LB-928

IMPROVING THE TRANSIENT RESPONSE OF TELEVISION RECEIVERS

RADIO CORPORATION OF AMERICA
RCA LABORATORIES DIVISION
INDUSTRY SERVICE LABORATORY

RADIO CORPORATION OF AMERICA

RCA LABORATORIES DIVISION

INDUSTRY SERVICE LABORATORY

LB-928

Improving the Transient Response of Television Receivers

Approved

Stuart Wm. Seeley

Introduction

Larger kinescopes and the advent of compatible color television have given increased importance to the problem of improving the transient response of television receivers. In terms of picture quality, this corresponds to improving the brightness transitions as well as the reproduction of fine detail.

Many contemporary receivers fall short of the goal of realizing the full electrical fidelity which the 4.5-Mc interval between picture and sound carriers makes possible. Frequently a wide bandwidth receiver will produce less crispness of picture detail than a narrower bandwidth receiver having lower wedge resolution. This is the direct result of the greater phase distortion inherent in the wide-band receiver's steeper attenuation slope and the increased difficulty of compensating for this distortion with minimum-phase-shift networks. When properly designed, a wide-band receiver gives the expected improvement in both reduced rise time and increased wedge resolution.

This bulletin is concerned with those receiver design factors that make for improved electrical fidelity, although the interlocking nature of the transmitter and receiver requires some consideration of the transmitter aspects of the problem. In the receiver, the electrical fidelity is primarily determined by the delay distortion produced in the video i-f amplifier, and the extent to which this distortion is compensated in the peaking circuits of the video detector and the video amplifier. Optimum results are obtained by working with both the phase-delay characteristic of the overall receiver and its corresponding amplitude characteristic. In general, correction of the phase-delay characteristic results in peaking the amplitude response in the mid-frequency range, unless non-minimum phase-shift networks are used. Where non-minimum phase-shift networks are employed, superior performance is obtained because compensation of delay distortion is achieved without distortion of the amplitude response.

In color receivers, prevention of delay distortion has been found to be of primary importance, as might be expected from the fact that the channel is being used to greater capacity. The factors which determine cross-talk between the I and Q color-difference signals and coincidence of the luminance and chrominance information are described in detail.

Variability in the quality of transmitted pictures has led to a tendency on the part of receiver designers to introduce more or less overshoot in the receiver. Widespread adoption of the standardization suggestions made by Kell and Fredendall[1] seems desirable, particularly with respect to the need for a standard monitor. In addition, more attention should be given to correcting at the transmitter for the distortion introduced by the vestigial sideband filter. Current FCC proof of performance includes no specification on the phase-delay characteristics of the transmitted signal.

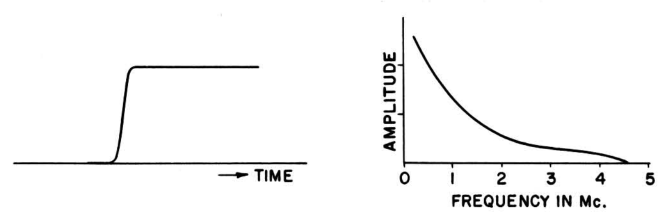
[1]Kell and Fredendall, "Standardization of the Transient Response of Television Transmitters", *RCA REVIEW*, March 1949.

General Considerations

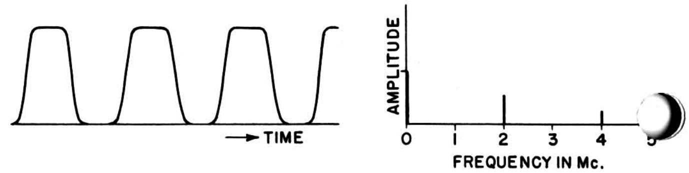
The response of a receiver may be evaluated by examining its response to the two test waveforms shown in Fig. 1. These transient waveforms are made up of frequency components which range from zero frequency to 4.5 Mc, the exact form of the spectrum varying with the particular transient. For example, the frequency spectrum of the step transition has an amplitude that varies inversely as the frequency, while the spectrum corresponding to a steady-state wedge pattern is made up of only discrete harmonics of the fundamental.

This viewpoint of going from the time domain to the frequency domain, as in Fig. 1, points up the need for transmitting the frequency spectrum making up the transient so that all components arrive at the output with the same time delay and the same relative amplitude. In the minimum phase-shift networks which are commonly employed in the i-f and video circuits of television receivers, the phase response of the network is uniquely defined once the amplitude response of the network is fixed. Aside from the use of non-minimum phase-shift networks, a more favorable delay characteristic can therefore be obtained only by distortion of the amplitude response. As can be seen intuitively, it is more important that the relative delays of the individual spectrum components be the same, than it is that their relative amplitudes be closely preserved. For this reason it is usually desirable to distort the amplitude response to make the phase delay more constant.

It is apparent from Fig. 1 that the reproduction of a step change in brightness requires that uniform time delay be maintained over a very wide range of frequencies, from a few tenths of a megacycle to 4.5 Mc. Reproduction of good wedge detail, however, is much less difficult; here only a few harmonics combine to form the wedge and uniformity of delay is required over a narrower band of frequencies. In the limit, wedge detail corresponding to greater than a 100-line resolution is reproduced as a single sinusoidal component and good reproduction is obtained even though



(a) – Step transition and its frequency spectrum.



(b) – High-frequency square wave and its frequency spectrum.

Fig. 1 – The transient response of a television receiver may be evaluated by its response to the two idealized waveforms. The corresponding frequency spectra of these waveforms are shown.

the delay characteristic of the circuit may be non-uniform and the unit step response correspondingly poor.

It is unfortunate that good resolution is so often advanced as being a criterion of transient response. Taken alone, the ability to resolve the vertical wedges of a test pattern is an exceedingly poor measure of transient response and is much inferior to the unit step characteristic, which shows up the presence or absence of smear, preshoot, overshoot, and ringing.

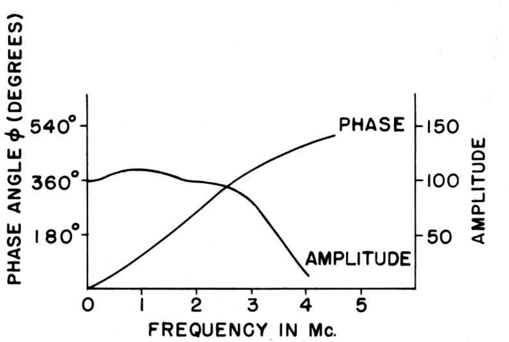
Phase and Envelope Delay

Various methods are used for evaluating the uniformity of transmission time through a television receiver. These include:

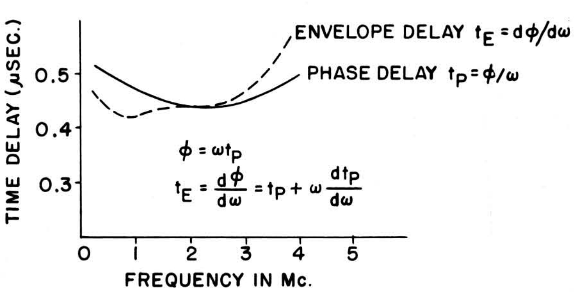
(a) Plotting phase angle against frequency

(b) Plotting phase delay $t_P = \varphi/\omega$ against frequency

(c) Plotting envelope time delay $t_E = d\varphi/d\omega$ against frequency.



(a) – Amplitude and phase.



(b) – Corresponding phase delay and envelope delay.

Fig. 2 – Typical curves to illustrate the relation between amplitude, phase angle, phase delay and envelope delay.

The first method is straightforward. However, the phase angle curve (Fig. 2a) is difficult to interpret because small changes in the slope cause large changes in the relative phase delay.

This difficulty can be partially overcome by plotting the phase delay vs frequency as in Fig. 2b. This delay curve is relatively easy to evaluate since deviations from the "average" delay can be interpreted in comparison with the rise time of the circuit.

The third method used for evaluating the phase distortion of the output video signal is to plot the envelope delay $t_E = d\varphi/d\omega$ vs frequency. This frequently is plotted by taking the slope of the phase angle φ vs frequency curve. More accurately, the envelope delay can be determined from the phase delay as follows:

Since

$$\text{phase delay} = t_P = \frac{\varphi}{\omega}$$

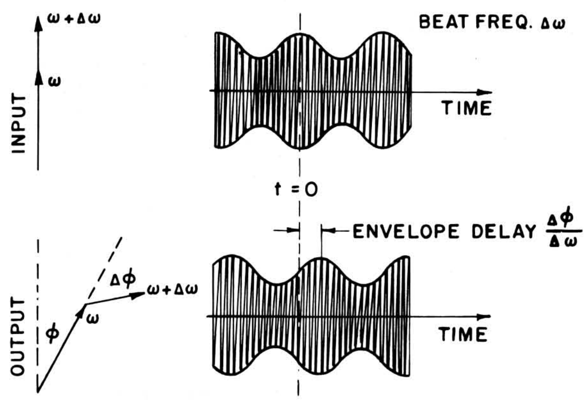
$$\varphi = \omega t_P$$

$$\therefore \text{envelope delay} = t_E = \frac{d\varphi}{d\omega} = t_P + \omega \frac{dt_P}{d\omega} \qquad (1)$$

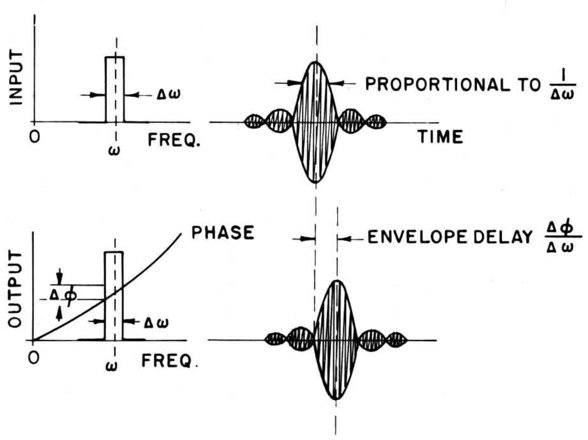
Thus the envelope delay is equal to the phase delay plus a term which takes into account how rapidly the phase delay varies with frequency.

The envelope delay of a video signal may be measured, as shown in Fig. 3, by taking two frequencies close together and noting that they form a low-frequency envelope. For example, 3-Mc and 3.1-Mc beat to form a 3-Mc signal the envelope of which varies at a 100-kc rate. The envelope delay $d\varphi/d\omega$ then represents the delay of this low-frequency envelope. In the classical case, rather than taking just two frequencies, a spectrum interval $\Delta\omega$ wide is taken to form a "low-frequency" wave packet; the envelope delay is then the delay of this low-frequency pulse.

The use of envelope delay in the analysis of television receiver distortion has probably received its impetus from the questionable concept that the transient response corresponding to any spectrum input can be determined by summing the wave packets or groups corresponding to each of the intervals $\Delta\omega$ in the spectrum. For distortionless transmission, the argument goes, one should expect that the envelope delay – the delay of all these packets – be constant. However, more careful consideration shows that the summation of these wave packets requires that the phase of the carrier within each group be known.[2] Since this phase depends upon the phase delay characteristic, it is clear that specifying only the amplitude and envelope delay from a few hundred kilocycles to four megacycles is insufficient to determine the transient response of the receiver. Furthermore, as Eq. (1, indicates, it is possible for the envelope delay of a receiver to be constant even though the phase delay is varying, and hence constancy of envelope delay is a necessary but not a sufficient condition for distortionless transmission.



(a) - Envelope formed by ω and ω+Δω.



(b) - Envelope formed by spectrum interval Δω wide.

Fig. 3 - Envelope delay may be interpreted as (a) the delay of the low-frequency envelope formed by ω and ω+Δω, or as (b) the delay of the envelope formed by a spectrum interval Δω wide.

[2]Hartley, R.V.L., "Steady State Delay as Related to Aperiodic Signals", *B.S.T.J.*, Vol. 20, p. 223, 1941.

For this reason, phase delay is a more usef tool than envelope delay in evaluating t transient response of television receivers.

Color Receiver Considerations

A critical examination of transient reponse is particularly timely, since the requirements of color receivers are more severe than those of monochrome.[3]

Fig. 4 shows the bandwidth occupied by the luminance and chrominance components. The I and Q color-difference signals are carried as in-phase and quadrature amplitude modulation of the 3.58 Mc color subcarrier. The fact that the chrominance components occupy overlapping bandwidths imposes the requirement that the amplitude and delay be constant over this region in order to prevent crosstalk between the I and Q signals. In addition, uniformity of delay is required to insure good coincidence of the luminance and chrominance contributions to abrupt transitions in color.

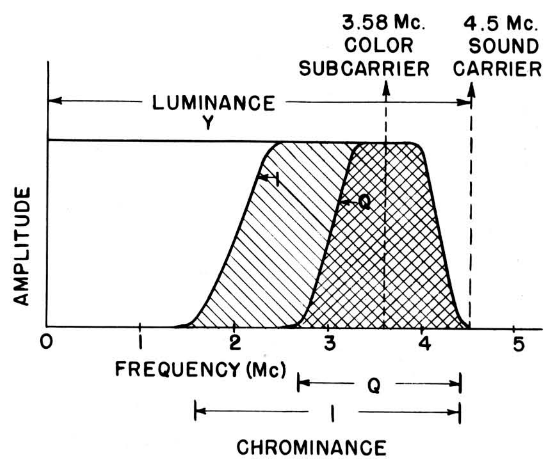


Fig. 4 - Bandwidth of luminance and chrominance components of a compatible color television signal. Factors determining crosstalk and relative delay are described.

The problem of insuring constant delay the color sideband region is eased by the phase predistortion proposed in the NTSC signal specifications. This predistortion, shown in

[3]Brown, G.H. and Luck, D.G.C., "Principles of Colo Television Systems", *RCA REVIEW*, Vol. 14, p. 14 June 1953.

Fig. 5, is based on the high-frequency phase distortion introduced by a typical receiver which has flat amplitude response to approximately 4 Mc. This correction is given in terms of envelope delay and thus its interpretation requires some care. In particular, while an overall constant envelope delay is sufficient to insure absence of cross-talk, it is not sufficient to insure time coincidence of the luminance and chrominance unless, at low video frequencies, the envelope and phase delays are equal. In general, however, these delays are not equal in the low-frequency range because of the distortion resulting from vestigial sideband reception. For this reason, it is necessary to compare the *phase delay* of the luminance information with the *envelope delay* at the color subcarrier in order to obtain a measure of the coincidence of luminance and chrominance information.

A measure of color cross-talk more useful than envelope delay may be obtained by measuring the delay of the envelope formed by beating a single color sideband of adjustable frequency

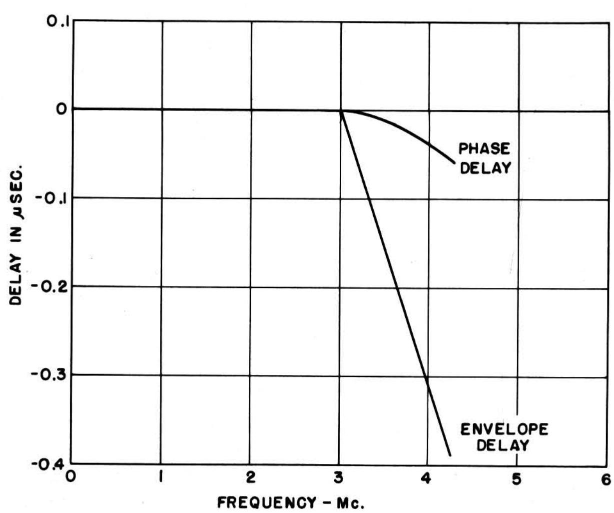


Fig. 5 - The NTSC specification for the envelope delay at the transmitter and the equivalent phase delay.

against the chrominance subcarrier. This "single-sideband" phase delay has the advantage over envelope delay of being the actual delay of the information contained in each color sideband, and hence of being a more direct indication of cross-talk.

In the absence of cross-talk, the overall color system may be viewed as one in which the color information is transmitted as separate amplitude modulation of in-phase and quadrature components of the color subcarrier. These modulations may then be considered as being synchronously detected at the receiver by a process equivalent to exalted carrier diode detection. This point of view indicates that the in-phase and quadrature signals are handled by the receiver in the same manner, and that there is only one equivalent response for both the in-phase and quadrature components. This response corresponds to low per cent amplitude modulation of the color subcarrier. The difference between the phase delay of this equivalent response and the corresponding phase delay of the luminance information is indicative of the lack of time coincidence of the chrominance and luminance.

IF Amplifier Considerations

Phase distortion in the i-f amplifier is a major factor in determining the electrical fidelity of the receiver. In general, however, the i-f amplifier must be designed primarily to provide the required selectivity. When the selectivity requirements are met, the phase response is correspondingly determined since,

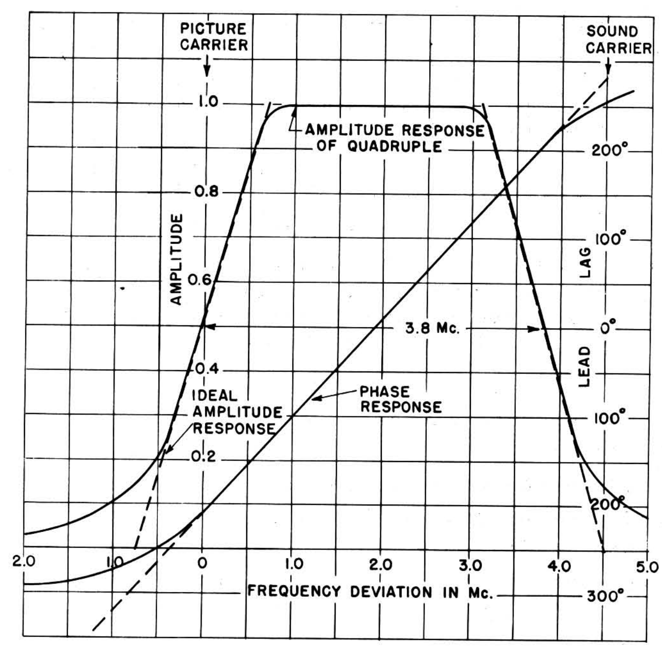


Fig. 6 - The amplitude and phase characteristics of a flat-staggered quadruple having a 6 db bandwidth of 3.8 Mc. The idealized amplitude characteristic of a receiver is shown dotted.

in general, conventional i-f circuits are minimum phase-shift networks. Despite the restrictions imposed by the need for adequate selectivity, typical i-f designs currently in use show considerable variation in phase response.

To show what can be expected of i-f amplifiers and the manner in which the phase characteristic is influenced by the amplitude characteristic, an investigation was made of the transient response of a flat-staggered quadruple. The bandwidth chosen was 3.8 Mc to approximate the amplitude characteristics of an ideal receiver, as shown in Fig. 6.

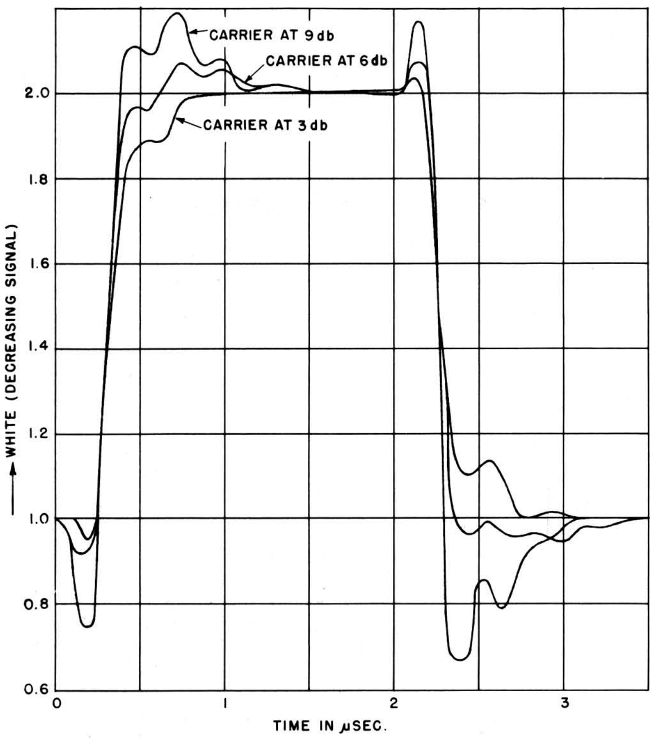


Fig. 7 – The square wave response of a flat-staggered quadruple having a 3.8 Mc 6 db bandwidth for low per cent (2:1) modulation when the carrier is at 3 db, 6 db, and 9 db respectively.

Using the Laplace transform, the square-wave response of this i-f amplifier was determined. Fig. 7 shows the response for a 2-to-1 step change when the carrier is located at the 3-db, 6-db, and 9-db points of the attenuation characteristic. In each case the transition is preceded by a preshoot or precursory transient. The carrier at 6 db yields about the best performance. Lowering the picture carrier to the 9-db response point has the underirable effect of increasing the preshoot and at the same time increasing the overshoot. Raising the carrier to the 3-db response point reduces th preshoot but has the disadvantage of increasin the rise time and introducing considerable smear.

The effect of increasing the per cent modulation is shown in Fig. 8. Here the transitions are shown with the carrier at the 6-db point for a 2-to-1 and for a 4-to-1 change in modulation level. If black level is taken as 72 per cent, then the 4-to-1 step would correspond to a variation from the 72 per cent carrier amplitude corresponding to black to the 18 per cent amplitude corresponding to (near white. The effect of the increase in per cent modulation is to slightly increase the rise time and the amount of smear. At the same time the difference between the black-to-white and

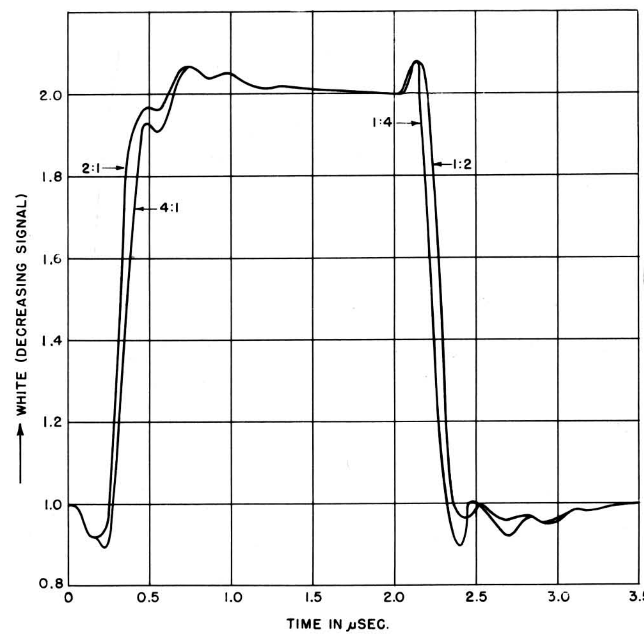


Fig. 8 – Square wave response of a flat-staggered quadruple having a 3.8 Mc 6 db bandwidth with the carrier at the 6 db point, for a modulation change of 2:1 and 4:1. The curves are normalized for comparison purposes.

white-to-black transitions is increased. The noticeably larger preshoot in the case of the black-to-white transition is evident. This difference between the black-to-white an white-to-black transitions with increasing pe cent modulation is characteristic of band-pass amplifiers in which the carrier is located off center.

The phase-delay curves for the quadruple of Fig. 6 are of interest and are shown i Fig. 9. The solid line shows the phase delay

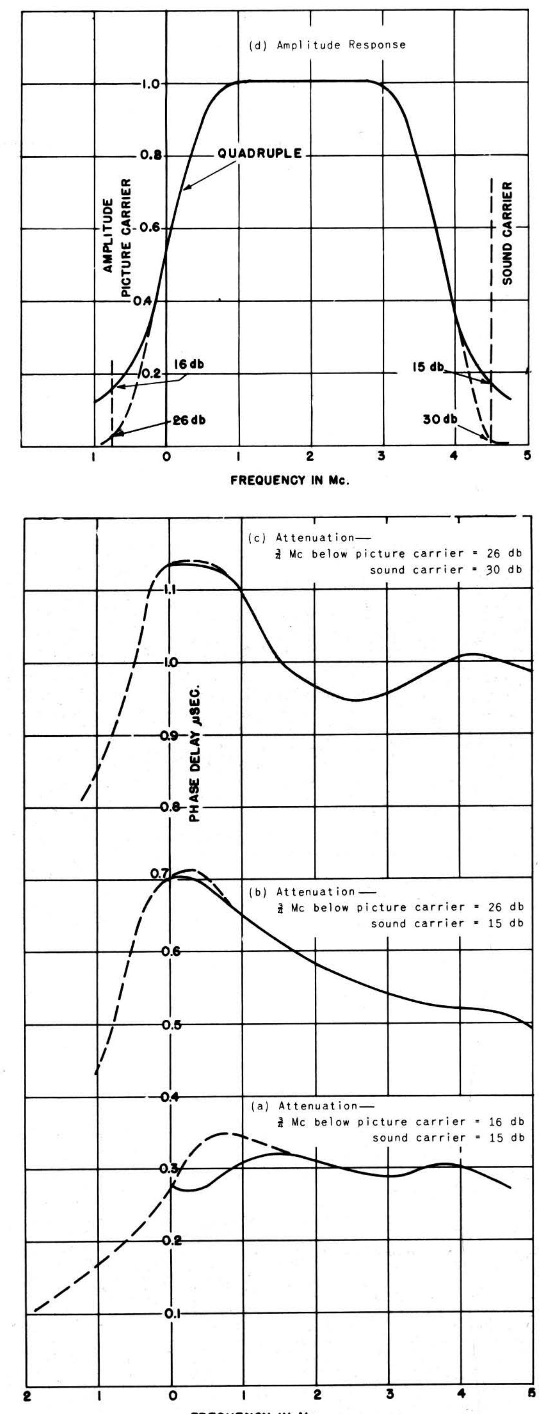


Fig. 9 – Phase delay of (a) quadruple, (b) quadruple with added attenuation below picture carrier, and (c) quadruple with added attenuation below picture carrier and added attenuation at the sound carrier. The solid line represents the response to a double sideband signal, the dotted line to a single sideband signal. The amplitude response is shown in (d). Normally the total delay is less than in (b) and (c) because of the "after response" characteristic of conventional traps.

associated with double-sideband modulation of the picture carrier. The dotted line shows the phase delay associated with single sideband modulation of the picture carrier. Above 1.25 Mc there is negligible difference between the two curves.

Since a quadruple 3.8 Mc wide without traps has considerably less than normal i-f selectivity, the effect of increasing the attenuation to 26 db at a frequency ¾ Mc below the picture carrier was investigated. The result as shown in Fig. 9b is greatly increased phase-delay distortion. If, in addition, the attenuation at the sound carrier is increased to 30 db, this added attenuation results in the "rise" in phase delay, characteristic of i-f amplifiers as the video modulating frequency approaches 4.5 Mc.

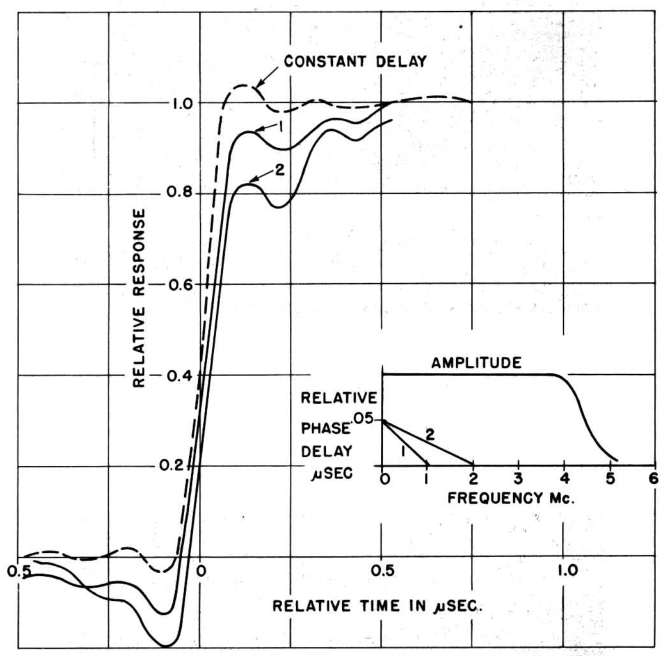


Fig. 10 – The effect of mid-frequency delay distortion on the response of an idealized filter having a 4% overshoot.

The total phase delay distortion of the quadruple with the added attenuation below the picture carrier and at the sound carrier is shown in Fig. 9c. The resultant phase distortion is significant since the change in phase delay exceeds the 0.1-μsec value corresponding to the duration of a single picture element. All of these curves are based on the minimum phase characteristic corresponding to the assumed amplitude characteristic.

To evaluate the effect of mid-frequency phase distortion on the reproduction of a square wave, an approximate calculation was made of the effect of a phase delay distortion that varies linearly (a) from 0.05 μsec at zero frequency to 0 μsec at 1 Mc and (b) from 0.5 μsec at zero frequency to 0 μsec at 2 Mc. As Fig. 10 shows, the effect of this distortion is to distort the step response of an "ideal" filter by increasing the amplitude of the preshoot and by introducing smear. Where the distortion extends as far as 2 Mc, the amount of smear is approximately 20 per cent, i.e., the initial steep rise reaches only to the 80 per cent response point.

The preceding description points out the magnitude of the phase distortion that can be expected in video i-f amplifiers and its effect in causing preshoot and smear. Investigation of the transient response of typical receivers has

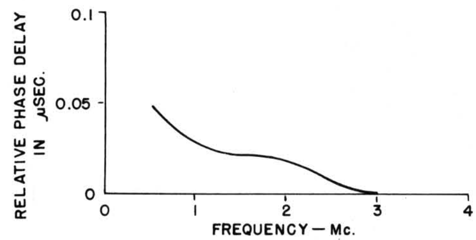
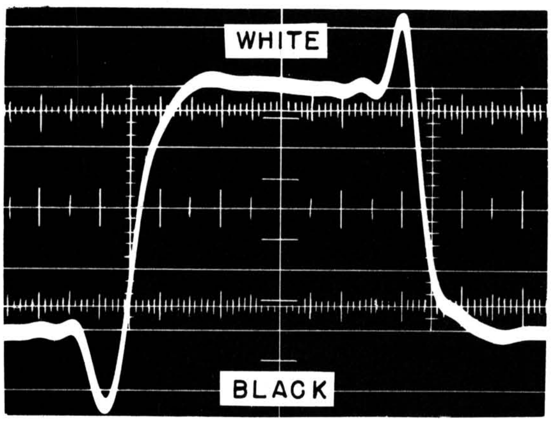
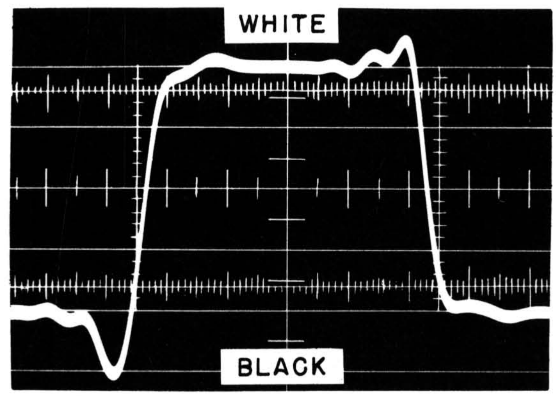


Fig. 11 - The distortion introduced by a typical vestigial sideband filter at the transmitter.



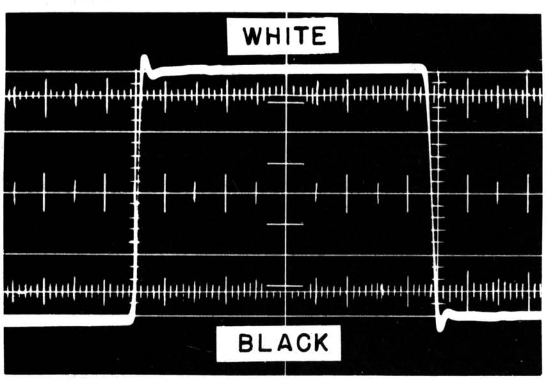
Black-to-white rise time = 0.29 μsec.
White-to-black rise time = 0.20 μsec.

(a) Response without the corrective network.



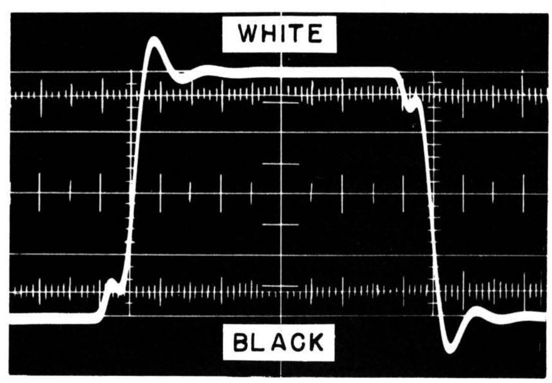
Black-to-white rise time = 0.18 μsec.
White-to-black rise time = 0.17 μsec.

(b) Response with the corrective network.



Black-to-white rise time = 0.04 μsec.
White-to-black rise time = 0.06 μsec.

(c) Input to corrective network.



Black-to-white rise time = 0.10 μsec.
White-to-black rise time = 0.10 μsec.

(d) Output from corrective network.

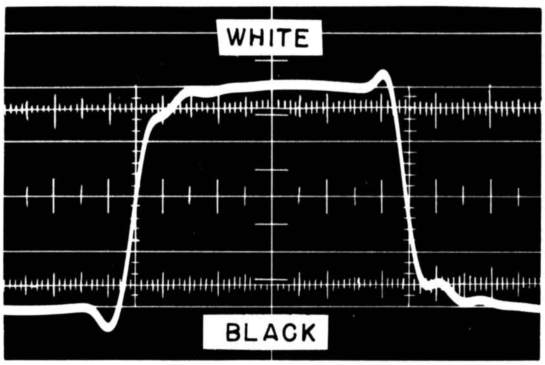
Fig. 12 - A typical receiver response (a) with, and (b) without phase correction for the distortion introduced by the vestigial sideband filter at the transmitter. The response of the transmitter corrective network is shown at (d).

shown extremely wide variations in performance ranging from excessive smear and preshoot, to excessive overshoot and ringing. The delay characteristics of these receivers show correspondingly large variations, particularly in the region from 0.1 Mc to 3 Mc. Correction of the delay distortion in this region has been found to be of far greater importance than correction of the distortion due to the high-frequency cutoff.

The distortion usually produced in the i-f amplifier is considerably more severe than that produced by failure to correct the distortion produced by the vestigial sideband filter of the transmitter. This might be expected since the vestigial sideband filter attenuation does not "bite" as close to the carrier as does the i-f amplifier characteristic.

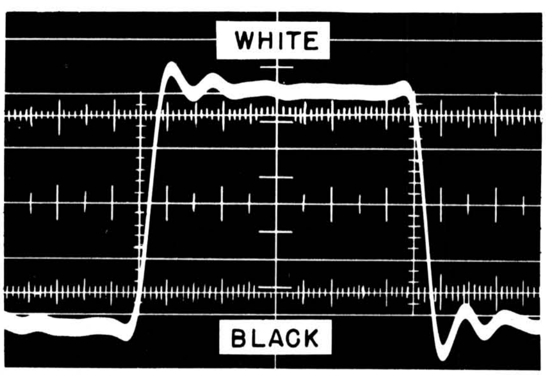
A transmitter vestigial sideband filter typically introduces distortion having a peak value of 0.05 μsec and tapering off to zero at approximately 3 Mc, as in Fig. 11. A typical receiver response, with and without correction at the transmitter for the vestigial sideband filter distortion, is shown in Fig. 12. Here again the effect of the distortion is excessive smear and preshoot.

The vestigial sideband distortion due to the transmitter is conventionally corrected in the transmitter by predistorting the modulating signal. The type of distortion required to provide this correction is shown in Fig. 12c. Note how the predistortion of the square wave is roughly "complementary" to that of the uncorrected distorted output.



Black-to-white rise time = 0.28 μsec.
White-to-black rise time = 0.22 μsec.

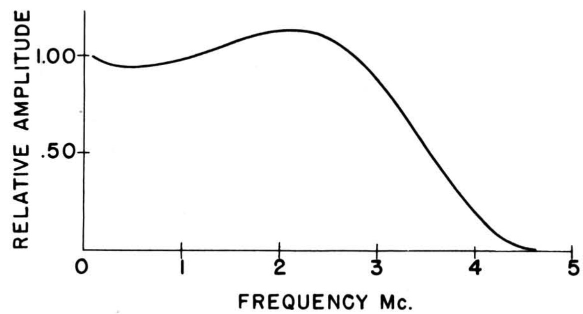
(a) Without peaking



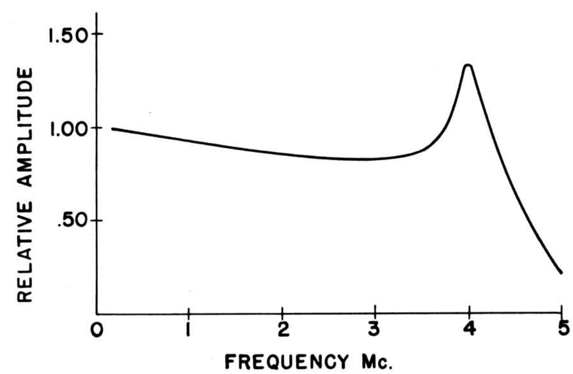
Black-to-white rise time = 0.14 μsec.
White-to-black rise time = 0.14 μsec.

(b) With peaking

Fig. 13 - The effect of detector peaking on the response of a typical receiver. (a) shows the response from the antenna to the detector with minimum shunt capacitance and no peaking; (b) shows the response with typical shunt capacitance and peaking adjusted for optimum.



(a) - Using video-modulated r-f signal.



(b) - Using video signal and 3300-ohm dummy resistor.

Fig. 14 - The effect of the video detector peaking can be determined by applying a modulated r-f signal and measuring the output (a) at the video amplifier grid. Although this includes the i-f amplitude response, it is a better index than the curve obtained by using a simulated 3300-ohm dummy detector output impedance as in (b).

I-f amplifier distortion is usually corrected by the video peaking. Ideally, it would be desirable to eliminate phase distortion in the i-f amplifier by using non-minimum phase-shift traps for the adjacent sound selectivity. However, in the present state of the art, the burden of providing the phase correction falls upon the video detector and video amplifier peaking circuits.

Lowering of the picture carrier to improve the overall transient response is occasionally used. This has the advantage of providing an overshoot which tends to compensate for the smear inherent in typical i-f phase-delay distortion. However, it also tends to cause excessive preshoot and therefore is not a basic solution to the problem where maximum fidelity is desired.

Video-Detector Peaking

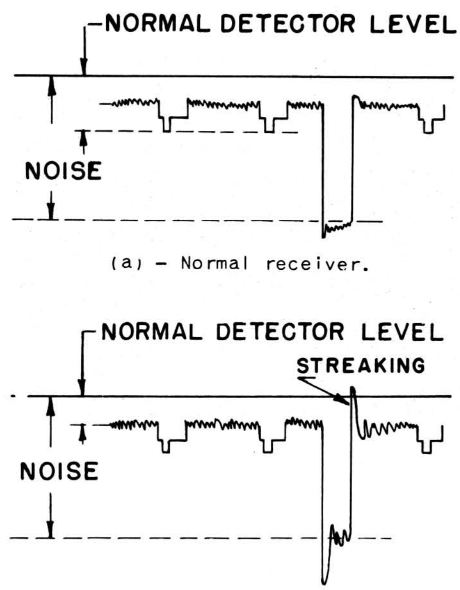
Peaking of the video detector presents a problem because of the many factors which influence the results. These include the characteristics of the tuned circuits driving the detector, their resonant frequency with respect to the carrier frequency, and the ratio of unloaded to loaded Q. Because of this, it has been found desirable to consider the video detector as an integral part of the i-f amplifier and to peak the video detector while observing the output corresponding to a modulated r-f signal applied to the antenna terminals of the receiver.

Fig. 13a shows a typical square-wave response measured from the antenna terminals to the video detector with no added peaking; the output is taken across the video detector load resistance with only enough added capacitance to serve as an i-f bypass capacitance. The improvement effected when the peaking is adjusted for optimum rise time and symmetry of transition is shown in Fig. 13b.

The frequently-used video detector generator impedance of 3300 ohms (resistive) has been found to yield results of doubtful value. When the results of peaking the video detector by considering it an integral part of the i-f amplifier are compared with the results obtained by peaking conventionally with 3300 ohms, there is a wide disparity in the peaking. This is illustrated in Fig. 14 which shows (a) the amplitude response measured through the i-f amplifier and video detector by using a video-modulated r-f signal and (b) the amplitude response of the video detector as measured with a video signal and a 3300-ohm dummy generator resistance. The difficulty of arriving at the optimum peaking by using the 3300-ohm dummy resistor is apparent.

In the actual mechanics of detector peaking a certain amount of cut-and-try is necessary because of variations in the detector input transformer circuitry and the detector capacitances. Peaking of the detector is particularly difficult where a triode video amplifier is used, because of Miller effect. In this case it is necessary to peak the video amplifier before peaking the video detector. In general, the time constant formed by the detector load resistance and the sum of the input and output capacitances must be sufficiently small to enable not only adequate amplitude response, but partial phase-delay compensation for the i-f amplifier delay distortion.

From the standpoint of obtaining maximum video signal output per milliampere of plate-supply current, it is desirable to use a relatively high video plate-load resistor in the video-output stage and to compensate for the



(a) - Normal receiver.

(b) - Overpeaked receiver.

Fig. 15 - The effect of excessive ringing in the video detector peaking is to cause white streaking as shown at (b).

resulting loss in high-frequency response by overpeaking in the video detector. If carried too far, this procedure results in white streaks following impulse noise. The mechanism is shown in Fig. 15. Even a relatively small percentage overshoot is effective in causing white level to be reached because of the absence of appreciable limiting up to the video detector. An equivalent amount of ringing in the video amplifier would not produce white noise because of the limiting in the video amplifier.

Video-Amplifier Peaking

In peaking the video amplifier of a television receiver, the normal data available for obtaining maximum gain-bandwidth product or maximum gain-rise time ratio is applicable only as a starting point. This is because the video amplifier must correct the phase delay distortion introduced by the video i-f amplifier. When this distortion is corrected, the video amplifier amplitude response when viewed separately will appear far from optimum, although the overall transient response is improved.

Experience has indicated that peaking of a narrow-band (3 Mc) receiver is less difficult than peaking a wide-band receiver. This is the result of two factors (1) the phase distortion produced in the i-f amplifier is less and (2) the narrow band of the video amplifier makes it possible to phase correct the residual i-f distortion without further compromising the overall receiver bandwidth. The calculated phase-delay response of a 3-Mc i-f amplifier composed of two flat-staggered pairs is shown in Fig. 16. This may be compared with the considerably higher phase distortion of the wideband staggered quadruple, with comparable skirt selectivity, shown in Fig. 9c.

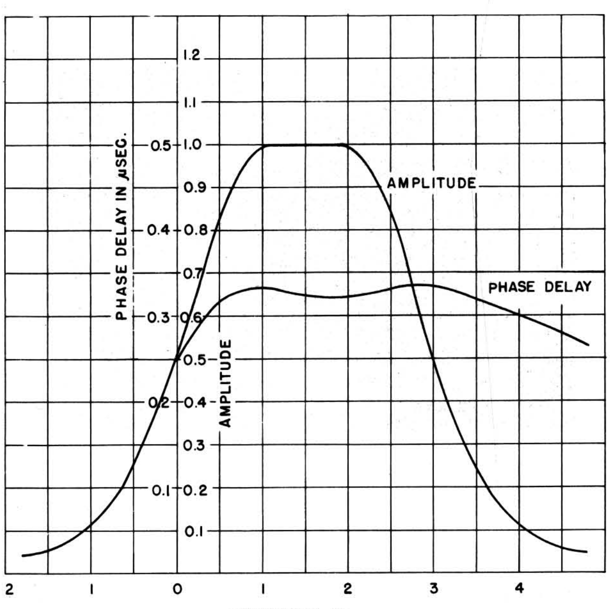


Fig. 16 - Equivalent video phase delay of two staggered pairs having a 3.0 Mc 6 db bandwidth.

Because of the need for correcting the i-f phase delay distortion, the choice of load resistor, peaking component values, and the location of the termination (with respect to the low-capacitance side of the filter) are more critical than in an ordinary video amplifier where correction for i-f phase distortion is not required.

The square-wave response alone is not an adequate index of video-amplifier peaking. It needs to be supplemented by examination of the swept amplitude response characteristic. Examination of both of these responses makes it possible to obtain the best square-wave response consistent with good high-frequency resolution, as indicated by the amplitude response.

Where only the square-wave response is used in determining the peaking constants, the tendency is for the amplitude response to be restricted at the high frequency end, and for the middle video frequency range to be boosted. This distortion of the amplitude response, when minimum phase-shift networks are used, makes possible correction of typical i-f amplifier phase-delay distortion.

An effective method of improving overall receiver transient response is the frequently used partially-bypassed cathode resistor. This introduces an overshoot in the square-wave response which is in the direction to compensate for i-f phase distortion and to crispen the picture. Where the cathode resistor is variable, a control with one or two taps enables the amount of overshoot to vary as a function of the contrast setting.

A partially-bypassed video-amplifier screen has been successfully used in obtaining a similar result. This is particularly advantageous in stages which have a variable cathode resistor since the overshoot tends to be independent of gain. The partially bypassed screen circuit has also been effective in compensating for long black-to-white smear.

Optimum results in peaking video amplifiers are obtained when non-minimum phase-shift networks are used. Thus, the use of such all-pass networks, for example, the bridged-T network shown in Fig. 17, makes it possible to correct the phase-delay distortion without restricting the bandwidth. The improvement obtainable in the unit step response of a receiver by means of such a network is shown in Fig. 18. These networks have the general disadvantage that they must be operated at relatively low impedance level because the shunt capacitances in the video amplifier must be rendered negligible. Promising results, however, have been obtained with modified non-minimum phase-shift networks which provide good gain-bandwidth as well as the desired phase correction.

A source of poor transient response which not infrequently crops up is phase distortion in the 1-Mc frequency range due to the disturbance caused by coupling the sync separator (and in some cases the AGC rectifier) to the video amplifier. As shown in Fig. 19, the distortion arises as a result of the variation in the impedance which the isolating resistor and the sync separator input capacitance shunts across the video peaking filter. The reaction on the filter is observed to be decreased by connecting the sync input to the kinescope feed point (A) rather than to the load resistor (B).

Where square-wave modulation of a carrier is employed without composite sync, it is essential that the sync separator not conduct during transition intervals and reflect an abnormally high capacitance across the circuit because of the Miller effect. Since the sync separator is normally cut off during the picture interval, it is necessary to artifically bias off the sync separator to avoid this excessive reflected capacitance.

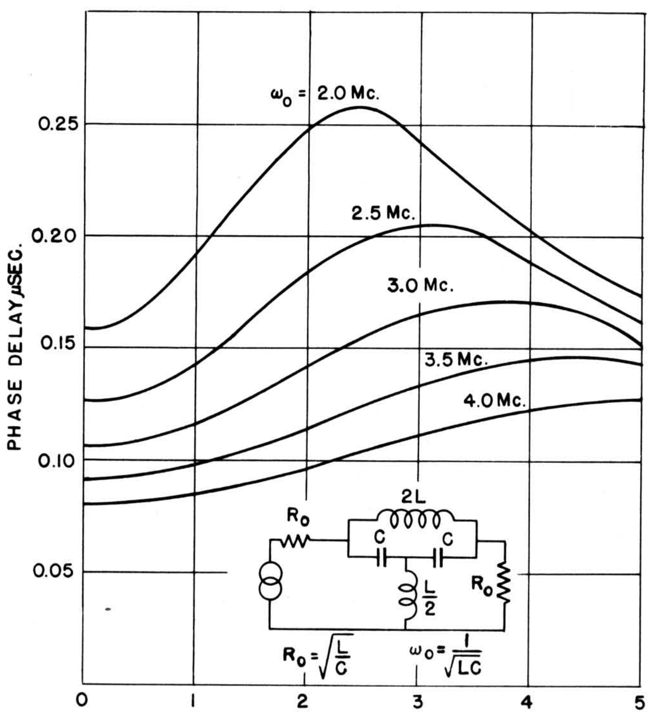
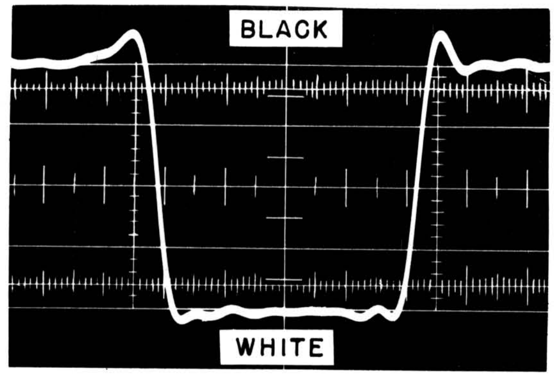
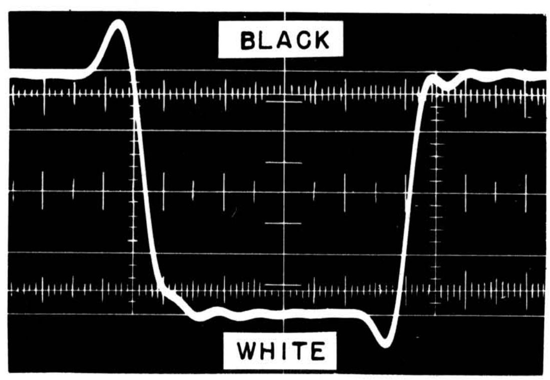


Fig. 17 – Phase delay of an all-pass phase-correction network.



Black-to-white rise time = 0.14 μsec.
White-to-black rise time = 0.14 μsec.

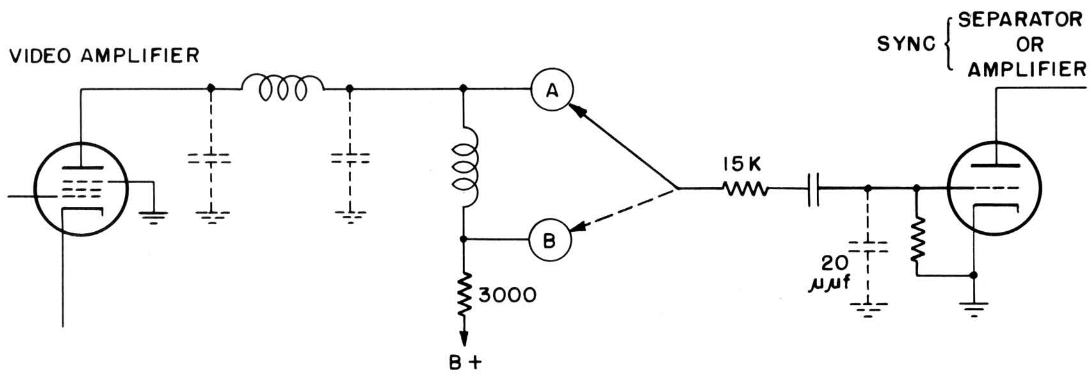
(a) With bridged-T network.



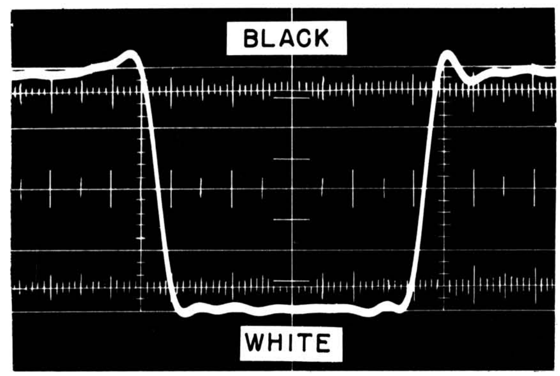
Black-to-white rise time = 0.20 μsec.
White-to-black rise time = 0.17 μsec.

(b) Without bridged-T network.

Fig. 18 – Effect of compensating for the i-f amplifier delay distortion by means of a bridged-T all-pass network in the video amplifier.

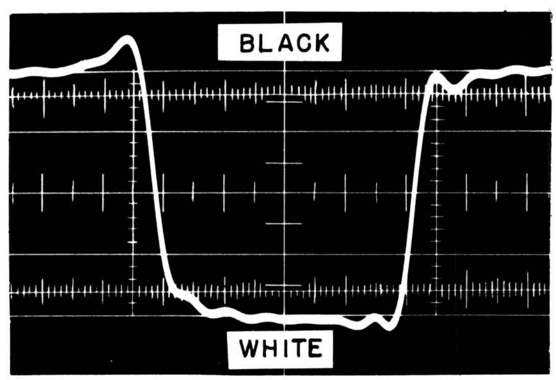


(a) - Video amplifier and sync take-off.



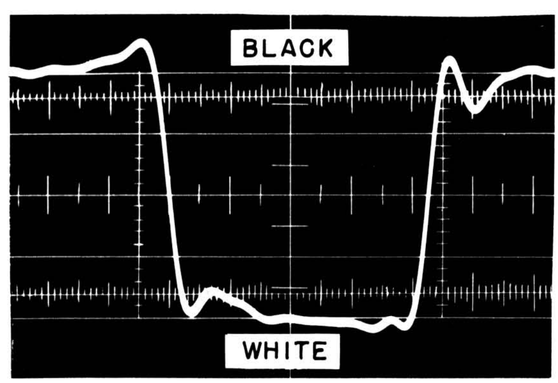
Black-to-white rise time = 0.14 μsec.
White-to-black rise time = 0.14 μsec.

(b) Sync separator disconnected.



Black-to-white rise time = 0.22 μsec.
White-to-black rise time = 0.16 μsec.

(c) Sync take-off at A.

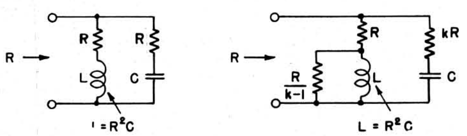


Black-to-white rise time = 0.15 μsec.
White-to-black rise time = 0.15 μsec.

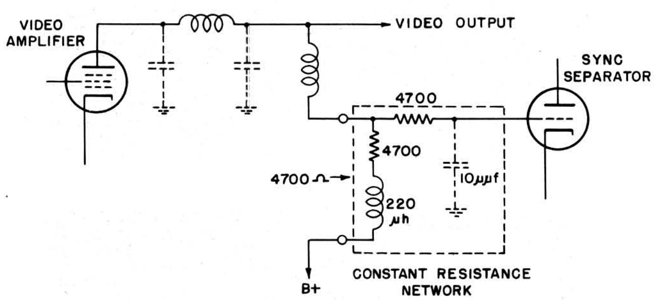
(d) Sync take-off at B.

Fig. 19 - The effect of coupling a sync separator (amplifier) to a video amplifier is shown. The smear is reduced when the sync input is connected to the junction of the peaking coils as shown in (c).

Distortion due to sync-separator or sync-amplifier coupling can be eliminated completely by making use of the classic constant resistance circuit shown in Fig. 20a. A more general form of this circuit which enables an increase in the value of the resistance in series with the sync separator input capacitance is shown in Fig. 20b. For each circuit the input is purely resistive over the complete frequency range and is equal to $\sqrt{L/C}$. This pure resistance is then used in place of the single load resistor normally employed. These circuits give excellent results and are relatively non-critical of tolerances in components.



(a) - Constant resistance networks.



(b) - Application to sync takeoff.

Fig. 20 - The use of a constant-resistance network to eliminate smear caused by the sync separator being coupled to the video amplifier.

Measurement Techniques

A modulated signal generator, wide-band oscilloscope, and phase-delay measuring equipment are essential in obtaining optimum receiver transient response. A low-capacitance probe is helpful in making it possible to examine the transient response at any point in the receiver without adding more than 2 μμf.

Signal Generator

The signal source for measuring the high-frequency electrical fidelity of a television receiver ideally should be a miniature television transmitter including the vestigial sideband filter and the associated video compensation networks. Almost invariably, however, a symmetrical double-sideband signal is used in testing television receivers. In addition to greatly simplifying the signal source, the use of a double-sideband generator simplifies monitoring of the signal since a diode video detector across the output terminals of the generator can be used.

The use of a double-sideband signal in receiver testing gives valid results under normal conditions where the receiver has negligible response to the sideband components which are normally absent in a vestigial sideband signal. Most i-f amplifiers satisfy this requirement, as has been observed by the similarity of response to a double-sideband signal and to a properly-compensated vestigial-sideband signal. An exception to this general rule occurs when the receiver does not sufficiently attenuate the unwanted sideband, possibly because of a rise in response beyond the adjacent sound-carrier trap frequency. In spite of this, however, use of the double-sideband signal is indicated in all instances where a vestigial sideband signal of *known* characteristics is not available.

The response of a normal receiver to a vestigial-sideband signal will be the same as the response to a double-sideband signal provided that the transmitter is properly compensated for the phase distortion introduced by the vestigial sideband filter. Proper compensation in this sense consists of compensating the transmitter (modulator) so that the phase delay is constant over the transmission range. This can be done through the use of all-pass non-minimum phase-shift networks (for example, Fig. 17) in the video amplifier of the transmitter modulator, so that the phase can be controlled independently of the amplitude.

Phase Delay

A set-up suitable for measuring phase delay is shown in Fig. 21. Here the delay through the receiver or video amplifier is matched against the delay of a calibrated continuously adjustable delay line. A scope with symmetrical horizontal and vertical amplifiers is used. This method is rapid and yields phase delay measurements accurate to 0.01 μsec. The data can be converted into envelope delay,

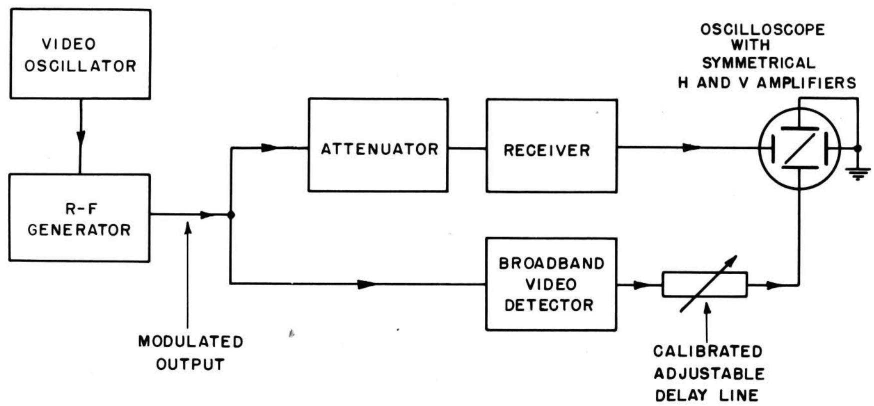


Fig. 21 – A set-up for measuring the phase delay of a receiver. The calibrated delay line is adjusted to equalize the delays as indicated on the diagonal line.

for comparison with data specified in terms of envelope delay, by adding to each measured value a quantity equal to the angular frequency multiplied by the slope of the phase delay at that frequency. This is considerably more accurate than measuring the slope of the phase angle φ vs frequency ω curve.

Standard Monitor

Ultimately, obtaining a uniformly high standard of receiver transient response depends upon industry agreement on a standard monitor. Variations in the phase and amplitude response of the transmitted signal with changes in transmitter adjustment make it highly essential that a standard monitor be available for judging the quality of the transmitted signal.

Two approaches are possible in tying down the performance of the standard monitor. One is to consider the monitor as a black box and to specify its performance on a step-modulated standard double-sideband signal, i.e., its preshoot, its rise time, overshoot, etc. The other approach is to specify in detail the amplitude and phase response of the monitor. Neither one of these approaches by itself is entirely satisfactory. A combination of the two is desirable so that its performance on a double-sideband signal can be relied upon to provide an accurate indication of the adjustment of a vestigial sideband transmitter.

Widespread use of a standard monitor should go far toward eliminating the present tendency for receiver designers to vary their designs to meet the peculiarities of particular transmitters.

Bernard Harris

Bernard Harris

John S. Horvath

John S. Horvath

Jack Avins

Jack Avins