

ELECTRONIC TECHNOLOGY SERIES

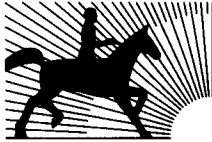
VIDEO AMPLIFIERS

a **RIDER** **publication**

\$1.80

VIDEO AMPLIFIERS

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JOHN F. RIDER PUBLISHER, INC.

116 West 14th Street • New York 11, N. Y.

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Library of Congress Catalog Card No. 59-8632

Printed in the United States of America

PREFACE

Video amplifiers are integral parts of a variety of communications and industrial systems. Voltages and currents of the frequency ranges comprising a video signal appear in radar transmission and reception, pulse amplification, television and within a complex of automation equipment. For this reason, it is essential that those concerned with electronics possess a working knowledge of the essential relationships pertaining to video amplifiers. Such knowledge and techniques are necessary and useful throughout the fields of communications, military and industrial electronics.

The intent of this book is to discuss and evaluate video amplifiers in general, presenting the fundamental concepts of the subject. The mathematical treatment employed has been kept simple, but the analyses are sufficiently extensive to permit the interested technician or student to develop a full comprehension of the pertinent theory. To insure this aim, an adequate amount of information is given relating to broad concepts and information designed for ready use; detailed descriptions of a small number of selected major topics are presented, rather than treating a larger body of less important material; and, through presentation of practical situations, equipment and problems the reader is afforded an opportunity to apply the principles he has learned.

Specific attention is given to the nature of the video signal; a description of the television video signal; the maximum frequencies in the video range; the Fourier theorems; the structures of square and saw-tooth waves; noise effects on the video signal; noise voltage calculations; the uncompensated video amplifier—its requirements; basic analytical approach; high-frequency behavior; phase shift at high frequencies; low-frequency response; phase shift at low fre-

quencies and pulse response; high-frequency compensation methods—analytical approach; analysis of shunt peaking; series-peaking; combination peaking; low-frequency compensation—analysis and transient behavior; video amplifier design procedure; calculation of a shunt peak network; calculation of a series-peaking network; performance and design data for high-frequency compensation systems; design of low-frequency performance; special measurement considerations; and measurements of gain- and phase-frequency characteristics. Thus, a foundation is provided upon which further concepts can be built.

Grateful acknowledgment is made to the staff of the New York Institute of Technology for its assistance in the preparation of the manuscript of this book.

February, 1959
New York, New York

A. S.

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Chapter 1

THE NATURE OF THE VIDEO SIGNAL

1. General Definitions

Video signals are generated, transmitted, and received in many varieties of communication and industrial systems. Although the formation of a picture is not always involved, and human vision does not always form a part of such a complete system, the signal is characterized by the term *video* because of its early and extensive use in television.

In terms of general usage, a video signal is any a-c voltage or current containing a wide range of frequencies from some low value such as 30 cycles up to and including several megacycles. (A pulsating d-c containing a-c components over such a frequency range may also be classified as a video signal.) Voltages and currents of this nature appear not only in television but also in radar transmission and reception, pulse amplification, and in some types of equipment used for automation.

This book discusses and evaluates video amplifiers in general. Since the television video signal contains a-c components carrying video intelligence as well as short and long pulses, any analysis of video amplifiers used in television is also applicable to the performance and design of video amplifiers found in radar and pulse equipment. For this reason, specific problems, solutions, and illustrations are applied to television usage; carry-over to other devices in

which wideband amplifiers are employed are discussed only in cases where essential differences in design and performance exist.

2. Description of the Television Video Signal

The standard television signal used in the United States consists of camera impulses carrying the video intelligence, blanking pulses, synchronizing pulses, and equalizing pulses. Figure 1 provides information concerning the relative amplitudes of these components. The polarity of the currents or voltages is intentionally selected so

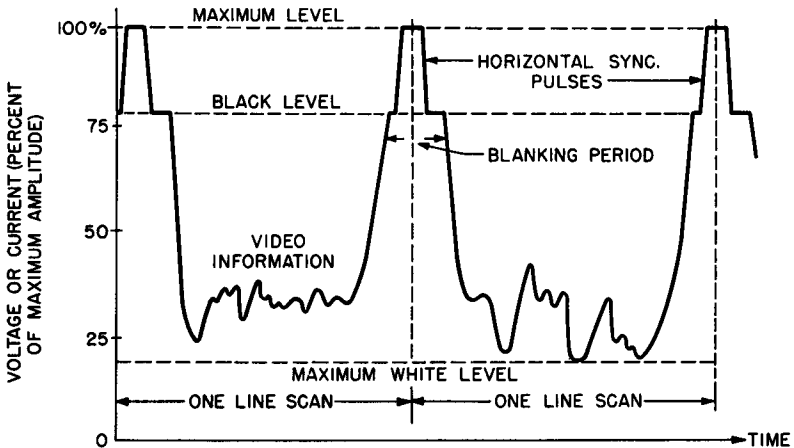


Fig. 1. Two scanned lines and the blanking periods of the standard video signal.

that the brilliance of the reproduced picture decreases as the amplitude increases. This transmission method is known as *negative transmission*. Positive transmission, or the system in which increasing amplitudes lead to increasing brilliance, is used in England and a few other foreign countries. Standardization of polarity is essential since it determines the tonal relationships in the final picture, that is, whether the picture reproduces as a photographic positive or negative. The relative merits of the two systems have been argued for a long time; both are used successfully but, since experience indicates some advantages inherent in the negative transmission method, this system has been firmly adopted in this country.

Figure 1 shows the video signal associated with the first two lines of a scanned image. When the picture is bright, the signal ampli-

tude is low, and conversely, when the picture is dark, the signal amplitude is high. The voltage variations that carry the video information are produced by the television camera as the scene is scanned. When the end of the first line is reached, the retrace occurs while the camera is inactive. During the period of inactivity, a

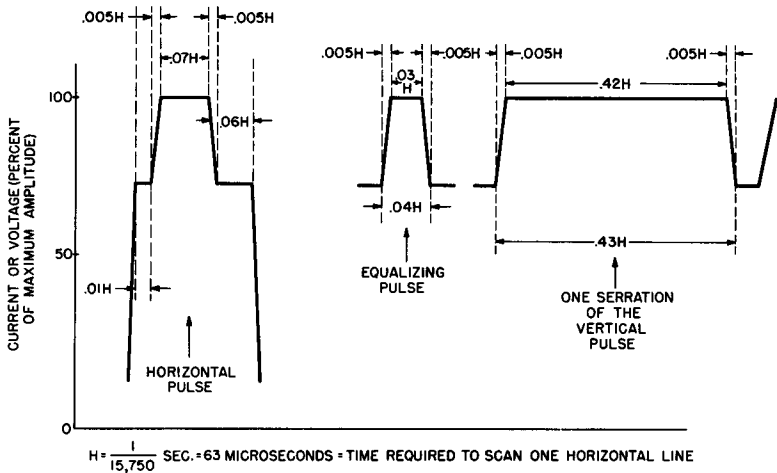


Fig. 2. Pulse width and other time intervals of pulses contained in the composite video signal in function of the time H required to scan one horizontal line.

blanking signal source superimposes on the signal a high-amplitude pulse that carries the relative voltage into the so-called black region.

At the beginning of the next scanned line, horizontal sync generators add an additional pulse to the top of the blanking pulse, so that synchronization between the transmitter and receiver is maintained. At the conclusion of each scanned field consisting of $262\frac{1}{2}$ lines, a serrated pulse of comparatively long duration is added to the video signal to preserve vertical synchronization from field to field. Equalizing pulses having very steeply rising and falling edges as well as short-duration periods, are inserted before and after the appearance of the serrated vertical sync pulse.

Thus, the standard video signal is a complex composite waveform containing many different frequencies and waveshapes all of which must be amplified without distortion to serve their purpose. This is the job of the video amplifier. Figure 2 illustrates the time intervals involved in producing the leading and lagging edges of the

various pulses, and the intervals between edges. These periods and pulse widths are important for the transient behavior analysis of video amplifiers. Time measurements are based upon percentages of the interval required for the scanning of one horizontal line, that is, $1/15,750$ second, or 63 microseconds.

3. Maximum Frequencies in the Video Range

The frequency range contained in the standard television video signal is discussed at length in many easily available books.¹ The important frequency components required in preserving the picture-signal waveform in accordance with American standards are those between 30 and 4,000,000 cycles. The lower limit is needed to handle changes in light that occur between successive frames. The upper limit is the minimum frequency that can produce good resolution between successive picture elements. Between these extremes are the frequencies that handle intermediate picture detail. Since any degree of detail between the upper and lower limits may be present in any scene, it follows that the transmission and reception systems must be designed to amplify all frequencies within this range without amplitude or phase distortion, to preserve the quality of the reproduced image.

Television studio equipment is generally built to handle a maximum video frequency of 5.5 mc or better. Thus when a reproduced image is viewed on the monitor with at least 5.0 mc effective in the picture, the definition of the picture is correspondingly high. On its way through the transmitter, however, sideband restriction causes a reduction of the maximum frequency to slightly over 4.0 mc. Since most commercial receivers insert an additional high-frequency attenuation, a maximum video frequency of about 3.5 mc might be expected. The definition of the picture is reduced accordingly. The entertainment market has shown remarkable tolerance for picture definition that is poorer than the best possible attainable with modern standards of transmission. This can be attributed to a demand of increased economy made possible in i-f and video amplifier systems where wideband demands can be reduced.

From the foregoing, it is possible to summarize the importance of uniform video amplification in this way: as the video frequency rises toward its upper limit, the detail and the definition of extended

¹Schure, A., *Basic Television*: New York: John F. Rider Publisher, Inc., 1958.

edges of images improve. The tone of extended edges becomes more uniform. These effects are very important in providing a reproduced image of good quality.

The frequency range from 30 cycles to 4 mc specifically applies to the American scanning standard of 525 lines per frame, 30 frames per second. The figures given below illustrate the differences in maximum video frequency for other possible scanning standards. (These figures are based upon an aspect ratio of 4:3 with equal vertical and horizontal resolution.)

343 lines, 30 frames per second . .	1.86 mc
525 lines, 30 frames per second . .	4.33 mc
1029 lines, 30 frames per second . .	16.7 mc

4. Structure of a Square Wave

According to the Fourier theorem, any continuous periodic function—such as the function of the picture current and time—may be represented by a sum of sine and cosine wave terms of different frequencies.

$$I = I_0 + I_1 \sin \omega t + I_2 \cos \omega t + I_3 \sin 2\omega t + I_4 \cos 2\omega t + \dots \\ + I_n \sin n\omega t + I_{n+1} \cos n\omega t$$

This theorem can be successfully applied to square waves to demonstrate their different frequency components. Sync and equalizing pulses used in television closely approximate the square waveform. The results of the Fourier analysis of the square wave are essential to an understanding of the video amplifier specifications. For a square waveform, the complete solution of the basic wave equation by the Fourier theorem provides a final series given by equation (1). In this equation, E is the instantaneous voltage, f is the fundamen-

$$E = \frac{2}{\pi} \left(\frac{\sin 2\pi ft}{1} + \frac{\sin 2\pi 3ft}{3} + \frac{\sin 2\pi 5ft}{5} + \frac{\sin 2\pi nft}{n} \right) \quad (1)$$

tal frequency of the wave, and t is time. As may be seen by inspection, the series contains components of the fundamental frequency (f), third harmonic (3f), fifth harmonic (5f), etc. Due to the divisors of each term, however, it is evident that the amplitude of each harmonic is inversely proportional to its frequency multiplying factors. For example, the amplitude of the tenth harmonic is only 1/10th of the amplitude of the fundamental frequency component.

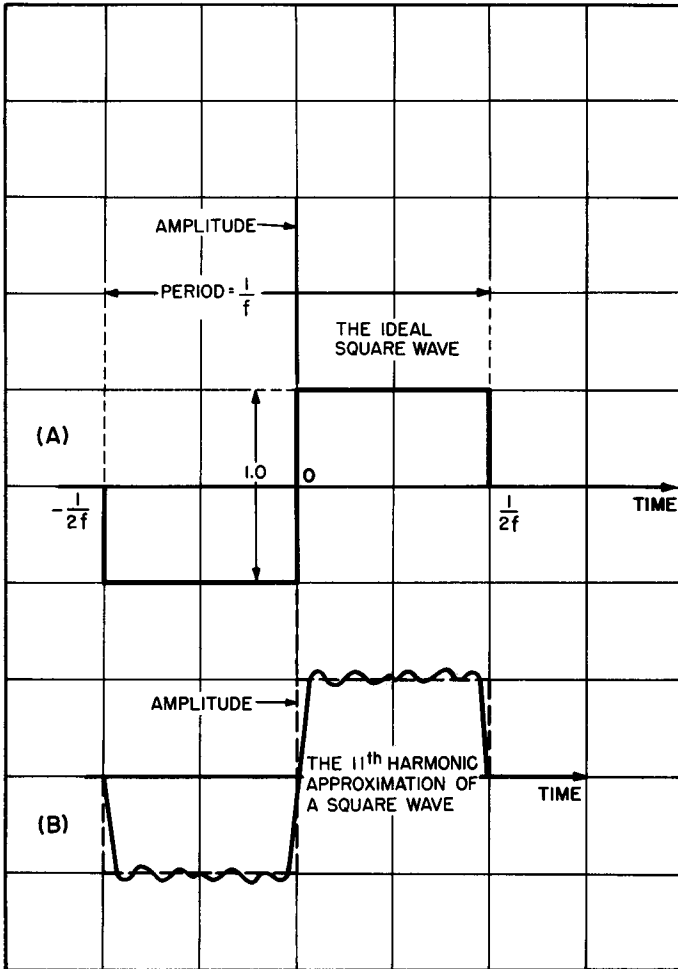


Fig. 3. (A) A square wave symmetrically placed around the origin, having a fundamental frequency f and an amplitude 1. (B) An approximate square wave obtained adding the fundamental and the 3, 5, 7, 9, and the 11th harmonic.

Thus, a square wave consists of the fundamental frequency plus a series of *odd* harmonics; the extent to which the harmonics are added in wave analysis is determined only by the precision with which the perfect square wave must be approached. In television work it is usually considered sufficient to include up to the eleventh harmonic. The approximation obtained is sufficient for all practical purposes. Such an approximation is shown in part B of Fig. 3.

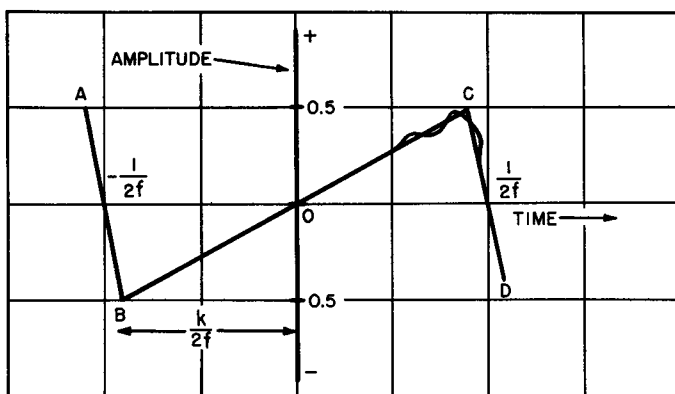


Fig. 4. A practical form of sawtooth wave used in television for horizontal and vertical scanning. Note that the retrace time is not zero and has a definite time relationship to the trace portion of the wave.

For example, the horizontal sync pulse shown in Fig. 2 has a pulse width of approximately:

$$.08H = .08 \times 63 \cong 5 \text{ microseconds}$$

Considering this duration as the period (T) of the square wave, the equivalent frequency is ($f = 1/T$) $10^6/5 \cong 200,000$ cycles. Using the 11th harmonic to approximate the square wave, or a frequency of $11 \times 0.2 \text{ mc} = 2.2 \text{ mc}$, it is evident that a wide-range video amplifier covering 30 cycles to 4.0 mc can easily handle this frequency. It is also clear, however, that if the bandpass of the video amplifier narrows for some reason, the picture definition suffers severely and synchronization is likely to be lost. For complete and reliable sync control of the horizontal sweep oscillator in a television receiver (or similar oscillator in radar), it is imperative that the square waveform be maintained without distortion throughout the video circuits of both the receiver and the transmitter. This, in turn, requires that the bandwidth be sufficiently great to pass the necessary high order harmonics.

5. Structure of a Sawtooth Wave

As a further application of Fourier's method of analysis to determine the harmonic content of specific waveforms, consider the sawtooth wave of Fig. 4. The total amplitude is taken as unity and the period as $1/f$ seconds. The retrace portions of the wave (AB and CD) are taken into account by assigning k as the fraction of the

total time occupied by the active or "forward" trace of the wave. The total active time of the wave is therefore k/f and the two half-intervals $k/2f$.

The series obtained by Fourier's theorem in this case is:

$$E = \frac{1}{\pi^2 (k - k^2)} \left(\frac{\sin \pi k}{1} \sin 2\pi ft + \frac{\sin 2\pi k}{4} \sin 2\pi 2ft + \frac{\sin 3\pi k}{9} \sin 2\pi 3ft + \dots + \frac{\sin n\pi k}{n^2} \sin 2\pi nft \right) \quad (2)$$

This series leads to an important conclusion, namely, that the amplitudes of the various harmonics diminish as the *square* of the harmonic order. Thus, high-order harmonics are evidently not nearly as important in bringing about a close approximation of a sawtooth wave as they are in the case of the square wave. As a matter of fact, a sufficiently good sawtooth wave for television scanning purposes may be obtained by adding to the fundamental sine wave all harmonics up to the sixth or seventh. We also note that, unlike the square wave, the sawtooth form contains both even and odd harmonics.

6. Video Signal Distortion

Wide frequency range amplifiers can distort the input signal in several ways:

(1) The amplifier may display imperfect frequency response that results in *amplitude distortion* within its assigned range;

(2) The dependence of phase angle on frequency may introduce phase distortion;

(3) The presence of noise or masking voltages imposes a limitation on the waveform of the reproduced signal; the extent of that limitation depends upon the magnitude of the masking voltages.

The structure of a square wave and sawtooth wave can be approximated by adding the proper number and types of harmonics to the fundamental sine wave. The computation of these series by Fourier's analysis is based upon the assumption that the transmitted harmonics are not acted upon by conditions which distort either amplitude or phase. In practical amplifiers, these conditions are never wholly met; this means that the conclusions previously drawn require certain modifications. Rather than attempt to establish these modifications through calculations, it is much more profitable

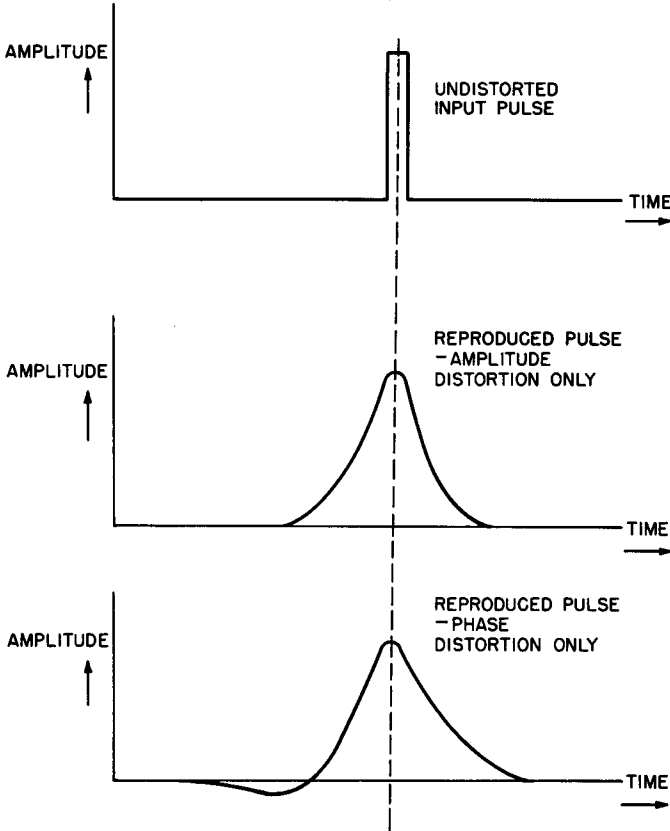


Fig. 5. Effects of amplitude and phase distortion on the shape of a non-periodic (transient) pulse.

to draw general conclusions about the effects from the shape of the non-ideal output wave characteristics of the amplifier.

The effects of amplitude and phase distortion can be best understood by considering what happens to a rectangular pulse of non-periodic (nonrepetitive) nature. It can be shown mathematically that when a pulse having a very large amplitude compared to its width is applied at the input,* two separate and distinct effects appear at the output if either type of distortion is present without the other. In the absence of phase distortion, amplitude distortion produces a form of symmetrical distortion in which the pulse is broadened and its sides take on a sloping appearance (Fig. 5). The

*This mathematical treatment involves the use of the Fourier integral and will not be discussed here.

peak of the reproduced hump coincides with the center of pulse flat-top. On the other hand, if phase distortion occurs in a system in which the amplitude response is perfect, the distortion is indicated by an asymmetrical hump around the central axis. Unfortunately, when both types of distortion appear simultaneously, it is very difficult to identify them merely by inspection of the output pulse shape.

Amplitude distortion may be analyzed comparatively easily by mathematical treatment; the same is not true of phase distortion. For this reason, little may be predicted about the general effects of phase distortion except in very simple cases. In specific transmission systems it can be measured. The rather important developments in this field during the past decade changed the accepted viewpoints concerning video amplifiers used in radar systems. In the early history of radar investigation and research, it was thought that the frequency response of a video amplifier should be flat and uniform over the covered frequency range, and that the cutoff outside this limit could be any convenient shape. As development continued, it was found that the phase distortion of the characteristic of a video amplifier is excessive when the cutoff is sharp. Such an imperfect phase characteristic produces overshoot oscillations of high amplitude, and may also lengthen the duration of the reproduced pulse. This led to the acceptance of the fact that a non-ideal frequency characteristic is often more tolerable than the troublesome phase characteristic produced by sharp cutoff. As a matter of fact, the most desirable design now appears to be one in which there may be some frequency discrimination, but where the linear phase characteristic is greatly emphasized by having gradual cutoff on each side of the range of frequencies to be handled by the video amplifier (Fig. 6). Gradual cutoff is far more conducive to linear phase response than sharp cutoff.

As a result, particularly in the design of radar video i-f amplifiers, transformer-coupled stages have been eliminated in favor of single-tuned circuits which have more nearly ideal phase characteristics. A similar set of design factors apply as well to the design of video amplifiers used in television.

7. Noise Effects on the Video Signal

The performance of a video amplifier may be seriously affected by certain types of *masking* or noise voltages developed within the

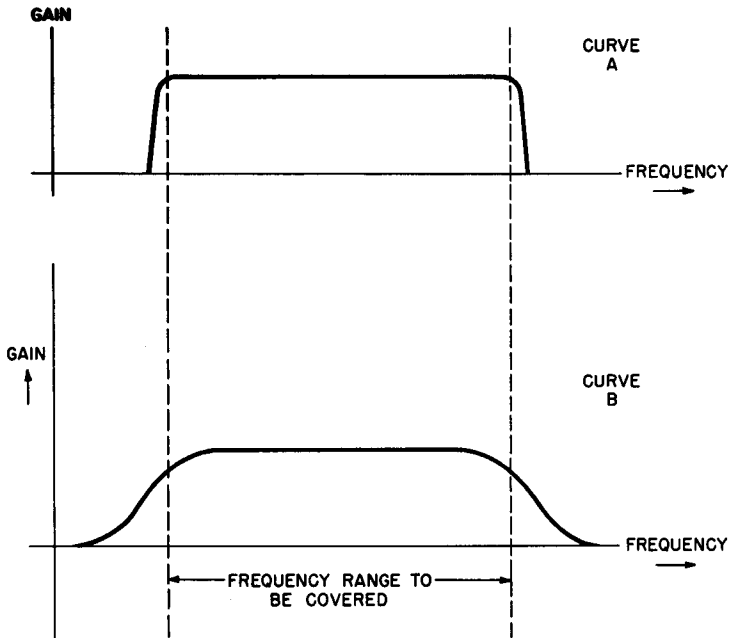


Fig. 6. Frequency response curves — Gain versus Frequency — Despite nonuniform frequency response, response of curve B is preferred over that of curve A because phase characteristic of the amplifier is improved.

video equipment. Other sources of noise voltages, such as those caused by carrier transmission of the video signal through an atmosphere in which electrical disturbances may give rise to masking voltages, are not discussed here.

The two important internally-generated types of noise are:

- (1) masking voltages due to thermal agitation in resistors
- (2) shot effect in the vacuum tubes.

The magnitude of a thermally-developed voltage is a function of the impedance of the conductor in which the noise arises, the absolute or Kelvin temperature, and the frequency range over which the amplifier is to provide essentially linear response. The exact relationship is:

$$E_{\text{eff}} = 7.42 \sqrt{TZ(f_2 - f_1)} \times 10^{-12} \quad (3)$$

in which E_{eff} is the effective thermal voltage, T is the Kelvin temperature, f_2 is the highest frequency to be attained in the amplification process, f_1 is the lowest frequency, and Z is the impedance of

the conductor (or the resistance of the resistor) in which the heat develops. To recognize the order of the magnitude of thermal masking voltage in the case of a television video amplifier, consider the following illustrative example problem.

Problem 1. What is the effective thermal noise voltage produced in a 100,000-ohm plate load resistor in a video amplifier of a television receiver at room temperature?

Solution. Room temperature is taken at 20°C which is the equivalent of 293°K. The range of frequencies covered in a television video amplifier may be assumed to be close to 4 mc (30 to 4,000,000 cycles). Substituting these values in equation (3):

$$\begin{aligned} E_{\text{eff}} &= 7.42 \times \sqrt{293 \times 10 \times 4 \times 10^{-12} \times 10^2 \times 10^3} \\ &= 7.42 \times \sqrt{11720} \times 10^{-7} \\ &= 80.8 \times 10^{-6} \text{ volts or } 80.8 \text{ microvolts} \end{aligned}$$

This rms thermal voltage (80.8 microvolts) becomes very significant in high-gain video systems. If the frequency response range is fixed by the nature of the video signal to be amplified, thermal masking voltage can be reduced only by reducing temperature, conductor impedance, or both.

Shot-effect noise is caused by the fact that the number of electrons emitted by a tube cathode varies from instant to instant. This variation generates a voltage in the circuit through which these electrons move.

The effects of masking voltage due to shot effect are generally more severe than those due to thermal origin. In the case of shot effect, the effective masking voltage depends upon the emission current, the frequency range, and the impedance of the coupling resistor, or other impedance, across which the shot-effect voltage appears. It is given by the equation:

$$E_{\text{eff}} = 5.64 Z \sqrt{I(f_2 - f_1)} \times 10^{-10} \quad (4)$$

The order of magnitude of shot-effect voltage may be seen from the following example.

Problem 2. What is the effective shot-effect noise voltage produced by the amplifier of Problem 1 if the emission current is 2.0 ma?

Solution. Substituting in equation (4), we obtain:

$$\begin{aligned} E_{\text{eff}} &= 5.64 \times \sqrt{2 \times 10^{-3} \times 4 \times 10^6 \times 10^3 \times 10^{-10}} \\ &= 5.64 \times \sqrt{20 \times 4 \times 10^5 \times 10^{-10} \times 10^{-2} \times 10^3} \\ &= 5.64 \times 8.95 \times 10^{-4} = 4.95 \times 10^{-3} \text{ volts } \cong 5 \text{ millivolts} \end{aligned}$$

Thus, even for a relatively low emission current, the shot-effect voltage is almost 5 millivolts—a much larger voltage than was caused by thermal agitation under similar circumstances.

From these examples we draw the following conclusion: The bandwidth of the amplifier should not be made any greater than is required to transmit the necessary video information, since the noise (both kinds) varies as the square root of the frequency range.

Noise problems become increasingly severe as the amount of video information increases. For example, in a television system, any increase in the number of picture elements to be passed through the video amplifier extends the upper video frequency range and results in an increase in the noise problem.

8. Review Questions

1. How is a *video signal* defined?
2. What is the difference in the appearance of the video signal for negative picture transmission systems and positive picture systems?
3. What are the chief characteristics of the television video signal that call for wideband video amplification? Fully explain.
4. List the pulse types present in the television composite video signal in the order of ascending pulse width.
5. Describe the relationship between the maximum video frequency transmitted and the resulting picture detail and definition.
6. Since it is agreed that better detail is obtained with 1000 scanning lines at 30 frames per second than with the present 525 lines at 30 frames per second, why has not a higher-number line structure been adopted in this country? (Give at least three cogent reasons.)
7. Explain how equation (1) shows that even-order harmonics do not play any part in the generation of a square wave.
8. Explain why a video amplifier of insufficient bandpass might satisfactorily amplify a sawtooth wave, but not a square wave of the same fundamental frequency.
9. Describe the separate effects of amplitude and phase distortion for a short-duration input pulse.
10. What are the two most important types of internally-generated noise voltages? Which one is the more severe? Does the noise problem increase or decrease when video amplifiers are designed for greater bandwidth? Explain.

Chapter 2

THE UNCOMPENSATED AMPLIFIER

9. Requirements of a Video Amplifier

A video amplifier must possess other fundamental properties aside from the necessary gain requirements, to provide *optimum* performance. The concepts that underlie some of these properties were developed in the previous chapter; we must add several important other specifications to complete the listing of requirements:

(a) The video amplifier must respond to all frequencies within its passband with approximately equal amplification. In the case of a television video amplifier system (comprising several individual stages), constant gain to within a few decibels must be provided from 30 cycles up to 4 mc to maintain good picture reproduction. Some quality of performance is sacrificed for the sake of economy, if the passband of the system is reduced to 3 mc or even to 2.5 mc. For portable, small-screen television receivers the loss of definition incurred by narrowing the amplifier bandwidth to this extent is not as evident as it is on the large screens of home receivers, and can be tolerated.

(b) There must be as little time-delay discrimination (phase distortion) as possible. For good television picture reproduction, the delay at the high frequencies must not be greater than approximately .05 microsecond. If, at the same time, there is a linear relationship between phase angle and frequency, then the time-delay distortion is further minimized.

(c) Noise must be reduced so that the signal-to-noise ratio is as

high as possible. Reduction of the detrimental effects of thermal agitation and shot effect is accomplished by keeping the operating temperatures as low as possible, selecting the best vacuum tubes for the task at hand, using low plate and screen currents to keep total emission current down, and restricting the amplifier bandwidth so that only the desired range is included in the passband.

(d) The video amplifier must be capable of yielding the required output voltage or power.

(e) The video amplifier must be designed to present the correct terminal impedances to the driving stage or to its load.

(f) The polarity of the output voltage or power of the amplifier must be correct for the task it has to perform. A television video amplifier must provide a signal of the correct polarity so that the stages that follow it will produce a picture having the right dark and light relationships on the picture tube screen.

(g) Particularly in television, video amplifiers must retain in their output signal a d-c component that represents the average brightness of the televised scene. This component is combined with the fixed bias applied to the picture tube to establish an average brightness level.

10. Basic Analytical Approach

Linear video amplification is a process in which the signal voltage, current, or power is amplified without significant change in waveform when the input is in the form of a pulse or a wide band of frequencies. The term *linear* implies that a direct proportionality exists between input and output voltage, current, or power.

The basic video amplifier circuit is illustrated in Fig. 7A and its constant-voltage-generator equivalent circuit in Fig. 7B. As high-gain pentodes are normally used in video amplifiers, this discussion will be related to these tubes except where otherwise noted. Although it is anticipated that the reader has studied the development of the fundamental amplifier equations dealing with gain and phase, a review of these essential concepts is presented below.

The voltage gain of an amplifier is defined as:

$$\text{Gain} = \frac{e_o}{e_g} \quad (5)$$

where e_g is the instantaneous signal voltage applied between the

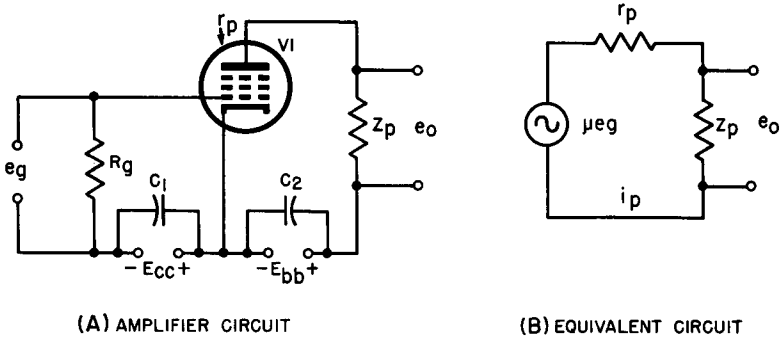


Fig. 7. (A) Fundamental uncompensated amplifier of the untuned type. (B) Constant-voltage-generator equivalent circuit.

grid and cathode of the tube, and e_o is the instantaneous voltage appearing across the plate impedance Z_p .

The phase shift introduced by the amplifier is defined as the angle between e_g and e_o . A normal amplifier produces a phase shift of 180° by inherent action; this angle is designated by the Greek letter “phi” (ϕ). Due to the presence of reactive components in the various tube circuits, however, additional phase shift almost always appears. If this phase angle is denoted by ϕ_1 , then:

$$\phi_1 = \phi - 180^\circ \tag{6}$$

The dynamic plate resistance r_p of a tube is related to its amplification factor μ and its transconductance g_m by the expression:

$$g_m = \frac{\mu}{r_p} \tag{7}$$

The instantaneous current i_p flowing in the plate circuit is readily obtained from the equivalent circuit (Fig. 7B) as:

$$i_p = \frac{\mu e_g}{r_p + Z_p} \tag{8}$$

The output voltage that develops across the plate impedance Z_p is, therefore:

$$e_o = \frac{\mu e_g Z_p}{r_p + Z_p} \tag{9}$$

Substituting this voltage of e_o in equation (5) gives:

$$\text{Gain} = \frac{\mu Z_p}{r_p + Z_p} \quad (10)$$

The additional phase shift caused by the reactive nature of Z_p is given by equation (11). In this equation, X_p is the reactive component and R_p the resistive component of Z_p .

$$\phi_1 = \tan^{-1} \frac{X_p r_p}{R_p^2 + R_p r_p + X_p^2} \quad (11)$$

Since high-gain pentodes have a dynamic plate resistance r_p 10 or more times greater than the plate load impedance Z_p , the latter may be neglected and equations (10) and (11) rewritten in simpler form:

$$\text{Gain} = \frac{\mu Z_p}{r_p} = g_m Z_p \quad (12)$$

$$\phi_1 = \tan^{-1} \frac{X_p}{R_p} \quad (13)$$

With respect to the performance of video amplifiers, in particular the manner in which they meet the requirements for uniform gain and small time-delay as given in Section 6 (a) and (b), equations (12) and (13) convey much information. For example, equation (12) demonstrates that the tube gain is a function of the transconductance and the plate load impedance. While transconductance is independent of frequency, Z_p certainly is not. Hence, an uncompensated amplifier cannot serve over a wide band of frequencies with the required degree of uniform amplification. Since Z_p contains (unavoidably) inductive and capacitive components as well as resistive components, means must be found to reduce the variations of Z_p with frequency. This procedure is called *amplifier compensation*, and is discussed in succeeding chapters.

The generally undesirable phase shift introduced in the amplifier circuit through the reactive load components must be linearly proportional to the frequency (Section 9b). Figure 8 shows the ideal phase-frequency characteristic of an amplifier in which the time delay of all frequency components is fixed. The phase angle must be linearly proportional to the frequency to effect a constant time de-

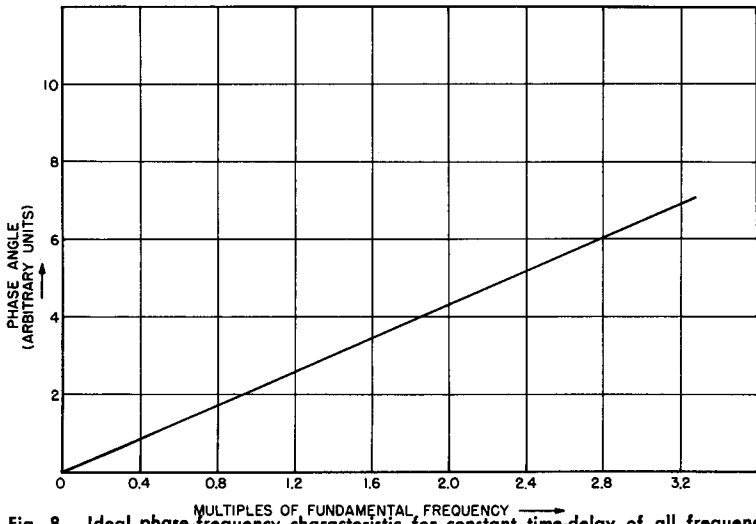


Fig. 8. Ideal phase-frequency characteristic for constant time-delay of all frequency components.

lay. For small phase angles, where the magnitude of the tangent approaches the angle itself, equation (13) may be rewritten:

$$\phi_1 = \frac{X_p}{R_p} \quad (14)$$

Assume for the moment that a capacitive reactance, lumped or distributed, appears across Z_p ; in effect, such a capacitance is equivalent to an inductive reactance in series with Z_p , and whose reactance is directly proportional to frequency, hence that ϕ too is also proportional to frequency. However, that this is only true for *small* phase angles. As the phase angle increases, there is greater and greater deviation from the required proportionality, accompanied by increasing phase distortion.

11. High-Frequency Behavior of an Uncompensated Amplifier

Video amplifier practice in modern equipment utilizes resistance-capacitance coupling almost to the exclusion of other types of coupling circuits. A few television receivers employ direct coupling. The advantages to be gained, however, are insufficient justification for the added circuit complications regarding d-c supply voltages.

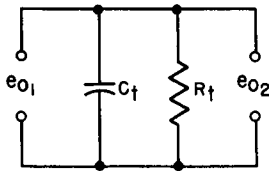
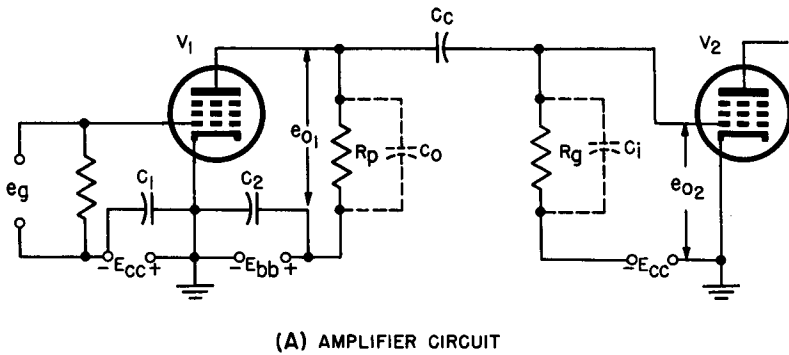


Fig. 9. (A) Normal uncompensated R-C coupled amplifier. (B) Equivalent circuit for high frequencies.

An R-C coupled circuit in an uncompensated amplifier and its equivalent circuit at high frequencies are shown in Fig. 9. At high video frequencies, the coupling capacitor has negligible reactance and can be considered as a short circuit. The equivalent plate impedance of V_1 consists of the normal plate load resistor R_p , shunt capacitances appearing across it due to plate-to-cathode capacitance and due to stray wiring capacitance C_0 , the grid return resistor of the following stage R_g , the input capacitance across this resistor C_i consisting of the normal grid to cathode, the Miller effect, and the stray wiring capacitances. In the equivalent circuit, all the capacitances are represented by C_t such that:

$$C_t = C_0 + C_i \tag{15}$$

and all the resistances by R_t such that:

$$R_t = \frac{R_p R_g}{R_p + R_g} \tag{16}$$

Therefore, the total equivalent load impedance in the plate circuit of VI consists of C_t and R_t in parallel and is given by the following expression:

$$\frac{1}{Z_p} = \frac{1}{R_t} + \frac{1}{jX_t} \quad (17)$$

in which jX_t is the vector representation of $1/(2\pi fC_t)$. The frequency f is the nominal operating frequency. Thus, the output impedance may be obtained by solving for Z_p :

$$Z_p = \frac{-jR_tX_t}{R_t - jX_t} \quad (18)$$

The expression in equation (18) may now be converted into the form given in equation (19) by dividing numerator and denominator by $-jX_t$:

$$Z = \frac{R_t}{1 + \frac{jR_t}{X_t}} \quad (19)$$

Removal of the j operator then provides an expression for the magnitude of Z_p :

$$Z_p = \frac{R_t}{\sqrt{1 + \frac{R_t^2}{X_t^2}}} \quad (20)$$

And substituting $1/(2\pi fC_t)$ for X_t in the above gives:

$$Z_p = \frac{R_t}{\sqrt{1 + 4\pi^2 f^2 R_t^2 C_t^2}} \quad (21)$$

In its form as given equation (21), the magnitude of the impedance cannot be interpreted easily. By utilizing a bit of legitimate mathematical manipulation, the significance of this development can be substantially clarified. Let us assume that we are dealing with a second frequency f_1 such that:

$$f_1^2 = \frac{1}{4\pi^2 R_t^2 C_t^2}$$

$$f_1 = \frac{1}{2\pi R_t C_t} \quad (22)$$

Substituting this value of f_1^2 into equation (21) yields this simple form:

$$Z_p = \frac{R_1}{\sqrt{1 + \frac{f^2}{f_1^2}}} \quad (23)$$

Finally, the gain of the amplifier may be obtained from equations (12) and (23):

$$\text{Gain} = g_m Z_p = \frac{g_m R_t}{\sqrt{1 + \frac{f^2}{f_1^2}}} \quad (24)$$

It is important now to interpret this equation in a physical sense. As previously defined, the frequency f is the operating frequency or range of frequencies for which the amplifier exhibits a gain of $g_m R_t$ since, if the load impedance is completely independent of frequency—that is, if it is purely resistive— $Z_p = R_t$. This is the highest possible gain.

In the actual case of a practical amplifier, it is possible to derive the meaning of f_1 . If the ratio $f^2/f_1^2 = 1$, then the frequency of operation must be f_1 since only for this condition can the ratio equal unity. But when this is the case, equation (24) may be solved for gain to obtain:

$$\text{Gain} = \frac{g_m R_t}{\sqrt{1 + 1}} = .707 \times g_m R_t$$

$$\text{Relative gain} = \frac{\text{Gain}}{\text{Gain at mid frequency}} = \frac{\text{Gain}}{g_m R_t} = .707$$

Thus, when the operating frequency is f_1 , there is a loss of gain of 29.3% or the voltage is “down” 3 db from its value when the operating frequency is equal to f . It is for this reason that f_1 is generally characterized as the frequency above which the amplifier response curve ceases to be *flat*, and is used as a reference by most authorities for calculating amplifier gain equations at high frequencies.

By drawing a curve (Fig. 10) of f/f_1 against gain, it can be seen that an amplifier provides a tolerably flat response through the middle frequencies up to the point when $f = f_1$. At this point, the relative gain drops below 70.7% of its maximum value for middle frequencies, making compensation necessary if higher frequencies

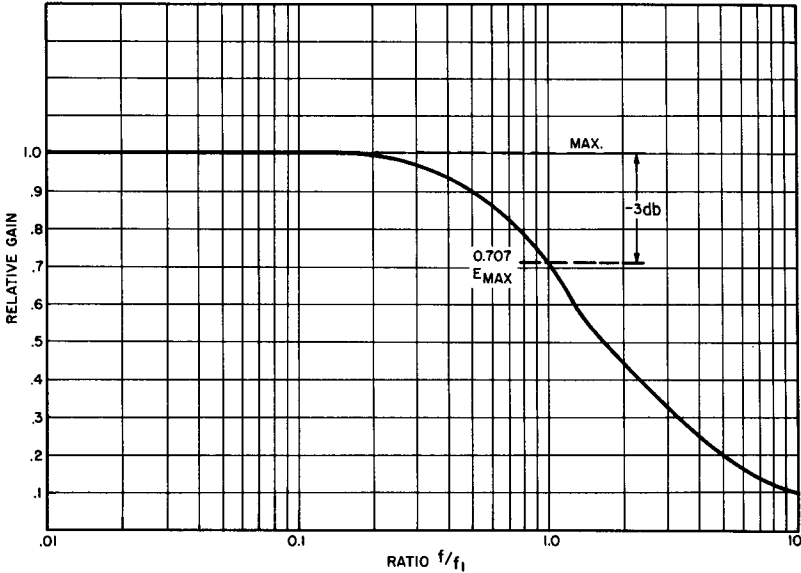


Fig. 10. Curve of relative gain versus the ratio of operating-to-reference frequency f/f_1 . This curve illustrates the importance of maintaining f_1 high by reducing C_t .

are to receive uniform amplification. Since the actual value of f_1 depends entirely upon R_t and C_t , reducing both these quantities will extend f_1 into the desirable higher ranges (equation 22). On the other hand, reducing R_t causes a proportional gain reduction (equation 24), hence the high-frequency response of an uncompensated amplifier is seen to improve if g_m is made as large as possible and C_t as small as possible.

12. Phase Shift of Uncompensated Amplifier at High Frequencies

The phase shift ϕ , added to the normal 180° phase shift produced by the tube is given by equation (18).

$$\phi_1 = \tan^{-1} \frac{X_p}{R_p}$$

To determine the equivalent values of X_p and R_p , equation (18) is first expanded as follows:

$$Z_p = \frac{R_t X_t^2 - j X_t R_t^2}{R_t^2 + X_t^2} \quad (25)$$

The resistive component of the output impedance Z_p is, therefore:

$$R_p = \frac{R_t X_t^2}{R_t^2 + X_t^2} \quad (26)$$

And the reactive component:

$$X_p = \frac{-X_t R_t^2}{R_t^2 + X_t^2} \quad (27)$$

If the ratio X_p/R_p is now set up in terms of equations (26) and (27), we obtain:

$$\frac{X_p}{R_p} = -\frac{R_t}{X_t} \quad (28)$$

Substituting the right hand member of the equation just obtained in equation (13):

$$\phi_1 = \tan^{-1} -\frac{R_t}{X_t} \quad (29)$$

By definition, we have:

$$X_t = \frac{1}{2\pi f C_t}$$

Equation (29) may be revised to read:

$$\phi_1 = \tan^{-1} - 2\pi f C_t R_t \quad (30)$$

However, we have previously defined f_1 as the frequency at which the amplifier gain is down 29.3% or 3 db from the operating frequency as:

$$f_1 = 1/2\pi R_t C_t \quad (\text{equation 22})$$

Making the substitution in equation (30) we may write:

$$\phi_1 = \tan^{-1} \frac{f}{f_1} \quad (31)$$

A physical interpretation of this equation is easily obtained by setting up a graph in linear coordinates in which the added phase angle ϕ_1 due to the components of the amplifier is shown as a function of f/f_1 . The results are shown in Fig. 11. An ideal linear phase characteristic is also plotted on the same axes for comparison, and to enable the reader to estimate the extent of phase shift for different values of f/f_1 .

For example, if we assume very small values for R_t and C_t then f_1 is quite large. This, of course, means that the amplifier handles a

large range of frequencies with virtually uniform amplification since there is appreciable *spectrum* between f and f_1 , the frequency at which the gain is down 3 db. At the same time, the ratio f/f_1 is proportionately small and approaches zero as R_t and/or X_t approach zero. Since the tangents of small angles can be approximated by the angles themselves, for the condition just described it is apparent that

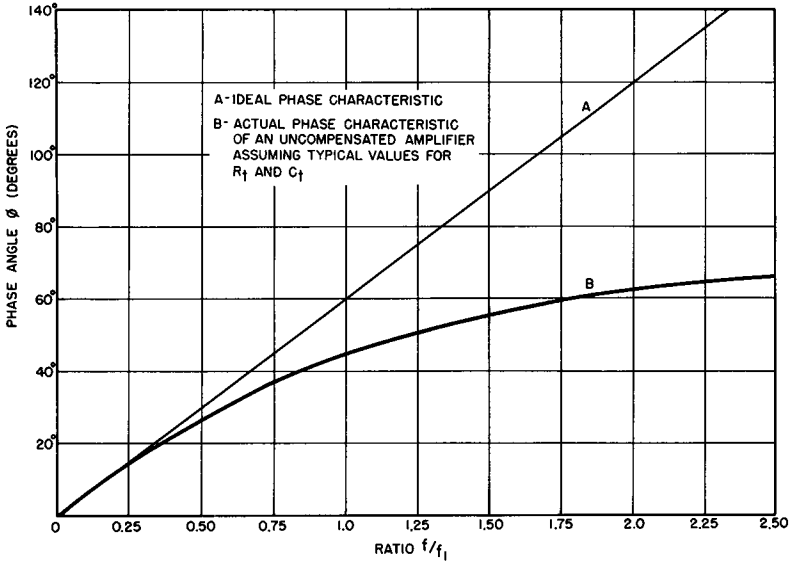


Fig. 11. Phase characteristics of uncompensated amplifier compared with ideal characteristic.

ϕ_1 increases with frequency for a time in a linear manner. Thus the ideal and actual phase characteristics coincide for small values of f/f_1 . However, when $f = f_1$ (the gain of the amplifier is down 3 db) there is an angle of approximately 16° difference between the ideal phase characteristic and the actual one.

It is apparent that the upper frequency limit of the useful performance of an amplifier without compensation is dictated by the choice of values of R_t and C_t . As R_t and C_t grow, the upper frequency limit becomes increasingly depressed; similarly, as these factors are made smaller, the upper frequency limit is extended further and further.

Consider what happens to gain, however, when R_t is reduced. Equation (24) shows that amplifier gain is directly proportional to

R_t , hence R_t cannot be reduced indiscriminately, otherwise the value of the amplifier may be destroyed altogether. Therefore, C_t is the factor which must be reduced if amplifier gain is to be maintained while bandwidth is increased. The total input capacitance C_t is composed of stray wiring capacitance, capacitance between tube elements, and an additional dynamic capacitance due to the Miller Effect. This is expressed by the equation:

$$C_t = C_w + C_i + C_{gp}(1 + \text{Gain}) \quad (32)$$

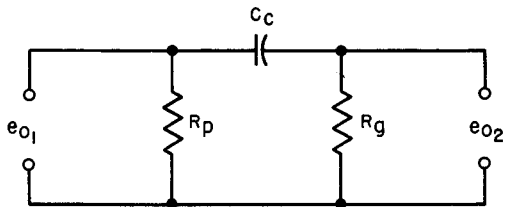
where C_w is the wiring capacitance, C_i is the total interelectrode capacitance not including the grid-plate capacitance of the tube, and C_{gp} is grid-plate capacitance. Thus, as a result of the Miller Effect, normal input capacitances are increased to a substantial extent, making it extremely difficult to achieve low values of C_t when high-gain tubes are used.

The foregoing considerations clearly demonstrate that a wide-band video amplifier can have good amplitude and phase characteristics at high-frequencies only if C_t is made very small. Since this appears to be very difficult, if not impossible in the present state of the art, compensatory measures must be adopted to make up for the unavoidable deficiencies existing in vacuum tube circuits.

13. Low-frequency Response of Uncompensated Video Amplifier

An analysis of the low-frequency response of an uncompensated video amplifier is based on a somewhat different equivalent circuit than that used for high-frequency analysis. The low-frequency equivalent circuit is given in Fig. 12;

Fig. 12. Low-frequency equivalent circuit of uncompensated video amplifier.



it is based upon the two-stage amplifier circuit presented in Fig. 9. We must remember that the input capacitance and output capacitance forming C_t in the high-frequency analysis no longer play a part here since C_t has too high a reactance at these low frequencies to affect the amplifier operation

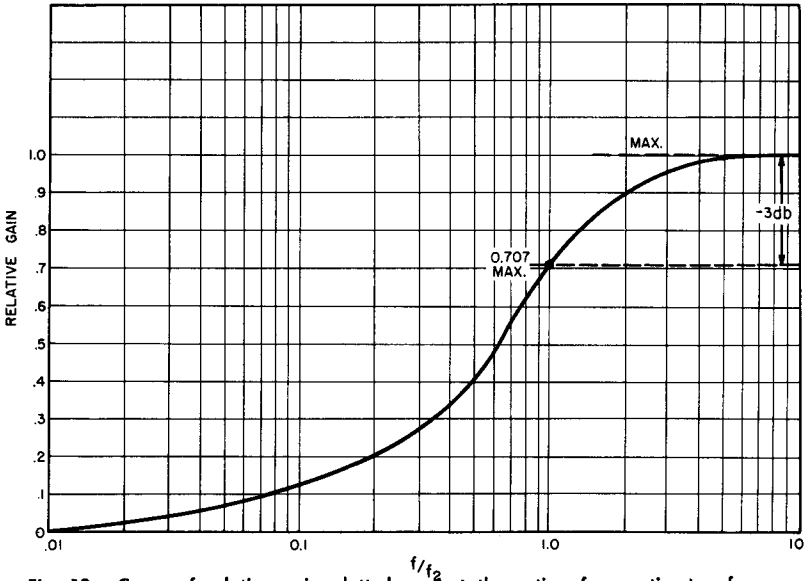


Fig. 13. Curve of relative gain plotted against the ratio of operating-to-reference frequency f/f_2 for the low-frequency gain equation (38).

to any significant extent. The coupling capacitor C_c , however, although considered a short circuit at high frequencies, must now be taken into account since its reactance is sufficiently great to affect the low-frequency response. Thus, the low-frequency equivalent circuit displays an output impedance consisting of the plate load R_p in parallel with the series-combination comprising the coupling capacitor C_c and the grid resistor R_g of the following stage. Since the driving voltage for the next stage (e_{o2}) appears across R_g , it is first necessary to determine how much of the input voltage developed across R_p is applied to the output circuit. C_c and R_g form a voltage divider. The output voltage is taken across R_g .

$$\frac{e_{o2}}{e_{o1}} = \frac{R_g}{R_g - jX_c} \quad (33)$$

in which X_c is the reactance of coupling capacitor C_c ($X_c = 1/2\pi f C_c$), and e_{o2}/e_{o1} is the fraction of the voltage developed across R_p that is delivered to the following tube grid. Taking into account

the transconductance g_m of the first tube, e_{o1} is a function of this g_m and the value of R_p as well as the original signal voltage to the first grid. The gain available across R_p is, therefore:

$$\text{Gain} = g_m R_p \quad (34)$$

Since only a fraction of the voltage developed across R_p is delivered to the next stage, the gain available at the second tube grid is:

$$\text{Gain} = g_m R_p \left(\frac{R_g}{R_g - jX_c} \right) \quad (35)$$

Substituting the component values of X_c and simplifying the equation, we can write:

$$\text{Gain} = \frac{g_m R_p R_g \times 2\pi f C_c}{\sqrt{1 + (2\pi f C_c R_g)^2}} \quad (36)$$

In order to arrange the terms of this equation in a more easily interpretable form, it is again helpful to define a second frequency f_2 as follows:

$$f_2 = \frac{1}{2\pi R_g C_c} \quad (37)$$

Then, using this definition of f_2 in equation (36), it is possible to write the latter in this form:

$$\text{Gain} = \frac{g_m R_p \left(\frac{f}{f_2} \right)}{\sqrt{1 + \left(\frac{f}{f_2} \right)^2}} \quad (38)$$

As in the high-frequency case, a curve of relative gain against f/f_2 may be plotted from this equation. This curve is drawn in Figure 13. At the frequency for which the ratio $f/f_2 = 1$ it will be observed that the gain has fallen off again by 3 db. Therefore, satisfactory low-frequency performance is obtainable without compensation only by keeping f_2 as small as possible; since f_2 is a function of R_g and C_c (equation 37), R_g and C_c should be made as large as other conditions permit to extend the low-frequency response of the amplifier. The effect of large values of R_g and C_c at high frequencies are evidently negligible since at these frequencies C_c behaves like a short-circuit and R_g shunts the comparatively low-resistance (R_p) without affecting it significantly.

There are three important factors that limit the values to which R_g and C_c may be raised:

(1) When R_g and C_c are made too large, the time constant of the combination may increase sufficiently to cause motorboating.

(2) When C_c is made too large, the stray capacitance to ground may increase sufficiently to spoil the high-frequency response of the amplifier.

(3) There is always danger of "gas" current in the grid circuit causing intolerable bias changes when this current flows through too large a grid resistor.

The last factor is the reason why tube manufacturers specify maximum resistance values for grid resistors for all tubes.

14. Phase Shift of Uncompensated Amplifier at Low Frequencies

By a method similar to the one used to determine the phase characteristic of an amplifier at high frequencies, it can be shown that the added phase shift produced by an uncompensated amplifier at low-frequencies is given by the equation:

$$\phi_1 = \cot^{-1} \frac{f}{f_2} \quad (39)$$

Figure 14 provides a graphical picture of the phase characteristic of the amplifier up to a coordinate where $f = f_2$. There is a significant departure from linearity. When the frequency is low, the relatively small phase shift measured in degrees equals the relatively large time delay measured in seconds. Here, again, an uncompensated amplifier may be expected to perform well at low frequencies only if f_2 is made very small by selecting R_g and C_c as large as possible, and taking into account the adverse effects of selecting these components too large.

15. Effect of Bypass Capacitors on Wideband Performance

In triode amplifier circuits where bias is obtained with either a cathode resistor or an external source, the signal current as well as the d-c flow through the bias impedance, unless adequate steps are taken to prevent it. The voltage drop across a cathode bias resistor, for example, consists therefore of a d-c and a signal component. The polarity of the signal voltage drop is such as to oppose the voltage of the applied signal and thereby reduce amplification (Fig. 15).

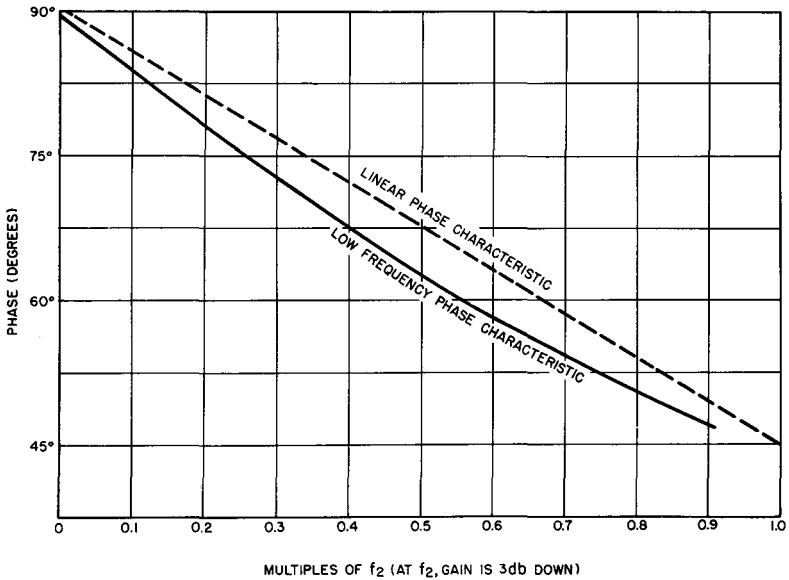
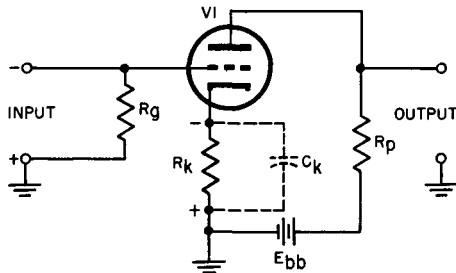


Fig. 14. Phase characteristic of an amplifier at low frequencies.

The particular instant in time selected in Fig. 15 shows the input signal going negative with respect to ground; a negative-going grid causes a reduction in plate current so that the voltage drop across R_k decreases. This makes the cathode more negative; thus, the cathode potential swings in the same direction as the input voltage, preventing the voltage difference between these input elements from reaching the same values that would be reached without this form of *degenerative feedback*.

If a loss of gain can be tolerated in a given amplifier and is due to this cause, the frequency response is improved rather than degraded since degenerative feedback serves to reduce distortion. On

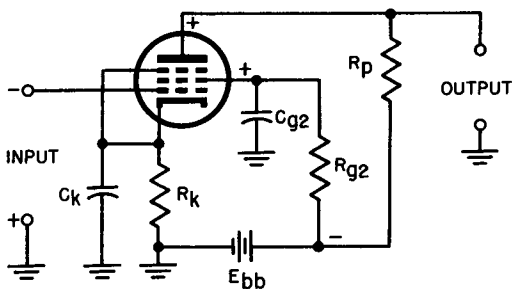
Fig. 15. Signal voltage drop across cathode-bias resistor acts in phase opposition to the applied signal, causing a gain reduction.



the other hand, should the gain be required, then R_k is normally bypassed by a large capacitor C_k (Fig. 15) to maintain the voltage drop across the resistance constant. Capacitors, however, are frequency-sensitive; C_k has a smaller reactance for high frequencies than for lows. Hence, although the high-frequency gain will be restored to approximately its original value by connecting C_k across R_k , the same is not necessarily true of the low frequencies for which C_k may not be an adequate bypass. This means that insufficient cathode bypass capacitance may result in a loss of gain at the low-frequency end of the amplifier's response spectrum with a consequent degradation of the response curve. Methods for determining the adequacy of cathode bypassing action will be discussed later.

Since most high-gain amplifiers now utilize pentodes, the screen-grid circuit assumes importance in any design procedure. As in the case of the cathode circuit, a signal voltage applied to the control grid of a pentode varies the screen current as well as the plate cur-

Fig. 16. Signal voltage drop across the screen resistor R_{g2} reduces gain.



rent. In Fig. 16, a similar condition to that of Fig. 15 has been established; a negative-going input voltage is assumed. This brings about an instantaneous reduction in screen voltage so that the voltage drop across R_{g2} decreases, making the net screen voltage higher than before. While a negative-going grid tends to reduce plate current, a positive-going screen tends to increase it, hence the drop across R_{g2} reduces the plate current swing. The net effect is one of gain reduction. A bypass capacitor such as C_{g2} connected from the screen grid to ground (cathode) is effective in restoring the loss of gain due to this cause for the high frequencies but may not be fully effective for the lows. The reason for this behavior difference is identical with that given previously for the possible inadequacy of a cathode bypass capacitor.

By analogy with the plate circuit action, we can write the mathematical value of the ratio of actual amplification with R_{g2} present to the ideal amplification if R_{g2} is short-circuited as follows:

$$\frac{\text{Actual Gain}}{\text{Ideal Gain}} = G' = \frac{r_{g2}}{r_{g2} + R_{g2}} \quad (40)$$

in which G' is the ratio under investigation, r_{g2} is the dynamic screen grid resistance of the specific tube, and R_{g2} is the resistance in series with the screen grid. Obviously, the condition we like to realize makes $G' = 1$ because in this case the actual gain approaches or equals the ideal gain. To do this, R_{g2} may be bypassed by a large capacitor so that equation (40) may be rewritten:

$$G' = \frac{r_{g2}}{r_{g2} + Z_{g2}} \quad (41)$$

Z_{g2} is the total screen circuit impedance consisting of the screen resistor and the bypass capacitor connected in parallel. Reducing the reactance of C_{g2} produces an equivalent decrease in Z_{g2} and, if the capacitive reactance is substantially smaller than the equivalent resistance of r_{g2} and R_{g2} , G' approaches unity. If the capacitance is large enough to give the required low reactance for high frequencies, but not sufficiently large to keep the reactance low for the low-frequency end of the amplifier range, the low-frequency response is degraded, as explained previously.

16. Pulse Response of Uncompensated Amplifier

As we have shown in Section 4 and in equation (1), a square wave consists of the fundamental sine-wave frequency plus a series of odd, harmonically related sine waves. The significance of harmonics of ascending number, determined by the relative amplitude of each, was shown to be function of harmonic number; the amplitude of the odd harmonic is inversely proportional to the harmonic number. In addition, all of the harmonics bear a zero-phase angle relationship to the fundamental frequency at the start and end of each fundamental cycle. Although *all* the odd harmonics play a part in the structure of a square wave (up to the harmonic having an infinite number), a good approximation of a square wave may be obtained by synthesizing the fundamental with all the odd harmonics up to 11 (or 15 or more, if the demands of the system are more rigorous).

VIDEO AMPLIFIERS

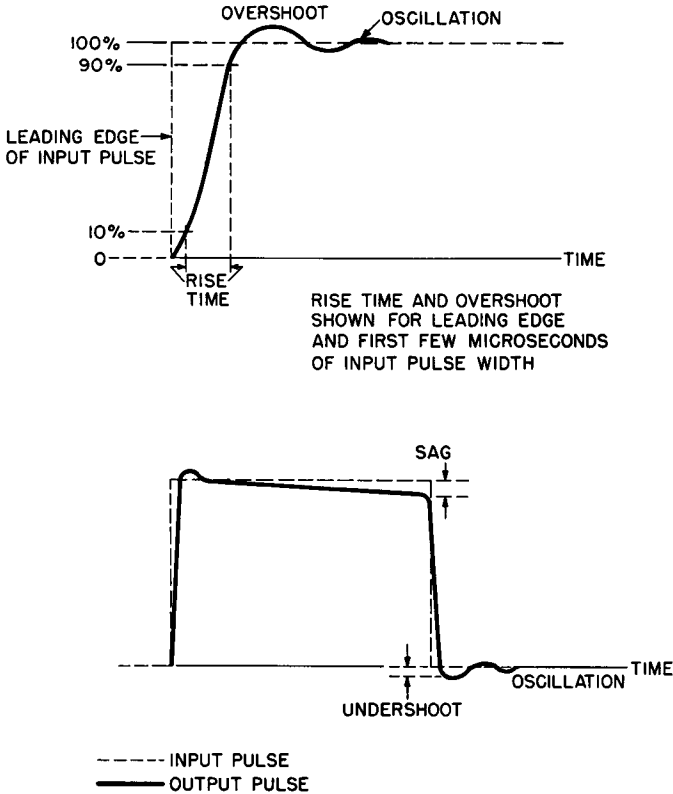


Fig. 17. Square waveform distortion resulting from passage through an amplifier with nonlinear amplitude (gain) and phase characteristics.

If the input signal to the video amplifier approximates a perfect square wave, it is delivered to the output impedance without change of form only if the amplifier has a very flat frequency response and a linear time-delay characteristic. We know that linearity of time-delay is appreciably more important in low-frequency amplification than in high-frequency work. Nonuniform frequency response or nonlinear time-delay, therefore, result in a change of the form of the input square wave. Thus, the use of square-wave input voltages for what is often called transient response analysis (the observation of the resulting output waveshapes on an oscilloscope) provides an excellent and rapid method of determining bandwidth and phase response of wideband amplifiers.

A perfect rectangular pulse is characterized by zero rise and fall times and a flat top without dips or humps. A reproduced pulse can have several defects (points of difference between it and the input wave shape), all of which are important in evaluating amplifier performance. These are illustrated in Fig. 17 and described as follows:

Rise time. A perfect square wave has zero rise time at its leading edge. An amplified pulse may have various rise times measured in microseconds from its 10% amplitude point to its 90% amplitude point. The greater the rise time the more serious is the distortion.

Fall time. A perfect square wave has zero fall time at its trailing edge. Actual fall time of an output pulse generally approaches its rise time in a well-designed amplifier. In addition, fall time is not as significant in amplifier analysis as rise time, and is generally omitted from such discussions.

Overshoot. A perfect square wave has no overshoot. The larger the overshoot, the more imperfect is the amplifier.

Undershoot. An undershoot follows after a rapid amplitude decline. Undershoots may or may not be associated with sags (described below); in that case, however, the undershoot amplitude is the same as the sag amplitude if the amplifier is linear.

Oscillation. Both overshoots and undershoots are sometimes accompanied by damped oscillations following their incidence. Such oscillations are particularly troublesome if their duration is greater than a small fraction of the pulse-width and must be minimized for good amplifier response.

Sag. Sag describes the fall-off of a flat-top pulse amplitude with time during a single period. Sag most often results from poor low-frequency performance on the part of the amplifier and is generally due to improper time-delay characteristics. The undershoot that follows a sag is generally of the same order of amplitude as the sag itself, if the amplifier approaches closely to linear performance.

Rise time is an extremely important characteristic of the performance of an amplifier intended to handle a wide range of frequencies. Analysis shows that rise-time is related to bandwidth by the following equation:

$$t = \frac{K}{B} \quad (42)$$

In this equation, t is the rise-time (10% to 90% amplitude) in seconds, K is a constant between 0.35 and 0.45, and B is the bandwidth

of the amplifier measured on the basis of the response range between 100% and 70.7% gain, as previously described. $K = 0.35$ is used when the overshoot is under 5% of the total amplitude of the input pulse and 0.45 when the overshoot is greater than this. The equation becomes invalid for amplifiers in which the overshoot exceeds 6 or 7%. Especially noteworthy is the fact that rise time is inversely proportional to bandwidth; this means that more perfect reproduction of the leading edge of a pulse may be expected as the bandwidth of the amplifier is improved.

The presence of excessive overshoot is an indication that the amplification falls off seriously at high frequencies and that the time-delay characteristics may be very nonlinear. Sag has already been attributed to poor low-frequency response due to nonlinear time-delay characteristics. Thus, the use of transient response analysis shows in a direct manner how the amplifier performs over its entire range and is very useful in designing television, radar, and other amplifiers in which abrupt changes of input amplitude are common.

A very useful relationship is given in equation (43).

$$\frac{\text{Gain}}{\text{Rise-time}} = \frac{g_m \text{ (mhos)}}{2.2C_t \text{ (Farads)}} \quad (43)$$

in which C_t is the total shunt capacitance consisting of the output capacitance of the first tube, the input capacitance of the second tube, and wiring capacitance. The gain to rise-time ratio is often called the *figure of merit* of the amplifier tube since the right hand member of equation (43) consists essentially of tube factors. For example, consider the gain to rise-time ratio of a 6AK5, a miniature pentode that enjoys wide usage in television and pulse amplifiers. For typical operation, its transconductance may be taken as 5000 micromhos and C_t as 12 $\mu\mu\text{f}$. For two R-C coupled 6AK5 stages, the gain rise-time ratio is:

$$\frac{G}{t} = \frac{5000 \times 10^{-6}}{26.4 \times 10^{-12}} \text{ amplification per } \textit{second} \text{ of rise time}$$

or $G/t = 190$ amplification per *microsecond* of rise time.

Depending on the amplifier bandwidth, two type 6AK5's can provide various gains and rise times. If, for example, the rise is 1/10 of a microsecond, the gain would be 20; or, if a short rise time of 1/100 of a microsecond is obtained, the gain would be only 2.

In general, the qualitative characteristics of transient analysis may be summarized as follows: (1) the high-frequency figures of merit of the amplifier determine the rise time and the overshoot amplitude and duration and (2) the low-frequency figures of merit of the amplifier determine the extent of sag and undershoot. A more extensive treatment will be given pulse-transmission characteristics in the chapters on the design of compensation systems for practical video amplifiers.

17. Tubes for Video Amplifiers

From equation (43) it is seen that the proper selection of a tube for use in a video amplifier system is extremely important. For the greatest gain and the shortest rise time—the equivalent of greatest gain with the greatest possible bandwidth—the g_m of the tube should be as high as possible while the net input and output capacitances between stages should be as small as possible. It is important to take input capacitance increase due to Miller Effect into consideration in such calculations, as well as stray wiring capacitance. If wiring capacitance is considered constant for various tubes, the gain/rise-time ratios can easily be calculated for comparison purposes. A few examples are given in Table 1 below.

TABLE 1
GAIN/RISE-TIME FOR VARIOUS PENTODES

<i>Tube</i>	g_m (μmhos)	C_t (μmf)	<i>Approximate</i> <i>Gain/Rise-time</i>	
	*	**	(<i>gain/microseconds</i>)	* g_m is taken for the approximate oper- ating conditions of the tubes
6AK5	5000	12	200	
6AC7	9000	21	190	
6AG5	4500	12.3	170	** Wiring capacitance assumed equal to
6AU6	4500	15.3	130	
6SJ7	1600	18	40	5 μmf .

The 6AK5, 6AC7, 6AG5, and 6AU6 are specifically designed for high-frequency wide band applications. Compare their gain/rise-

time ratios with that of the general purpose pentode, for example, the 6SJ7 type.

18. Review Questions

1. List and discuss the requirements that must be met by an amplifier if it is to serve adequately in wideband video applications.
2. Which components of an amplifier introduce phase shift which is measurable as variable time-delay for different frequency components?
3. Explain why the ideal phase-frequency characteristic of an amplifier appears as a straight line (Fig. 8). Why is such a curve called "ideal" in terms of amplifier performance?
4. In which equivalent circuit, high or low frequency, is the coupling capacitor considered a short circuit? For which equivalent circuit is the stray-wiring capacitance considered negligible? Explain.
5. Explain in detail the significance of f_c , the frequency defined by the reciprocal of $2\pi R_c C_c$.
6. It can be shown that the reduction of both R_c and C_c results in superior amplifier behavior at high frequencies. Which of these factors do designers attempt to decrease without limit? Which one cannot be decreased without limit? Give the reasons for both statements.
7. Improvement of low-frequency response can be effected by increasing the values of R_g and C_c in Fig. 12. What are the conditions that establish limits to which either or both of these factors can be raised?
8. Explain what is meant by the statement: When the frequency is low, a relatively small phase shift measured in degrees is the same as a relatively large time delay measured in seconds.
9. What is the effect of inadequate bypassing of cathode and screen impedances on low-frequency response. Explain each of these effects.
10. What are the important evaluation factors involved in transient-response analysis in which the frequency and phase characteristic is determined from a study of the reproduction of rectangular pulses?
11. Explain what is meant by rise-time and overshoot; by sag and oscillation as applied to the reproduction of a square wave.

Chapter 3

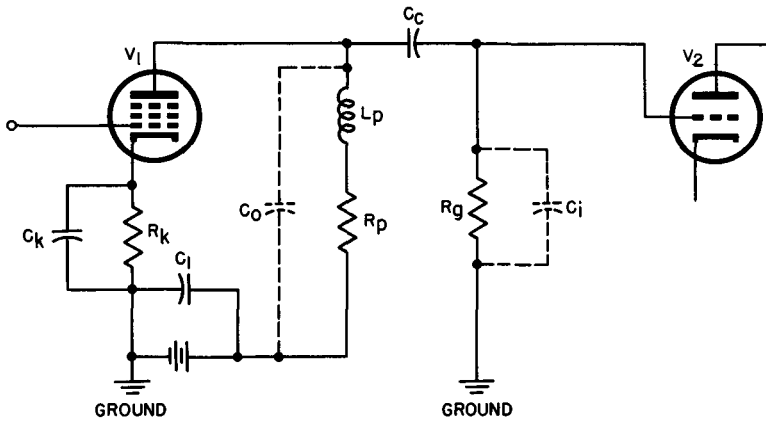
HIGH-FREQUENCY COMPENSATION METHODS

19. Analytical Approach to High-Frequency Compensation

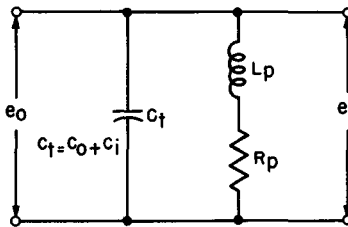
The analysis of the uncompensated video amplifier described in Chapter 2 indicates that the high-frequency response of any amplifier is degraded by the shunting action of C_t which is made up of input and output tube capacitances and stray-wiring capacitances. It has also been shown that C_t cannot be reduced below very definite values and that at these values very wideband performance is not possible.

The principal reason for this behavior of C_t at high frequencies is that it offers reactance (therefore, is frequency sensitive) offering less reactance (and greater reduction in gain) as the frequency rises. This at once suggests the addition of some component, which combines with C_t in such a way as to make the combination non-reactive. The sensitiveness of the load impedance to frequency changes is significantly reduced. For example, can an inductance be so selected as to produce the desired effect without establishing a condition of resonance in which a single high-frequency voltage would be peaked at the expense of others? In short, can an inductance value be found such that C_t is made nonreactive over a wide range of frequencies? The analysis presented in the next section answers these questions.²

²This analytical approach was first proposed by G. D. Robinson in an article entitled "Theoretical Notes of Certain Features of Television Receiving Circuits" in *Proc. IRE*, June 1933.



(A) AMPLIFIER CIRCUIT



(B) EQUIVALENT CIRCUIT

Fig. 18. (A) Amplifier circuit containing a shunt-peaking coil. This arrangement is given this name because the inductance is effectively in shunt with C_c . (B) High-frequency equivalent circuit of shunt peaking system.

20. Analysis of Shunt Peaking

An inductance L_p may be connected in series with the plate resistance R_p of a given amplifier. The equivalent circuit at high frequencies has the appearance given in Fig. 18. The total shunt capacitance C_t is the sum of the output capacitance of the driver amplifier (the amplifier producing e_o) and input capacitance (including Miller Effect capacitance) C_i of the stage to which compensation is being applied. The total load impedance then becomes the combination shown in Fig. 18B; C_c may be ignored because at high frequencies it is the equivalent of a short circuit; R_g need not be considered because its value is very high compared to X_{Lp} and R_p .

If an attempt is made to set up an equation for the total impedance of this combination of C_t , L_p , and R_p , the solution is found

to be quite complex and difficult to handle. On the other hand, if *susceptance* is operated upon rather than impedance, a rather simple solution can be obtained. The effective susceptance of any portion of a network is the factor by which the rms value of E_z , the voltage drop across the total impedance, must be multiplied in order to yield the value of the reactive component of the current through the impedance. If the current through the impedance is symbolized by I , reactance by X , and susceptance by B , then by definition:

$$I \sin \phi = B E_z$$

where ϕ is the phase angle between the impedance and the resistance of the branch.

Since $\sin \phi$ is defined by X/Z , then we have:

$$\frac{IX}{Z} = BIZ$$

Solving for B , we obtain:

$$B = \frac{X}{Z^2}$$

If the circuit in question is composed of a resistance and inductive reactance, then:

$$B = \frac{2\pi fL}{R^2 + (2\pi fL)^2}$$

Just as an impedance becomes resistive if the inductive and capacitive reactances are equal, the same impedance becomes resistive if the inductive susceptance equals the capacitive susceptance. In the case of the equivalent circuit in Fig. 18B, the total impedance becomes resistive when $B_{Ct} = B_{RL}$. Writing the equation for B_{Ct} we have:

$$B_{Ct} = 2\pi fC_t \quad (44)$$

The susceptance of the resistive-inductive branch containing R_p and L_p was shown to equal:

$$B_{RL} = \frac{2\pi fL_p}{R_p^2 + (2\pi fL_p)^2} \quad (45)$$

If the inductance of the added coil is assumed to be small compared to the load resistance R_p , then the factor $(2\pi fL_p)^2$ is too small to have significant effect upon the solution of equation (45) and may be dropped.

$$B_{RL} = \frac{2\pi fL_p}{R_p^2} \quad (46)$$

To have the impedance purely resistive, B_{C_t} must equal B_{R_L} . Equating equations (44) and (46), we obtain:

$$2\pi f C_t = \frac{2\pi f L_p}{R_p^2} \quad (47)$$

Dividing through by $2\pi f$, and solving for L_p , we obtain:

$$L_p = R_p^2 C_t \quad (48)$$

The significance of equation (48) may be stated as follows: Based upon the initial assumption that L_p is of small value, small enough so that the factor $(2\pi f L_p)^2$ can be ignored in equation (45) even when the frequency is reasonably high, equation (48) states that *an inductance L_p can be found which can make the equivalent circuit*

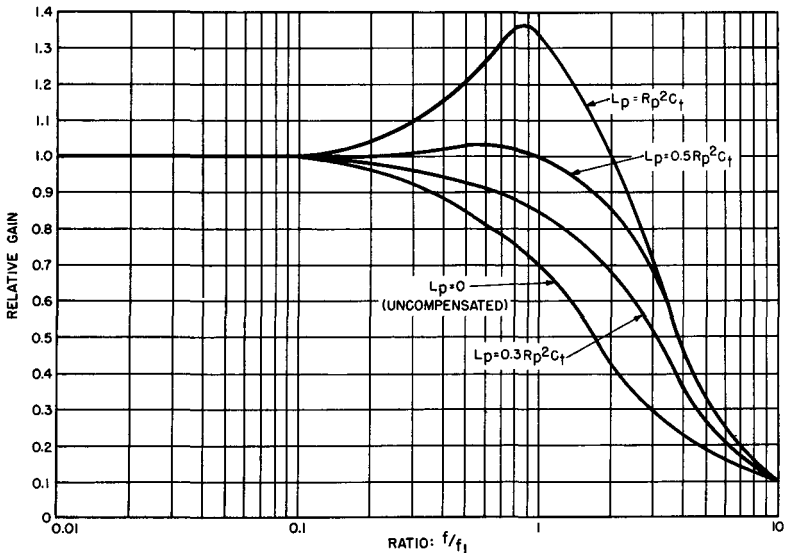


Fig. 19. Relative gain of a shunt-peaked amplifier for various values of L_p plotted against ratio f/f_1 .

impedance purely resistive regardless of the frequency. We note at this point that when the frequency rises sufficiently so that $(2\pi f L_p)^2$ can no longer be neglected, the impedance again becomes subject to variations with changing frequency. It is interesting to observe that the shunt peaking coil L_p *does not* form a simple resonant circuit

with C_t when both L_p and f are small: by properly selecting the inductance value of L_p , it is therefore possible to appreciably flatten out the video amplifier response curve appreciably at the high frequencies. When L_p is made too small, the high frequency amplitude rise is insufficient; if L_p is made too large, a resonance peak develops as shown in Fig. 19. As a matter of fact, the latter condition is the one in which the factor $(2\pi fL_p)^2$ has grown too much to be dropped out in equation (45).

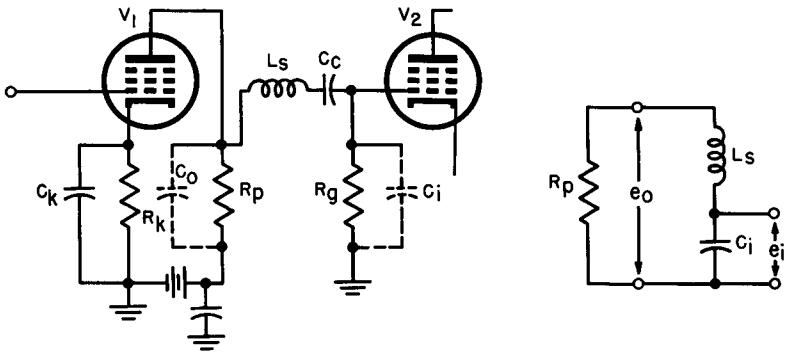
The various values of L_p shown in the curves of Fig. 19 are measured in terms of their fractional relation to $R_p^2C_t$. For example, if L_p is made equal to $R_p^2C_t$, a very large peak develops in the vicinity of $f/f_1 = 1$. It will be recalled that this is the point where $f = f_1$, or the frequency at which the amplifier gain *without compensation* normally drops to 70.7% of its mid-frequency range value. As L_p is made smaller, the peak subsides. When $L_p = 0.5R_p^2C_t$, the curve is essentially flat right up to $f/f_1 = 1$. This is a significant improvement over the performance of the uncompensated amplifier.

The use of a shunt peaking coil also improves the time-delay characteristic of a single stage video amplifier. When $L_p = 0.5R_p^2C_t$, an excellent compromise between linearity of gain and phase response is obtained. This value is most often used in television and radar practice.

21. Series Peaking

Although shunt peaking is adequately effective for many video amplifier applications, somewhat superior high-frequency compensation may be obtained by changing the position in the circuit of the peaking inductance and its value as shown in Fig. 20. By placing the peaking coil L_s between C_1 and C_p , these capacitors become isolated from each other so that C_1 which is normally considerably larger than C_p no longer affects the output circuit of the first amplifier. Therefore R_p may be made of substantially larger value than in the circuit where C_1 and C_p shunt each other; therefore, an improvement in gain results without degradation of high-frequency response.

With C_c considered a short circuit at high frequencies, and R_g large in value, the series peaking arrangement results in the equivalent circuit shown in Fig. 20B. The uncompensated output voltage developed across R_p and C_p of the preceding stage is impressed across



(A) AMPLIFIER AND SERIES PEAKING CIRCUIT

(B) EQUIVALENT CIRCUIT

Fig. 20. (A) Series peaking coil in series with the coupling capacitor. This arrangement is sometimes called filter coupling. (B) Equivalent circuit.

the series combination of L_s and C_i . R_g has negligible effect since it is shunted by the comparatively low reactance of C_i , hence it does not appear in the equivalent circuit. The input voltage to the following tube e_i is, therefore, the voltage that appears across C_i .

The resonant frequency of L_s and C_i is selected somewhat above the high-frequency limit that has been chosen for the particular amplifier. As the operating frequency approaches the high-frequency end of the range, the voltage across C_i increases and acts to neutralize the loss of gain at high frequencies caused by the lack of com-

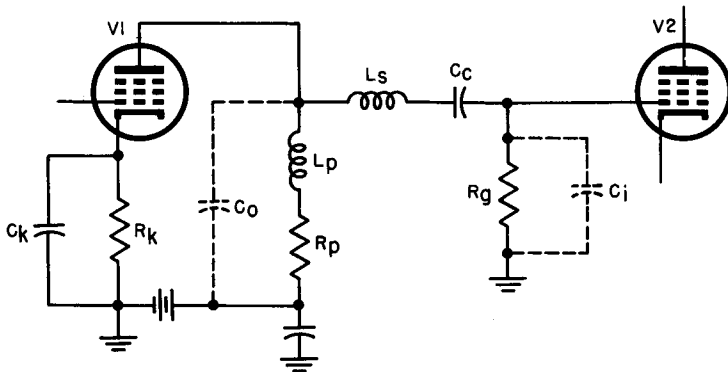


Fig. 21. Combination peaking using shunt and series inductances. This circuit is capable of better gain than either series or shunt peaking alone, but does not provide the same uniformity of response.

pensation in the preceding output circuit. Due to the greater overall gain possible when R_p is made large, the series peaking circuit provides about 1.5 times the amplification obtained with shunt peaking, if all other factors remain constant. As an additional advantage, the phase-characteristic for series peaking is definitely superior to that of shunt peaking. The derivation of the actual figures from the phase-delay function is a rather complicated and tedious procedure and is not attempted here. The reader will be interested in the fact, however, that the time delay of a series-peaked circuit can be made constant up to f_1 with a variation of approximately $0.11/f_1$ microseconds. This is only slightly more than 50% of the variation in time delay obtained with shunt peaking.

22. Combination Peaking

The working advantages of shunt and series peaking may be made additive by combining them in a single circuit as shown in Fig. 21. Combination peaking can be made to yield 80% to 85% greater amplification than simple shunt peaking. When correctly adjusted, the phase characteristic of combination-peaking does not differ materially from that of series peaking. Table 2 below indicates the essential performance characteristics of all three types of peaking compared to that of an uncompensated amplifier.

TABLE 2

<i>Compensation Type</i>	<i>Relative Gain at f_1</i>	<i>Time-Delay Variation</i>
Shunt only	1.0	$0.021/f_1$
Series only	1.5	$0.011/f_1$
Combination	1.8	$0.014/f_1$
Uncompensated	0.707	$0.035/f_1$

A more detailed discussion of component values in all three forms of peaking circuits will be given in Chapter 5.

23. Compensation Problems in Multistage Amplifier Circuits

Television studio and transmitter design, as well as radar, almost always call for more than one stage of video amplification. While

modern television receivers employ only one video amplifier as a rule, as many as 25 video amplifiers follow the camera in a television studio and transmitting room. Compensation for good high-frequency response cannot be handled merely as a series of independent stage modifications in such cases since the effects of compensation — particularly certain undesirable ones — are cumulative and require separate study.

The amplitude-frequency discrimination resulting from several stages of video amplification in cascade is *the product of the individual stage frequency discriminations*. Consider two amplifiers having amplifications of A and A' respectively. Assume that the first one discriminates between f and f_1 by a factor k , and that the discrimination of the second is equivalent to k' . If f and f_1 are applied to the input of the first amplifier, their relative output voltages are A and Ak . These voltages are then fed to the second amplifier so that the relative output in this case is AA' and $AA'kk'$. The total discrimination of the two stages in cascade is then obtained from the ratio of the voltage which, in this case, is kk' or the product of the individual discrimination factors.

The phase-shift characteristic, on the other hand, is *additive* in cascaded stages since the time delay resulting from inconstant phase shift in the first stage sets up the reference for measuring the time delay in the second stage, the second establishes the reference for the third, and so on.

Figure 22 illustrates the cumulative effect of peaking cascaded stages with the same peaking coil values, while maintaining other identical circuit conditions. Assuming that a small peak occurs in the vicinity of $f/f_1 = 1$ for a single stage, this condition is aggravated by the cascade multiplication and introduces a serious nonuniformity in the amplitude-frequency characteristic of the multistage video amplifier. A similar effect is observed in video amplifiers which do not produce a peak at the high-frequency end of the response range but which fall off in amplification at this point. In this instance, although the bandwidth of the individual stage may be great enough for the desired range, the overall bandwidth of several cascaded stages is actually *smaller* than required because amplitude drop-off is also cumulative in effect. For instance, if the bandwidth of 10 cascaded stages is to be constant to within 3 db between a reasonably low-frequency such as 30 cycles and a high-frequency

limit of 4 mc, then each individual stage response curve must be equally flat between 30 cycles and *better than 7 mc!* In general, television studio practice dictates that peaking circuits be designed for maximum flatness rather than high gain. In other words, the bandwidth of each amplifier is specified at no less than 7 or 8 mc, regard-

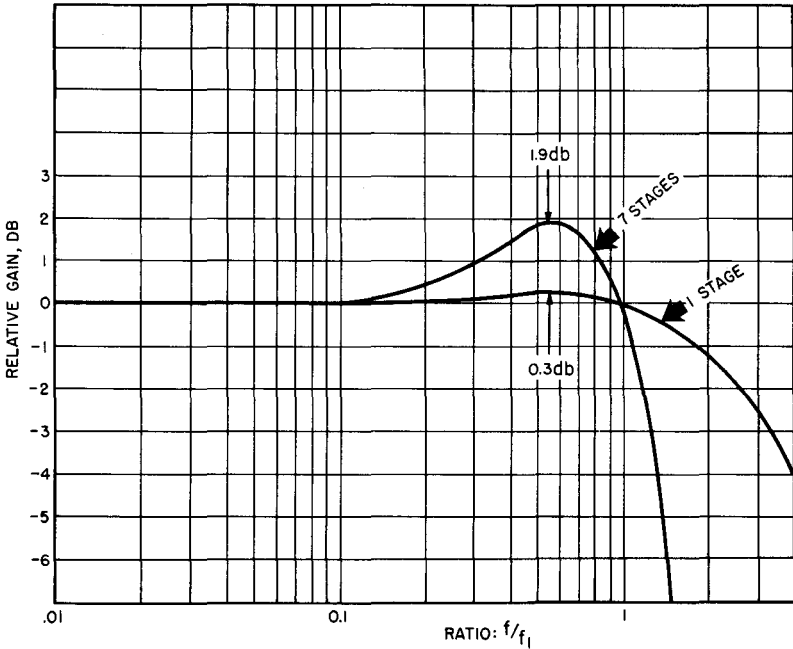


Fig. 22. The effect of identical peaking used in seven cascaded amplifier stages. Note the improper emphasis on the resonance peak near f/f_1 .

less of the loss of gain so incurred; then additional stages, also designed for maximum flatness, are inserted to make up for the loss of amplification attributable to extreme wideband design. Under these initial design considerations, it is then possible to construct a cascaded amplifier of many stages which will exhibit uniform amplitude and constant phase shift characteristics from 30 cps to 4 megacycles.

Note: We are concerned only with high frequency compensation in this chapter. Details relating to compensation down to low frequencies of 30 cycles or less are covered in the next chapter.

24. Review Questions

1. What would happen to the high-frequency response of a video amplifier if the output impedance could be made purely "resistive" in nature (nonreactive)? Explain.
2. With the aid of a diagram, explain why shunt peaking is effective in increasing the high-frequency response of a video amplifier?
3. Define susceptance. Prove that susceptance becomes the reciprocal of reactance in a nonresistive circuit.
4. What are the conditions under which equation (48) shows that an inductance L_p can be found which can make the equivalent circuit impedance purely resistive regardless of frequency? Explain fully.
5. Discuss the reason for the statement that the gain of a series-peaked compensated amplifier can be made appreciably larger than the gain of a shunt-peaked one.
6. How does the phase characteristic of a series-peaked compensated amplifier compare with that of a shunt-peaked type?
7. If you were about to select a peaking circuit, which would you choose to obtain the greatest gain? If you were specifically interested in the most constant time-delay, which one would you choose?
8. If each of two identical amplifiers display a degradation of 2 db at f_1 , what degradation would you expect when these amplifiers are connected in cascade?
9. Explain why the total time delay for several amplifiers in cascade is the sum, rather than the product, of the individual time delays.
10. Describe the procedure used in television transmitter design to insure adequate high-frequency compensation of many video amplifiers in cascade.

Chapter 4

LOW-FREQUENCY COMPENSATION METHODS

25. General Information

It was shown in Chapter 2 that the low-frequency response of an amplifier is improved by using reasonably large values of grid resistor R_g and coupling capacitor C_c , and by adequately bypassing the cathode and screen resistors. The values of R_g and C_c , however, cannot be increased without limit because of the increased shunting effect of these components and the consequent degradation of high-frequency performance; also, relaxation oscillation is an ever-present danger when the R-C time constant of the coupling circuit is made too large.

Aside from the proper selection of the components mentioned above, further improvement in the low-frequency range of a video amplifier may be effected by adding a low-frequency compensation circuit in series with the plate resistor R_p . When correctly designed, the compensation circuit consisting of C_1 and R_1 in Fig. 23 can extend the low-frequency end of the video amplifier response curve to a substantial degree.

26. Analysis of Low-Frequency Compensation Circuit

The analysis of the coupling circuit and the compensating filter is most easily handled by considering the tube to be a constant-cur-

rent generator (rather than constant-voltage type). That is:

$$i_p = e_i g_m \quad (49)$$

in which i_p is the instantaneous plate current, e_i is the instantaneous input voltage, and g_m is the transconductance of the tube in mhos. This relationship is particularly applicable to pentodes since it assumes that both R_p and R_1 are small compared to the dynamic plate resistance r_p of the tube. Let us recall that R_p is seldom larger than 10% the value of r_p in practical pentode circuits; R_1 is even smaller than that. These assumed conditions are correctly met by a pentode amplifier. Since the plate current of a pentode is virtually independent of the plate voltage, small variations of voltage drop due to signal currents across R_p and R_1 have little effect upon plate current; hence, the phrase "constant current."

The constant signal current in the output circuit of the tube sees a branched output impedance Z_p : one branch consists of R_p and R_1 (shunted by C_1) in series, while the other comprises the series combination of C_c and R_g . In practice, R_g is very much larger than $R_p + R_1$, hence most of the constant current flows through the R_p - R_1 - C_1 branch. Symbolizing the amplification developed across this branch by A' , we can write:

$$A' = g_m Z_p \quad (50)$$

in which Z_p is the total impedance of the plate circuit branch described above. The value of Z_p may be stated in the usual manner as:

$$Z_p = R_p + \frac{R_1}{j2\pi f C_1 \left(R_1 + \frac{1}{j2\pi f C_1} \right)} \quad (51)$$

Equation (51) may be arranged for simplification thus:

$$Z_p = R_p + \frac{R_1}{1 + j2\pi f R_1 C_1} \quad (52)$$

Substituting this value for Z_p in equation (50):

$$A' = g_m \left(R_p + \frac{R_1}{1 + j2\pi f R_1 C_1} \right) \quad (53)$$

In this case, as in any R-C coupled amplifier, C_c and R_g form a signal voltage divider in which only a part of the gain developed across the

plate impedance is transferred to the grid circuit of the next tube. The fraction transferred is given in general terms as:

$$\text{Fraction} = \frac{R_g}{R_g + X_{Cc}} \quad (54)$$

in which X_{Cc} is the capacitive reactance of the coupling capacitor.

Thus, the net gain delivered to the grid of the second tube is equal to A' multiplied by the fraction in equation (54). The complex equation obtained by this process may be finally reduced to the form:

$$\text{Net Gain} = \frac{2\pi f g_m R_p \left(2\pi f - \frac{j}{C_1 R_2} \right)}{\left(2\pi f - \frac{j}{R_1 C_1} \right) \left(2\pi f - \frac{j}{R_g C_c} \right)} \quad (55)$$

In this expression, it will be noted that a new term appears: R_2 . For simplicity, we set $R_2 = R_p R_1 / (R_p + R_1)$. This ratio makes its appearance in the intermediate multiplication of A' by transferred fraction previously described. By substituting a simple symbol such as R_2 for this ratio, the final equation as given in (55) can be more easily interpreted.

A striking feature of this equation is that *both the numerator and denominator contain terms in which frequency appears as a multiplier*, suggesting therefore that the proper selection of components might result in partial, if not total, cancellation of the frequency component. Suppose, for example, that $R_2 C_1$ is made equal to $R_g C_c$. (This is the equivalent of saying that the time constants of these combinations are arranged for equality.) For this condition the entire parenthetical term in the numerator is cancelled by the right-hand parenthetical term in the denominator so that the net gain becomes:

$$\text{Net Gain} = \frac{2\pi f g_m R_p}{2\pi f - \frac{j}{R_1 C_1}} \quad (56)$$

The importance of the compensating circuit becomes immediately apparent from this equation. If the time constant $R_1 C_1$ is made increasingly larger—keeping $R_2 C_1 = R_g C_c$, as demanded by the assumption that made the equation possible—the right-hand term in

the denominator approaches zero. Under these circumstances, the $2\pi f$ terms in the numerator and denominator cancel leaving the net gain equal to $g_m R_p$. In this form, the amplifier is not frequency-sensitive at all.

Therefore, from a theoretical point of view, the low-frequency response of the video amplifier may be indefinitely extended by increasing the time constant of the compensation network without

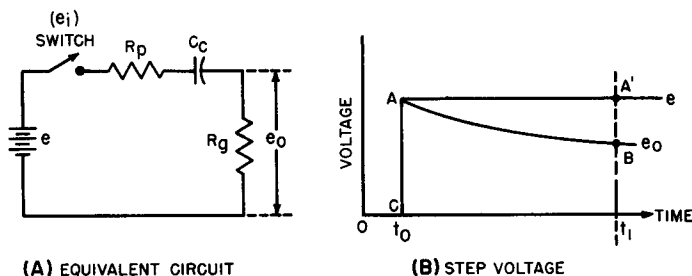


Fig. 23. A step voltage acting between the grid and cathode of an amplifier tube compared to an equivalent circuit in which a switch is closed.

limit. This ideal result cannot be achieved in practice, of course, since an excessive value of R_2 reduces the plate voltage of the tube to the point where it has no gain at all. The analysis shows, however, that compensation brings about substantial improvements in performance.

From another point of view, the identity $R_2 C_1 = R_g C_c$ demands that, if $R_2 C_1$ is made large, so must $R_g C_c$ be raised in value. This requirement is an additional limitation on the magnitude of the time-constant product of the compensating network since $R_g C_c$ cannot be increased greatly before motorboating begins. The actual values of the various components used in practical video amplifiers will be derived in the discussions contained in Chapter 5.

27. Transient Behavior at Low Frequencies

Some of the characteristics of video amplifiers as related to pulse-transmission response were defined in Section 16. It was noted in this section that *sag* in the reproduction of a square wave was due to improper low-frequency performance of the amplifier. (See Fig. 17.)

The application of a step voltage (the leading edge of this pulse has nearly instantaneous rise time and comparatively large pulse width) to the grid circuit of an amplifier tube can be closely compared with the act of closing a fast-acting switch in a simple R-C circuit as shown in Fig. 24.

Assuming that C_c is initially discharged, the charging current that flows immediately after the switch is closed causes the largest possible voltage drop e_o across the grid resistor R_g for the particular values of the circuit. As C_c charges, the current diminishes thereby causing e_o to diminish in proportion. In Fig. 24B, waveform CAA' is the applied step voltage while CAB is the waveform of the voltage e_o from time t_o to t_1 . Sag is defined as the ratio of A'B/AC and may be shown by basic d-c theory to be:

$$\text{Sag} = 1 - \epsilon^{-D} \quad (57)$$

where:

$$D = \frac{t}{C_c (R_p + R_g)}$$

in which ϵ (epsilon) is the base of the natural logarithms. If R_g is much greater than R_p , as is normally the case in practice, then equation (57) may be simplified as follows:

$$\text{Sag} = 1 - \epsilon^{-E} \quad (58)$$

where:

$$E = \frac{t}{R_g C_c}$$

Thus, the sag shown in Fig. 24B is due to the presence of the coupling capacitor C_c . It is also evident from equation (58) that sag can be made small only by increasing the resistance of R_g and the capacitance of C_c . This is the same conclusion reached by the preceding study of amplitude and phase characteristics at low frequencies.

Using a similar approach, it can be demonstrated that sag may also be caused by inadequate bypassing of screen and cathode impedances. A compensation network such as that formed by R_1 and C_1 in Fig. 23 reduces the net sag by causing a *lift* (or negative sag) in the flat top of the reproduced pulse. In theory, it is possible to proportion R_1 and C_1 so that they fully compensate for any one of the sags introduced by the coupling capacitor, the screen impedance, or the cathode impedance. Compensation of practical amplifiers rarely approaches such perfection, however, for the reasons previously

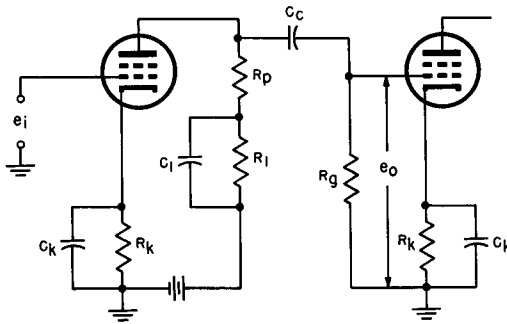


Fig. 24. Low-frequency compensation circuit consisting of capacitor C_1 and resistor R_1 in series with R_p .

given: any attempt to make R_1 large enough to approach the ideal results in excessive plate voltage loss due to the voltage drop in this resistor; the component values are very critical so that compensation action does not remain constant as the components age; finally, when several sources of sag are present at the same time, one R-C network cannot compensate for all of them simultaneously.

Thus, the performance of a practical compensating system never produces an output waveform identical to the input waveform when the input is a step voltage. System designs are based upon close approximations only. Normally, a compensating network having the desired characteristics is first calculated and connected into the associated amplifier. Its behavior with an applied square wave is then observed on an oscilloscope; excessive sag still present indicates that the compensation is insufficient while a rising flat top points towards overcompensation. The values of C_1 and R_1 are then adjusted until the best possible compensation is obtained.

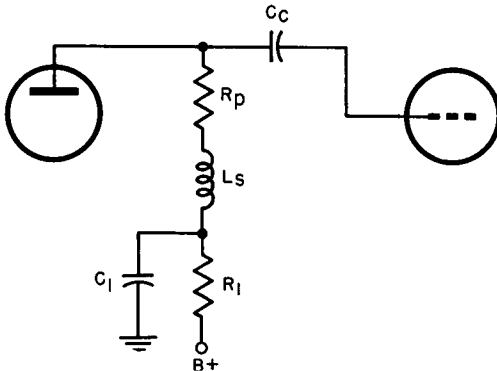


Fig. 25. Low-frequency and high-frequency compensating circuit components in a single video amplifier stage.

The final video amplifier design must include both high- and low-frequency compensating networks so placed as to avoid mutual interference. It is customary to connect a high-frequency shunt peaking coil between the plate resistor and the low-frequency compensating network as shown in Fig. 25. At very high frequencies, C_1 is a short-circuit to ground, effectively removing itself and R_1 as significant factors in the circuit. As the frequency diminishes sufficiently to make C_1 effective, the inductive reactance of L_s becomes so small as to render the peaking coil inoperative, reducing the plate load to R_p . Thus, neither of the compensating systems causes interference with the action of the other.

28. Review Questions

1. What are the chief causes of attenuation of the low frequencies in a video amplifier?
2. Explain why a pentode can be considered as the equivalent of a constant current generator.
3. Why cannot the low-frequency response of an amplifier be extended indefinitely by increasing the time constant of the compensation network without limit?
4. What factor or factors make it impossible to increase the coupling capacitor value without limit? the value of the grid resistor?
5. What is meant by a step voltage?
6. Define *sag*. What is the value of measuring the sag in the reproduced voltage of a step-voltage input signal?
7. Explain on the basis of R-C time-constant theory why sag appears in the reproduced pulse of an uncompensated amplifier when the input is a step-voltage.
8. How can sag be reduced? Why is perfect compensation for sag virtually impossible?
9. Draw a schematic diagram of a circuit containing high-frequency compensation in the form of a shunt peaking coil and a low-frequency R-C compensating network. Repeat for series peaking and combination peaking.
10. Explain why high- and low-frequency compensation circuits act independently of each other.

Chapter 5

VIDEO AMPLIFIER DESIGN PROCEDURE

29. Outline of Design Procedure

The fundamental limitations of an uncompensated amplifier are:

(1) The nonuniform gain over the portion of the spectrum to be amplified.

(2) The varying time delay for different portions of the amplifier's frequency range. It has been pointed out that an inconstant phase shift characteristic causes more degradation of low-frequency response than nonuniform gain, but that the latter defect is appreciably more significant at the high frequencies. Although it is impossible to realize both uniform gain and constant time delay over the video range, good approximations resulting in satisfactory amplifier performance can be achieved by proper design.

The basic approach to video amplifier design procedure can be summarized as follows:

(a) The bandwidth of the amplifier is decided upon so that f_1 is known for both the high- and low-frequency ends of the uniform response range of the amplifier. The amplifier tube or tubes are then selected. Normally this selection is determined by gain requirements, space and size considerations, availability, etc.

(b) The uncompensated stage is then wired in *prototype form*. That is, a duplicate of the amplifier to be installed in the final equipment is constructed so that it can serve as an electrical *and* mechanical prototype. The purpose of this step is to insure that

the shunt capacitances of the experimental model exist in the final equipment.

(c) The total shunt capacitance C_t is then calculated and/or measured by methods to be described.

(d) The high-frequency compensation system, using either shunt, series, or combination peaking, is then calculated and/or determined by a specific experimental method which is discussed later.

(e) Screen circuit and power supply impedances are then reduced as much as practicable.

(f) The low-frequency compensating network is then calculated to inhibit the detrimental effects of the cathode impedance and coupling-circuit action.

(g) The performance of the compensating filter R_1C_1 is checked by square-wave analysis. Finally, the behavior of the amplifier with respect to uniformity of gain and constancy of time-delay is tested using both square-wave and variable-frequency input methods.

Although the design procedure described in the remainder of this chapter has been confined to numerical examples dealing with a specific objective, the approach is quite general and can be applied to any suitable tubes or circuits. The specific attack has been selected because it is felt that much additional clarification of both the problem and the solution is possible in this way.

Let us state the design problem and then proceed with the solution in accordance with the outline above.

A two-stage video amplifier consists of a 6AH6 miniature sharp-cutoff pentode and a 6V6 beam power video output tube. The gain of the first stage is to be uniform from 30 cycles to 4 mc. Electrode voltages may be selected to conform with tube ratings according to standard practice.

30. Gain at Middle Frequencies

The mid-frequency gain of a pentode such as the 6AH6 in which r_p is ten or more times greater than the plate load impedance Z_p is given in equation (12) as:

$$\text{Gain} = g_m Z_p = g_m R_p \quad (R_p \text{ is the resistive load}) \quad (12)^*$$

At the inception of design, it is informative to calculate the mid-frequency gain of the stage on the basis of some reasonably assumed

*Equations that appear earlier in the book are identified by the original number.

value of plate load impedance. The 6AH6 is designed to operate with 300 volts between its plate and cathode and, at this voltage, its plate resistance is given as 0.5 megohm (screen potential = 150 volts, cathode resistor = 160 ohms). Under these conditions, the g_m of a 6AH6 is 9000 micromhos. We have seen that f_1 rises as the plate load impedance is made smaller and, since the desired high frequency f_1 is quite high in this case (4.0 mc), it is best to start with a relatively low value for R_p . If the assumed value does not give the required gain, then R_p will have to be increased in size. It would be reasonable, then, to start with Z_p equal to, say, 5000 ohms. Thus:

$$\text{Mid-frequency Gain} = 9000 \times 10^{-6} \times 5 \times 10^3 = 45$$

Assuming that this gain is adequate, we will temporarily accept 5000 ohms as the design value for R_p . In accordance with the outline, a prototype of the first amplifier stage is then wired, using values for screen dropping and bias resistors such that the rated voltages are obtained. The cathode resistor is given in the tube manual as 160 ohms for a transconductance of 9000 micromhos; similarly, the screen voltage is stated as 150 volts and the screen current as 2.5 ma. Thus, the screen dropping resistor required is:

$$R_{g2} = \frac{150}{.0025} = 60,000 \text{ ohms.}$$

The finished prototype must also contain the input to the 6V6 stage as well as the components required for the 6AH6 amplifier portion. The circuit thus far is given in Fig. 26. Since we are interested at the moment in high-frequency performance, the values of other components such as R_{g1} , C_k , C_c , C_{g2} , and R_k for the 6V6 are selected merely on the basis of ordinary, narrow-band audio amplification as obtained from the R-C coupling data charts given in receiving tube manuals.

31. Establishing the Value of C_t

The value of C_t should be calculated and then measured. From equation (32):

$$C_t = C_w + C_i + C_{gp}(1 + \text{Gain}) \quad (32)$$

in which C_t is the shunt capacitance of the system, C_i is the input capacitance consisting of the sum of the output capacitance of the

6AH6 and the input capacitance of the 6V6, C_w is the stray-wiring capacitance, and C_{gp} is the grid-plate capacitance of the 6V6. The gain factor is that associated with the 6V6. Using manufacturers' values and an estimated C_w for a well constructed amplifier, we have:

$$C_i = 2\mu\mu\text{f} + 10\mu\mu\text{f} \quad (2\mu\mu\text{f}, 6\text{AH6 output C}, 10\mu\mu\text{f}, 6\text{V6 input C})$$

$$C_{gp} = 0.3\mu\mu\text{f} \quad 6\text{V6 gain} = 10 \text{ (approx)} \quad C_w = 5\mu\mu\text{f} \text{ (roughly)}$$

$$C_t = 5 + 12 + 0.3(1 + 10) = 20.3\mu\mu\text{f}$$

Thus, C_t , the total shunting capacitance acting across the plate load impedance in this case equals $20.3 \mu\mu\text{f}$.

Normally, the relatively simple calculation above is followed by a measurement of C_t . Such a measurement is used as the calculation's check since C_w is an estimated factor based upon previous experience and listed in the literature. There are two standard methods now in use for measuring C_t ; these will be described in the following paragraphs.

A very convenient approach to the measurement of C_t first described by Seeley and Kimball³ is based upon the relationship:

$$f_1 = 1/2\pi R_t C_t \quad (22)$$

in which R_t is the total plate load resistance as defined in equation (16), C_t is the total shunt capacitance, and f_1 is the high frequency at which the amplifier gain is down 29.3% from its mid-frequency gain value. To measure C_t , a signal generator is used to apply a mid-frequency of 10,000 cycles to the input of the amplifier; the amplifier output voltage is measured by a vtm or a cathode-ray oscilloscope (CRO). The frequency is then gradually raised while the input voltage is maintained at its initial value. When the output voltage has dropped to .707 times its initial value, the frequency of the signal generator must be f_1 and C_t can be found by solving equation (22) for C_t . R_t may be taken as the value of the plate load resistor (in our example, 5000 ohms) since the shunting effect of R_g — a very large resistor — is negligible. If the calculated value for $C_t = 20\mu\mu\text{f}$ is correct, then the gain drops by 3 db when $f_1 = 1.57$ mc as shown below:

$$f_1 = \frac{1}{6.28 \times 5 \times 10^3 \times 20.3 \times 10^{-12}} = 1.57 \text{ mc}$$

³Seeley and Kimball, "Analysis and Design of Video Amplifiers," *RCA Review* January 1939.

It is interesting to note at this point that this f_1 is quite low and that such an uncompensated amplifier can not be used for video purposes, even in the poorest of television receivers.

A second practical method often used for measuring C_t consists of replacing the plate load resistor with a known value of inductance and then finding the frequency at which this inductance resonates with C_t . An r-f signal generator is normally used; its unmodulated output is applied to the grid of the first video amplifier and the voltage developed across the plate circuit coil measured by a wide-range vtvm. The value of C_t is then determined from the equation:

$$C_t = \frac{1}{(2\pi f)^2 L_p}$$

in which L_p is the known value of plate circuit inductance. Suppose a peaking coil of $80 \mu\text{h}$ was used to replace the 5000-ohm plate resistor in our example and that it was found to resonate with C_t at a frequency of 4 mc. The measured value of C_t would then be:

$$\begin{aligned} C_t &= \frac{1}{(6.28 \times 4 \times 10^6)^2 \times 80 \times 10^{-6}} \\ &= \frac{1}{630 \times 10^{12} \times 80 \times 10^{-6}} \\ &= 19.9 \mu\mu\text{f} \end{aligned}$$

Since the value of C_t obtained is a function of the reciprocal of the square of the frequency, it is essential that the precision of the signal generator used for this measurement be good enough to provide a reliable answer to at least three significant figures. Note that the figure obtained by measurement differs from the computed value by approximately 4 parts out of 200 for an error of about 2%. An error of this magnitude is normally insignificant in ordinary video amplifier design.

32. Calculation of Shunt Peaking Network

In Section 30, a plate load impedance $R_p = 5000$ ohms was tentatively selected for the construction of the amplifier prototype. Now that C_t and f_1 are both known, the actual value of plate load

resistance required for compensation up to 4 mc can be determined from equation (22). Thus (assuming R_g large compared to R_p):

$$R_p = R_t = 1/2\pi f_1 C_t$$

$$R_p = \frac{1}{6.28 \times 4 \times 10^6 \times 20.3 \times 10^{-12}} = \frac{10^6}{510}$$

$$= 1960 \text{ ohms or } 2000 \text{ ohms approximately}$$

As previously pointed out, one unavoidable consequence of this reduction of R_p is the accompanying decrease in gain. Where the uncompensated gain of this amplifier with $R_p = 5000$ ohms was 45, the gain will now be down to $2/5 \times 45 = 18$.

The theoretical considerations leading up to the selection of L_p as the shunt peaking coil are presented in Section 20 and Fig. 19. The discussion shows that an excellent compromise between uniformity of gain and phase shift is obtained by setting $L_p = 0.5R_p^2C_t$. This equation may be rephrased in somewhat more convenient form by using the fact that R_p has been selected by making it equal to $1/2\pi f_1 C_t$. Thus, we may write:

$$L_p = \frac{0.5R_p}{2\pi f_1} \quad (59)$$

Substituting the known values from the above example, we obtain:

$$L_p = \frac{0.5 \times 2000}{6.28 \times 4 \times 10^6}$$

$$= 39 \text{ microhenries}$$

Thus, when a shunt peaking coil of $L_p = 39$ microhenries is used with a plate load resistor of 2000 ohms, the amplifier gain equals approximately 18 and remains reasonably uniform from the middle frequencies up to 4 mc.

33. Calculation of Series Peaking Network

Section 21 shows that an advantage of series peaking (Fig. 20) consists in the isolation of C_i and C_o , which makes it possible to realize greater gain by basing the selection of R_p on the value of C_o rather than that of C_t .

Analysis shows that the ratio of C_1 to C_0 plays an important role in the operation of a series peaking circuit. C_1 is almost always larger than C_0 ; by judiciously placing components in the wired circuit, it is possible to make the ratio $C_1/C_0 = 2$. This is usually the basis for design of practical series peaking circuits. Similarly, the resonant frequency of the series peaking coils L_s and C_t is generally chosen to be about 1.41 times f_1 . That is:

$$f_{LC} = 1.41f_1 \quad (60)$$

From the relation: $f_{LC} = \frac{1}{2\pi\sqrt{L_s C_t}}$ we can write:

$$f_1 = \frac{1}{1.41 \times 2\pi\sqrt{L_s C_t}} \quad (61)$$

since $1.41 = \sqrt{2}$, squaring both sides of equation (61) yields:

$$f_1^2 = \frac{1}{8\pi^2 L_s C_t} \quad (62)$$

and solving for L_s :

$$L_s = \frac{1}{8\pi^2 f_1^2 C_0} \quad (63)$$

Equation (63) can be used as a design equation for determining the size of the series peaking coil. Also, with C_1 and C_0 isolated from each other, the plate load resistor R_p can be made appreciably larger than in the shunt-peaking case. An increase of 1.5 times is a good compromise, so that:

$$R_p = \frac{1.5}{2\pi f_1 C_t} \quad (64)$$

which is the design equation for the plate load resistor.

Using series peaking for the amplifier having the specifications given at the beginning of this chapter, the following results are obtained:

Assuming that the physical wiring of the circuit is adjusted to make $C_1 = 2C_0$, then

$$\begin{aligned} R_p &= \frac{1.5}{6.28 \times 4.0 \times 10^6 \times 20.3 \times 10^{-12}} \\ &= 1.5 \times 1960 = 2940 \text{ or approximately } 3000 \text{ ohms} \end{aligned}$$

Thus, the gain of the series peaked stage is roughly 1.5 times greater than the shunt peaked one.

The value of the peaking coil L_s is found from equation (63). (Total C_1 of 6V6 including Miller Effect capacitance is $13.3 \mu\mu\text{f}$; thus $C_o = 6.6 \mu\mu\text{f}$.)

$$\begin{aligned} L_s &= \frac{1}{8 \times (3.14)^2 \times 16 \times 10^{12} \times 6.6 \times 10^{-12}} \\ &= \frac{1}{8 \times 9.86 \times 16 \times 6.6} \\ &= .000120 \text{ henry} \\ &= 120 \text{ microhenries} \end{aligned}$$

Equation (63) sometimes proves difficult to use because C_o is not always known or cannot be measured accurately. A second expression that is often more useful is given in equation (65):

$$L_s = 0.67 C_t R_p^2 \quad (65)$$

If the substitutions from the example are made in this equation, taking $R_p = 3000$ ohms, we obtain:

$$L_s = 0.67 \times 20.3 \times 10^{-12} \times 9 \times 10^6 = 122 \text{ microhenries}$$

It is important to note that these design equations are valid only for the condition in which $C_t/C_o = 2$. The reason for selecting this ratio is that it provides for the best compromise between uniform gain and constant phase characteristics. We must never lose sight of the fact that the phase characteristic is of primary importance in determining the performance of the video amplifier, particularly at low frequencies.

34. Calculation of Combination-Peaking Network

Reference is made to the combination peaking network shown in Fig. 21 throughout this discussion.

Much attention has been given combination peaking design because of its importance in television receivers. It is generally conceded that the greatest advantage to be expected from a well designed combination peaked circuit is increased gain, but that series-peaking is capable of more uniform amplification and more constant time-delay. Thus, the type of compensation used must be determined by the specific needs of the design. In television video

amplifiers—particularly in modern receivers in which a single video amplifier is used—gain is a very important factor. For this reason, combination peaking is quite common in this application.

The conditions that are met in the formulation of the equations are:*

$$\frac{C_1}{C_o} = 2 \quad (66)$$

$$R_p = \frac{1.8}{2\pi f_1 C_t} \quad (67)$$

$$L_p = 0.12 C_t R_p^2 \quad (68)$$

$$L_s = 0.52 C_t R_p^2 \quad (69)$$

The design procedure follows the same series of steps as previously outlined: assuming that f_1 is selected at 4.0 mc, C_o and C_1 are determined by measurement or calculation, C_1 is made twice as great as C_o , and R_p is determined from equation (67). With this value of R_p , the gain equals 1.8 times that of a simple shunt-peaking system. L_p and L_s are then calculated from equations (68) and (69) respectively.

Applying these equations to our example, the value of R_p is obtained merely by multiplying the value of R_p as obtained for the shunt-peaking case (i.e. 1960 ohms) by 1.8 or:

$$\begin{aligned} R_p &= 1.8 \times 1960 = 3528 \text{ ohms or} \\ &= 3500 \text{ ohms approximately} \end{aligned}$$

The value of the shunt-peaking coil L_p is obtained by substituting in equation (68) as follows:

$$\begin{aligned} L_p &= 0.12 \times 20.3 \times 10^{-12} \times 12.25 \times 10^6 \\ &= 30 \text{ microhenries} \end{aligned}$$

Similarly, L_s is calculated using equation (69) :

$$\begin{aligned} L_s &= 0.52 \times 20.3 \times 10^{-12} \times 12.25 \times 10^6 \\ &= 130 \text{ microhenries} \end{aligned}$$

Or, more simply, the value of L_s can be obtained directly from the value of L_p by means of the simple relationship:

$$L_s = \frac{13L_p}{3} \quad (70)$$

*Design factors originally given by E. W. Herold, August 1938.

since L_s is larger than L_p by a factor of 52/12 or 13/3.

The variation in time delay for combination peaking is very slightly larger than the variation for series peaking.

Table 3 which follows summarizes the design data developed in this chapter.⁴

TABLE 3

Performance and Design Data for High-Frequency Compensation Systems

	<i>Uncompensated</i>	<i>Shunt Peaking</i>	<i>Series Peaking</i>	<i>Combination Peaking</i>
R_p	$\frac{1}{2\pi f_1 C_t}$	$\frac{1}{2\pi f_1 C_t}$	$\frac{1.5}{2\pi f_1 C_t}$	$\frac{1.8}{2\pi f_1 C_t}$
L_p		$0.5C_t R_p^2$		$0.12C_t R_p^2$
L_s			$0.67C_t R_p^2$	$0.52C_t R_p^2$
Relative gain at f_1	0.707	1.0	1.5	1.8
Time delay variations up to f_1 (sec)	$0.035/f_1$	$0.023/f_1$	$0.011/f_1$	$0.015/f_1$

35. Selection of Screen Circuit Impedance for Optimum Low-Frequency Performance

The effect of bypass capacitors on wideband performance is explained in Section 15. This discussion shows that failure to bypass the screen dropping resistor, or inadequate bypassing, results in increasing degeneration as the frequency is reduced. Such degeneration causes a serious degradation of the low-frequency response of the amplifier.

The screen dropping resistor R_{g2} in Fig. 26 is calculated from specified values of screen voltage and screen current as given in the receiving tube manual (see Section 30); its value is 60,000 ohms. It can be shown experimentally that a video amplifier screen resistor, when bypassed by a capacitor having a reactance 1/10 that of the

⁴Time delay variations abstracted from *RCA Review*, January 1939.

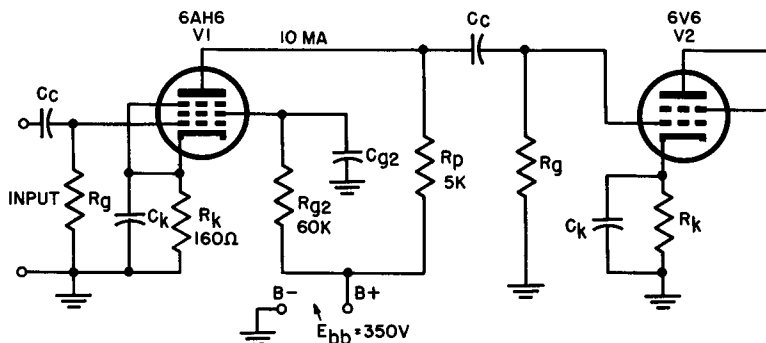


Fig. 26. Prototype wiring for a typical amplifier.

resistor (or less), performs well enough for the required bandwidth in television. On the other hand, rigorous transient analysis methods applied to very wideband amplification indicate that the screen bypass capacitor must have a reactance in the order of $1/200$ of the screen resistance. Using the first ratio ($X_c = 60,000/10 = 6000$ ohms), the figure obtained for the bypass capacitor is:

$$\begin{aligned}
 C &= \frac{1}{2\pi f X_c} \\
 &= \frac{1}{6.28 \times 30 \times 6000} \\
 &= .89 \mu\text{f}
 \end{aligned}$$

and working with the second ratio, the bypass capacitor would have to be:

$$C = \frac{1}{6.28 \times 30 \times 300} = 17.5 \mu\text{f}$$

From the practical constructional point of view, with the easy availability and low cost of electrolytic capacitors made possible by modern manufacturing methods, a $20\text{-}\mu\text{f}$ bypass capacitor can be used just as well as one of lower value. This then meets the specifications imposed by the transient analysis; it must be noted, however, that electrolytic capacitors are generally not as reliable over long periods of time as are paper or oil-impregnated capacitors. It is for this reason that one often finds circuit designers making use of methods of obtaining screen bias other than that of screen

dropping resistors. Many radar video amplifiers, where the frequency-response and phase-shift characteristics are necessarily more rigorously controlled, utilize fixed voltage sources for screen potential or voltage-divider arrangements where the resistance of the source is very small compared to the dynamic screen resistance of the tube, making a bypass capacitor totally unnecessary.

For practical television purposes, however, screen voltage is generally obtained by an R-C combination as in Fig. 26, the bypass capacitor being selected in the vicinity of 10 to 20 μ f.

36. Design of the Low-Frequency Compensation Network

Section 26 showed that low-frequency distortion caused by the capacitor C_c (Fig. 23) can be minimized by making the time constant R_1C_1 of the combination as large as possible. An important part of the design procedure states that the product R_1C_1 be equal to the product R_gC_c . This fact limits the value of the time constant R_1C_1 . Too large a time constant R_1C_1 can lead to instability and relaxation oscillations. Various experiments have given design figures for maximum R_gC_c products. Their results suggest that a time constant R_gC_c greater than 0.01 can lead to instability in a multistage system. However, values as high as 0.5 have been used successfully in two-stage circuits. To allow a margin of safety in our example, it seems justifiable to select a product such as $R_gC_c = R_1C_1 = 0.1$.

The tube manual provides the information that the maximum recommended resistance in the grid circuit of a 6V6 is 0.5 megohm. For an R_gC_c time constant = 0.1, the coupling capacitor would be:

$$C_c = \frac{0.1}{0.5} = 0.2 \mu\text{f}$$

when $R_g = 0.5$ megohm.

In conformity with the design equations, the product R_1C_1 equals 0.1. Since R_1 is connected in series with the plate circuit of the 6AH6, its value cannot be made too large, otherwise the voltage drop across R_1 becomes excessive. The value $R_1 = 25,000$ ohms is considered a representative value. Assuming this value for R_1 , C_1 is determined:

$$C_1 = \frac{0.1}{0.025} = 4 \mu\text{f}$$

37. Design of the Cathode Impedance for Optimum Low-Frequency Performance

With the screen impedance selected for optimum low-frequency performance, and the design values for R_1C_1 calculated, all that remains is to compute the cathode impedance components that will support the compensation design thus far completed. The cathode resistance required by the tube is 160 ohms (see Section 30). This resistor is normally bypassed by a relatively large capacitor to minimize degenerative effects. It is this capacitor, however, that causes the cathode impedance to affect frequency response and time delay, particularly if the capacitor is not sufficiently large.

There are two ways out of this dilemma, short of using some form of fixed bias—a rather undesirable procedure in television receivers. One of these is to use a capacitor of extremely large size—in the order of 3000 to 5000 μf ; this figure is necessary for good low-frequency performance in sensitive radar circuits by straightforward transient analysis. The other alternative is not to bypass the cathode resistor. The loss of gain so incurred may be tolerable or not, depending upon the particular design. If the loss in gain can be tolerated then R_k is left unbypassed; if gain is very important (as in simple television video systems) then R_k must be bypassed by as large a capacitor as conditions permit. Satisfactory operation can usually be obtained with $C_k = 20 \mu\text{f}$ or more. For wideband amplifiers having rigorous performance requirements, the capacitor has to be very much larger than this (3000 to 5000 μf) or some other method of biasing must be used.

The completely designed video amplifier including high- and low-frequency compensation circuits appears in Fig. 27. Since this solution of the design problem can be applied to a television receiver, compromise values obtained in the design procedure are included.

38. Functional Test of Complete Video Amplifier

No design procedure is complete without operating tests. One of the most appealing tests of a complete video amplifier consists of the application of square-wave input and oscillographic study of the output; as previously explained, this type of transient analysis affords adequate information concerning both frequency response and time delay.

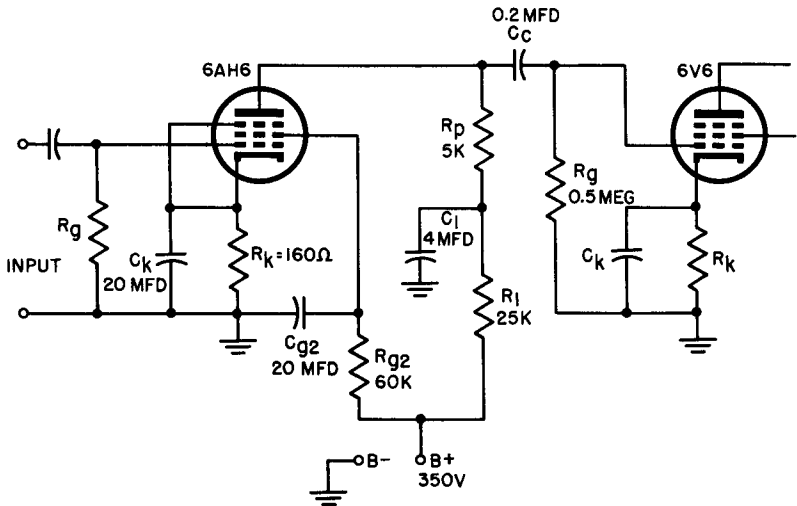


Fig. 27. The complete two-stage video amplifier with values suitable for television application.

If an electronic switch is available, both the square-wave input waveform and the resulting output waveform can be superimposed so that they may be viewed simultaneously; small irregularities and distortions in the reproduced wave are then easily detected.

Regardless of the precision with which the design is carried on, small unanticipated distortion factors often make their appearance and become visible in the analysis. The worker then has the opportunity to make corrections, slight as they may be, in the constants that make up the compensation circuits. If an electronic switch is not available, the input square wave is first displayed on the oscilloscope screen and note made of its coordinates on the quadrille plastic template on the CRT face.

Care must be exercised in keeping the input voltage low enough to avoid overloading the voltage amplifier stage. Both the high- and low-frequency performance is checked with square waves of appropriate fundamental frequencies. For the low frequencies in a television audio amplifier, testing is carried out using a square wave of 30 cycles; for the highs, the square-wave frequency should be between 200 and 400 kc to make certain that the harmonic content of the input rises to at least 4 mc. A series of oscillograms illustrating some typical video amplifier response defects are shown in Fig. 28.

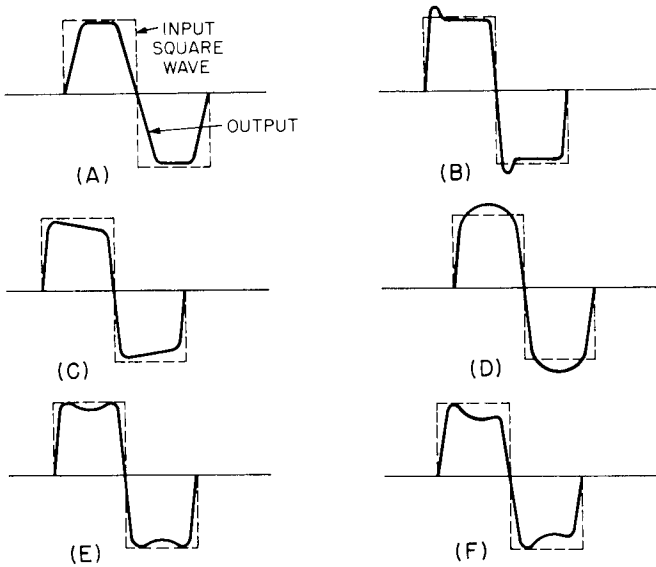


Fig. 28. Oscillograms showing response defects in improperly compensated video amplifiers.

These responses are necessarily idealized in structure but are easily recognizable in operating amplifiers. The dashed lines in each case represent the input square wave while the solid lines picture the waveform of the video amplifier output.

- (a) *Defect.* Rise time too slow; corners rounded.
Cause. Poor high-frequency response, no evidence of time-delay trouble.
- (b) *Defect.* Excessive overshoot.
Cause. Time delay and poor high-frequency response. An amplifier with this characteristic shows acceptable gain for most of the mid-frequency range, but a relatively sharp drop-off at the high end.
- (c) *Defect.* Excessive sag.
Cause. Time-delay distortion, excessive leading angle at the low frequency end of the amplifier's response range. This usually indicates an incorrect time constant in the low-frequency compensating network.
- (d) *Defect.* Convex peak.
Cause. Excessive compensation for the fundamental square-wave frequency resulting in overemphasis of the lows.

- (e) *Defect.* Concave peak.
Cause. Insufficient low-frequency compensation, but no phase-distortion.
- (f) *Defect.* Sag coupled with peak concavity.
Cause. Both low-frequency attenuation and phase-delay distortion. Correction required in low-frequency compensating network.

39. Review Questions

1. Outline the steps included in the design procedure for a compensated video amplifier.
2. Why is it advisable to build a wired prototype of a proposed video amplifier early in the design process?
3. Describe one method in common use for measuring the value of C_t in a video amplifier.
4. Find the mid-frequency gain of an amplifier stage in which the tube has a transconductance of 6500 and a plate circuit impedance 8500 ohms.
5. Describe the steps involved in the calculation of the inductance of a shunt-peaking coil.
6. Why may the plate impedance of a series-peaked amplifier stage be made larger than the plate impedance of a shunt-peaked stage?
7. Explain how you can tell by examining equation (67) that the gain of a combination-peaked stage is substantially greater than either series or shunt peaking alone.
8. Suppose that you wanted to design a compensated video amplifier having the best possible time-delay constancy for low frequencies. Which compensation method would you select? Why?
9. Explain in detail the procedure used to design the low-frequency compensating network R_1C_1 in Fig. 23.
10. Why does insufficient bypassing of screen dropping resistor and cathode resistor lead to poor low-frequency response? Why is it often suggested that cathode bias is not advisable in high-gain, wideband amplifiers?

Chapter 6

SPECIAL AMPLIFIERS AND MEASUREMENTS

40. Reduction of Amplifier Noise

Weak-signal amplifiers are limited as to the weakest signal amplified by the noise voltage amplitude present in the system. Extended discussions of the causes of circuit noise are presented elsewhere in this series; a summary of noise sources is given below, however, for review purposes.

Tube noise may be classified in six groups:

1. Shot effect due to random or spotty emission from the tube cathode.
2. Secondary emission noise resulting from variations in the number and velocity of electrons emitted from tube elements due to electron bombardment.
3. Gas noise arising from the variation in the number of kinds of ions produced in residual gas in the tube by electron and ion collisions.
4. Grid noise due to induced voltages that appear in the grid circuit as a result of the fact that the grid is immersed in an electron stream.
5. Current division or partition-noise that appears when currents going to different positive elements in the tube divide in random fashion, causing absolute electrode currents to vary slightly from moment to moment.

6. Cathode noise due to a relatively slow but repetitive variation of cathode emission. This is characteristic of oxide-coated cathodes.

Thermal noise originates in wires and other conductors as a result of thermal agitation of free electrons serving as current carriers. Thermal action causes the potential drop across the terminals to vary unpredictably and irregularly with time, producing very often a large noise voltage.

In designing amplifiers for service where low noise is an important consideration, it can be shown analytically and experimentally that a low value of coupling resistance and a small plate current magnitude can result in significant reductions in tube and thermal-noise voltage. Since the plate-coupling resistance is already small in the interests of wideband performance in video amplifiers, not much can be done in this direction without seriously reducing the gain. By starting the design with tubes having inherently low plate currents, the detrimental action of shot effect can be minimized.

High-gain, low-level input stages are most subject to noise induction. In such stages, thermal noise can be shown to be proportional to the square-root of the amplifier bandwidth. From the point of view of noise reduction, therefore, it is important that the amplifier bandwidth be no greater than necessary to provide the performance required.

As a final step, the grid-return resistor following the high-gain stage should be made as large as other considerations permit. Noise voltage developed in a grid impedance is proportional to the square-root of the impedance while the signal voltage varies directly as the value of this impedance. Increasing its size brings about an improvement in the signal-to-noise ratio.

41. High Voltage and Power Output from Video Amplifiers

In many applications involving video amplifiers, including the video system in television receivers, a relatively large output voltage across a small load resistor is required. Since power = E^2/R , more power may be required than can be handled by an ordinary voltage amplifier. In such cases it is customary to employ a power amplifier pentode or beam power tube to follow the voltage amplifier stages

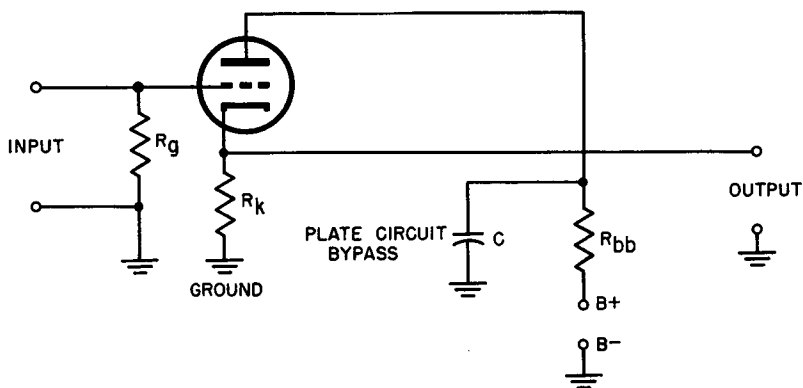


Fig. 29. A triode cathode-follower circuit. Note the use of a bypassed plate circuit resistor.

as in the example of Chapter 5. For this kind of a power amplifier the gain can be found with the equation:

$$\text{Gain} = g_m R_p \left(\frac{r_p}{r_p + R_p} \right) \quad (71)$$

The portion of equation (71) in parenthesis is a factor made necessary by the normally low dynamic plate resistance of power tubes (as compared to voltage amplifier tubes). Thus, r_p becomes an important factor in determining gain. In addition, its presence in the equation implies that compensation is also affected. Relatively-simple changes in the compensation network can easily correct for it.

42. Amplifiers Having Low-Impedance Output

Most transmission line systems, such as grounded-sheath coaxial cable used for carrying video signals over long distances, require a low-impedance source. Impedance transformation from the high output impedance of a normal video amplifier to the low input impedance of the cable cannot be handled by an impedance matching transformer since such units simply do not have a wide enough passband.

Reducing the plate circuit impedance of a video amplifier for impedance-matching purposes is not a desirable solution to the problem because of the drastic reduction in gain that accompanies such a step. Another unsatisfactory expedient that has been attempted has been the utilization of several tubes in parallel to

reduce the output impedance. The lack of success of the method is largely due to the great increase of shunt capacitance that occurs, with severe degradation of high-frequency response.

The best answer by far to the problem of matching the tube's high-output impedance to low-input impedance is the cathode follower. A representative triode cathode-follower circuit appears in Fig. 29. In addition to the low-output impedance feature, the cathode follower has an inherently good high-frequency response characteristic due to the extremely low equivalent plate resistance typical of this circuit; its low-frequency response is likewise excellent because the output can be coupled directly to the load without the use of a coupling capacitor. The gain and output impedance equations associated with cathode followers are:

$$\text{Gain} = \frac{g_m R_k}{1 + g_m R_k} \quad (72)$$

$$\text{Output } Z = \frac{R_k}{1 + g_m R_k} \quad (73)$$

From the gain equation (72), it is evident that the gain of a cathode follower can never exceed unity, although it can approach it closely if both g_m and R_k are relatively large. Hence, a cathode follower is not an amplifier in the strict sense of the word but serves as a virtually distortionless impedance-transformation device. The impedance-equation (73) is obviously of the general form indicating that two quantities ($1/g_m$ and R_k) are connected in parallel to produce an impedance lower than either. As g_m is made larger, its

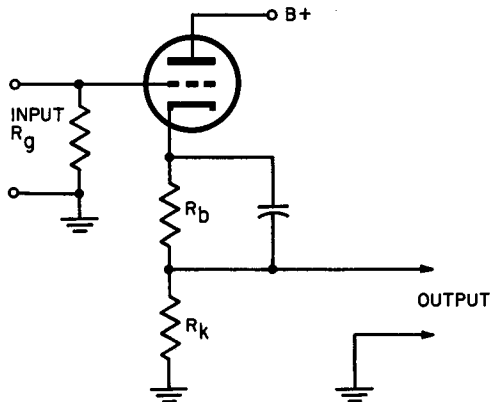


Fig. 30. Method of obtaining proper bias without raising output impedance above the calculated or desired value. R_k is the output resistor and R_b is the additional bias resistor.

reciprocal decreases, placing a lower impedance in parallel with R_k . Hence, the net resistance is considerably lower than R_k , often going down to 50 ohms or less. Equation (73) also shows that the output impedance is completely controllable by the designer, since R_k may be selected in conjunction with the tube g_m to give any output impedance less than itself. If, in this selection process, R_k becomes too low to provide adequate bias, then a second resistor may be added in series with R_k to establish proper bias conditions. In that case, the output signal is taken across the junction of the two resistors and ground, while the upper resistor is bypassed to prevent degenerative feedback from this source (Fig. 30).

43. The Measurement of Gain-Frequency Characteristic

The gain-frequency characteristic of a video amplifier can be most simply determined by connecting a reliable signal generator having a range from about 20 cycles to 5 mc to the input of the amplifier, and either a vtvm or a cro to the output as shown in Fig. 31. Capacitors C_1 and C_2 form a voltage divider in which C_2 is made variable; the respective capacitances should be selected so that it is possible to meet the conditions of equation (74) with reasonable capacitive values. Under these circumstances, both the ratio of input to output voltage for various frequencies can be read

$$\text{Gain of amplifier} = \frac{C_1 + C_2}{C_2} \quad (74)$$

directly from the vacuum-tube voltmeter or oscilloscope, and the gain of the stage can be obtained for any frequency with equation (74). A capacitive voltage divider (rather than resistive) is desirable in this arrangement because the large capacitances involved neutralize stray capacitances that may happen to be present. This is not true of resistive dividers since the stray capacitance between wires and within the potentiometer itself become important factors at high frequencies.

A complete series of input and output signal voltages is obtained by reading the vtvm first in one position of the switch and then in the other as the frequency is changed in convenient steps. It is essential, of course, that the frequency calibration of the signal generator be within the tolerances demanded by the nature of the measurement and the accuracy of the desired results.

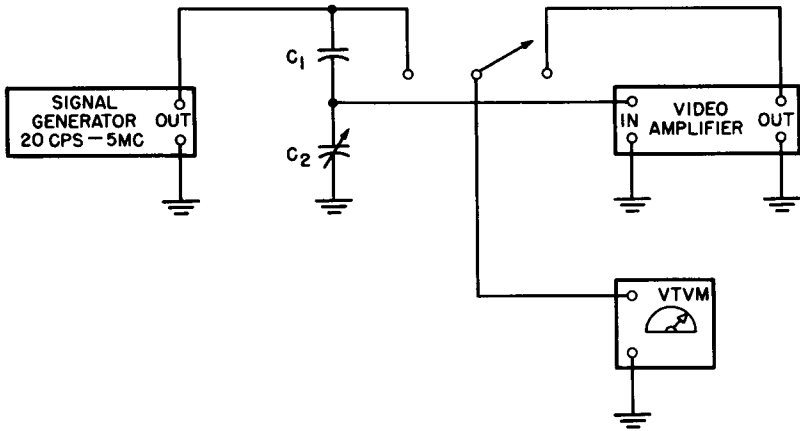


Fig. 31. Block diagram showing connections for measuring the gain-frequency characteristic of a video amplifier.

Gain is easily measured by first connecting the vtvm across the output terminals of the generator, by moving the switch to the left and reading some convenient value of signal voltage directly. The full output of the generator may be utilized, if required, to provide a quarter to three-quarter scale reading. The switch is then moved to the right-position and C_1 adjusted until the initial reading is again obtained. The gain is then computed with equation (74). This should be repeated over the entire video range, so that a plot of gain versus frequency may be obtained.

44. Determination of Phase Characteristic by Measurement

A quantitative estimate of time-delay amplifier characteristics may be obtained by using the arrangement shown in Fig. 32. Starting with the lowest frequency for which the amplifier is designed (say, 20 cps), C_1 and C_2 are adjusted so that the voltage applied to the horizontal deflecting plates is approximately the same as the voltage applied to the vertical deflection plates from the output of the amplifier with C_3 and R_1 set in their respective mid-positions. If the input and output voltages are in phase (phase difference = 0°), the Lissajous figure displayed on the screen is a straight line. This is also true for $\phi = 180n$ where n is any integer, but this offers no difficulty in measuring single stage amplifiers where the phase shift is always less than 90° . If a phase shift exists, the Lissajous

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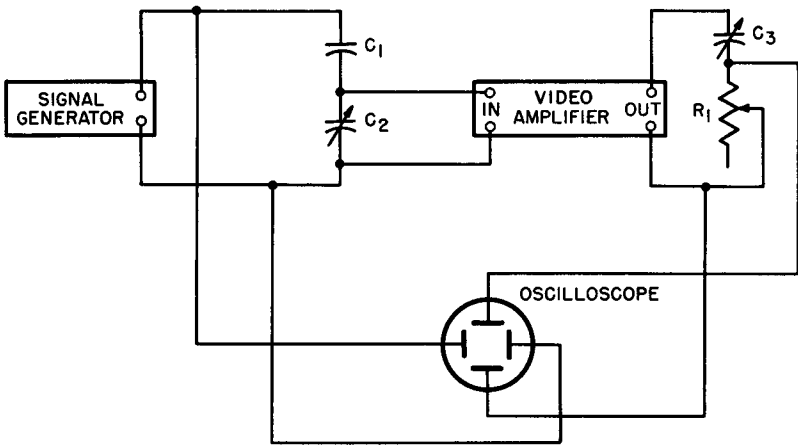


Fig. 32. Block diagram showing connections to determine amplifier phase-frequency response.

figure is an ellipse whose eccentricity depends upon the magnitude of the phase angle. By adjusting C_3 and R_1 (or either one alone, if possible), the display changes to a straight line. When this condition is obtained, the phase shift produced by the C_3R_1 network is given by:

$$\theta = \arctan \frac{1}{2\pi f R_1 C_3} \quad (75)$$

in which θ is the phase shift produced by the network comprising C_3 and R_1 , f is the frequency of the signal from the generator, and R_1 and C_3 are the values of the resistance in ohms and capacitance in farads, respectively.

The phase shift produced by the network R_1C_3 differs from the amplifier phase shift by 180° in the circuit as given. Thus, not counting the normal vacuum-tube phase shift of 180° , the added phase shift produced by the amplifier is:

$$\phi = \theta - 90^\circ \quad (76)$$

It must be observed that connection is made directly to the deflecting plates in the CRT in Fig. 32. This assumes that the output of the signal generator is sufficiently high to produce a practical deflection without requiring further amplification. The amplifiers built into the oscilloscope may be used if this condition cannot be obtained, provided, however, that both the V and H systems are identical in every respect. The solutions to equations (75) and (76)

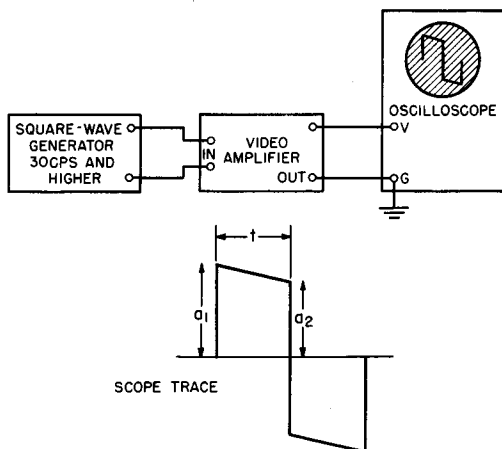
are not affected if both oscilloscope amplifiers have *equal* phase shifts even if the individual displacements are not equal to zero.

The final measurements are made by increasing the frequency in convenient steps; for each step, C_3 and R_1 are adjusted for a straight Lissajous line and the values for these settings substituted in equation (75). The signal generator (or generators, if more than one is used to cover the entire range from 20 cps to 5 mc) must be precisely calibrated for reliable results.

45. Response Measurements at Very Low Frequencies

The circuit shown in Fig. 33 together with the square-wave pattern as it appears on an oscilloscope screen (see insert in Fig. 33) enables the operator to obtain a quantitative value for the low-

Fig. 33. Method of determining low-frequency response in terms of fractional change of signal level.



frequency response in terms of a percentage change in signal level. Starting with a generator that produces a perfect square wave, the distances a_1 and a_2 are accurately measured and substituted in equation (78). This relationship is obtained from:

$$R_g C_c = t \frac{a_1}{a_1 - a_2} \quad (77)$$

in which R_g is the coupling resistor in the video amplifier grid circuit, C_c is the coupling capacitor, t is the length of the square wave half-period, a_1 is the amplitude of the leading edge, and a_2 is the amplitude of the lagging edge.

The ratio given in equation (78) is often termed the signal-level percentage because it represents the percentage change of television screen brightness between the top of a scanned frame and the bottom of the same frame. Thus:

$$\% \text{ change} = \frac{a_1 - a_2}{a_1} \quad (78)$$

Obviously, if there is no sag in the flat-top of the square wave, $a_1 = a_2$ and there is no change of signal level. In television reception, a change of more than 5% as given by equation (78) would become perceptible as a change in average background illumination going from the top to the bottom of the frame. Generally, the response is specified in manufacturing engineering to be no greater than 3% to provide uniform screen illumination over the entire screen when a plain background is being scanned.

46. Review Questions

1. Briefly describe the source of each of the following tube noises:
 - (a) Secondary emission noise
 - (b) Grid noise
 - (c) Shot effect
 - (d) Gas noise
 - (e) Partition noise
 - (f) Cathode noise
2. Which of the above is the most serious insofar as video amplifiers are concerned?
3. What are the customary steps in the design of a video amplifier necessary to reduce thermal and tube noise to a minimum?
4. Draw the diagram of a basic triode cathode follower circuit.
5. If the source voltage must be reduced by a series plate resistor, why must this resistor be bypassed in a cathode follower?
6. Prove that the gain of a cathode follower can approach unity, but never exceed it.
7. Describe a direct method for determining the amplitude-frequency characteristic of a video amplifier.
8. Describe a method for obtaining a quantitative figure for the phase shift produced by a video amplifier at different frequencies.
9. Under what circumstances may phase-shift measurements be made through oscilloscope amplifiers rather than directly to the deflecting plates? Why?
10. How may a square wave reproduced on an oscilloscope screen be analyzed to determine the percentage of signal drop-off from input to output at low frequencies?

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