18 mmf. Then, since the g_m of a 12AT7 tube is 4,000 micromhos,

$$f_u = \frac{4{,}000}{6.28 \times 18} = 35 \text{ mc (approx)}$$

Thus, this triode will have unity gain at approximately 35 mc.

For the preceding example, the gain-bandwidth formula is:

$$\text{Gain} \times \text{Bandwidth} = 35 \text{ mc}$$

Thus the bandwidth for a gain of 10 is:

$$\text{Bandwidth} = \frac{35}{10} = 3.5 \text{ mc}$$

We see that achieving a gain of 10 (in an uncompensated amplifier) will limit frequency response to 3.5 mc.

Video amplifiers may be compensated by means of peaking circuits to achieve a broader bandwidth for a given gain. Because of the resultant phase shift across the passband, the problem becomes one of compromising between frequency response and ideal transient response. Also, ideal transient response can only be achieved by an infinite bandwidth system, which cannot exist in actual practice.

Remember the bandwidth and rise-time relationship discussed in Section 1-2, and remember that the rise time (a transient response) is equal to the square root of the sum of the squares of the individual rise times throughout a series of amplifiers.

A video amplifier resembles the average audio amplifier only in the basic RC method of coupling the stages. The bandwidth of an audio amplifier is based upon a concept that does not apply to the video amplifier. That is, the mid-frequency gain is given a value of unity, and the upper and lower limits of the bandwidth are taken as those points where the gain falls to 0.707 (or 70.7%) of the mid-frequency gain.

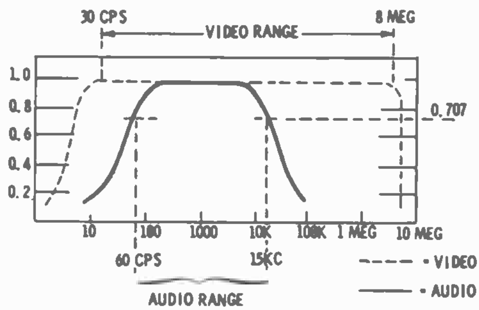


Fig. 5-1. Comparison between the ideal response curves of a video amplifier (dotted line) and an audio amplifier (solid line).

Fig. 5-1 illustrates the comparison of a good audio-amplifier response curve with that of a good video amplifier. Notice that the mid-frequency gain is given as unity, or 1. The points on the audio response curve which correspond to a gain of 0.707 of the mid-frequency gain are 30 cps and 15 kc respectively. These points correspond to a power loss of 3 db, or 50%, and they generally define the effective bandwidth of an audio amplifier.

Note, however, that the desired passband of the video amplifier is taken over that portion of the curve which is essentially flat (in practice, within 1 db, or 10% on voltage scale), denoting constant gain. Over this region of constant gain in an RC-coupled circuit the angular phase shift should be proportional to frequency, resulting in an equal time delay for all input frequencies within this range. In the upper and lower end regions where the gain changes rapidly, phase shifts cannot be proportional to frequency, and the time delay will therefore not be constant, result-in phase distortion in the video amplifier.

Since this type of distortion is detrimental to picture quality, its effects will be analyzed. A study of phase shift and the effect of circuit elements on this phase shift will serve to clarify the overall requirements of video amplifiers in bandwidth characteristics and transient response.

At the mid-frequency range of an amplifier, the shunt capacitances and coupling capacitances may be considered to have negligible effect on the amplification, and they may be represented by an equivalent circuit as in Fig. 5-2. In this range of frequencies, the gain of the amplifier may be assumed to be ap-

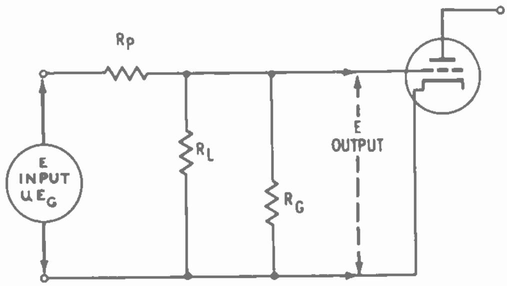


Fig. 5-2. Equivalent circuit for RC-coupled amplifier in mid-frequency range.

proximately the product of the tranconductance of the tube and the load resistor R_L. This is expressed as:

$$G_m = g_m R_L$$

where,

G_m is equal to mid-frequency gain,
g_m is equal to tube transconductance,
R_L is equal to load (coupling) resistor.

At the higher frequencies, shunt capacitances across the load resistor become effective; this attenuates the amplifier response with increase in frequency. An equivalent circuit at the higher frequencies is illustrated in Fig. 5-3. At the lower frequencies, the impedances of cathode- and screen-bypass capacitors and the coupling capacitor serve to attenuate the lower frequencies. The equivalent circuit at low frequencies is shown in Fig. 5-4.

To increase the passband of the video amplifier so that frequency and phase distortion may be held to a minimum, low-frequency and high-frequency boosting circuits are used. In addition to these special circuits, a relatively low value of R_L is used, at the sacrifice of gain, to achieve a broader flat response than is possible with conventional values of load resistors. This is illustrated by the curves in Fig. 5-5. In commercial equipment the coupling resistors generally range from 680 to 2,000 ohms.

Therefore, the internal plate resistance for pentode TV video amplifiers is much greater than R_L, and the grid resistor for the

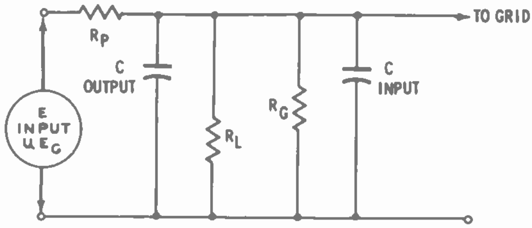


Fig. 5-3. Equivalent circuit for RC-coupled amplifier at high frequencies.

following stage is also much greater than R_L. With these conditions prevailing, the current change through R_L is in phase with the generator voltage, since the internal resistance (pure resistance) of the tube is actually the principal impedance in the load circuit (see Fig. 5-4). Also the internal plate-current change is essentially equal in magnitude and phase to the current

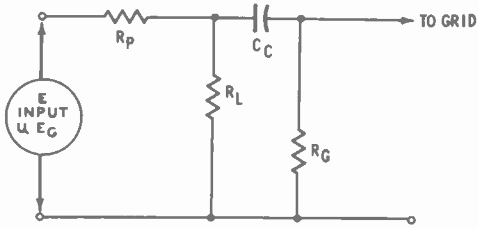


Fig. 5-4. Equivalent circuit of RC-coupled amplifier at low frequencies.

change through R_L, and the voltage drop across R_L is in phase with the plate-current change. However, since coupling capacitor C_c is reactive at the low frequencies, the current through R_g will lead the voltage change applied and is therefore displaced in phase. The amount of leading phase-shift is determined by the ratio of the capacitive reactance of C_c to the resistance of R_g, and it may therefore be seen to increase with a decrease in frequency

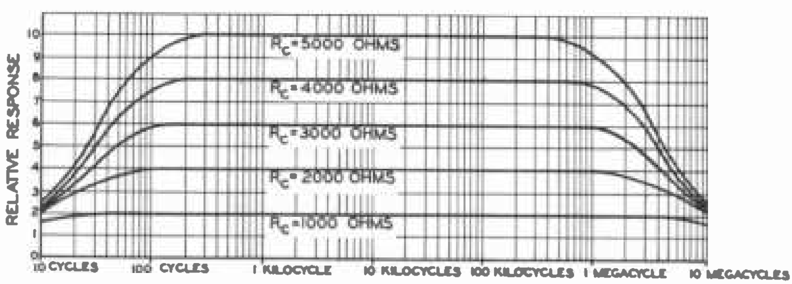


Fig. 5-5. Effect of plate coupling resistor on gain and bandwidth for typical high-transconductance pentode.

(larger capacitive reactance with a decrease in frequency). Since the current change through resistor R_g is displaced in phase, the corresponding voltage drop across this resistor is likewise displaced in phase, and low-frequency phase distortion is prevalent.

To keep the voltage change across the grid resistor in phase with the plate-current change through R_L, the time constant of the coupling network (C_c, R_g) must be effectively increased so that the reactance of the coupling capacitor becomes negligible even at the lowest frequencies in the passband. The two most obvious means of accomplishing this are *not* practical in design. First, the coupling capacitor could be made extremely large in value to provide negligible attenuation at the lowest frequencies. Larger capacitors, due to their physical size, however, increase the effective shunt capacities to ground, severely attenuating the higher frequencies. Also, circuit instability in the form of motorboating may occur. The grid resistor (R_g) could be greatly increased so that the relative reactance of the coupling capacitor would be very small. This cannot be done beyond the limits determined by the maximum allowable grid resistance given in the manufacturers tube data for the particular tube used. Too much grid resistance allows gas current (positive-ion current) to accumulate on the grid, resulting in excessive average plate-current. The practical solution, therefore, is to shift the phase of the voltage changes across R_L so that, in conjunction with the coupling capacitor, the current changes through R_g are in phase with the current changes through R_L.

This is the function of the low-frequency boosting circuit shown in Fig. 5-6. The desired relative phase shift across R_L results when the two parallel branches of the load circuit are similar in impedance characteristics insofar as equal phase angles are concerned. This is accomplished when the product of the plate load resistance and the decoupling capacitance is equal to the product of the coupling capacitance and the grid resistance (equal time constants). This is written:

$$R_L C_d = C_c R_g$$

The current changes through the grid resistor will be in phase with the current changes through the plate load resistor. Decoupling resistor C_d must be used to provide a DC path to the plate of the tube, and it must be much larger than the plate load resistor. Since this value must be limited in size by the available B supply voltage (plate current must flow through this resistor as well as R_L), if R_d cannot be made large in ratio to

$$(\tfrac{1}{2}\pi f \times C_d)$$

then compensation is made by shunting the coupling capacitor with a suitable value of resistance. This value is such as to restore the similarity of the two parallel branches, and the maintenance engineer will find this method used in some cases. To avoid a DC path to the grid of the following stage, an extra capacitor is used, as shown by the dotted lines of Fig. 5-6.

We have seen how low-frequency phase distortion is avoided by a properly designed decoupling circuit that extends the response at the low frequency. High-frequency phase distortion is avoided by properly designed peaking circuits used to extend the passband through the highest frequencies to be passed. At these frequencies the total shunt capacity across the load results in a

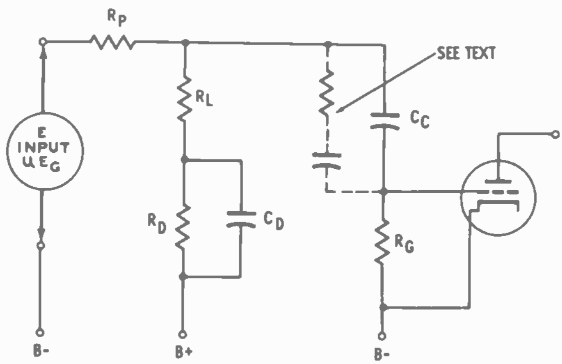


Fig. 5-6. Equivalent circuit of video amplifier at low frequencies with addition of low-frequency decoupling (boost) circuit. To provide equal time constants of the two parallel branches of the load circuit; $R_L C_D = C_C R_G$.

lagging phase displacement in comparison to the mid-frequency range. A properly adjusted peaking circuit will provide a phase shift to compensate that of the effective load network.

Tube amplifying action in itself causes a normal 180° phase shift (exact phase inversion) in the output circuit compared with the input signal. This means that a uniform time delay occurs for all frequencies in the passband. If each frequency is considered separately, it may be seen that for a uniform time delay a different phase shift must occur at each frequency, so that the resulting phase displacement is proportional to frequency. At the horizontal scanning frequency, the time of scan of one line is 63.5 microseconds. Across a 20-inch monitor this corresponds to about 6.35 microseconds per inch. Should any part of the video signal be delayed only one microsecond above the normal phase inversion, that portion of the picture would be displaced approximately one-sixth of an inch. Obviously if all frequencies in the passband were delayed one microsecond, the entire picture would simply be displaced about one-sixth of an inch, but it would result in a satisfactory picture that could be centered by the centering controls. However, the resistance-capacitance elements of the coupling networks (when uncompensated or when the compensating circuits are out of adjustment) cause a shift in phase which differs both in direction and number of degrees for different frequencies. Phase distortion may be seen to be directly related to amplitude distortion, since minimum phase distortion is ob-

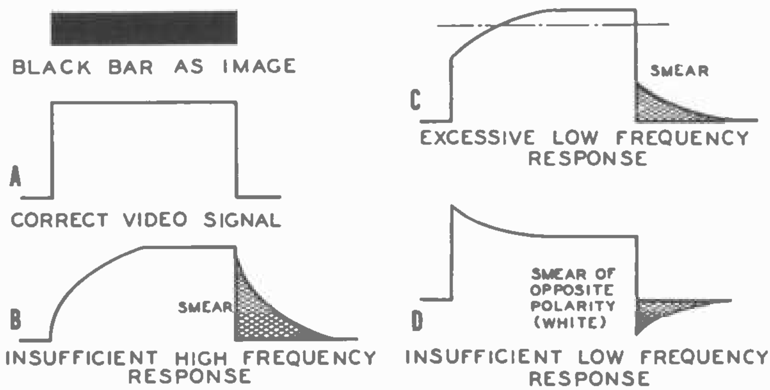


Fig. 5-7. Square-wave video signals showing the effect of amplifier deficiencies.

tained only by a long, flat-topped response over the desired passband and with no sharp cutoff above the passband.

This is emphasized by considering the content of the video signal when the image to be scanned is the extreme case of a black bar on a white background (Fig. 5-7). In Fig. 5-7A is illustrated the ideal response where an abrupt rise in the tube-current of an image orthicon would occur as the scanning aperture encountered the leading edge of the blackbar. At the trailing edge of this bar, the current should abruptly fall to the "white" level. Amplifier circuits that must faithfully reproduce such current changes will have exceptional amplitude and phase characteristics. If the amplifier has insufficient high-frequency gain, the leading edge of the amplified wave becomes a gradual slope, instead of a sharp rise as in Fig. 5-7B. The reproduction is that of a gradual shading from gray to black on the leading edge of the black bar, and a gray to white "smear" on the trailing edge. Fig. 5-7C shows the effect of overcompensation of the lower-frequency response. The effect is similar to that of insufficient high-frequency response, but it is not as pronounced. Insufficient low frequency response with attendant phase shift is shown in Fig. 5-7D. Since loss of lows causes the flat top of the ideal square-wave response to become tilted as shown, black to gray shading occurs at the leading edge, while a white to gray shading smear results at the trailing edge.

One possible effect is illustrated in Fig. 5-8. Such an effect may also be produced in peaking circuits by transient oscillations due to shock excitation of suddenly changing square-wave currents through the coils. Damping resistors are used across series peaking-coils, which are often adjustable in commercial circuits for proper peaking and damping characteristics.

Fig. 5-8. Station test pattern showing the effect of reversed-polarity smear on transient oscillation.

The curves in Fig. 5-9 show the correct relationship of phase shift being proportional to frequency (uniform time delay at all frequencies), in comparison to that of an average amplifier not compensated for flat, high-frequency response. The relation of time delay, phase shift, and frequency is the following:

$$\text{Time Delay} = \frac{\text{Phase Shift } (\theta) \text{ in degrees}}{360^\circ \times \text{Frequency in cps}}$$

From observation of the desired characteristic curves at a frequency of two megacycles, the phase shift (θ) is 30°. Therefore:

$$\text{Time Delay} = \frac{30^\circ}{360\,(2 \times 10^6)} = \frac{30^\circ}{720 \times 10^6} = 0.041 \text{ microsecond}$$

From the same curve at 3 megacycles, $\theta = 45^\circ$. Therefore:

$$\text{Time Delay} = \frac{45^\circ}{360\,(3 \times 10^6)} = \frac{45^\circ}{1{,}080 \times 10^6} = 0.041 \text{ microsecond}$$

From the same curve at 4 megacycles, $\theta = 60^\circ$. Therefore:

$$\text{Time Delay} = \frac{60^\circ}{360\,(4 \times 10^6)} = \frac{60^\circ}{1{,}440 \times 10^6} = 0.041 \text{ microsecond}$$

Thus for phase shift proportional to frequency, a uniform time delay occurs throughout the video amplifier at any frequency. This results in a uniformly shaded picture, other factors being equal. Let's see now the effect of the uncompensated amplifier phase-shift curve on time delay at various frequencies. From the uncompensated curve at 2 mc, $\theta = 27°$.

$$\text{Time Delay} = \frac{27°}{360\,(2 \times 10^6)} = \frac{27°}{720 \times 10^6} = 0.037 \text{ microsecond}$$

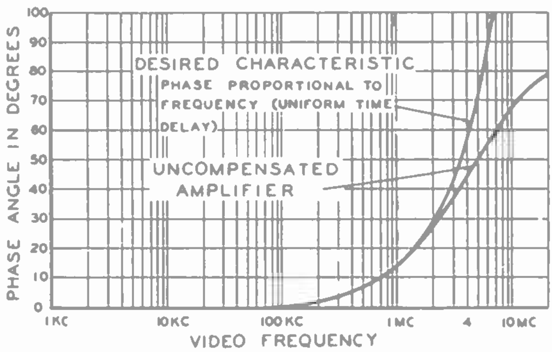
At 3 mc, $\theta = 38°$.

$$\text{Time Delay} = \frac{38°}{360\,(3 \times 10^6)} = \frac{38°}{1{,}080 \times 10^6} = 0.035 \text{ microsecond}$$

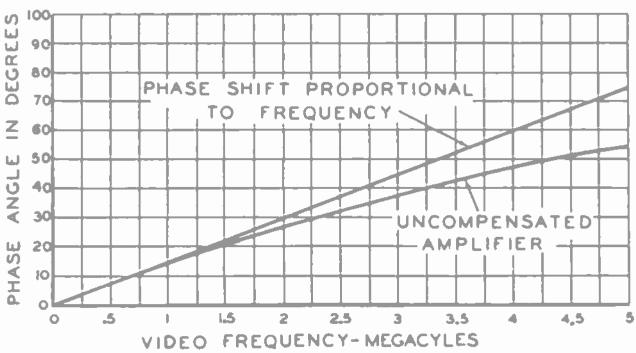
At 4 mc, $\theta = 48°$

$$\text{Time Delay} = \frac{48°}{360\,(4 \times 10^6)} = \frac{48°}{1{,}440 \times 10^6} = 0.033 \text{ microsecond}$$

Thus, phase distortion occurs, and the ringing effect in the picture is apparent. It is noted that for high-frequency phase shift, the time delay decreases as frequency increases. This is the effect of a lagging phase shift across the coupling network; it is



(A) Ideal phase characteristics plotted in logarithmic coordinates.



(B) Same curve as in A but on a linear basis.

Fig. 5-9. Phase-shift requirements of a video amplifier for perfect reproduction.

due to shunt capacitances. At the low frequencies, the leading phase shift results in time delays that increase as the frequency decreases. Most of the ringing effect (smearing after black bars in the picture) is due to low-frequency phase distortion. This may also be noticed as a gradual shading in backgrounds from top to bottom of the reproduced picture, which cannot be corrected by vertical shading controls. High-frequency phase distortion results in general deterioration of resolution.

The *figure of merit* of any video amplifier tube is a ratio expressing general capabilities in amplifying high frequencies. It is well known that gain at the higher frequencies is proportional to the transconductance (g_m) of the tube and inversely proportional to the total shunt capacitance (C_t). The mathematical expression is:

$$\text{Figure of merit} = \frac{g_m}{C_t}$$

where,

g_m is in micromhos,
C_t is in micromicrofarads.

The shunt capacitances of any given tube increase when it is used as an amplifier. For example, the rated input capacity of a 6AC7 tube is 11 mmf. Because of the *Miller effect*, the effective input capacitance depends on the stage gain in the following manner:

$$C_t = C_{gk} + C_{pk} + C_{gp}\ (1 + \text{gain}).$$

where,

gain $\cong g_m R_L$.

Thus the total shunt capacity (in addition to stray capacitance) is the sum of the grid-cathode, plate-cathode and grid-plate capacities times the quantity 1 plus stage gain. Since gain is the product (approximately) of the transconductance and load resistance, it may be seen that the higher is the value of R_L (greater stage gain for given tube), the higher is the effective shunt capacitance. Thus pure resistance loads reflect back to the input as capacitance. When the plate load is complex, the resistance reflects as capacitance, while the reactive portion of the load reflects as resistance.

The performance of a video amplifier must be measured in both gain and bandwidth. It has been shown how the choice of the value of R_L makes these factors mutually dependent. Therefore, as one factor is made to increase, the other factor is made to decrease, and the product will remain constant. This characteristic therefore is used to express the *figure of merit* of a video amplifier and is stated: figure of merit = gain × bandwidth.

Thus if a given amplifier has a gain of 40 with a bandwidth of 2 mc, the load impedance could be halved, resulting in a gain of 20 and a bandwidth of 4 mc. If a given amplifier has a gain of 20 and a bandwidth of 8 mc, the load impedance could be doubled, resulting in a gain of 40 and a bandwidth of 4 mc. It is imperative for the maintenance engineer to understand these relationships so that circuit characteristics may be correctly interpreted.

5-2. VIDEO AMPLIFIER HIGH-FREQUENCY COMPENSATION

Nearly all modern video amplifiers used in distribution systems employ negative-feedback circuits for wide bandwidth with minimum phase distortion. This notebook is primarily concerned with amplifiers of this type, where the units provide isolation of feeds and are normally operated as unity-gain devices. However, video peaking-circuits are encountered in the modulator stages of transmitters, video monitors, and elsewhere. Therefore, both types of frequency compensation are covered. In general, it should be understood that amplifiers used primarily for isolated feeds at unity gains employ negative-feedback designs, while those used to increase the level of the video signal employ peaking coils. Some units use a combination of each.

Fig. 5-10 shows basic method of high-frequency adjustable compensation in negative-feedback amplifiers. Adjustment of the frequency compensating trimmer capacitor varies the amount of high-frequency negative-feedback; hence, it compensates the response curve. Usually, several such circuits in cascade are employed.

Series and shunt peaking have been mentioned several times. The maintenance engineer should be familiar with such circuits used to extend the bandwidth of amplification.

Fig. 5-11 illustrates the shunt-peaking method for compensating usual high-frequency losses. Electrically, this is a parallel resonant circuit, designed so that the resonant frequency is approximately 1.41 times the highest frequency to be amplified. Thus a boosting of the "high-pass" frequencies is affected, with no effect on the lower pass frequencies.

In practice, peaking coils vary in value between one and several hundred microhenries. Ten to fifty microhenries is the average range found in commercial equipment for shunt peaking.

Fig. 5-12 illustrates the series-type peaking circuit. The series coil, in combination with the effective circuit capacitance, forms a low-pass filter network. At first though it might appear that such a circuit defeats the purpose intended; that is, to increase the efficiency of amplification at higher frequencies. A basic analysis is therefore important.

It is necessary to make C1 twice the capacity of C2. The load resistor R_L is connected to the low capacitance side of the circuit. In practice, a physically small capacitor may be found in the circuit where the effective capacity (C1) would appear. It is

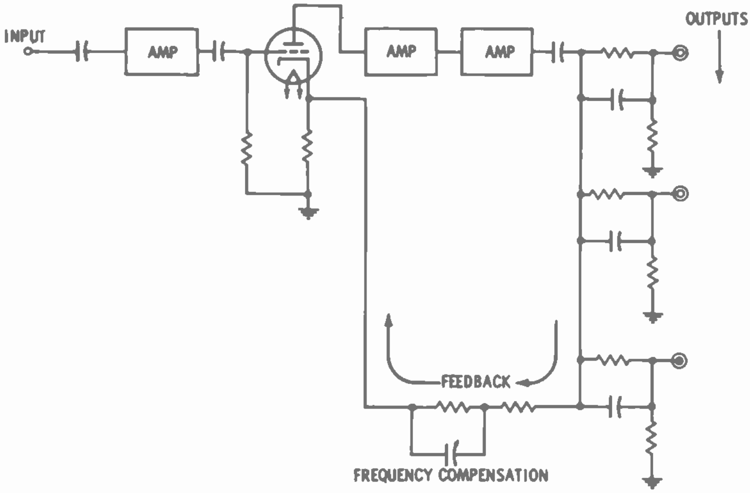
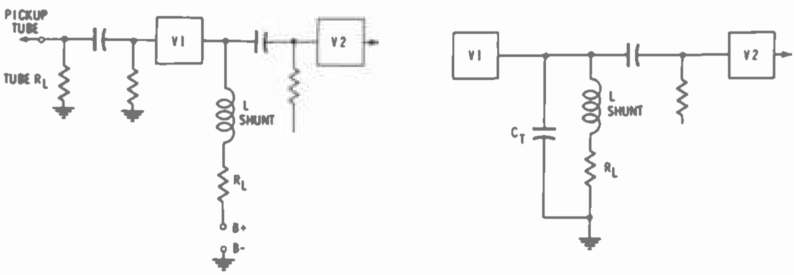


Fig. 5-10. Typical video amplifier circuit with negative-feedback frequency compensation.

therefore considered to be in parallel with C1, effecting a 2:1 ratio in effective capacities.

Inductor L and capacitor C1 form a series resonant circuit with an effective resonant increase in current as the frequency increases. C2 is separated from C1 by inductance L, with a result-

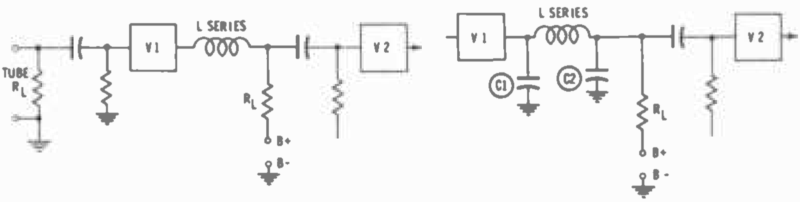


(A) Circuit.

(B) Equivalent circuit.

Fig. 5-11. Shunt-peaking compensation.

ing reduction in shunting effect across L at high frequencies in the passband. Since the voltage drop in the series resonant circuit increases with increase in frequency and is applied across load resistor R_L shunted by C2, the voltage developed in the load will



(A) Circuit. (B) Equivalent circuit.

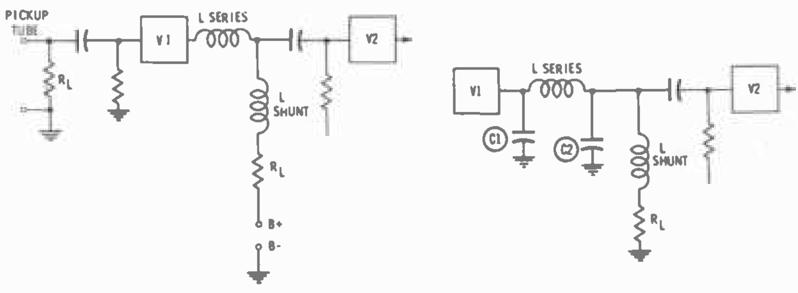
Fig. 5-12. Series peaking.

likewise increase for frequencies in the high passband. This video voltage is then coupled to V2 in the orthodox manner. The increase of higher frequency voltages resulting from the resonant effect of L and C1 more than offsets the decrease in the reactance of C2 with increasing frequencies.

Fig. 5-13 illustrates the most efficient design used in video amplifiers to increase the high-pass range. This is the shunt-series peaking circuit and is a combination of the two methods just discussed. The increased voltage gain from such a circuit is materially aided by the fact that a load-resistor value approximately 80 per cent greater than is possible with a simple shunt-peaked circuit may be used. Since the gain of a stage is equal to the transconductance of the tube times the value of R_L, the gain is appreciably increased by this factor alone. It should be remembered that the value of plate load resistance is limited in ordinary amplifiers by the bandpass required; too great a value of load resistance reduces the bandpass capabilities of the stage.

The relative gains of video amplifier circuits may be tabulated as follows:

1. Uncompensated 0.707
2. Shunt peaked 1.0
3. Series peaked 1.5
4. Shunt-Series peaked 1.8



(A) Circuit. (B) Equivalent circuit.

Fig. 5-13. Shunt-series peaking.

Video-amplifier output stages assume one of two general forms: either the cathode follower or a specially designed plate-loaded circuit. The cathode follower is so termed because the load is connected across the cathode resistance, and the load signal is therefore of the same polarity (follows) as the grid signal. In the conventional plate-loaded circuit with the load coupled to the plate of the preceding stage, phase inversion of the signal results. This is because as the grid signal voltage swings in the negative direction, plate current is decreased (more grid bias), and the signal voltage drop is decreased across the plate resistor. This causes the coupled point in the plate circuit to become more positive as the grid becomes more negative; thus, the signal phase is 180° out of phase with the applied signal.

The cathode follower is shown in equivalent circuit form in Fig. 5-14. The primary function is to match a high impedance to the 75-ohm standard impedance of interconnecting coaxial lines and to preserve the proper waveshapes in the video signal. As the

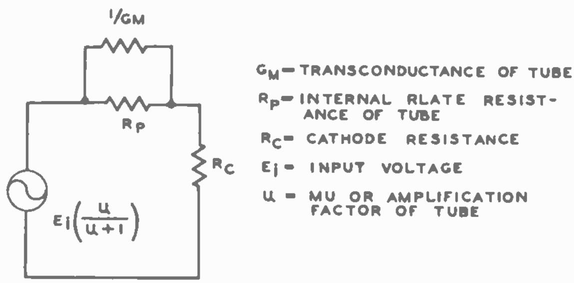


Fig. 5-14. Equivalent circuit of cathode follower.

grid signal swings in the positive direction, plate current through the cathode resistor increases and the I_pR_c drop therefore increases. As the grid signal swings in the negative direction, the I_pR_c drop is decreased. Thus, the load signal follows the input signal, and no phase inversion takes place.

The amplification factor (mu) in Fig. 5-14 is multiplied by the input voltage (E_i) to obtain the generator voltage. The term within the parenthesis is the generator impedance for this equivalent circuit. If the tube is a pentode, the internal plate resistance (R_p) is very large with respect to g_m, and amplification factor is large. The approximate expression for gain may then be expressed:

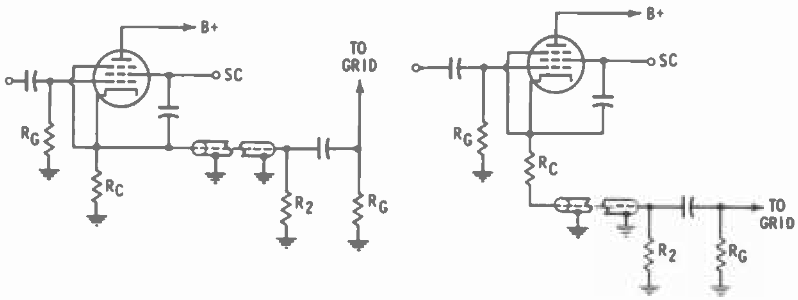
$$\text{Gain} = \frac{R_c}{R_c + 1/g_m}$$

Thus in the case of a pentode cathode-follower, the equivalent generator voltage is approximately the input voltage (E_i), and the internal impedance is $1/g_m$. Thus the quantity of $1/g_m$ is effectively in parallel with R_c.

From the preceding expression for gain of a cathode follower, it is noted that the denominator must always be greater than the numerator. Therefore the gain may approach unity (same output as input) but, it can never exceed unity. Amplification is not attempted in a cathode follower.

It is important for the maintenance engineer to understand which of the operating characteristics of the cathode follower may affect frequency response and hence phase distortion. From the preceding discussion it was pointed out that the output impedance is the effective cathode resistance (effective R_c is the actual cathode resistance in parallel with load resistance at the receiving end of the coax), in shunt with $1/g_m$. Any factor that would affect the transconductance of the tube with applied signal would simultaneously affect the output impedance. As an illustration, suppose the grid received an instantaneous signal of sufficient negative value to cut off plate current. At this instant, g_m becomes zero, and theoretically the generator impedance is infinite. Thus the output impedance is the effective R_c, and the value of load impedance has changed with applied signal. *It is very important that a cathode-follower stage be operated within its normal limits.* While this is obviously true of all video stages, the cathode follower is extremely sensitive to limits over which it obtains optimum results. It follows that any slight change in circuit components or the tube itself may require priority attention of the maintenance department.

Since the internal impedance of the cathode follower is $1/g_m$, the effective R_c should nearly match this value for good power-transfer. The type 6AG7 tube is sometimes used as a cathode-follower output in video amplifiers. From tube data sheets it is found that the g_m is 11,000 micromhos. Thus 1/11,000 micro units is approximately 90 ohms. We may see now what effects the "effective" value of the cathode resistance. Since the transconductance of a 6AG7 is relatively high, the value $1/g_m$ is a very small quantity which is in shunt with R_c. As shown before, this value is about 90 ohms. Fig. 5-15 shows the two common methods of coupling the load to a cathode follower. If the internal impedance of the tube is high in ratio to the required network (for proper matching and optimum power output consistent with required bandwidth), the circuit of Fig. 5-15A is used. If the internal impedance is low in ratio to the required coupling network, the circuit of Fig. 5-15B is used. In either case, for the one that properly matches the cathode-follower to the transmission line, the DC component may be seen to be on the line. The coupling capacitor is almost universally used at the receiving end as shown,



(A) Internal impedance high with respect to coupling network R_c-R_2.

(B) Internal impedance low with respect to coupling network R_c-R_2.

Fig. 5-15. Two popular types of cathode-follower circuits for video amplifiers.

connected directly to the high-impedance grid circuit. This results in optimum performance with negligible capacitive effects across the line.

Whenever the cathode follower may be directly-coupled to the line as illustrated (a 75-ohm load coupled to a high-impedance amplifier at the receiving end), the circuit when properly adjusted exhibits excellent frequency characteristics down to and including DC. DC on the line is a disadvantage when an amplifier is used to feed transmission circuits that terminate in 75 ohms. The use of a cathode follower to feed such lines would require an extremely large coupling capacitor (to block DC) to couple such a low resistance. The value of such a capacitor would necessarily be in the neighborhood of 2,000 mfd to achieve good low-frequency response down to 60 cps between two low-impedance devices. Therefore, the engineer will find many amplifiers incorporating a special type of plate-loaded output stage.

Fig. 5-16 illustrates a typical plate-coupled video output stage. In the preceding video amplifier stages described, the plate re-

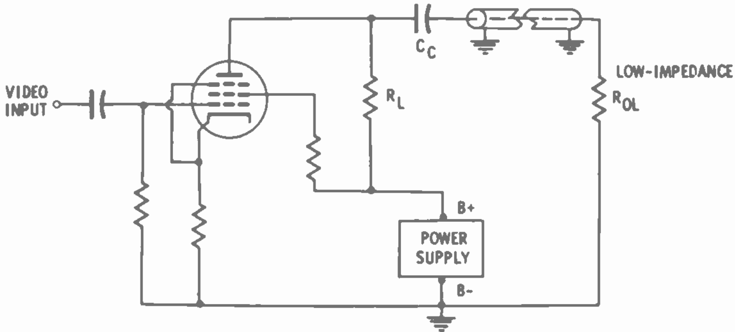


Fig. 5-16. Schematic of typical plate-loaded, video output stage.

sistance (R_L) was very low in magnitude compared with the following grid resistor)R_g). The time constant of the coupling network therefore was essentially C_cR_g. In the circuit of Fig. 5-16, however, R_L is not negligible, since C_c couples into a relatively low resistance of about 75 ohms. Therefore, the time constant upon which good low-frequency response depends is largely determined by the values of R_LC_c. In this instance, R_L may be made as large as possible consistent with the power-supply voltage available.

Design engineers ascertain the figure of merit of any coupling circuit at low frequencies by simply adding to the time constant of the circuit the lowest frequency to be passed with good fidelity. This is given as: figure of merit at lowest frequency (LFM) = RfC. In practice, this value should not be less than 20. From this relationship we may see how the minimum value of the coupling capacitor C_c (Fig. 5-16) may be determined. The output load (R_{OL}) will have a low value of around 75 ohms and need not be considered here.

Consider an output stage using an R_L of 5,000 ohms. Solving for the value of C_c from the previous formula, and considering 20 as the minimum value of LFM (for 30 cps):

$$C = \frac{20}{RF} = \frac{20}{5{,}000 \times 30} = 0.000133 = 133 \text{ mfd}$$

Thus C_c in this case is approximately 133 mfd. Consider now the necessary value of such a coupling capacitor when used at the cathode of a cathode follower that must block DC. From our previous discussion if cathode resistor R_c is to match the internal impedance of the tube, and the tube used is a 6AG7, then:

$$R_c = \frac{1}{g_m} = \frac{1}{0.011} = 90 \text{ ohms}$$

If we choose the same minimum figure of merit at 30 cps, then:

$$C = \frac{20}{90 \times 30} = 0.007400 \text{ (approx.)} = 7{,}400 \text{ mfd}$$

The size of such a capacitor which could withstand even the relatively small voltages encountered would be prohibitive. Thus, the reader should now have a reasonable insight as to why cathode followers are not used to feed lines coupling two low-impedance devices between which the DC must be blocked from the line.

Most modern plate-coupled, low-impedance output stages take the form of the single-ended, push-pull circuit shown in Fig. 5-17. The phase-inverted signal at the V2 plate coupled to the grid of V1 provides a configuration resulting in the advantages of push-pull amplification while supplying a single-ended output for the coaxial cable. Typical values of output coupling capacitors range from 200 to 400 mfd.

The use of transistors in video and pulse amplifiers is complicated by two basic, inherent factors: low input-impedance of transistors and frequency-dependence of the input-impedance.

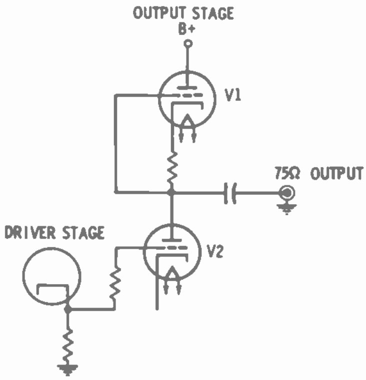
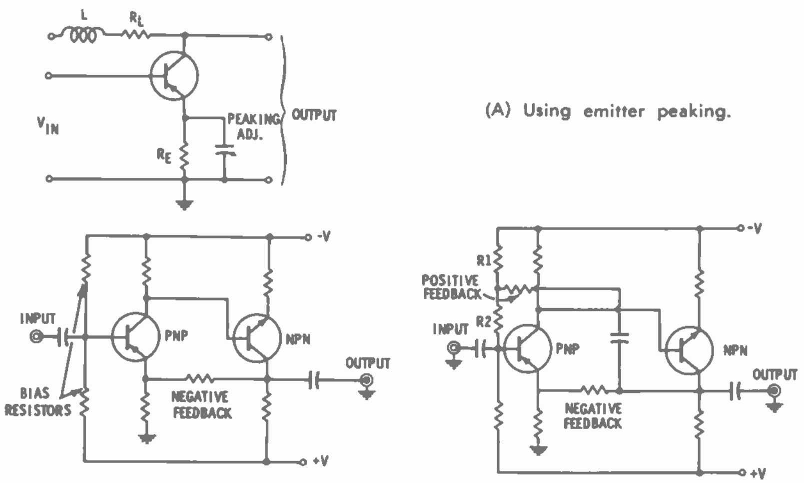


Fig. 5-17. Single-ended, push-pull, output stage.

Since, in general, the response of typical high-frequency transistors is influenced more by input parameters than by output characteristics, peaking or high-frequency compensation is generally associated with the base-to-emitter circuit. Fig. 5-18A shows a typical input-stage transistor circuit which employs emitter peaking. This type of peaking improves both the linearity and stability of the circuit and minimizes the effect of varying



(A) Using emitter peaking.

(B) Using negative feedback.

(C) Using positive feedback.

Fig. 5-18. Typical transistor circuits.

internal parameters of the transistor. Transistor circuits also employ conventional series and shunt peaking-coils which function in the same way as for the vacuum tubes previously described.

In video distribution a large number of amplifiers may be necessary, requiring a "loop-thru" on the inputs. Vacuum-tube amplifiers provide the high-impedance necessary in such cases, but transistors are inherently low-impedance input devices. It is possible, however, to raise the equivalent input-impedance to any value required by the driving source by the use of negative feedback or a combination of negative and positive feedback. Actually, such circuits exchange gain for high input-impedance.

The negative-feedback resistor of Fig. 5-18B returns a portion of the output voltage to the emitter of the first transistor. Since this voltage is out of phase with the input voltage, the net voltage drop beween the emitter and the base is reduced, and input loading is correspondingly reduced. With one transistor a PNP type and the other an NPN type, larger bias resistors may be used to reduce the loading in the bias network.

A positive-feedback loop is sometimes used to still further offset the bias-resistor loading effect (Fig. 5-18C). The value of the positive-feedback resistor is chosen to produce a voltage at the junction of R1-R2 which is in phase with and equal to the input voltage. This reduces the voltage drop across the bias resistors, which is equivalent to raising the resistance of the input network.

5-3. WHAT TO EXPECT IN PICTURE RESOLUTION

There are two resolution factors for a television picture; (1) vertical resolution, which is independent of system bandwidth, and (2) horizontal resolution, which is directly related to system bandwidth.

Vertical resolution determines how well horizontal lines in a picture are resolved. The maximum vertical resolution is fixed by the number of active scanning lines. The United States standards call for a total of 525 lines. Vertical blanking time is approximately 7.5% of the total frame time; therefore: $525 \times 0.075 = 39.375$, or about 40 lines blanked out. This leaves a total of 485 active picture lines scanning from left to right and top to bottom of the image.

This would appear to indicate that 485 horizontal lines spaced vertically in the image would be resolved. But in practice the slight spacing between the scanning lines and the fact that the scanning spot will straddle some of the lines both tend to reduce the utilization of the maximum number of active lines by a factor which can be taken as 0.7 times the total active lines.

Thus 485 × 0.7 = 340 lines, approximately. Fig. 5-19 shows the horizontal wedges of a test pattern in which the lines merge at a point which represents 340 black and white horizontal lines in the total image height. This is a typical value of vertical resolution at both studio and transmitter outputs.

Horizontal resolution is the ability to define vertical lines in the image. The essentially round shape and the fact that the scanning spot is not infinitely small both place an immediate limitation on the ability to reproduce rapid picture transitions. In Fig. 5-20 where the beam suddenly encounters a sharp vertical line representing transition from black to white, the resulting signal is not a square wave but more nearly a sine wave.

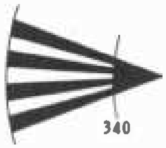


Fig. 5-19. Wedges for visual interpretation of vertical resolution.

This is the effect not only from the round shape and finite size of the scanning beam (often termed "aperture"), but also from the fact that a straight vertical line represents an infinite rise time—which would require infinite bandwidth. But an infinite bandwidth is impossible to obtain in practice, and the rise time of the signal representing the instantaneous transition is limited to the practical system-bandwidth available. When the total rise times (rise plus decay times) equal the spacing between lines, they are not visible as separate picture elements, and they are not resolved (Fig. 5-21).

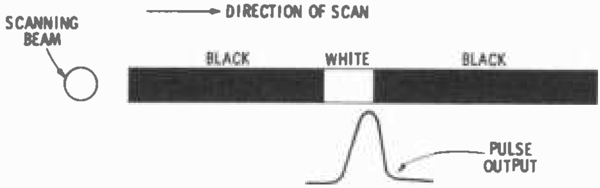


Fig. 5-20. Scanning beam encountering sharp vertical line.

Assuming that the scanning beam is properly focused, the limitation on horizontal resolution is the system bandwidth. The pulse rise time representing an instantaneous transition in the picture is directly related to the system bandwidth. As a rule of thumb, 80 TV lines require a 1-mc bandwidth, as explained in Chart 5-1.

The maximum video bandwidth of the transmitter is 4.18 mc. Therefore the maximum H-resolution to be expected from the transmitter is 4.18 times 80, or 334 lines, about equal to V-resolution. Thus the transmitter operator will observe essentially the same resolution on the vertical and horizontal wedges of the test chart for an optimized tuning condition. The test chart transmitted to the home viewer is normally arranged to have a minimum

Fig. 5-21. Wedges for visual interpretation of horizontal resolution.

Chart 5-1. Derivation of 1 Megacycle = 80 TV Lines

1. The aspect ratio of the picture is 4 units wide to 3 units in height. This requires the horizontal resolution of a test chart to be related to the height.
2. This is to say, if black and white lines with the same thickness as those indicated at the 340 position on the horizontal wedge were placed adjacent to one another, a total of 340 could fit into the height of the chart. (Width of each line equals 1/N × picture height.) Or, 340 × 4/3 or 452 of the same thickness lines could be placed in the width of the chart.
3. Lines of H resolution per cycle equals 2. One cycle consists of two alternations (positive and negative) therefore two picture elements—1 white, 1 black.
4. Then the H resolution factor equals 2/1.33 (Note: 4/3 = 1.33).
5. Since 2/1.33 equals 1.5 the H resolution factor is 1.5, and this factor times the active line interval specifies the number of horizontal-lines resolving power per megacycle of bandwidth.
6. The total line interval is 63.5 microseconds. Horizontal blanking is usually 11 microseconds so that the active line interval is 63.5 — 11 or 52.5 microseconds.
7. 52.5 × 1.5 = 80 TV lines/mc (approx).

wedge equivalent to 320 lines horizontal resolution. Table 5-1 relates bandwidth to rise time and horizontal resolving power in TV lines.

Although the transmitter is limited to essentially a 4-mc video bandwidth, it is well known that due to accumulative factors

Table 5-1. Relationship Between Bandwidth, Rise Time, and Horizontal-Resolving Power

Bandwidth (mc)	Rise Time (us)	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

of frequency-phase distortions, the better the picture going into the transmitter, the better is the picture received in the home. For this reason, studio facilities are normally maintained to approximately twice the bandwidth employed by the transmitter, or 8 mc. Specifications of modern television studio equipment are well within 1 db to 8 mc, and within 3 db to 10 mc. Thus the studio operator will normally observe about 640 lines horizontal resolution ($8 \times 80 = 640$) which approaches the upper limit of the resolving power of most picture monitors and which is almost a 2:1 ratio to the vertical resolution possible. This system bandwidth slightly exceeds the practical day-to-day resolution capabilities of the image orthicon and vidicon pickup tubes, with the exception of the newest 4½-inch image orthicon.

In addition, the best monochrome stations are maintained within color standards so that studio facilities are practically 100% perfect in amplitude linearity, within 5% differential gain (at 3.58 mc), and 3° differential phase. The overall (from studio inputs to transmitter output) differential gain should be within 20%, and differential phase should be within 10° at 3.58 mc (covered in Section 6).

5-4. SPECIAL TEST EQUIPMENT ACCESSORIES

It is often desirable to employ "keyed" test signals phased by the station sync generator to eliminate the test amplitude during horizontal and vertical blanking intervals. This enables checking the many types of amplifiers incorporating line-to-line clamps which otherwise need to be modified if straight test signals are used. Although commercial equipment is available for keyed sine waves, video sweep, stairstep signals, etc., there is an apparent scarcity of available units that will properly process a square-wave signal.

Fig. 5-22 shows a simple transistor devised for this purpose. The 2N384 drift-field transistor is very reasonably priced and is also very effective for video use. The equalization network (R1-C1) at the signal input corrects for changes in input resistance with changes in frequency and holds the output constant over a 10-mc bandwidth. The interlead shield is connected to the case, and this center lead should be grounded.

Introduction of sync and/or blanking pulses of negative polarity drives the transistor to cutoff for the input signal, and the amplitude of the pulse as adjusted by R6-R7 appears across the output load (input of system to be checked).

Fig. 5-23A illustrates the keyed output when the test signal is a 60-cycle square wave. The setup (blanking) level is adjustable by R6 to the desired amount of pedestal. This type of signal

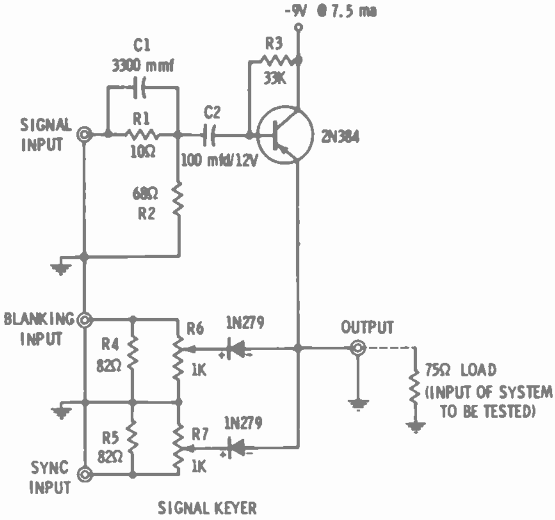
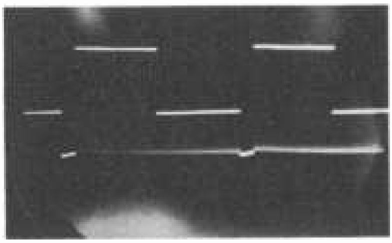


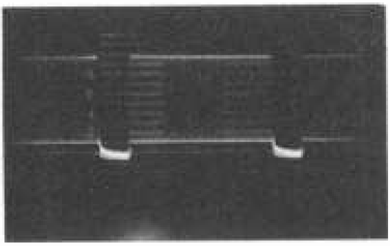
Fig. 5-22. Signal keyer.

results in a clean, composite blanking interval, and with the addition of sync no modification is necessary for units employing clamps.

The keyer may also be used for sine waves, as shown by Fig. 5-23B with only blanking inserted. Fig. 5-23C illustrates the



(A) 60-cycle square wave with station blanking.



(B) 1-megacycle sine wave with station blanking.

(C) Same as B with addition of station sync.

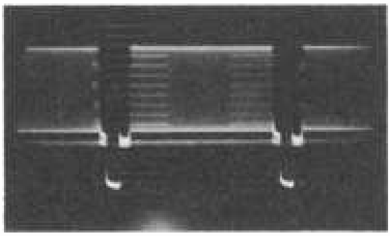


Fig. 5-23. Output of keyer in Fig. 5-22 for various test signals.

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.

Another useful accessory is the simple differentiating network of Fig. 5-24. The switch selects the proper time constant for horizontal or vertical drive, and delivers a positive trigger from the trailing edge of the input pulse. Square-wave generators that accept external sync inputs are more stable with a positive

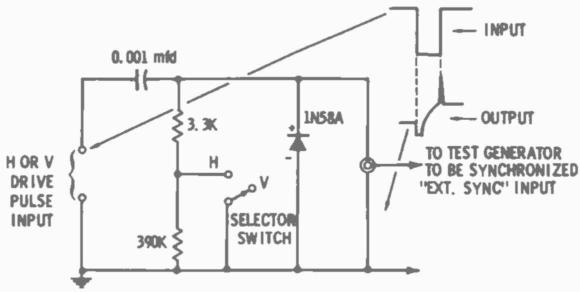


Fig. 5-24. Differentiating circuit designed to obtain positive sync trigger from trailing edge of horizontal or vertical drive.

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 in measurements.

5-5. AMPLITUDE VERSUS FREQUENCY REQUIREMENTS

The amplitude versus frequency response of the television system must be such that the *overall* characteristic (through the transmitter) is reasonably flat over the required passband of 4 mc. It is realized, however, that the pickup device, whether it be an image orthicon, vidicon, or flying-spot scanner, 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 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 affect the gain at middle and low frequencies of the passband.

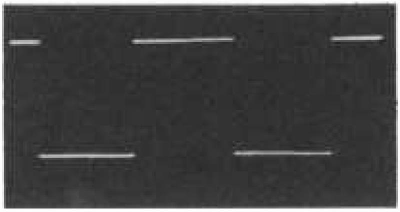
Since each pickup device varies over a limited range in characteristics, each camera chain incorporates the necessary correction circuitry for the pickup device. These special problems will be covered in future notebooks on the image orthicon and vidicon camera-chains. At this time we are concerned primarily with distribution amplifiers, stabilizing amplifiers, microwave relays, 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 "predistort" the signal in an inverse relation to the nonlinearities of the transmitter. This is covered in Section 8.)

A word of caution is in order at this point. Technicians have been known to peak-up video amplifiers to obtain a sharp and crisp reproduction on a master monitor. This seems to be particularly tempting on certain types of video amplifiers employing a single boost or peaking circuit which is adjustable from the front panel. Overpeaking is most apt to occur in stations where personnel are divided permanently between studio and transmitter, and the overall system function is not continually borne in mind. As will be evident in following discussion, the practice of overpeaking to obtain a crisp picture on a studio monitor will sometimes result in a deteriorated picture in good home receivers.

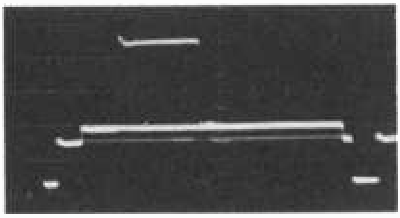
Video monitors themselves may easily be checked for resolution capabilities with a sine-wave generator. The most convenient method is to feed the generator output through a keyer, such as that shown in Fig. 5-22, so that pedestal and sync may be inserted for stable monitor operation. (For a monitor driven by external sync or drive, only blanking pedestal is inserted). As the frequency is increased, the vertical lines on the picture tube become thinner (and fainter) until nothing but raster remains. By noting the maximum frequency at which the lines are just visible, the cutoff frequency of the monitor may be determined as evident by Table 5-1. The effect of brightness and contrast ratios on resolution may also be observed, as well as effect on the comparative resolution capability between various areas of the raster.

5-6. TYPES AND APPLICATIONS OF TEST SIGNALS

Figs. 5-25 through 5-29 illustrate the most popular and accepted types of test signals for television system analysis. The application of each type accompanies the photographs. There is no valid substitution of one for another; each individual type will describe the performance of the TV system over a limited gamut for which it is intended. Due to the nature of their composite signal, the window (Fig. 5-25), keyed burst (Fig. 5-26), and the stairstep (Fig. 5-29) are the most convenient and readily applied sources for overall distribution and transmission tests. However, advantages and disadvantages exist for any one test signal as compared to another.



(A) Square wave—very effective for amplitude and phase response checks and setting tilt controls.

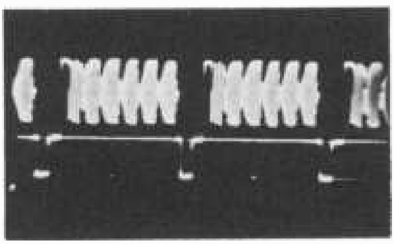


(B) Window—convenient for checking low and middle-frequency response at multiples of line scanning frequency.

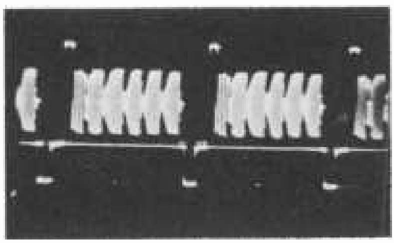
Fig. 5-25. Popular test signals for television analysis.

5-7. USE OF THE SQUARE WAVE

The square wave is a versatile test signal when the user follows the precautions mentioned with respect to the oscilloscope as outlined in Section 1 of this book. When the system includes



(A) As proportional for normal AT&T line checks and studio runs.

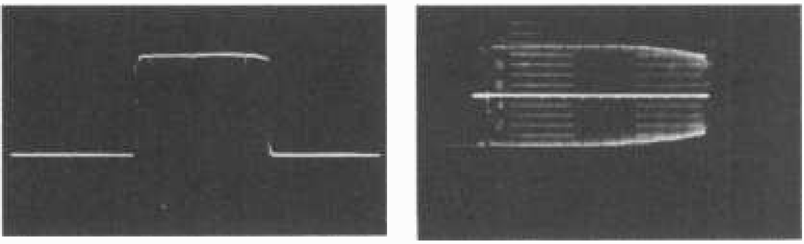


(B) As proportional for transmitter and video-tape recorder runs.

Fig. 5-26. Keyed burst signals normally used as overall transmission check for amplitude versus frequency.

a unit incorporating clampers (such as a stabilizing amplifier or the transmitter), the square wave should be keyed and sync inserted as discussed previously. Fig. 5-30 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 free) picture. Overall tilt (studio to transmitter output) should be held under 2% for complete freedom from visually observable effects. In some cases the stabilizing amplifier will give a flat output with an input tilt of up to 10%.



(A) Detected. (B) Undetected.

Fig. 5-27. Video-sweep test signals widely used for checking amplitude versus frequency response from 100 kc to 10 mc.

When this type of overall distortion is noted, individual units should be checked alone. This waveform at 60 cycles is very useful for setting coupling circuit time-constants (where used), usually designated as *tilt controls*. For example, the RCA TVM-1 series of microwave systems employ such controls in the modulator, monitor amplifier, and receiver amplifier. The square wave

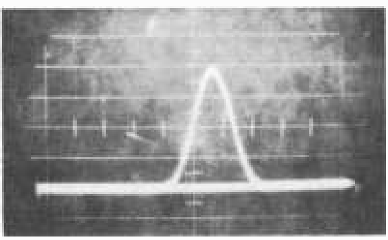


Fig. 5-28. The sine^2 pulse used for phase distortion and transient response checks.

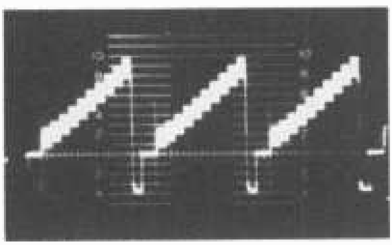


Fig. 5-29. The stairstep test signal (with 3.58 mc on steps) used most widely for checking the amplitude linearity of video systems. May also be used to check differential gain and phase, when a single-frequency sine wave is superimposed on the steps as shown.

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, most "Lap-Dissolve" amplifiers (mixing amplifiers) used with switching systems 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 cycles well within 1 or 2 db of the reference frequency, and yet fail to pass a 60-cycle square wave with less than 20% to 30% or more tilt. Loss of effective coupling circuit time-constants, or clamping failure, will cause this type of distortion.

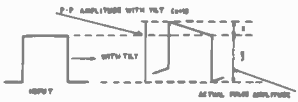
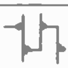
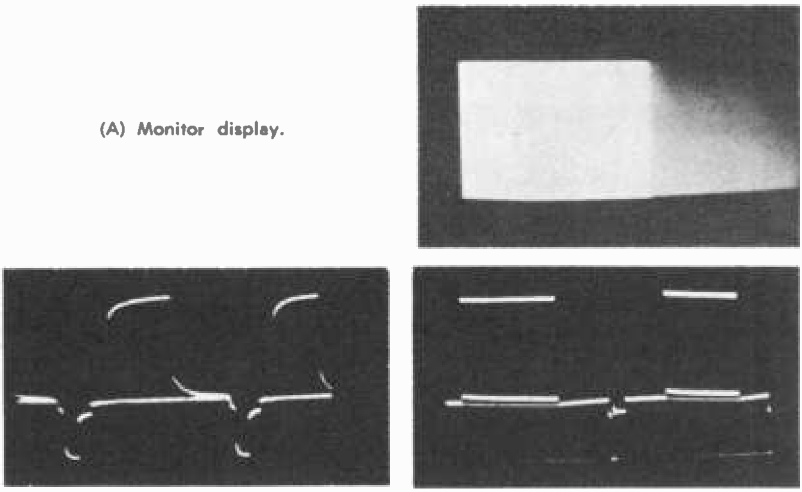
TEST SIGNAL	CRO INDICATION AT SYSTEM OUTPUT	EFFECT ON PICTURE	DEFECT	CAUSE
 60 Cycle		Normal picture.	No defects. Excellent low frequency response. Negligible phase shift.	
		Shading top to bottom of picture. Loss of "setup". (see text).	Loss of low frequency gain with leading low frequency phase shift.	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.
		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).
				% tilt = $\frac{x}{y} \times 100$ Example: $x = 1$ cm, $y = 4$ cm then: $1/4 = 0.25 \times 100 = 25\%$ tilt
 15,750 Cycle (Line Scanning Frequency)		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.
		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).	Mid & low frequency; too high with lagging low frequency phase shift.	Misadjustment of phase correction circuitry or low frequency compensation circuit.
 75 kc		Normal.	No defects. Good high frequency & transient response.	
		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 image.	Excessive high frequency response; non-linear time delay; high frequency cut-off too rapid.	Overpeaking from improper adjustment; plate loads decreased in value; peaking coils or open shunts on coils; poor lead dress.

Fig. 5-30. Use of the square wave.

5-8. USING THE WINDOW SIGNAL AND SINE² PULSE

The window signal is the most effective test signal available for determining absolute values of picture streaking. It is also useful in a more limited sense to evaluate low-to-high frequency-response ratios which, in the final analysis, are a major contributor to the phase distortion that results in streaking. Depending on the rise time of the window signal, an indication of overpeaking, or excessively rapid cutoff may also be revealed.



(A) Monitor display.

(B) Horizontal-rate CRO display.

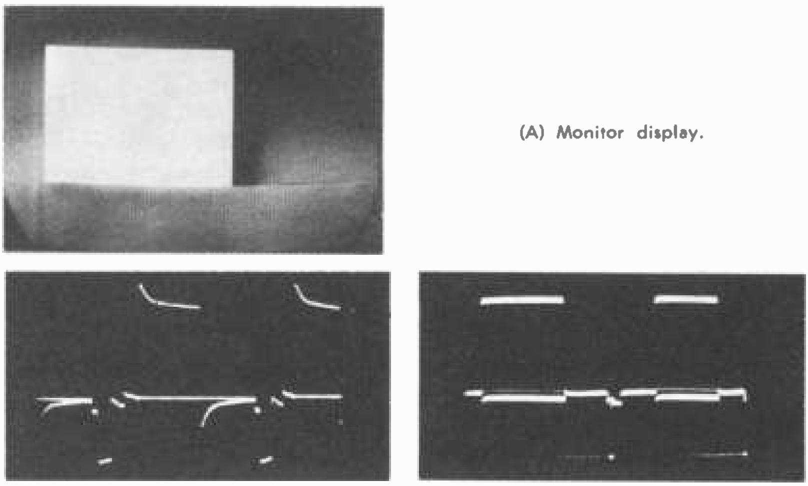
(C) Vertical-rate CRO display.

Fig. 5-31. Examples of positive streaking of a window signal.

Fig. 5-31A is a picture monitor presentation of a window signal with heavy positive streaking. (Polarity is given as positive, since white follows white. If black follows white, the polarity of streaking is termed *negative*.) It is important to understand that the degree of streaking observed on a picture monitor depends not only on the monitor amplifier characteristics but also on ratio of brightness and contrast control settings. There is almost always some amount of visible streaking on a picture monitor when displaying a window signal (or any white bar on black background when the bar extends an appreciable length of a scanning line) and particularly where the picture tubes draw grid current. This was the primary reason for development of the white window: so that a truly accurate measurement could be obtained from the CRO presentation in quantitative terms.

The CRO presentation of the window of Fig. 5-31A at a horizontal rate time base is illustrated in Fig. 5-31B. Note the excessive rate of time required for the white pulse to return to the blanking level. In this extreme case, it never quite makes it. Fig. 5-31C is 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 will cause an increase in setup level, in addition to the streaking effect from the attendent low-frequency phase shift. Such distortion is usually the result of a defective equalizer on long lines, or overcompensation of low-frequency compensation controls or tilt controls.

Fig. 5-32A illustrates a picture monitor display of a form of streaking known as *negative streaking*. Figs. 5-32B and C show



(A) Monitor display.

(B) Horizontal-rate CRO display.

(C) Vertical-rate CRO display.

Fig. 5-32. Examples of negative streaking of a window signal.

the horizontal and vertical rate (respectively) of the CRO presentation. This type of phase distortion is the result of insufficient gain at low frequencies, which may be taken as all scanning frequencies below approximately the tenth harmonic of the 15,750-cps line rate. It will normally be found in practice that the loss of gain is occurring below the first few harmonics, or approximately 50 kilocycles.

Fig. 5-33 is the CRO display of the window signal when the low-frequency response is normal, but overpeaking of the higher frequencies occurs. Such severe edge effects can also result when a video sweep shows a perfectly flat response over the normal video passband, but cuts off sharply immediately above this passband. This is the reason for using the square wave as a final check for video amplifier alignment.

The sine^2 technique essentially involves transmission of a pulse at the line repetition rate with a half-amplitude diameter (abbreviated h.a.d.) equal to the time of either one or two picture elements. It is important to remember here that one TV cycle is equal to two picture elements. This is to say that the pickup-tube scanning-beam sweeping across a vertical black-to-white bar of the image on the photocathode (or target) will produce one cycle of the frequency representing the fineness of transition.

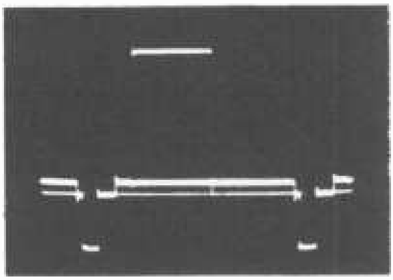


Fig. 5-33. CRO display at horizontal rate showing excessive high-frequency response.

One cycle occurs in a time equal to the reciprocal of its frequency; for example:

$$1 \text{ cycle @ } 4 \text{ mc} = \frac{1}{4\,(10^6)} = 0.250 \text{ microsecond}$$

This says that a black-to-white transition of a vertical bar with a width representing 4 mc 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-mc system is 0.125 microsecond (one alternation of the complete cycle). In the sine^2 technique, a time duration of one picture element is given the symbol T, whereas a time duration of two picture elements (for the system bandwidth under test) is symbolized by 2T. Fig. 5-34 shows the preceding definition in terms of T and system bandwidth. Fig. 5-35 shows the horizontal- and vertical-rate CRO display of the Model TMC 1073-D2 Sine^2-Pulse Window Generator manufactured by Telechrome. The sine^2 pulse appears as a thin, vertical, white line on the left of the raster immediately following blanking, the white window is on the right side, equally spaced vertically on the raster as shown by the field display. This instrument is used frequently for system and line checks. The window signal has a rise time equivalent to that of the sine^2 pulse.

The frequency spectrum of the sine^2 pulse is such that at a frequency where f = 1/T the spectrum amplitude remains at least 35 db under the fundamental. Thus for a 2T-pulse of a 4-mc system (0.250 microsecond) no harmonics beyond 4 mc are present, and the system is checked over the intended frequency

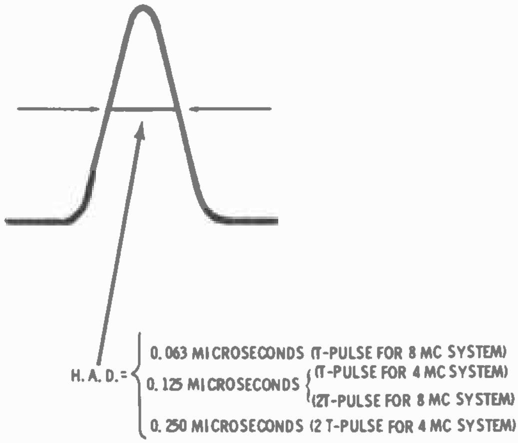


Fig. 5-34. The sine² pulse in terms of T and bandwidth.

range. Conversely, a T-pulse (0.125 microsecond) will contain frequencies to 8 mc and will reveal the characteristics of a 4-mc system (such as the TV transmitter) when being "hit" with the usual 8-mc signals from studio gear. The characteristics of the pulse are always the same and fixed by definition, just as for the VU meter in audio work, and from this standpoint they appear to suggest a step in the right direction for obtaining a "standard" test signal.

The pulse measurement through a system under test is made in terms of the first lobe (negative) and second lobe (positive), by the ratios of the leading- and trailing-edge lobe amplitudes, by

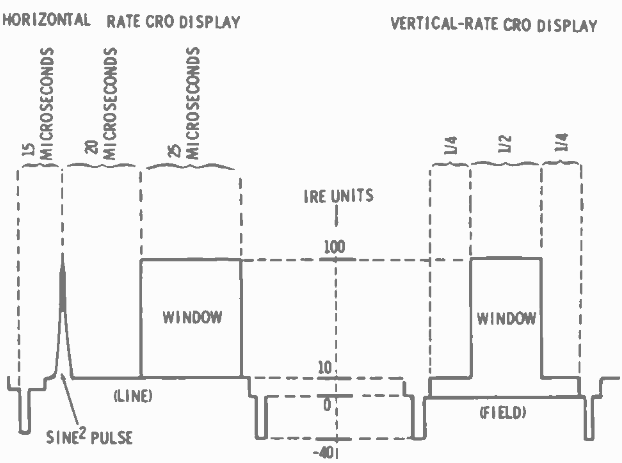


Fig. 5-35. The horizontal and vertical-rate CRO display of the Telechrome Model TMC 1073-D2 Sine² Pulse Window Generator.

the h.a.d., and (with the combination window and pulse) by the relative heights of the pulse and window. (NOTE: Be sure to use the wideband (FLAT) position on the scope.)

Fig. 5-36 illustrates the terminology used above. In general, the T-pulse measurement for a given complete system may be considered satisfactory if the h.a.d. is within 0.18 microsecond, the first (negative) lobe overshoot within 12% and the second (positive) lobe overshoot is within 8%.

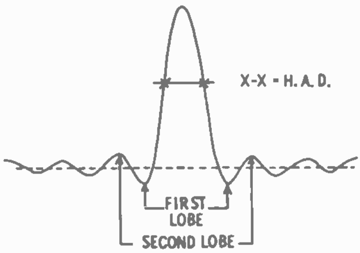
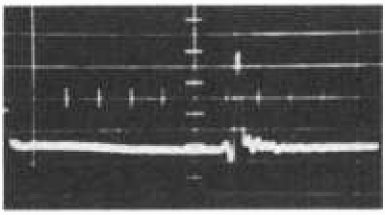
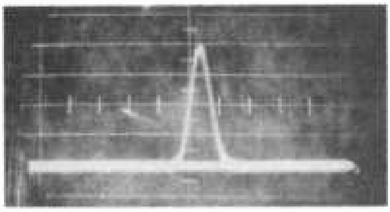


Fig. 5-36. Illustration of terminology used in preceding figures.

An increase in attenuation such as that produced by a sharp cutoff above the desired passband will cause increased phase distortion below the upper limit of the passband. This is indicated by a reduction in T-pulse height, increase in h.a.d., and a large amplitude ring on the right-hand side of the pulse. Fig. 5-37A shows the expanded CRO display of a T-pulse through a 4-mc



(A) T pulse display (high-frequency delay indicated).



(B) 2T pulse display (satisfactory response to 4.0 mc).

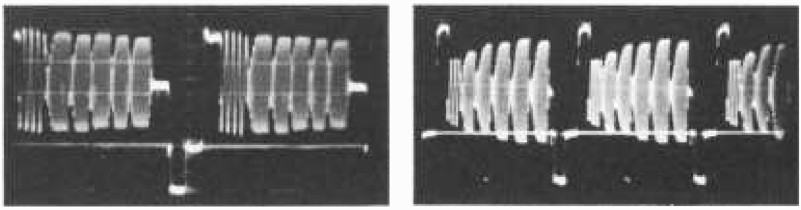
Fig. 5-37. Typical expanded CRO displays when T and 2T pulses are passed through a 4-mc system with a sharp cutoff at approximately 4.5 mc.

system with sharp cutoff at about 4.5 mc. Fig. 5-37B shows a 2T-pulse through the same system. Relative low- and high-frequency gains are also indicated by the window to T-pulse amplitude ratios.

5-9. THE KEYED BURST SIGNAL

The keyed burst signal is the most convenient line or system check for frequency response from 0.5 mc to the upper limit of the system passband. The individual sine-wave bursts should be read peak-to-peak in voltage or IRE units. The setup, of course, will change with attenuation of the burst frequency and should be disregarded in readings.

Fig. 5-38A illustrates a gradual roll-off with increasing frequency as would occur on a long unequalized line. For example,

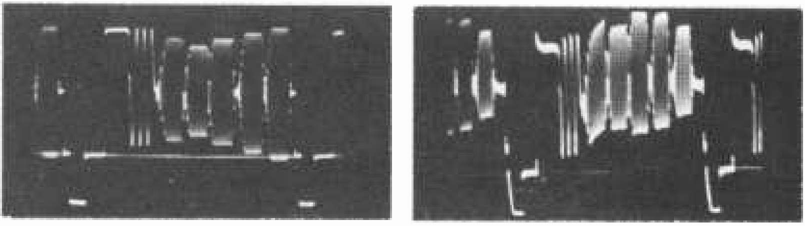


(A) Showing gradual roll-off of highs with increasing frequencies.

(B) Showing increase of gain with frequency.

Fig. 5-38. Keyed-burst signal displays.

the attenuation at 4 mc of RG 11/U cable is 0.4 db per 100 ft. Fig. 5-38B shows the rising response that usually is the result of overpeaking. Fig. 5-39A 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 actually results in a picture much inferior to that obtained from the gradual roll-off in Fig. 5-38A, since middle-frequency "holes" affect pic-



(A) Poor mid-frequency response (hour-glass effect).

(B) Harmonic distortion (shifted axis) due to overload or overpeaking.

Fig. 5-39. Keyed-burst signal displays indicating other types of distortion.

ture resolution to a drastic amount. This effect can also be produced by an open-shield ground on one end of the coax cable transmitting the signal or by faulty terminations. Fig. 5-39B illustrates a shifted axis along the individual bursts resulting from frequency-selective harmonic distortion which can be caused by overloads at selective frequencies or by overpeaking.

An actual loss of high-frequency response or the axis-shift effect is sometimes the result of deliberate overpeaking in an attempt to obtain a sharp picture. But if an off-air monitor where placed side-by-side with a studio monitor displaying the overpeaked or overequalized signal, the modulation effects of the main transmitter, and any studio-to-transmitter links (usually involving either microwave or equalized lines) would be most revealing. Most transmission employing FM for video relay employ low-frequency attenuation circuits to prohibit excessive swings of the carrier frequency at the low video frequencies. This effectively limits the frequency excursion in the microwave receiver IF strip so that differential phase at 3.58 mc (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-cycles with gradually decreasing attenuation to 6 mc. The video is restored in the receiver restoration network. With any such networks, an overpeaked signal with the higher frequencies 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.

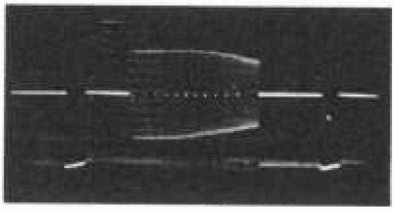
The amount of compression, of course, is also dependent on the peak-to-peak video level used at the modulator to obtain the 100% reference modulation. When this is held within the design limits of a particular system, the degree of compression from an overpeaked signal can be quite small. In this case, the major cause of severe edge effects is the ringing occurring at the main transmitter low-pass filter that employs a rapid cutoff above 4.18 mc. It is also known that when high-frequency energy is appreciable (as is the case with sine-waves or keyed video sweeps), vacuum-tube circuits can exhibit considerable overloading at these frequencies while passing lower frequencies at normal gains.

5-10. KEYED VIDEO SWEEP

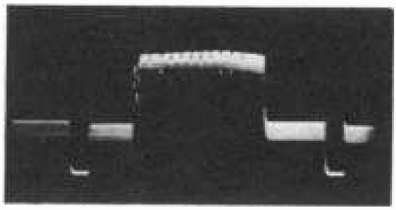
Complete systems may be checked with keyed video sweep without removal and modification of clamper circuits as is required with straight video sweep. A schematic diagram of a simple keyer was shown previously in Fig. 5-22. Precautions in setup should be taken as shown by Fig. 5-40A, which is the wide-band CRO display through the keyer mentioned before. Adjust the video gain and blanking gain so that the swept 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. 5-40B shows the signal of Fig. 5-40A after detection. Video sweep technique and interpretation is the same as that used for the keyed burst, except that a detector probe can be used to eliminate any effects of oscilloscope high-frequency roll-off.

NOTE: If, when using the keyer, 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.

All of the precautions on CRO calibration and use given in Section 1 should be scrupulously followed in all frequency and transient response measurements.



(A) On wideband scope.



(B) Signal of "A" detected.

Fig. 5-40 Keyed video sweep displays.

5-11. VIDEO SWEEP TECHNIQUES

The most efficient procedure in using the video sweep is to employ keyed sweep for overall system checks, and normal video sweep (without blanking and sync) if it is necessary to service individual amplifiers not incorporating clamping circuits. (Clamping circuit operation is covered in Section 5-12.)

For the keyed-type video sweep, insert blanking only 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 (unkeyed) for amplifiers or systems 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 whose output is beat with a sweep-oscillator frequency modulated at 60 cps. The frequency modulation is such as to cause the beat-frequency to swing over a usable range of about 100 kc to approximately 10 or 20 mc. The frequency swing may be produced by a conventional reactance-tube circuit with 60-cps excitation from the power line or, in many cases, by a motor-drive capacitor in the oscillator tank circuit. A 3,600-rpm motor provides a 60-cps 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 over the usable range.

The fundamentals of checking the high-frequency characteristics of video amplifiers are illustrated in Fig. 5-41. The frequency-modulated output of the sweep generator consists of a video envelope containing a sweep at uniform rate and amplitude over a range of 100 kc to 10 or 20 mc, with a tunable frequency marker (notch) placed at any desired frequency. The wave is repeated at regular intervals at 60 cps. This test signal is applied to the amplifier to be tested. A detector of the type shown in Fig. 5-41 is connected to the output of the amplifier. This rectifies the signal output as shown (in this case the amplifier is considered as a theoretical ideal: no distortion has occurred), and the output of the detector is fed to the vertical input of the oscilloscope. The scope for this test should have excellent low-frequency response so that no distortion of the 60-cps square wave takes place. High-frequency response need be no greater than 50 kc. By this means the oscilloscope traces a graph of output voltage vs frequency over the passband above 100 kc.

It is very important not to overload the amplifier(s) when using video sweep. If the normal output is a 1-volt (p-p) signal, feed just enough input sweep level to result in 0.5 volt (p-p) amplifier output. Remember to calibrate your detector probe (as outlined in Section 1) so you will know how much loss occurs in the probe. For example, if the detector probe gain is 50%, a 0.5-volt (p-p) actual output level reads 0.25 volt (p-p) through the probe to the oscilloscope screen.

All modern video sweep generators for broadcast service employ an internal output-impedance (sending-end impedance) of 75 ohms. When feeding the input of the system or individual amplifier, the proper 75-ohm termination should be used. When aligning response circuits, 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.

When 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. 5-42 illustrates the general technique used in the alignment procedures:

1. Connect the sweep generator to point 1. Adjust L4 for the flattest response.
2. Note that in stage V3 it is necessary to temporarily bypass the small capacitor with one of approximately 0.25-mfd.

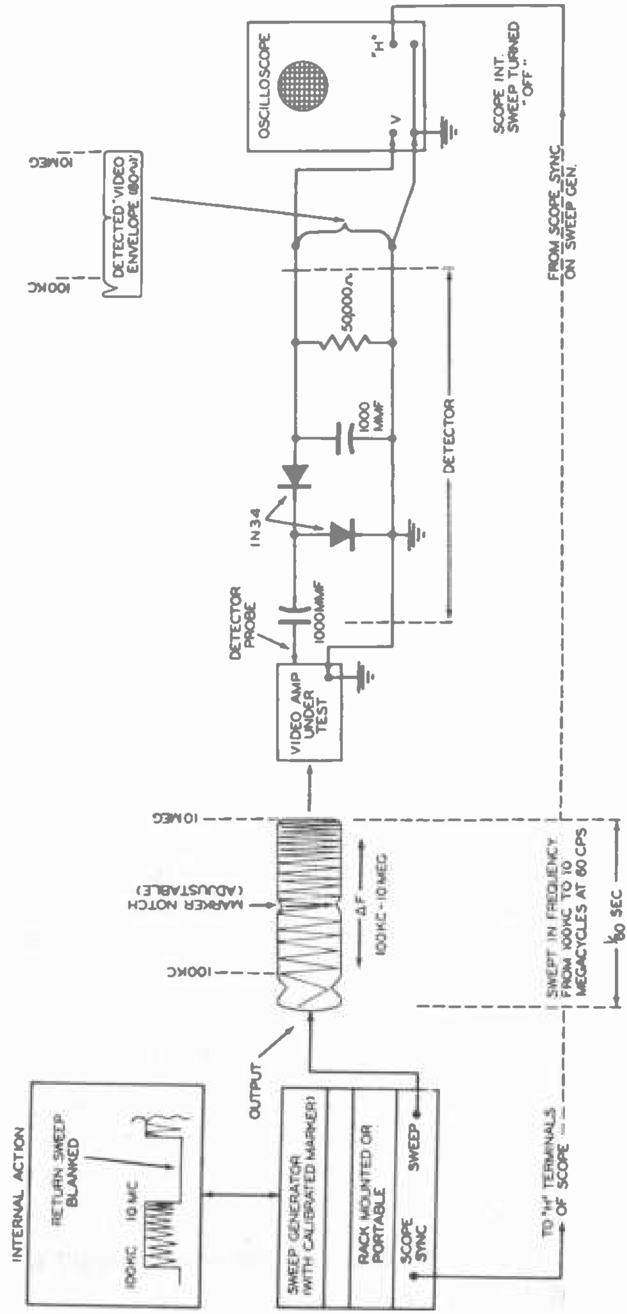


Fig. 5-41. Basic method of testing video amplifiers for high-frequency characteristics.

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-L3 for flatest overall response.

3. Remove the temporary bypass. Note that the grid of V2 has a DC value for 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-mfd 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 equipment instruction manuals in sweep procedures when given and available.

It is assumed that in all following examples of troubles the back-to-back response of the sweep generator and scope (detected) is at least as good as curve 1 in Fig. 5-43.

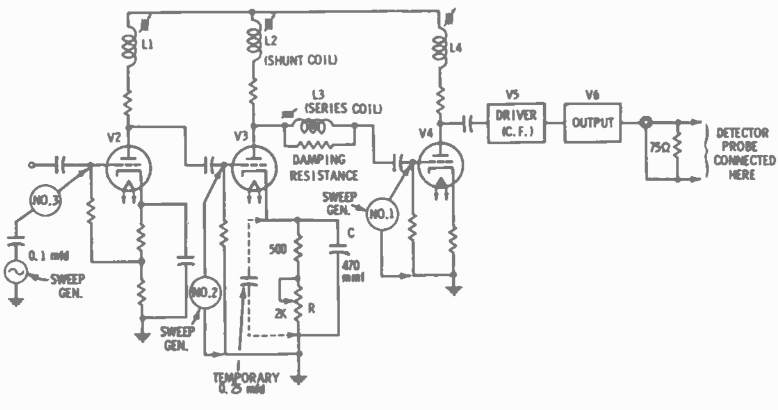


Fig. 5-42. General outline of video amplifier alignment.

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 mc, the response is flat from 100 kc to 8 mc. A minimum of 8 mc is considered necessary in broadcast station equipment to ensure minimum frequency and phase distortion, although a flat response to 5 mc is tolerated when necessary. In many instances, most modern commercial equipment tests well over 8 mc.

If plate-load resistor R_L should increase from the normal value to a higher resistance, the trace obtained would appear similar to curve 2 in Fig. 5-43. From the curves shown previously in Fig. 5-5 it was noted 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 kc to 8 mc, 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 pictures. Although this effect might be caused by an actual change

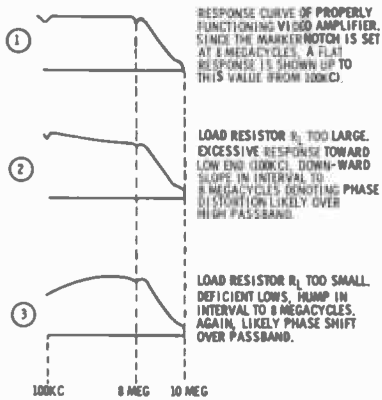


Fig. 5-43. Effect of R_L on demodulated sweep-generator test signal.

in the value of R_L, this is not necessarily the sole cause. Anything that would affect the dynamic plate load so as to increase its effective load impedance over the passband would have the same results. For example, observe curve 9 in Fig. 5-46. This is essentially the same trace as curve 2, and it is caused by a reduced value of the inductance of the shunt peaking coil. Since this coil

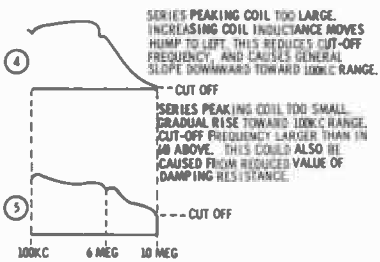


Fig. 5-44. Effect of series-peaking coil on response curve.

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 going further.

When engineers design a shunt peaking-coil, its reactance at the highest pass frequency is made to equal approximately one-half the value of the reactance of the total output and input shunt capacitance whose effect the peaking coil must nullify.

The total plate-load impedance is a complex quantity rather than pure resistance, and takes the conventional value:

$$Z_L = \sqrt{R^2 + (X_L - X_C)^2}$$

Assume just as an illustrative example that the total shunt capacitive reactance at the highest pass frequency is 2,000 ohms. With X_L equal to one-half this value, or 1,000 ohms (with a 1,000-ohm R_L):

$$\begin{aligned} X_L &= \sqrt{(1000)^2 + (1000 - 2000)^2} \\ &= \sqrt{10^6 + 10^6} \\ &= 1{,}415 \text{ ohms (approx.)} \end{aligned}$$

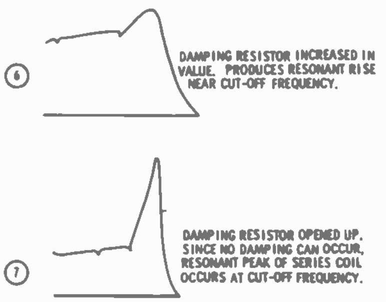


Fig. 5-45. Effect of damping resistor on response curve.

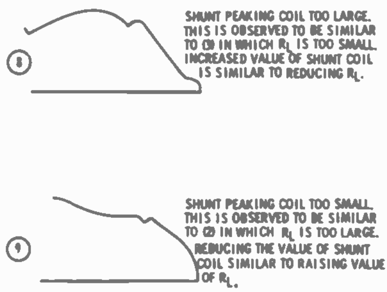


Fig. 5-46. Effect of shunt-peaking coil on response curve.

Thus with an R_L of 1,000 ohms, and normal shunt peaking-coil reactance of one half the value of total X_C at the highest frequency, the total plate impedance at this highest frequency is 1,415 ohms.

Now consider what occurs should the inductive reactance be reduced by one-half its normal value. Obviously the difference in the quantity enclosed in the reactive parenthesis of the equation becomes larger, since X_C remains the same. To be more explicit:

$$\begin{aligned} Z_L &= \sqrt{1{,}000^2 + (500 - 2{,}000)^2} \\ &= \sqrt{10^6 + (2.25 \times 10^6)} \\ &= 1{,}802 \text{ ohms (approx.)} \end{aligned}$$

Observe that Z_L has increased with a decrease in shunt X_L.

Understanding of these basic circuit relationship materially aids the maintenance engineer in interpreting resultant scope traces. Curve 3 of Fig. 5-43 is a typical trace when R_L has decreased from its normal value. Observation of curve 8 in Fig. 5-46, in which L2 is larger than the optimum value, reveals the effect on plate-load impedance at the higher frequencies, effectively decreasing the value of R_L. The traces are therefore similar.

The effect of the series peaking-coil is shown in curves 4 and 5 of Fig. 5-44. When the series coil is larger than optimum value (curve 4), a gradual upward slope occurs from 100 kc (low end of sweep) toward the mid-range of sweep. The larger the value of inductance, the farther the hump is shifted to the left. Compare this with curve 3 in Fig. 5-43, which indicates R_L is too low in value. The major difference in the resulting traces is the extremely reduced cutoff point (maximum downward slope of curve) indicated in curve 4, with the series coil too large. This causes the notch to "slide down" on the sloping portion of the curve at the high end. From the slope of response curve 4, it may rightly be inferred that the effective R_L is reduced as the series peaking-coil is increased in value, just as in the case of the shunt peaker. Increasing the value of the series coil lowers the resonant frequency (greater LC ratio) at which the series peaker performs. This causes the hump in amplitude response to move to the left (lower in frequency), and the effective Z_L at the highest frequencies in the desired passband is reduced.

If the series peaker is too small in value (curve 5 in Fig. 5-44), the response from 100 kc to mid-range of the sweep is too large, and the cutoff frequency is larger. This condition may also be caused by a reduced value of damping resistor shunted across the series coil.

The effects of increased values of damping resistance is noted in curves 6 and 7 of Fig. 5-45. Curve 6 is displayed when the resistance value has increased to the point where inadequate damping of the resonance peak occurs. This is one possible symptom of transient oscillation. Curve 7 indicates an open damping resistance that allows the resonant peak to appear.

The peaking circuits of most video amplifiers are adjustable, as indicated by the variable inductances in the schematic of Fig. 5-42. The proper alignment of these stages constitutes an important function of the maintenance engineer on both initial set-up of the amplifiers and in routine and priority checks of equipment. With the sweep generator connected at point 2 in Fig. 5-42, the effects of varying the adjustments of L2 and L3 may be noted. It will be observed that varying series peaker L_3 mostly affects the trace at the right of the pattern, and varying L2 will mostly affect 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 mc. 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. L2 is similarly varied while observing the scope pattern, and is adjusted to obtain the ideal response characteristic (curve 1 in Fig. 5-43). No more than approximately 2% variation should occur over this high-pass band from 100 kc to the limit of the flat portion of the curve. Always compare results with manufacturers specifications.

The typical traces shown in Figs. 5-43 through 5-46 assume only one component error, as is usually the case in preventive maintenance or in trouble occurring during operation. If a number of amplitude variations show up in the pattern, several errors 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 very 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 normal and abnormal appearance of any slight spurious trace.

If a great number of pronounced humps or wiggles occur on the scope screen, chances are there 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.

5-12. THE STABILIZING AMPLIFIER

Modern stabilizing amplifiers greatly influence the overall frequency and transient response by built-in, signal-processing circuitry.

The primary function of a stabilizing amplifier is to correct any fault existing in the video signal. Such faults may be hum, switching surges, noise, or sync tip modulation. It may also be used to insert sync pulses into the composite signal, or in some instances it is used to separate sync from the composite signal for genlock purposes.

Stabilizing amplifiers may be found in the control room as the first amplifier for incoming network programs, incoming remote relay position, and following the switcher unit that was described in the preceding section. This type of amplifier is also found at the transmitter room, being used as the first amplifier from the coaxial cable connecting the studio or from the studio-to-transmitter RF relay receiver. The unit contains video amplifiers, sync separators, sync insertion circuits, keyers, shapers, and clamping circuits.

The first two or three stages usually provide linear video amplification and a means of inserting the composite (horizontal and vertical) sync pulses. Such insertion is accomplished by means of two input tubes with a common plate or cathode load.

The combined video and blanking signal is fed to one tube, and the composite sync is fed to the other tube. The signals are then mixed in the common load. This stage is then followed by one or two linear video and sync amplifiers.

At this time a phenomenon known as *sync stretching* takes place. Fig. 5-47 illustrates the electrical function of a sync-stretcher circuit. This type of circuit might employ both a Class-A stage and Class-C stage. Over the normal video-signal range up to the blanking level (points A to B on the total plate curve), the amplification is linear. At the blanking level, the Class-C stage also begins conduction, adding to the total signal from blanking

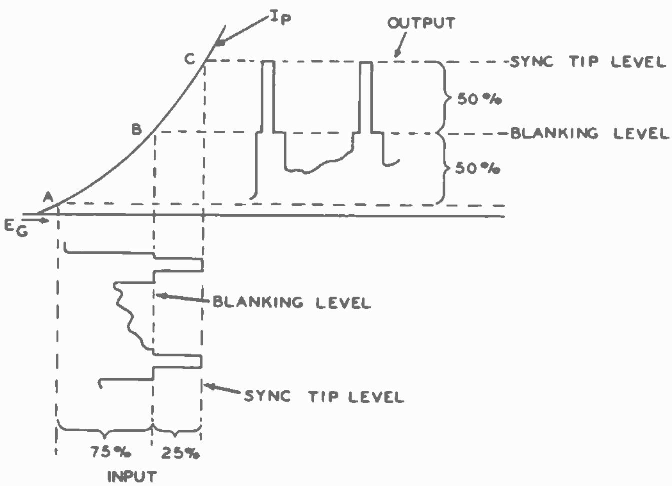


Fig. 5-47. Basic action of sync stretcher.

level to sync tips (points B to C). In this manner, a composite signal input of 25% sync and 75% video may have its output sync-region expanded as shown. There are several reasons why this function is desirable. For example, if the sync tips should become modulated by any stray noise pickup, sync stretching allows a following clipper stage to remove the modulated tips. Also, in the case of incoming network lines or remote signals of any source, the sync region is apt to become compressed, thus requiring stretching to restore the proper sync-to-signal ratio.

The blanking level at point B is held at this constant level by a clamping circuit that functions independently of the picture amplifier. The clamp circuit clamps the peak of each blanking pulse at the correct point on the amplifier curve and eliminates spurious low-frequency components from the signal.

The resulting amplifier composite signal is then fed to a clipper stage. Here the sync is clipped by an amount determined by grid-bias adjustment or whatever type of clipping circuit is used in this stage. The proper sync-to-signal ratio is thus restored after the faults have been removed. The grid circuit of this clipper is also usually clamped at blanking level so that the predetermined sync amplitude is independent of variations in the average video signal amplitude. Thus, in this stage also, spurious signals are eliminated from the video and blanking portions of the composite signal.

The output of a stabilizing amplifier consists of two stages: one to feed the line or transmitter and the other to feed a monitor bus.

The sync-stretcher stage is independent of the video signal in the amplifier. This means that 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

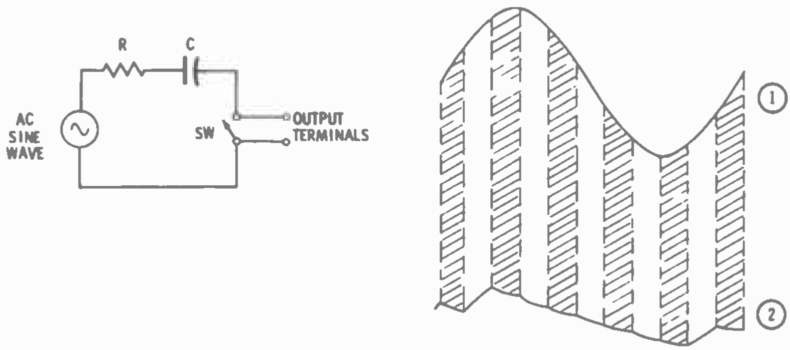


Fig. 5-48. Clamp circuit theory for sine waves.

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 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 compression 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-cps AC from the amplifier response.

How an electronic device may eliminate one low-frequency signal and improve the response of another low-frequency signal comprises the study of the clamping circuit which, due to the nature of the TV waveform itself, functions at high efficiency. Consider first the drawing of Fig. 5-48. This is an AC signal source with a series RC circuit and a switch. Curve 1 in this figure is the signal voltage that appears at the output terminals if 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. Curve 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 described: opening and closing a switch at the horizontal-line frequency (15,750 times per second), so that any 60-cps sine wave (such as would occur from a stray field) is greatly attenuated.

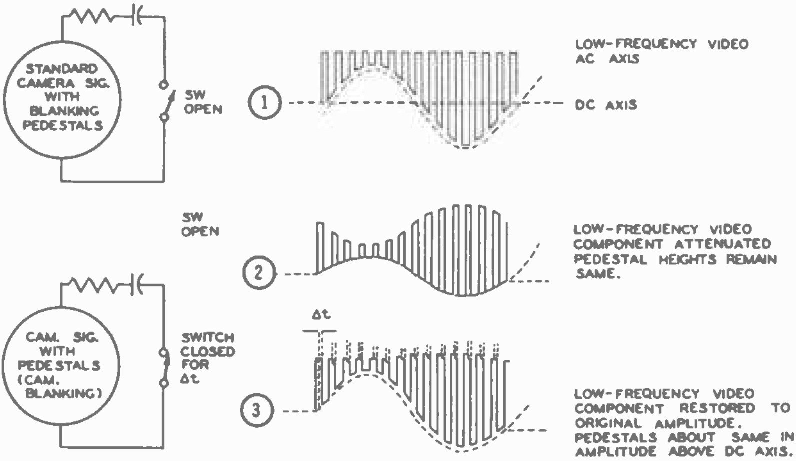


Fig. 5-49. Clamp circuit action for waves broken up by pedestals at a fixed level.

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. Drawing 1 in Fig. 5-49 shows the output waveform with switch SW open, showing that the same waveform appears as that applied. Assume at this time that some circuit action occurs which results in poor low-frequency response. Drawing 2 in Fig. 5-49 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. Thus if the switch is closed for the Δt intervals shown, the output waveform will appear as in drawing 3 (Fig. 5-49), 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 upon 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.

The effectiveness of line-to-line clamping at 60 cps may be evaluated by the following analysis:

It takes 1/120 second for hum signal to pass through the amplitude extremes (see Fig. 5-50).

$$1/120 \text{ second} \cong 131 \text{ lines.}$$

Amplitude change at peaks of sine-wave hum pickup is negligible. Therefore, where the slope of the hum signal is maximum, the amplitude change during a single line may be considered as: $1/131 \cong 0.00764$, or 0.764%.

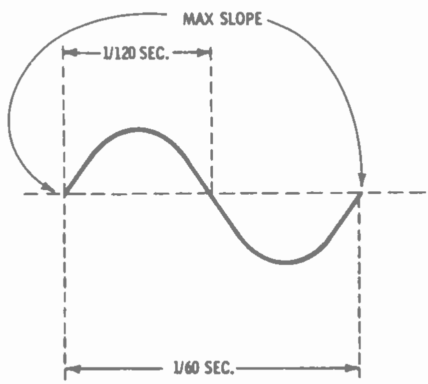


Fig. 5-50. Time-amplitude relationship for 60-cycle hum.

Since the signal passes through maximum slope twice in this time:

$$2 \times 0.764 = 1.528, \text{ or approximately } 1.5\%.$$

1.5% is 38-db attenuation of the 60-cps hum component.

Effectiveness at higher frequencies of sine-wave pickup decreases until complete lack of effect is noticed at approximately 2 kilocycles.

It may be seen that for 10 times the 60-cps frequency (600 cps) the wave passes through maximum slopes at a rate 10 times greater, or $10 \times 1.5\% = 15\%$. This is only approximately 16.5-db attenuation. Table 5-2 lists the effectiveness of the clamper at frequencies between 60 and 2,400 cps.

Table 5-2. Effectiveness of Clamper at Frequencies Between 60 and 2,400 cps

Frequency (cps)	db attenuation (approx.)
60	38
120	31
300	22.5
600	16.5
900	13
1200	10.5
1500	8.5
1800	6.9
2100	5.6
2400	2.2

Either the stabilizing amplifier or some other amplifier in the system employs circuits where blanking and sync signals are injected and where clamping is employed. These stages require a special technique in testing procedure. Fig. 5-51 presents a simplified diagram of a sync insertion circuit, feeding a cathode-follower output stage whose grid is clamped. This might be a typical arrangement of a mixer-amplifier unit following the switcher stage or of a line output amplifier in which sync insertion takes place. Point 1 in the clamper stage, is our first consideration. From the inherent nature of clamping tubes, considerable capacity is added to the circuit at that point. Therefore, these tubes cannot be removed without upsetting circuit constants, which would seriously affect normal operation of the output stage. Neither can they be left in unless the keyed test signal is employed as described earlier.

This is true because the resultant clamping pulses would give spurious response in the output, since the unkeyed sweep gen-

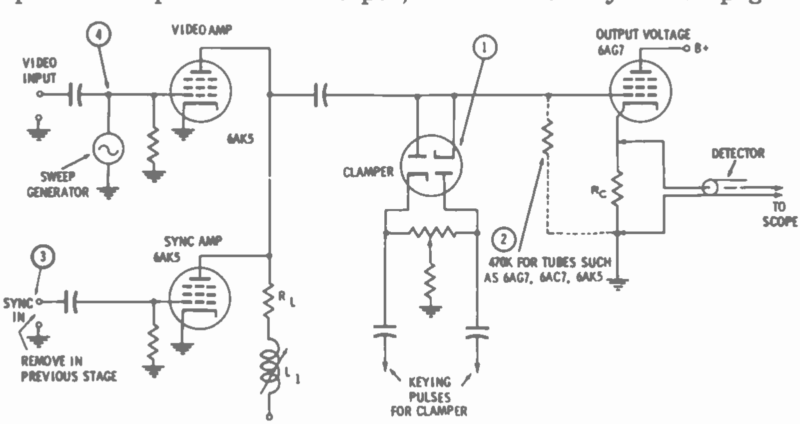


Fig. 5-51. Typical sync injection and clamping circuit.

erator contains no blanking pulses upon which the clamper normally operates. For this reason it is necessary to replace the clamper with the same type tube, but with the heater circuit opened by cutting off the heater pins. These "dummy" tubes should be plainly marked in some fashion such as with paint or finger-nail polish so that they will not inadvertantly be left in place after testing. When the clamp is immobilized in this way, the grid of the output stage is left "floating." It is then necessary to temporarily insert a grid resistor at point 2 in the circuit (Fig. 5-51). A value of 470,000 ohms is proper for most tubes.

The sweep generator may be connected at point 4 for the purpose of checking this stage and aligning shunt peaking-coil L_1. In this stage, another condition must also be considered. Sync pulses are usually inserted (as shown) by an amplifier sharing a common plate load with the video amplifier. Aside from the capacitive effects of the sync amplifier on the video amplifier, the video amplifier steady-state plate voltage is dependent on the load current drawn by the sync amplier stage. Sync pulses injected into the video amplifier, however, must not be allowed, since the resultant patterns would be meaningless for the purpose of this test. To maintain normal circuit conditions, the sync amplifier tube obviously cannot be removed. If the amplifier simply has a plug-in connection for the composite sync signal, this may be removed during the test. If the amplifier is rack-mounted and receives sync from a distribution bus common to a number of amplifiers, the signal should be "killed" in the stage prior to point 3 by use of a dummy tube. Should point 3 obtain its drive directly from a sync distribution bus, it is necessary to break the connection temporarily at this point.

A very common form of clamping at both studio and transmitters is known as *back-porch clamping*, in which the clamped point is held at the blanking level of the signal. This is done by keying the clampers from a sync pulse of the line waveform in such a manner as to cause the clamping tubes to refer to the reference level which immediately follows, this being the blanking or pedestal voltage upon which the sync pulse is constructed. Fig. 5-52 is a simplified schematic of one of the clamping circuits in the General Electric stabilizing amplifier. Tube V18 is a sync-clipper stage which has removed the sync pulses from the composite picture signal. This amplified sync is fed to the grid of V19A, which has its grid resistor returned to the well regulated B+ supply of 275 volts. The coupling capacitor is very small (47 mmf), so that the resulting grid current of V19A charges the coupling capacitor during the rise of each sync signal. The output of the sync amplifier at point 1 is positive. At the trailing edge, during the negative going excursion, a large negative overshoot occurs, which drives the grid of V19A momentarily below cutoff (point 2). Between sync pulses the tube returns to zero bias by charging toward B+ through the grid resistor. This condition prevails until the next sync signal occurs. During this interval, a positive pulse is formed at the plate of V19A. From the negative overshoot at point 2, it is seen that the positive pulse occurring at the plate of this stage is coincident with the trailing edge of the sync pulse (see the waveform at point 3). This pulse is amplified, inverted and clipped by V19B (see waveform at

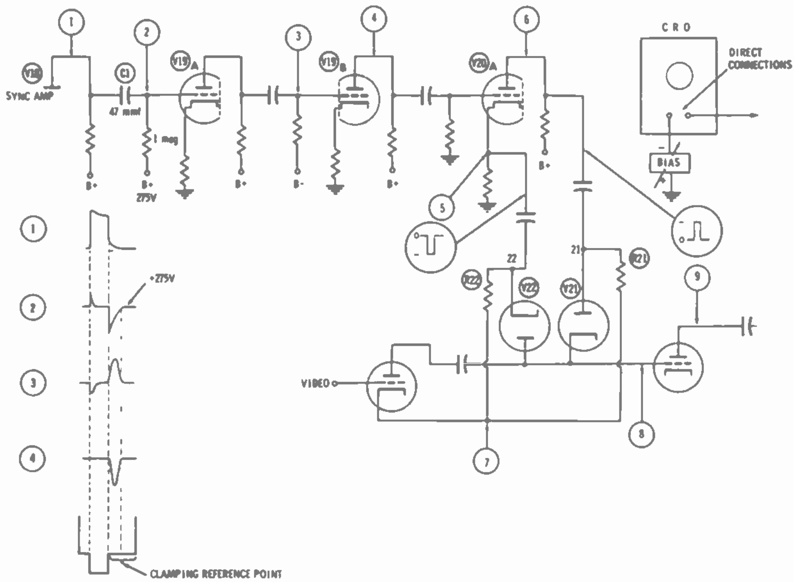


Fig. 5-52. Simplified schematic of one of the clamping circuits in a stabilizing amplifier.

point 4), and used to drive clamp keying-tube V20A. Since the pulse applied to V20A is negative, the signal at the plate of this stage (point 6) is positive, while that at the cathode (point 5) is negative. These pulses are applied simultaneously to clamp diodes V21 and V22.

The function of these diodes is to short the grid at point 8 momentarily at the line frequency of 15,750 cps. The keying pulses at point 21 are positive, while the simultaneous keying pulse at point 22 is negative. At the instant these pulses arrive, if point 8 is the same potential as 7, current will flow through the diodes and resistors R21 and R22, since the condition for conduction (plus charge on plate, minus charge on cathode) has occurred. Since the diodes and resistors are of equal values in the two branches, equal voltage drops occur. The potential at point 7, is therefore midway between that at point 21 and 22, and the grid potential at point 8 remains equal to the potential at point 7.

If, when the clamping pulses are applied, point 8 is at a higher potential than point 7, V22 will conduct while V21 is cutoff. This drains the excess potential from point 8. When point 8 again becomes equal to 7, the action becomes that described previously. Conversely, if point 8 is at a lower potential than point 7 when clamp pulses occur, V21 conducts and restores the grid potential to that at point 7.

In checking the action of a clamper, since the important characteristic is the DC component, the oscilloscope must use DC amplifiers or the direct connection to the CRO. The most convenient point of checking overall action is point 9, which is the plate circuit of the stage whose grid is clamped. It is necessary to center the sweep beam (with sync and blanking signal applied as in a composite all-black signal), and if the plate potential deflects the CRO sweep off the screen and the beam cannot be centered, it is necessary to use an external bias as shown. This, of course, removes the scope chassis from ground and caution must be exercised in its use to avoid grounding contact. Only enough external positioning bias is needed to sufficiently neutralize the static potential so that the centering control in the scope is effective.

With the beam thus centered, and with blanking and sync pulses at their respective reference lines for the all-black signal, a mixed black and white video-signal is applied and the scope sweep-frequency is adjusted to display the line frequency of 15,750 cps. The blanking level should occur at the same center position on the screen, provided the clamper is properly working. If clamping pulses are out of sync, the video will be hopelessly chopped to pieces by the random clamping pulses. If the clamp is completely inoperative, the grid at point 8 is effectively shorted, and no signal occurs. Emergency operation such as might be necessary in the field could be carried on by removing the clamper tubes and inserting a fixed bias at point 8. Obviously this condition would require constant attention to all picture controls and overall performance is poor.

Faulty clamping action may be checked back through the clamp keying tube circuits and clamp pulse-forming circuits by following the points in Fig. 5-52. The vertical amplifier of the scope is used with sweep set to display the line frequency of 15,750 cps. Tubes are always suspected first, and it is advisable to change all tubes in this chain to ascertain their effect on the clamping action. If this does not remove the trouble, the scope should be connected at the points shown until the first serious deviation from the indicated waveforms occurs. The resistors and capacitors for that circuit, and in some cases in the preceding or succeeding stage, are checked point to point for component failure by conventional methods.

The color signal differs from a monochrome signal in two major respects, both of which pose problems. 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 subcarrier 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 infrablack excursions are not clipped off.
2. *Burst distortion*—To insure 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.

These two problems, subcarrier clipping and burst distortion, are avoided in the RCA TA-9 stabilizing 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 simplified block shown in Fig. 5-53 illustrates the major circuit features of the TA-9 stabilizing amplifier. Note that the composite picture signal traverses three paths—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 insure stable and accurate clipping over a wide range of signal level variations.

A noise 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 cir-

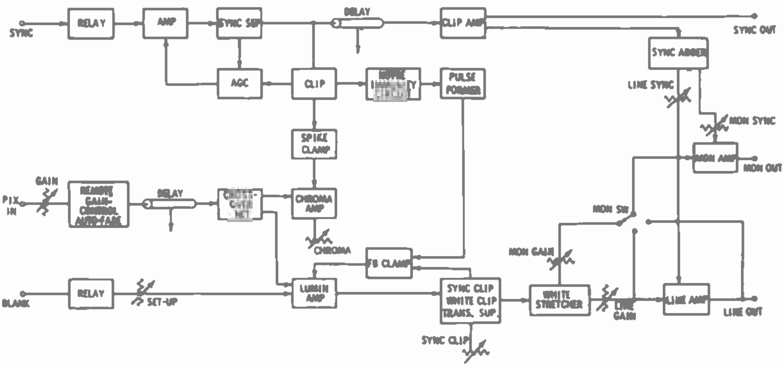


Fig. 5-53. Simplified block diagram of the RCA Model TA-9 stabilizing amplifier.

cuit 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 greatly attenuate the narrow noise pulses, so that the sync pulses can trigger the pulse former.

In the picture channel the signal is again split into two paths, one carrying mid-band frequencies or chrominance information and the other carrying the luminance information. The crossover occurs at the color subcarrier frequency, 3.58 mc, 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: to maintain clipping at exact black level over long periods of time without readjustment; to automatically set the clipped signal at the proper position on the white stretcher characteristic; and 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 damaging 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 intercarrier buzz in receivers 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 is usually 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 for predistortion of the transmitter, when necessary, as described in Section 8.

SECTION 6

AMPLITUDE AND PHASE LINEARITY

From preceding sections it is apparent that amplitude and phase linearity is important to monochrome television. For black and white pictures, the amplitude-phase relationship has been shown to depend on frequency response and coupling-circuit time constants as well as clamping processes to obtain correct black level reference.

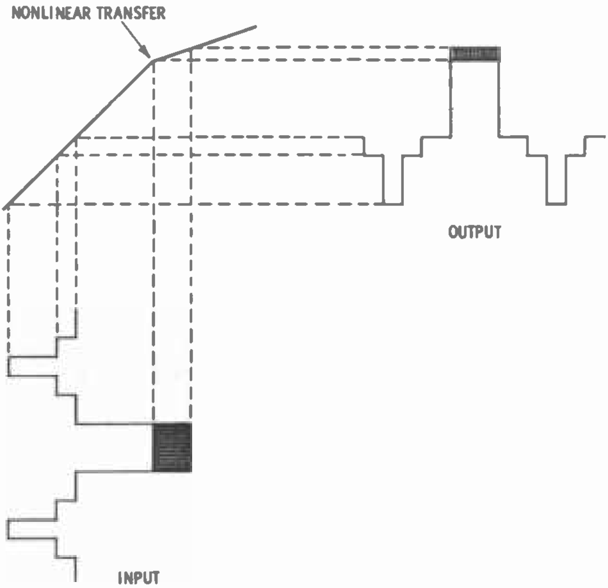
Amplitude and phase linearity is doubly important to the transmission of color pictures due to the added chrominance-carrier sidebands around the 3.58-mc region. Since most stations (even those currently involved only with monochrome) maintain the station facilities within the color specifications, it is important to be aware of the modern techniques in testing and maintenance.

6-1. DEFINITION OF TERMS

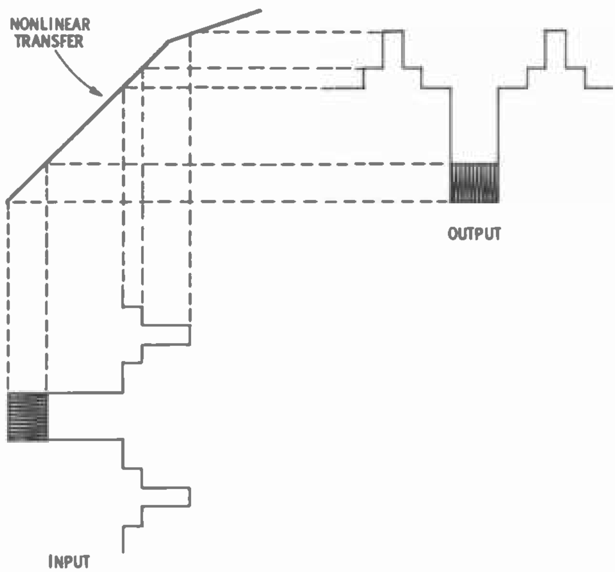
Incremental Gain Distortion

This is the basic cause of amplitude and phase nonlinearity (Fig. 6-1). 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 compression may occur in either the white or sync direction, depending on the input signal polarity.

The shaded area of the signals in Fig. 6-1 represent an RF component, such as the color subcarrier in the region of 3.58 mc. If the tip of the white pulse in Fig. 6-1A 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:



(A) Compression of RF component in the white direction (no sync compression).



(B) Sync compression (no compression of RF component in white compression).

Fig. 6-1. Whether compression occurs at white level or sync level depends on polarity of input signal at the nonlinear stage.

(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 will exist (defined below) which actually shifts 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 green 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 video transmitters), both white and black is compressed, and grays are stretched.

In Fig. 6-1B 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 described in Section 5; and incremental gain distortion, resulting in a nonlinear transfer.

Differential Gain

This 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.

Note 1: Differential gain may be expressed as a percentage by multiplying the above difference by 100.
Note 2: Differential gain may be expressed in db by multiplying the common logarithm of the ratio described in (a) above by 20.
Note 3: In this definition, level means a specified position on an amplitude scale applied to a signal waveform.
Note 4: The low- and high-frequency signals must be specified.

The preceding definition is that of the IRE subcommittee 23.4 and is 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 cps; the high-frequency superimposed component is a sine wave at the color subcarrier frequency of 3.579545 mc, which, for simplicity, we refer to as 3.58 mc.

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

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.

Note 1: Notes 3 and 4 that were applied to differential gain also apply to differential phase.

The above definition is that of the IRE subcommittee 23.4 as approved by the ASA.

Differential phase is measured in degrees using the same type of test signal as for differential gain. Again, in this case the term distortion is redundant, since any amount of differential phase is distortion.

Summary of Definitions

Incremental gain distortion will result in both differential gain and phase. When the color signal (the high-frequency component) is clipped or compressed, saturation errors in the colors results. Even though the gain at 3.58 mc 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. Differential phase at 3.58 mc for a monochrome picture is not important, except that the measurement is an excellent criterion of overall monochrome system performance.

6-2. TYPES OF TEST SIGNALS AND MEASURING GEAR

Test signals must be able to convey information at low and high frequencies over every possible pictorial value likely to be encountered. This pictorial 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) as:

$$\text{Duty cycle} = (\text{Pulse Duration}) \times (\text{PRF})$$

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.

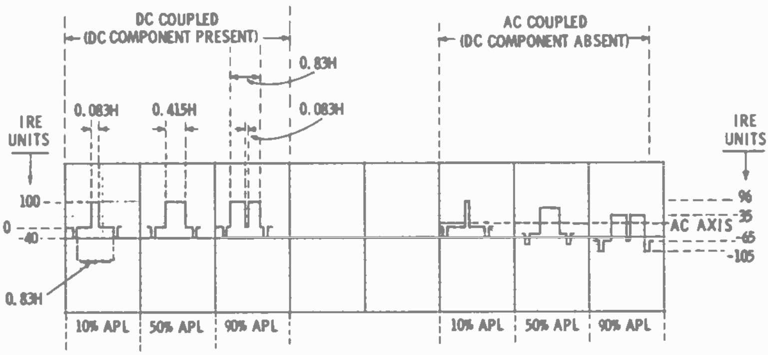


Fig. 6-2. Variation of signal excursions with APL.

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 (see Fig. 6-2). 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 stabilizing amplifiers, are AC-coupled. Table 6-1 tabulates IRE units to APL for AC-coupled amplifiers when the average AC axis is arbitrarily assigned zero IRE units. Note that the total excursion of the signal is 201 IRE units instead of the 140 normal IRE units.

Waveform monitors such as 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 525) allow switchable operation, either clamped or unclamped. Even though the monitoring CRO is clamped, "bounces" of a momentary duration will occur upon drastic scenic changes in APL, and this is normal. Most scopes employed for routine test and service do not use DC restorers or clamping circuits.

Table 6-1. IRE Units to APL for AC-Coupled Amplifiers With AC Axis (The No-Signal Operating Point) at 0 IRE 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 IRE units.			

Fig. 6-3 illustrates a sine wave of 15,750 cps, keyed with sync and phased as shown to represent one cycle of transition per line. The 3.58-mc wave is superimposed at 20 IRE units. This signal is often used by the Long Lines Radio & TV Division of the AT&T for testing purposes.

With the sine-wave test signal (Fig. 6-3) or the sawtooth test signal without APL (Fig. 6-4), the data are processed so that the results reported for each APL condition correspond to the levels shown in Table 6-2. In practice, the 50% APL condition is obtained with either signal repeated on successive lines at the normal transmission level. With the presence of sync pulses with 40 IRE units peak amplitude, the 15,750-cps sine-wave signal should have a capability of 184 IRE scale units as shown by Table 6-2 for tests to 10% APL. For the sawtooth wave with horizontal and vertical blanking accounted for, the corresponding peak am-

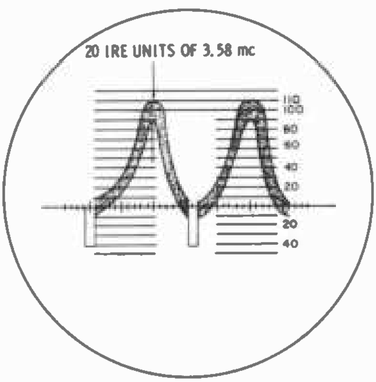


Fig. 6-3. A test signal with a sinusoidal waveform phased with sync to represent a black-to-white transition (and return to black) within a single-line period.

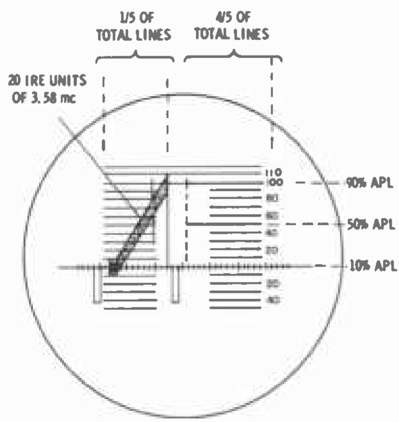


Fig. 6-4. Sawtooth (15.75 kc) with superimposed 3.58 mc at 20 IRE units. When provided with variable APL, the saw occurs once every five lines with fixed levels on alternate sets of four lines to simulate required APL.

**Table 6-2. Significant Range (IRE Units)
(Relative to the 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 mc to be retained at 20 IRE units for all levels of low-frequency test signal.		

plitude should be 178 IRE scale units. Note in any case that the 20 IRE units of superimposed 3.58 mc signal results in an additional 10 IRE scale units over the peak low-frequency level.

Fig. 6-5 presents the specifications for the popular stairstep generator test signal, when provided with variable APL. This generator normally includes provisions for inserting horizontal

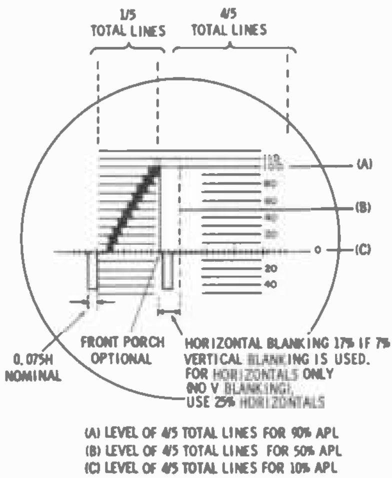


Fig. 6-5. Standard stairstep signal setup with 3.58 mc superimposed at 20 IRE units.

sync only. In this case, the blanking width (which is adjustable internally) should be set for 25% 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 (maximum of 11.4 microseconds horizontal with 7% vertical blanking).

All of the waveforms presented (sine wave, sawtooth, stairstep) are inherently 50% duty cycle signals when presented line for line. The required variation in APL is obtained by presenting this 50% APL signal for only one-fifth of the total active scanning 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% APL, at 50% of reference white for 50% APL, and at reference white level for 90% APL. The stair-step generator is most commonly used in station practice, and this type of waveform is considered in following material.

Equipment necessary to measure amplitude linearity (low-frequency) and differential gain at 3.58 mc is quite simple. All

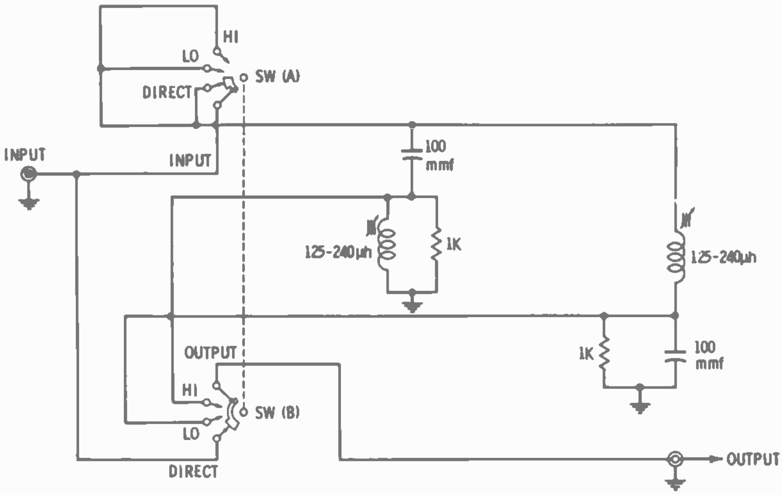
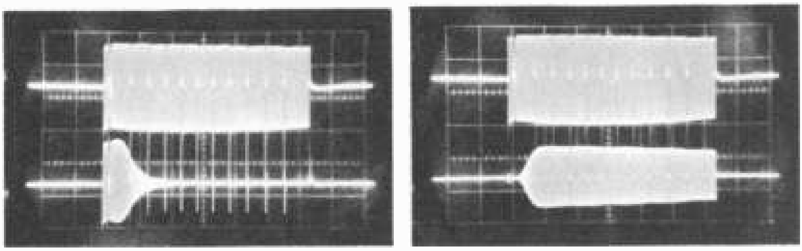


Fig. 6-6. Standard crossover filter for 3.58-mc frequency checks on stairstep.

that is required is a wide-band scope and a crossover network of the type shown in Fig. 6-6. This filter incorporates a switch for selecting Direct, Low-Pass, and Hi-Pass positions. The standard response of the various positions should be as shown in Fig. 6-7. The markers are at 1-mc intervals to 10 mc. Note that in the low-pass position the response is negligible at 3.58 mc, passing only the low-frequency 15,750-cps component of the signal (Fig. 6-7A). In the high-pass position, the response is negligible at frequencies below about 0.5 mc, while passing 3.58 mc (Fig. 6-7B). Note that any loss in amplitude of the 3.58-mc component is not of interest, since only the *difference* of levels at the steps between black and white is important.

The measurement of differential phase requires special equipment, such as the Tektronix 526 *Vectorscope*, the RCA Color-Signal Analyzer, or the Telechrome Video Transmission Test Signal Receiver. 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-mc oscillator is located at the studio, while 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 that received from the studio.

Fig. 6-8 illustrates the typical functions required. The test signal is applied to two paths: (a) a crystal filter circuit that converts the 3.58-mc signal component into a constant amplitude, constant phase reference and (b) a bandpass amplifier that amplifies only the 3.58-mc signal (and its associated sidebands when a color signal is used) so that the amplitude and phase of each of the 3.58-mc steps is preserved as received. (The test signal is the same as used for differential gain as described previously.)



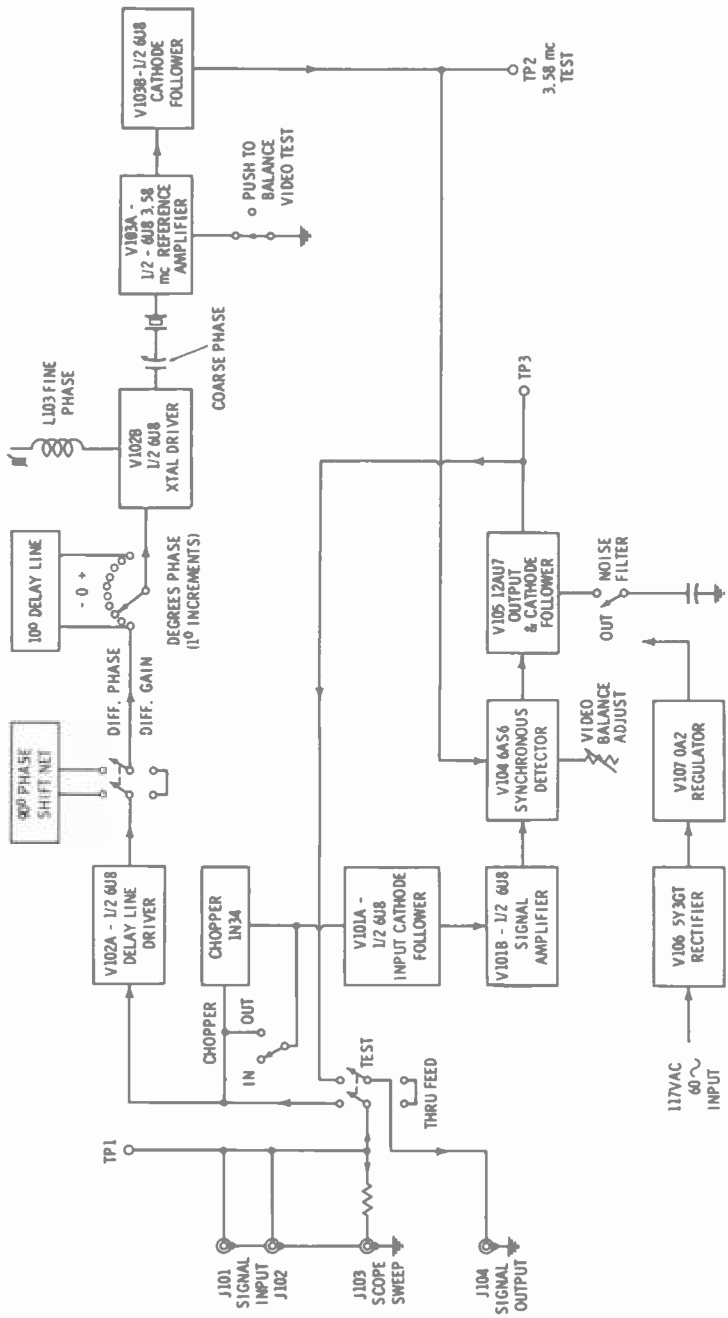
(A) Filter on "LO" position. (B) Filter on "HI" position.

Fig. 6-7. Characteristics of standard crossover filter for differential gain measurements (markers at 1 mc intervals to 10 mc).

The reference sine-wave and the amplified signal are applied to a phase demodulator whose output is amplified and viewed on the external scope. To measure differential phase the reference signal phase is adjusted so that it is nominally 90° out of phase with the signal. Under this condition, zero output from the demodulator is obtained. Slight differences from 90° result in an output that is proportional to amount of phase difference. In this instrument, the horizontal sync pulse is used as a reference line, since no 3.58-mc signal exists during this interval and therefore represents zero output. 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 calibrated dial.

6-3. AMPLITUDE LINEARITY AND DIFFERENTIAL GAIN

As in every type of testing, the very 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



Courtesy Telechrome

Fig. 6-8. Block diagram of a video-transmission, test-signal receiver.

on the oscilloscope, and use the internal adjustments of the stair-step generator to obtain a step on every 10 IRE 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. 6-9). This enables a most precise adjustment on linearity of the steps. Note that for an uncluttered view the pattern has been vertically displaced slightly from the IRE graticule lines in Fig. 6-9.

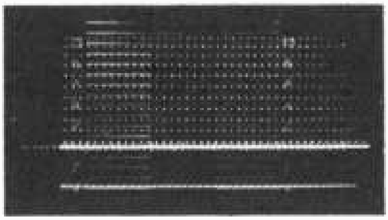


Fig. 6-9. The output of a stairstep generator at the vertical rate on a CRO to permit greater accuracy in setting the internal adjustments for linearity.

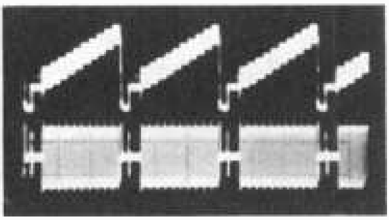


Fig. 6-10. Displays used to check the stairstep generator itself for differential gain on superimposed 3.58-mc signals. Upper trace is 10 steps with 3.58 mc on steps; lower trace is signal through hi-pass filter and increased gain on CRO.

Next, be certain the generator itself does not furnish differential gain on superimposed 3.58-mc signals. The upper trace of Fig. 6-10 shows the direct generator output, whereas the lower trace is the 3.58-mc component obtained through the high-pass filter described in Section 6-2. Differential gain from the generator should be zero.

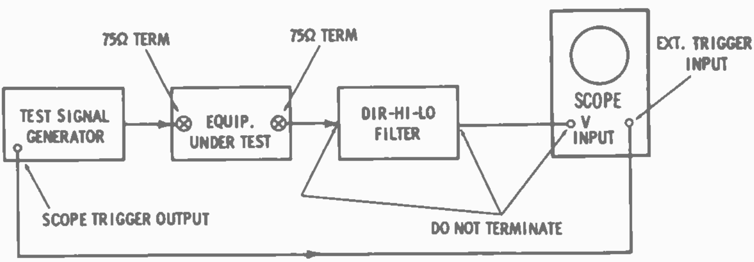


Fig. 6-11. Test setup for checking amplitude linearity and differential gain.

Fig. 6-11 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) is used.

Amplitude linearity normally refers to the luminance scale represented by the low-frequency step component. The upper trace of Fig. 6-12 shows the stairstep input signal, either without the 3.58 mc superimposed or as viewed through the low-pass filter with 3.58 mc present. The lower trace shows an almost complete clipping of the last (white) step.

The upper trace of Fig. 6-13 shows the output of a system under test, viewed with the filter in the Direct position, indicating obvious compression in the white region. The actual amount of differential gain is measured by viewing this signal through the high-pass filter as in the lower trace. Note that the transients (spikes) which are due to the fast rise-times of the steps provide a convenient marker to indicate the step number associated.

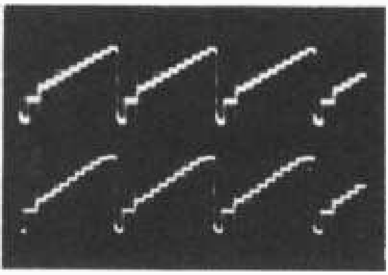


Fig. 6-12. CRO displays through a low-pass filter. Upper trace is input signal; lower trace shows filter output with white compression.

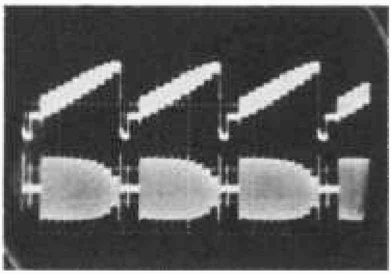


Fig. 6-13. Differential gain in white region. Upper trace shows output signal with filter in "Direct" position; lower trace is output signal with filter in "Hi-Pass" position with increased scope gain.

Differential gain is normally measured in percentage. On a wideband scope the 3.58-mc component through the high-pass filter may be "blown up" so that the highest block is 100 IRE units or 100%, and the amount of compression (smallest block) read in percent response. The peak-to-peak block 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 measurement of differential gain:

COMPRESSION:

$$\text{Differential Gain} = (1 - b/a) \times 100$$

where,

a is equal to height of uniform (linear) portion of signal,
b is equal to height of *smallest* block of signal.

EXPANSION:

$$\text{Differential Gain} = (b/a - 1) \times 100$$

where,

a is equal to height of uniform (linear) portion of signal,
b is equal to height of *largest* block of signal.

An example to illustrate the precaution required in computation of compression and expansion follows. Assume a equals 20 IRE units and b equals 15 IRE units. Then, for compression, we have:

$$
\begin{aligned}
\text{Differential Gain} &= (1 - 15/20)\ 100 \\
&= (1 - 0.75)\ 100 \\
&= (0.25)\ 100 \\
&= 25\%\ \text{differential gain (compressed)}
\end{aligned}
$$

Now assume a equals 15 IRE units, and b equals 20 IRE units. Then, for expansion, we have:

$$
\begin{aligned}
\text{Differential Gain} &= (20/15 - 1)\ 100 \\
&= (1.33 - 1)\ 100 \\
&= (0.33)\ 100 \\
&= 33\%\ \text{differential gain (expanded)}
\end{aligned}
$$

Fig. 6-14 illustrates the output of a stairstep generator with variable APL adjusted to equal 10% APL (upper trace). The lower trace is the same signal through the high-pass filter. The generator output should show zero differential gain. The upper trace of Fig. 6-15 shows the adjustment for 50% APL; the lower trace is the adjustment for 90% APL. The direct output of the

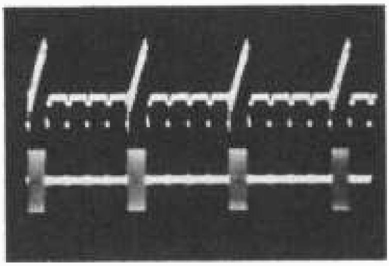


Fig. 6-14. Output of stairstep generator with variable APL adjusted to equal 10% APL. Upper trace shows 1-line-in-5 stairstep signal; lower trace is same signal through high-pass filter.

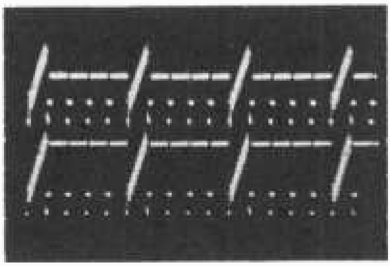


Fig. 6-15. 1-line-in-5 stairstep signal at 50% APL (upper trace) and same signal at 90% APL (lower trace).

generator should indicate zero differential gain at all three levels of APL.

It is obvious that the 50% APL condition most nearly simulates average transmission. Where time or facilities are limited, the 50% APL condition should be selected. However, a complete story of system performance is permitted only by tests at 10%, 50%, and 90% APL.

The tabulation of differential gain is normally expressed in either of two ways:

(A) The extreme values of differential gain with respect to that portion of the differential gain function judged to be most nearly constant. Plus implies expansion; minus implies compression. See Table 6-3 for typical data.
(B) The maximum range of the differential gain which is the difference of the extreme values. Note that if method B is used with the same data as in Table 6-3, differential gain would simply be recorded as +9%.

Some stations use a combination of the above methods by simply stating the maximum range of differential gain (method B) for each APL (method A).

Differential gain is caused by excessive response at high frequencies (overpeaking), by lagging transconductance of vacuum tubes or other tube vagaries, or by low plate or bias voltages. In a complete system, coaxial cables and terminations must also be checked. The existence of differential gain at 3.58 mc may or may not indicate luminance nonlinearity as pointed out by Fig. 6-1A. This is easily checked by the low-pass filter position presentation of the steps. It is obvious that if the steps themselves are compressed, differential gain will exist.

Modern stabilizing amplifiers designed for color systems normally employ predistortion networks (adjustable) for both differential gain and differential phase. However, these predistortion circuits are most useful at the transmitter installation, and they are covered in Section 8.

Table 6-3. Differential Gain Data (Method A)*

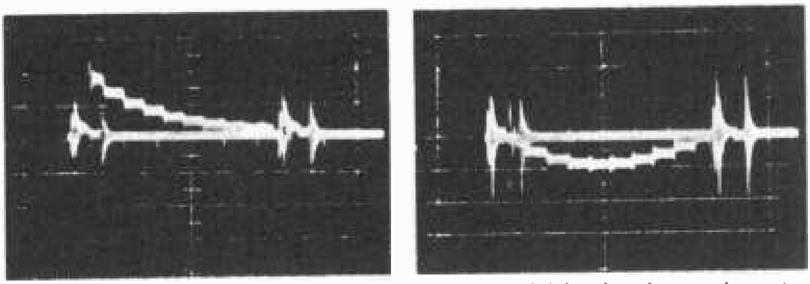
APL	ΔG%	AMPLITUDE REGION
10%	+2 0 −7	Black Gray White
50%	+2 0 −2	Black Gray White
90%	+5 0 −2	Black Gray White
Where: Black = block 1 + neighbors Gray = block 5 + neighbors White = block 10 + neighbors		

* Differential Gain Data (Method B): ΔG = +9%

6-4. DIFFERENTIAL PHASE

Differential phase is measured with the same type of test signal as discussed in previous sections, with special measuring equipment basically described in Section 6-2.

The output of the system is fed to the measuring device to determine the phase shift in degrees. Fig. 6-16 is a typical trace obtained from the measuring equipment with stairstep signal. In Fig. 6-16A, the right-hand edge of the trace has been adjusted to the



(A) With right-hand edge of trace adjusted to the reference line.

(B) With left-hand edge and center positioned on reference line.

Fig. 6-16. Two possible traces at the output of a color signal analyzer.

reference line by means of a position phase-control. A calibrated control is then adjusted to bring the left-hand edge of the trace to the reference point, and the differential phase in degrees is read directly from the calibration on the control (Fig. 6-16B). In this instance, the positioning phase control is adjusted to place both edges at the reference; then the calibrated control is used to position the center on the reference.

In general, a system which has little differential gain will also have minimum differential phase. Compression or clipping of the

Table 6-4. Differential Phase Data

APL	$\Delta\theta$ in degrees	Amplitude Region
10%	0 +0.5 −3	Black Center White
50%	0 +1 −2	Black Center White
90%	0 +1 −1	Black Center White

Note that 0° reference is at blanking level.

3.58-mc 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 chroma and luminance channels and impedance elements which may have constant impedance at line scanning frequencies, but are variable in the 3.58-mc region, can result in differential phase. The latter is more likely to occur in the transmitter than at the studio.

Differential phase is normally tabulated by either of two expressions as follows:

(A) The values of differential phase with respect to the value at the blanking level. "Plus" implies leading phase; "minus" implies lagging phase (see Table 6-4).

(B) The maximum range of the differential phase (difference of extreme values. Note that for the tabulated data of method A, the method B 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 3° differential phase at 3.58 mc. The transmitter should show a maximum of 7° for an overall allowance of 10°.

SECTION 7

MICROWAVE SYSTEMS

Television relay microwave systems operate in the 2,000-mc (1,990–2,110), the 7,000-mc (6,875–7,050) and the 13,000-mc (12,750–13,250) bands. The 7,000-mc region is the most popularly employed band of frequencies for remote pickups and studio-to-transmitter (STL) links.

This notebook 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 book. The manufacturers of the particular equipment used normally furnish either the complete service or the data necessary to plan the initial installation.

7-1. 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 by a series of tests in various weather conditions. Methods of measuring signal-to-noise ratio are described in Section 7-3.

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 equal to the path attenuation. 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 with the units themselves, you may proceed with the notion that either antenna misalignment has occurred (as might happen with loose mountings) or that the path is at fault. Antenna alignment should, of course, be tried first. If nothing is gained, investigate the path profile (see Fig. 7-1 for example). If a 100-ft clearance exists at the nearest point to the center of the beam, the path is not in trouble, since this provides an adequate safety margin.

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 used. Since it is more convenient for limited use, you may use ordinary linear graph paper and the data of Fig. 7-2. Paper with ten squares to the inch is ideal for this purpose.

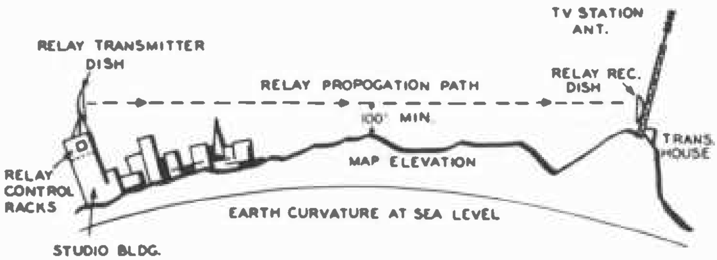


Fig. 7-1. Basic principle of STL.

2. The path profile and obstructions on the path may be charted from topographic maps. The topographic map gives the height above sea level above the 4/3 earth contour, to which is added the height of major obstructions. Maps for specific areas may be obtained from U. S. Geological Survey, Washington 25, D. C. For maps of areas west of the Mississippi write to the U. S. Geological Survey, Denver 15, Colorado.
3. The clearance from the tallest obstruction in the path should be at least that shown in Table 7-1.

The above technique will enable you to roughly determine 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.

You can also determine approximately what should be expected in signal-to-noise ratio by use of the antenna system gains and free-space path attenuation. Table 7-2 shows the gain of parabola reflectors as a function of size and operating frequency. Table 7-3 is the system gain for the designed parabola and reflector size at 7,000 mc. (These data are obtainable from the equipment manufacturer of your microwave unit.) Table 7-4 is a tabulation of the free-space loss for the microwave bands up to a 30-mile path

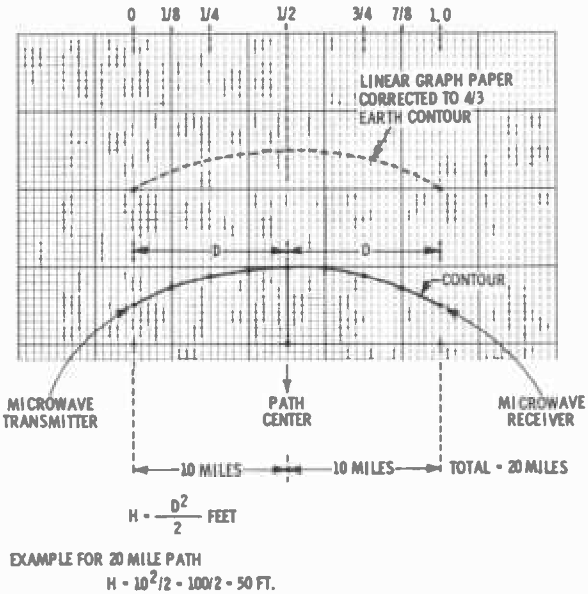


Fig. 7-2. Method of plotting a profile of the transmission path.

Table 7-1. 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

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: 7,000 mc.
Path length: 20 miles.

Table 7-2. Approximate Gain (Parabola Only)

	DB GAIN		
DISH	2,000 mc	7,000 mc	13,000 mc
4 ft.	25	37	42
6 ft.	28	40	45

Table 7-3. Approximate Antenna System Gain (Parabola & Reflector) in db Power—Frequency: 7,000 mc

NOTE: ALL DIMENSIONS IN FEET

H DIMENSION	D_2 DIMENSION 4	6	
80	35	39	
100	34	38.5	
120	33	38	
140	32	37.5	D_1 = 4 ft Parab.
160	31	37	
180	30	36.5	
200	29	36	
80	36	40.5	
100	35	40	
120	34	39.5	
140	33	39	D_1 = 6 ft Parab.
160	32	38.5	
180	31	38	
200	30	37.5	

Table 7-4. Approximate Free-Space Loss in db

MILES	FREQUENCY 2,000 mc	7,000 mc	13,000 mc
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 = $37 + 20 \text{ Log } f_{mc} + 20 \text{ Log } D_{(miles)}$
f_{mc} = operating freq. in mc
D = distance in miles

Antenna system gain at each end: 35 db. (This is a total of 70 db gain.)

The net path loss is:

$$A = (a) - G_t - G_r$$

where,

A is equal to net path loss,
(a) is equal to free-space loss (from Table 7-4),
G_t is equal to transmitter system gain (Table 7-2 or 7-3),
G_r is equal to receiver antenna system gain (Table 7-2 or 7-3).

From the example in the above data:

$$\begin{aligned} A &= 139 - (+35) - (+35) \\ &= 139 - 70 \\ &= -69 \text{ db} \end{aligned}$$

The receiver power input is:

$$P_r = P_t - A$$

where,

P_r is equal to receiver power input,
P_t is equal to transmitter power output,
A is equal to net path attenuation.

The net path loss (in db) subtracted from the transmitter power output (for 1 watt this is expressed as 0 dbw, or 0 db above 1 watt) will give the power input to the receiver in −dbw (db below 1 watt). As:

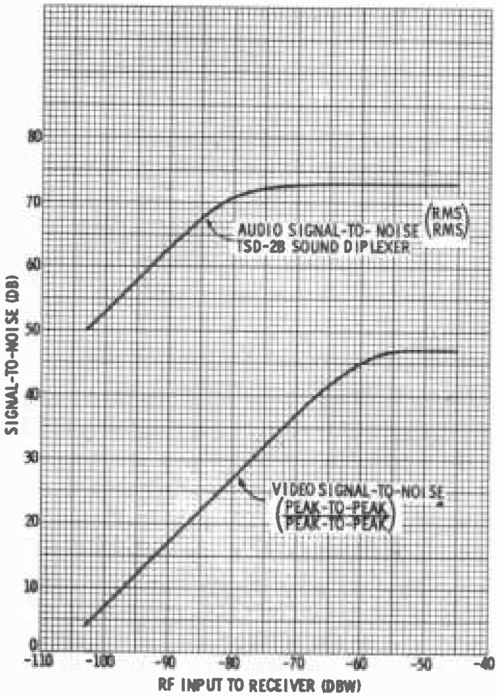
$$P_r = 0 - 69 = -69 \text{ dbw.}$$

Fig. 7-3 is a graph of the expected signal-to-noise ratio versus power input to the receiver (dbw) for the RCA-TVM-1 microwave relay system. Note that for −69-dbw power input, the video signal-to-noise ratio should be 38 db on a peak-to-peak video to a peak-to-peak noise basis. This is approximately 56 db on a peak-to-peak video to rms noise basis, as covered under techniques of measurement in Section 7-3.

Note also from the graph of Fig. 7-3 that this receiver input power should result in better than a 72-db signal-to-noise ratio in diplexed sound. Techniques of checking diplexed sound are given in Section 7-4.

In evaluating microwave relay performance, the following fundamentals serve as the basic guide:

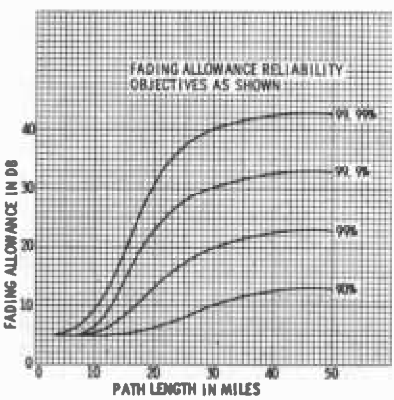
1. In general, noise will be visible in the picture when the signal-to-noise ratio deteriorates to less than 24 db.
2. Since a microwave beam is sometimes bent and scattered by atmospheric conditions, not only adequate clearance, but adequate *fading margin* must be provided. The picture becomes unusable at a signal-to-noise ratio of 8 db. This 8-db figure establishes a basis for computing the fade margin.
3. For example, if a 24-db SN ratio exists, the fade margin is only 16 db ($24 - 8 = 16$). From the graph in Fig. 7-4, a 16-db fade margin indicates only about 98% 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%, the fade margin for a 25-mile path should be 37 db. This requires a SN ratio of $37 + 8 = 45$ db.
4. From Fig. 7-4, compute your allowable fade margin for 99.99% reliability, especially if the service is STL. Assume your path is 20 miles. This requires a fade margin of 31 db. Then you should obtain a minimum of $31 + 8 = 39$-db SN



Courtesy RCA

Fig. 7-3. Graph showing signal-to-noise ratio.

ratio average on measurements made under average weather conditions. Normally, measurements made daily or nightly over a period of a week or two will give a reliable indication of the practical SN ratio of your system.



Courtesy RCA

Fig. 7-4. Graph showing fading allowance.

7-2. MICROWAVE CIRCUIT FUNDAMENTALS

A microwave system employs conventional circuits with the exception of two major items: (A) A special type of tube (normally either a magnetron or klystron) is used at the transmitter and receiver heads. (B) A *wave guide* and *buttonhook* are used to couple the RF energy into the parabola dish at both transmitter and receiver.

Fig. 7-5 illustrates typical microwave plumbing. The wavemeter and monitor cavity are extensions of the same type of waveguide used for the specific application.

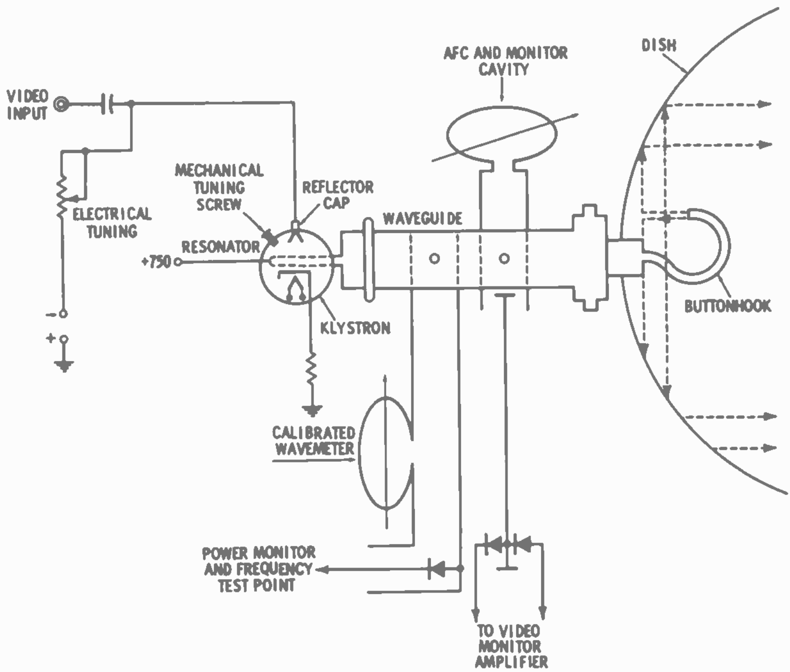


Fig. 7-5. Basic microwave "plumbing."

The reflex klystron tube has a single resonant cavity. In conventional tubes with negative grids the ideal operating characteristics are: constant velocity stream of electrons from cathode to plate, varied in intensity by a variation of the grid bias according to the signal voltage applied. At extremely high frequencies, however, such action becomes practically impossible, due to the transit time of the electrons from cathode to plate. This difficulty becomes increasingly evident in the larger tubes for handling large power outputs. The electrons which traverse the grid during the negative swing of the signal voltage are slowed down, and those which traverse the grid region during the positive signal swing are speeded up. Thus the electrons reach the plate at random speeds, decreasing the efficiency of the tube in accomplishing any amplification of the applied signal. The klystron tube, however, is designed to use the applied signal to control the *velocity* of a *constant-current* beam, instead of attempting to vary the intensity of a constant-velocity beam.

The cavity is designed for resonance in the operating band. Electrons bouncing back and forth in this cavity pass through several gridlike structures into a drift space, which is terminated in the reflector electrode. The resonator is operated at approximately 750 volts positive, and the reflector electrode is operated at a negative potential. Thus the electrons are reflected back in to the resonator, resulting in a bunching effect of the electrons, sometimes referred to as *velocity modulation*. The resultant RF oscillations are extracted by a probe in the waveguide and fed to the coaxial output line through a wide-band, coaxial, transducer coupling unit. This coupler efficiently couples the coax output line to the waveguide. The socket through which this line passes is an ordinary octal socket.

Although the reflex klystron is designed for a particular band of operation, both mechanical and electrical tuning within this range are available. A mechanical screw turned clockwise (into) the cavity decreases the frequency; a counterclockwise turn increases the frequency. As described later under tuning techniques, the klystron is centered for best operating characteristics by this mechanical adjustment in conjunction with the variable negative potential on the reflector electrode. The klystron oscillator is *frequency modulated* by varying the reflector voltage with the video signal. The modulation sensitivity of a type 220C klystron averages about 400 kc per volt of signal.

At radio frequencies above approximately 300 mc, special techniques must be applied in getting the RF energy from unit to unit or from amplifier to antenna. The entire UHF band, as well as the microwave region, uses specially designed connection devices for transferring the RF signals. Although it is not important that the TV maintenance engineer understand all of the theory associated with transmission lines and wave guides, he should be familiar with the fundamentals, so that the major elements of mystery in such devices are removed.

Ordinary transmission lines are practically unusable on such frequencies, due to the severe attenuation along the line. This occurs because of the high series inductance and low shunt capacitive reactance between the inner and outer conductors. Although this problem has been solved by the development of a special transmission line for transmitters in the UHF range, wave guides are generally used at the microwave frequencies of TV relay operation.

A wave guide is a hollow metallic tube, usually rectangular in shape, although it may be round or oval. Its purpose is to pass high radio frequencies with a minimum of attenuation within the boundaries of the tube. When radio waves are radiated into free space, electromagnetic and electrostatic fields exist at the point of propagation—the antenna. When this energy is directed into an ordinary transmission line, the lower-frequency energies are evenly distributed throughout the conductors as in free space. At UHF and microwave frequencies, however, most of the current is concentrated on the outer surface of the conductor (skin effect),

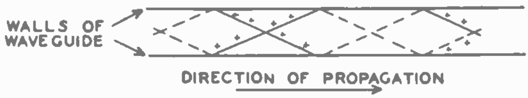


Fig. 7-6. One electrostatic field along a wave guide (looking down on wave guide from top). The fields cancel along the walls and add through the center of the tube, thereby eliminating "skin effect" which occurs at high frequencies.

and a high resistance is presented to the passage of this current through the line. In waveguides, the dimensions of the tube are such that the concentration of energy takes place in the center, with very little electric field existing at the walls.

Fig. 7-6 illustrates the electrostatic field along two wavepaths and how the positive maxima and negative maxima meet at the walls, effectively cancelling at these points. The positive maximum and negative maximum meet at the center of the tube, adding to the field of force through the center.

The engineer encounters terms such as TE and TM, with various subscripts such as TE_{01} and TM_{01} in literature concerning wave guides. TE stands for *transverse electrostatic* mode of operation, and TM stands for *transverse magnetic* mode of operation. It should be recalled that all radio energy contains both electrostatic and magnetic fields which are at right angles to each other. When the electrostatic field is as shown in Fig. 7-7A (that is, across the guide or transverse to the direction of propagation), the TE mode is designated. When the magnetic field is transverse to the direction of propagation as in Fig. 7-7B, the TM mode is designated. The mode of operation is determined by the manner in which the RF energy is fed into the waveguide. Note that when the RF energy is fed by a probe or a quarter-wave dipole in the manner shown in Fig. 7-7A, the TE mode results. When fed by a loop in the manner shown in Fig. 7-7B, the TM mode results. Power is extracted, depending on the mode of operation, in the same way.

The subscripts 0 and 1 designate the number of half-wave patterns of the electrostatic field along the B and A sides of the structure. The frequency at which dimension A is ½ wavelength is termed the *cutoff frequency*, and it therefore determines the lowest frequency that may be propagated by the waveguide. Frequencies higher than this cutoff frequency are readily passed. However, if the frequency is very much higher, other modes of operation, such as TE_{11} or TM_{11}, occur. This simply means that

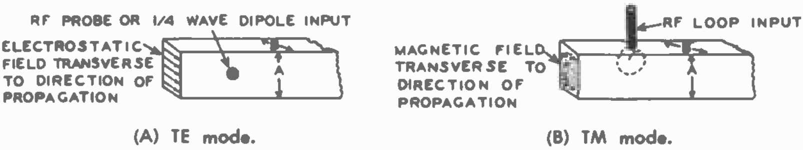


Fig. 7-7. The method of introducing RF energy in a wave guide determines the mode of operation.

more than one half-wave pattern occurs across the tube. In practice, the operator will find that these higher modes of operation are not used in TV relays, and TE_{01} or TM_{01} modes are predominant. The short (B) side is one-half the dimension of the (A) side. A wave guide used for the 2,000-mc band is approximately 3″ × 1½″ and for the 7,000-mc band is approximately 1″ × ½″.

The TE mode is normally used with the klystron tube, the direction of polarization depending on the direction of the short (B) side. A "straight" buttonhook normally gives horizontal polarization; a "half-twist" is imparted to the waveguide to give vertical polarization.

Fig. 7-8 illustrates the signal-path block diagram of a typical microwave system. Sound diplexing is optional, but it is normally employed when the system is used STL. A sound diplexer employs a frequency-modulated subcarrier (normally between 5.5 and 7 mc) to which the picture signal is added (not mixed) in a passive combining network.

The cable between the control unit and the head is normally a standard camera cable which carries all voltages, control circuits and video. When the coax of the camera cable is used to carry the video signal, a 75- to 50-ohm matching network is necessary, as shown in Fig. 7-8. When the camera cable must be longer than 50 feet, video equalization is necessary; otherwise, response falls off about 0.85 db per 100 feet at 6 mc. With runs greater than 200 feet it is normal practice to use an external video cable, either RG-8/U or RG-11/U. The RG-8/U provides a 50-ohm line. If RG-11/U is employed, the matching network is eliminated and

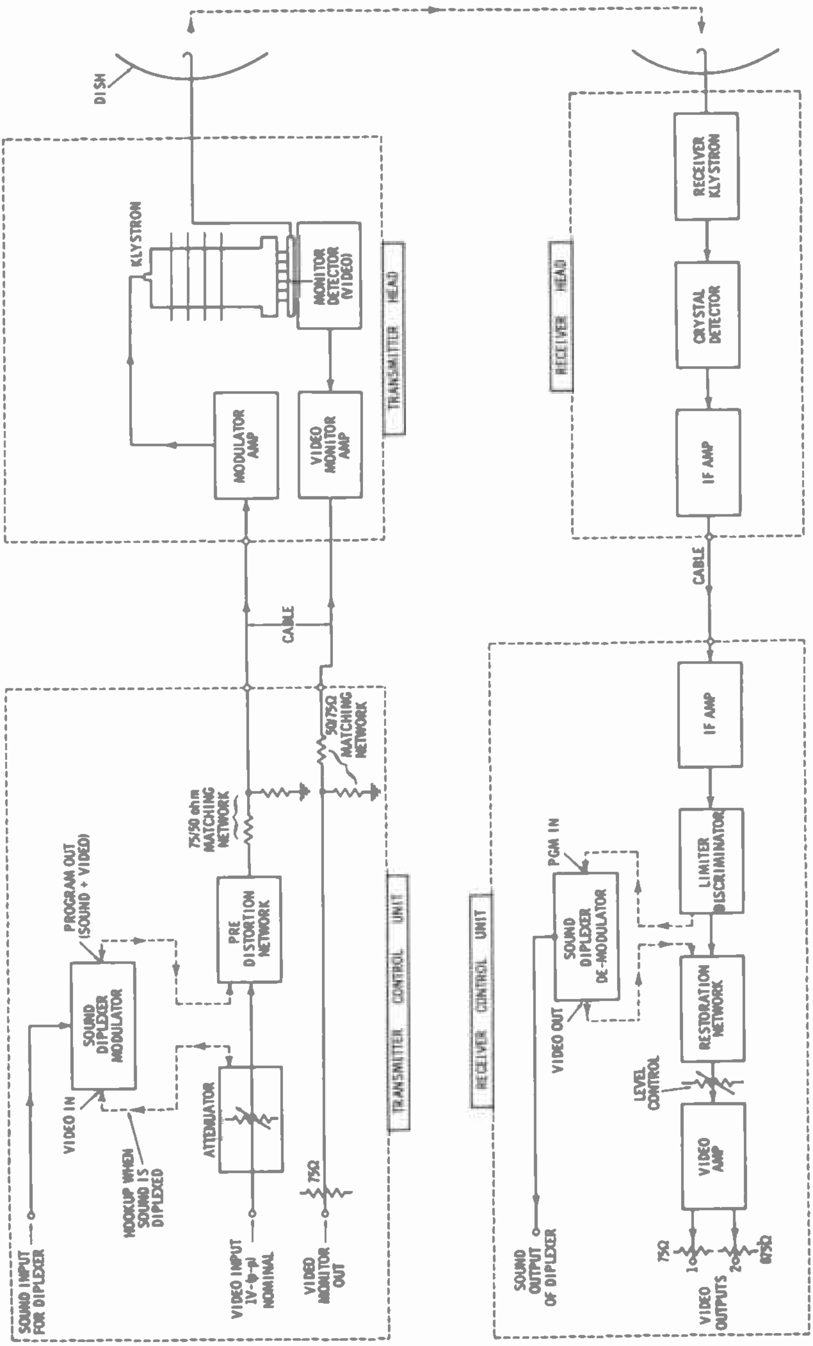


Fig. 7-8. Block diagram showing the signal path in a typical microwave system.

the modulator is modified to provide a 75-ohm termination, rather than 50 ohms. When either of the external lines is used, equalization is required at the rate of 0.4 db per 100 feet at 6 mc.

Predistortion is employed when either sound diplexing or color signals (or both) must be handled. A typical circuit provides a video insertion loss of 8 db at 60 cps. This loss remains 8 db through approximately 100 kc, then tapers to 4 db at 1 mc, 0.9 db at 3 mc and 0.25 db at 6 mc. This reduces transmitter frequency deviation at low frequencies, reducing frequency excursion in the IF section of the receiver, with a marked reduction in differential phase shift at the 3.58-mc color subcarrier and the higher-frequency sound subcarrier.

A restoration network with a reverse attenuation characteristic, is provided in the transmitter monitor and a similar network is used in the receiver. This restores the output to a flat response.

7-3. VIDEO MEASUREMENTS

Measurements on a microwave system that has been properly installed are directly affected by transmitter and receiver tuning; therefore, this subject is of initial importance.

1. Set the wavemeter in the transmitter head on the exact operating frequency desired, noting any calibration correction data for the frequency setting that may have been supplied by the manufacturer.
2. Apply 60-cps sine-wave modulation. This is normally provided by a TEST position on the video selector switch at the control unit. Note from Fig. 7-5 that 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 readily be monitored 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.
3. In Fig. 7-9 the applied sine wave is of such amplitude as to cause a 6-mc deviation of frequency (± 3 mc from the operating frequency, f_0). The repeller voltage and the mechanical screw on the klystron are alternately varied to obtain maximum crystal current and minimum AC component, as observed on the external CRO. Note from curve 1 in Fig. 7-9

that the power output varies at twice the rate of the modulating voltage, which results in a 120-cps pattern. (The wavemeter absorbs power each time the frequency swings through f_o; thus the notches occur at a 120-cps rate for a 60-cps 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 gain.

The most sensitive indication of klystron "centering" is the AC component. Fig. 7-10 illustrates how the AC component is increased with the klystron off-center (Fig. 7-10A). When the klystron is not properly centered, either white or sync compression will occur, particularly at high modulation levels. When

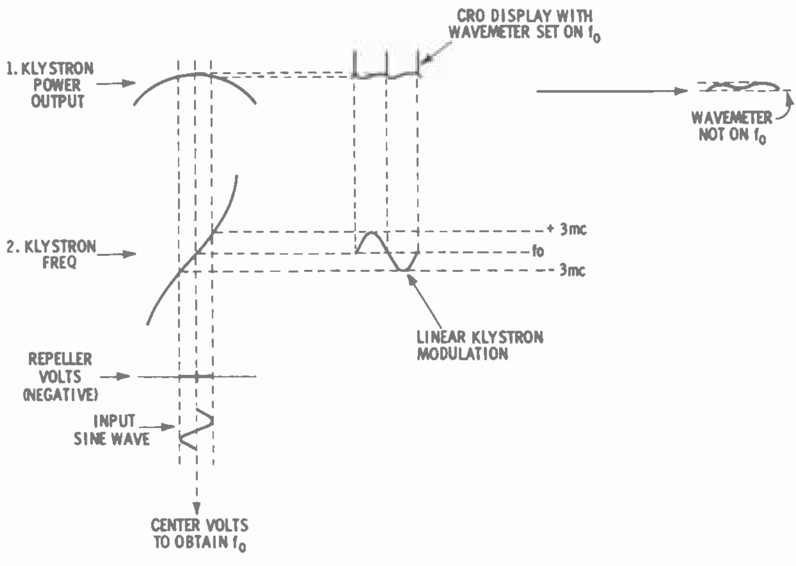


Fig. 7-9. Correct tuning of a transmitter klystron.

modern klystrons are properly centered (Fig. 7-10B) (maximum crystal current and minimum AC component at the Freq. Test jack), no compression will occur at levels up to almost twice the normal modulation level. The limiting factor in most cases is the modulator amplifier.

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 (5.8 mc in the Raytheon system; 6.8 mc in the RCA) will cause a certain degree of sync buzz in the sound if the receiver is detuned. With normal video applied but no sound, observe the audio output of the sound demodulator with an oscilloscope set for vertical-rate display. This enables precise tuning of the receiver klystron for minimum crosstalk into the audio channel. With high scope gain, the 60-cps

(A) Off center. (B) Properly centered.

Fig. 7-10. Waveform displays indicating klystron centering.

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 modulation-frequency deviation must also be obtained before making video measurements. The normal procedure using a wavemeter is as follows:

A. With the test 60-cps signal applied, proper tuning is indicated by the 120-cps signal shown in the upper trace of Fig. 7-11.
B. Unlock the calibrated dial of the wavemeter and rotate the dial until the notches just merge into the 30-cps signal (lower trace of Fig. 7-11). Read the frequency at which this occurs on the dial.
C. Now rotate the dial in the opposite direction until the notches again just merge. Read this frequency. If the normal 100% modulation is to be 6 mc (p-p), the deviation from each side of center frequency should be 3 mc.

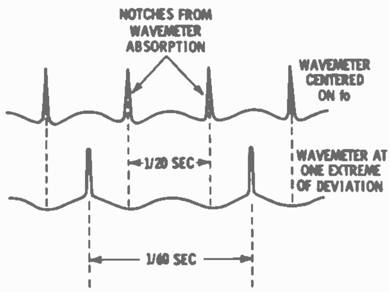


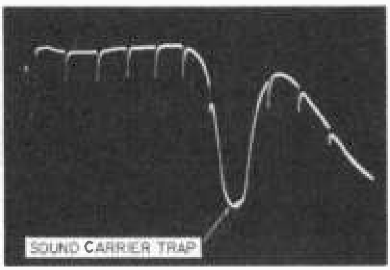
Fig. 7-11. CRO display of wavemeter indication of deviation from center frequency with 60-cycle test signal.

D. If the total p-p swing is not correct, adjust the amplitude of of the 60-cps test signal until the required deviation is obtained. Measure the p-p value of this sine wave at the klystron repeller terminal with a low-capacity scope probe. For example, the type 220 klystron will require approximately 15 volts (p-p) of signal (modulation sensitivity about 400 kc per volt) to obtain a deviation (p-p) of 6 mc.

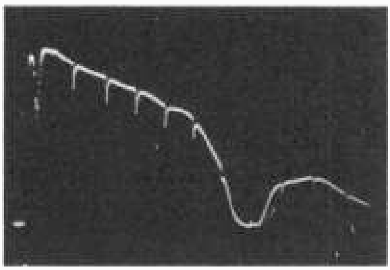
E. Now bypass the predistortion network (if used), and feed a normal video level to the modulator. Adjust the signal-modulation gain to obtain the value (p-p) required as determined by step D. Reinsertion of the predistortion network will reduce the swing at 60 cps to 2.4 mc if the usual 8-db network (described previously) is employed.

Microwave Frequency Response

Initially, the transmitter monitor amplifier should be measured for frequency and amplitude linearity, so that overall measurements can be conveniently diagnosed as to individual characteristics of 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 techniques are the same as those outlined in Sections 5 and 6.



(A) Proper overall response when sound diplexer is used with a subcarrier frequency of 6.8 mc.



(B) Display showing detected sweep on system with excessive roll-off characteristics.

Fig. 7-12. Video sweep of STL system (markers at 1-mc intervals from 1 to 10 mc).

The video sweep is most conveniently used for wide-band response measurement. NOTE: If the system employs line-to-line clampers, keyed sweep should be used as described in Section 5. Fig. 7-12A shows the proper overall response when a sound diplexer is used with a subcarrier frequency of 6.8 mc. The sound notch should be centered on the subcarrier frequency with a width (normally) of about 300 kc. This adjustment is usually provided by trimmers in the passive mixing network for video and audio at the transmitter.

Fig. 7-12B shows the detected sweep on a system with excessive roll-off. If this occurs on the overall system measurement at the receiver video output and the transmitter monitor output indicates the waveform of Fig. 7-12A, the roll-off is actually at the receiver. In STL installations, two units at both 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. 7-12A, while the response through receiver 1 and transmitter 2 is as shown by Fig. 7-12B, then transmitter 2 needs servicing. The same response should be evident from the transmitter No. 2 monitor video.

Poor frequency response is usually the result of sagging transconductance in modulator tubes, faulty coax lines, or bad terminations. Complete alignment of the modulator circuits may be necessary at intervals of a year or two.

Similarly, microwave receiver IF circuitry seldom needs complete alignment procedures, since modern units employ IF bandwidths 18 to 30 mc wide, which is far over the normal 4- to 8-mc deviation swing of the carrier. Therefore, poor frequency response is almost always the result of tubes or plate loads, as previously discussed.

Amplitude and Phase Linearity

Amplitude and phase linearity is checked in the conventional manner with the stairstep signal as described in Section 6. Always check the maximum allowable p-p value of signal input to where

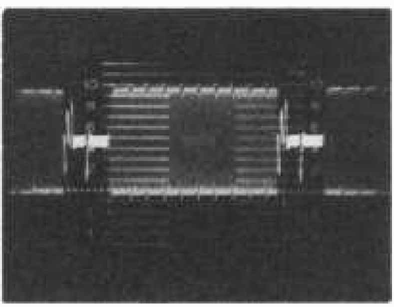


Fig. 7-13. Display permitting measurement of differential gain.

compression just starts. This provides a warning flag as to need for further checks. For example, if the normal p-p video input 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 mc must also be measured. (The techniques are the same as those in Section 6.) Always adjust the transmitter monitor cavity for zero differential gain, as shown in Fig. 7-13. This will result in minimum differential phase. Remember that for overall differential-phase measurements, a burst-controlled oscillator unit must be employed at the receiver (Section 6) unless the units are measured back-to-back on the bench.

Causes and cures of poor amplitude and phase linearity are the same as those outlined in Sections 5 and 6. 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.

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 video-to-random noise ratio 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 to peak-to-peak noise, or peak-to-peak video to rms noise. The former is most often used in practice, since it requires a wide-band scope (10 mc) readily available at the station. The noise may be specified in terms of rms when a wide-band (6 mc) VTVM is available, calibrated in rms of a sine wave. 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.

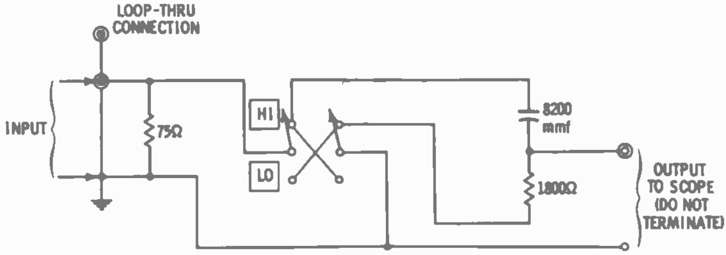
EIA standards call for a p-p video to rms noise minimum of 58 db. Thus to meet this standard, the p-p video to p-p noise measurement should be at least 38 db. In any event, the measurement should indicate a signal-to-noise ratio of at least the minimum required for the conditions outlined in Section 7-1. 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 by Fig. 7-14 should be constructed and used. The input circuit provides a termination for the receiver video output. The filter output should go directly to the scope vertical amplifier input without termination.

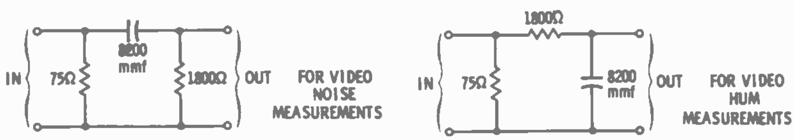
A convenient method of hookup is shown in Fig. 7-15. The reference video level fed from the microwave transmitter is measured by scope input No. 1. The filter output at No. 2 input is then available at the flick of a switch. This assumes a dual input (switchable) as available on the Tektronix 524 scope. The filter provides zero attenuation at 60 cps in the LO position, and zero attenuation above 100 kc in the HI position. No insertion loss is involved and the readings are direct.

A suggested method of signal-to-noise measurement is as follows:

1. If sound diplexing is used, turn off the sound modulator and demodulator.
2. Leaving any predistortion and restoration networks employed in place, feed the 60-cps test sine wave through the transmitter at the normal deviation (100% modulation).



(A) Filter circuit.



(B) Equivalent circuit for "HI" position. (C) Equivalent circuit for "LO" position.

Fig. 7-14. Video filter for STL signal-to-noise measurements.

3. Set the receiver video output level to the station standard (normally 1 volt p-p). If the hookup of Fig. 7-15 is used, the signal should be applied at the No. 1 scope input. Use vertical-rate sweep.
4. Remove the test signal at the transmitter, and remove the coax line feeding the video input. Substitute a termination.
5. At the receiving position, observe the No. 2 scope input with the filter in the HI-pass position. Increase the gain of the scope to maximum and read the p-p voltage of the random noise. The resulting voltage ratio is converted to db from a db table or by computing 20 log of the ratio.

NOTE: Some engineers prefer to use the scope internal 60-cps sweep with the Sweep Attenuator adjusted to collapse the trace to about a quarter-inch width. This will give a more readily measured noise indication at very low noise levels.

The video signal-to-hum measurement is made exactly the same as the above except that the filter is placed in the LO-pass position for the hum content trace. Hum content should be at least 40 db down, or at least that which is 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.

7-4. DIPLEXED SOUND MEASUREMENTS

It is imperative that the engineer be familiar with the practical use of the audio noise and distortion meter. If not, the reader should refer to the authors book, *AM-FM Broadcast Maintenance*, Sections 1 and 4. In this respect, this book can be considered a companion book to this one. The techniques are the same for

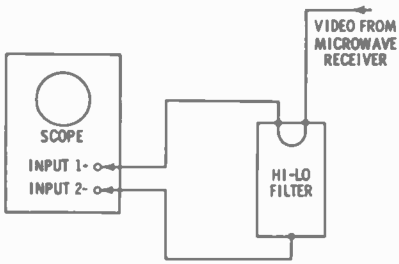


Fig. 7-15. Suggested method of using high-low filter.

regular FM broadcasts and aural TV transmitters. The only difference is that FM broadcast employs ±75-kc deviation for 100% modulation, as contrasted to ±25-kc deviation in the aural TV transmitter.

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 Section 8.

Frequency Response

A suggested procedure for frequency-response measurements is as follows:

1. Feed a 1-kc tone to the modulator at the specified program input level. Adjust the modulator gain to 100% modulation (0 VU on the meter).
2. At the receiver, adjust the sound-demodulator gain for ref-

erence output (0 VU). Calibrate the measuring equipment, and run response measurements at 50, 100, and 400 cps. 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 kc. 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 kc.

Since the overall measurement includes de-emphasis, the response curve should be relatively flat. In any event, the response must be adequate to meet FCC Proof-of-Performance runs on the overall system.

Distortion Measurement

1. With a 1-kc tone applied to the modulator and the gain adjusted for 100% modulation, apply proper termination to the demodulator output and calibrate the noise-distortion meter connected across this termination.
2. Make the distortion measurement at 1 kc. Then reduce the tone generator frequency to 50 cps and *use the same generator output level as for the 1-kc tone.* (The modulation meter will fall considerably due to 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 kc (hold demodulator output constant; hold signal generator output constant).
4. At all frequencies above 1 kc, *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-distortion meter at each new frequency. Make distortion measurements at the required spot frequencies up to 15 kc.

Distortion is normally produced by the reactance tube and the tubes in the modulator used for audio amplification. In the demodulator, it 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 Section 8), and a spot check of the STL only indicates sound diplexing to 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 can normally be completed in about four hours during any nonbroadcast time. The manufacturer's instructions regarding adjustments affecting 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 affects the discriminator secondary adjustment for balance (zero output at no modulation). When such adjustments

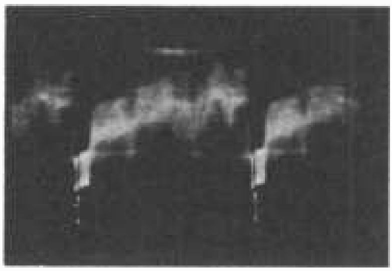


Fig. 7-16. Wideband CRO display of video signal with diplexed sound (vertical-rate time base). The amplitude of the sound subcarrier is normally adjusted to almost fill in the sync region.

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. Due to the wide variance of circuitry, it is necessary to follow implicitly 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 kc (Fig. 7-12). If the sweep is observed on a wide-band display (undetected), the flatness of the notch is more evident.

Aural Signal-to-Noise Ratio

1. Feed a 1-kc tone to modulator at normal program level. Adjust the modulator gain control for 100% (0 VU) modulation.
2. Adjust demodulator gain for 0 VU output, with proper termination applied and noise meter connected across this termination. Calibrate the noise meter.
3. Remove tone modulation and substitute a resistor equal to

the input impedance (normally 150 or 600 ohms). Measure the residual noise.

The noise level of modern diplexing equipment should be between −70 and −85 dbm when measured back-to-back on the bench. In service, this level depends on the effectiveness of the path. Note that to expect a signal-to-noise ratio in sound of 70 db, it is necessary to obtain at least 25-db signal-to-noise ratio in video on a p-p video to p-p noise basis.

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 on the picture output. Fig. 7-16 illustrates the sound subcarrier level adjusted just below an amplitude that would completely fill in the sync region. In practice, the subcarrier level is adjusted to approximately 20% of the p-p video level.

SECTION 8

TRANSMITTER MAINTENANCE AND PROOF OF PERFORMANCE

Details of transmitter circuits with which the operator is particularly concerned are presented in this section. Engineers concerned with new station planning and details of transmitter location, field intensity surveys, etc., should obtain the current FCC Rules and Regulations. This information is obtainable from the Superintendent of Documents, Government Printing Office, Washington 25, D. C., at prevailing prices, which must be obtained from that source before ordering.

8-1. FUNCTIONAL DESCRIPTION OF VIDEO STAGES

The function of the *visual exciter* of a TV transmitter is to generate the carrier wave at the assigned frequency, and to amplify the power to the designated power output. For the utmost in stability, a crystal-controlled oscillator is used. A typical arrangement for channels 2 through 6 consists of a crystal oscillator stage, followed by a tripler and two doublers, resulting in a frequency multiplication of 12 ($3 \times 2 \times 2$). For example, channel 6 has its video carrier frequency at 83.25 mc. Thus, the crystal frequency would be 6.93750 mc (83.25/12). The multipliers step up this oscillator voltage in both frequency and amplitude sufficient to drive the modulated stage. Thus the visual exciter consists of the conventional narrow-band RF circuits, which are tuned to the crystal frequency by adjusting the tank-circuit capacitors (or inductors when inductively tuned) for minimum plate current and maximum grid current to the following stage. All necessary frequency multiplication takes place prior to the stage being amplitude-modulated by the video voltage.

The incoming video from the studio is usually fed to a stabilizing amplifier to minimize the effects of hum, noise, or sync compression. Incorporated in the stabilizing amplifier are circuits designed for control of the relative sync-to-video amplitude (sync stretcher) and, in some cases, linearity controls to precorrect the transfer characteristics of the transmitter amplifiers. The transmitter transfer characteristic is the ratio of the RF output voltage to video input voltage and is generally linear within 10%. This is desired at the transmitter, since the *gamma* at the studio (signal sources) is adjusted to result in optimum picture quality as observed by the television audience; hence, the remaining portion of the overall system should be as linear (gamma = 1) as possible.

The final *video* amplifier stage in the transmitter is the video modulator. The DC component is reinserted at the grid of this stage by clamper circuits and the modulators are then coupled by direct coupling to the RF stage being modulated to maintain this DC component.

The RF circuits that follow the modulated stage are essentially linear RF amplifiers adjusted for maximum power output *consistent with a flat frequency-response throughout the upper sideband.* Proper adjustment of these amplifiers results in partial cancellation of the lower sideband. If low-level modulation is used (which may be either grid or plate modulation) a sufficient number of linear amplifiers are used to obtain the desired vestigial sideband response. This action is aided by inserting a notching filter adjusted to 1.25 megacycles below the video carrier frequency. For high-level modulation (which must be grid modulation), a vestigial sideband filter must be used. The standard transmission signal from this filter is then fed to either a *bridge diplexer* or *notch diplexer*, into the transmission line to the antenna system.

In the visual exciter section of Fig. 8-1, typical values are given for a transmitter working channel 6. The incoming video is shown with DC restoration taking place at the grids of the modulator tubes. The modulator stage of most commercial video transmitters is made flat to 5 mc to insure freedom from phase distortion. Observed in the illustration is the partial removal of the lower sideband at the output of the (in this example) final modulated stage.

The operator must become acquainted with the correct interpretation of picture and waveform monitors, since limitations in both transmitter and monitoring devices are very prominent, due to the vestigial sideband characteristic of the output waveform. The details of interpretation are included later; while the physical arrangement and description of the electrical characteristics are given here.

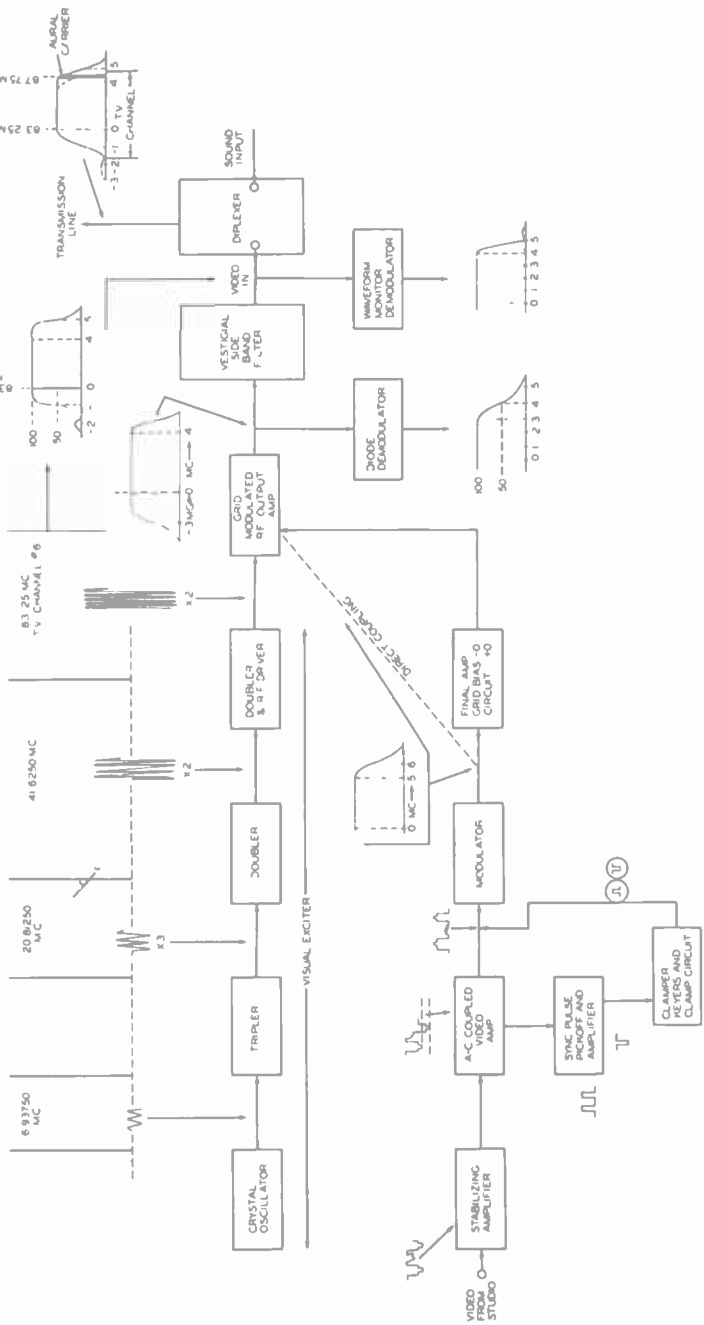


Fig. 8-1. Functional block diagram of a TV Channel 6 transmitter.

Two monitoring points with typical bandpass response are shown in Fig. 8-1. One is a diode demodulator at the output of the modulated final amplifier, and the other is a vestigial sideband demodulator at the output of the vestigial sideband filter. The typical output response of the ordinary diode demodulator is reduced to approximately 50% at 4 mc with a gradual roll-off from about 2.75 mc to 5 mc. The display observed on a picture monitor driven from this source will inherently lack sparkle or detail, due to high-frequency roll-off, and should be so interpreted by the operator. This diode curve is due to the partial cancellation of the lower sideband in the final amplifier, since the resultant addition of the upper and lower sidebands in the detector produces the typical curve shown.

Because of this characteristic, the ordinary diode detector cannot be used (for picture monitoring) at any point after the vestigial sideband filter, or, in low-level modulation, at any point past the first modulated stage, since the sharp attenuation of the lower sideband results in a worthless diode response curve for observing picture detail. Therefore, the vestigial sideband monitor shown connected at the output of the filter is a special insensitive-type receiver circuit for picture monitoring having the typical response shown in the illustration. This provides a longer flat-top response, while the sharp cutoff at the high end enables observation of any ringing effects in the picture. For waveform monitoring, the output of this demodulator is fed to a keyer circuit, then to an oscilloscope. The purpose of this keyer circuit (also termed *vibrator* or *chopper*) is to intermittently short-circuit the output of the detector, providing an additional line on the scope screen representative of zero output. Keep in mind the two components of the standard composite signal, namely, the DC and AC signal axis. The DC axis must be constant; the AC video signal axis is variable, depending on light or shade in the original sense. Periodic shorting of the demodulator produces a zero reference level representing no signal. The basic equivalent circuit of such waveform monitoring is shown in Fig. 8-2, illustrating application to measuring modulation characteristics. If an all-black video signal is fed into the transmitter with a sync pulse height S1 above pedestal level, the resultant scope pattern from such an arrangement is shown in 2. The ratio of the amplitudes S1 to E1 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 3. The ratio of the amplitudes of E3 to E2 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 within 10% of the amplitude of an all-black picture. When functioning properly, modern transmitters hold well within 5% in going from black to white.

The percent of variations under such conditions may be determined as follows:

$$\text{Blanking level variations} = \frac{(E2 - S2) - (E1 - S1)}{E1 - S1} \times 100\%$$

$$\text{Sync level variations} = \frac{E2 - E1}{E1} \times 100\%$$

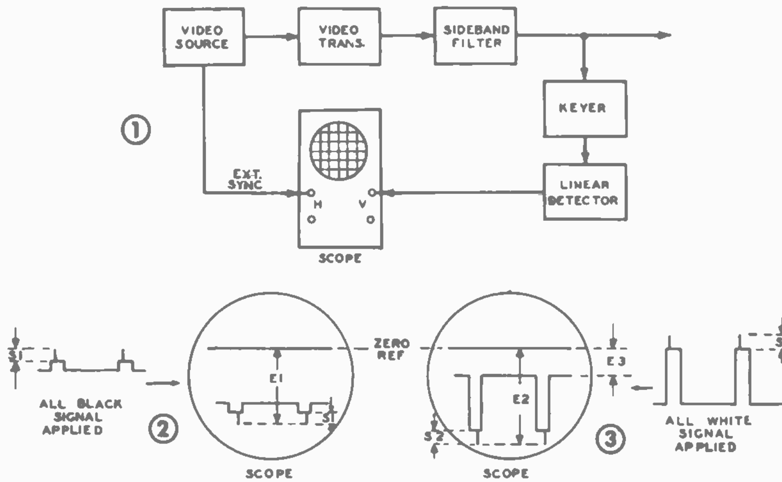


Fig. 8-2. Measuring modulation characteristics.

The preceding arrangement also enables the operator to set the maximum white level of the video signal to 12½% (±2½%) of the peak sync amplitude.

In actual practice the operator will find the picture and waveform monitors incorporating input selector switches so that monitoring is accomplished at points other than those shown in Fig. 8-1. Usually the switches provide for insertion of the monitors at the stabilizing amplifier output and modulator output. This permits observation of the signal at a sufficient number of points to aid in determining stages where trouble may occur. Some stations utilize an ordinary receiver monitor as an overall check, in which case precautions must be taken not to overload the receiver circuits from the high signal strength prevailing at the transmitter location.

At the present time, video transmitters have not reached the stage of development where overall frequency response is equal to that of studio equipment. Since distortion is additive, studio equipment must be operated with as wide a band as it is possible to obtain with the equipment used.

The vestigial sideband characteristic in itself is a source of picture distortion which, in any present type of demodulation system, produces slight leading whites and trailing smears upon transition from white to black regions of the picture. Such defects may be made very slight, however, in comparison to the advantage realized in gaining maximum use of the available frequency spectrum. The inherent distortion of vistigial sideband transmission can be minimized by predistortion of phase and amplitude characteristics in portions of the transmitter circuits or in the stabilizing amplifier. Where transmitters are concerned with color signals, the transmitter and average receiver characteristics are equalized, as described under Measurements.

The overall response up to the video modulator stage in the transmitter is essentially flat to 4.2 mc. From this point on, the response is a compromise in economic design of circuits and the inherent nature of the standard transmission signal. The final clamping point for DC reinsertion is ordinarily found at the grid of the modulator stage. This necessitates some form of direct coupling between modulator plate and modulated grid in order to maintain the DC component. The reader may wonder why the clamping action is not inserted at the grid of the modulated stage rather than the modulator grid, which requires direct coupling to maintain the direct current. This is the first point of compromise in design. The clamper *keying pulses* must have a greater peak-to-peak value than that of the actual clamping pulses and the video voltage applied to the clamped grid. If this were not the case, one of the clamper diodes might be brought into conduction during the video voltage signal, rather than the blanking interval. Any advantage that might be gained by clamping the grid of the modulated stage would be offset by the larger power-handling modulator stages required, which would increase initial cost beyond any possible advantage to warrant it. If the grid of the RF modulated stage were AC-coupled, approximately 60% greater signal amplitude (peak-to-peak would exist at that point than in the case of a DC-coupled arrangement.

To meet the standard transmission characteristics of negative modulation, an *increase* in light content of the signal must cause a *decrease* in the amplitude of the carrier wave. This requires that the grid-modulated radio-frequency stage receive a *black positive* video signal as indicated in 1 of Fig. 8-3. As the signal swings in the black (positive) direction, grid bias is decreased and plate current is increased, resulting in a greater amplitude of signal carrier envelope. As the video swings in the negative (white) direction, grid bias is increased, decreasing the plate current; and less amplitude of RF carrier envelope results. Thus the signal input to the modulator grids must always be black negative, as shown. In 2 of Fig. 8-3 is illustrated a typical transfer characteristic of the modulator stage. As the signal swings in the positive (white) direction, grid bias is reduced and modulator plate current increases. Increasing the plate current causes a greater voltage drop across the modulator load, reducing the voltage at the coupled point. This results in the familiar phase reversal of 180° between plate voltage and grid voltage swings. As the grid voltage is increased in the positive direction, the plate voltage coupled to the modulated grids goes in the negative direction. Also notice that there is a change in DC restoration potential accomplished by the clamper or restorer stage, so that pedestal and sync levels result at the same modulator plate current point in either all-black or all-white signal conditions.

The transfer characteristic of the modulated RF stage is also illustrated in Fig. 8-3. A grid-modulated stage is operated in Class-B as are any following linear amplifiers, being biased close to the cutoff point with no excitation. For negative modulation, when no signal is received, the radio-frequency excitation from the driver stage is sufficient to drive the plate current to its maximum value. For an all-white picture signal, the bias on the grids is maximum and plate current is reduced, except during the blanking and sync intervals, to a point at least 15% of maximum level and no more than 10%. For an all-black video signal, such as the application of pedestal and sync only, the bias on the grids is at minimum, resulting in maximum amplitude of carrier envelope.

In practice the quiescent grid bias is adjusted so that video excursion about that point maintains the output waveform over the linear portion of the grid-plate transfer characteristic curve. Excessive bias will push the operation down around the lower knee of the curve and result in compression of the signal representing white in the picture information. Insufficient bias will not allow full advantage of the linear portion of the curve without sync compression, since the resulting operation along the upper part of the curve will cause the sync region to fall on the bend of the curve unless the amplitude of the applied video is held to an unreasonably low level.

From the foregoing analysis of the video modulator action, the importance of proper clamping function may be observed. This DC restoration action at the modulator grids is fundamentally illustrated in 2 of Fig. 8-3. Fig. 8-4 illustrates the clamper action.

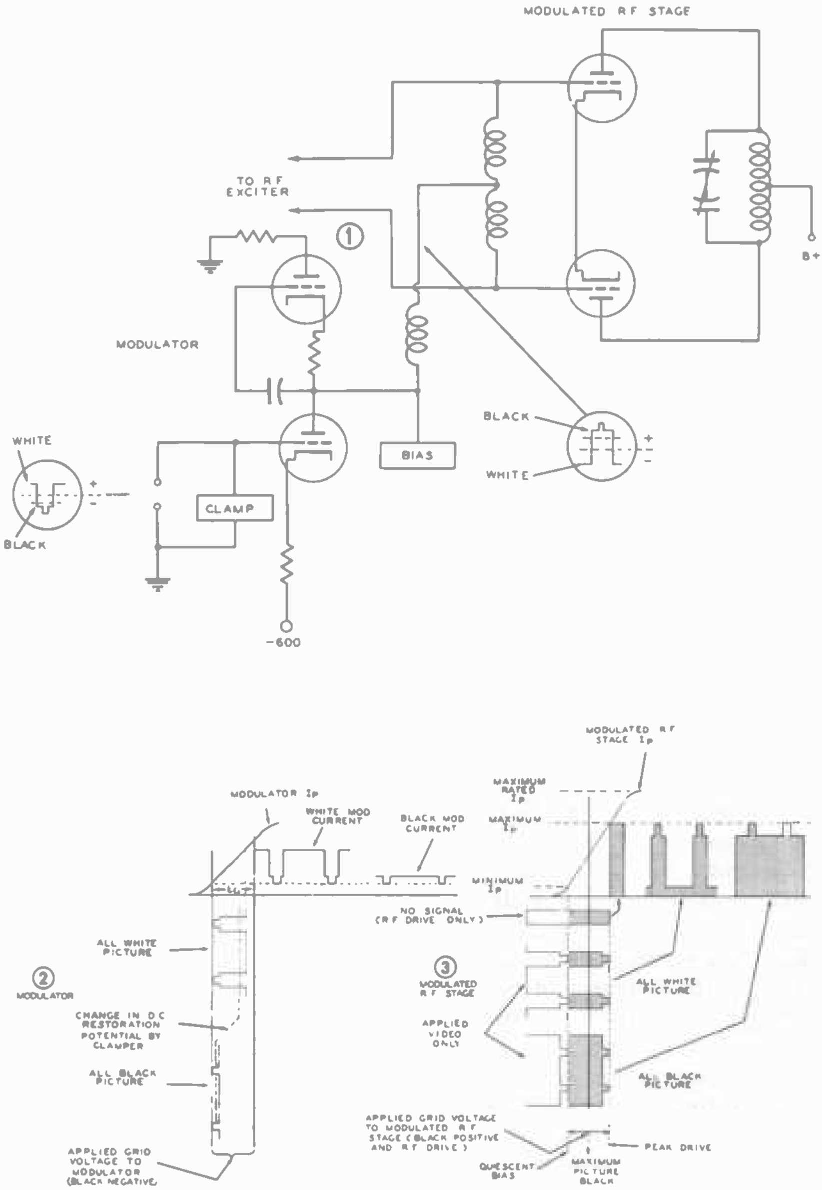


Fig. 8-3. Video modulation fundamentals.

The average AC axis for a symmetrical waveshape is as shown at points 1 and 2 in this figure, where an equal area of wave occurs above and below the zero axis. At point 3 is a video signal equivalent to an all-white signal. It is necessary for the clamper or DC restorer circuit to shift the AC axis in the positive direction to hold the pedestal and sync levels at the predetermined reference point. Compare this to 2 of Fig. 8-3. An all-black video signal is shown at point 4 in Fig. 8-4. Notice that the AC axis must be shifted in the negative direction to hold the peaks at the previously mentioned reference level. In this case the quiescent bias of the modulator grids is shifted in the negative direction. This shifting of the AC axis so that the reference level always occurs at the same point on the grid-voltage, plate-current transfer curve, re-

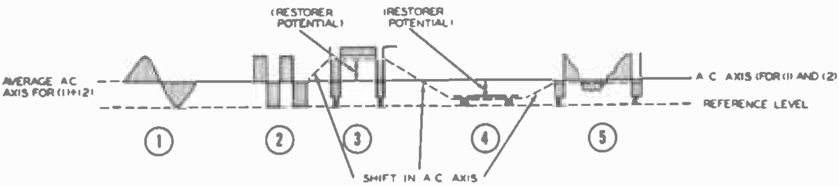


Fig. 8-4. Basic clamper operation.

gardless of waveshape, is equivalent to restoring the DC signal component. As shown at point 5 in Fig. 8-4, a video signal consisting of an exact balance between black and white would have its AC axis very nearly equal to that of a symmetrical waveshape. The slight difference occurs due to the *setup* of the video maximum black to pedestal level, and the existence of the sync peak level. (To eliminate any possible confusion it would be helpful at this point for the reader to review Section 2, where amplitude calibration and maintenance of levels are discussed.) In Fig. 8-5 is a representation of a CRO display set to indicate depth of carrier modulation with the chopper establishing the zero carrier line. Note that the actual video-to-sync ratio transmitted is identical to that received from the studio. The 25% sync refers *only* to 25% of the carrier power, *not* to the percentage of the transmitted composite signal.

In the process of amplitude modulation by the grid-bias method, a small amount of incidental phase modulation may be introduced. Extra precautions are taken in the initial design of transmitters to minimize this effect, and the operator must also observe precise relationships of adjustment to maintain operation within the allowable phase modulation in the picture signal. The net effect of incidental phase modulation is greater in the vestigial sideband type of carrier than would be the case in double sideband. This is so because in a double sideband amplitude-detector the extra set of sidebands produced cancel out, whereas they add directly to the AM sideband in a single sideband detector. Observation of the vestigial sideband characteristic reveals that lower video frequencies up to 750 kc (0.75 mc) are actually transmitted double-sideband, whereas higher video frequencies from 0.75 mc to the upper limit are transmitted vestigially. This is one of the limitations on video frequency characteristics which are fixed by transmission standards, and it has a direct bearing on the amount of allowable incidental phase modulation in the transmitted signal.

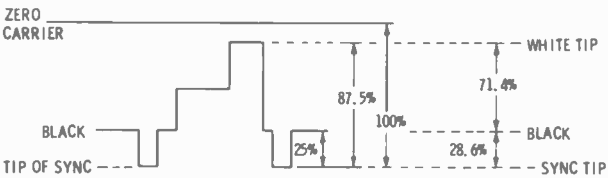


Fig. 8-5. Representation of a CRO display set to indicate depth of carrier modulation. Actual video-to-sync ratio transmitted is same as received, unless compression occurs.

The most important point for the operator to understand is how to determine in a practical manner this allowable amount. An intercarrier-type receiver in good working order provides a most reliable basis for judgment. In this receiver an intermediate frequency of 4.5 mc is obtained as a beat between video and audio carriers. If the visual carrier contains incidental phase modulation from the picture signal, buzz and noise will result from picture modulation in the sound portion of the receiver. This test must assume that other factors that would also produce noise in an intercarrier receiver, such as overmodulation in the white direction, are not present. In general, it may be understood that picture distortion caused by phase modulation will be negligible if sound distortion in an intercarrier-type receiver is also negligible from this effect.

Picture-phase modulation might be caused by any condition resulting in an undue amount of RF feedback, such as would occur in stages that are improperly neutralized.

Video currents flowing in the grid, screen, and plate circuits of amplifiers will produce a corresponding voltage drop across the internal impedance of the associated power supplies. This drop obviously affects the DC potential applied to the electrodes and will result in picture distortion. Thus power supplies for video circuits are designed with extremely low internal impedances, and it is found that many tubes are paralleled in regulator circuits not only to take advantage of their current-handling capabilities but also to decrease this impedance value to an absolute minimum.

The screen voltages of RF stages are also usually regulated by electronic means.

8-2. VISUAL TRANSMITTER POWER OUTPUT

The video transmitter is never required to develop an average power output greater than the average of the combined pedestal and sync levels. The rating of a video transmitter is given in peak power capability, and the measurement of this output power must be determined under these conditions. Thus, in practice the average power of a standard black signal is found, and the peak value is computed from this measurement.

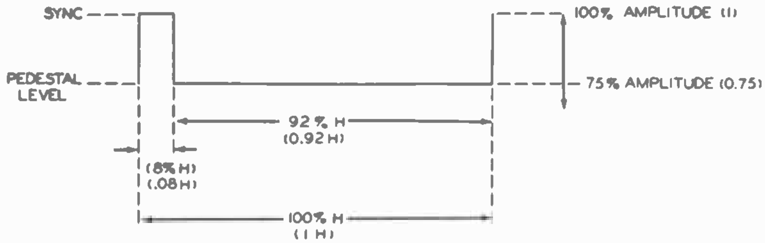
A dummy load and RF wattmeter consist of a resistor element for terminating the transmission line in its characteristic impedance and a current-indicating meter for measuring the amount of power dissipated. The power dissipating section (dummy load) consists of a resistor unit immersed in a liquid which is cooled by air in the low-power units, by tap water in the medium-power units, and by forced water in high-power units. In order to prevent excessive use of tap water during the time the RF power is at a low level, a water saver is used in most instances. This consists of a thermostatically controlled solenoid valve which allows the water to flow only when needed.

The power-measuring section consists of a short length of transmission line (Thruline), a meter, and a wattmeter element. A socket is provided on the side of the transmission-line coupling section to accommodate a calibrated wattmeter element which, when coupled to the transmission line, develops a DC current approximately proportional to the forward-wave voltage across the load resistor. This current is supplied to a remote meter calibrated to indicate directly the power dissipated in the load.

The wattmeter element is a reflectometer which consists of a coupling loop, a crystal detector and a filter network. The wattmeter element may be rotated 180° in the transmission-line housing. This permits it to indicate the incident power to the load or the reflected power from the load.

The transmitter is operated into this load for about fifteen minutes to obtain equilibrium of temperature. This is the *average* power output with the transmitter modulated by a standard black signal. This means that the modulation consists only of pedestal and sync voltages, with pedestal level carefully adjusted to 75% of the peak output. To obtain the *peak* power output, the above-measured average power level is multiplied by the factor 1.68. This measurement is made at the output of the vestigial sideband filter whenever used. In low-level modulated transmitters, the measurement is made at the output of the final linear amplifier.

The reader should understand how this multiplying factor of 1.68 is obtained, since the visual transmitter power must be specified in peak value. Fig. 8-6 is presented here to aid in this understanding. A standard black signal is a transmission of sync pulses with the blanking (pedestal) voltage carefully adjusted to 75% of peak sync value. In this standard signal the sync pulse occupies 8% of the line interval, or 0.08H. The pedestal level then occupies the remaining line interval of 92%, or 0.92H. The entire line interval (100% or 1H) is from the leading edge of one H sync pulse to the leading edge of the next H sync pulse. These relation-



Pulse	Degree of Modulation (Amplitude)	Corresponding Power Level in % of Peak Power Level (I^2R)	Time Interval in % of H	Ratio of Av/Pk Power
Sync.	100%	$(1)^2$ X 100 = 100%	100 X 0.08H = 8%	8%
Pedestal	75%	$(.75)^2$ X 100 = 56%	56 X 0.92H = 51.5%	+51.5%
				59.5%

Since: Ratio of Average Power/Peak Power = 59½% or 0.595

Then: Ratio of Peak Power/Average Power = 1/0.595 = 1.68

Therefore the multiplying factor to obtain peak power output from average power output of a standard black video signal is 1.68.

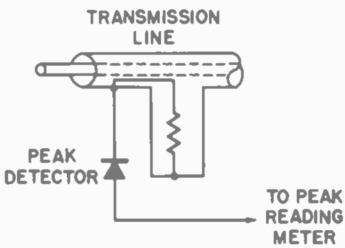
Fig. 8-6. Relationship between peak power and average power.

ships are shown in the diagram. The derivation of average to peak power ratio is also shown, as well as the conversion of this relationship to peak to average power. Remember that *average power* is related to I^2, or E^2/R, where current and voltage in rms values are those measured through or across a known resistance. The relative *peak power* is then dependent on the nature or shape of the power curve.

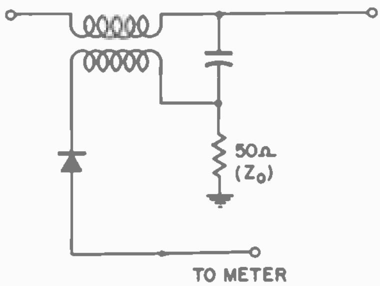
Following the data given with the drawing, note that the peak sync pulse is the level at which 100% modulation of the carrier wave occurs, and therefore the corresponding power level in percent of peak power level is the same, or 100%. The duration of this power level in terms of the line interval (H) is 0.08H, or 8%. The pedestal level is that level which produces 75% amplitude modulation of the carrier wave. Since the corresponding power level for any given value of resistance depends on the square of the current or voltage, the average power level in this case is 56%, as shown. Since this power level occurs over 92% of the line interval (0.92H), the pedestal power interval is 51.5%. The total of the time intervals of the respective powers (8%) plus (51.5%) is then the ratio of the average to peak powers. Therefore, to find the ratio of the peak to average powers, the reciprocal of the above is used.

It may also be computed that the ratio of the rms voltage or current of the carrier wave during H sync interval to the rms voltage or current of the carrier wave during the entire H time is equal to 1.295. The square of this factor is $(1.295)^2 = 1.68$.

To meet the requirements of the FCC Rules & Regulations, all installations must include some indicating device that shows peak power output of the video transmitter during the operating schedule. This meter is initially calibrated by transmitting a known power as determined, for example, by the previously described method. The indicator must then be checked at periodic intervals by rerunning the dummy-load power computation and comparing the peak indicator-meter reading with the computed power output.



(A) Coupling arrangement.



(B) Equivalent circuit.

Fig. 8-7. Directional coupling principle.

The peak power indicator usually takes the form of a *reflectometer*. This is a combination of a directional coupling device and a peak-reading diode detector circuit. This indicator provides a constant check on power output as well as showing condition of transmission line and antenna system as they affect standing waves on the line.

A directional coupler is shown in Fig. 8-7. The loop may have both magnetic and capacitive coupling to the transmission line. The capacitive coupling is small, with a large reactance at the carrier frequency. Therefore, the current that flows through the resistor is in quadrature (90°) with the line current. The loop and resistor voltage drops are in series. For a wave traveling in one direction, the voltage across the transmission line and the current in the line are in phase at a magnitude set by the characteristic impedance of the line. Since both the coupled voltage and the resistor voltage are in quadrature with line current, the loop voltage is in phase with the resistor voltage, and the addition represents the directional-coupler output voltage.

In the event of a standing wave (wave on transmission line from opposite direction), the loop-induced voltage is out of phase with the resistor voltage drop. Now, if the loop is adjusted so that these voltage drops are made equal, the coupler output voltage will be zero. In this way the directional coupler can distinguish between waves of opposite directional flow. It can be calibrated to measure power output, or, by comparing voltages of opposite current flow, it can measure load mismatch or voltage standing-wave ratio.

It is necessary to calibrate, then ensure maintenance of correct calibration of the reflectometer in terms of peak power output. This is calibrated against the dummy load and RF wattmeter measurement described before.

8-3. GENERAL TUNING PROCEDURES

The tuning of television transmitters is not unduly complex, but the procedures are necessarily more involved and interdependent than is the case of the conventional audio transmitter. It is necessary for the AM operator to gear his thinking to the requirements of circuit function in relation to the nature of the standard television signal.

For the basic adjustment of a grid-modulated RF stage, 3 of Fig. 8-3 should be observed during the following discussion. This stage operates essentially as a Class-B amplifier with the fixed bias such that the tubes are operated near plate-current cutoff. The general procedure is as follows:

1. The modulated RF amplifier grid-bias is adjusted without RF drive or video signal to a point allowing only a small plate current to flow; in other words, it is adjusted to very near the cutoff value. This fixes the lowest point of operation along the most linear portion of the transfer curve, shown as minimum plate current in 3 of Fig. 8-3. This is the quiescent or static bias of the stage and varies with the tube and circuit conditions of the particular transmitter. Normal plate voltage and loading must be used on the stage during this adjustment.
2. The RF drive (no video) is increased sufficiently to drive the plate current to the upper knee of the transfer curve. Note that the indicated minimum and maximum values are the

operating values of plate current. In practice, the maximum operating value is approximately one-half the maximum *rated* plate current of the particular tubes used in the modulated RF stage. In general, therefore, the RF drive is increased to a value about one-half that required to drive the tubes to the *maximum rated* plate current.

3. A maximum white video signal of approximately 30% sync and 70% video is applied to the modulators. The video gain is advanced until the modulation envelope shows white modulation between 15% and 10%. The monitoring device may be either a diode pickup and a scope with chopper reference line or a special RF waveform-analyzer using a calibrated screen. It is noted from 3 of Fig. 8-3 that application of a white signal produces negative modulation from the condition of the maximum carrier amplitude under no-signal conditions. This is to say that with proper RF drive applied and *no* video signal the peak output of the transmitter prevails. Also notice from Fig. 8-3 that when video is applied the adjustment is such that the sync peaks of the *applied* video signal fall at the quiescent bias (minimum operating plate-current) of the tube. Now considering sync tips only, the carrier amplitude is the same as with RF drive only, with no signal, or maximum value. At the blanking (pedestal) level, the bias is increased, and plate current decreases by that amount. The large negative swing of the picture voltage then increases the bias still further, and the carrier amplitude accordingly decreases to the minimum plate-current value. The same reasoning is applied to the all-black signal (or sync and blanking levels only), and the plate current (hence carrier amplitude) is reduced to the pedestal level over the 92% of the line interval. The sync tips represent 100% modulation, the blanking level 75% modulation, and the maximum white level is 10 to 15% modulation (12.5% nominal).

The preceding has considered adjustment of the modulated stage, and it has been assumed that the exciter supplying the drive to this stage has been properly tuned by conventional methods. The remaining tuning procedures concern the output circuit of the modulated stage and any following linear RF amplifiers where this method is used. If the preceding modulation adjustment is carried out before adjustment of following linear amplifiers, the monitor pickup used must be from the modulated stage.

The TV operator is concerned with circuits in which both plate and grid circuits are tuned, commonly referred to as double-tuned circuits in an overcoupled condition to achieve adequate power output with satisfactory bandwidth. Since tuning procedures of such circuits are unconventional, a brief review from fundamental theory is in order.

Fig. 8-8 illustrates a double-tuned circuit arrangement and response curves corresponding to several factors or conditions. When the coefficient of coupling is small, the secondary response is small for a constant-current AC in the primary, and it has the typical shape of a single-peak resonance curve. As the coupling is increased, the secondary response rises in amplitude and broadens in response. If this process is continued until the resistance that the secondary couples back into the primary is just equal to the primary resistance at resonance, the point of *critical coupling* is reached. At this point the secondary response attains its maximum possible amplitude. The shape factor (S) of this curve is

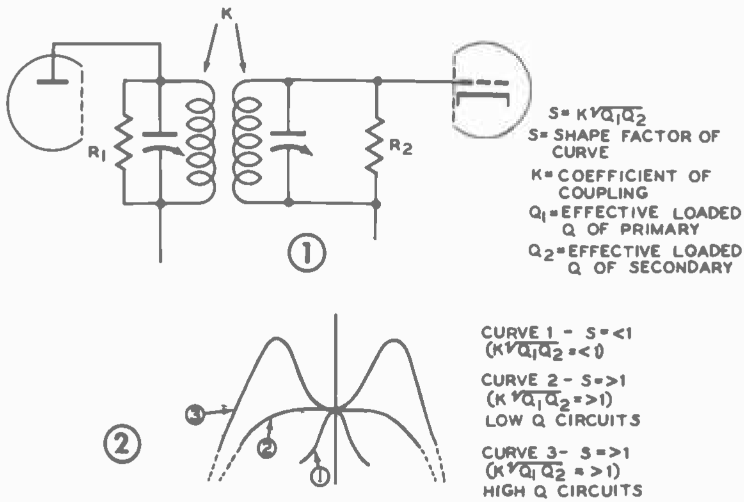


Fig. 8-8. Variable factors of double-tuned circuits.

still less than I, even with high-Q circuits, as indicated in the diagram in Fig. 8-8. The primary or secondary Q is the ratio of the energy stored in that circuit to the energy dissipated per cycle (X/R). The transmitter operator tunes the primary and secondary for maximum secondary response, indicating resonance and optimum loading condition simultaneously.

From this point on, conditions differ from conventional AM circuit action. When the coupling is tightened beyond the critical value, the secondary response begins to show double humps. When this occurs, the shape factor becomes greater than 1, even with low-Q circuits. For a given circuit, the peaks of the humps become greater in amplitude and farther apart as the coupling is increased. Thus, the peaks may become quite pronounced with a decided valley between them, as shown by curve 3 in Fig. 8-8. This is typical of a tightly overcoupled circuit with high circuit Q and a resulting shape factor much greater than 1.

In order to obtain the more desirable response curve shown by 2 in Fig. 8-8, we may examine the possibilities afforded the operator. The shape factor depends on the coupling and circuit Q. The Q in itself will depend on loading of the circuit. The operator has no control over the "designed Q." Therefore he has only two variables: (1) coupling and (2) loading.

In practice, circuit constants have been designed so that the shape of the response curve will be correct when the circuit is adjusted for optimum bandpass characteristics. For a given value of coupling in a given overcoupled circuit, increasing the secondary load (by *decreasing* the effective value of R2 in Fig. 8-8) will decrease the amplitude of the humps and at the same time provide a more flat-topped response as shown by curve 2 in Fig. 8-8.

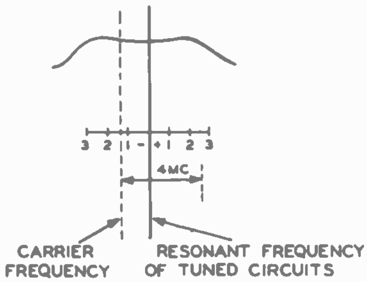


Fig. 8-9. One method of obtaining the required video bandpass while maintaining the proper attenuation of the lower sideband.

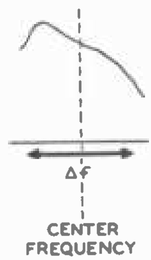


Fig. 8-10. Resultant frequency response curve of an over-coupled double-tuned circuit when the primary and secondary are tuned to different frequencies.

The primary and secondary of such a circuit must both be tuned *on resonance*. This resonant frequency need not be the actual carrier frequency; indeed, this practice is seldom used in video transmitters. The reason is that the tuned RF circuits are adjusted so that the lower sidebands of the video passband is attenuated by the required amount. Thus the carrier frequency is actually lower than the resonant frequency by about 1.5 mc, as illustrated in Fig. 8-9. This is accomplished by adjusting the resonant frequency of the tuned circuits of the modulated stage and RF linear amplifier to a value higher than the carrier frequency.

Since the primary and secondary of the double-tuned, overcoupled circuit must be tuned to resonance, the operator cannot follow the conventional practice of tuning for maximum power output as in AM circuits or broad-band single-tuned circuits. What actually happens when this is attempted is that the primary and secondary are tuned to different frequencies in order to find a load impedance favorable to maximum power output. The result is shown in Fig. 8-10. When the primary and secondary are tuned to the same frequency, resulting in a symmetrical bandpass characteristic, the input impedance at the center of the band (which determines the maximum power the tube can develop) is at a minimum value.

Therefore, if he is tuning strictly by the meter method without the aid of an oscilloscope, the operator performs adjustments with the above characteristic in mind of obtaining minimum load impedance for a given value of coupling. When minimum load impedance on the driver is obtained, a minimum peak in grid current of the driven stage will occur upon rocking the primary capacitor back and forth through resonance. Upon initial adjustment the plate voltage is lowered on the driven stage, and the primary tuning adjustment is rocked through resonance as indicated by the grid current meter in the driven stage. If this stage uses a tetrode tube, more accurate indication may be observed by watching the screen-current meter for the peak screen current. When primary resonance is found by this procedure, the secondary is adjusted so that a minimum peak in grid or screen current occurs as the primary is varied back and forth through resonance. This ensures that both primary and secondary are tuned to the same frequency.

In stages using variable coupling, it may occur that the driver load-impedance from the foregoing procedure is too low. This will be revealed by excessive plate current compared to the effective power output, indicating high internal anode power dissipation. Under these conditions the bandwidth is usually greater than required. The situation is remedied by using reduced coupling and repeating the above procedure. This increases the load impedance, decreasing the tube loading condition and hence reducing the plate current for a given power output.

It should be noted here that, under some conditions, increased driving power may result from *reducing* rather than increasing the coupling, as is necessary in conventional AM transmitters. This is a characteristic of double-tuned, overcoupled RF transformers. It should also be borne in mind that when a single-tuned broadband circuit is used between stages, the circuit is tuned in the conventional way for maximum grid current in the driven stage.

Coupling adjustments on ordinary link-coupled circuits are obvious; moving the links farther apart decreases the coupling, and vice versa. Adjustment of circuits using resonant lines is not so obvious, however. In general, it should be understood that moving the connection on the resonant line toward the open end results in increased loading. For example, consider the common case of a driver stage that is coupled to the following stage operating grounded-grid by tapping onto the cathode resonant line. The driver stage would be loaded more heavily by adjusting the point of cathode connection toward the open end of the line, and the loading would be *decreased* by moving this connection toward the cathode terminal of the driven amplifier. In the instance of loading a final stage to the transmission line, the final amplifier is loaded more heavily by moving the transmission-line connections toward the open end of the final plate resonant-line output circuit.

For a properly tuned overcoupled circuit, bandwidth (separation between humps in the response curve) is determined primarily by the degree of coupling, whereas flatness across the top of the response curve is affected mostly by loading.

After the circuit has been properly tuned, the entire resonant frequency may be shifted the required amount above the carrier frequency by using a sweep generator, markers, and a scope.

8-4. TYPICAL TRANSMITTER OPERATIONS

Transmitter operations may be roughly divided as follows: pre-sign on; regular operating day of program transmission; shut-down period after sign off; and preventive maintenance, which also takes place after sign-off.

The sign-on man generally arrives at the transmitter an hour or so before the first test pattern signal is to be put on the air. In systems using water-cooled tubes in the high-level stages, the water pumps are usually started before any other operation. In air-cooled systems, the blowers are ordinarily actuated upon application of filament voltages. Any adjustable autotransformer must be set on the tap giving the proper primary voltage indication for the particular installation.

Modern transmitter circuits employ an orderly control-circuit system for two purposes: (1) to prevent improper transmitter functioning as to overloaded circuits or inadequate time-delays for application of high voltages, and (2) to protect the operating personnel from contacting high-voltage terminals. The former function also serves to prevent certain applications of potentials to elements not receiving a normal flow of cooling medium, such as water or forced air. The operator must be thoroughly familiar with the functioning and operating sequence of control at any particular installation.

A typical sequence of operation is as follows. The Start button is pressed, which applies filament voltages to the tubes and starts a number of other relay circuits to functioning. Blower motors are started, and, until the air stream is of sufficient strength to actuate mercury switches on the air vanes, full filament voltage is not applied. At the same instant, time-delay relay motors are started. These do not close the high-voltage circuits until a specified time has elapsed, such as 30 seconds or one minute. Some transmitters employ switches that automatically apply high voltages upon timing-out of the delay relay. When the transmitter is first placed on the air, however, these switches are normally set to the manual position so that other adjustments may be made or checked before this occurs. Filament voltages should be checked (as indicated by their respective meters) and any adjustments necessary to bring the potentials to normal value should be made. Door interlock switches also prevent application of high voltage if any door is open in a cubicle containing high voltage. An open door also usually actuates grounding switches that short the high-voltage supplies to ground so that large capacitors cannot discharge to ground through an operator's body.

After all filament meters have been checked and normal operation is obtained, the low voltages are usually applied to the video amplifier stages and RF exciter stages. This permits checking their operation before application of high voltages to the final stage or series of high-level linear amplifiers. Any necessary adjustments such as touch-up of tuning controls are made, and the grid current of any high-level stage is observed to ascertain normal driving power to that stage before application of the final high voltage. Fig. 8-11 illustrates the aural and visual panel controls and meters for the RCA TT-50AH transmitter, as installed at WISH-TV.

When the operator is assured that the transmitter is functioning properly into the antenna system, he generally removes the high voltage and places the dummy load on the final in lieu of the antenna (Fig. 8-12). He is then in a position to check performance with test signals from either equipment at the transmitter or from the studio line. These tests include test-pattern signals, checking of waveform from the studio and after passing through the transmitter, adjustment of video and audio levels, etc.

Due to the nature of the FM aural transmission, some operators are at first confused on obtaining proper interpretation of the aural modulation monitor-meter readings. Recall that pre-emphasis of the audio spectrum in accordance with FCC standards is used at the input of the aural transmitter. The monitor circuit which provides audio voltage to drive the sound-monitoring amplifiers then includes a de-emphasis circuit to restore the audio amplitude-frequency curve to its normal value. It must be remembered, however, that some modulation meters show percent of modulation in terms of this pre-emphasis curve, which means that about 17 db less audio is required at 15,000-cps pure tone to modulate the transmitter 100% than at any pure tone below approximately 400 cps. This action in terms of average program material should be taken into consideration by the operator. A program that has a number of highs in the signal content should modulate the transmitter 100%. However, film or recordings often contain sound which is definitely lacking in highs, and 100% modulation will cause noticeable distortion in lows, due to the overloaded condition in the audio amplifiers of the station and especially in the receiver circuits following de-emphasis. Voice transmission should seldom exceed 30 to 40% modulation. The studio operator may be peaking his meter close to the 100% indi-

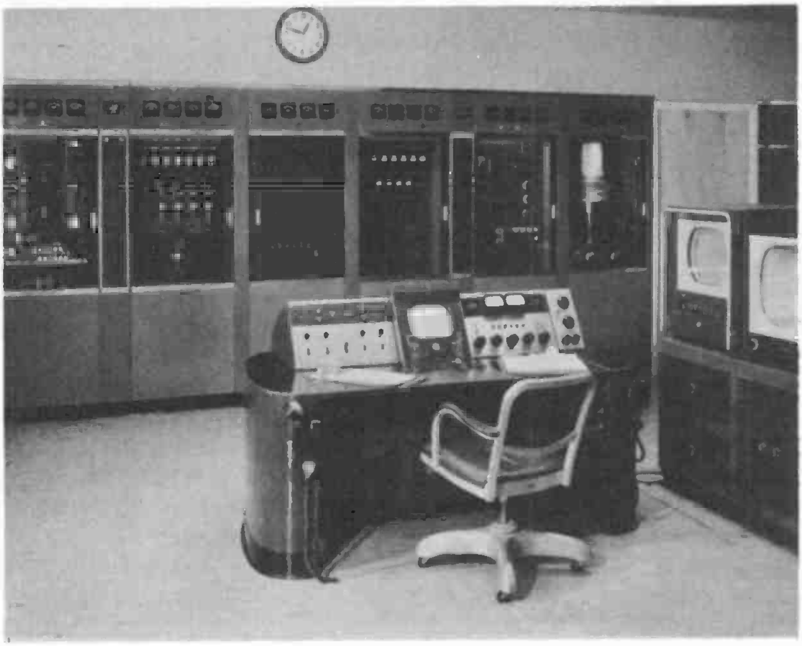
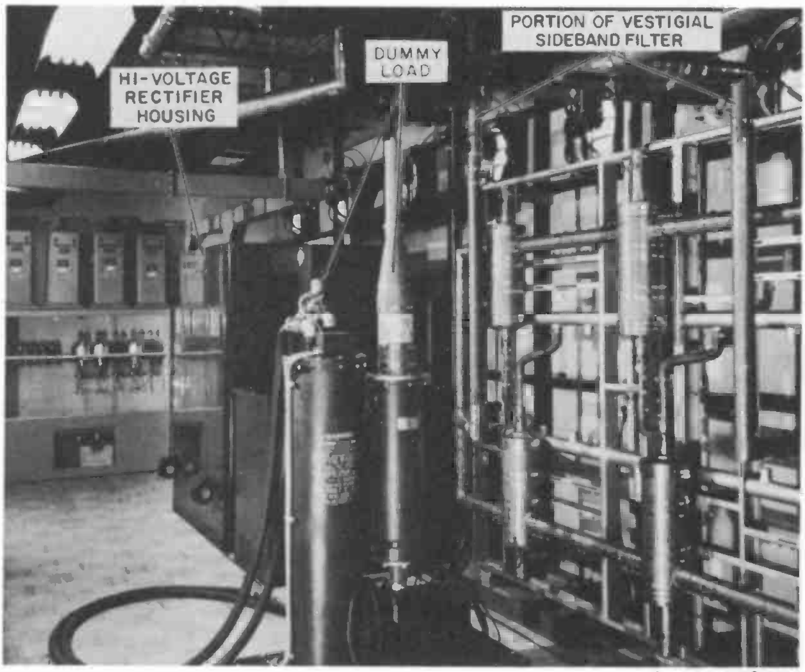


Fig. 8-11. Transmitter and control console at WISH-TV.

cation, but the transmitter modulation-indication cannot be expected to follow the studio meter under all varying program conditions.

While the transmitter is on the air, a competent operator will be continually alert to picture quality, waveform level, amount of *setup* and proper sync-to-blanking ratio, meter readings, temperatures, and even to his *sense of smell*. As he observes the meters and makes adjustments on the line-voltage autotransformer, he should be sensitive to the characteristic operating odors of resistors. relay solenoids, capacitors and transformers. This practice often indicates the general location of impending trouble, even if visual observation is impossible because of the location of the component. (Meeting emergencies is more fully discussed in Section 8-5.) Meter readings must be recorded on the transmitter log at 30-minute intervals, or as required by current FCC Rules.

After sign-off the high voltage is removed and the rear doors are opened for visual observation of all components. The operator should, at this time, become thoroughly familiar with the *feel* of filament, grid and anode connections of high-level stages, important capacitors, and other components where temperature indi-



Courtesy WISH-TV

Fig. 8-12. Dummy load (foreground) and associated station equipment.

cation as revealed by this feeling process is important in case of trouble or impending failure. He should *always* ascertain that interlock switches and high-voltage shorting relays are properly functioning before he touches any component.

After shutting off the low-voltage and filament potentials, blower motors or water-cooling systems sometimes continue to operate by keep-alive relays for a specified time (generally 4 to 7 minutes) to cool the high-level tubes. This is the end of the operating day and should be the start of any preventive maintenance schedules.

8-5. EMERGENCY PROCEDURES

The variety of corrective measures that might be called for in getting a transmitter back on the air or in clearing defective pictures is so great that a natural limitation is immediately placed upon the thoroughness of presentation in this section. The treatment is therefore very general in nature.

The most important phase of meeting emergencies is in training the entire staff to be *mentally prepared* for corrective procedures. Due to the complexity and unusual expense, complete standby transmitters for emergency use are a rarity in TV broadcasting, particularly in smaller stations. It is the duty of each individual operator to thoroughly study the technical aspects of his particular installation, and to prepare himself to analyze malfunctioning equipment with a certain coolness and deliberateness. Such a psychological preparedness actually minimizes the time necessary to clear transmitter faults. The chief engineer or other supervisory personnel of every station should conduct classes during off-hours in which most likely and typical faults are simulated for observation of effect on meter readings, waveform and picture content, etc. Transmitters that have been on the air over a period of one year will inevitably have certain pecularities that are revealed in case histories and these should be pointed out to all operators.

Within a very few seconds after trouble has occurred the operator should be able to analyze the fault as being in one of the following general classes: control circuits or power supplies, video amplifiers used as incoming line amplifiers from studio or network, video modulator section, or radio frequency section. If trouble is in the sound, he will immediately place the possible source as either the line amplifier from the studio, modulator section of the aural transmitter, or the frequency multiplier and final RF stages.

It is most helpful upon the first instant of trouble in picture or sound (or both) to observe the respective frequency and modulation monitors. This is most important for the following reasons:

(1) A picture, for example, may disappear from the picture monitor screen, yet still be transmitted over the air. In this instance the picture monitor itself is obviously at fault. If this should be the case, the modulation monitor showing the modulated RF envelope will be indicating as usual, and a spare picture monitor is merely substituted for the defective one.

(2) Assume that the picture disappears from the picture monitor. A quick glance at the modulation monitor CRO shows

no modulation taking place. At this time it is possible to get a preliminary idea as to possible trouble by observing the RF input indication to the frequency monitor. For example, in some monitors, any deviation from the normal RF input is indicated by a lamp. If this lamp indicates a fault in the RF input level, the operator has a pretty fair hunch that the trouble is in the RF portion of the transmitter. If the RF indication to the monitor is normal, the operator should suspect the video section (which includes the stabilizing amplifier on the incoming line, video amplifier or modulator stages in the transmitter) or no incoming signal from the studio. This is quickly checked in most installations by a switch on the control panel which places a monitor across the incoming line terminations, output of stabilizing amplifier, or output of modulator stage in transmitter. Thus the signal may be traced in this manner to quickly isolate the faulty stage.

In instances where the transmitter goes off the air either due to tripping of the overload relays or failure of a power supply, the previous procedure is obviously unnecessary. Visual observation of the transmitter rectifier tubes, overload relay indicators and meters is the initial act of the operator. He then mentally analyzes the evidence and decides what is necessary to place the unit back in operation. Overload relays or thermal switches may have to be reset. If this results in another quick shut-down of the transmitter, the fault must be cleared up before the high voltage is reapplied.

Sometimes visible or audible arcing occurs to give an indication of the general stage being overloaded. If the arcing is not visible, aural perception is usually sufficient to tell the operator which rear doors to open to observe for visible signs such as blackened spots on the frame next to a capacitor or high-voltage terminal. Insulators must be observed for cracks or signs of breakdown in this case, and high-voltage leads to tubes or components should be examined for bad insulation. If the arcing cannot be located by either of the above methods, it will be necessary to carry out an emergency procedure which must be exercised with the utmost caution and preferably with another operator standing by.

This procedure consists of opening the rear doors of the suspected unit, strapping the interlock circuits for that cubicle closed, and applying high voltage while the operator watches for the point of arcing. To do this it is simply necessary to jumper the proper terminal numbers associated with that particular door lock so that voltage may be applied with the door open. *This is an emergency procedure* only; it is never done except when absolutely necessary.

The operator should be so familiar with the control-circuit diagram that he can locate terminal numbers with the minimum of delay. For example, it would be entirely possible for the contacts on a time-delay relay to open up, either from improper adjustment, dirt or corrosion between contacts, or a faulty relay itself. This would be indicated by some light usually on the control panel designated, for example as READY. This means that the time-delay interval has expired after the transmitter is turned on, and the high-voltage circuit is capable of being energized when the operator is ready. If this light should go out, the time-delay relay would be one possible cause, and the operator should be able to locate the proper terminal board numbers to jumper for this emergency. This, of course, is only one example of many possibilities.

Overloads which trip the AC overload indicators are usually caused by mercury-vapor rectifiers arcing back. Many transmitters employ arc-back indicators on each rectifier tube, which indicate this reverse-firing condition so that that particular tube may be replaced. If such indicators are not used, it is wise to replace the entire component of mercury-vapor tubes with rectifiers known to be good, preheated, and air-tested. Proper preventive maintenance usually prevents this occurrence.

Some transmitters have built-in emergency provisions for occurrences such as improper clamping of signals in the video modulator section. This is the case in the GE TT-10-A. Several different types of trouble may cause the modulator to stop functioning in its normal, clamped manner. One instance is an input signal (from the studio) that is defective in certain particular ways. Another instance is failure of some of the tubes or components in the sync-operating and pulse-forming part of the modulator.

It should be noted, for example, that as long as the failure does not involve the video amplifier stages in the transmitter, operation may be resumed in the manner designated as emergency operation. This emergency operating procedure is outlined by General Electric as follows:

(1) If the input sync-voltage is too low, there will not be enough sync to be properly separated, and keying pulses will be formed in an erratic manner. Note that lack of sufficient input sync-voltage results from either too low a total peak-to-peak composite signal-input (considerably below one volt) or too low a sync percentage, even though the peak-to-peak value of the composite signal is one volt or greater. (GE specifications call for at least 10% sync at any input voltage over one volt.)

(2) If the incoming signal has back porches narrower than standard, or split pulses resulting in narrow slots in the sync going down to black, normal operation will not be obtained. Notice that certain other types of defective signals such as poor low-frequency response or hum are improved by the clamp operation of the modulator.

If, in the previous transmitter, such a failure occurs, the modulator may be switched to Emergency Operation. In this change, simple diode DC insertion on sync peaks is substituted for the back-porch clamp-type of normal operation. The switch from normal to emergency operation is accomplished by changing the two 6AL5 diodes from their Normal to their Emergency sockets as indicated by the front-panel marking on the modulator. The switch is left in the Clamp position. Readjustment of the RF Gain, Sync, and Visual PGM controls on the control panel will then be required.

On Emergency Operation, the two stages where the diodes are used for DC insertion are being operated far outside of the normal grid-resistance ratings of the tubes. Therefore it is recommended that the Emergency condition be used no longer than absolutely necessary. It is necessary to repair whatever caused the switch and return to clamp operation as soon as possible.

Troubles in RF stages of transmitters may generally be isolated by observing the meter readings of the individual stages. For example the stage nearest the oscillator showing lack of proper grid current indicates insufficient drive from the preceding stage, or a defective tube or component in the observed circuit. Tubes are always the first component to be suspected.

A word of caution is in order at this point. When first placing a transmitter on the air at the start of operations, and during preliminary overall checks on the sideband analyzer or waveform indicator, an indication of complete detuning might result from a defective sweep generator or indicating device. Should the RF waveform be defective, yet all meters are indicating "on the nose," the operator should first check his test equipment before suspecting the transmitter. This is done by checking the output trace of the sweep generator on a scope that is known to be good. If the traces are normal at the terminations and inside the sweep generator, and the RF waveform is defective as displayed on the substitute scope, the trouble may be assumed to be in the RF stages of the transmitter. Most generally, this results in abnormal meter reading on the RF stages.

The same technique should be observed if, at any time during the operating day, the frequency meter for either visual or aural transmitter indicates a frequency outside the authorized deviation. Remember that the station monitor is a secondary standard, and before suspecting the transmitter, the frequency monitor itself must be ascertained to be in proper working order. This may be done by checking with a commercial primary-standard frequency-measuring service authorized by the FCC. When the station monitor has been calibrated against this primary standard, as it should be whenever its operation is suspected, then the transmitter may be adjusted accordingly. The monitor is usually checked with such a service once a month as part of routine maintenance procedures.

Push-pull tubes in RF linear amplifier stages should have their plate or cathode currents balanced within 10%. When an unbalance greater than this amount is revealed by meter readings, the thought immediately occurs as to whether this is caused by tubes or by components in the stage. This question can be settled quickly by temporarily removing the RF drive. If the unbalance remains, the *tubes* should be suspected and replaced with balanced pairs. If the currents are balanced with removal of RF drive, the *circuits* should be suspected and examined for the cause. For example, many RF stages use cathode or filament bypass-capacitors. Should one of these capacitors be defective, currents would obviously be unbalanced with RF drive. If one is shorted, the currents would be unbalanced with or without RF drive, and the cathode-current meter readings would be abnormally low. If one is open, excessive tilt across the tops of the horizontal sync pulses would be observed from the output, whereas the waveform at the modulator would be normal. Obviously, unbalanced currents could also be caused by such things as defective screen or plate bypasses, bad connections, misaligned link couplers, etc.

Since any emergency changing of tubes in the RF portion during the operating day will affect to some degree the tuning of the stage, it is wise to try all the spare tubes in these stages during regular preventive maintenance schedules, and keep a record posted as to dial settings for proper tuning with these tubes. This saves the time of alignment that would otherwise be required. The transmitter may then be aligned exactly with the proper test equipment after the end of the regular operating day.

8-6. PREVENTIVE MAINTENANCE PROCEDURES

The importance of a rigid preventive maintenance schedule at TV transmitters (where standby units are a rarity) should now 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 30-minute meter readings, make daily reports on any peculiarities in meter readings, time and duration of any abnormal waveform observations, and any unusual deviations in frequency-deviation readings and water-temperature readings in water-cooled tubes. Record 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, 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 air around all high-voltage rectifier tubes.
5. Check 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. Get the habit of observing along with feeling for normal apeparance 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 would be absolutely necessary about every two months for optimum transmitter results.
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 and inspect all terminal boards for tightness of connections. After dusting, clean the entire transmitter with a vacuum cleaner.
4. Check blower motors and 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 for proper operation.
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 reflectometer against dummy-load reading.

Monthly (in addition to weekly)

1. Remove and test all receiving-type tubes with a good dynamic tester. Any tubes falling below 10% of their normal transconductance value should be replaced. Be sure to check the 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 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 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 all automatic control equipment, such as time-delay relays, overload relays, etc.
8. Where water-cooling systems are employed, check 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 preheated and stored upright as they should be, a 15-minute period of filament voltage only should be observed. Check for high-voltage operation and run them for several hours before storing 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, aural noise and distortion, and keep accurate records of tests. Any undue deviation from normal should be run down by stage isolation.

Mercury-vapor rectifier tubes should not be neglected in maintenance schedules. Unless proper precautions are taken, a major portion of lost airtime will be due to 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.

Due to the importance of foreseeing such trouble and due to the lack of familiarity of the average operator with testing methods of this type tube, the reader should become familiar with the maintenance procedure illustrated by Fig. 8-13. Since cathode-ray oscilloscopes are common at TV transmitter installations, the operator may conveniently use this most accurate check. An isolation transformer of at least 300 volt-amp rating should be used and a series current-limiting resistor of 50 ohms as shown. The mercury-vapor rectifier tube is left in its regulator 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-cps power line and power applied to the filament of the tube being checked, the scope pattern will show both the AC half of the nonconducting cycle and the conducting half which gives the DC potential. The sharp peak at the start of conduction reveals the tube condition under operating conditions. A good tube will fire at between 10 and 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 from 30 to 40 volts, the tube must be tested at more frequent intervals, preferably once a week. When this firing peak reaches close to 50 volts, the tube must be replaced with a new rectifier. Operators following this procedure will greatly minimize off-the-air time caused by rectifier arc-backs and otherwise defective tubes. Always remember that mercury-vapor rectifiers must have their filaments operated at normal voltage for a minimum of 30 minutes, then 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.

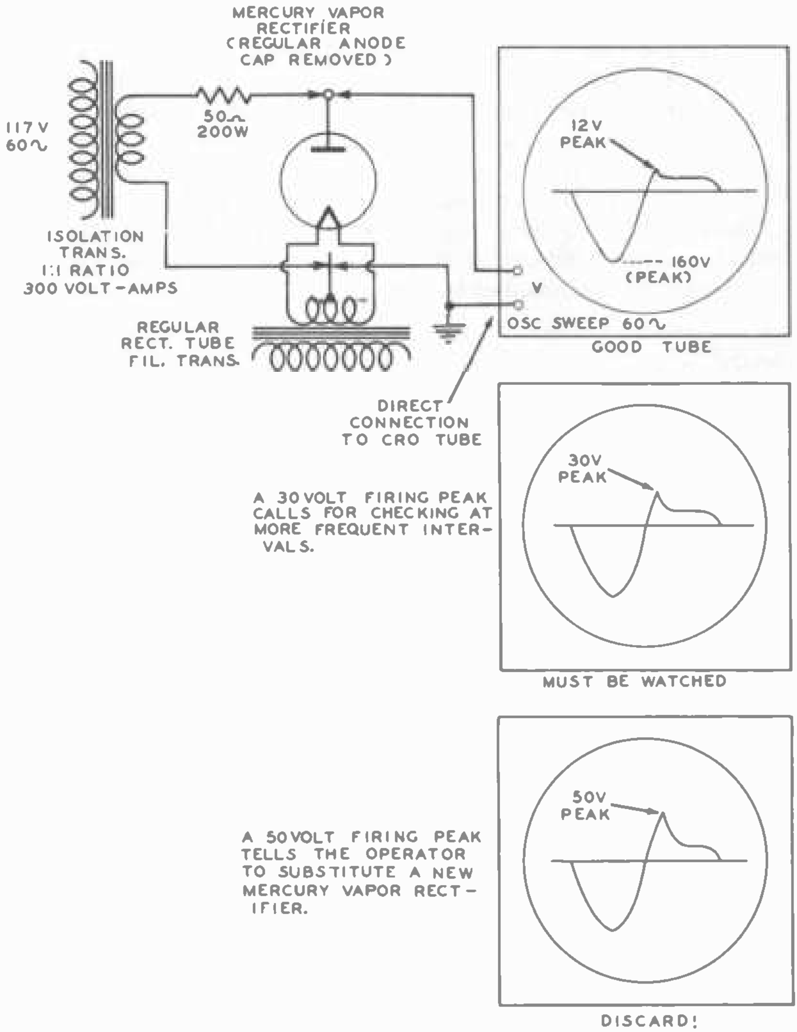
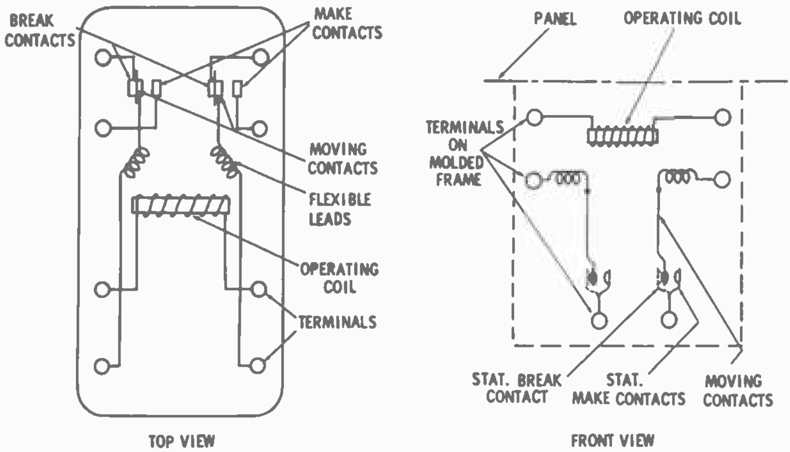


Fig. 8-13. Method of checking mercury vapor rectifier tubes and typical scope displays indicating tube conditions.

A number of contactors and relays are used in transmitters to provide automatic or remote-control closing and/or opening of various electrical circuits. The term *switch* is usually confined to manually operated assemblies that open and/or close circuits.

Relays are found in two basic types: front-connected open types, and rear-connected types solidly enclosed in dust-proof cases. In addition to this basic classification, a large variety of operational functions are encountered, such as normally-closed, normally-open, and relays which break certain circuits upon "making" other circuits. An example of the latter is a relay that completes a circuit to a supervisory pilot light when not energized, and completes a primary high-voltage circuit while opening the circuit to the pilot light when energized. Thus if this pilot light should start indicating during operation, the operator knows which relay has dropped out of operation.



(A) Open type. (B) Enclosed type.

Fig. 8-14. Westinghouse type SG relay.

One basic type of relay is the Westinghouse Type SG Relay illustrated in Fig. 8-14. Both open and enclosed types are shown. In either case the relay consists of four essential parts: core, yoke, armature, and coil. The open-type relay is normally supplied with two contacts and is shipped with both stationary contacts arranged to close when the relay is energized. However, either or both contacts can be converted quickly into a break contact merely by removing the screw which holds the stationary contact bracket, and turning the bracket over. After tightening the screw, the contact bracket may be bent slightly with the fingers, if necessary, to change the back-contact follow or alignment. When the make contacts are closed, the moving contact fingers should be deflected approximately 3/64", measured at the contacts, or slightly over 1/32", measured at the upper edge of the molded armature block. The assembly of the moving contact fingers on the armature block is arranged to provide spring-follow with either make or break stationary contacts. The closed-type relay is provided with two make and two break stationary contacts with the moving contacts common, and the open-type relay may also be provided with such a contact arrangement for applications that require it.

In examining relays of this type, observe the following:

1. Relay assembly for dirt, dust or other foreign matter.
2. All connections for tightness.
3. The coil for any signs of overheating revealed by charred insulation.
4. Coil and all wiring for defective insulation.
5. Moving parts for freedom of travel and follow.
6. Contacts for dirt, burns, pits or corrosion.
7. Proper line-up and follow of contacts; correct spacing.
8. Contact springs for proper tension and function.

Preventive maintenance of relays consists chiefly of cleaning the contacts with a strip of crocus cloth dipped in carbon tetrachloride. This operation should be followed by pulling a clean, dry, linen cloth between the contacts, which are held lightly closed with the fingers. Never use emery cloth on relay contacts, since abrasive granules may be left imbedded in the contact surfaces, tending to raise contact resistance and to encourage a tendency to weld. Slightly pitted contacts should be cleaned with a fine file. Badly pitted contacts should be replaced with new ones. Before any contact surfaces are worn to one-half their original thickness, they should be replaced.

Time-delay relays take several forms that operate on different principles to obtain a delay in function, but one of the most common is the motor-type illustrated in Fig. 8-15. This is the Westinghouse-type TD timing relay. This is an AC relay suitable for applications that require a time delay of from about five seconds to several minutes between the closing of an AC circuit and the closing or opening of a second circuit, either AC or DC, through the contacts of the relay. Many such applications are found in transmitters.

The relay should be mounted with its long dimension horizontal, so that the gear shafts are vertical and the motor terminals are at the top. It will not operate properly if mounted in any other position. An external resistor is used when the circuit to which the motor of the relay is connected has a voltage higher than 240 volts AC. This resistor should be mounted near the relay and connected in series with the motor circuit.

The type TD relay consists of a small 600 rpm self-starting synchronous motor, a gear train, and a set of silver contacts of the bridging type. When the motor is de-energized, the rotor rests in a position somewhat lower than the pole pieces of the stator. In this position the pinion on the rotor shaft is out of mesh with the gear on the countershaft that is mounted in the motor frame. When the motor is energized, the rotor is lifted by magnetic attraction, and the pinion is brought into mesh with the gear. The

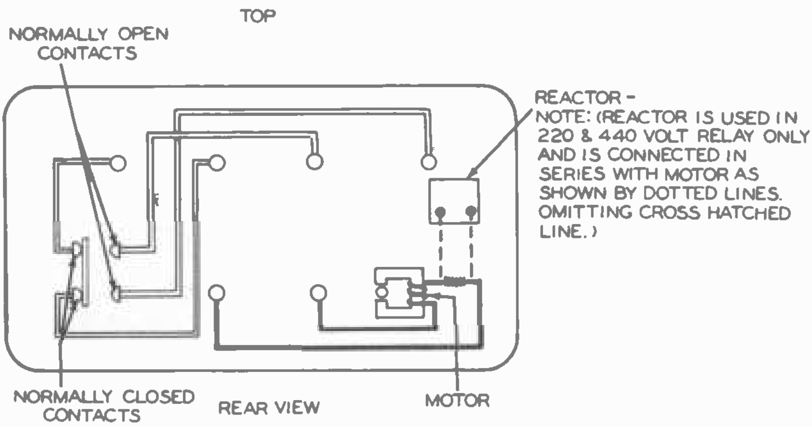


Fig. 8-15. Internal connections of Westinghouse six-terminal type TD (time delay) relay.

pinion on the motor countershaft drives a train of three reduction gears. An arm pressed on the shaft of the last gear is used to operate the contacts. When the motor has operated to open or close the relay contacts, the arm on the last shaft strikes a stop and the motor stalls. However, the motor can remain connected to the line without injury when stalled, and the locked rotor torque provides very good pressure on the closed contacts.

A spiral spring fastened to the shaft of the last gear causes the arm to reset to its initial position when the motor is de-energized. Since the pinion on the rotor shaft drops out of mesh when the motor is de-energized, the gear-train ratio is reduced, and the control spring will reset the arm very quickly. The time for maximum time delay is less than 5% of the operating time. Because of the inertia of the gear train, the resetting time is not directly proportional to the operating time. Consequently, with a time-delay setting of about one scale division the resetting time may be about 10% of the closing time.

An adjustable backstop for the arm on the last gear shaft is clamped between the upper bearing plate of the gear-train assembly and the bearing screw for the last shaft. A scale on the upper bearing plate is used in conjunction with an index line on the Micarta portion of the arm when it is desired to make an approximate setting of the relay. The motor will drive the arm over the entire scale travel in approximately 1.5 minutes. Thus each of the ten small scale divisions corresponds to approximately 9 seconds. Where a very accurate setting is desired, the time interval should be checked with a stop watch and the position of the backstop should be adjusted for the exact time required. The backstop is then clamped securely by means of the bearing screw after the desired setting is obtained.

For applications that require a time delay longer than 1.5 minutes, the TD relay is available with a maximum delay of approximately 3 minutes. This longer delay is accomplished by a change in the gearing of the motor only. The two styles of relays are identical in other respects.

The TD relay may also be found with several different contact arrangements. The contacts themselves are made of chemically pure silver. The four-terminal relay can be furnished with bridging contacts that are either opened or closed at the end of the time delay. The six-terminal relay has one set of contacts that open after a time-delay and a second set of contacts that close a few seconds later. In this case a silver strip that is held bridged across the normally-closed contacts by a spring is moved away from these contacts by the arm on the last gear shaft and is forced against the normally-open contacts. Therefore, the transition is not instantaneous, but depends on the spacing between the two sets of contacts and on the gear ratio. The spacing between the contacts can be varied a small amount by adding or removing washers between the heads of the contact screws and the mounting blocks.

A six-terminal TD relay is also available with one set of contacts that open as soon as the motor is energized, and a second set of contacts that either open or close at the end of the time-delay; but in this case the time-delay would not be adjustable.

The minimum time delay obtainable with the TD relay with normally-open contacts depends entirely on the minimum contact gap permissible. If the back stop is set so that the contact arm is one-half of a scale division from the contact-closed position, the contact gap will be approximately 1/16″, and the time delay will be about 4.5 seconds for the 1.5-minute relay. Because of the bridging type of contact used, there are two gaps in series in the contact circuit, and the total gap in this case will be about ⅛″. If the contacts are required to break only a small amount of current when the motor circuit is de-energized, the time delay could be decreased by a further reduction in the contact gap.

The normally-open contacts of the type TD relay can be used to close circuits carrying as much as 10 amperes at 125 volts, either AC or DC. They will open such a circuit satisfactorily if it is AC, but they should not be used to open a DC circuit carrying more than 1.5 amperes at 125 volts. The normally-closed contacts have less contact pressure than the normally-open contacts and should not be used to carry more than 5 amperes. Because they open slowly, they should not be required to break more than 2.5 amperes at 125 volts AC or 0.5 ampere at 125 volts DC.

All relays of either type, follower or plunger, should operate with a definite "snap" upon application of current to the operating coil. Friction resulting from dirt or other foreign matter, improper tension on contact spring arms or retaining springs, etc., must be remedied at the first opportunity.

Manually operated switches should also operate with a decided snap. The maintenance engineer soon becomes accustomed to the feel and sound of switch action and is aware as to when replacement is desirable. Switch contacts should be kept clean by constant wiping action under operation, but this should be reinforced by an occasional cleaning in the manner described previously.

Maintenance of antenna systems consists mainly of routine servicing of gas or dry-air dehydrating systems for gassing the coax lines and testing the lines for gas leaks should pressure be noted to drop at short intervals.

If it is apparent that gas leaks exist in the line, all terminals and line connectors should be coated with soapy water while the line is under normal gas pressure and then observed for the characteristic soap bubbles which reveal even slight amounts of gas leakage. Lines which do not retain gas pressure well are subject to rapid deterioration in feeding characteristics due to moisture and must therefore be repaired as quickly as possible. Most manufacturers recommend that lines and connections be thoroughly checked and tightened at least once a year.

To anticipate outage, capacitors in the high voltage, high-level circuits should be given the "feel" treatment each night after shutdown. The operator soon becomes accustomed to the proper operating temperature of large capacitors and in some instances can anticipate trouble by this procedure. Observation should be made for any signs of undue terminal strain, cracks, leaking cases, etc. Smaller capacitors should always be inspected closely for physical appearance, tightness of mounting and connections, etc. Electrolytic capacitors should be observed for bulging sides and other signs of breakdown.

One of the primary functions of preventive maintenance schedules is to keep all components clean. Newcomers are likely to minimize the importance of this item, but the old-timers realize the utmost importance of a clean transmitter. Operation in the VHF and UHF spectrum makes this item extremely important as compared to standard broadcast frequencies.

In routine inspection procedures, note the appearance of all resistors. Blistered or blackened cases are danger signs. Run periodical ohmmeter measurements on all important resistors, and note any deviations over a period of time. Considerable trouble may be prevented by this simple precaution.

Transformers should also be observed for any physical signs of deterioration. Feel the coil insulation at periodic intervals for any softness or charring tendencies. Adjustable transformers such as line-voltage autotransformers are particularly subject to troubles, without good preventive maintenance. Examine brushes at regular intervals and replace them with the type recommended by the manufacturer *before* undue wear occurs. Test brush springs for tension and tightness of screws holding the brush springs. The brushes and commutator surfaces should be periodically cleaned with a strip of crocus cloth and dusted with an air stream followed by vacuuming.

Remember that every maintenance schedule is extremely important, and that alertness and calm deliberation followed by positive action pays off with a minimum of operating-time emergencies.

Many TV transmitter installations include a standby power generator similar to that of Fig. 8-16 in case of loss of commercial AC power. The emergency power source should be checked at least once a week, and starting batteries (if used) should be kept fully charged.

8-7. MEETING FCC PROOF-OF-PERFORMANCE REQUIREMENTS

A proof of performance is required before each application for renewal of license. It is a good practice to run these tests at least twice annually, even though not required by the FCC.

For monitoring of either the actual picture or the modulated envelope at the input to the antenna system (output of vestigial sideband filter or output of final stage in low-level, modulated transmitters) special considerations are involved. Fig. 8-17A shows the response of an ordinary diode at this point. The sharp cutoff of the lower sideband frequency components in the RF signal results in a boost of the frequencies below 1.25 mc, as shown. It is obvious that the high-frequency response will be approximately 50% of the response below 0.75 mc. Thus, in the case of the visual transmitter where the input is the video-band frequencies and the output is in the radio-frequency spectrum with vestigial sideband characteristics, monitoring devices must be arranged and interpreted accordingly. If an ordinary diode is used ahead of the sideband filter or at the first modulated RF stage (where both sidebands are present to an appreciable amount), the device provides a suitable signal either for picture

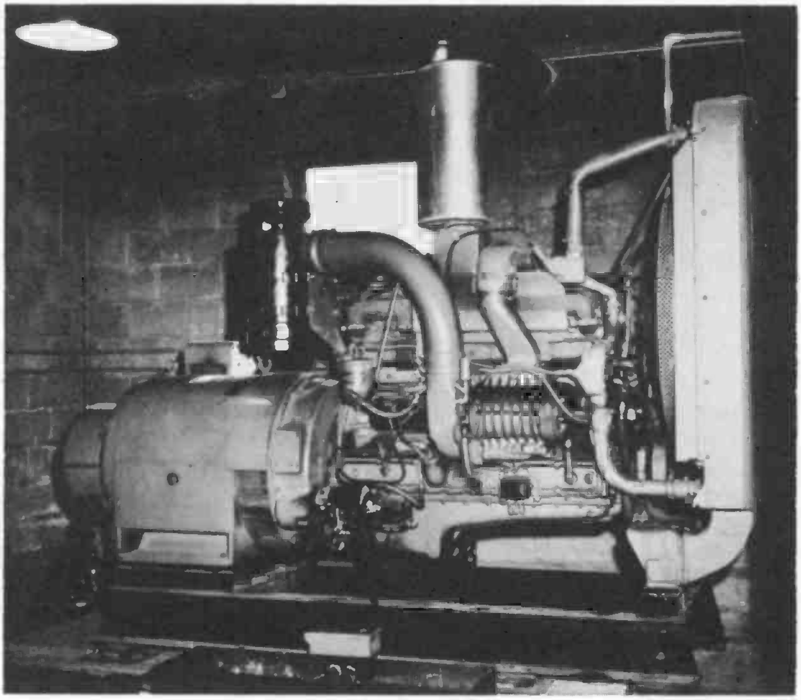
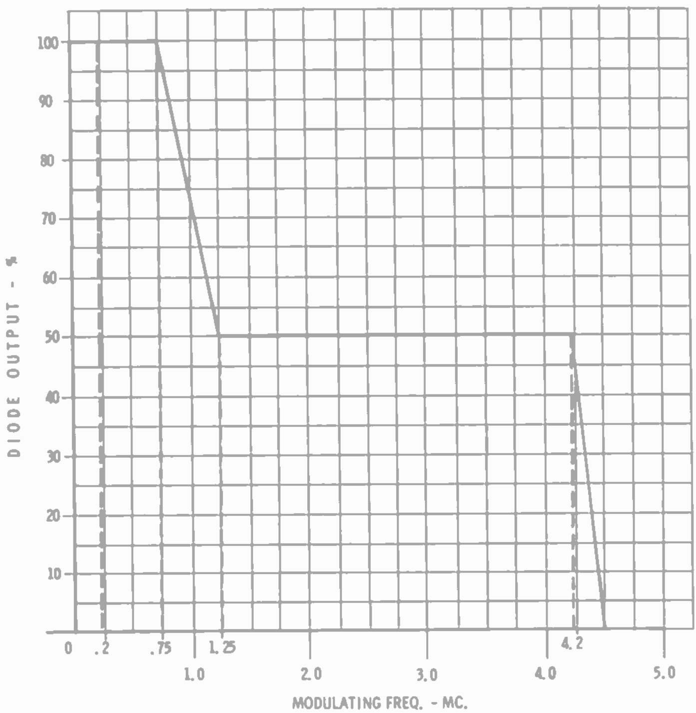


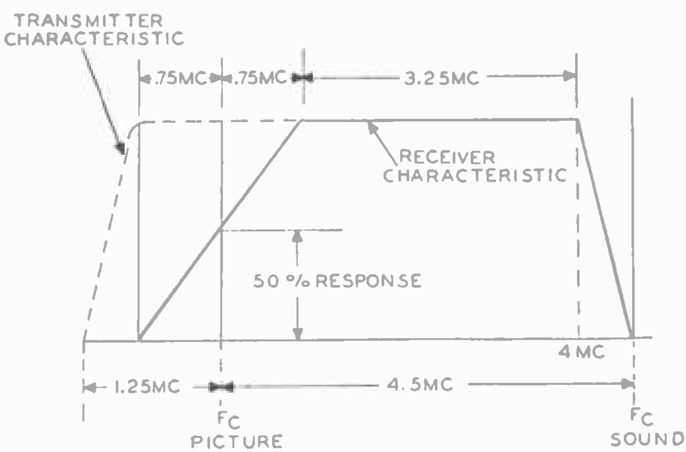
Fig. 8-16. Diesel-driven generator used to supply emergency power for the transmitter at station WTAE.

monitoring or waveform monitoring. However the reader should understand that such monitoring does not provide an accurate check at the antenna system input, since the effects of vestigial sideband transmission are not being monitored.

The vestigial RF output signal from the transmitter must be properly returned to the video frequency band for accurate monitor interpretation. To do this the demodulator must approach as nearly as possible the ideal receiver response, as illustrated in Fig. 8-17B. This characteristic serves to equalize the unbalanced sideband energy distribution of the transmitted signal by a 6-db



(A) Ordinary diode detector connected at output of vestigial sideband filter.



(B) Ideal receiver response characteristic.

Fig. 8-17. Response curves.

attenuation of the picture carrier. This takes place over a range of 3 db for 0.75 mc above the picture carrier frequency and 3 db for 0.75 mc below the picture carrier frequency. A linear detector is preceded or followed by suitable selective circuits to provide this overall response characteristic. The amplitude-frequency characteristic is flat from 0.75 mc to 4 mc above the carrier and substantially zero at frequencies lower than 0.75 mc below the carrier. The response at the carrier frequency should be one-half the value of the response in the upper sideband above 0.75 mc.

Many stations use a modified receiver chassis for monitoring the picture, which gives an accurate final check on the radiated signal if precautions are taken to keep the receiver circuits in proper order. Such a receiver may also be used for modulation-percent indication by incorporating the chopper or vibrator across the second detector to establish the zero-carrier reference line (preceding chapter). The short-circuit intervals are made suffi-

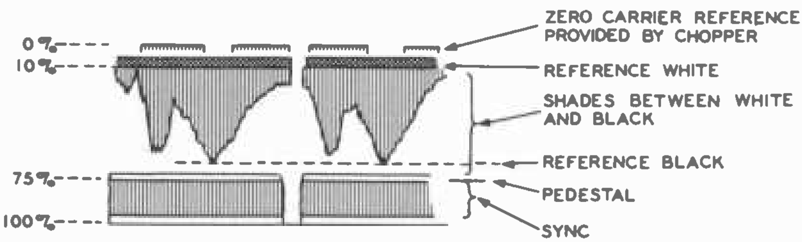


Fig. 8-18. Drawing of typical CRO display of half-tone video signal chopped at 120-cycle rate and displayed with scope sweep of frame frequency (30 cps).

ciently short so as not to minimize the usefulness of the oscillographic display, and they have a repetition rate sufficiently high to establish the zero level at least twice during the displayed trace. When displayed at frame frequencies, a common chopper rate is 120 cps, which establishes four zero-reference lines (Fig. 8-18). When displayed at line frequency, a chopper rate of 31.5 kc provides two zero-reference lines.

Amplitude-Frequency Response Measurements (visual transmitter)

When stations are equipped with a device known as a *sideband analyzer*, this characteristic is normally monitored daily. This type of test signal also reveals any need for transmitter tuning adjustments.

The basic functional diagram of the RCA sideband-response analyzer is presented in Fig. 8-19. The wobbulator section consists of the conventional arrangement of a fixed-frequency oscillator and a sweep oscillator that varies above and below the fixed

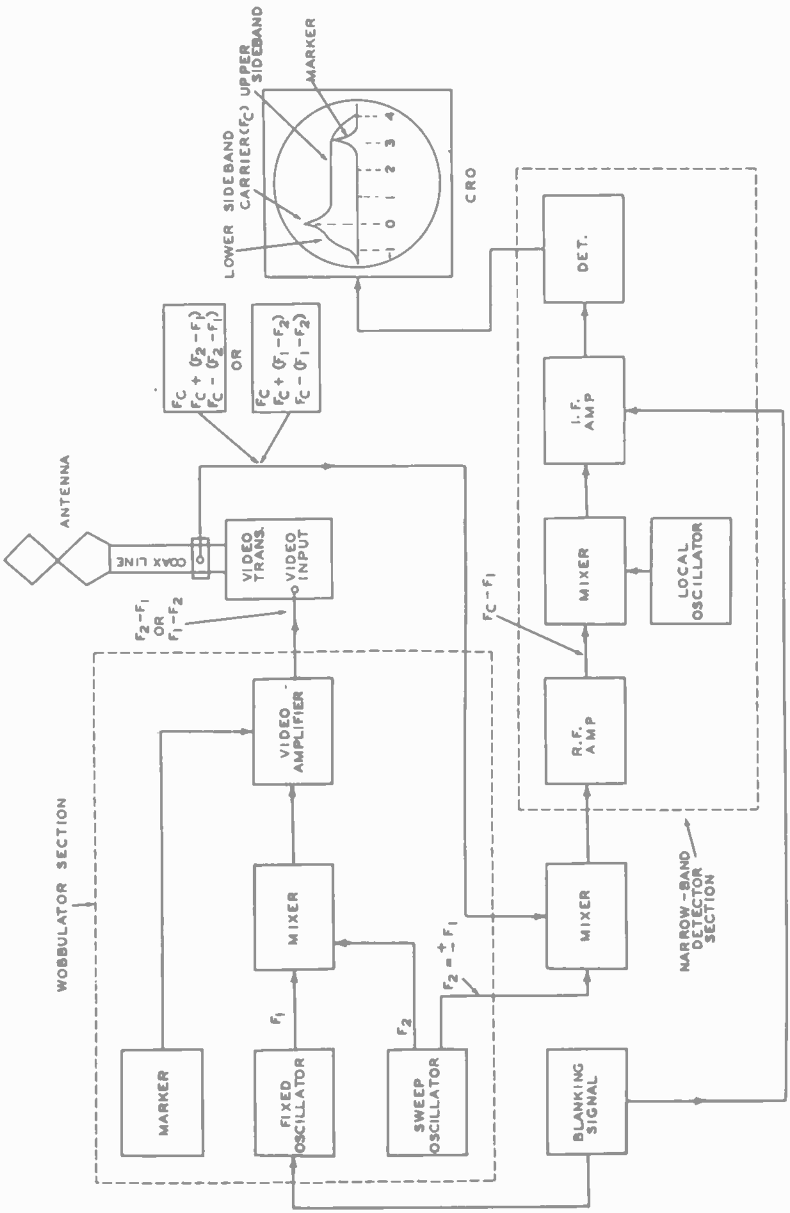


Fig. 8-19. Basic block diagram of RCA sideband response analyzer.

frequency by approximately equal amounts. When the frequency of the sweep oscillator (f2) is higher than the fixed frequency (f1), the radio-frequency carrier (f_c) is modulated by the frequency f2-f1. During this interval the signal that is fed to the antenna (and to the sideband analyzer) contains three frequencies as follows:

Carrier frequency = f_c
Upper sideband = $f_c + (f2 - f1)$
Lower sideband = $f_c - (f2 - f1)$

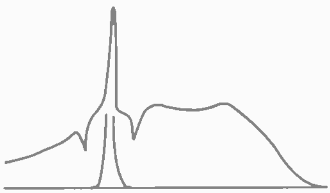
This signal is fed to the mixer stage, which heterodynes it with the signal (f2) from the sweep generator. The resulting heterodyned signal is fed to the RF amplifier, which is the first stage of a narrow-band detector portion of the analyzer. As shown in the diagram, this detector accepts only the frequency $f_c - f1$. The output is then proportional to the upper sideband response when f2 is greater than f1. Similarly, the output is proportional to the lower sideband when f2 is lower in frequency than f1. Using a sufficiently high sweep rate on the CRO displays the frequency versus amplitude characteristic of the transmitter, as shown in the typical curve in the illustration.

The net effect may be seen to separate the upper and lower sideband response for the purpose of simultaneous presentation on the screen of an oscilloscope. The theory of operation is very similar to that described for checking the high-frequency response characteristics of video amplifiers at the studio. In this application the method is adopted to the transmitter action of vestigial sideband response. Its primary function is to check and adjust the broadband, overcoupled RF circuits used in most visual transmitter circuits. The display with markers obviously permits optimum adjustment of stages to obtain the proper standard transmission characteristics.

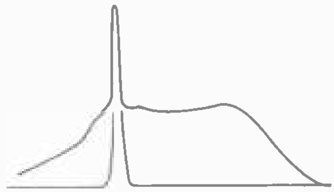
Fig. 8-20 illustrates typical traces obtained with the analyzer for two common transmitter tuning misadjustments. In Fig. 8-20A is shown the effect of cathode-lead resonance, and Fig. 8-20B shows the trace after this resonance is damped out. In Fig. 8-20C the effect of improper neutralization is shown by the inequality of the upper and lower sidebands in the immediate carrier vicinity, whereas Fig. 8-20D shows the trace obtained upon correction of the neutralization.

Fig. 8-21 illustrates how the trace proves useful in proper tuning of the driver stage for vestigial sideband transmission. Fig. 8-22 is a photograph of a typical trace monitored following the vestigial sideband filter in a properly adjusted transmitter.

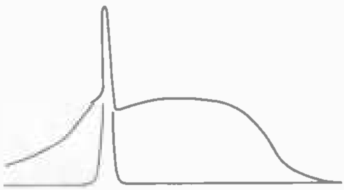
The primary usefulness of the sideband-response analyzer is in keeping the transmitter properly tuned and adjusted for normal



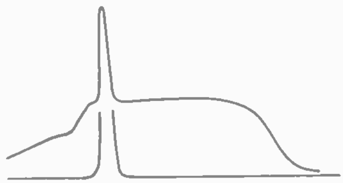
(A) Cathode-lead resonance.



(B) Cathode resonance damped out.



(C) Upper and lower sideband inequality due to improper neutralization.

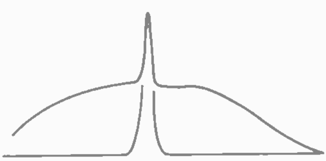


(D) Response indicating proper neutralization.

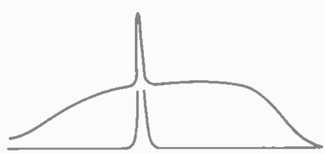
Fig. 8-20. Transmitter adjustments by means of sideband response analyzer.

operations. The FCC proof-of-performance data are normally compiled from single-frequency sine-wave runs with 200 kc (0.2 mc) as the reference frequency. Current FCC rules specify the condition for this measurement as follows:

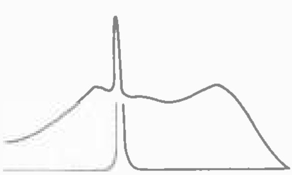
> The attenuation characteristics of a visual transmitter shall be measured by application of a modulating signal to the transmitter input terminals in place of the normal composite television video signal. The signal applied shall be a composite



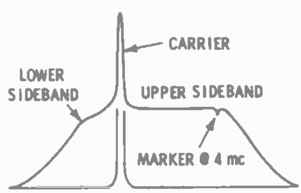
(A) Incorrect driver alignment (double-sideband response).



(B) Driver correctly aligned for carrier offset to achieve partial suppression of lower sideband.



(C) Overcoupled RF circuits.



(D) Curve obtained in most modern transmitters ahead of sideband filter.

Fig. 8-21. Transmitter tuning with sideband response analyzer.

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 in this section means either

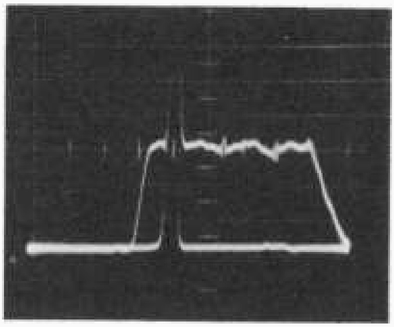


Fig. 8-22. Typical sideband response after adjustments within FCC specifications.

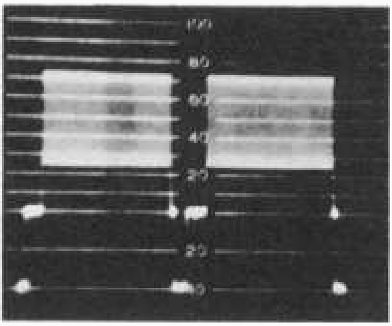


Fig. 8-23. Proportionment of sine wave and sync combination for transmitter runs.

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-kilocycle 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 inseration 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. 8-23. The sine wave occupies the region of 25 to 75 IRE units for a peak-to-peak value of 50 IRE units. The transmitter modulator gain is left adjusted for the original normal input of 10 to 100 IRE Units of video. Note, therefore, that full video modulation of the transmitter does not occur on single-frequency sine-wave runs.

Fig. 8-24 illustrates a typical test setup for transmitter frequency-response (attenuation characteristic) runs. Although this meets the FCC requirements (video test signal to input of transmitter terminals) it is good operating practice to send the test signal via 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.

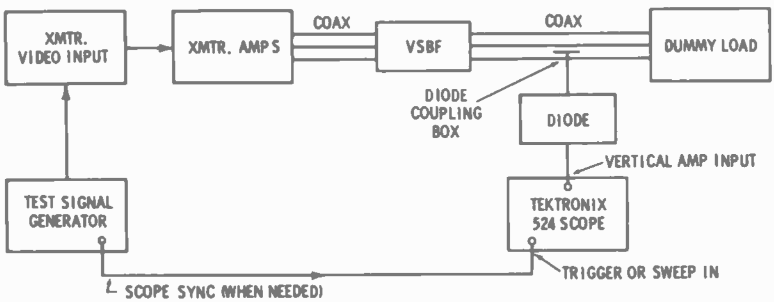


Fig. 8-24. Setup for measuring transmitter frequency response.

The 0.2-mc sine wave (Fig. 8-23) is the reference level on the scope at the VSBF output and is designated 0 db. The frequency is then increased in steps, holding the same input amplitude to the transmitter, and the response tabulated as in Table 8-1. These data may then be transferred to the diode-demodulator curve of Fig. 8-25.

The FCC visual transmitter frequency-response requirements for monochrome and color are tabulated in Chart 8-1.

Table 8-1. Tabulation of Video Frequency Response Data

(Example of typical readings taken on WTAE Main Visual runs)

FREQUENCY (mc)	DIODE RESPONSE	
	%	DB
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

Chart 8-1. FCC Frequency Response Requirements (Visual Transmitter)

MONOCHROME	*COLOR*
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. 2 db at 0.5 mc 2 db at 1.25 mc 3 db at 2.0 mc 6 db at 3.0 mc 12 db at 3.5 mc The curve shall be substantially smooth between these specified points, exclusive of the region from 0.75 to 1.25 mc. Output measurement shall be made with the transmitter operating into a dummy load of pure resistance and the demodulated voltage measured across this load.	A sine wave of 3.58 mc 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 R. F. transmission line supplying power to the antenna), which is down 6±2 db with respect to a signal produced by a sine wave of 200 kc. In addition, the amplitude of the signal shall not vary by more than ±2 db between the modulating frequencies of 2.1 and 4.18 mc.

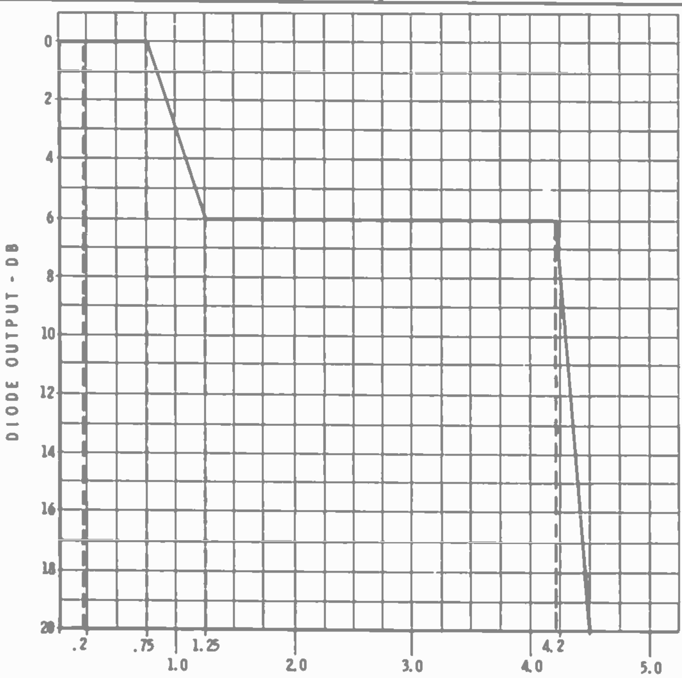


Fig. 8-25. Graph plotted from data in Table 8-1.

Fig. 8-26 illustrates how you can plot your own curve for the particular application involved. This can be drawn on a master and reproduced by the station's reproducing equipment when available.

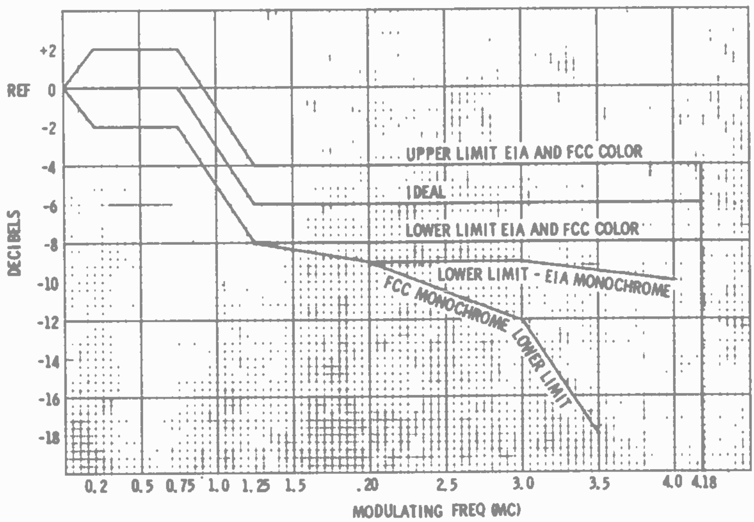


Fig. 8-26. Example of graph plotted to show frequency response characteristics of monochrome and color transmitters.

Note that with vestigial sideband transmission (see Fig. 8-27) the characteristic of the sideband attenuation (upper and lower) must be considered.

Always check current FCC rules—The FCC specifications 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 mc or greater and in addition, for color, shall not be greater than −42 db for a modulating frequency of 3.579545 mc (the color subcarrier frequency). For both monochrome and color, the field strength or voltage of the upper sideband as radiated or dissipated and measured as described shall not be greater than −20 db for a modulating frequency of 4.75 mc or greater.
>
> NOTE: Field strength measurements are desired. It is anticipated that these may not yield data which are consistent enough to prove compliance with the attenuation standards prescribed above. In that case, measurements with a dummy load of pure resistance, together with data on the

antenna characteristics, shall be taken in place of over-all field measurements.

When field-strength readings of the upper and lower sidebands are made, the test signal of Fig. 8-23 is used. The field strength (for example) with a modulating signal of 2 mc is measured at the carrier frequency plus 2 mc for the upper sideband, and the carrier frequency minus 2 mc 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. 8-25. A video sweep, or video sweep incorporated with the sideband analyzer is employed for the lower-sideband attenuation characteristic. The transmitter is

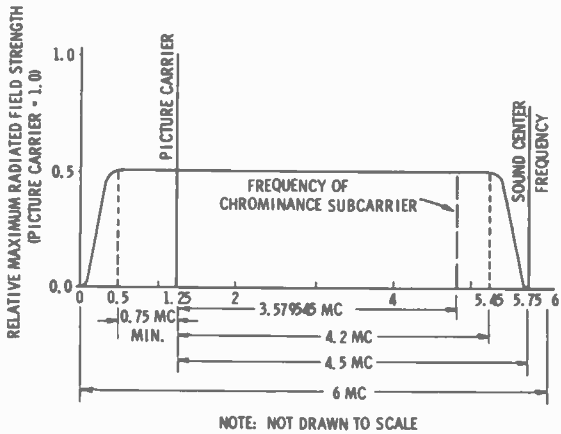


Fig. 8-27. Ideal picture transmission amplitude characteristic.

fully modulated (87.5%) and the marker is moved to 0.2 mc for the reference 0 db. Maximum scope gain must normally be used to measure the lower-sideband response at frequencies above 1.25 mc. Note that when color standards must be met, the lower-sideband energy at 3.58 mc 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 entirely 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.

Color-Transmission Equalization

Time-delay distortion in color signals must be held to a minimum to avoid the "funny paper" effect of misregistration. (This term is a carry-over from the early days of printing color comics, where 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 mc, 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 mc, the corresponding phase distortion must be brought into tolerable limits for good color transmission. At higher frequencies up to 4.18 mc, 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.

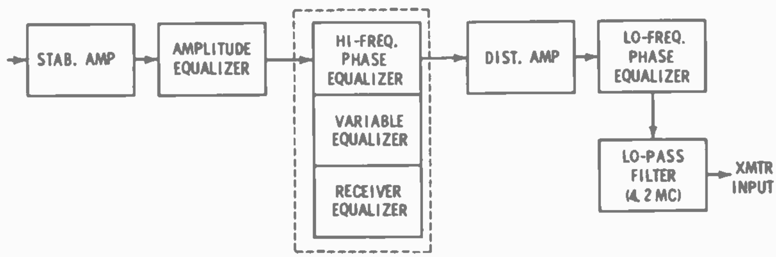


Fig. 8-28. Units employed for color transmitter equalization.

Fig. 8-28 illustrates a typical series of units employed to meet the preceding requirement. Due to the general lack of suitable measuring equipment, most engineers have relied on square-wave response to judge the transmitter radiation phase-characteristics. Fig. 8-29 illustrates the FCC specification for the transmitter envelope delay curve for color TV transmissions. The specifications themselves read as follows:

> 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 mc, of zero microseconds up to a frequency of 3.0 mc; and then linearly decreasing to 4.18 mc so as to be equal to −0.17 microseconds at 3.58 mc. The tolerance on the envelope delay shall be ±0.05 microseconds at 3.58 mc. The tolerance shall increase linearly to ±0.1 microsecond down to

2.1 mc, and remain at ±0.1 microsecond down to 0.2 mc. (Tolerances for the interval of 0.0 to 0.2 mc are not specified at the present time.) The tolerance shall also increase linearly to ±0.1 microsecond at 4.18 mc.

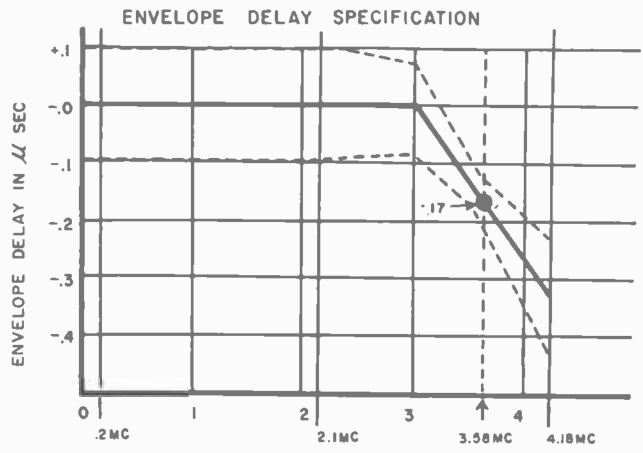
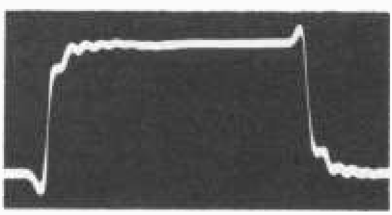
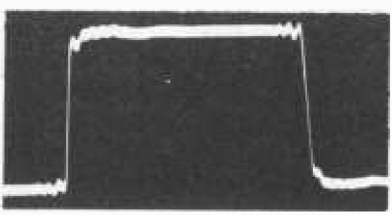


Fig. 8-29. FCC specifications for the transmitter envelope delay curve for color TV transmissions.

Fig. 8-30A shows a typical uncorrected transmitter response to a 100-kc square wave. The transients preceding the transitions result from low-frequency phase distortion inherent with the attenuation of the lower sideband. Ringing after the transitions is phase distortion from attenuation of the upper sideband. These defects indicate delay distortion outside the tolerance for color transmission.



(A) Uncorrected response.



(B) With phase correction.

Fig. 8-30. TV transmitter response to a 100-kc square wave.

The networks of Fig. 8-28, when properly adjusted, can minimize but not completely eliminate the phase distortion. Fig. 8-30B shows the response to the 100-kc 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 by special equipment only, such as the recently developed RCA BW-8 Envelope-Delay Measuring Set. 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 mc have been accepted by the FCC.

Amplitude Linearity

Current FCC rules simply state that, for color transmission the amplitude linearity "shall be substantially linear." Industry has generally established a maximum allowable nonlinearity of 10%. It must be realized that this degree of nonlinearity is the *total overall scale*, and cannot all occur, for example, in either the white or sync region.

Amplitude linearity of the transmitter is measured with the stairstep signal as previously described (Section 6). However, a special problem exists at the transmitter, which calls for a modified procedure.

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 stabilizing amplifier, are provided to predistort the signal so that the overall transfer curve is linear.

When these adjustments are provided in the external stabilizing amplifier, this amplifier becomes an integral part of the measuring path and the stairstep signal is fed to the stabilizing amplifier input. The "white stretch" circuit is adjusted so that the demodulated signal (after the sideband filter) indicates linear transfer. Full modulation should be used.

Adjustments are also provided to compensate transmitter differential gain and phase. The techniques are the same as already covered in Section 6, with the added function of the precorrecting circuits. Modern installations can readily be adjusted to within 5% differential gain and 5° of differential phase at 3.58 mc.

Assembling Data on Proof of Performance (visual)

Exhibit 1. Draw a block diagram of the test-signal path. Record type, manufacturer, and serial numbers of each unit employed in the measurement.

Exhibit 2. Tabulate the frequency response (upper and lower sidebands).

Exhibit 3. Plot the data of (2) on suitable graphs. This should include a plot on the ideal diode response-curve for the upper sideband (Fig. 8-25). The lower-sideband response can be plotted on suitable linear graph paper.

Exhibit 4. Tabulate the pertinent data as in Fig. 8-31. Regulation measurements were described in Section 8-1. The video signal-to-noise ratio is measured as described for the STL in Section 7. NOTE: The signal-to-noise ratio at the transmitter output can normally be considered to better that obtained on the STL alone. This is because the STL is measured with a wide-band scope response. While this is also true of the main transmitter, the attenuation characteristic (sharp roll-off above 4.18 mc) erases 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. However, it is a good practice to include this and following photos on each proof.

EXHIBIT 4

(With typical measurements at WTAE Transmitter)

REGULATION: Black: 100
White: 103
DIFF. GAIN (3.58 mc): 5%
DIFF. PHASE (3.58 mc): 50% APL: 2°
VIDEO SIGNAL/NOISE RATIO: 40 db
FINAL PLATE VOLTS: 3.65 kv
FINAL PLATE CURRENT: 8.2 amps, Black Level
INDICATED POWER IN DUMMY LOAD: 13.2 kw; Black Level
FREQUENCY METER READING: 100 cps

Latest Calibration Date: ______________
By: ______________________

Date: ______________
Engineer: ______________

Fig. 8-31. Example of tabulation of pertinent data for proof-of-performance test.

Exhibit 6. Photograph the 100-kc square-wave response (color installations only). Or, tabulate envelope-delay *versus* frequency if direct measurements can be taken.

Exhibit 7. Photo of video waveform (after VSBF) at normal line-rate, showing 7½% setup and full modulation.

Exhibit 8. Photo (expanded scale) of H sync interval with 0.005H markers.

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 preceding data, and attaches the exhibits to the forms.

Aural Transmitter Proof (Current FCC Rules)

(1) The transmitter shall operate satisfactorily with a frequency swing of ±25 kc, which is considered 100% modulation. It is recommended, however, that the transmitter be designed to operate satisfactorily with a frequency swing of at least ±40 kc.

(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 cps. 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 cps. The lower limit shall be uniform from 100 to 7,500 cps, and 3 db below the upper limit; from 100 to 50 cps the lower limit shall fall from 3-db limit at a uniform rate of 1 db per octave (4 db at 50 cps); from 7,500 to 15,000 cps the lower limit shall fall from 3-db limit at a uniform rate of 2 db per octave (5 db at 15,000 cps).

(3) At any modulating frequency between 50 and 15,000 cps and at modulation percentages of 25%, 50%, and 100%, the combined audio-frequency harmonics measured in the output of the system shall not exceed the rms values given in Table 8-2.

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 kc.

NOTE: Measurements of distortion using de-emphasis in the measuring equipment are not practical at the present time for the range 7,500 to 15,000 cps for 25% and 50% modulation. Therefore, measurements should be made at 100% modulation and on at least the following modulating frequencies: 50, 100, 400, 1,000, 5,000, 10,000, and 15,000 cps. At 25% and 50% modulation, measurements should be made on at least the following modulating frequencies: 50, 100, 400, 1,000 and 5,000 cps.

Table 8-2. Modulation Frequency Versus Distortion Percentage

Modulation frequency	Distortion per cent
50 to 100 cycles	3.5
100 to 7,500 cycles	2.5
7,500 to 15,000 cycles	3.0

It is recommended that none of the three main divisions of the system (transmitter, studio to transmitter circuit, and audio 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 cps shall be at least 55 db below the audio-frequency level representing a frequency swing of ±25 kc.

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 cps shall be at least 50 db below the level representing 100% 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.

Chart 8-2. Summary of Station Proof of Performance Measurements

(After VSBF with transmitter operating into dummy load)

VISUAL	*AURAL*
1. CARRIER FREQUENCY (By station monitor; with notation of latest calibration; date, and by whom made.) 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 notation on log.) 3. XMTR REGULATION: BLACK-TO-WHITE 4. SIDEBAND RESPONSE AND ATTENUATION CHARACTERISTIC (If calibrated receiver such as RCA BW-7 available, run measurements on point-to-point basis with single-frequency sine waves and sync combined. Or use sideband analyzer such as RCA BW-5. Photos of CRO display should be filed.) 5. SIGNAL/NOISE RATIO AND SIGNAL/HUM RATIO 6. LINEARITY AT FULL MODULATION 7. DIFFERENTIAL GAIN (3.58 mc) AT FULL MODULATION 8. DIFFERENTIAL PHASE (3.58 mc) AT FULL MODULATION 9. PHOTOS OF: Test Pattern from Monitor ½ line-rate sweep of above on CRO. ½ filed-rate sweep of above on CRO. Expanded H blanking interval with 0.005H markers.	1. CARRIER FREQUENCY (By station monitor, with notation of latest calibration; date, and by whom made.) 2. RF OUTPUT POWER(rms) $E_p \times I_p \times F$ where: E_p is equal to plate volts of final stage, I_p is equal to plate current of final stage, F is equal to efficiency factor supplied by manufacturer. THE FOLLOWING TO BE MADE WITH VISUAL TRANSMITTER OFF: 3. SIGNAL/NOISE RATIO (a) FM (b) AM 4. AUDIO FREQUENCY RESPONSE (Measuring input level required to hold reference modulation percentage.) 5. HARMONIC DISTORTION (50-15,000 cps Distortion meter must measure harmonics to 30 kc.)

NOTES

On engineering portions of FCC forms under heading "Types of Emission," one of the following applies:

A3: Amplitude modulation; telephony.
F3: Frequency modulation; telephony.
F5: Frequency modulation; television.
F5/F3; Frequency Modulation; TV and telephony, such as TV STL with aural subcarrier.

Under heading "Communication Bandwidth" enter only the maximum bandwidth authorized for the frequency concerned. Following is complete tabulation. (ALWAYS CHECK CURRENT FCC RULES):

Frequency Band	*Service*	*Emission*	*Bandwidth (KC)*
1606 to 1646 kc	Remote Pickup	A3	10
25.87 to 26.03 mc	Remote Pickup	A3 or F3	40
26.07 to 26.47 mc	Remote Pickup	A3 or F3	20
152.87-160.89 mc	Remote Pickup	A3 or F3	60
161.64-161.76 mc	Remote Pickup	A3 or F3	30
166.25-170.15 mc	Remote Pickup	A3 or F3	60
450.05 455.95 mc	Remote Pickup	A3 or F3	100
942.5-951.5 mc	AM or FM STL	F3	400
1990-2008 mc	TV STL	F3/F5	18,000
2008-2500 mc	TV STL	F3, F5 or F3/F5	17,000
6875-13,225 mc	TV STL	F3, F5 or F3/F5	25,000

(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 diplexing, follow the procedure given in checking the STL described in Section 7. Note that when measuring the overall system, which includes the main TV aural transmitter, the gain must be dropped at the studio by 20 dbm when the 1-kc 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 (review Section 7).

The aural frequency response should be tabulated and plotted within the pre-emphasis curve. (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. Date and sign the forms. Chart 8-2 summarizes the visual and aural proof-of-performance methods.

INDEX

TELEVISION SYSTEMS MAINTENANCE
by Harold E. Ennes

Television Systems Maintenance combines the latest techniques in the standardization of television systems maintenance procedures with a complete analysis of system malfunctions. A certain amount of fundamental systems theory has been included where important in maintenance practice.

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ABOUT THE AUTHOR

Harold Ennes has been associated with various phases of radio engineering since 1930. He entered the broadcast field in 1936 as a staff engineer with station WIRE, Indianapolis. Later he installed the first FM broadcast station in Indianapolis—noncommercial WAJC for Jordan College of Butler University—and was the station's chief engineer for four years. In addition, he taught radio and television at Butler University for five years. Since 1958 Mr. Ennes has been maintenance supervisor for Television City, Inc. (WTAE-TV Pittsburgh). He has written numerous articles and books on the various aspects of radio and television broadcasting. Other SAMS broadcast references by Mr. Ennes include: *AM-FM Broadcast Operations*, *AM-FM Broadcast Maintenance*, and *Television Tape Fundamentals*.

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