Color TV Circuit Operation Guide

Revised and Expanded

An in-depth analysis of existing tube and solid-state color circuitry that begins with a review of how the color signal is developed and proceeds stage by stage through each section of the color receiver.

Edited by the publishers of **Electronic Servicing** *magazine*

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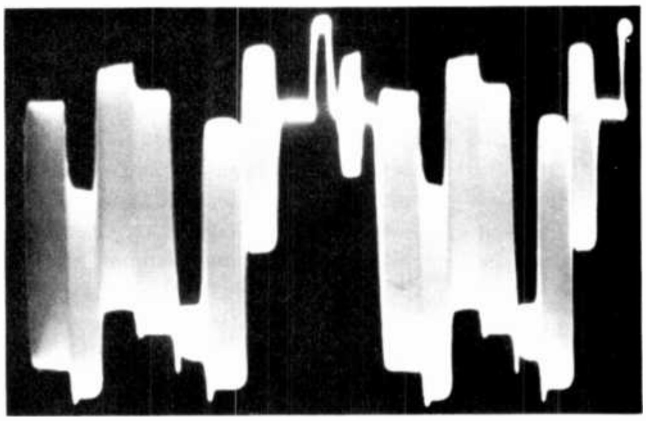


Chapter 1

Developing the color signal

The video signal from a color-TV transmitter actually consists of two separate signals. These are the luminance and chrominance signals. To properly understand the color-TV transmission system, we must study the make-up of these signals at the transmitter.

The scene to be televised is viewed by a tricolor camera which receives white light from the scene and produces three signals representative of the red, green, and blue components of that white light. Fig. 1 shows the basic components of the tricolor camera.

The white light from the scene is focused by the optical system and is directed onto the dichroic mirror at A. This type of mirror is designed and manufactured to pass all light frequencies of the spectrum except those of one particular primary color. Thus the mirror at A reflects blue light and passes all the other light. The beam of blue light is directed onto a front-surface mirror at B and thence onto the photosurface of the blue camera tube.

The light passing through the mirror at A strikes another dichroic mirror at C, which is designed to reflect red light. The red beam is directed to a front-surface mirror at D and thence to the red camera tube. The light beam passing through mirror C is devoid of blue and red and contains only green light; thus it goes directly to the green camera tube. We can see that the light from the scene being televised has been broken up into the three primary colors.

The three camera tubes and their associated amplifiers develop three voltages which are representative of the three colors. These voltages we will designate as E_R for red, E_G for green, and E_B for blue. We will go on to see how the luminance and chrominance signals are formed from these three voltages.

The Luminance Signal

In order for the color signal to be compatible with black-and-white receivers, one portion of the signal must represent the televised scene in terms of brightness only and must be very similar to the video signal specified for monochrome transmission. This signal is called the luminance signal.

When the specifications for this luminance signal were being drawn up, the response of the human eye to light of different frequencies was considered. The human eye does not see all colors with equal brightness. We can best illustrate this with three projectors for red, green, and blue beams. If these projectors are adjusted so that their outputs are equal, as measured by photoelectric instruments, white light will be produced when the three beams are superimposed upon each other. When viewed separately, the green light will appear to the average person to be almost twice as bright as the red light and from five to six times as bright as the blue light. We can see that the human eye is most sensitive to green, less sensitive to red, and least sensitive to blue.

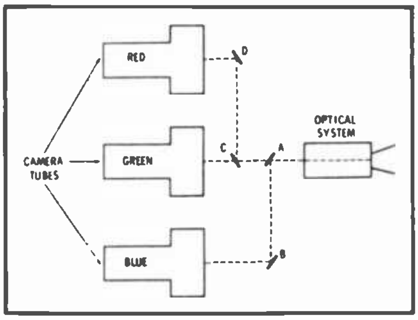


Fig. 1. Arrangement of basic components of a tricolor camera.

This varying sensitivity of the eye was considered when the luminance signal was designed. The color signals from the camera are combined in definite proportions to form the luminance signal: 59 per cent of the total green signal, 30 percent of the total red signal, and 11 per cent of the total blue signal.

We frequently call the luminance signal by the term, Y signal, and express its signal voltage as E_Y. The equation for expressing the content of E_Y can be derived from the percentages given previously and is written as:

$$E_Y = .59\,E_G + .30\,E_R + .11\,E_B \quad (1)$$

where,

E_G = the voltage of the green signal,

E_R = the voltage of the red signal, and

E_B = the voltage of the blue signal.

The actual formation of a luminance signal can be understood by considering Fig. 2. Assume the scene to be televised consists of a card having four vertical bars. These are, from left to right: white, red, green, and blue. We adjust the camera so that each of the color signals from the camera is one volt when the camera is focused on the white bar. As equation 1 states, the luminance, or Y signal, will also be one volt at this time. A monochrome receiver tuned to this signal would produce a bright white bar on its screen.

When the camera is shifted to scan the red bar, the red signal remains at one volt and the blue and green signals drop to zero. If E_B and E_G in equation 1 are zero, E_Y must equal .30 volts. A monochrome receiver will produce a gray bar with this signal.

Scanning of the green bar will produce a green signal of one volt and zero for the red and blue signals. The luminance or Y signal will have a value of .59 volts and will produce a very light gray bar on the monochrome receiver. Scanning of the blue bar will cause the blue signal to be one volt and the red and green signals to be zero. The Y signal will be .11 volts and the monochrome receiver will produce a dark gray bar.

The Chrominance Signal

In addition to the luminance signal which conveys the black-and-white information, we must transmit a signal conveying the color information. This is called the chrominance signal and, since it conveys only color information, we must remove the black-and-white or luminance information from it. We can do this by subtracting the luminance voltage, E_Y, from each of the three color signals produced by the color camera. Fig. 2 shows how this is done. The polarity of the luminance signal is inverted and combined with each of the three camera signals. The result is three signals representing red minus luminance, blue minus luminance, and green minus luminance. These are the important *color-difference signals* and are designated as: $E_R - E_Y$, $E_B - E_Y$, and $E_G - E_Y$. These are often shortened to R−Y, B−Y, and G−Y; but it should be remembered that these represent voltages and not colors.

The term E_Y in each of the color-difference signals has the value given in equation 1. If we substitute this value, we obtain three equations for the color-difference signals in terms of the three primary colors.

$$E_R - E_Y = E_R - (.30\,E_R + .59\,E_G + .11\,E_B)$$

$$E_R - E_Y = E_R - .30\,E_R - .59\,E_G - .11\,E_B$$

$$E_R - E_Y = .70\,E_R - .59\,E_G - .11\,E_B \quad (2)$$

Similarly

$$E_G - E_Y = .41\,E_G - .30\,E_R - .11\,E_B \quad (3)$$

$$E_B - E_Y = .89\,E_B - .59\,E_G - .30\,E_R \quad (4)$$

A further understanding of these color-difference signal equations can be derived from Fig. 2. If the red bar is being scanned, the R−Y matrix receives one volt of red signal. The luminance signal is .30 volt at this time (from equation 1). The polarity inverter changes this value to −.30 volt which is also applied to the R−Y matrix. The two voltages applied to the R−Y matrix form the R−Y signal: 1.0 volt minus .30 volt or .70 volt. Note that this agrees with the coefficient of E_R in equation 2.

The luminance signal is .59 volt when the green bar is being scanned and the red signal is at zero. The R−Y signal becomes −.59 volt. This agrees with the coefficient of E_G in equation 2. During the scanning of the blue bar, the signals to the R−Y matrix are zero and −.11 volt; the resultant R−Y signal becomes −.11 volt. This is the coefficient of E_B in equation 2. The voltages of the color signals in equations 3 and 4 can be derived in a similar manner.

The voltages of the color-difference signals all become zero when the white bar is scanned and no chrominance signal is developed. This is as it should be, because only a luminance signal is needed for black-and-white information. This is also true for any shade of gray.

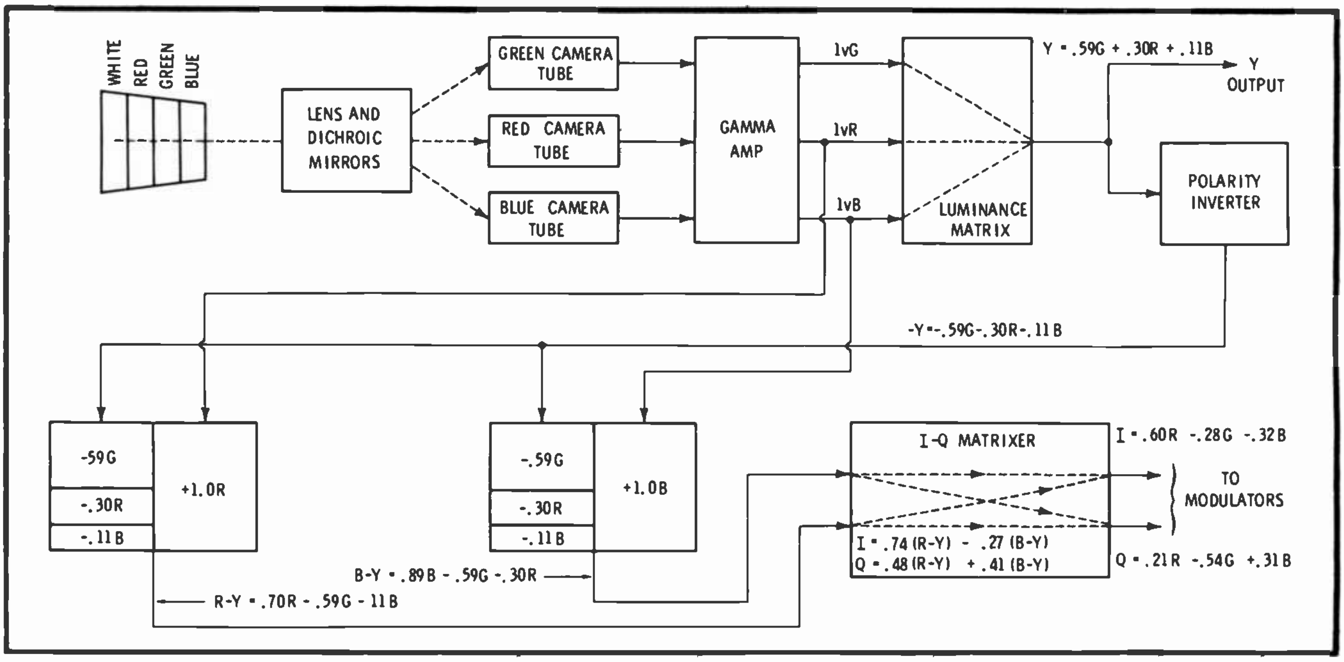


Fig. 2. Developing the luminance and color-difference signals in the transmitter.

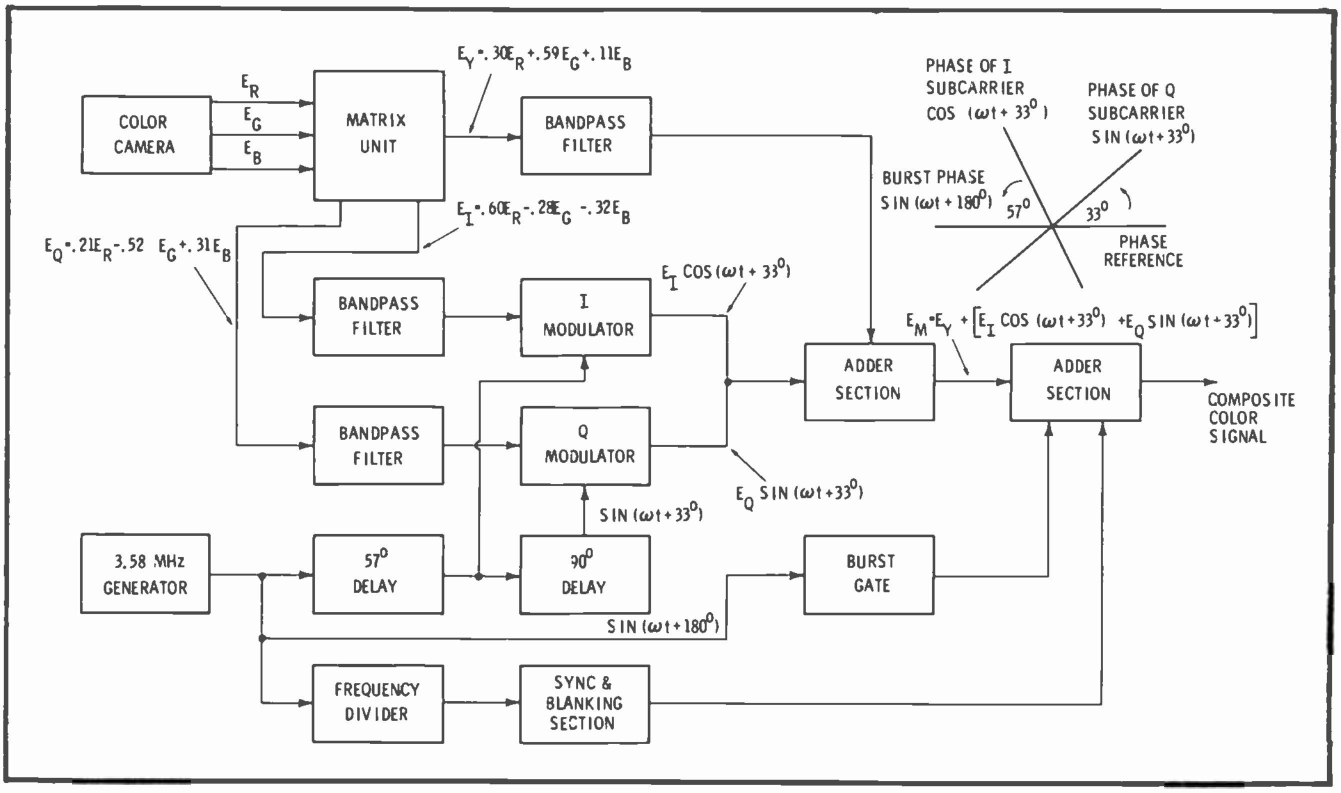


Fig. 3. Matrix unit in transmitter mixes three camera signals to form luminance, I, and Q signals.

Assume that the brightness of the white bar has been reduced to 50 per cent. This means that the camera tube output signals would be reduced to .50 volt. If each coefficient in equation 1 is reduced by 50 per cent, the luminance signal voltage would equal .15 + .295 + .055 or .50 volt; and the color-difference voltages would still be zero. The only signal would be the luminance signal representing the bar but its amplitude would be only 50 per cent. The monochrome receiver would reproduce a gray bar halfway between white and black.

For the sake of economy, the designers of our color-TV system decided that only two signals could be used to transmit the three color-difference signals. It was found that a signal representative of $E_G - E_Y$ could be formed in a receiver by combining suitable proportions of $E_R - E_Y$ and $E_B - E_Y$; therefore there was no need to transmit $E_G - E_Y$. The proportions of $-.51$ $(E_R - E_Y)$ and $-.19$ $(E_B - E_Y)$ will produce a signal equivalent to $E_G - E_Y$.

In the actual transmission of color, two signals are used to convey color information. These are the I and Q signals and they are formed by combining specific proportions of $E_R - E_Y$ and $E_B - E_Y$, as shown in Fig. 2. It was found that better reproduction of color could be obtained with this system.

By combining .74 $(E_R - E_Y)$ and $-.27$ $(E_B - E_Y)$, the I signal is formed. The Q signal is composed of .48 $(E_R - E_Y)$ and .41 $(E_B - E_Y)$. If we subtract the value of E_Y (equation 1), we obtain E_I and E_Q in terms of the three color camera signals.

$$E_I = .60\,E_R - .28\,E_G - .32\,E_B \quad (5)$$
$$E_Q = .21\,E_R - .52\,E_G + .31\,E_B \quad (6)$$

The matrix unit in the transmitter forms the luminance, I, and Q signals by mixing the three camera signals as can be seen in Fig. 3. The luminance signal passes through a bandpass filter and is fed into the adder section. The I and Q signals also pass through bandpass filters but then are fed to the modulators. The I signal modulates a subcarrier with a phase of cos $(\omega t + 33°)$; the Q signal modulates a subcarrier with a phase of sin $(\omega t + 33°)$. The phase reference ωt is the phase of the color burst plus 180°. The phase angles between the two subcarriers and between them and the color burst are also shown in Fig. 3. The I subcarrier is leading the Q subcarrier by 90° and lags the color burst by 57°.

A very close tolerance is required for these phase differences. This is accomplished by using a common source at 3.58 MHz for all signals including the synchronization.

The chrominance signal is formed by the outputs of the I and Q modulators, and is then merged with the luminance, sync, and blanking signals to form the composite color signal ready for transmission.

The Composite Color Signal

The original 4.25-MHz bandwidth of the monochrome (luminance) signal had to be retained in the color TV system in order to meet the requirements of compatibility. The chrominance signal also had to be included within the alloted 6-MHz channel. Meeting these requirements posed quite a problem to the system designers but was accomplished by placing the chrominance signal at the proper place within the band of video frequencies and by limitation of its bandwidth.

An interleaving process makes possible the inclusion of both the chrominance and luminance signals within a 4.25-MHz video band. The energy of the luminance signal, as with any signal, concentrates at definite intervals in the frequency spectrum. Consider an amplitude modulated signal of 10 MHz. The sidebands of the signal would exist relatively close on either side of 10

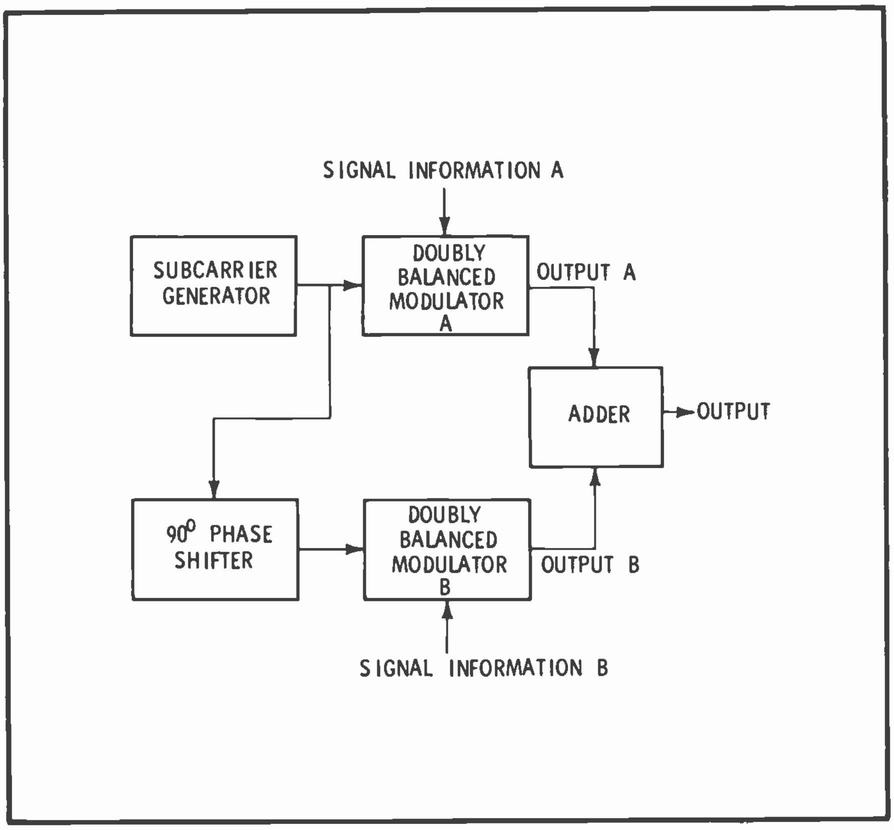


Fig. 4. Divided-carrier system modulates subcarrier with I and Q signals.

MHz. There would be no energy existing between the upper sideband and the edge of the lower sideband of the second harmonic at 20 MHz. It is in these blank areas that the chrominance signal energy is concentrated. There is no objectionable interference between the luminance and chrominance signals.

The luminance signal is transmitted by conventional amplitude modulation of the video carrier. The chrominance is transmitted by means of a subcarrier, the frequency of which was chosen to interleave properly with the luminance signal. This subcarrier frequency (3.579545 MHz) is high in the video band so that its sidebands, when they are limited in bandwidth, do not interfere with monochrome reproduction of the luminance signal.

The subcarrier is modulated by a process known as divided-carrier modulation since two different signals (I and Q) must be placed on the same subcarrier. Fig. 4 illustrates the basic principle of the system. A generator supplies the subcarrier frequency of 3.58 MHz which is applied to modulator A. The same signal is shifted in phase by 90° and applied to modulator B. The schematic in Fig. 5 represents either modulator. The modulating signal (I or Q) is split in phase and applied to the control grids of the two tubes. The subcarrier is also split in phase by the transformer and applied to the suppressor grids of the two tubes.

Tube No. 1 increases in conduction when the signal at its control grid goes positive and a subcarrier signal of increased amplitude is produced. Tube No. 2 receives a negative signal at the same time and its conduction and output signal decrease. On the next half cycle of the input signal, these conditions are reversed.

The plates of the two tubes are tied together so the waveforms shown at the plates cannot exist separately but are actually combined. The signals at the plates can be seen to be 180° out of phase, and the actual output sidebands will consist of a combination of the two plate signals. Because they are out of phase, a cancellation takes place, and the output amplitude becomes the difference in the amplitudes of the two signals. Additionally, the phase of the output is the phase of the largest signal.

A change in the amplitude of the input signal causes amplitude change of the output signal. If the phase of the input signal changes, the phase of the output signal will change 180°. The output signal therefore represents the input signal but in terms of the phase and amplitude of the subcarrier frequency.

Modulator B in Fig. 4 operates in an identical manner except that the subcarrier input is delayed by 90°. Thus, the output of modulator B is 90° from that of A. This can be either a lead or lag depending upon the polarities of the modulating signals.

The output signals from the modulators are combined in the adder stage, Fig. 4, and become a single waveform varying in amplitude and phase in accordance with the amplitudes and phases of the I and Q signals. Two signals have been placed upon a single subcarrier and can be recovered by reversing the modulation process in the receiver.

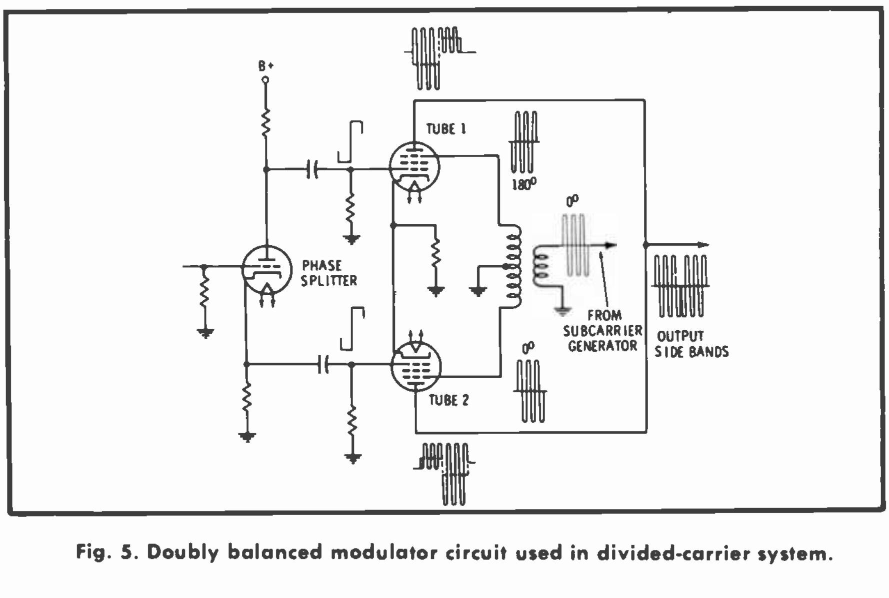


Fig. 5. Doubly balanced modulator circuit used in divided-carrier system.

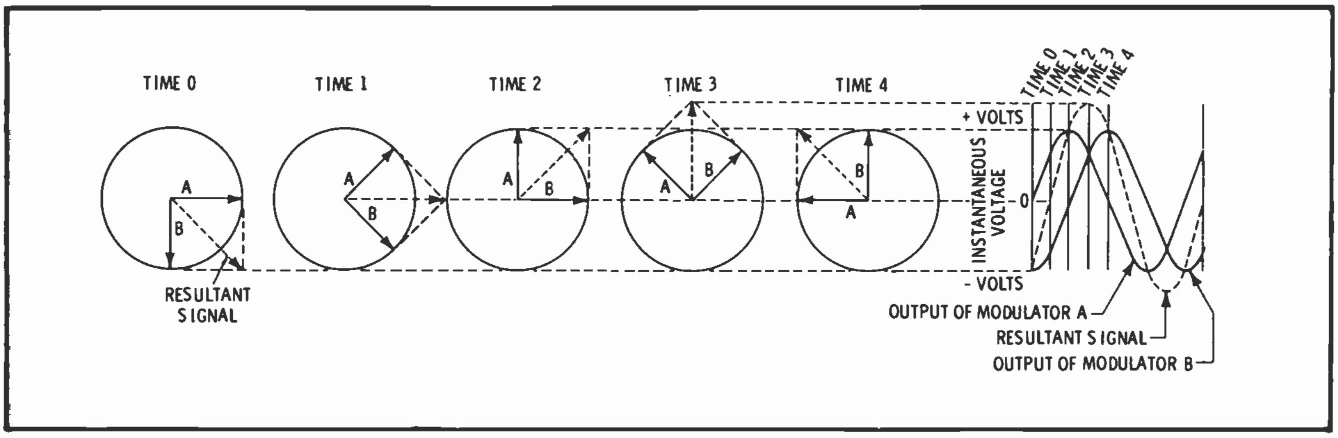


Fig. 6. Deriving a resultant waveform by use of vectors.

So far, we have considered only simple, unchanging waveforms applied to the modulators. In actual practice, the waveforms are quite complex because the picture may be composed of many different hues and may be constantly changing as well. Consequently, it is necessary to consider the modulation characteristics in terms of vectors rather than in terms of sine waves. Referring to Fig. 6, we have two vectors which are displaced from each other by 90° just as the I and Q axes are separated by this amount in a color transmitter. The first pair of vectors are shown at zero time. The vector which represents the instantaneous value of A leads vector B by 90° (Vector motion is arbitrarily in the counterclockwise direction.) At this point in time, the value of A is zero while B is maximum negative. The resultant, or vector, sum of the two voltages is equal to the value of B.

If we cause the vectors to rotate 45°, the magnitudes of A and B will be equal in value (70.7% of maximum) but opposite in polarity. In the sine wave diagram, the resultant is shown to be zero. At time 2, A is maximum positive while B has decreased to zero and the resultant is equal to A. We see that at time 3 the resultant has reached its positive maximum and both A and B are 70.7% of their maxima. At time 4, the value of the resultant is the same as it was at time 2, but the amplitudes of A and B have been interchanged. It is possible to show the value of the resultant for every value of A and B by simply performing the calculations for every degree of rotation of the two vectors. However, the above illustration should prove adequate for our purposes.

The voltages from modulators A and B of Fig. 4 are combined in the adder to form a single waveform which varies in phase and amplitude in accordance with the amplitudes and phases of the two signals (I and Q axes) which produced it. Thus, the two modulating signals are able to modulate a single carrier in such a manner that both signals may be recovered separately in the receiver. As we shall see later on, it is not even necessary that the demodulation be accomplished on the same axes as the ones on which the modulation was impressed. In fact, receivers in use today demodulate on I and Q, X and Z, R − Y and B − Y axes, and even on three axes (R − Y, B − Y, and G − Y).

Video Spectrum

The discoveries of Pierre Metz and Frank Gray were the key to the problem of inserting the color information into what seems to be the already-crowded 6-MHz TV channel. Their studies of the scanning process used in telephotography and television proved that the energy in the video spectrum is concentrated at certain discrete points in the bandwidth. These points are situated at frequencies which are whole multiples of the scanning rate or frequency. The actual proof which they developed is far too complicated for inclusion in this series, but the following illustration should help to understand the principles involved.

It can be proven mathematically that any video signal is the combination of a number of pure sine waves which are multiples or harmonics of a fundamental frequency. Thus, a video waveform consists of a series of sine waves all of which are multiples of the horizontal scanning frequency. There is a second series of frequencies to be considered, the vertical scanning frequency and its harmonics. This energy is concentrated in "sidebands" of each multiple of the horizontal scanning frequency.

From the foregoing, it is apparent that there is actually a considerable portion of the video spectrum of the TV channel which is not used in passing the monochrome information. Further analysis shows that these "holes" in the spectrum occur at the odd harmonics of one-half the horizontal scanning frequency. The color subcarrier frequency was carefully chosen to occupy one of these "holes."

In radio transmission and in monochrome TV, the modulation is recovered by a simple detection process. In color TV, however, the color subcarrier sidebands are made a part of the video modulation and must be processed further after video detection in order to recover the chroma information. Since the sidebands of the chroma subcarrier are a portion of the composite video signal, the frequency of the chroma subcarrier has to chosen not only for minimum degradation of the existing monochrome information but also so that it will not add undue interference to the picture. In addition, it must be chosen so that the chroma information which is modulated on it will not be outside the receiver bandpass.

By keeping the chroma subcarrier frequency as high as possible, the interference will be held to a minimum. In this manner, the reduced response near the upper limit of the video passband of most monochrome receivers will reduce the amplitude of the interference which reaches the picture tube, and the relatively fine "grain" produced by a high frequency is less objectionable. On the other hand, the subcarrier frequency must be low enough so that the upper sideband of the chroma information will pass through a color receiver.

Experimentation proves that it is not necessary for the pure chroma information to have frequency components which extend as high as the components of the monochrome signal. Stated another way, the R, B, and G signals do not require rise times which are as steep as the rise time of the Y signal. While luminance information must have frequency components above 3MHz to produce a pleasing picture, there will be no noticeable decrease in quality of a color picture if the upper limit of the chroma information is only 600kHz. Thus, the chroma subcarrier frequency should be approximately 3.6MHz if we consider that the practical video bandwidth for color transmitters and receivers is approximately 4.2MHz.

It has been shown that the signal energy which is produced by scanning an image is concentrated around the harmonics of the scanning frequencies. By the same reasoning, we find that the energy contained in the chroma subcarrier sidebands appears at frequencies which are the sum and difference of the subcarrier and the line and frame frequencies. Therefore, a frequency which is an odd multiple of one-half the horizontal frequency must be used for the chroma subcarrier so that the "interleaving" characteristic may be utilized. This subcarrier may be tentatively established by the following formula:

$$f_s = \frac{15{,}750X455}{2} = 3{,}583125\text{Hz}$$

This frequency was not adopted because of an objectionable feature which became apparent. Monochrome receivers which employ an intercarrier sound system develop a 4.5-MHz signal at the output of the video detector. When this 4.5-MHz signal beats with the color subcarrier, a difference frequency of approximately 900kHz results and this produces an objectionable pattern on the screen. When the beat frequency is an odd multiple of one-half of the horizontal scanning frequency, it is least objectionable.

Since it would have been impractical to change the 4.5-MHz intercarrier frequency, another method had to be found. This was done by changing very slightly the horizontal scanning frequency for color. This new scanning frequency was computed as follows:

$$f = \frac{4.5 \times 10^6}{286} = 15.734.264 \text{ Hz}$$

Thus, the 286th harmonic of the new line frequency equals the 4.5-MHz intercarrier frequency.

Since each frame must consist of 525 lines, the color field frequency has been changed to 59.94 Hz. Finally, the color subcarrier frequency becomes 3.579545 MHz.

The new scanning frequencies are slightly below those which were chosen for black-and-white transmission; however, the changes amount to less than 1%. These are within the tolerances that are allowed and do not affect the operation of black-and-white receivers when receiving a color signal. For the purpose of maintaining close synchronization of color receivers, the tolerance for the subcarrier frequency is held to ±.0003%, or about ±10Hz; and the rate of change may not exceed 1/10Hz per second. The same percentage of tolerance also applies to the line and frame frequencies, and it is standard practice to develop all these frequencies from a common source.

Now let's see how all this theory works out on a real, live TV set. Since the chroma signal falls within the bandpass of the video amplifier, a dot pattern is actually produced on the screen of a receiver which is tuned to a colorcast. However, the interference is not visible because of a phenomenon known as "cancellation effect." This cancellation effect is actually nothing more than a way of taking advantage of a characteristic of the eye which has been known for a long time. The eye has the ability to actually integrate a series of visual signals and, consequently, it can detect a signal of very low intensity if it recurrs at corresponding points in each of a group of adjacent traces on a scope. This was discovered in radar technology where it was found that an observer could detect a target whose amplitude was actually less than the amplitude of the receiver background noise or "grass."

Since it has been established that all of the sine wave components of the luminance signal are harmonics of the line and field frequencies, these components may be considered to be in phase during successive lines or frames. Thus the signals on adjacent lines are additive and tend to reinforce each other. The eye tends to integrate these successive signals and consequently it "sees" not individual lines and frames, but a composite of a number of them.

Conversely, the interference signal, which is the result of the 3.579545-MHz subcarrier, is applied to the picture tube in opposite polarity on adjacent scanning lines. The eye tends to reject this signal because it averages the light produced by the spots, and, of course, the average of the alternate dark and light spots is zero.

Summary

In this first chapter we have discussed the fundamental operation of the color camera and the method by which the three color signals from it are used to develop the luminance and chrominance signals. These have to be transmitted separately so that a black-and-white receiver can receive a color signal. The chrominance signals from the color camera are ultimately impressed on a subcarrier in such a manner that they may be recovered by the color receiver.

The chroma subcarrier sidebands can be transmitted in the same spectrum along with the luminance or Y information because the luminance signal energy is not evenly distributed in the spectrum, and so the two signals may be "interleaved" without interference.

To achieve compatibility, slight modifications had to be made in the horizontal and vertical scanning frequencies, but these are so slight that the normal tolerances of receivers will accept either scanning rates. ▲

Chapter 2

Tuner to video amplifier

A color receiver may be divided into sections for the purpose of study. One of these is composed of those circuits which are more or less similar to their counterparts in a b-w receiver. As we shall see, some of these circuits are almost identical to their b-w ancestors, while others have been modified extensively to meet the more demanding requirements of color reception. The remaining circuits of the color receiver have functions which are related specifically to color and have no b-w counterparts. This lesson will deal principally with the circuits in the first category. Fig. 1 is a functional block diagram of a color receiver. The circuits which are practically unchanged from their b-w counterparts are shown unshaded, those which have b-w counterparts but have been extensively modified are shaded, and those circuits which are peculiar to color are crosshatched.

Receiver Bandpass

Before discussing the specific circuits used in RF and video IF amplifiers of color receivers, let's consider the bandpass requirements of a color receiver as opposed to black-and-white. Present-day b-w receivers of good design usually have an overall response from the mixer to the video detector output that is 3 to 3.25 MHz wide at the 6dB points. This means simply that a video signal component at 3 MHz would produce only one-half the output voltage at the detector as one having a frequency of perhaps 2 MHz.

Typical color receivers of modern design have responses which are 3.58 to 4.0 MHz wide at the −6dB points. This is necessary to allow the color burst to pass through the IF strip and, more important, to allow the upper sideband of the color subcarrier to pass. From the first lesson, you will recall that the color burst has a frequency of 3.58 MHz and that the upper sideband extends to about 4.2 MHz. While it may appear that these frequencies cannot pass the IF strip, we shall see later that the relative attenuation of these frequencies in the IF strip is compensated by the tuned circuits of the chroma bandpass amplifier.

Still another requirement of the IF and RF bandpass of a color receiver is the shape of the curve. Because of the tendency of the color subcarrier and the 4.5-MHz intercarrier audio signal to beat in the chroma circuits and produce interference in the picture, the output stage of the video IF must have excellent 4.5-MHz trapping. (The sound is recovered from a point preceding this trap.) It is also important that the slope of the IF response curve on each side of the lower −6dB point be constant so the overall response through the chroma amplifier will be smooth from 3.0 to 4.2 MHz.

RF Tuner

The function of the RF amplifier and mixer of a color receiver is the same as it is in a b-w receiver. It must tune to the 12 VHF channels and to the output of the UHF tuner with a reasonably constant noise figure across the band. However, the requirements for bandpass are more stringent, and the need for low noise is more pronounced in color, especially in fringe areas. Thus, a tuner which was entirely satisfactory for b-w reception might prove inadequate for color.

Although a tilt or sag in the response curve of a monochrome tuner might be compensated in the IF amplifiers, it is necessary to provide uniform response across the band for proper reception of color signals. A tuner having a response similar to Fig. 2 would produce excellent results in a color receiver. It is interesting to note that since color receivers have become so popular, some of the manufacturers are adapting their color tuners to their black-and-white sets. Two examples of this are the RCA tuner Model KRK118 and Zenith tuner Model 175-503.

With the exception of tube failures, both color and b-w tuners present only a moderate number of service problems. It is important

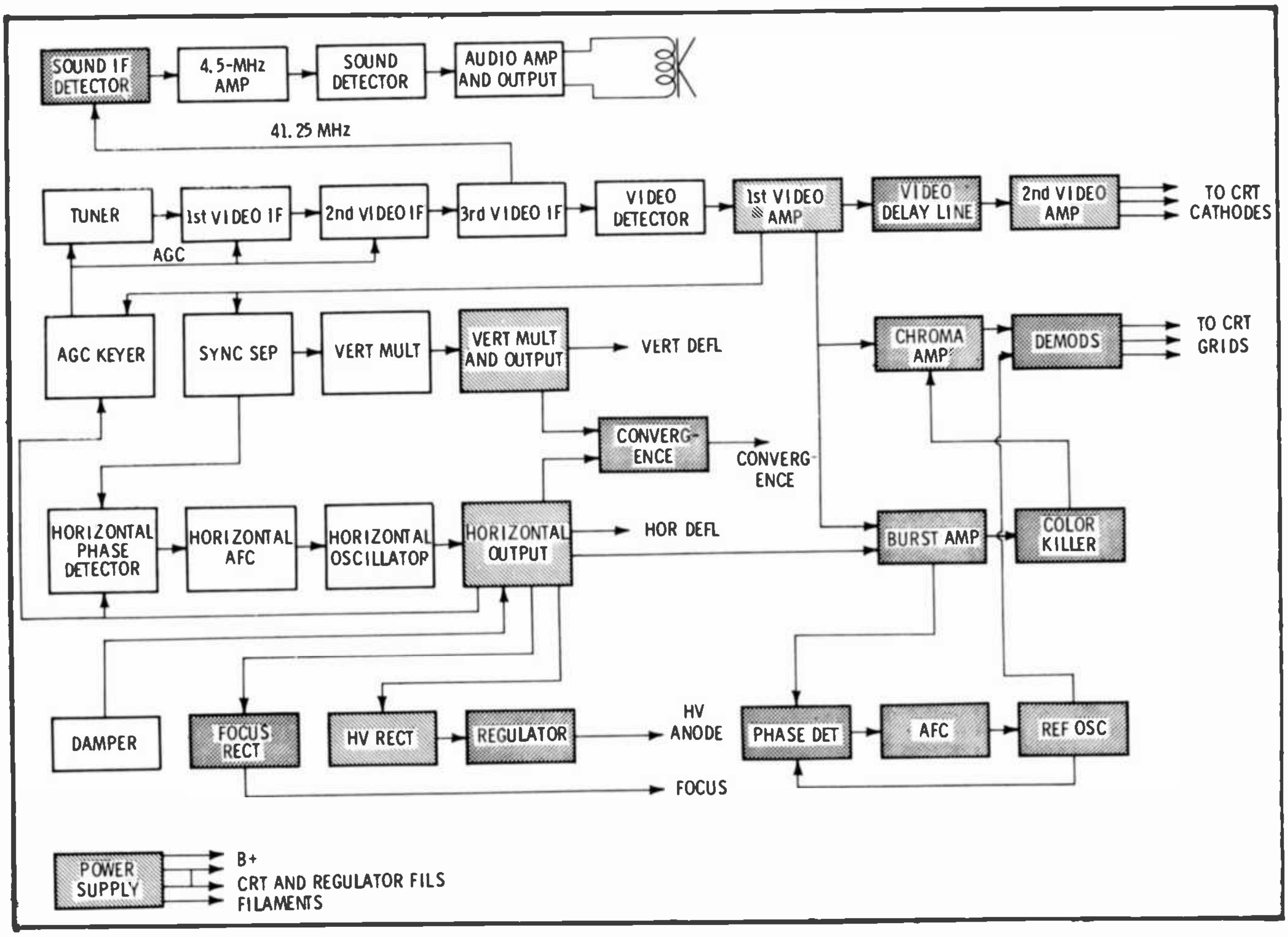


Fig. 1. Functional block diagram of a typical color receiver.

that the technician appreciate fully the necessity of accurate alignment and careful repair techniques when dealing with color tuners. It is possible to "repair" a tuner using slipshod methods or improper alignment procedures only to find that color reception has been seriously impaired even though b-w reception is reasonably good. Many of the compromises which have become common practice in b-w receivers cannot be made in tuners used in color receivers.

Video IF Amplifiers and Detector

Although the function of the IF amplifiers in a color receiver is essentially the same as for a monochrome set, the bandpass requirements are more stringent just as they were in the tuner. First of all, the purpose of the section is to amplify and select a certain band of frequencies as shown in Fig. 3. Notice that the sound carrier frequency is lower than the video carrier. This is due to the fact that the local oscillator in the tuner is tuned to a frequency above the channel and thus the sideband positions are reversed. The transmitted upper sideband becomes the IF lower sideband and vice versa.

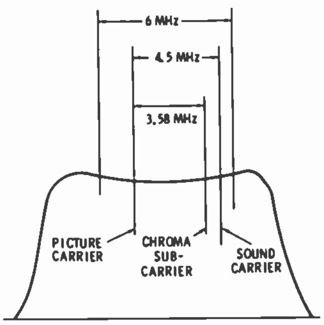


Fig. 2. Ideal tuner bandpass curve.

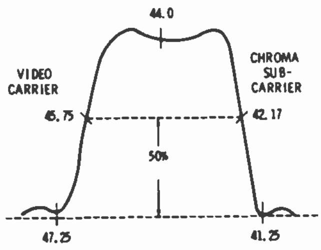


Fig. 3. Typical IF bandpass curve.

In the past dozen or more years, the video IF amplifier has evolved from a very complex circuit with six stages of amplification to contemporary designs with as few as two stages only slightly more complex than those used in present-day black-and-white receivers. The IF amplifier and video detector (RCA chassis CTC19) shown in Fig. 4 is a typical two-stage circuit.

The mixer plate coil, L202, and the coupling transformer, L2, are tuned to approximately the center of the passband, and C14 controls the bandwidth of the combination by changing the amount of coupling. When properly adjusted, these components produce a passband at the grid of V1 having −6dB points at 42.17 and 45.75 MHz. A2 is a sound trap which attenuates the sound carrier and produces a "notch" at 41.25 MHz. The combination of the tuned circuit, L1 and C13, and R19 form a second trap tuned to the sound carrier of the adjacent channel. Fig. 5 shows the actual bandpass curve seen at the plate of V1. At the time of observation, L3 was swamped and,

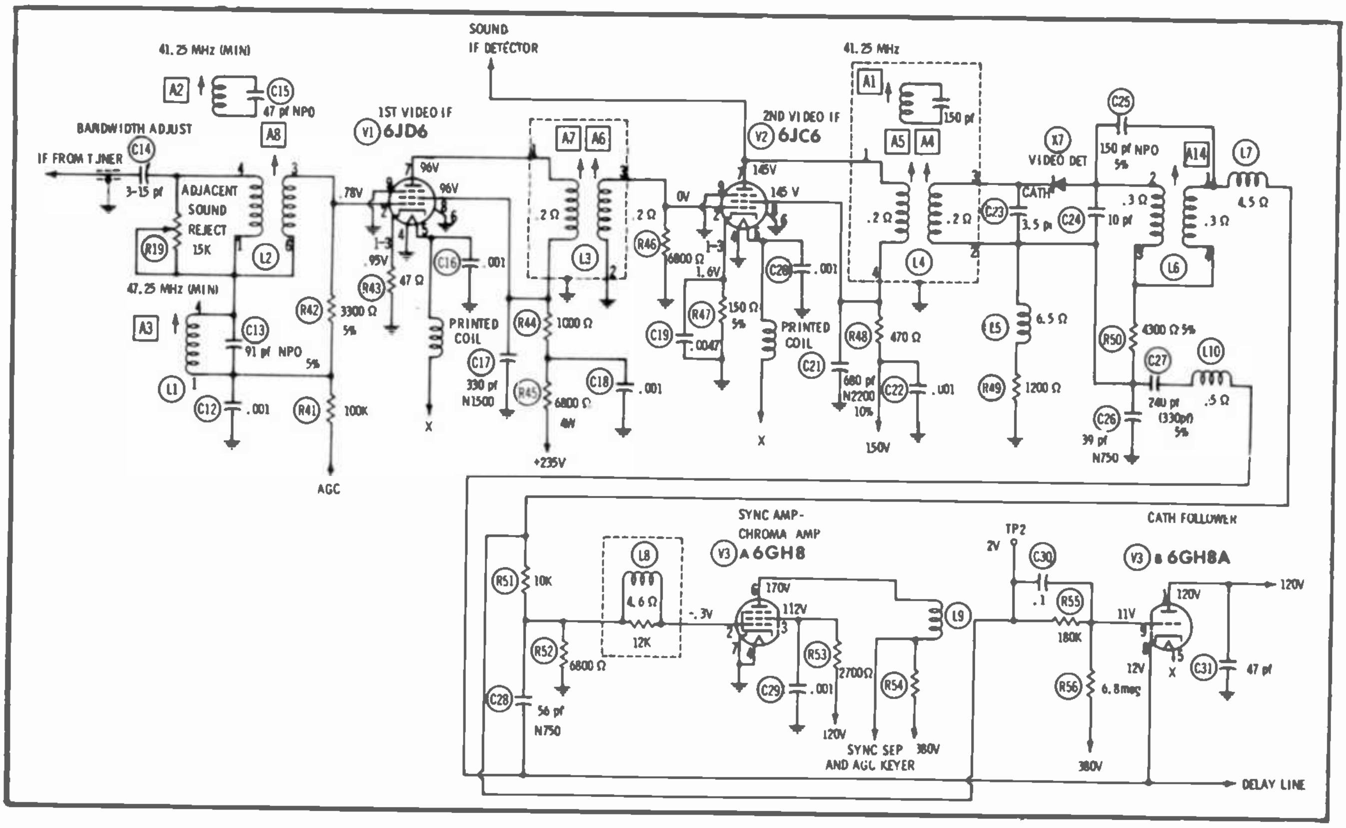


Fig. 4. IF amplifier and detector of RCA CTC 19 chassis.

therefore, had no effect on the curve.

L3 is the interstage coupling circuit between V1 and V2. It is a fairly conventional double-tuned circuit and is adjusted so that a symmetrical curve having a 3.58-MHz bandpass appears at the plate of V2 when the plate circuit is "swamped."

The output of V2 is developed across two separate loads. The shunt-fed sound takeoff coil, L14, is resonant at the sound IF (41.25 MHz). The audio must be recovered in front of the video detector because the sound IF freuency is trapped out before video detection. The remainder of the audio circuit will be discussed later.

The second plate load of V2 is L4 which consists of a double-tuned coupling transformer and a 41.25-MHz sound trap. The coupling transformer is overcoupled to increase its bandwidth and the two slugs are adjusted for symmetry and bandwidth so that the video output appears as shown in Fig. 6. Naturally, the sound trap (A1) is tuned to trap out the 41.25-MHz sound carrier.

A final 4.5-MHz sound trap, A14, is located in the output of the video detector. Notice that there is a total of three sound traps in the receiver. These are necessary to completely remove the 4.5-MHz beat from the video section, as even a minute signal will beat with the chroma subcarrier in the color demodulator and cause interference.

The IF section of Admiral's 1G1155-1 chassis (Fig. 7) is typical of the three-stage IF amplifier currently in use. The interstage circuit between the mixer plate and V1 is similar to the circuit employed in the RCA chassis except that it does not have a 41.25-MHz sound trap. L4, L6, and L7 comprise a staggered triple with the three transformers single-tuned to 45.75, 42.5, and 43.8 MHz, respectively Notice that the sound IF takeoff is from the last IF amplifier plate as it was in Fig. 4.

The input to the video detector is trapped at 41.25 MHz to attenuate the sound carrier, and a 4.5-MHz trap in the output of the detector further attenuates the sound. Although it is not shown in Fig. 7, there is one more 4.5-MHz trap which is located in the grid circuit of the first chroma bandpass amplifier. Thus, both the Admiral and RCA have the same number of traps, one adjacent-channel trap and three sound traps, although their locations in the circuit are different.

Notice that the first and second IF amplifiers are "stacked." That is, the plate supply of the first tube is taken from the cathode of the second. This circuit is used in many receivers, both color and monochrome, and has the following advantages: (1) Since the plate voltage does not have to be dropped by a bleeder, the load on the power supply is decreased. (2) The second IF amplifier acts as a voltage regulator for the first IF stage to improve its stability and signal-to-noise ratio. (3) The AGC voltage which is applied to the grid of the first IF amplifier is amplified and applied to the cathode of the second

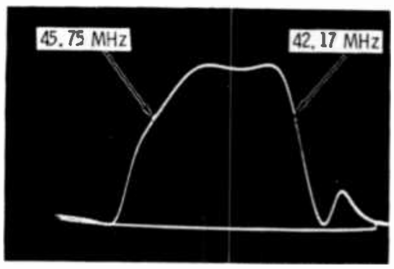


Fig. 5. Response curve at V1 plate.

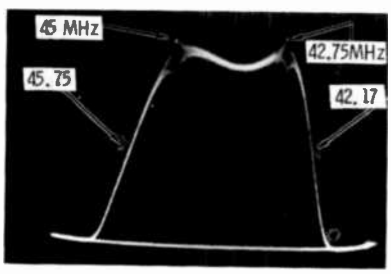


Fig. 6. Overall IF response curve.

stage to improve the overall AGC action.

Referring once more to Fig. 4, there are two outputs from the video detector. One of these drives a cathode follower and its output ultimately reaches the CRT cathodes as the Y, or luminance, signal. The second output from the detector drives a sync and chroma amplifier, and this amplifier, in turn, supplies signal to the sync and AGC circuits as well as to the chroma bandpass amplifier.

The video detector employed in the Admiral chassis (Fig. 7) has only one output which goes to the first video amplifier. Then the video signal is distributed to the sync, AGC, Y, and chroma circuits.

Sound IF and Audio

Except for the point of takeoff and the separate sound IF detector, the sound system of a color receiver is essentially the same as those in monochrome receivers. The sound takeoff point in a color receiver is normally the plate of the final video IF amplifier because one or more sound traps following this point severely attenuate the sound signal before it reaches the video detector. Fig. 8 shows the sound IF and audio section of Zenith Chassis 24MC32 .

The 4.5-MHz sound signal is developed by the sound and sync detector which is connected to the plate of the third IF amplifier. The detector output is amplified by V6A and the sync pulses are supplied to the AGC and sync circuits. The 4.5-MHz signal is amplified further in V6B and then detected by the quadrature detector, V8A. The remainder of the circuit is so familiar that no further explanation is warranted.

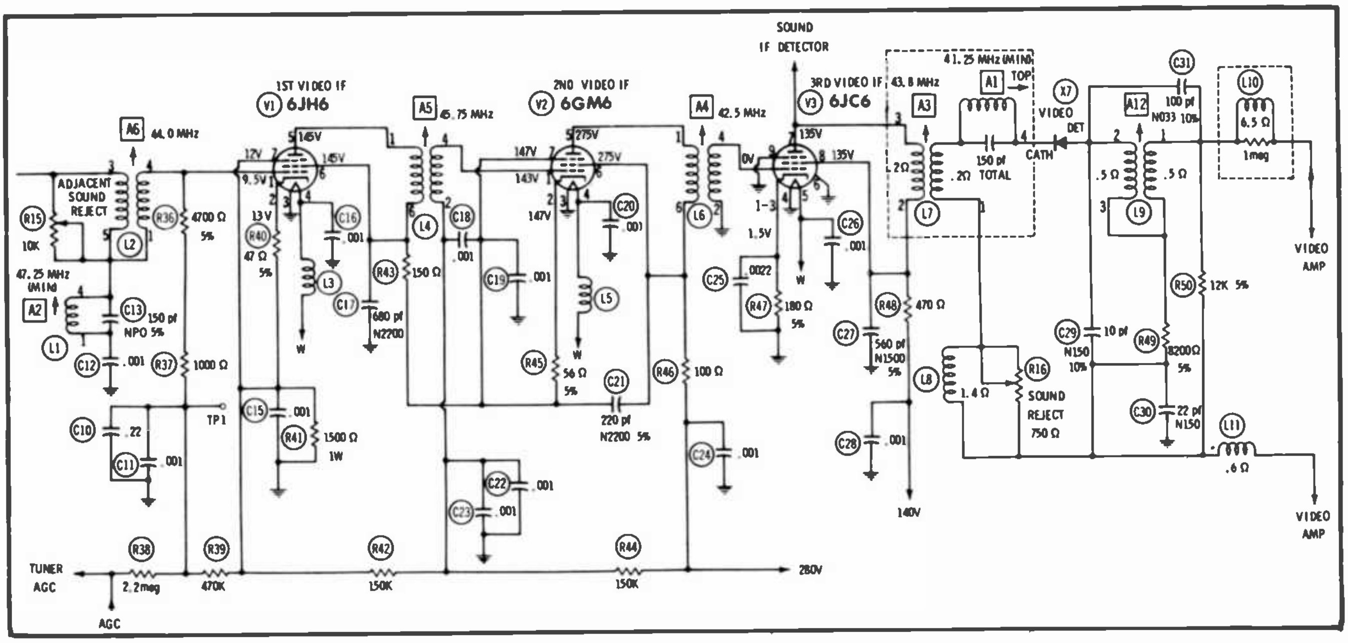


Fig. 7. IF amplifier and detector of Admiral 1G1155-1 chassis.

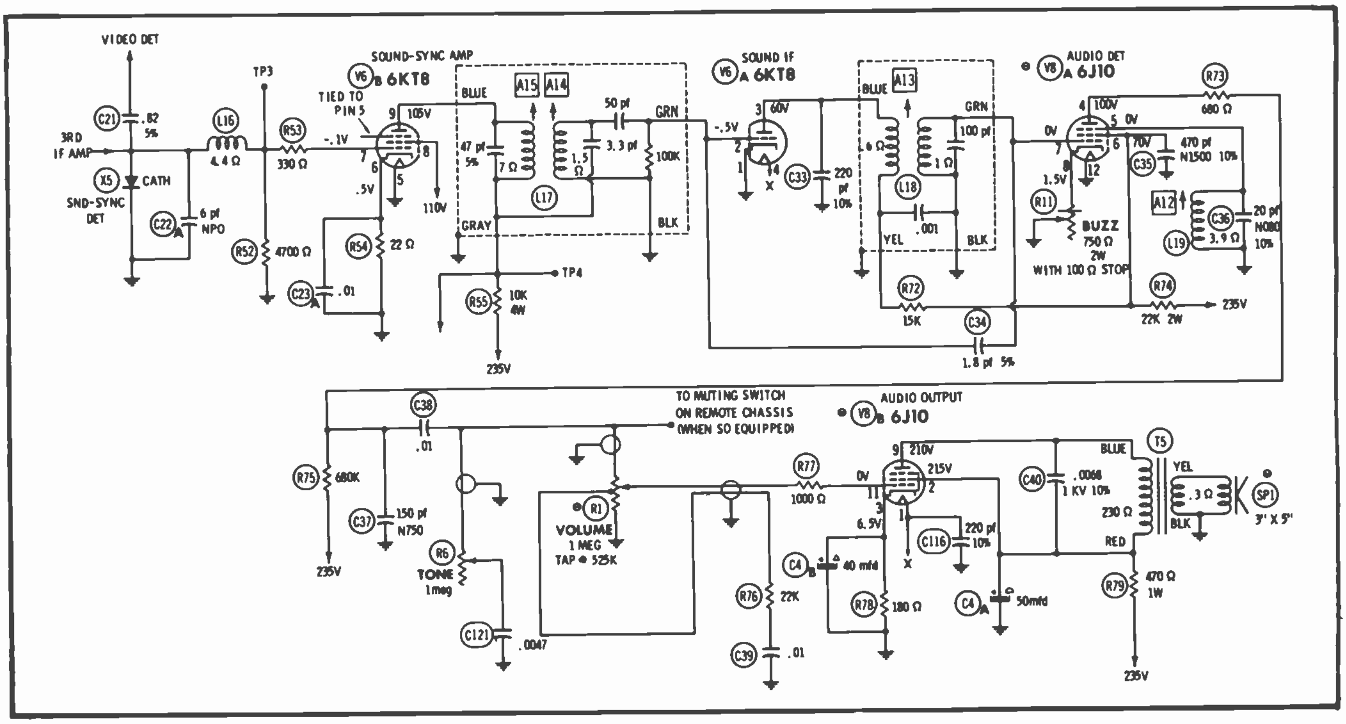


Fig. 8. Sound IF and audio section of Zenith 24MC32 chassis.

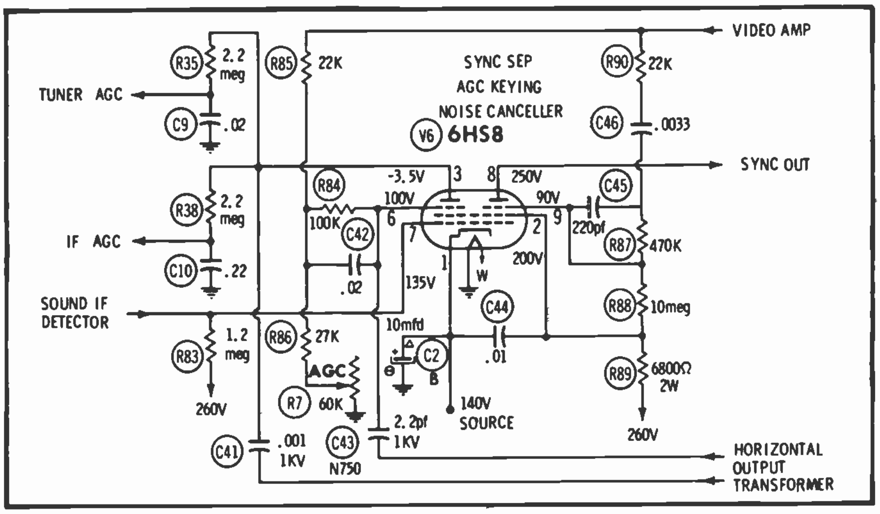


Fig. 9. AGC keyer, sync separator, and noise canceller of Admiral chassis.

AGC and Sync

Although it is conventional in operation, the AGC circuit is very important in color reception since variations in signal level in the video and chroma circuits will cause the color of the image, as well as the brightness, to vary.

The AGC keyer, sync separator, and noise canceller circuit of the Admiral Chassis 1G1155-1 is shown in Fig. 9. During retrace time, the plate (pin 3) is supplied with a positive pulse from the high-voltage transformer. At the same instant, a positive sync pulse is applied to the suppressor grid (pin 6). Also, at the same instant, a negative sync pulse is applied to the control grid; but its normal amplitude is insufficient to hold the tube in cutoff and we shall ignore it for the moment.

With both the plate and suppressor grid driven positive, V6 conducts and charges the top of C41 negative. C41 tends to discharge between pulses, but the time constant during discharge is so long that the voltage remains substantially constant between sync pulses. Since the amplitude of the pulse applied to the plate is constant, the amount of charge on C41 is determined by the amplitude of the sync pulses applied to the suppressor grid and, of course, the bias set by the AGC control. Thus an increase in the amplitude of the sync pulses causes a greater negative charge to accumulate on C41 and this decreases the gain of the receiver.

If a noise spike should appear on top of (coincident with) a horizontal sync pulse, it seems that the charge on C41 would be excessively negative. This would cause a radical increase in AGC voltage and result in varying contrast on the CRT. To preclude this, negative sync pulses are applied to the control grid of V6. If a noise spike appears on top of the sync pulse, the tube simply does not conduct and the AGC voltage remains unchanged until the next normal sync pulse arrives.

The voltage developed on C41 has high-amplitude pulses superimposed on it and these must be filtered out before the voltage can be used for AGC. This is accomplished by the long time constants of R35, C9 and R38, C10.

The sync circuit is conventional. The right side of V6 remains cut off except when positive sync pulses are applied to its suppressor grid. The sync pulses are amplified and inverted by V6 and fed to the deflection circuits. The control grid of V6 is common to both sides, and a noise spike "riding" a sync pulse also holds this side of V6 in cutoff. As a result, one sync pulse does not appear at the output, but this does not cause the deflection circuits to fall out of sync.

Luminance Circuits

The main function of the luminance channel is to amplify the output of the video detector to a level which is sufficient to drive the CRT cathodes. The channel may have one, two, or three stages depending on the video output level, the delay-line loss, gain per stage, etc. Fig. 1 shows two stages since this number is a reasonable compromise among several modern sets.

While the luminance channel is similar in many respects to the video amplifier of a b-w receiver, it performs some additional functions. In many sets, the first luminance (or video) amplifier is used to amplify the signal fed to the sync and AGC circuits. In other sets, these signals are amplified separately or in conjunction with the sound IF signal. Some manufacturers take the chroma signal directly from the video detector to the chroma bandpass amplifier while others take this signal from the output of the first video amplifier. The luminance channel also introduces a specific delay in the brightness signal, contains the brightness and contrast controls, and has means for setting the signal levels at each of the individual CRT cathodes.

Fig. 10 was chosen to illustrate the luminance channel because it performs all of the functions mentioned above. The output of the video detector passes through a 4.5-MHz trap and a peaking coil to the grid of V4A. The level of the composite video at this point is approximately 2.5 volts.

Four outputs are derived from the plate circuit of V4A. A signal having a sync-pulse amplitude of 60 volts is taken from the bottom of L10 and fed to the chroma section. A 35-volt signal from the bottom of R59 goes to the sync separator, and one having an amplitude of 20 volts goes to the AGC keyer. The fourth output is applied to the grid of V4B through L12 and R61. These components reduce the high-frequency response to attenuate the chroma subcarrier sidebands. A portion of the output of the vertical output tube is also coupled to this grid to provide vertical blanking. In modern color receivers, it is common practice to inject the vertical blanking pulse into the luminance channel.

The delay line driver, V4B, is a triode because a fairly low output impedance is required. The type of delay line driver varies among sets of different make, but it is characterized by relatively low output impedance. Notice that the plate load resistor of V4B is only 6800 ohms. To drive the delay line, some manufacturers use a cathode follower,

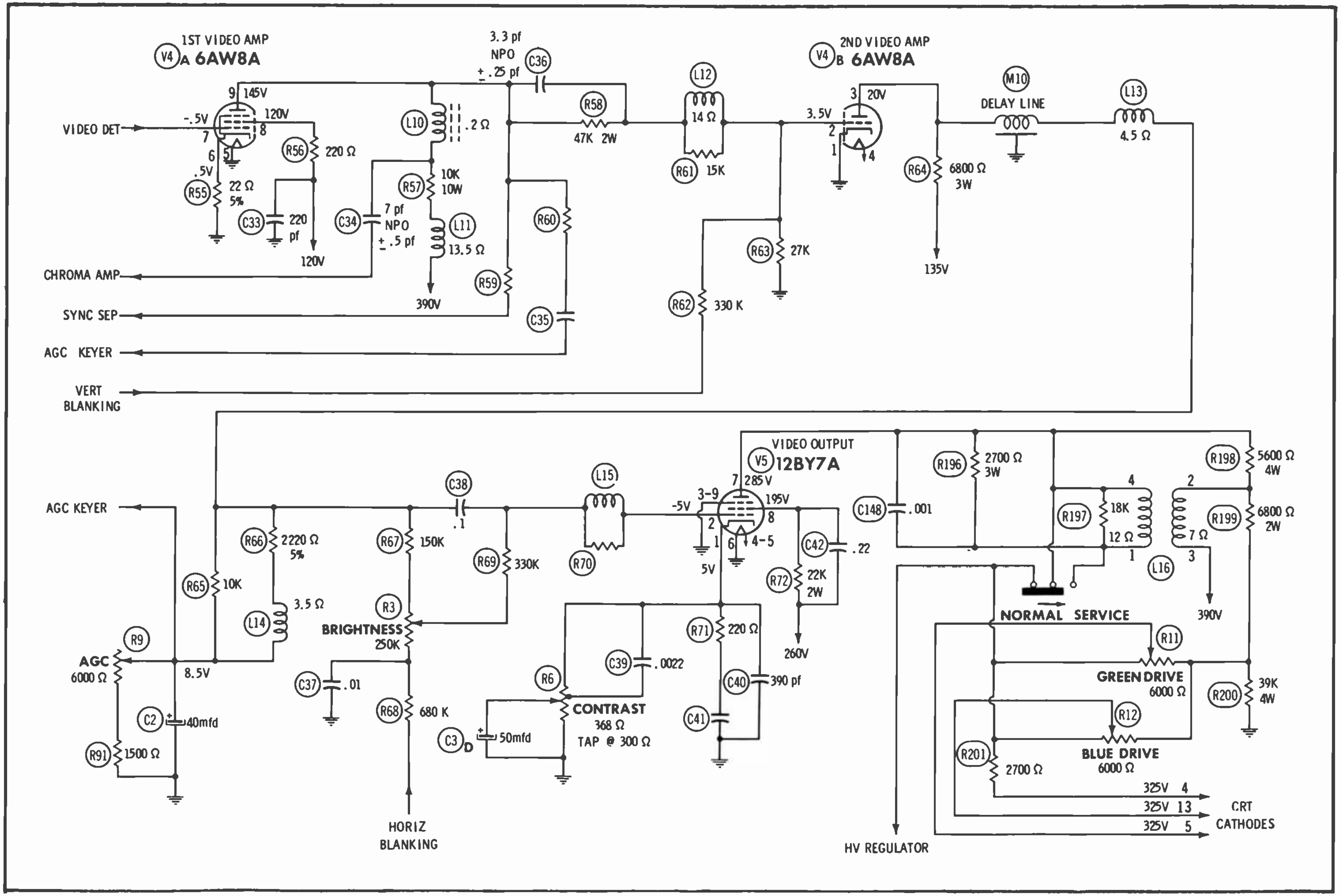
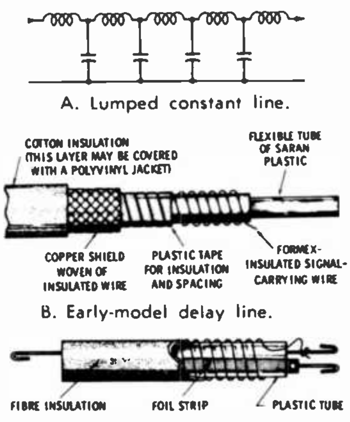


Fig. 10. Luminance channel of General Electric FY chassis.

others use a bootstrap amplifier, and still others employ a transistor.

One characteristic of any amplifier circuit is that signals passing through it are delayed by an amount which is inversely proportional to the bandpass. Since the bandpass of the luminance channel is more broad than that of the chrominance channel, the transit time is about 1 microsecond less. If no correction were made, the Y signal would be presented on the CRT before the chrominance signal and the color would "trail" the black-and-white image approximately N/64 inches, where N is the width of the raster in inches.

Although the term "delay line" may be new to some technicians, all have worked with transmission lines and coaxial cables which are, in reality, delay lines. As we all know, the velocity of propagation (speed) in a line is less than it is in free space, and so any transmission line could be used as the delay line in the luminance channel. The velocity of an electromagnetic wave is 300 meters per microsecond in free space but only 200M/usec in RG-59/U cable. Thus 200 meters (656 feet) of RG/59U would produce the desired delay.



A. Lumped constant line.

B. Early-model delay line.

C. Modern delay line.

Fig. 11. Evolution of delay lines.

Unfortunately for the cable manufacturers, more convenient delay lines have been developed. Delay lines consisting of lumped constants (Fig. 11A) have been used in many applications, but a "distributed constants" line is somewhat more economical. Fig. 11B is a drawing of the cable used in an early General Electric receiver (Model 15CL100). The inner conductor is wound around an insultating tube to increase its inductance, and a coating of powdered aluminum and styrene increases the shunt capacity. In this manner, a 1-microsecond delay is achieved in only 18 inches of this cable. Fig. 11C shows a more recent delay line which is only about 5 inches long.

Four signals are applied to V4 (Fig. 10). In addition to the delayed luminance signal and vertical blanking pulses, horizontal blanking pulses are inserted through R68 to the bottom of the brightness control. The brightness control determines the DC potential at the grid which, in turn, controls the plate voltage of V4 and, ultimately, the CRT cathode voltages. Notice that the lower ends of R65 and L14 are connected to the slider of the AGC control and the cathode of the AGC keyer. As the cathode bias of the AGC keyer tube is made more positive, the AGC voltages is decreased, resulting in greater receiver gain and a darker picture. However, this same bias which is applied to the cathode of the AGC keyer is applied to the grid of V5. Thus, as an increased signal tends to darken the picture, the CRT brightness is automatically increased.

The contrast (video gain) control located in the cathode circuit of V5 functions in much the same fashion. As the resistance is decreased, the contrast is increased and the plate potential is reduced. This reduces the CRT cathode potentials to increase the brightness.

The network in the plate circuit of V5 splits the signal into three separate signals which, in turn, are fed to the three CRT cathodes. While the red drive is fixed, the blue and green cathodes are connected through drive controls to provide for gray scale adjustments. In the "service" position, the service-normal switch shorts the primary of L16 and removes the luminance signal from the CRT. (Another switch section is used to remove the vertical sweep.)

An output is taken from the common ends of R11, R12, and R201; and, after it is integrated or filtered, it is used as the control voltage for the high-voltage regulator. For example, if the brightness control is set for less brightness, the DC level of the CRT cathodes is made more positive. This increased positive level increases the conduction of the shunt regulator to compensate for the reduced conduction of the CRT. Thus, the load on the high-voltage power supply remains constant.

Some of the receivers which were checked incorporated some type of video peaking control in the luminance channel. This control changes the high-frequency response of the video amplifier to give a sharper or softer picture. If it is a back-of-set control, it should be set according to the owners preference.

One important characteristic of all luminance amplifiers is often ignored by many technicians. Since all the stages in the channel are direct coupled, a moderate change in voltage level at one point in the circuit may have drastic effects elsewhere. For example, suppose that R55 in the cathode circuit of V4A increases in value enough to make the cathode potential rise to 1 volt. If this were a conventionally coupled amplifier, there would be very little change in operation since the tube would continue to function reasonably well with a moderate increase in bias. Not so in a chain of direct-coupled amplifiers! The increase in positive cathode bias will cause the plate voltage to rise. Since the two succeding plates are direct coupled, this change in level is amplified just like any other signal. The voltage gain from the grid (or cathode) of V4A to the CRT is about 40, so the .5-volt change at the cathode of V4A will cause the CRT cathodes to change about 20 volts. The change at the CRT cathodes would be positive, so a noticeable decrease in brightness would result.

Because of this characteristic, it is very important to be sure that an abnormal voltage near the output of the channel is actually being caused by a component in that specific area. Often the cause of trouble is far removed from the point where the symptoms are first detected. It is sound practice to check back towards the input until a point is reached where everything is normal and then examine the components just after this point.

Power Supply

The power supplies of color receivers are quite similar to those

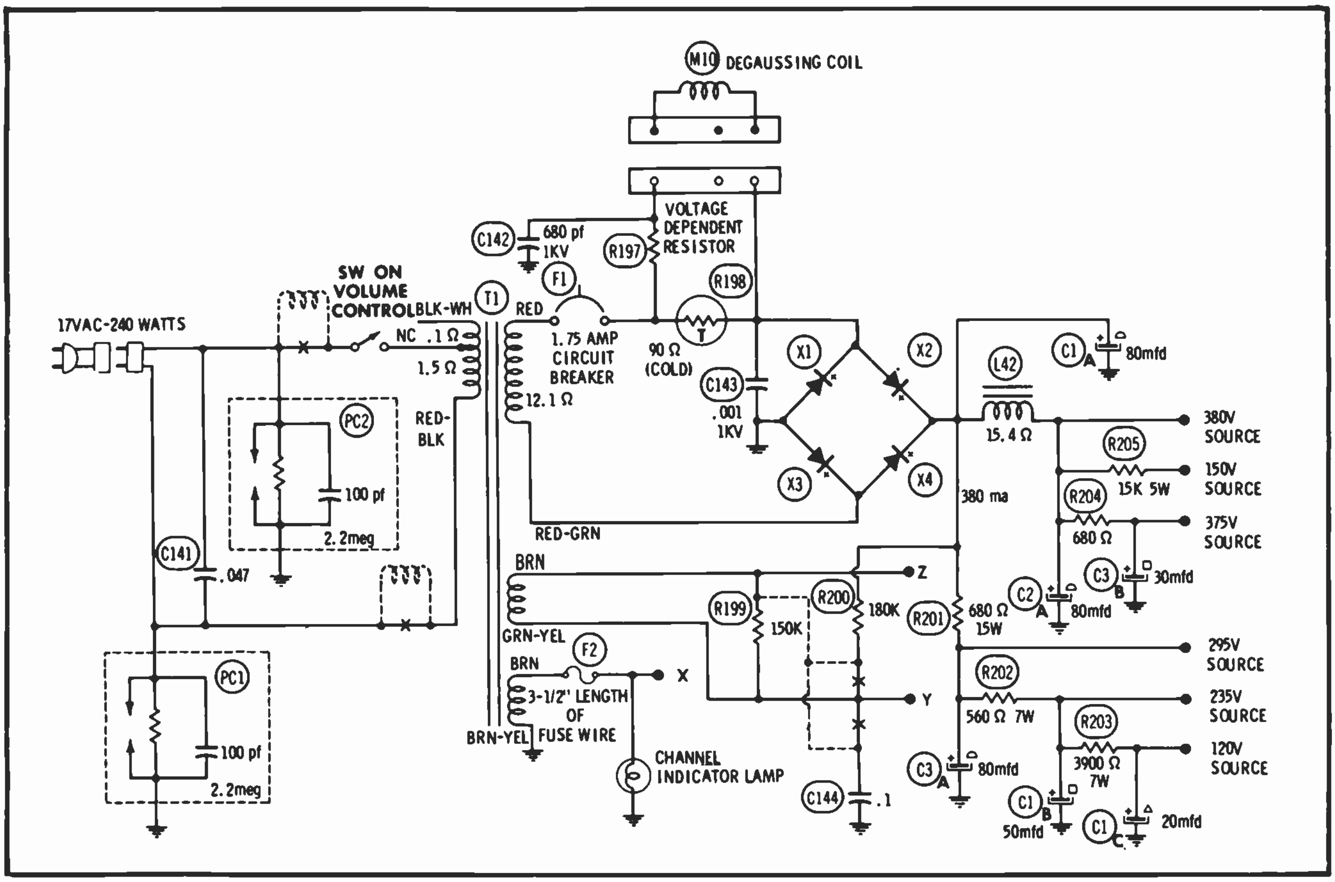


Fig. 12. Power supply of RCA CTC19 chassis.

found in many b-w receivers. Fig. 12 is the power supply of the RCA CTC19 chassis. While this power supply uses a full-wave bridge rectifier, full-wave doublers are used by some manufacturers. Some receivers use B+ voltages as low as 280 volts, but a B+ source nearer 400 volts is more popular. All the sets checked used a power transformer with a step-up winding for the rectifier and a separate filament winding for the shunt regulator and CRT. This winding is connected to B+ to reduce the heater-cathode potential of these tubes.

Nearly all present-day sets feature automatic degaussing and the circuit in Fig. 12 is typical. When the receiver is off, the resistance of R198 is high while the resistance of R197 is low. When the set is turned on, AC flows through the rectifier and the degaussing coil which are effectively in series. A small current flows through R198, causing it to become warmer. This warming action decreases the resistance of R198 and this decreases the voltage drop across M10 and R197. R197 is voltage sensitive, that is, its resistance increases as the applied voltage decreases. The action of R197 and R198 is cumulative, and so the current through the degaussing coil falls to zero after a few seconds.

Summary

Referring to Fig. 1, the color receiver circuits may be divided into three groups. One group comprises those circuits which are similar or identical to circuits found in b-w receivers: tuner, IF strip and video detector, most of the sound circuit, AGC keyer, sync separator, and the deflection oscillators. The video amplifiers, power supply, high-voltage supply, and deflection output circuits have b-w counterparts; but these circuits have been modified extensively to adapt them for color. The third group includes the circuits which do not have counterparts in a b-w receiver. These are the chroma circuits, sound IF detector, convergence circuits, video delay line, high-voltage regulator, and the focus rectifier.

The tuner and IF circuits are modified only slightly to increase the response near the upper edge of the video passband. Additional traps are required to prevent interaction between the 4.5-MHz sound carrier and the 3.58-MHz chroma information. The additional traps make it necessary to move the sound takeoff point "forward" and use a separate sound IF detector. Otherwise, the sound system is conventional. The AGC and sync circuits are nearly identical to the ones found in monochrome receivers.

The luminance channel performs the same functions as the monochrome video amplifier and also it has some additional features. It provides for video delay, performs the retrace blanking, incorporates the brightness control, and, of course, it drives three CRT cathodes instead of one. The luminance channel is direct coupled and this can lead to service problems for the unwary technician.

The power supply is essentially a "beefed up" b-w power supply. The usual B+ output is about 400 volts with a current drain near 500 ma. Automatic degaussing is accomplished by the power supply as an additional function. ▲

Chapter 3

Deflection and High-Voltage Circuits

In Chapter 2 of this book, the tuner, video IF amplifier, sound IF amplifier and output, luminance (or video) amplifier, and the sync and AGC circuits were discussed. All of these circuits are similar to circuits having the same functions in a black-and-white set.

The tuner and IF circuits are modified only slightly to increase the response near the upper edge of the video passband. Additional traps are required to prevent interaction between the 4.5-MHz sound carrier and the 3.58-MHz chroma information. The additional traps make it necessary to move the sound takeoff point "forward" and use a separate sound IF detector. Otherwise, the sound system is conventional. The AGC and sync circuits are nearly identical to the ones found in monochrome receivers.

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Vertical Deflection Circuit

The circuit most frequently used to generate the vertical deflection current is a modified free-running multivibrator. The modifications are incorporated to change the usual square-wave output into a waveform which produces a sawtooth current through the deflection coil or yoke. Since the impedance of the yoke is inductive, current lags the voltage; and so the multivibrator output waveform rises very steeply, decays steeply to about one-half its maximum, and then decays at an almost linear rate.

The circuit shown in Fig. 1 is from the Zenith 24MC32 chassis. This circuit is typical of many receivers although several sets examined used the old, familiar twin-triode configuration. We also noted that some circuits have the hold control in the cathode circuit instead of the grid circuit of the normally cutoff tube.

Consider the circuit at a time when the vertical sweep is approaching the bottom of the CRT. The grid of V9 is swinging in a positive direction as current flows upward through R97 to charge the right side of C45. The current through V9 and the deflection coils is increasing at a linear rate. V7 is cut off by the negative charge at the top of C46 although this voltage is approaching the level where the tube can go into conduction.

Shortly before V7 would go into conduction spontaneously, the negative, integrated vertical sync pulse is applied to the grid of V9. The following things all happen in a very few microseconds:

1. The current through V9 is diminished.
2. The field around the deflection coils begins to collapse, causing a positive spike to begin forming at the plate of V9.
3. This positive pulse is coupled to the grid of V7 causing it to conduct.
4. The plate of V7 swings in a negative direction and this negative spike is coupled to the grid of V9.
5. V9 is cut off and the plate rises to a peak of several hundred volts as a result of the total collapse of the field around the deflection coils.
6. The collapsing field drives the electron beam to the top of the CRT, ready to begin another downward scan.
7. The positive spike is coupled from the plate of V9 to the grid of V7. This drives V7 into saturation, the grid draws heavy current, and a negative charge collects at the top of C46. When the spike from the plate of V9 is ended, C46 begins discharging through R93 and R4 and this cuts off V7.
8. As soon as V7 returns to cutoff, its plate swings positive. This positive-going pulse partially discharges C45 and allows V9 to resume conduction.

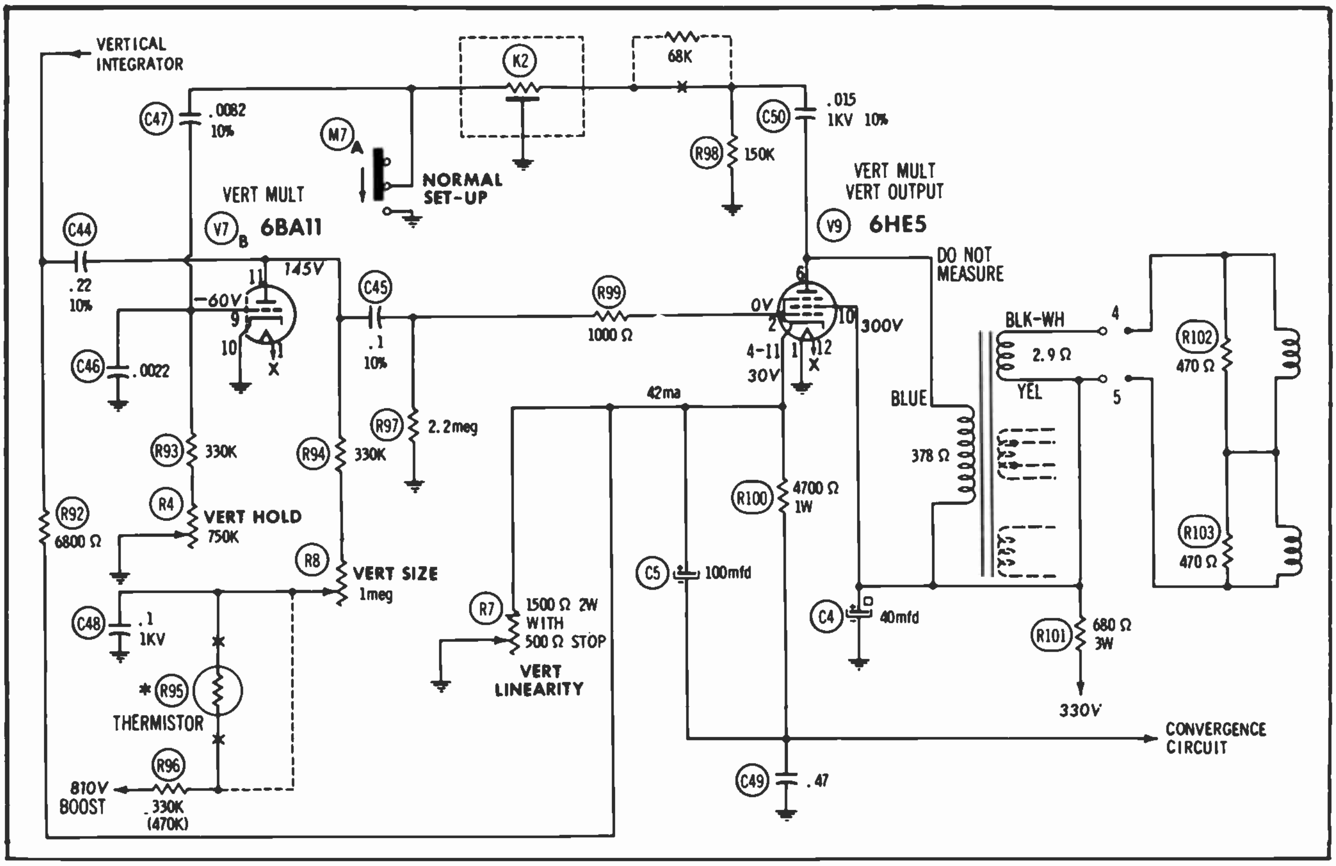


Fig. 1. Vertical multivibrator of the Zenith 24MC32 chassis.

When V9 begins conduction, the trace begins its downward deflection. The grid voltage of V9 gradually becomes more positive as current flows upward through R97 to charge C45. This increases the current through V9 and the deflection coils to produce the vertical sweep.

There are three controls in the vertical circuit: hold, size, and linearity. The hold control determines the time required for C46 to discharge to the point where V7 can conduct. If this discharge time is too short, (free-running frequency too high) a new vertical scan will have started before the arrival of the next sync pulse and the picture will roll down. If the free-running frequency is slightly lower than the scan rate, the vertical multivibrator synchronizes normally.

The size control determines the plate saturation voltage of V7. The difference between saturation voltage and B+ is the amplitude of the signal which is coupled to the grid of V9 and this ultimately determines the amplitude or size of the sweep.

The voltage at the grid of V9 is not completely linear since it is a portion of the exponential charging curve of a capacitor; however, V9 is not linear either (gain is different at various bias levels). By proper adjustment of the linearity control, a level of bias is obtained that causes these two inherent nonlinearities to be equal and opposite.

The network consisting of C47, K2, R98, and C50 shapes the feedback pulse to the grid of V7. The shape of this pulse determines the conduction time of V7 and this determines the retrace time. M7 disables the multivibrator for color setup purposes. R102 and R103 are loading resistors across the deflection coils. They prevent the coils from oscillating at the end of the retrace interval. The resistance of R95 decreases with temperature rise to compensate for the increase in resistance of R96, R8, and R4 as the temperature rises.

Some sets use a vertical centering circuit to control a small DC current which flows through the vertical deflection coils. This current either aids or opposes the vertical deflection current to shift the entire raster up or down.

Horizontal Phase Detector

The horizontal oscillator and its synchronizing circuits used in color sets are no different than those used in b-w receivers. The schematic shown in Fig. 2 is the horizontal phase detector circuit used in the Philco 16QT85 chassis. The two inputs to the phase detector are compared to determine their relative timing and a correcting voltage (error signal) is developed to correct the oscillator frequency.

Negative sync pulses from the sync separator are applied to the junction of the cathodes of the phase detector diodes and both diodes conduct. This places a negatie charge on the tops of C75 and C52. Between sync pulses, these capacitors discharge as follows: C75 discharges downward through R113 causing the anode of X20 to be negative with respect to ground. C52 discharges downward through R112 causing the bottom of R112 to be positive with respect to its top. Since the voltage at the bottom of R112 cannot be negative with respect to the top of R113 and the voltage drops across R112 and R113 are nearly equal, the voltage measured from the top of R112 (or C52) to ground is nearly zero.

At about the same time that the

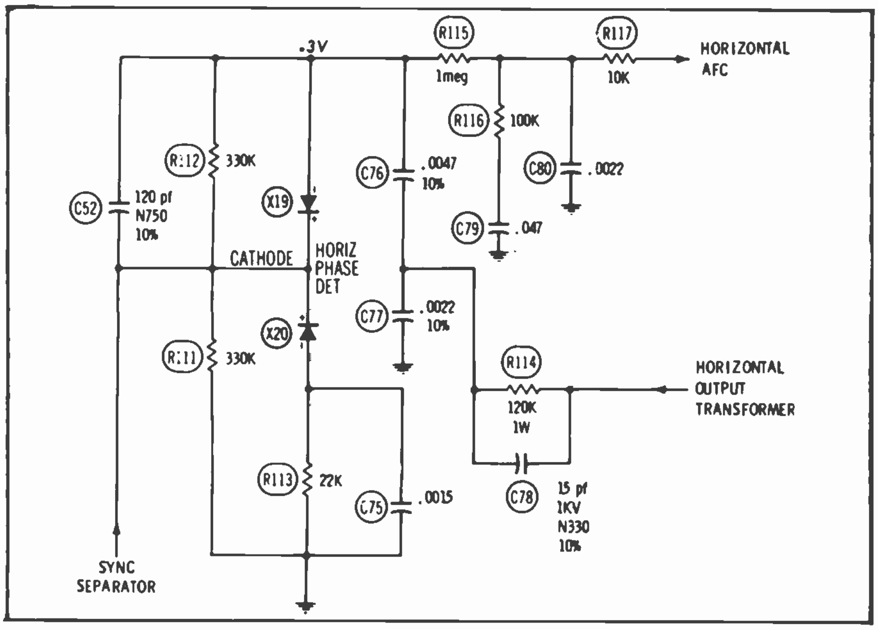


Fig. 2. Horizontal phase detector of the Philco 16QT85 chassis.

sync pulse is applied to the junction of the diode cathodes, a negative pulse from the horizontal output transformer is supplied to the junction of C76 and C77. This drives electrons away from the top of C76 and down through R112 and R111. During the interval between pulses, the current flows back up through R111 and R112 producing a positive-going sawtooth at the top.

Now consider the interaction of the two separate actions which result from the sync pulse and the feedback pulse. Referring to Fig. 3A, we see the phase relation of the two voltages when the oscillator is operating at the correct frequency. The sync pulse arrives shortly after the start of the negative-going feedback pulse. Since X19 can conduct only until its anode becomes negative with respect to its cathode, only a portion of the electrons that could be supplied by the sync pulse actually get to the top of C52. During the long interval between sync pulses, some of the electrons required to recharge the top of C76 are supplied from C52 and the remainder flow upwards through R111 and R112. Since the total-current through R111 and R112 has been diminished by the action of the feedback pulse, the average-voltage at the top of C76 is less positive than it would be in the absence of the sync pulse.

Fig. 3B illustrates the action of the phase detector if the oscillator frequency is too high. The feedback pulse is shifted to the left from its normal position and the anode of X19 is driven farther negative before the sync pulse arrives. As a result, fewer (or none) of the electrons supplied by the sync pulse are actually deposited on the top of C52. During the interpulse interval, all of the electrons must flow upwards through R111 and R112 and the voltage at the top of C76 is more positive.

Inspection of Fig. 3C shows that if the feedback pulse arrives too late (frequency low), a greater portion of the sync pulse electrons reach C52 and the average voltage at the top of C76 is less positive than before.

Referring once again to Fig. 2, R115, R116, R117, C79 and C80 form the anti-hunt network. This is essentially an integrating circuit which smooths the voltage at the top of C76 into an almost constant level. If the integrations are too great (time constants too long), horizontal pulling will result because the AFC tube will not correct the frequency quickly enough. On the other hand, too little integration will cause horizontal jitter because the AFC tube will tend to overcorrect the frequency (hunt).

From this discussion of the phase detector circuit, we learn that an increase in oscillator frequency results in a positive-going error signal at the grid of V7. While this is true of this specific circuit, some circuits are designed so that the exact opposite is true. That is, increased frequency produces a negative-going error signal in some receivers even though the circuit configuration is quite similar.

Horizontal AFC and Oscillator

Referring to Fig. 4 which is the schematic of the AFC and oscillator circuits of the same Philco 16QT85 chassis, the error signal is applied to pin 9 of V7. However, before attempting to understand the operation of the AFC tube, the operation of the oscillator must be thoroughly understood.

The oscillator is basically a Hartley type although the tank circuit capacity, as well as the inductance, is tapped. The waveshape of the oscillator output is modified by R-

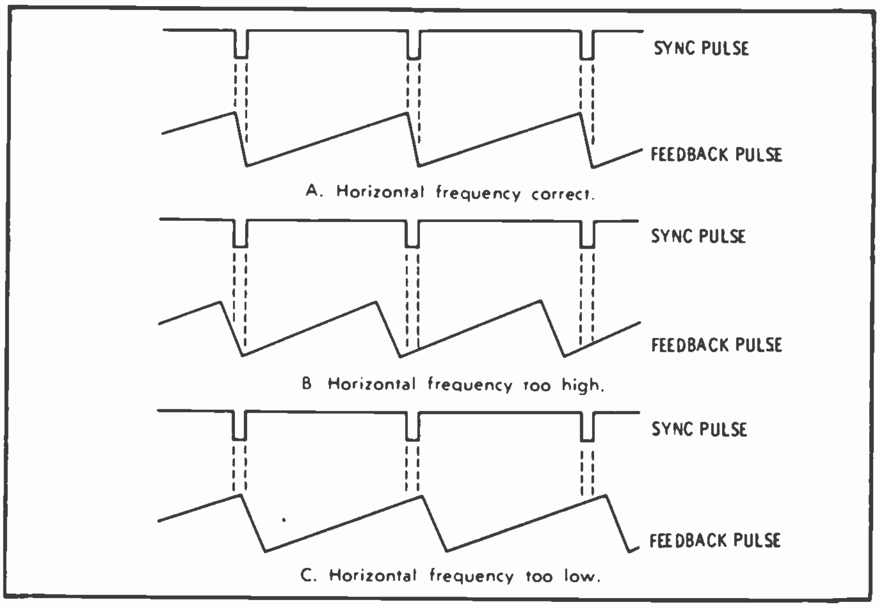


Fig. 3. Horizontal sync and feedback pulses.

124 and C88 before it is applied to the output tube. V7B is grid-leak biased by R122 and C84.

The frequency-determining circuits of the oscillator are L37, C85, C86, and the AFC tube. Insofar as an AC signal is concerned, the AFC tube is shunted across C86, and so the current through V7A is part of the total current of the tank circuit.

V7A has two inputs, the error signal from the phase detector and also the 15,750-Hz signal present at the cathode of V7B. This second signal is shifted in phase by the combination of C81, R117, and C80 so that the voltage at the grid of V7A is leading the voltage at the cathode of V7B. This leading voltage causes a leading current to flow through V7A and through the tank circuit. From basic electronic theory, we recall that a leading current is a capacitive current, and so the current of V7A appears as an additional capacitive current in the tank. Thus, as the error signal goes positive, the capacitive current increases and the resonant frequency decreases to reduce the error.

R4 and R120 set the bias of V7A and determine the free-running frequency of the oscillator. The error signal at the grid of V7A causes the capacitive current to vary above or below this level to maintain horizontal sync.

It was pointed out earlier that the *direction* of the error signal generated by the phase detector can be either positive or negative for a given frequency error. It is also possible to change the AFC circuit from the one shown in Fig. 4 so that a positive error signal at the grid will increase the frequency rather than decrease it. One way of doing this is to interchange the values of C80 and C81. Now, the feedback would lag and the current through V7A would appear inductive rather than capacitive.

There are numerous other circuits which are used to generate the horizontal time base. Space limitations do not allow an exhaustive analysis of each of them in this course. For most service problems, the technician will be able to solve the difficulties if he will simply take time to identify the frequency-determining components and determine the *direction* of the error signal from the phase detector.

Horizontal Output and High-Voltage Circuits

Regardless of the oscillator and AFC circuits used, the designs of the horizontal-output sections used by various manufacturers are similar. In many respects, the circuits are the same as ones used in monochrome receivers, but the whole horizontal deflection system has been "beefed up" for several reasons:

1. The ultor (high-voltage anode) voltage (about 25 kv) is higher than it usually is in b-w receivers, and the CRT beam current is increased three-fold.

2. The shunt regulator draws additional current from the ultor supply.

3. The boosted B+ has a higher potential and the load is greater.

4. The focus rectifier and bleeder, the convergence circuits, and the pincushion circuits extract power from the horizontal output circuit.

The schematic of the horizontal output and high-voltage circuit used in the Zenith 25LC30 chassis is shown in Fig. 5. The yoke is driven in the conventional manner by taps on the primary of the high-voltage transformer. A secondary winding is used to obtain the voltages necessary for convergence, horizontal AFC, and keying of the AGC, color killer, and burst amplifier.

The anode supply for the focus rectifier is obtained from a tap on the primary of the high-voltage transformer. The output of V15 is divided by the bleeder consisting of R130, R131, R132, and R20; and the focus voltage is controlled by the setting of R20. C68 is a filter for the focus supply and R133 is an arc protector.

The ultor voltage is developed in the usual manner but, unlike b-w supplies, it is regulated by a shunt regulator. Regulation is accomplished by the special regulator tube (V16). R19 is used to set the bias on V16 so that it is near cutoff when an all-white picture is being displayed. As portions of the picture are made black or gray, the CRT beam currents decrease and the ultor supply voltage tends to rise. The boosted B+ also rises and this positive-going voltage is used to decrease the bias on V16. Thus, V16 conducts more, the load on the ultor supply remains constant, and the ultor voltage is stabilized.

Some manufacturers use the average potential of the CRT cathodes as a control voltage for the ultor regulator. A black picture is produced by a positive-going voltage at the CRT cathodes, and this voltage (in lieu of the boosted B+) is used to increase the conduction of the regulator tube.

A rather unique approach to the problem of high-voltage regulation is illustrated in Fig. 6. This is a schematic of the horizontal output and high-voltage section of the Ad-

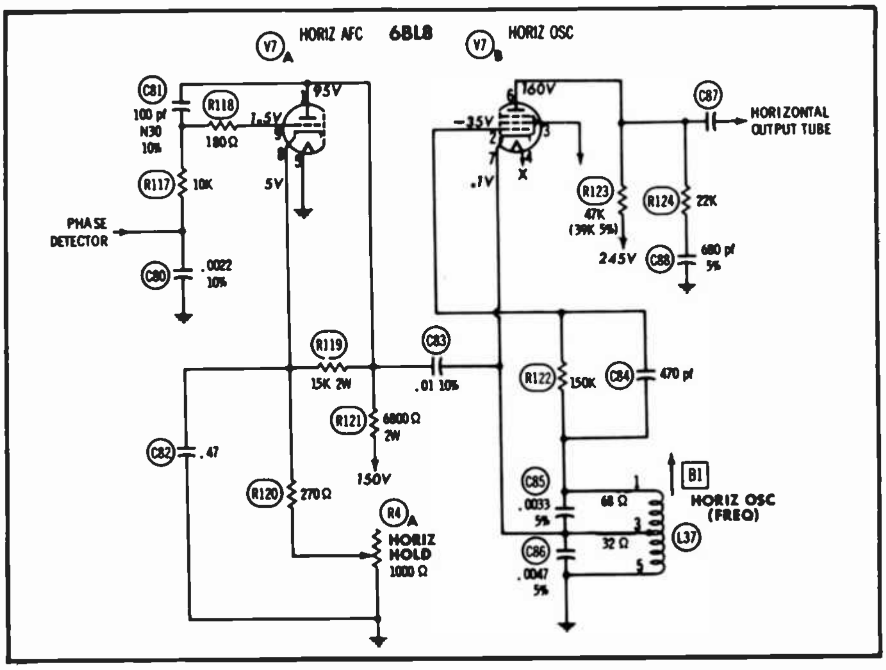


Fig. 4. Horizontal oscillator and AFC of the Philco 16QT85 chassis.

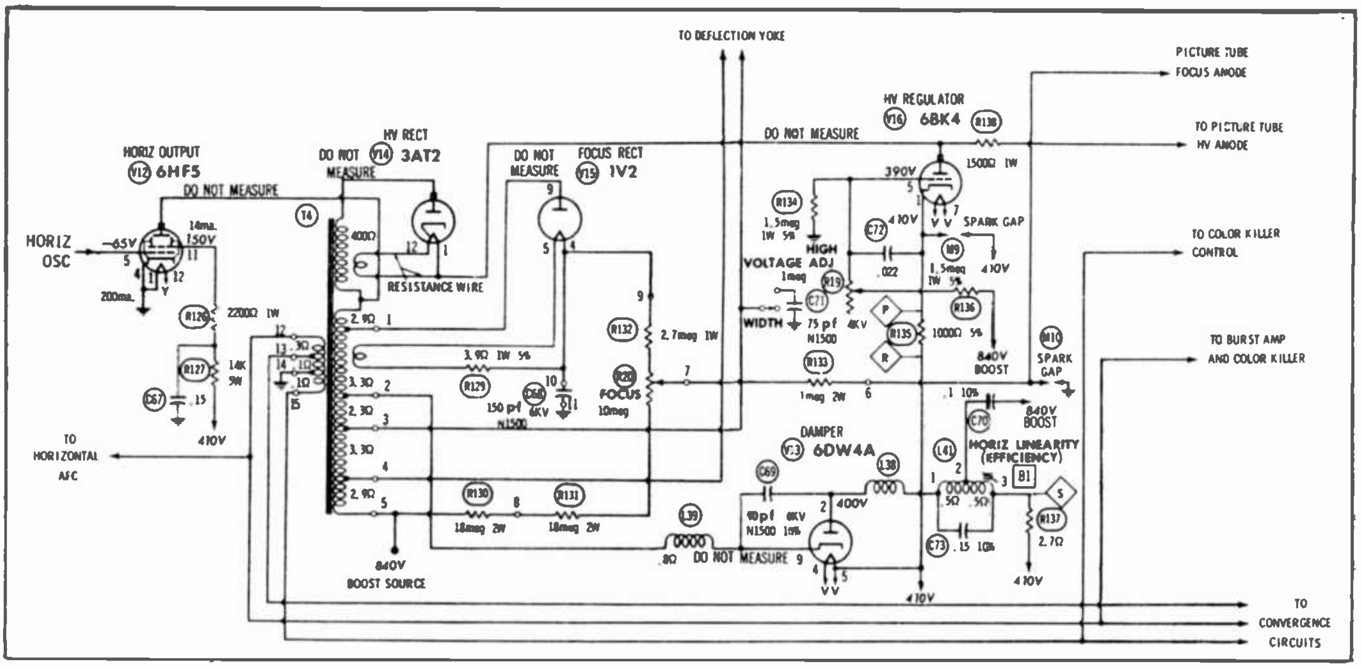


Fig. 5. Horizontal output and high-voltage circuit of the Zenith 25LC30 chassis.

miral 1G1155-1 chassis. Notice that no shunt regulator is used.

A feedback pulse from the secondary winding (terminals 1 and 2) of T4 is rectified by X10 and the resulting voltage is added to the bias supply for the horizontal output tubes. This total bias determines the amount of drive to the high-voltage transformer and, finally, the potential of the ultor supply.

As the CRT beam currents increase and load the supply, the amplitude of the feedback pulse is decreased and this results in a lesser bias voltage at the grids of V10 and V11. This causes V10 and V11 to conduct more heavily to increase the ultor voltage.

While this system is adequate, it does allow some variations in the high voltage. In order to maintain good focus over a range of ultor voltages, the focus voltage is made to "track" the ultor voltage. This is accomplished by R141 and R142 along with their filter, C99. As the CRT beam currents increase, the voltage across R141 and R142 increases and this voltage is added to the output of the focus rectifier to adjust the focus automatically for changes in CRT current. A slight change in width as the brightness control is rotated is normal in receivers using this circuit.

The horizontal output and high-voltage circuits of the RCA CTC16 chassis are shown in Fig. 7. In this circuit, the focus rectifier is connected to the plate of the horizontal output tube and develops 4 to 4.5 kv. The ultor-voltage supply and regulator are similar to the Zenith chassis discussed earlier, but the regulator control voltage is derived from the CRT cathodes.

The horizontal-centering problem is solved by driving the yoke with two identical windings of the high-voltage transformer. A 10-ohm potentiometer used as a centering control is connected between terminals 3 and 4, and a small DC potential exists across it. One end of the yoke is connected to the center arm of the control. Depending

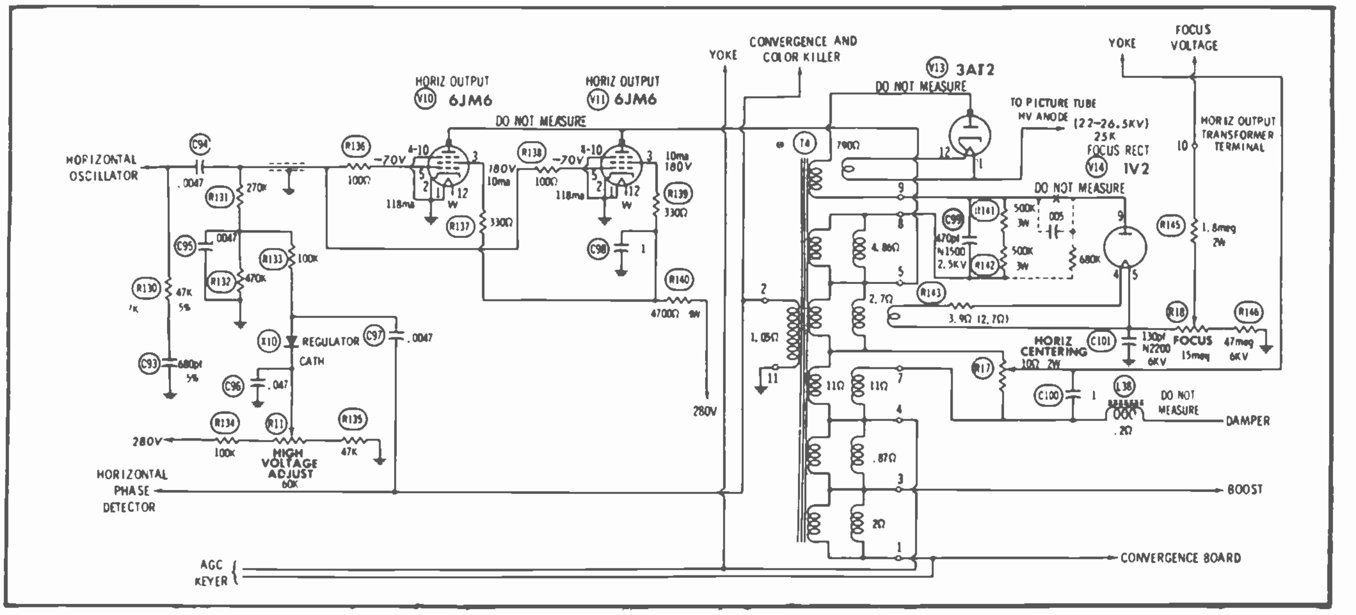


Fig. 6. Horizontal output and high-voltage circuit of the Admiral 1G1155-1 chassis.

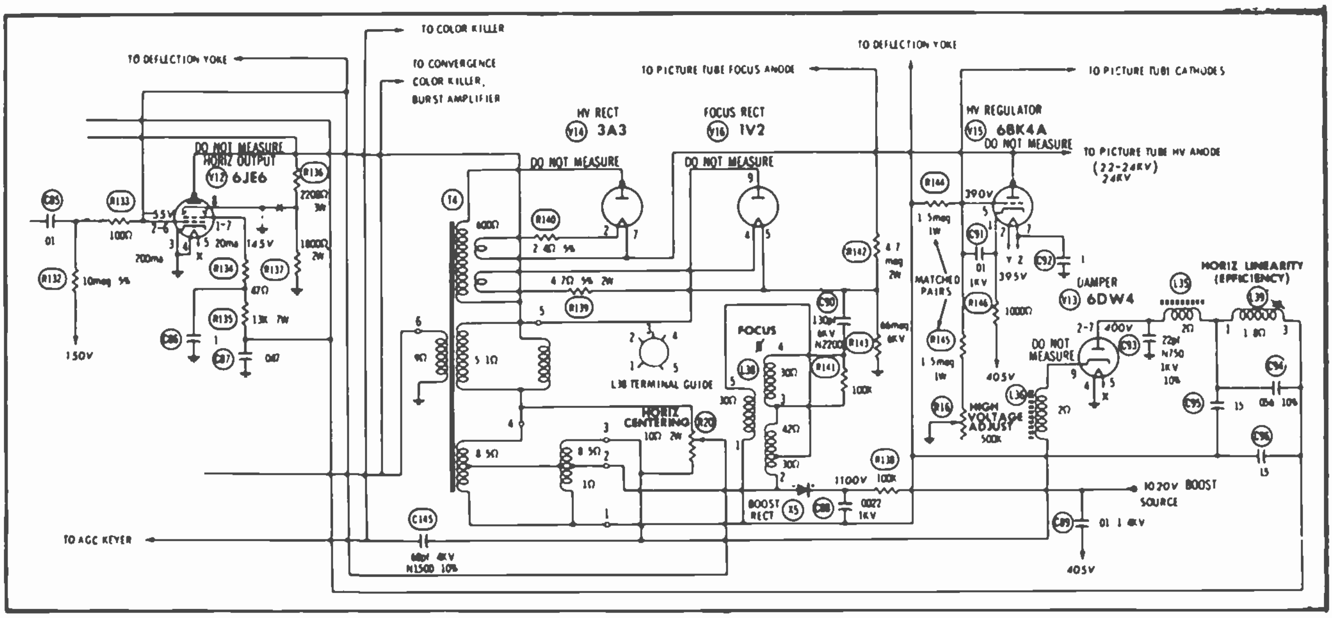


Fig. 7. Horizontal output and high-voltage circuit of the RCA CTC16 chassis.

on the setting of the control, a small DC current can be caused to flow in the yoke to change the centering.

A third method of high voltage regulation has been developed in the Zenith 23XC38 chassis. The horizontal output circuit is shown in Fig. 8. This circuit uses a type 6HS5 regulator (V21) whose cathode and plate are connected to B+ and a tap of the horizontal-output transformer, respectively. Since the tube is operating at potentials which are much lower than those encountered in the conventional shunt regulator circuit, there is no need for radiation shielding or a double-ended envelope for the tube.

The control signal for V21 is derived from a divider network consisting of R82, R83, and R84 connected between a portion of the boosted B+ supply and ground. R83 is used to set the bias level of V21 and this ultimately adjusts the ultor supply voltage. A second voltage is also applied to the grid of V21. This is a positive pulse taken from the cathode circuit of the horizontal oscillator.

Assume a white raster which results in maximum loading of the ultor power supply. The boosted B+ potential is minimum and the bias on V21 is maximum. Under these conditions, the conduction of V21 is minimum and the regulator has little effect on the output voltage of the ultor supply.

Under black-raster conditions, there is no load on the ultor supply and this voltage tends to rise. The boosted B+ also tends to rise and the bias of V21 is reduced. The horizontal-oscillator pulse causes V21 to conduct during the retrace interval and this conduction loads the horizontal output transformer. Stated another way, V21 clips the positive excursion of the "ringing" or flyback pulse of the horizontal-output transformer and this tends to reduce the ultor supply voltage. The amount of clipping (or loading) is determined by the boosted B+ potential which, in turn, is determined by the load on the ultor supply. As the raster changes from black to white, the conduction of V21 becomes progressively less and so the ultor supply voltage is stabilized.

The regulator tube is pulsed on only during horizontal retrace for two reasons:

1. Since V21 is cut off during the forward scan, it does not affect the width of the raster.

2. Since V21 conducts only during retrace, the duty cycle is low and only a minimum amount of power is extracted from the horizontal-output circuit. This also minimizes the plate-dissipation requirements for V21.

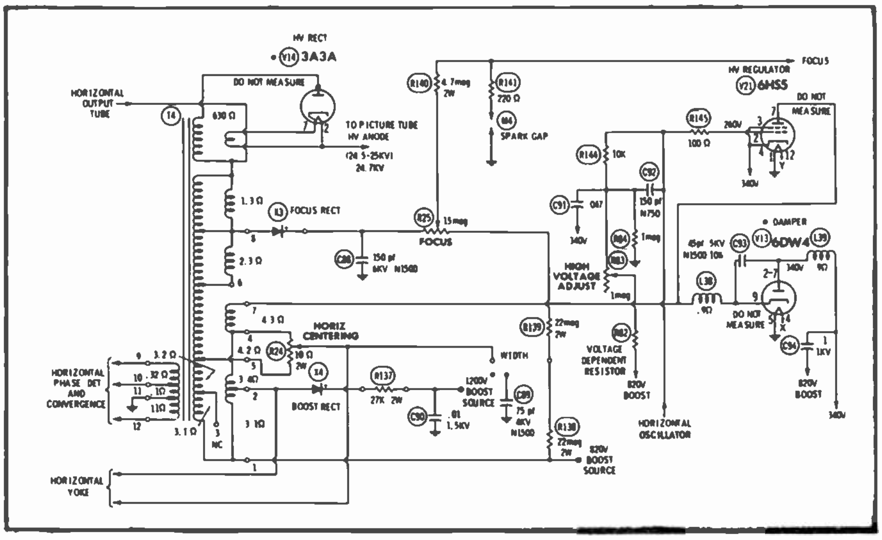


Fig. 8. Horizontal output circuit in Zenith 23XC38 chassis.

Summary

In this chapter, we have analyzed the power supply and deflection circuits. The principal difference between these circuits and their b-w counterparts is that greater power-handling capability and greater stability are required in the circuits of **color receivers.** ▲

Chapter 4

Block Diagram Analysis of Chroma Circuitry

As we have seen from Chapters 2 and 3 of this book, much of the circuitry of a color receiver is similar to that which is found in a well-designed monochrome set, and many of the modern circuit refinements are nothing more than the result of the normal progress of the art of electronics. The two sections of a color TV whose circuits differ from black-and-white circuitry are the chrominance channel and the picture tube itself.

In considering the chrominance circuits and the color picture tube, it is well to remember one important point: Black-and-white TV designers in the late '40's had a wealth of information from which to draw, including a great amount of practical experience obtained from WW II electronics devices similar to television. On the other hand, many brand-new concepts were involved in the design of the first color sets. Thus, the chroma circuits are of much more recent design and are still undergoing a certain number of "growing pains."

To cite two examples, consider the continuing development of the color CRT and the numerous variations in demodulator designs in present-day sets. In this discussion of the chrominance circuits, we will consider the more popular present-day designs, realizing that radically different designs are a possiblity.

Block Diagram

Figs. 1 and 2 are block diagrams of two chrominance circuits. The essential difference is whether high-level or low-level chroma demodulation is used. Low-level demodulation (Fig. 1) is an earlier design but it is by no means obsolete. At present, all the major manufacturers, with the exception of Admiral, Motorola, and Zenith, are using low-level demodulation. As shown in Fig. 2, the Zenith design uses two tubes in the high-level demodulator while Admiral and Motorola use only a single tube.

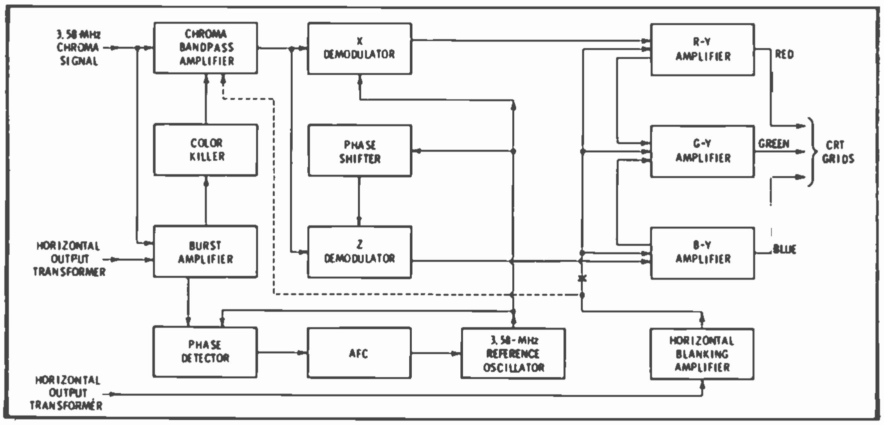


Fig. 1. Functional diagram of a chroma circuit using low-level demodulation.

Most of the sets that use low-level demodulation use two demodulator tubes operating on the X and Z axes. Three color-difference amplifiers are used, R − Y, B − Y, and G − Y. The input for the G − Y amplifier is derived from a matrix circuit driven by the other two difference amplifiers.

General Electric has developed a demodulator circuit using diodes in a modified phase-sensitive detector. Some of their sets use three-axis chroma demodulation, but others demodulate only two axes and derive a third signal, G − Y, from a matrix. In either case, R − Y, B − Y, and G − Y color-difference amplifiers are used. Electrohome also uses diode demodulators in some of their models.

Referring to Fig. 1, there are four major functions which the chroma circuits must perform. They must reconstitute the 3.58-MHz color reference signal, amplify the chroma sidebands, demodulate these sidebands to develop the R − Y, B − Y, and G − Y color-difference signals, and amplify the difference signals to a level which is sufficient to operate the picture tube. (These last two functions are combined in circuits using high-level demodulation.) The chroma circuits perform two auxillary functions, color killing and horizontal blanking; however, these are not essential to color operation. In fact,

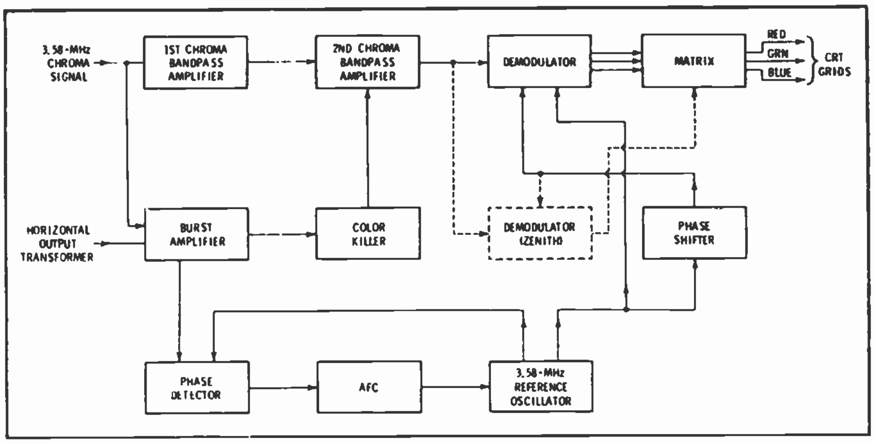


Fig. 2. Functional diagram of a chroma circuit using high-level demodulation.

the color killer only operates during b-w reception, and not all chroma circuits have blanking.

The Reference Signal

From Part 1 of this book, we recall that the 3.58-MHz chroma subcarrier was suppressed at the transmitter. There are two reasons for suppressing this subcarrier:

1. The energy contained in the subcarrier would be part of the total available transmitter power and the useful signal power would be decreased accordingly.
2. The presence of a strong 3.58-MHz signal at the video detector of the receiver would make it very difficult to prevent a 920-kHz beat from appearing in the picture. This interference could have been trapped out of sets built subsequent to the date when the color transmission standards were implemented, but b-w receivers already in service at that time would have been almost useless during color broadcasts.

Since the chroma subcarrier is suppressed at the transmitter, it must be regenerated in the receiver. This is accomplished by the 3.58-MHz reference oscillator. This is a crystal-controlled oscillator whose phase is controlled by the color burst signal which is part of the transmitted composite video.

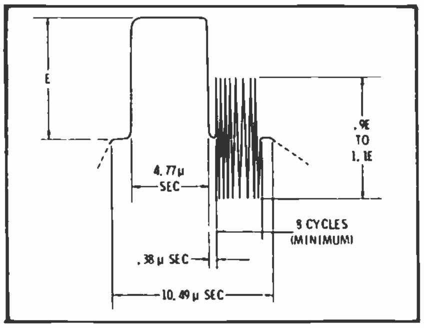


Fig. 3. Sync pulse and color burst.

The color burst is a short burst of the 3.58-MHz signal generated in the color modulator. A minimum of eight cycles of this signal is transmitted immediately following each horizontal sync pulse. Fig. 3 shows the position and amplitude of the color burst relative to the horizontal sync pulse. This waveform may be observed at the output of the video detector of the receiver and this is a convenient starting point when troubleshooting a no-color condition.

Referring again to Fig. 1, the composite video signal, including the color burst, is applied to the input of the burst amplifier which is biased below cutoff. A second input of the burst amplifier is an "on" gating pulse, usually taken from the horizontal output transformer. Thus, the burst amplifier is a coincidence gate which can amplify only the color burst.

After the color burst is separated from the remainder of the composite video by the burst amplifier, it is applied to the phase detector. The function of the phase detector is similar to the familiar horizontal phase detector in that it compares a synchronizing signal with a locally generated signal and develops an error voltage proportional to the difference in phase of the two. This error signal is used by the AFC circuit to correct the phase of the 3.58-MHz reference oscillator and synchronize it with the color burst.

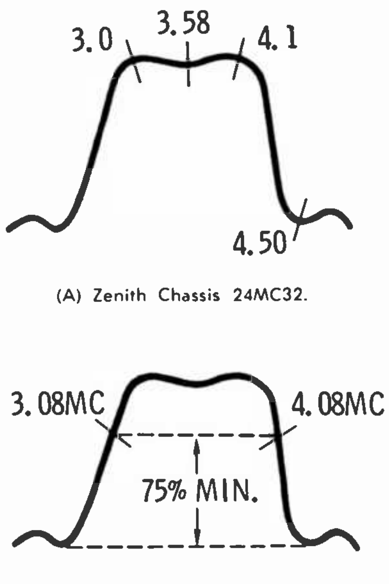
The reference oscillator is a crystal-controlled oscillator which operates at the exact frequency of the reference oscillator at the transmitter. The FCC standards for the oscillator at the transmitter are 3.579545 MHz ± .0003% with a maximum rate of change of frequency not to exceed .1 Hz per second. In a well-designed receiver, the reference oscillator output approaches this same precision because it is corrected at the start of each horizontal trace—15,734 times per second.

Returning for a moment to the burst amplifier, let's consider its second function. During a colorcast, a portion of the burst signal is used to cut off the color killer tube. During a black-and-white broadcast, the color killer conducts and biases the chroma bandpass amplifier below cutoff to prevent colored "snow" from appearing on the CRT.

Bandpass Amplifier

The chroma bandpass amplifier is similar to a conventional IF amplifier having a center frequency of 3.58 MHz. Fig. 4A is the chroma amplifier response curve of Zenith Chassis 24MC32 which uses two stages of amplification, and Fig. 4B shows the response of the single-stage amplifier of the RCA Chassis CTC19. Notice that the Zenith response is somewhat broader. Nearly all current sets have response curves which lie within the limits set by these two curves.

The curves obtained in Figs. 4A and 4B were obtained by injecting a signal from a sweep generator at



(A) Zenith Chassis 24MC32.

(B) RCA Chassis CTC19.

Fig. 4. Bandpass amplifier response.

the video detector output and observing the response at the signal grid of a demodulator. An interesting variation in the method of determining the response is discussed in an article by Carl Babcoke which appeared in TEST EQUIPMENT CYCLOPEDIA, Howard W. Sams & Co., Inc.; Cat. #50007. Using this method, known as "video sweep modulation" (VSM), a signal at the receiver intermediate frequency (45.75 MHz) is modulated by a video-frequency sweep generator operating between approximately 2 and 5 MHz with suitable markers injected. The response curve shown in Fig. 5 is typical of the curve that should be observed at a demodulator grid. This method of observing the response of the chroma bandpass amplifier takes into account any misalignment of the video IF strip and is a valuable "quick - check" for overall alignment of the receiver.

This rather lengthly discussion of the bandpass characteristics of the chroma amplifier and the reference to overall bandpass was included because of the importance of correct bandpass in obtaining a good color picture. Even the rather moderate change in response shown in Fig. 6 (same method of observation as Fig. 5) will seriously degrade the color performance of a set. Indeed, there are cases on record where a too sharply tuned antenna has made color reception impossible even though the receiver bandpass was well within tolerance.

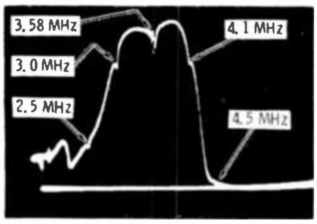


Fig. 5. Typical overall response of IF and chroma (VSM method).

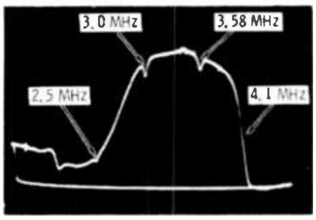


Fig. 6. Incorrect overall response curve (VSM method).

Again referring to Figs. 1 and 2, the input to the chroma bandpass amplifier is the 3.58-MHz chroma information. The 4.5-MHz audio subcarrier is rejected by traps, discussed in Part 2 of this series, and the luminance or Y signal and the sync pulses are blocked by a small coupling capacitor between the video detector and the bandpass amplifier. The design of some receivers (the RCA CTC12 chassis, for example) incorporates a circuit which cuts off the bandpass amplifier during horizontal retrace.

Generally, a single-stage bandpass amplifier is used in conjunction with low-level demodulators, and two stages of amplification precede a high-level demodulator. At any rate, the output of the bandpass amplifier, which is applied to the demodulators, is a 3.58-MHz signal whose phase is determined by the hue of the picture being transmitted at that particular moment. The amplitude of this same 3.58-MHz signal is determined by the intensity, or degree of saturation, of the hue.

Chroma Demodulation

In part 1 of this series, it was explained how the three primary colors could be modulated on a single carrier without losing any of the information. By a reversal of this modulation process, the color information may be obtained from the chroma signal. It is not necessary to demodulate on three separate phase axes since the third color-difference signal usually (G − Y) may be derived from a suitable combination of the other two (R − Y and B − Y). The G − Y voltage is a combination of −.51 R − Y voltage and −.19 B − Y voltage. $E_G - E_Y = -.51\ (E_R - E_Y) - .19\ (E_B - E_Y)$.

It can be proven mathematically that the chroma signal can be demodulated on nearly any pair of axes, but considerations of economy in design have resulted in the use of the X and Y axes in most sets. However, some manufacturers (Admiral and Zenith, for example) use the R − Y and B − Y axes. Fig. 7 shows the more important phase angles involved in color modulators and demodulators. The I and Q axes are no longer being used for demodulation, by the way.

An inspection of Fig. 7 will show that the R − Y and B − Y axes are not far removed from the X and Z axes. Thus, the axes of demodulation of nearly all sets of current design are approximately the same.

Referring to Fig. 1, the chroma signal is applied to the X and Z demodulators in the same phase. A potentometer in the output of the bandpass amplifier is used to adjust the amount of color signal which is fed to the demodulators. This is a front-panel control and is usually labeled "Color"

A second input, from the reference oscillator, is also fed to the demodulators; however, the phase of the reference signal is **not** the same at each demodulator. If the X and Z axes are being used, the reference signals at the two demodulators differ by 63.9°; or, in the case of R − Y and B − Y demodulators, the phase difference is 90°.

In any event, the function of the X demodulator is to produce an output which is proportional to the chroma information present along this particular axis. By the same token, a Z demodulator produces an output corresponding to the chroma information on the Z axis, an R − Y demodulator detects the information present on its axis, etc. We will take up the specific circuitry of the various demodulators later in the series.

While many texts explain the operation of the demodulators by proving what outputs will be present for a number of hypothetical input-signal phases, it is perhaps more meaningful to show the demodulator outputs which result when a keyed-rainbow signal is fed

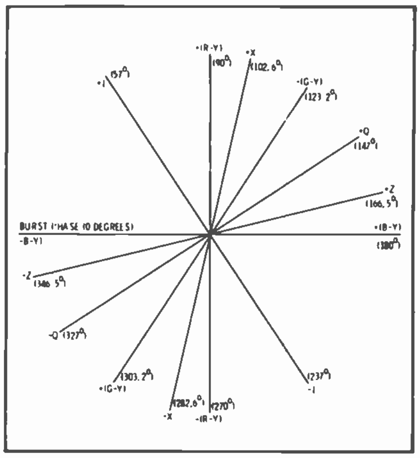


Fig. 7. Chroma demodulator axes.

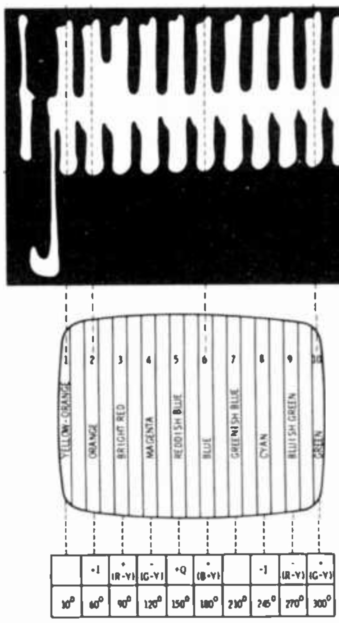


Fig. 8. Keyed-rainbow chroma signal, CRT presentation, and phases.

into the set. Fig. 8 shows the waveform that will be observed at the signal grids of the demodulators. Notice that there are eleven energy pulses, in addition to the color burst, although only ten bars are normally seen on the color receiver. The eleventh pulse occurs during retrace and cannot be seen, and, of course, the burst pulse is invisible on the CRT. Fig. 8 also shows the color bars produced by the first ten energy pulses along with the phase angle of the chroma signal at the center of each color bar. The various axes which were shown in Fig. 7 also appear again in Fig. 8.

Fig. 9 shows the outputs of I, Q, R − Y, B − Y, and G − Y demodulators. Notice that the output of each demodulator reaches a positive maximum when the signal phase is the same as the axis of the particular demodulator and reaches a negative maximum 180° later. Also notice that the output of a demodulator is zero when the signal is displaced 90° from its axis. These zero-output points, or nulls, are more frequently discussed since they are more easily observed than the maximum output points.

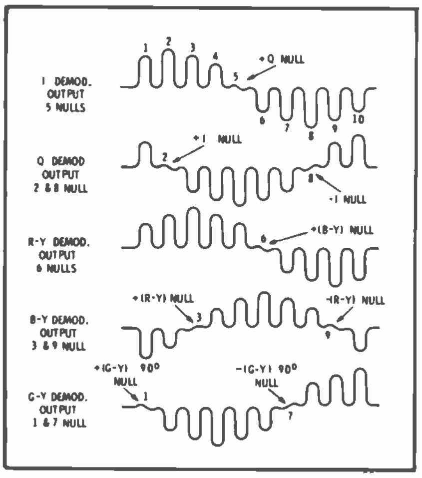


Fig. 9. Demodulator waveforms.

It is customary to observe the waveforms of the X and Z demodulators after their outputs have been amplified by the color-difference amplifiers. Under these conditions, the X and Z signals will be identical to the R − Y, B − Y, and G − Y signals because the points of observation are the same; i.e., the CRT grids. If the output of the X demodulator is observed directly, the waveform is similar to an inverted R − Y output waveform; but the maximum negative output will accur at a point between the third and fourth bars, the null is between the sixth and seventh bars, and the maximum positive output is between the ninth and tenth bars. Fig. 10 shows the output from the X and Z demodulators when a keyed-rainbow signal is fed into the receiver. Notice that the output of the Z demodulator is similar to the inverted B − Y waveform of Fig. 9, but that it is shifted slightly in phase (13.5°).

Color Difference Amplifiers

Referring again to Fig. 1, the outputs of the X and Z demodulators are fed to the R − Y and B − Y difference amplifiers, respectively. These amplifiers and their associated circuitry amplify and invert their inputs and feed them to the CRT control grids. It is common practice to connect the cathodes of the three color difference amplifiers together and to leave the cathode resistor unbypassed. Thus, a portion of both the R − Y and B − Y signals are mixed together and fed to the cathode of the G − Y amplifier. In addition, a portion of the R − Y amplifier output is fed to the grid of the G − Y amplifier. In this manner, R − Y and B − Y are combined to form G − Y. Also notice that, because of the common-cathode arrangement, some B − Y voltage is fed to the R − Y difference amplifier and vice-versa. The effect of this is to cause the output of the R − Y amplifier to actually be the R − Y voltage although the apparent input to the amplifier is the X-axis voltage. A similar action takes place in the B − Y difference amplifier.

As noted previously, not all manufacturers incorporate horizontal blanking in the chroma circuit. When horizontal blanking is used, the usual arrangement is to apply a positive pulse to the cathodes of the difference amplifiers to cut them off during horizontal retrace. In some sets, horizontal blanking is accomplished in the chroma bandpass amplifier.

A variation from the conventional low-level demodulator design was noted previously in this article. This is the circuit used by General Electric in which three separate demodulators (R − Y, B − Y, and G − Y) were used. Obviously, since the G − Y axis is demodulated, it is not necessary to derive this voltage from a matrix.

Referring to Fig. 2, notice that the outputs from the demodulators are of sufficient amplitude to drive the CRT. By definition, this is high-level demodulation. Since there are no color-difference amplifiers, it is somewhat more convenient to use the R − Y, B − Y, and G − Y axis for demodulation. The circuitry used by the three principal proponents of high-level demodulation (Admiral, Motorola, and Zenith) will be discussed later in this book.

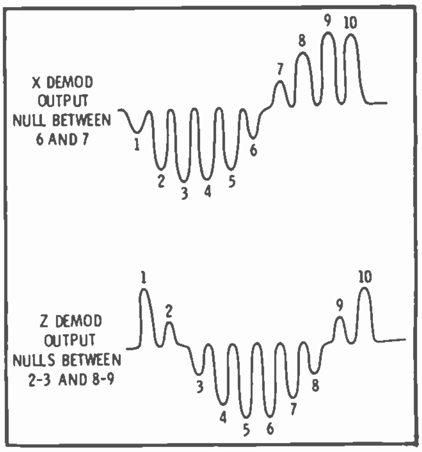


Fig. 10. Demodulator waveforms.

Chapter 5

Circuit Analysis of Chroma Circuitry

Circuit Analysis of Reference Oscillator Circuits

Fig. 1 shows the portion of the RCA chassis CTC25 chroma section which reconstitutes the 3.58-MHz reference signal. These circuits are the burst amplifier, chroma sync phase detector, chroma reference oscillator control, and the chroma oscillator. Two signals are applied to the control grid of V19, a positive gate pulse and the output of the first video amplifier. Because of the filters which are incorporated in the first video amplifier and the very small value of the coupling capacitor, C25, the low-frequency components of the video signal have been removed and only the chroma information remains.

Between pulses from the horizontal output transformer, which are fed to its grid, V19 is cut off by a positive pulse from the horizontal output transformer applied through a divider to the cathode. This pulse is integrated by C116, and the average cathode potential is maintained at about 35 volts. The positive gate pulse applied to the grid of V19 is sufficient to overcome the cathode bias and the tube amplifies the color burst which is applied at this same instant.

The output of V19 is developed across the primary of L31 which is tuned to 3.58 MHz. Transformer coupling is used between the burst amplifier and the phase detector to block the enabling pulse from the circuits which follow. Thus, the input to the phase detector, which comes from the burst amplifier, is eight or nine cycles of the 3.58-MHz signal which originated at the transmitter, and nothing else.

In the absence of a signal from the reference oscillator, the output of the phase detector at the junction of R173 and R174 is zero. Since the signals at the extreme

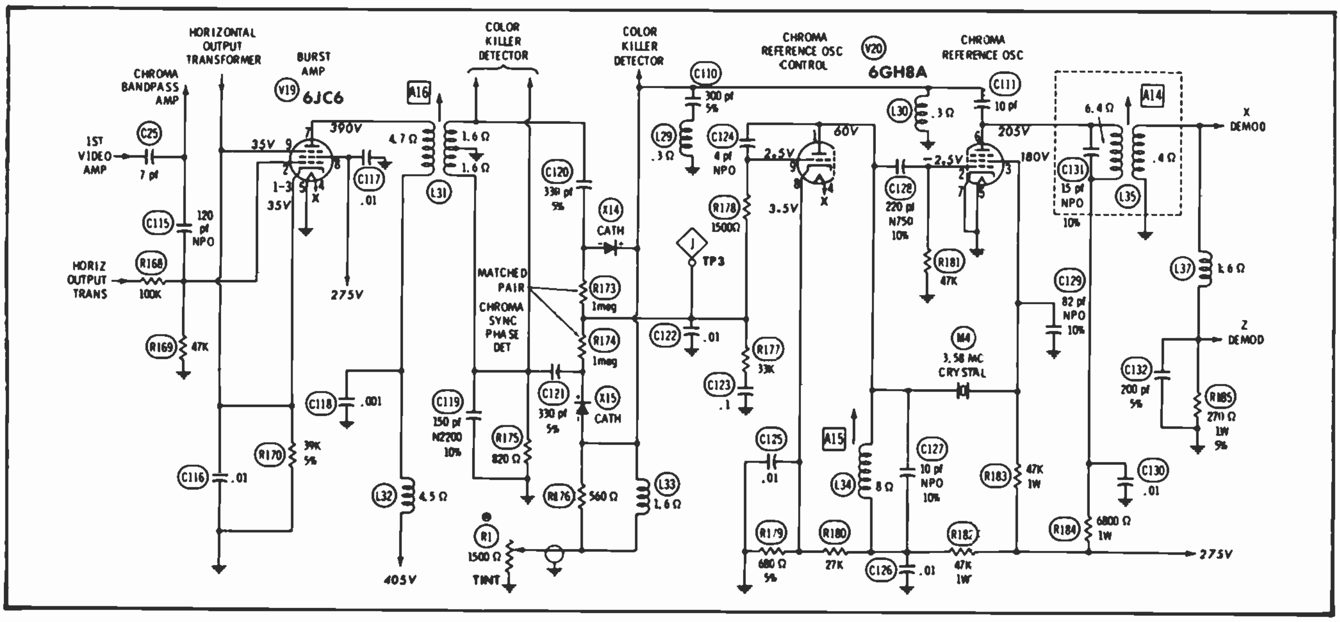


Fig. 1. Reference signal circuits of RCA Chassis CTC25.

ends of these resistors are equal and opposite, and the center of the secondary of L31 is grounded, the junction of R173 and R174 remains at ground potential during the half-cycle when the diodes are conducting—as well as when they are cut off.

Since the reference oscillator signal is fed to the cathode of X14 and the anode of X15, one diode is forward biased at the same instant that the other is reverse biased. During the half-cycle when the burst signal causes the diodes to conduct, it is possible for X14 to be reverse biased or forward biased depending on the phase of the signal from the reference oscillator. At the same instant, X15 will be biased in the opposite direction. Unless the reference oscillator is operating at the correct phase, the amounts of conduction in X14 and X15 are no longer equal and the voltages at the extremes of R173 and R174 are no longer equal and opposite. Therefore, the voltage at the junction of R173 and R174 can swing either positive or negative depending on the phase relationship of the reference signal and the color burst.

The network consisting of R176, L33, and R1 along with C110, C111, L29, and L30 determines the phase shift of the feedback signal from the reference oscillator to the phase detector. By changing the setting of R1, this phase shift may be varied. This generates an error signal in the phase detector which eventually changes the phase of the reference oscillator. The range of this control is adequate to shift the reference oscillator about 30° in either direction. This will shift a keyed-rainbow pattern one complete color bar from normal in either direction.

The output of the phase detector is integrated, or filtered, by C122, R177, and C123 and fed to the grid of the chroma reference oscillator control tube. This is essentially an AFC tube and the operation is very similar to that of the horizontal oscillator AFC tube discussed in Part 3 of this series. The principal differences are the frequency of operation and the fact that the DC component of the plate current flows through the oscillator tank circuit.

The reference oscillator is a crystal-controlled electron-coupled oscillator. The tuned circuit, L34 and C127, is tuned to the exact frequency of the color burst and the crystal stabilizes the frequency. Grid-leak bias is developed by C128 and R181. The output is developed across the primary of L35 which is tuned to the oscillator frequency. The output of the X demodulator is taken directly from the top of the secondary of L35. The reference signal for the Z demodulation is shifted in phase by L37, C132 and R185.

Color Killer

Fig. 2 is the schematic of the color killer circuit of the RCA CTC25 chassis. This circuit is immune to noise because it is actuated by the phase difference between the reference oscillator and the color burst rather than by the mere presence of the burst signal.

First, consider the operation of the circuit when no color burst is present. Notice that its operation is quite similar to a keyed AGC circuit. A positive pulse from the horizontal output transformer causes current flow through V14 to charge the right side of C103. Between pulses, C103 must discharge through R172 to ground, developing about −15 volts at the top of R172. The long time constant of C108 and R172 maintains this voltage at a steady level. This filtered voltage is used to hold the chroma bandpass amplifier in cutoff.

The actual amount of conduction of V14, and hence the amount of bias at the chroma bandpass amplifier, is determined by the amplitude of the positive pulse, also from the horizontal output transformer, applied to the grid of V14. The magnitude of this grid pulse is set by the color killer adjustment.

In the absence of the color burst, only the reference oscillator signal is applied to X12 and X13. Both diodes conduct the same amount and the output at the junction of R158 and R159 is zero. When a color burst is present, the conduction of X12 and X13 is unequal and a negative output appears at the junction of R158 and R159. This negative voltage opposes the positive pulse voltage at the grid of V14 and the tube remains in cutoff. As a result, no bias is developed in the plate circuit and the chroma bandpass amplifier is allowed to amplify the chroma signal.

Unlike the chroma sync phase detector whose output can be either positive or negative, the output of the color killer detector is always negative. This is true because the phase relationship between the reference oscillator output and the color burst is a constant in the color killer detector but a variable in the chroma sync phase detector.

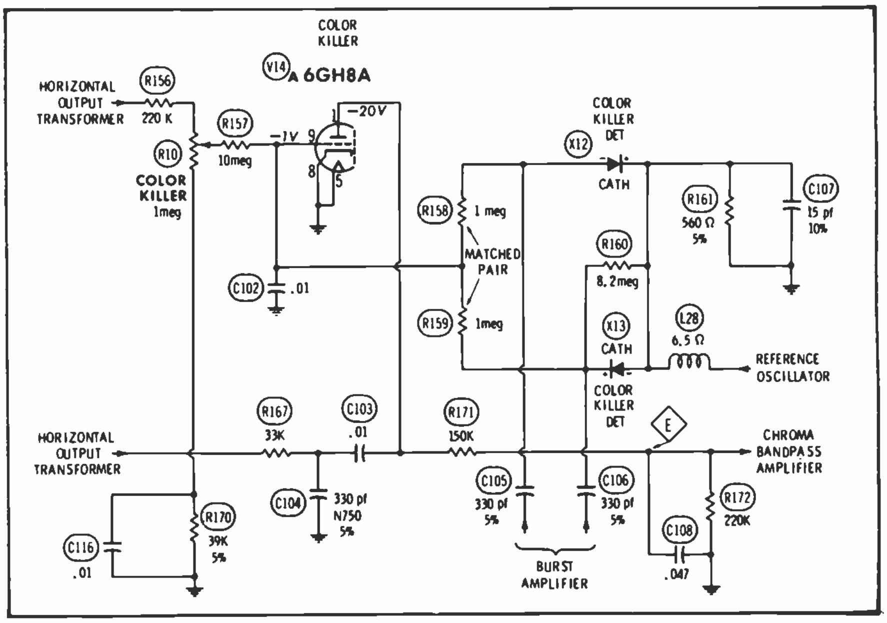


Fig. 2. Color Killer circuit of RCA Chassis CTC25.

Automatic Chroma Control

Another refinement of the chroma system has been incorporated in some sets. Called "automatic chroma control" (ACC), this circuit is essentially an AGC circuit for the chroma bandpass amplifier. A means of automatically controlling the gain of the chroma bandpass amplifier is desirable for two reasons:

1. The receiver AGC circuit is controlled by the amplitude of the horizontal sync pulses, not the color burst level. In theory, the relation between these two levels is constant, but this may not be true under all circumstances This condition is especially noticable in fringe areas. Accordingly, the level of the chroma signal may vary even though the sync pulse level is maintained fairly constant by the AGC.
2. No AGC circuit, or any other closed-loop correcting system for that matter, is 100% efficient. (If it were, no error signal could be developed.) Therefore, an "auxiliary" AGC circuit will maintain a more nearly constant output with varying inputs.

The ACC detector circuit used in RCA Chassis CTC21, CTC28, and CTC30 has about the same configuration as the color killer detector circuit shown in Fig. 2. The phasing is set so that an increase in the level of the color burst produces a negative-going output and a decrease in the level of the color burst produces a positive-going output. This output voltage is used to raise or lower the bias of the chroma amplifier to maintain a more nearly constant level of chroma signal. This helps to prevent changes in color intensity and precludes the owner having to readjust the color control everytime the input signal level changes.

RCA Closed-Loop ACC

The previously described ACC (Automatic Chroma Control) used in RCA chassis CTC21, 28, and 30 is an open loop system. That is, the output of the ACC circuit is not used to control the gain of the amplifier which feeds it. By contrast, the ACC circuit used in the RCA CTC31 chassis is a *closed loop* system. The loop is from the grid of the first chroma amplifier, through the burst amplifier, through the ACC amplifier, and back to the grid of the first chroma amplifier. Fig. 3 is a simplified schematic of this circuit.

The color burst as well as the chrominance information are amplified by the first chroma bandpass amplifier. The plate load is the primary of the double-tuned transformer and one of the outputs from the secondary is fed to the burst amplifier. The burst amplifier amplifies the color burst and injects it into the reference oscillator circuit to control its phase.

First, consider the circuit with no burst signal present. The oscillator operates at its natural frequency and develops approximately 3.5 volts of negative bias at its grid. This voltage is applied to the emitter of the ACC amplifier and a positive potential of about 35 volts is present at the collector. Because of the voltage drop across R739, the DC potential at the grid of the first chroma bandpass amplifier is about +5 volts. (This bias voltage may vary considerably from set to set.)

During color operation, the color burst from the first chroma bandpass amplifier is fed through the burst amplifier, which is gated on during horizontal retrace, to the grid of the oscillator. This signal increases the drive and causes the bias to increase to about −8 volts. This 5-volt change in voltage at the grid of the reference oscillator is amplified by the ACC amplifier transistor and causes a 31-volt swing at its collector. The normal collector voltage is about 4 volts when a nominal 80-volt burst signal is applied to the oscillator grid. The bias voltage at the grid of the first chroma amplifier is approximately −5 volts under these conditions. If the amplified color burst signal increases in amplitude, the grid of the reference oscillator becomes more negative and the emitter current increases. This, in turn, causes the collector current to increase and the collector potential to swing in a negative direction. Finally, this negative-going voltage is used to increase the bias of the first chroma amplifier and reduce its gain.

Conversely, if the amplified color burst decreases in amplitude for any reason, the emitter and collector voltages of the ACC amplifier become less negative, decreasing the bias on the first chroma amplifier and increasing its gain. Thus, the burst amplitude at the grid of the oscillator is maintained at a constant 80 volts. This is the optimum level to properly phase the reference oscillator. Since the first chroma bandpass amplifier also amplifies the chrominance signal, it, too, is maintained at its optimum level, This, of course, is the more important function of the ACC circuit.

The principal advantage of the closed-loop ACC circuit is its ability to maintain a more nearly constant level of chrominance signal. The

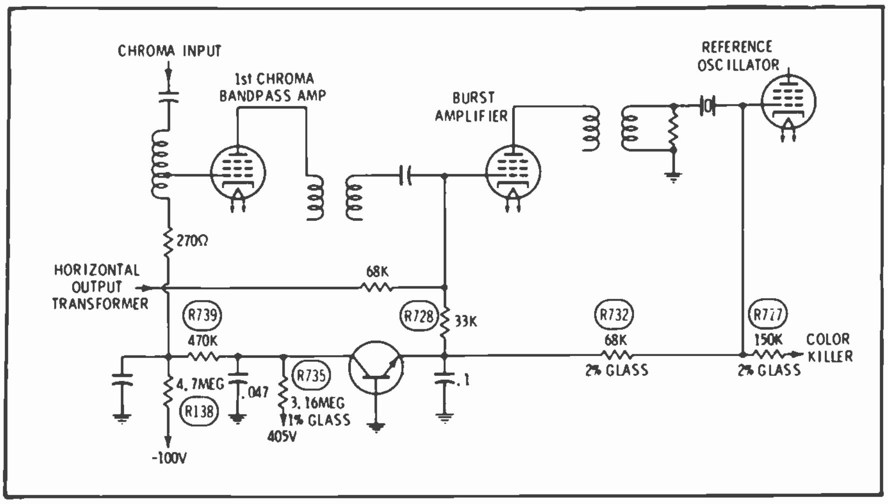
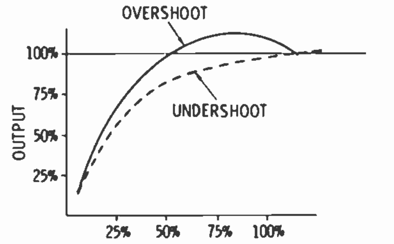
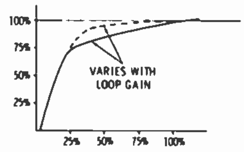


Fig. 3. Simplified ACC circuit of the RCA CTC31 chassis.

curves shown in Fig. 4 demonstrate the characteristics of the two types of control, open-loop and closed-loop. Bear in mind that these curves illustrate the characteristics of the two basic systems and do not apply to any specific circuits.



A. Open-loop system.



B. Closed-loop system.

Fig. 4. Characteristics of open-loop and closed-loop control systems.

Notice that R732, R777, and R735 are quite critical as to value, drift characteristics, and temperature coefficient. For this reason, glass resistors having close tolerance and low temperature coefficients are used.

RCA Color Killer

Incorporation of closed-loop ACC made it necessary to revise several other circuits in the RCA chroma system. Since the color burst is amplified by the first chroma amplifier, color-killer bias had to be fed to a different stage; the demodulators were chosen. The use of transistors in the ACC and color killer is also a significant departure from earlier RCA designs.

A simplified schematic of the color-killer circuit used in the RCA CTC31 chassis is shown in Fig. 5. The base voltage of the killer transistor is established by the setting of the killer control and the potential at the grid of the reference oscillator. Under no-color conditions, the base potential is about .5 volt positive with respect to the emitter and the transistor is cut off. Since the transistor is cut off, the collector voltage is determined by the voltage divider, R737 and R749, connected between the blanker grid and ground. Since the blanker grid is about −100 volts, the collector voltage is about −25 volts. This voltage is also present on the screen grids of the demodulators and keeps these tubes below cutoff.

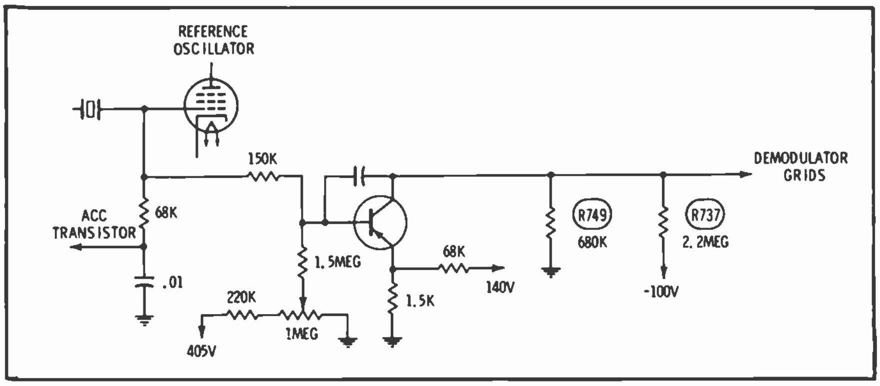


Fig. 5. Simplified color-killer circuit of the RCA CTC31 chassis.

When a color burst is applied to the reference oscillator, the grid swings negative and the killer transistor is driven into saturation. This clamps the collector voltage to the emitter potential, raising the screen voltage of the demodulators to about 2 volts, well above cutoff. Notice that the color-killer transistor operates either at saturation or cutoff. Thus, variations in color-burst amplitude have no effect on the bias supplied to the demodulators.

RCA Reference Oscillator

The RCA CTC31 chassis uses an injection type reference oscillator which is similar to the one used in the CTC18, CTC20 and CTC24 chassis. Fig. 6 is a simplified schematic of the oscillator used in the CTC31. Basically, the oscillator is of the tuned-plate, tuned-grid, electron-coupled type. The frequency is determined by the crystal in conjunction with the small trimmer capacitor shunted across it.

As with any TPTG oscillator, the oscillator plate tank (L704) is tuned slightly above the oscillator frequency. In this circuit, it is adjusted so that the self-bias developed at the oscillator grid is −3.5 volts with no burst signal applied. In earlier models using the injection oscillator, the counterpart of L704 was not adjustable. In the absence of a color burst, the oscillator runs at the reference frequency, but with a random phase. When the burst is injected through T702, this signal pulls the oscillator into phase with it. The oscillator is stable enough to remain properly phased until the arrival of the next burst.

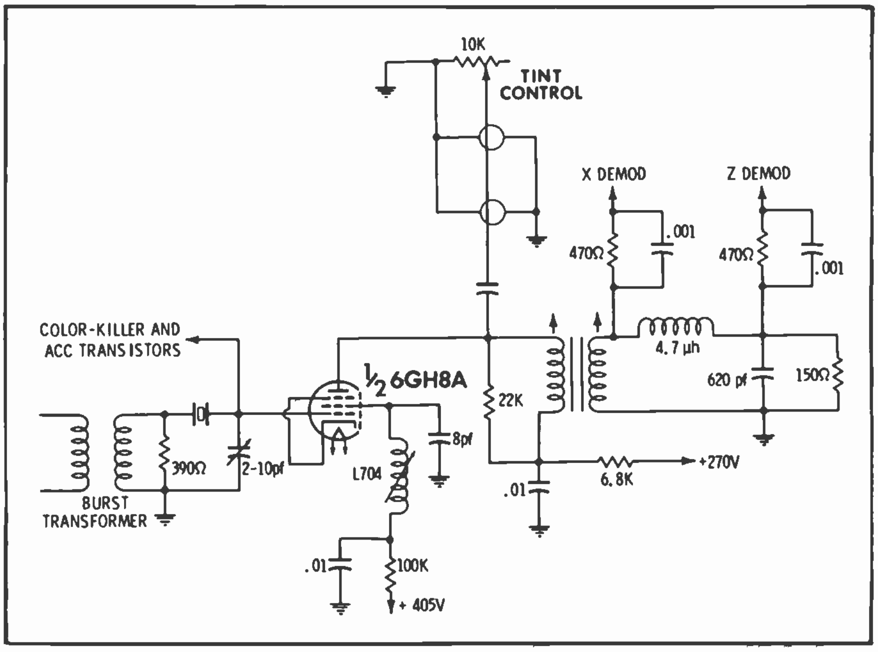


Fig. 6. Simplified reference-oscillator circuit of the RCA CTC31 chassis.

Admiral Burst Amplifiers and Reference Oscillators

Fig. 7 shows the burst amplifier and reference oscillator circuit of the Admiral 1G1155 and other chassis. The more recent chassis, 3H10, 4H10, 5H10, and 4H12, use essentially the same burst amp and oscillator circuitry except for the tube type. The chroma signal from the plate of the first chroma bandpass amplifier and a positive enabling pulse from the horizontal output transformer are both fed to the burst-amplifier grid. The enabling pulse, having an amplitude of 50 volts, brings the tube out of cutoff and the color burst is amplified to a peak-to-peak amplitude of about 170 volts. The large value of cathode resistance, 39K ohms, prevents saturation. During the time that V5 is conducting, the drop across R166 charges C120. Between pulses, current flows upwards through R166 to discharge C120, maintaining an average cathode bias of about +45 volts. This prevents the chrominance signal from appearing in the plate circuit of the burst amplifier.

The reference oscillator is an injection type, electron-coupled oscillator and its operation is similar to that of the RCA circuit discussed above. The self-bias under free-running conditions (no color) is —.3 volt. When a burst signal is injected, this bias swings negative and the negative excursion is used to operate the color killer and the ACC circuit. The output of the reference oscillator is coupled through the plate transformer, L35, to the phase shifting circuits, L36, C129, and L37, which establish the correct phase of the reference signal for R − Y and B − Y demodulation. C128 and R6, the tint control, are used to shift the phase of the reference signal without disturbing the phase displacement between the R—Y and B—Y axes.

Admiral ACC and ASC Circuits

As shown in Fig. 7, the ACC circuit used in the chassis series 1G1155-1, 2G1156-2, 2G1157-1, 3G1155-2, and 3G1155-3 has two variations. In the solid-line drawing, the ACC control voltage is taken from the grid of the reference oscillator and, after filtering, is used as bias voltage for the first chroma amplifier. Since the oscillator grid swings more negative as the color burst increases in amplitude, the chroma-amplifier gain is reduced as the level of the composite chrominance signal (chroma and color burst) increases. The operation of the dashed-line circuit is much the same, although a diode detector has been added. Since this addition increases the amount of the control voltage, a divider network, 180K ohms and 270K ohms, is also added to the circuit.

The ASC (Automatic Saturation Control) circuit used in the 3H10NC57-1 chassis is shown in Fig. 6. This circuit, as well as the one described above, is a closed-loop system. A portion of the output from the first chroma amplifier is rectified by X11 and the negative voltage which is produced is used, after filtering, to control the gain of the first chroma amplifier. Thus, as the chrominance level increases, the bias increases to reduce the amplifier gain and vice versa. If this were the only control voltage, the dynamic range of saturation would be seriously limited (the degree of color saturation of the picture would remain constant regardless of the degree of saturation of the scene being televised). To prevent this, a second voltage is combined with the bias derived from X11. The negative voltage at the grid of the reference oscillator, which is proportional to the color-burst amplitude, is also fed to the grid of the first chroma amplifier. Thus, sufficient bias is always available to maintain the desired dynamic range of saturation.

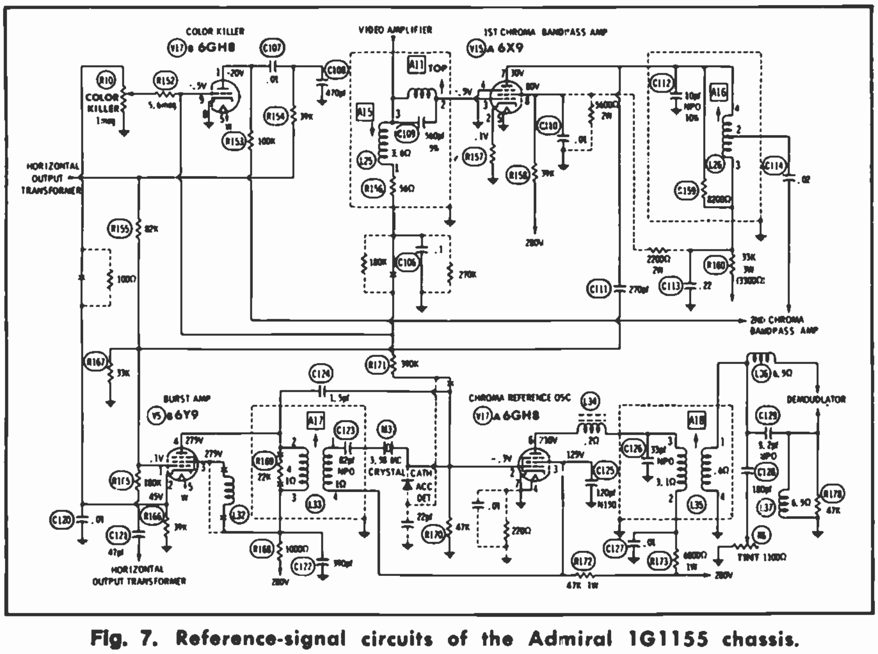


Fig. 7. Reference-signal circuits of the Admiral 1G1155 chassis.

Admiral Color Killer

The color-killer circuit shown in Fig. 8 is typical of many of the circuits used in late-model Admirals. The positive cathode bias of the burst amplifier is divided across the threshold control, R9, and negative voltage is obtained from the grid of the reference oscillator. In the absence of a color burst, this negative voltage is slight and the color-killer tube will conduct if plate voltage is supplied. The source of plate voltage is the positive pulse from the horizontal-output transformer which is fed to the left side of C128. Current flows through V15A, charging the right side of C128 to a negative potential. Between pulses, C128 partially discharges through R170 and the negative voltage which is developed holds the second chroma amplifier below cutoff. When a color burst is received, the negative voltage at the grid of the reference oscillator increases and cuts off the color-killer tube. This allows C128 to completely discharge and the cutoff bias is removed from the second chroma amplifier.

Notice that the setting of R9 affects the bias of the first chroma amplifier and, if R9 is misadjusted, the operation of the ASC circuit will be impaired. To properly set R9, adjust all front-panel controls for

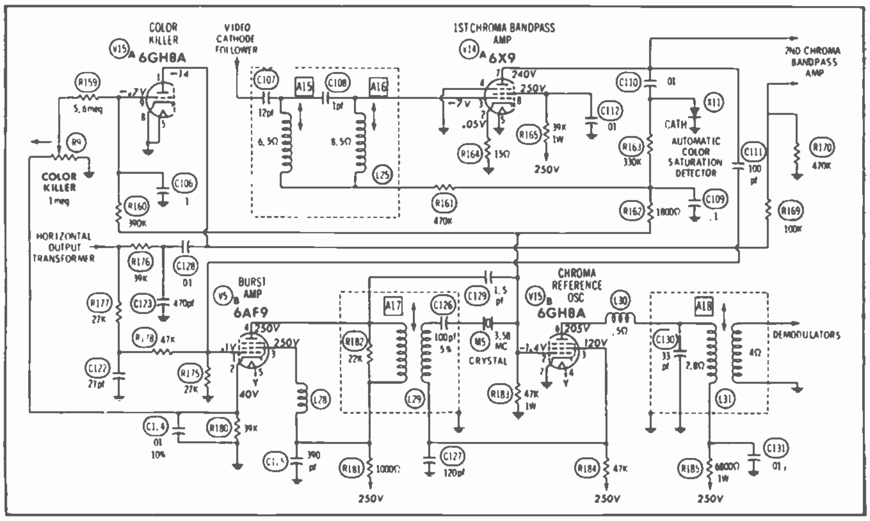


Fig. 8. Reference-signal circuits of the Admiral 3H10NC57-1 chassis.

the chroma reference-oscillator control tube. The operation of this type of circuit was explained in Part 3 of this series.

The reference oscillator is typical of the type of oscillator used in conjunction with an AFC tube. Since the system of chroma demodulation used by Zenith requires four reference signals in quadrature, a special output transformer is used instead of the usual RLC phase-splitter network.

Zenith Color-Killer and ACC Circuits

The color-killer and ACC circuit of the Zenith 20X1C36 chassis is also shown in Fig. 9. The color-killer and ACC detector is a conventional phase detector, but, since the phase relationship of two inputs is constant, the amplitude of the output becomes a function of the amplitude of the color burst. When no burst is present, the output is −.7 volt, but, during normal color reception, this potential increases to approximately −6 volts. If the amplitude of the color burst decreases from its normal value for any reason, the detector output also decreases.

The output of the detector is filtered by C64 and used as bias for the grid of the first chroma amplifier, V4B. Under no-color conditions, V4B is near saturation and the screen potential is about 75 volts. This voltage is at one end of a series network consisting of R165, R17, and R164. The opposite end of R164 is connected to the grid of

proper operation and set the color control at mid-range. Turn to an unused channel, set the color-killer control fully clockwise, and then adjust it until the color in the snow almost disappears.

Zenith Burst Amplifier and Reference Oscillator Circuit

The chroma-reference circuits of Zenith's 20X1C36 and 20X1C38 are shown in Fig. 9. In many respects, they are similar to the circuits of the RCA CTC25 chassis discussed under the heading "Circuit Analysis of Reference Oscillator Circuits" in Part 5. The composite chrominance signal from the plate of the first chroma amplifier and a positive enabling pulse from the horizontal-output transformer are fed to the grid of the burst amplifier. Since the color burst is coincident with the enabling pulse, the burst is separated from the remainder of the chrominance signal and amplified. The positive cathode bias of about 45 volts is developed by conduction through R176 while the tube is gated on, and this voltage is sustained between pulses by the charge stored in C127.

The output of the burst amplifier is fed to the chroma-sync phase detector and to the ACC and color-killer detector. The hue control, R3, in conjunction with C131 allows the viewer to vary the phase of the amplified color burst.

The chroma-sync phase detector compares the relative phases of the reference oscillator signal from L30 and the color burst from the burst amplifier. Any phase error is converted to a voltage error which is used to change the conductance of

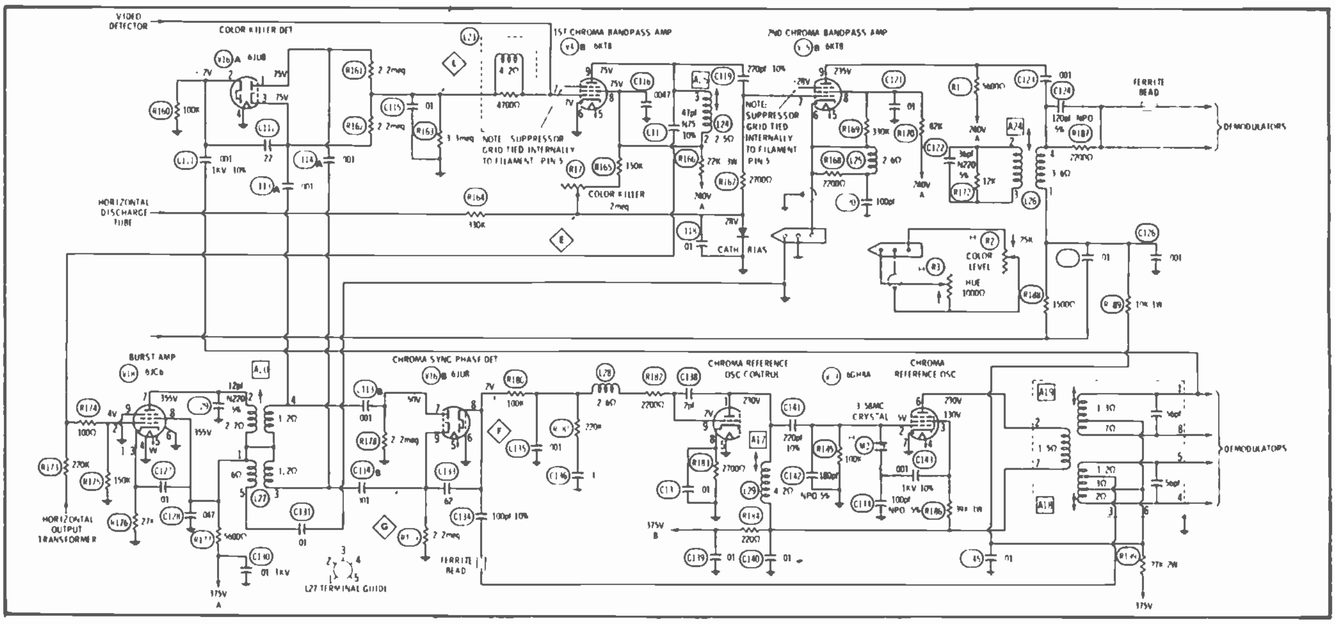


Fig. 9. Reference-signal circuits of the Zenith 20X1C36 chassis.

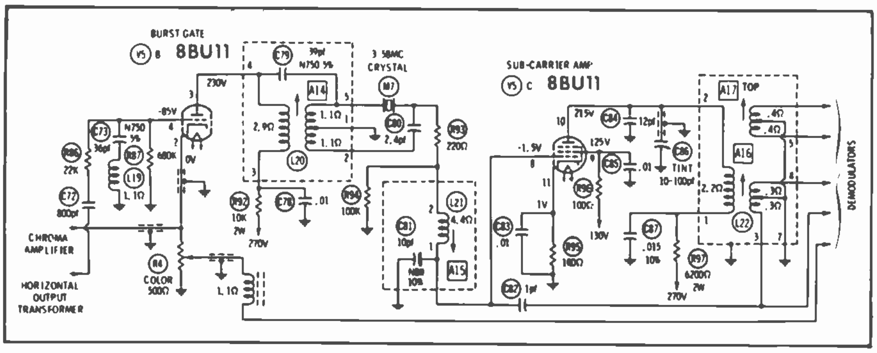


Fig. 10. Burst gate and subcarrier amplifier of the General Electric HC chassis.

the horizontal discharge tube which is 65-volts negative. When R17 is properly adjusted, the voltage at its junction with R164 is about −28 volts. This voltage is used to bias the second chroma amplifier below cutoff. C118 is a bias filter which integrates the horizontal pulses from the horizontal discharge tube.

During color reception, the grid bias of V4B increases to −6 volts and the screen and plate voltages rise to 225 volts. This would cause the voltage at the grid of the second chroma amplifier to rise to a positive potential if it were not for the clamper diode connected across C118. The actual bias of the second chroma amplifier is 0 volt.

As stated before, the output of the color-killer and ACC detector is −6 volts under conditions of normal color reception. If the chrominance level varies from its normal value, the detector output will also change. Thus, a decrease in chroma level reduces the negative bias on V4B, increasing its gain. Conversely, an increase in the chroma level increases the bias on V4B to reduce its gain. Notice that small variations in bias on V4B do not affect the bias of the second chroma amplifier because of the action of the clamper diode. This, too, is a closed-loop system.

General Electric Reference Circuits

Fig. 10 shows the burst gate and subcarrier amplifier circuits of the General Electric HC chassis. The output of the chroma-bandpass amplifier is fed to the cathode of the burst gate tube, V5B, and the 100-volt enabling pulse from the horizontal-output transformer is fed to the grid. This allows the color burst to be separated from the composite chroma signal and amplified. The positive enabling pulse causes the grid of V5B to draw grid current, charging C72 and C73. Between pulses, these capacitors discharge through R87, developing about 85 volts of bias. This bias, of course, holds the tube below cutoff between pulses.

The output from the burst-gate tube is coupled through L20 and excites the 3.58-MHz crystal, causing it to ring. Because of the high Q of the crystal, this ringing continues throughout the interval between color bursts. Each successive color burst rephases the crystal if there has been any drift. The amplitude of the ringing signal at the grid of V5C is large enough to overdrive the tube, and thus the output remains constant throughout the interval between bursts.

C86, connected between the plate of V5C and ground, shifts the phase of the reference signal to provide tint control. Quadrature reference signals are required, so a transformer having two secondaries is used as the plate load of V5C.

No color-killer circuit, as such, is used in this chassis. Since the demodulators have no output unless there is a reference-signal input, and since the 3.58-MHz crystal "rings out" if there is no color burst, the modulators are, in effect, cut off during b-w operation.

Motorola Reference Circuits

The burst amplifier, chroma sync amplifier, color killer, and demodulator of the Motorola A22TS-918A are depicted in Fig. 11. The chroma cathode follower (not shown) drives both the chroma bandpass amplifier and the burst amplifier. An enabling pulse from the horizontal-output transformer turns on the burst amplifier, V16A, during the horizontal retrace interval, allowing the color burst to be separated from the composite chrominance signal.

The interstage transformer between V16A and V16B is tuned to the burst frequency. The network consisting of R4, L31, and C149 is a phase-shifting network which allows the phase of the burst to be adjusted for correct hue. V16B further amplifies the color burst and feeds it, via the 3.58-MHz crystal, to the chroma-demodulator tube. Notice that the positive pulse applied to the screen grid of V16B

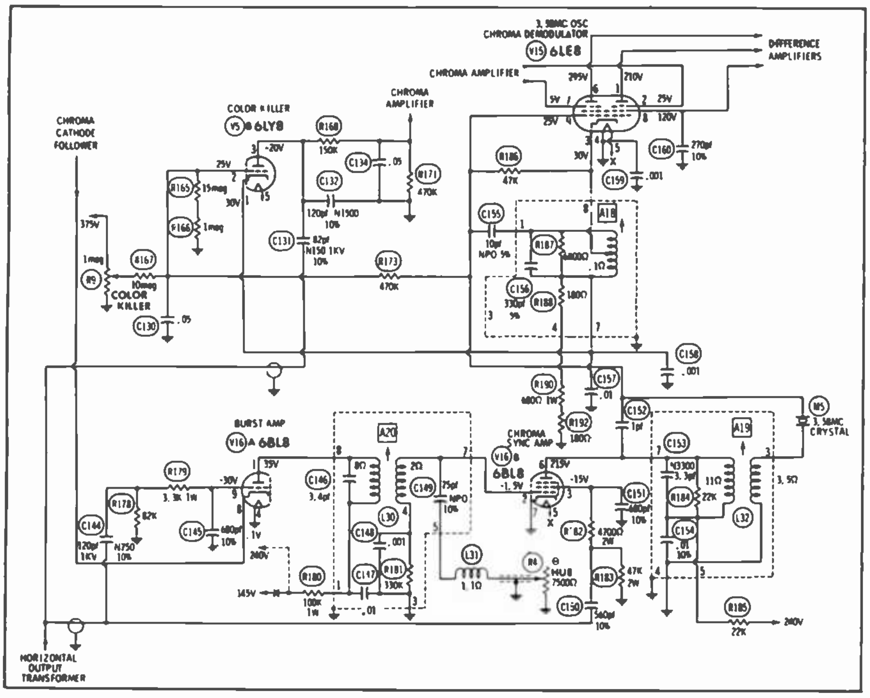


Fig. 11. Reference-signal circuits of the Motorola A22TS-918A chassis.

gates this tube on during the horizontal retrace interval only.

The chroma-demodulator tube not only demodulates the chroma signal, but serves as the reference oscillator as well. This combination of both functions in a single tube was not noted in any of the other makes of sets examined. Since this particular portion of the color training series is limited to reference-signal and associated circuits, the method of demodulation will be discussed at a later time. At present, we will consider only the functions of the cathode, control grid, and screen of V15.

Consider V15 as an electron-coupled Hartley oscillator. In the absence of color bursts, oscillations are sustained by virtue of the split-inductance tank typical of a Hartley oscillator. During color reception, the amplified color burst is injected through the crystal to the grid of V15 and rephases the tank circuit to synchronize it with the burst signal. In this respect, the oscillator is similar to the injection-locked oscillator used in a number of other sets.

Since the oscillator is an integral part of the demodulator circuit, there is no way to split the oscillator phase prior to demodulation. As we shall see later, chroma demodulation may be achieved so long as the phase of either the reference signal *or* the chroma signal is split. Motorola's decision to split the phase of the latter instead of the former is unique but equally acceptable.

The color-killer circuit is similar to the ones used in many of the sets discussed in the preceding pages. In the absence of color, a pulse from the horizontal-output transformer causes V5B to conduct, charging C131 and producing cutoff bias for the chroma amplifier. During color reception, the cathode current of V15 increases (because of the increased oscillator activity) and this drives the cathode of V5B positive into cutoff. Since C131 cannot charge when V5B is cut off, the bias is removed from the chroma-bandpass amplifier.

Since the function of the chroma bandpass amplifier (or amplifiers) is simply one of increasing the level of the chrominance signal, there is very little to be said about it. The bandpass considerations were discussed in Part 4 of this series. The means by which it is gated by the color killer and controlled by the ACC circuit were covered in Parts 4 and 5.

Malfunction in the chroma bandpass amplifiers will normally result in insufficient color or no color at all. If the gain of the amplifier is reduced, color saturation will be decreased; if the amplifier fails completely, there will be a complete loss of color. Improper alignment of the amplifier will cause color smearing or "grainy" color, but alignment problems are more likely to be the result of tampering than drift. Realignment should be attempted only if the necessary test equipment is available. "Eyeball" alignment will usually result in further degradation of picture quality.

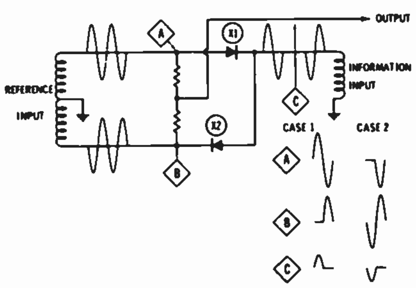


Fig. 12. Simplified phase-sensitive detector using diodes.

Diode Chroma Demodulators

In essence, a chroma demodulator is simply a phase-sensitive detector —no more, no less. Although phase-sensitive detectors (PSD) have been used in the majority of b-w receivers built in the past 20 years, the use of a pair of phase detectors to extract the color-difference signals from the chroma sidebands seems to excite a great deal of interest. Thus, a thorough discussion of each of the four popular types of chroma demodulators is included here. Since the circuit which utilizes diodes is perhaps the most easily explained, we will begin this discussion with it.

Fig. 12 shows a simplified phase-sensitive detector and associated waveforms. The input from the reference transformer is constant, while the phase of the information input varies. Cases 1 and 2 show the information signal at two possible phase angles. In case 1, diode X1 cannot conduct since the instantaneous cathode and anode voltages are equal throughout the cycle. Consequently, the voltage at point A is a simple sine wave whose first excursion is positive.

During the first half of the cycle, X2 conducts because its cathode is negative with respect to its anode, and the instantaneous voltage at point B is equal to the sum of the two applied voltages. These are equal in amplitude, but opposite in polarity; the potential at point B is 0 during the time that X2 is conducting. During the second half-cycle, X2 is cut off because the cathode is positive with respect to the anode. The voltage at point B is the positive excursion of a sine wave.

The voltage at point C, the output, is the sum of the instantaneous voltages at points A and B. During the first half of the reference sine wave, the voltage at point A completes a positive half-cycle while the voltage at point B is clamped to zero. Therefore, the voltage at point C is a positive half-cycle having an amplitude equal to one-half the amplitude of the half-cycle at point A. During the second half of the sine wave, the voltages at points A and B are equal in amplitude but of

Fig. 13. Typical output curve for a phase-sensitive detector.

opposite polarity, and the voltage at point C is zero. Thus, if the input signals are in phase, the circuit functions as a half-wave rectifier with a positive output.

Case 2 shows the instantaneous voltages that are present when the information signal is 180° out of phase with the reference signal. Now it is X2 which never conducts, and X1 acts as a half-wave rectifier. But, X1 is connected in the opposite polarity from X2, so its output is negative instead of positive. Thus, if the input signals are out of phase, the circuit functions as a half-wave rectifier having a negative output.

Additional waveforms to show the output of the PSD for intermediate phase relationships could be included, but they add little to the discussion. Instead, Fig. 13, showing the output for various phase relationships, is presented. Notice that the curve has the form of the familiar sine wave.

The polarity of the output from a PSD may be reversed by two means: reversing the diodes or reversing either of the transformers that supply the signals. Another characteristic which is of particular interest is this: With the exception of the positive and negative maximums, any output amplitude (including zero) may be the result of two different phase relationships. These statements become meaningful when the PSD we have been discussing is renamed a chroma demodulator and placed in a TV set.

Since the phosphors used in a color CRT are red, blue, and green, a minimum amount of circuitry will be used if one demodulator produces a maximum output when a red chroma signal is received. If we arrange the circuits preceding the demodulator so that a chroma signal representing red reaches the demodulator *in phase with the reference signal,* we have a red (R − Y) demodulator. (The red axis is at 76.6° and the R − Y axis is at 90°. This discrepancy is more apparent than real since the color difference amplifiers shift the axis slightly and also because the various red phosphors in use have slightly different colors. Depending on these variables and the position of the "tint" control,the true axis of operation of the R − Y demodulator may be any angle from, perhaps, 60° to 120°).

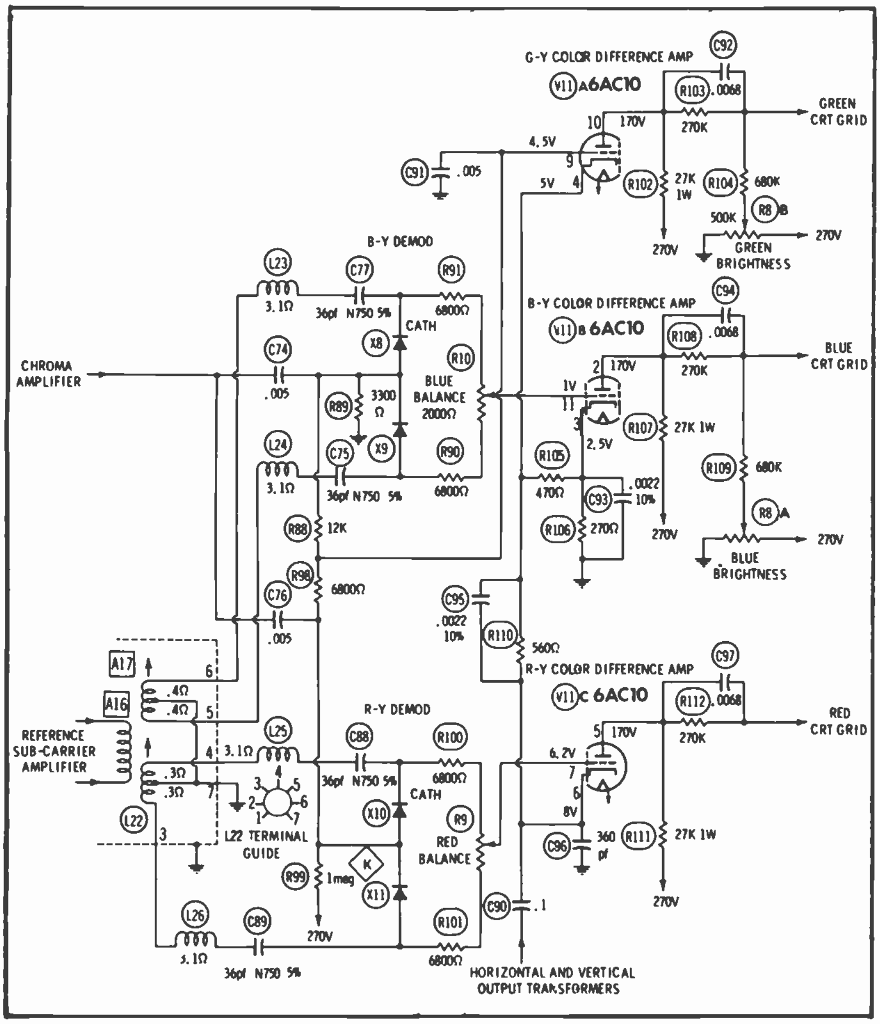


Fig. 14. Chroma demodulators and color-difference amplifiers of the General Electric HC Chassis.

In actual practice, the R − Y demodulator of a TV receiver may be followed by an amplifier. This inverts the signal so that the R − Y demodulator output must be maximum negative, instead of maximum positive, to produce red on the CRT. To be absolutely correct, we must refer to an R − Y demodulator followed by a difference amplifier as a −(R − Y) demodulator.

A second demodulator might well be connected so that its maximum outputs occur when the chroma signal is "all blue" and "no blue." Notice that we may cause this demodulator to have either a positive or negative output (for an "all-blue" signal) merely by reversing the phase of the reference signal.

Fig. 14 shows the chroma demodulators and difference amplifiers of the General Electric HC chassis. Consider the R − Y demodulator. The phases of the two inputs are such that a "red" (R − Y) chroma signal produces a negative output. This output is fed to the R − Y color-difference amplifier where it is amplified and inverted. The output from the R − Y difference amplifier is the R − Y signal which is fed to the red control grid of the CRT gun.

The B − Y demodulator is identical to the R − Y demodulator but the phase of the reference signal has been changed. In the B − Y demodulator, a "blue" (B − Y) chroma signal produces the maximum negative output. This is amplified and inverted in the B − Y color-difference amplier and finally appears as a positive signal at the blue grid of the CRT, turning on the blue gun and causing a blue field.

It was stated previously that a specific amplitude and polarity of output from a demodulator may be the result of either of two phase relationships. For example, observe from Fig. 13 that phase relationships of 160° and 200° each produce a negative output which is 94% of the 180° output. However, since the reference signal applied to the B−Y demodulator is shifted 90° from the reference signal at the R−Y demodulator, the 160° signal at the R−Y demodulator becomes a 70° signal at the B−Y demodulator. Again referring to Fig. 13, this signal will produce an output from the B−Y demodulator which is positive with an amplitude that is 34% of maximum. By the same token, the 200° signal at the R−Y demodulator becomes a 110° signal at the B−Y demodulator, and a negative output with an amplitude which is 34% of maximum is produced.

From the above, it is apparent that even though two chroma signals having different phase angles can produce the same output from a single demodulator; when these same chroma signals are fed to the second demodulator, they cause radically different outputs. Therefore, any phase of chroma signal may be described in terms of the outputs it produces from the two demodulators. Stated another way, two demodulators are sufficient to extract all of the color information from the chroma signal.

In spite of this, three color difference signals are necessary to operate the color CRT because the colors from three phosphors are required to produce all the visible hues. There are two methods of producing control voltage for the excitation of the third phosphor. A third demodulator operating on the color axis of the third phosphor (green) may be used, or this voltage may be derived from the outputs of the R−Y and B−Y demodulators. This latter method is more popular although the former method is often used.

The generation of the G−Y signal by combining portions of the R−Y and B−Y signals is the method used in the General Electric HC chassis shown in Fig. 14. Since the signals at the plates (pins 5 and 2) of V11 are the R−Y and B−Y color-difference signals, respectively, the grid signals of these two triodes are the −(R−Y) and −(B−Y) signals. These signals also appear at the cathodes of the respective triodes since C96 and C93 have significant impedance at the frequencies contained in the color difference signals. (At .5 MHz, C93 has an impedance of 145 ohms.)

In part 1 of this series it was stated that

$$G-Y = -.51(R-Y) - .19(B-Y)$$

Thus, by combing suitable portions of the cathode signals of V11B and V11C, a G−Y signal is fed to the cathode of V11A. Since the V11A is operated as a grounded-grid amplifier, there is no inversion and the G−Y signal at its plate may be connected directly to the green grid of the CRT.

Since the cathode-to-ground resistance of each section of V11 is different, the bias present on each cathode is also different. To develop the desired bias on each section of V11, the grids are returned to a bias-bleeder network, consisting of R89, R88, R98, and R99, which is connected between ground and the B+ supply.

R8A and R8B are the blue and green brightness controls, respectively. Since the red phosphor is the least brilliant, the red grid is operated at the maximum positive potential and no adjustment is required. R8A and B are used to set the blue and green conduction to produce reference white.

Blanking of the CRT during horizontal and vertical retrace is accomplished by the signals fed through C90 to the cathodes of V11. These signals are negative pulses taken from the horizontal and vertical output transformers. Since there is no signal inversion in a driven-cathode amplifier, the amplified pulses at the plates of V11 are negative and cut off the three guns of the CRT.

Low-Level Triode Demodulators

Any attempt to prove that the use of the diode phase-sensitive detector is inferior or superior to the use of amplifying devices in a phase-sensitive detection system is inconclusive. Each system has inherent advantages and limitations. The use of amplifying devices (tube or transistor) has the advantage that some gain is contributed to the system, but increased complexity and lowered gain-stability is the price of this amplification. Similar arguments apply in choosing from among the various types of amplifying detectors: triode, pentode, sheet-beam, and twin-pentode. The use of transistorized demodulators offers some interesting possibilities, although no such circuits were used in the 1967 and early 1968 product lines that we examined.

The demodulator and difference-amplifier circuits used in the Philco 17QT85A chassis are shown in Fig. 15. Although pentodes are used as demodulators, they are connected as triodes, the plates and screens being tied together. Fig. 16 is a plot of the combined plate and screen currents of the pentode section of a 6BL8 when it is connected as a triode. The circuit used in developing the curve was quite similar to the Z demodulator of Fig. 15 although bias was applied to the grid instead of the cathode. This has the effect of increasing the plate potential 8 volts, which effectively increases the steepness of the curve a slight amount.

In the Z demodulator (V14A of Fig. 15) the reference signal which drives the cathode has a p-p amplitude of 4.5 volts. In the absence of any other signal, the conduction of V14A and V15A produces a positive bias of 8 volts on each tube. Since this is the approximate cutoff bias of the tubes, V14A conducts only during the negative half-cycles of the cathode signal. At the instant when the reference voltage is peak negative, the tube bias is reduced to 3.5 volts and the tube current is about 8.2 ma. The unfiltered plate waveform would consist of negative-going half-cycles. However, the pi filter (C145, L35, and C146) integrates the pulses, and the voltage at the output of the pi filter is at a DC level of 235 volts.

When a chroma signal is fed to the grid of V14A, the instantaneous current is determined by the instantaneous grid-cathode voltage. This voltage is the sum of three potentials: (1) the cathode bias, (2) the instantaneous reference-signal voltage, and (3) the instantaneous chroma-signal voltage. For example, if the chroma and reference signals are 180° out of phase, the control grid is nominally 1.25 volts positive at the same instant that the cathode potential is + 3.5 volts (bias less

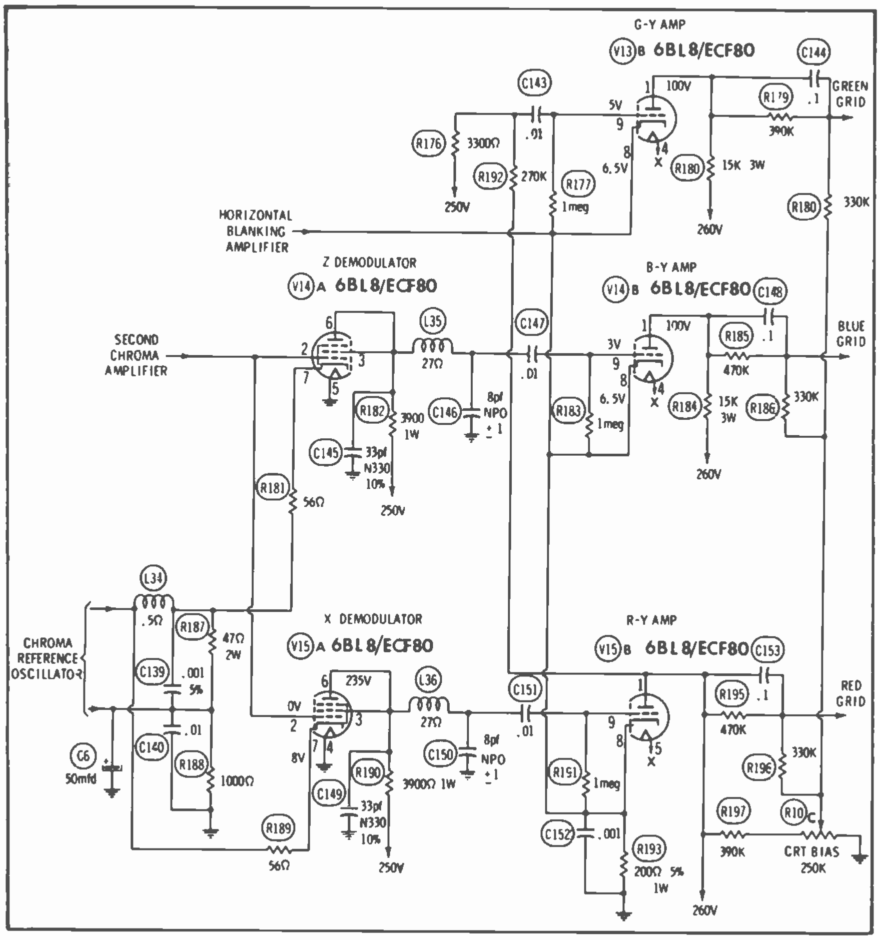


Fig. 15. Chroma demodulators and color-difference amplifiers of the Philco 17QT85A chassis.

the peak reference signal of 4.5 volts.) Thus, the total grid-cathode voltage is -2.25 volts. From Fig. 5, the tube current is about 12.7

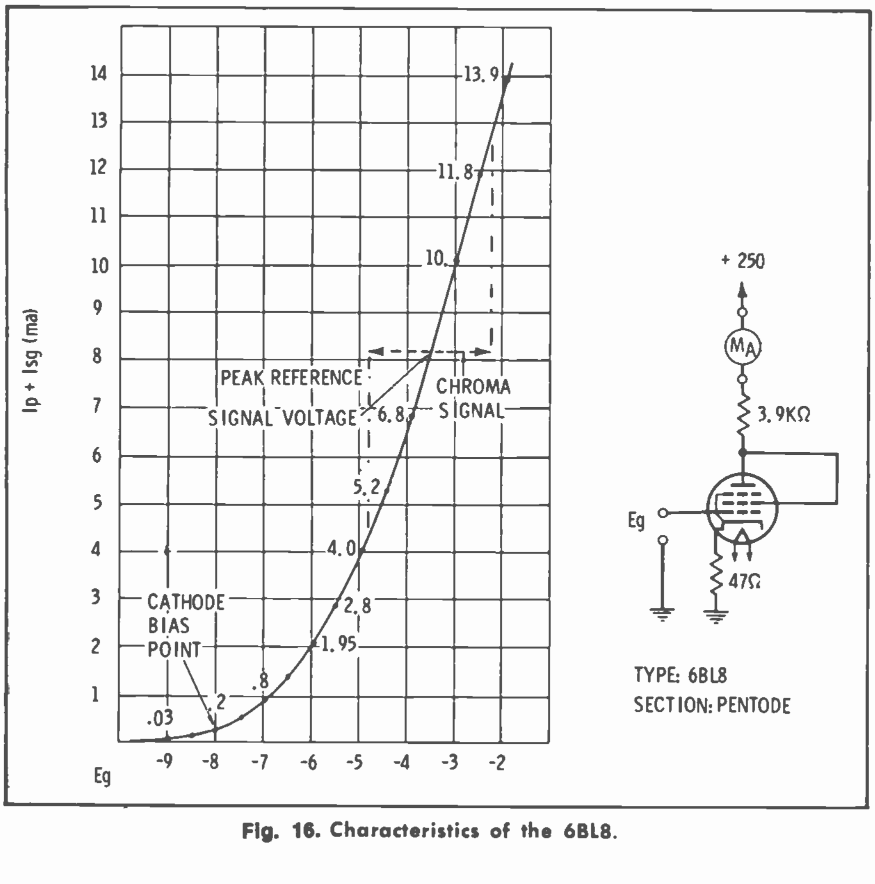


Fig. 16. Characteristics of the 6BL8.

ma at this instant. This is an increase of about 4.5 ma from the peak current with no chroma signal applied. After a number of cycles of voltage, the plate voltage will stabilize at a new voltage which is less positive than it was when no chroma signal was applied.

In the example just cited, the instantaneous peak tube current reaches its maximum value. The minimum instantaneous peak current may be approximated by considering the effect on tube conduction of in-phase chroma and reference signals. In this case, the grid is maximum negative at the same instant that the cathode is maximum negative (minimum positive) and the instantaneous grid-cathode potential is -4.75 volts. Again referring to Fig. 16, the current is approximately 4.5 ma. A third condition which is easily described is the 90° or 270° phase relationship. In this case, the chroma signal is passing through zero when the reference signal is maximum negative, and the peak tube current is the same as if no chroma signal were applied, 8.2 ma.

From the above examples, we find that the maximum and minimum instantaneous peak tube currents are 12.7 and 4.5 ma, respectively. Since the plate load resistor is 3.9K ohms in the actual circuit of Fig. 15, the swing in plate voltage may be predicted to be 32 volts. Since the chroma reference-signal amplitude at the grid of V14A is 2.5 volts, p-p, during normal operation with a keyed-rainbow generator as a signal source, the predicted "gain" is 12.8. Actual observations showed this "gain" to be 10.

Bear in mind that the above examples are an oversimplification of the operation of the circuit since only the instantaneous peak currents were considered. In actual practice, the tube conducts throughout approximately 180° of the reference-signal sine wave, and the magnitude of the current may be represented by the positive half of a sine wave. Nevertheless, we can conclude that the tube conduction is proportional to the sum of the instantaneous amplitudes of the reference signal and chroma signal during the intervals when the tube is out of cutoff. In the circuit of Fig. 15, these intervals correspond to the negative half-cycles of the reference signal. The operation of the X demodulator of Fig. 15 is similar to that of the Z demodulator just described.

Now, consider the operation of the two demodulators if the signal source is an unkeyed-rainbow generator. Using an unkeyed-rainbow signal, there is a constant rate of change of phase difference between the chroma signal and the reference signal. Since the chroma signal is fed to both demodulators at the same phase angle, the outputs of the two demodulators differ in phase by the same amount as the phase difference of the reference signals applied to the two demodulators. In a system using X and Z demodulation, the phase of the reference-oscillator signal applied to the two demodulators differs by 63.9°. Observations of the outputs of the two demodulators when an unkeyed-rainbow generator is used as a signal source confirm that the sinusoidal output of the Z demodulator passes through zero about 64° later than the output of the X demodulator passes through zero.

The operation of the color-difference amplifier shown in Fig. 15 is typical of most color-difference amplifiers used in conjunction with low-level, vacuum-tube demodulators. The difference signals from the demodulators are coupled through C147 and C151 to the B−Y and R−Y amplifiers, respectively. These signals are amplified and fed to the respective control grids of the CRT.

Notice that there are two apparent discrepancies in the foregoing discussion. First, since the signals at the red and blue control grids of the CRT are R−Y and B−Y respectively, the signals at the inputs to the difference amplifiers must bear the negative sign, −(R−Y) and −(B−Y) Hence, the demodulators are actually −X and −Z demodulators. However, it is common practice to ignore this reversal of polarity.

Also observe that while the demodulators operate on the X and Z axes, the difference amplifiers operate on the R−Y and B−Y axes. Although this seems unlikely, it is actually the case. Since the difference amplifiers have a common cathode circuit, a portion of the −Z signal which is fed to the B−Y amplifier appears at the cathode of the R−Y amplifier and vice versa. Thus, a small portion of the −Z signal is added vectorially to the −X signal, shifting the true axis of the difference amplifier to R−Y. Also, a portion of the −X signal is added vectorially to the −Z signal in the B−Y amplifier to establish its true axis.

The means of deriving the G−Y signal is similar to the method described in the explanation of the General Electric circuit. In the circuit of Fig. 15, an additional portion of the R−Y signal is fed from the plate of V15B to the grid of V13B where it is added to the −(R−Y) signal being fed to the cathode of V13B. A sample of the −(B−Y) signal is also fed to the cathode of V13B. The total of these three signals produces a voltage at the plate of V13B which is −.51 (R−Y) −.19(B−Y), equal to G−Y.

Low-Level Pentode Demodulators

The operation of a low-level pentode demodulator is quite similar to the triode demodulator just discussed. The demodulator circuit used in Packard-Bell Chassis 98C15 is shown in Fig. 17. The 6GY6 tubes used as demodulators belong to the family of tubes having two independent control grids. That is, both the suppressor grid and the control grid have considerable effect on the plate current.

The plate current of V18 and V19 are controlled by the instantaneous values of chroma and reference signals and detection takes place at the plate of the tube. The operation of the color-difference amplifiers is essentially the same as the ones discussed previously.

Sheet-Beam Demodulator

The sheet-beam demodulator used in Zenith receivers is actually a variation of the pentode demodulator. However, there are two significant departures from the conventional design. Since the amplitude of the output from sheet-beam demodulators is sufficient to drive the CRT directly, they are known as high-level demodulators. The more important difference is the means by which detection of the chroma signal is accomplished. In the triode and pentode circuits just discussed, the amount of cathode current is controlled jointly by the chroma and reference signals. In the sheet-beam demodulator, cathode current is controlled solely by the chroma signal, but the selection of which plate receives the current is controlled by the reference signal.

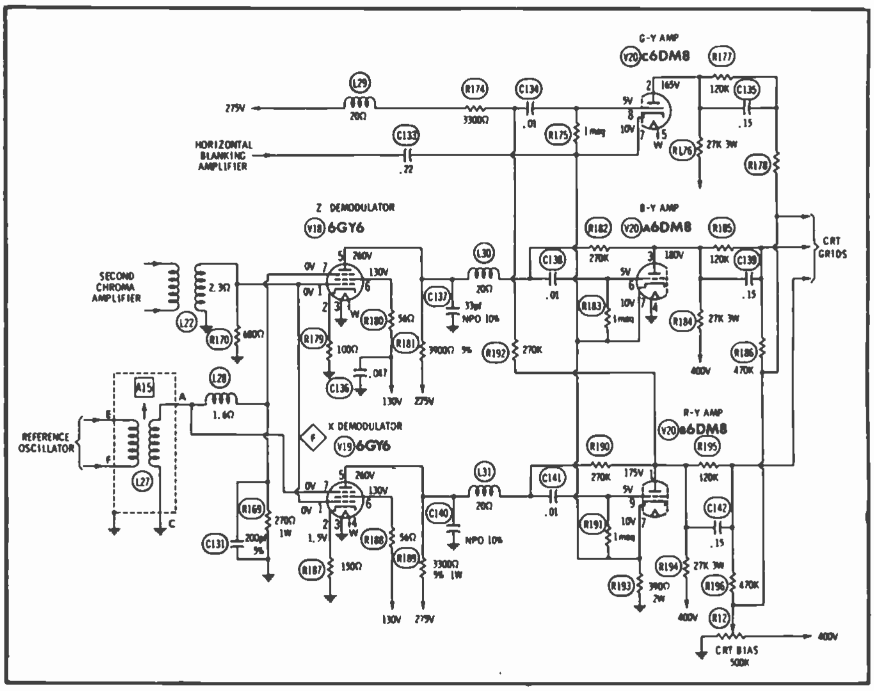


Fig. 17. Chroma demodulators and color-difference amplifiers of Packard-Bell Chassis 98C15.

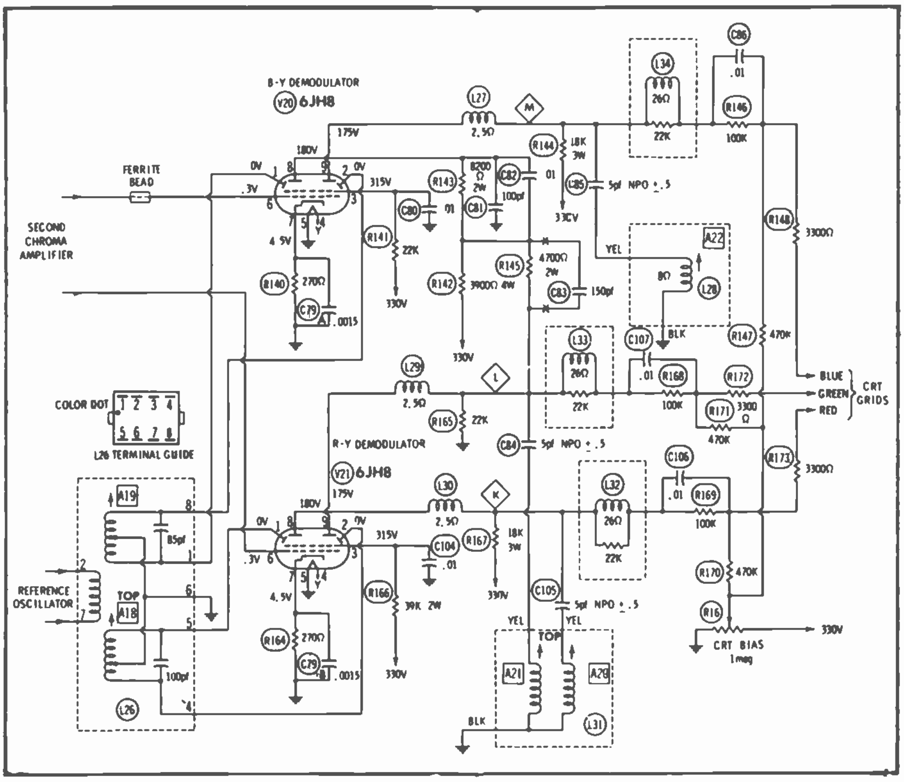


Fig. 7. Sheet-beam demodulator of Zenith Chassis 24MC32.

Fig. 18 shows the demodulators used in the Zenith 24MC32 and 42 chassis. Consider the B − Y demodulator under no-signal conditions. The reference signal is fed to the two deflection plates in opposite phases and so the cathode current is directed alternately to each of the plates. Since the amplitude of the reference signal is considerably greater than the minimum required to deflect all of the current to one plate, the current at each plate consists of a series of square-wave pulses of constant amplitude.

During color reception, the chroma signal is fed to the control grid and determines the amount of instantaneous current that is available to the plates. Assume, for example, that the left deflection plate, pin 1, is maximum positive at the same instant that the control grid is maximum positive. In this case, the current of the left plate is maximum and the plate voltage is minimum. Since the right deflection plate is positive one-half cycle later, the control-grid voltage is negative during the time that the right plate is gated on. Thus the current of the right plate is minimum and the voltage is maximum.

A 90° phase difference between reference and chroma signals causes one plate to be gated on while the chroma signal swings from peak positive to peak negative. The other plate is gated on while the chroma signal swings from peak negative to peak positive. Therefore, the average of all the values of instantaneous grid voltage present while one plate is gated on equals zero and the average value of plate current is equal to the amount of current under no-color conditions. For intermediate phase relations of the reference and chroma signals, the current of a specific plate varies in the same manner as in any of the other demodulators described. The curve shown in Fig. 13 applies to this circuit except that the phase relationship is shifted 180°.

The phase of the reference signal is established so that the left side of V21 demodulates on the R − Y axis and the output is fed to the red grid of the CRT. The right side of L31 is a filter which attenuates the 3.58-MHz ripple. Since the right deflection plate of V21 is driven 180° out of phase with the left deflection plate, the right side of V21 is operating on the − (R − Y) axis, another way of saying that its output polarity is reversed. In the same fashion, the right side of V20 operates on the B − Y axis and drives the blue grid of the CRT while the left side produces the negative or − (B − Y) output.

As we have pointed out previously, the G − Y signal is composed of − .51(R − Y) and − .19(B − Y).

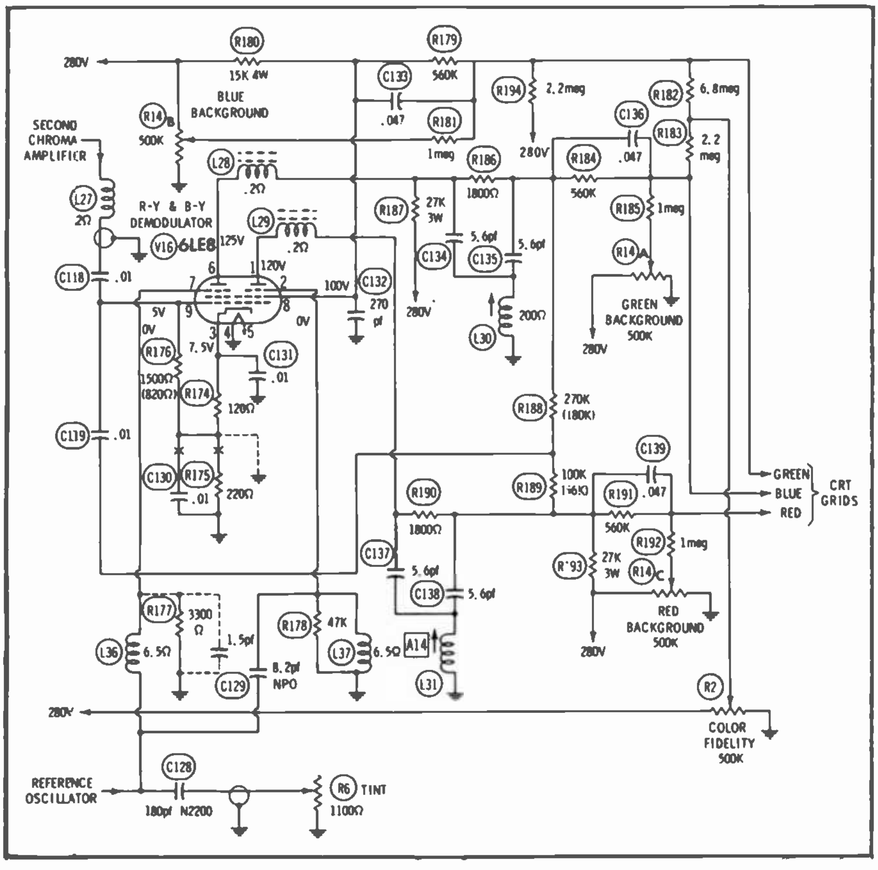


Fig. 19. Chroma demodulator of Admiral Chassis 1G1155-1.

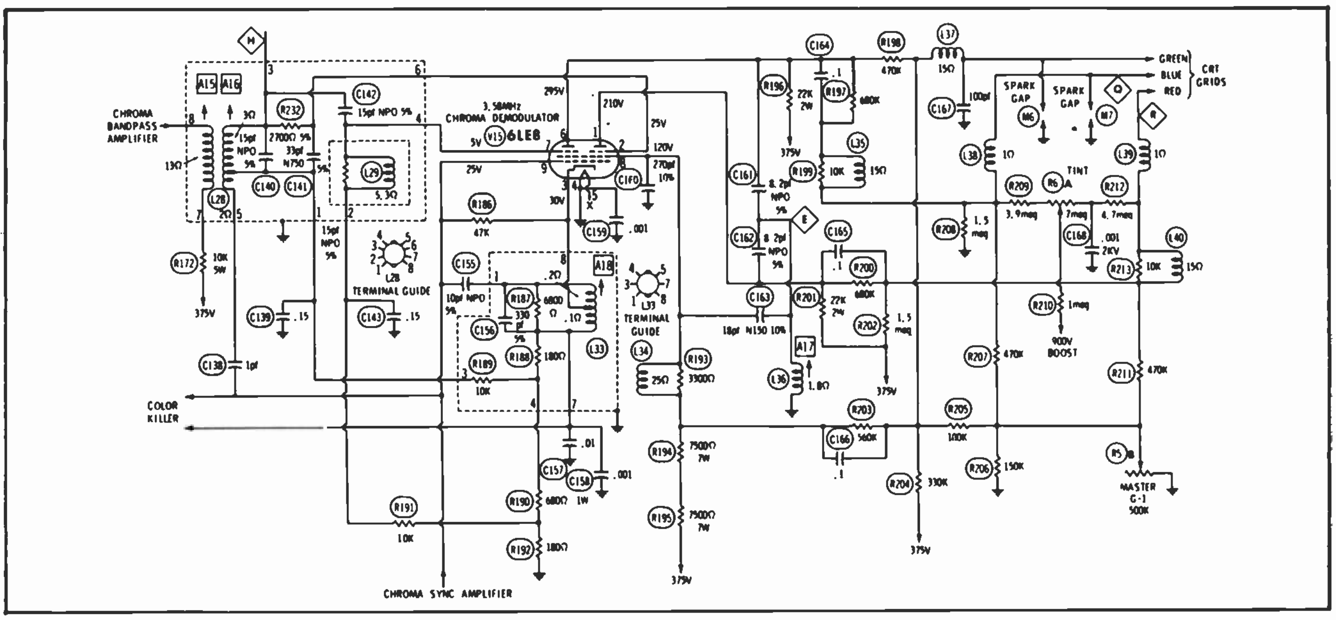


Fig. 20. Chroma demodulator of Motorola Chassis A22TS-918A.

The signals from the left side of V20 and the right side of V21 are combined in the proper proportions and the G−Y signal which results is fed to the green grid of the CRT.

Twin-Pentode Demodulators

The twin-pentode demodulator used in Admiral and Motorola receivers is still another significant variation of the conventional pentode demodulator. The circuit shown in Fig. 19 is used in the Admiral 1G1155-1 chassis.

Insofar as demodulation of the R−Y and B−Y signals is concerned, the operation of this circuit is not greatly different from the operation of a circuit employing a pair of pentode demodulators. The reference signal is fed to the suppressor grids at the correct phase to cause the left and right halves of the tube to demodulate on the B−Y and R−Y axes, respectively. After the 3.58-MHz ripple has been filtered, these plate signals are used to drive the blue and red grids of the CRT.

Before attempting to explain the method used to develop the G−Y signal, we must digress somewhat to consider the distribution of the cathode current between the screen grid and plate of a pentode. An oversimplified explanation may be developed by considering what happens if the plate voltage is removed from a pentode having a low-impedance screen supply. There is an immediate increase in screen current followed by a gradual decrease as the screen-grid structure melts and flows to the bottom of the envelope. While this demonstration is, perhaps, more dramatic than useful, it does indicate that the screen current increases if the plate current decreases.

Actually, the screen current depends to a great degree on the velocity of the electrons passing between its conductors. If the electrons are travelling at very high velocities, the positive potential of the screen grid deflects them from their path only slightly. Thus, the only electrons which impinge on the screen grid are the ones which happen to be travelling directly towards the wires of the grid. However, if the velocity is decreased, the screen grid causes more deflection of the electrons. Since the screen grid is positive, the electrons are deflected toward its conductors and more of the electrons actually strike the wires. This, of course, causes an increase in screen current.

Now, consider the electron velocity with various suppressor grid potentials. If the voltage on the suppressor is driven negative, electrons in the region between the screen and the suppressor are decelerated to a lower velocity and the screen current increases. Conversely, if the suppressor is driven positive, electrons are decelerated a lesser amount, and the screen current is reduced.

A positive suppressor voltage which decreases the screen current also increases the plate current, and, while the plate swings negative, the screen swings positive. Thus, a signal on the suppressor grid may be amplified in both the plate circuit and the screen circuit. The signal is inverted between suppressor and plate, but, since a decrease in plate current is attended by an increase in screen current, the signal at the screen is in phase with the suppressor grid and out of phase with the plate.

Returning to the circuit of Fig. 19, since the signals at the two plates of V16 are B−Y and R−Y, the common screen grid has elements of the −(B−Y) and −(R−Y) signals on it. However, the suppressor-to-screen transconductance is comparatively low and the G−Y signal at the screen grid is insufficient to drive the green grid of the CRT. To increase the G−Y signal at the screen, portions of the R−Y and G−Y signal from the plates of V16 are coupled back to the control grid. This signal is amplified in the control-grid/screen-grid circuit and the amplified signal adds to the G−Y signal already present at the screen. Since the (R−Y) + (B−Y) signal at the control grid is a video frequency and the chroma-signal frequency is much higher, the interaction between them is too small to be objectionable. The concept of amplfying two dissimilar signals in the same tube (reflex amplification) is by no means a new one. For example, a

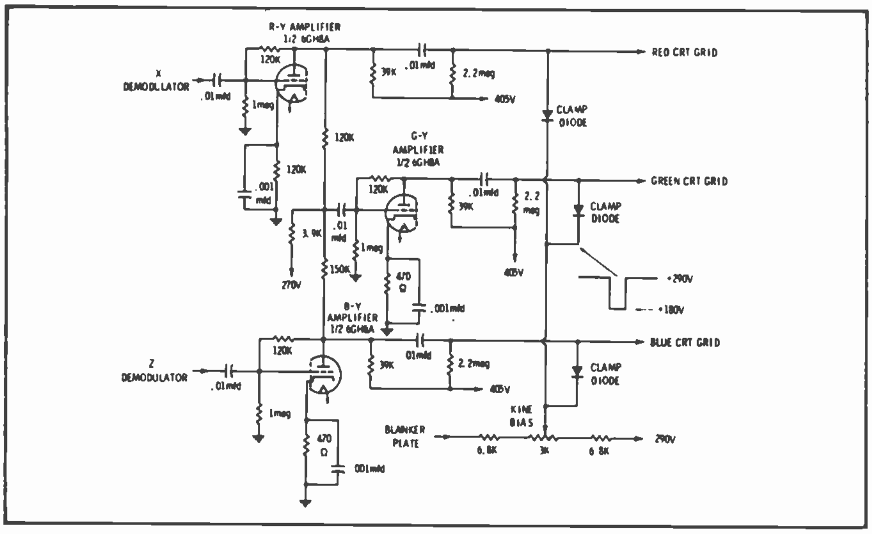


Fig. 21. CRT grid clampers of RCA Chassis CTC27.

single tube is used as the second video IF amplifier and first sound IF amplifier in the Muntz J chassis (PHOTOFACT Folder 444-2).

The demodulator used in Motorola chassis A22TS-918A (Fig. 20) uses the twin-pentode which serves as the reference oscillator to demodulate all three color axes. The explanation of the division of current between plate and screen grid which was given above also applies to this circuit.

In this part of this book, it was pointed out that the cathode, control-grid, and screen-grid circuits of V 15 (Fig. 20) function as a reference oscillator in a Hartley configuration. In considering V15 as a demodulator, we need only remember that a high-level reference-oscillator signal is being fed to the control grid.

Since this control grid is common to both pentode sections, the reference signal cannot be split prior to demodulation. Consequently, the phase of the chroma signal is shifted between the two suppressor grids. Since the axis of a particular demodulator depends on the *relative* phases of the reference and chroma signals fed to it, it is not important where in the reference or chrominance circuits the phase-shifting network is located. In Fig. 20, the phases of the chroma signal are established by L28 and its associated components so that the left plate of V15, pin 6, demodulates on the B − Y axis and the right plate, pin 1, demodulates on the R − Y axis.

Because the division of current between plate and screen grid is determined by the instantaneous voltage of the suppressor grid, elements of the −(R − Y) and −(B − Y) signal are present on the screen grid of V15. A portion of the B − Y signal from the junction of R208 and R207 is added to the screen-grid signal, reducing the −(B − Y) content. The total of the three signals is the G − Y voltage fed to the green grid of the CRT. C161, C162, and C163 in conjunction with L36 are used to attenuate the 3.58-MHz ripple which is present on the plates and screen grid of V15.

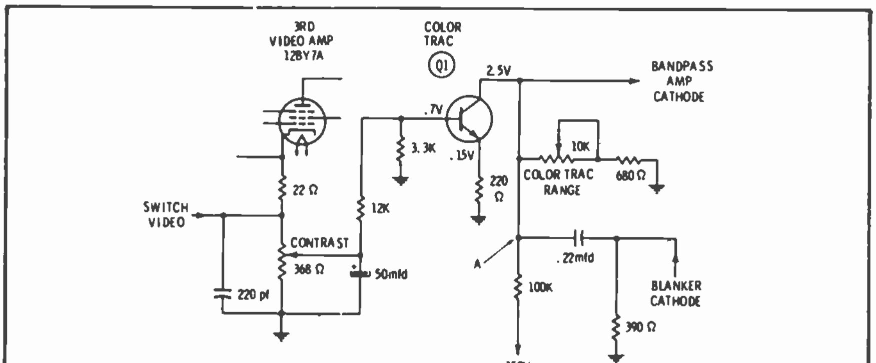


Fig. 22. Color-tracking circuit of Hoffman Chassis 913-187486.

Miscellaneous Circuits

Three circuits which are being used more frequently are CRT grid clampers (DC restorers), color tracking circuits, and tint controls. Fig. 20 shows an example of the latter. R6A is connected between the red and blue grids of the CRT and allows the customer to adjust the relative conduction of these two guns.

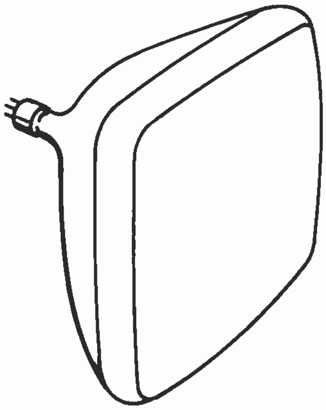
Fig. 21 shows the CRT grid clampers used in the RCA CTC27 and CTC31 chassis. By inserting coupling capacitors between the plates of the difference amplifiers and the CRT grids, it is possible to operate these two elements at nonrelated DC potentials. This allows simplification of the circuits and also precludes changes in color temperature caused by changes in the static conduction levels of the difference amplifiers. Thus, the CRT grids are not affected by aging of the difference-amplifier tubes and components.

Although the use of coupling capacitors is desirable for the reasons stated above, a means of maintaining the correct CRT grid voltages during b-w reception had to be devised. Otherwise, the right sides of the coupling capacitors would slowly charge through the 2.2-megohm resistors to 405 volts, saturating the CRT. To prevent this, a negative-going pulse with a negative peak of 180 volts positive above ground (refer to waveform in Fig. 21 is fed through the clamper diodes to the CRT grids during horizontal-retrace time. Between pulses, the CRT grid voltages rise from 180 volts to about 181 volts, but this is insufficient to cause a noticable change in brightness across the screen.

A method of increasing color saturation, as well as contrast, with a single control is illustrated in Fig. 22. As the contrast control is adjusted for greater contrast, the forward bias on Q1 is increased. The increased current through Q1 causes the collector voltage to decrease, reducing the positive cathode bias of the chroma bandpass amplifier and increasing its gain. Positive blanking pulses are also fed to the cathode of the chroma-bandpass amplifier. ▲

Chapter 6

The Color CRT



The circuitry used to extract the luminance and chrominance information from the composite video signal has been explained in Chapters 2, and 5 of this book. From these discussions, we learn that there are four video inputs to the color CRT: (1) the luminance, or Y, signal which contains only brightness information; (2) the R−Y signal containing, essentially, only the red information without any brightness information, (3) the B−Y signal containing the blue information, and (4) the G−Y signal which contains the green information only.

It is customary to assign a notation of polarity to each of the last three of these signals. Thus we have the quantities −(R−Y), −(B−Y), and −(G−Y) as well as their positive counterparts expressed above. These arbitrary designations can become confusing when considering the operation of the CRT itself since a specific gun, red for instance, conducts even when a −(R−Y) signal is fed to its grid. Actually, the quantity 0 (R−Y) is defined as the amount of conduction of the red gun which, when the G−Y and B−Y signals are also at zero level, will produce reference white. Stated another way, zero levels of R−Y, B−Y and G−Y are present during b-w reception. Negative R−Y, B−Y, or G−Y refers to an amount of conduction of the specific gun which is less than that which is required to produce reference white.

Notice that the conduction of a specific gun is not entirely dependent on the color-difference signal fed to its grid. Conduction is determined by the grid-to-cathode voltage of the gun, and since the cathode is being driven by the luminance, Y, signal, the absolute value of conductance of each gun is determined by the sum of the color-difference and Y signal being fed to it. But, the cathodes of the three guns receive the same signal; therefore, brightness is determined by the total conductance of all three guns, but hue is determined by the relative conductance of each of the three.

Unlike the picture tube used in a monochrome TV, which must reproduce only brightness information, a color CRT must also translate the various input signals into colors as well as brightness. In the earliest color TV's, the color-difference signals (R−Y, B−Y, and G−Y) were recombined with the luminance (Y) signal outside the CRT, but nearly all modern sets depend on the CRT itself to perform this function. This, of course, makes the design of the color CRT even more critical. The Y signal is fed to the three cathodes of the CRT in phase and at approximately the same amplitude. The three color-difference signals are fed to the control grids of the red, blue, and green guns. Therefore, the beam current from the blue gun is controlled by combining the B−Y signal with Y to obtain blue (B) information, R−Y is combined with Y in the red gun, and G−Y is combined with Y in the green gun.

The no-signal conduction of each gun is preset by adjusting the grid No. 2 (screen) voltages of each gun for reference white. If the characteristics of the three guns were identical, no further adjustments would be necessary. However, the transconductance, $\frac{\Delta \text{ beam current}}{\Delta \text{ grid voltage}}$, of the three guns may vary slightly, causing the overall tint of the picture to change with changes in brightness. This condition (known as improper gray-scale tracking) is corrected by inserting gain controls in the cathode circuits of two of the guns. These are adjusted so that the hue of the picture is not affected by its brightness.

Fig. 1 is a graph showing the brightness produced by each of the three guns in a hypothetical CRT. These curves show no specific quantities but only a relationship between the three guns. In this hypothetical tube, if white were produced at 12 brightness units, the screen would tend to become blue with increased brightness. At brightness levels less than 12, it would be predominantly green. By decreasing the transconductance of the blue gun and increasing the transconductance of the green gun, reference white can be obtained at all brightness levels. This, of course, is known as gray-scale tracking.

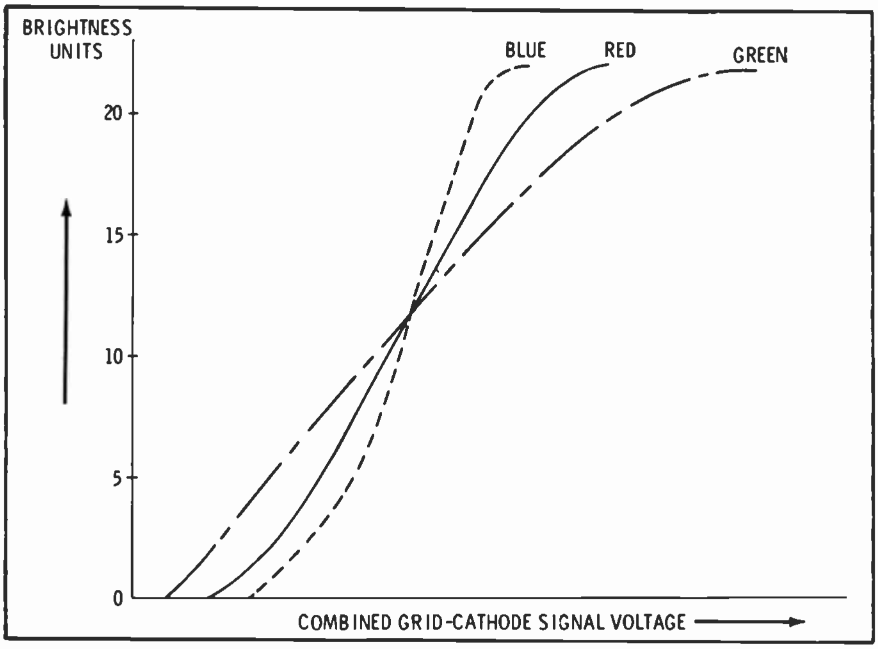


Fig. 1. Possible gain variations of the three CRT guns.

CRT Geometry

At the present state of the art, the color tube most commonly used has three guns spaced at equal angles around the axis of the tube, a shadow mask with color-dot triads, electromagnetic deflection and convergence, and electrostatic focus.

Fig. 2 is a cross section of the neck of a CRT showing the position

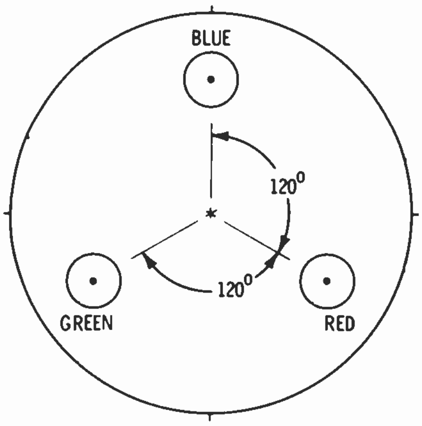


Fig. 2. Cross section of CRT neck.

of the three guns. Notice that the red and green guns share a common horizontal plane while each gun lies in its own vertical plane. A view of two gun assemblies is shown in Fig. 3. A close examination of these gun assemblies would show that each of the three guns is tilted so that the three beams will converge at a common point at the center of the shadow mask. This simplifies the convergence circuitry associated with the tube.

As shown in Fig. 3, each gun consists of a heater, cathode, control grid (No. 1), accelerating grid (No. 2), focusing grid (No. 3), and a final accelerating grid (No. 4) which, since it is connected to the neck coating and shadow mask, is part of the ultor or high-voltage anode of the tube. A pair of pole pieces are located at the end opposite the cathode of each gun. These are used in conjunction with external fixed magnets and electromagnets to position the beams to provide proper convergence. In some CRT's, an additional set of pole pieces are situated adjacent to the focusing grid of the blue gun. These, too, are used in conjunction with an external magnet (the blue lateral magnet) to aid in converging the blue beam with the other two.

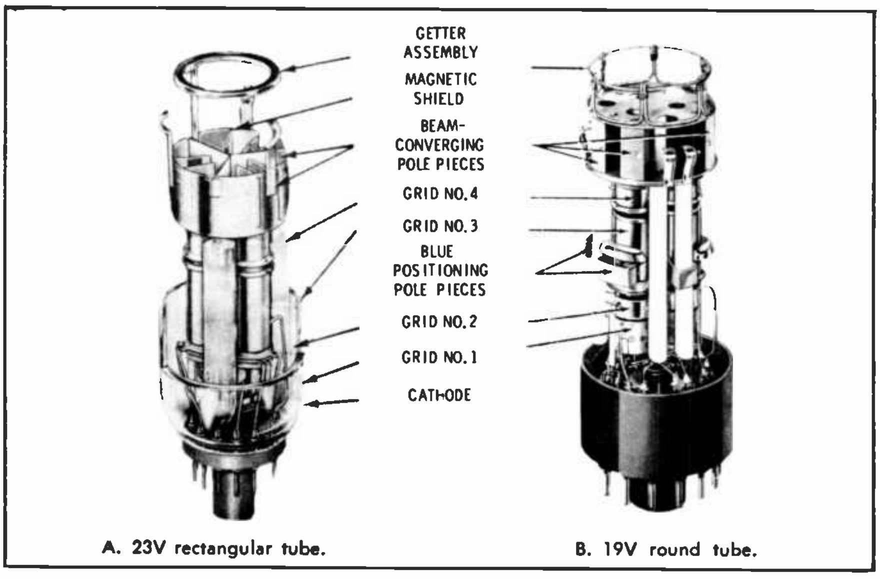
Fig. 4A shows a side view of a typical 23V color CRT with the deflection components in place, and Fig. 4B is a cutaway view of a 19V tube. Of particular interest is the arrangement of the shadow mask and phosphor screen. The shadow mask is a thin sheet of metal which has been perforated with a series of very small holes by a photoengraving process.

For each hole in the mask, three bits of colored phosphorescent material are deposited on the face plate of the tube in a pattern called a triad. These dots are arranged so that electrons emanating from the red gun pass through a specific hole in the mask and excite the red phosphor dot. At the same time, electrons from the blue gun pass through the same hole in the shadow mask to excite the blue phosphor dot; and, of course, the electron beam originating at the green gun at this same instant passes through the identical hole and excites the green phosphor dot.

Color Purity

The degree of color purity depends on how well the beams from the guns actually pass through the holes in the shadow mask to strike the phosphor dots. The two variables which affect purity are the distance from the deflection plane to the shadow mask and the points on this plane through which the beams pass. Fig. 5 illustrates these two variables.

The relative positions of the phosphor dots and the holes in the mask



A. 23V rectangular tube. B. 19V round tube.

Fig. 3. Typical electron-gun assemblies.

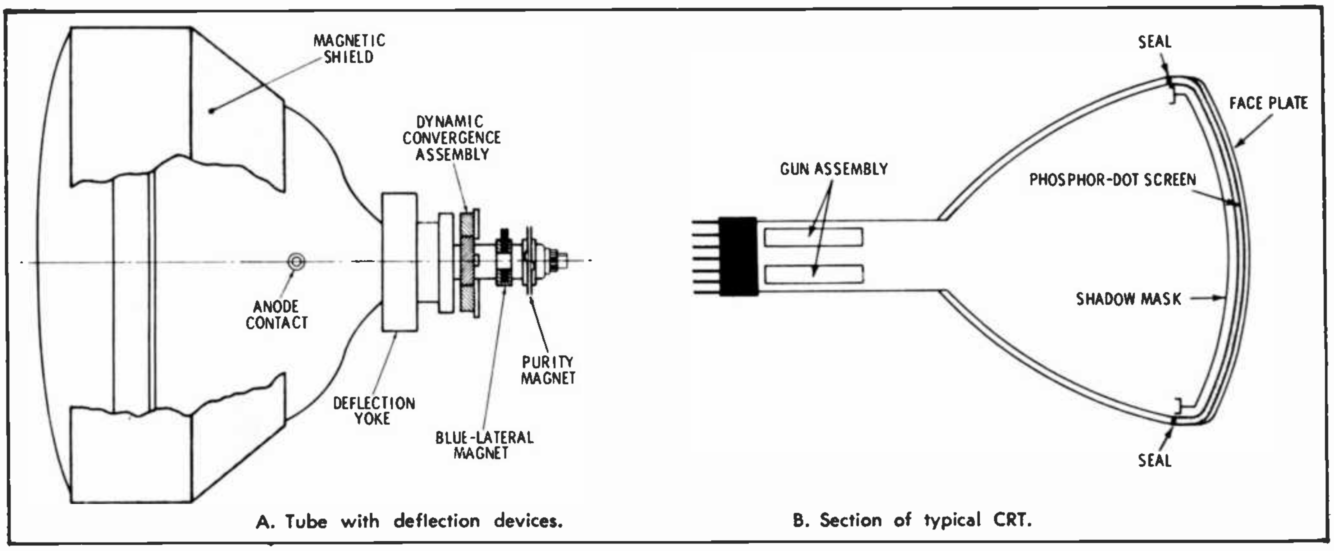


Fig. 4. Details of CRT construction.

were fixed at the time of manufacture of the CRT. These positions are such that if lines were to be drawn from each red dot, through their respective holes in the mask, toward the neck of the tube, all of the lines would come together at one precise point. In making the purity adjustments, the beam from the red gun is made to emanate from this same point, which is somewhere on the correct deflection plane.

Referring to Fig. 5, notice that the correct point and two incorrect points of apparent beam origin are shown. If the beam emanates from an incorrect point on the proper deflection plane (point A), it will miss the red phosphor dots at the edges of the screen and also at the center. On the other hand, even though the deflection plane is incorrect, if the point of origin is correct (point B), the beam will hit the red dots near the center of the screen but miss them at the edges.

The correct procedure for adjusting purity may be inferred from this discussion. If the deflection yoke is moved toward the rear of the CRT as far as possible, it will be in the vicinity of the incorrect deflection plane shown in Fig. 5. This allows the purity magnet to be adjusted for the proper point of apparent beam origin (point B of Fig. 5), as evidenced by a pure red area at the center of the raster. Then the yoke is moved forward until red purity is obtained over the entire raster.

The purity magnet must be capable of producing a uniform magnetic field of adjustable intensity

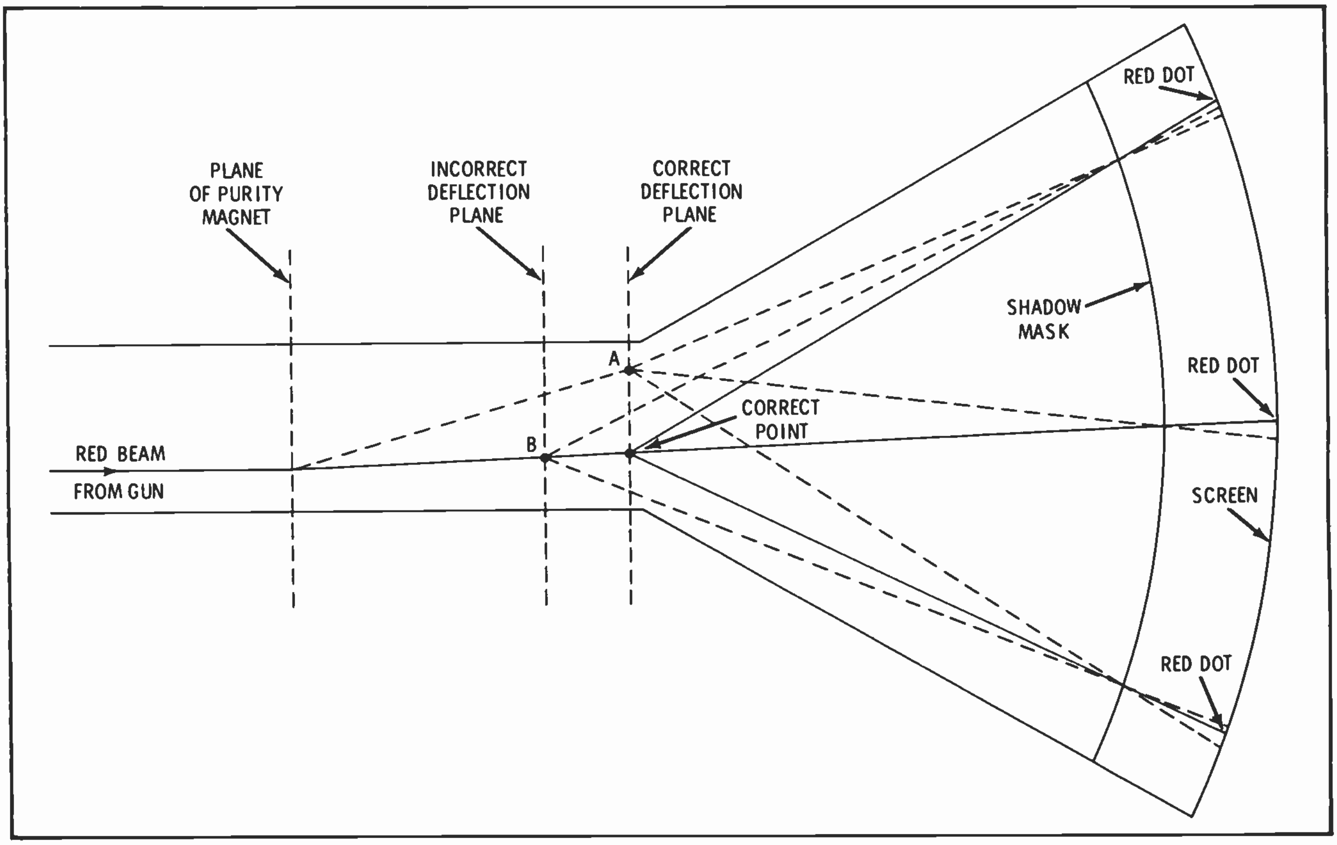


Fig. 5. Deflection geometry showing effects of purity adjustments.

perpendicular to the neck of the CRT. Also, this field must be capable of being oriented in any direction. To accomplish this, two magnetic rings, each polarized as shown in Fig. 6, are mounted on the CRT neck. Depending on the relative position of the two rings, their fields will either aid or oppose each other to determine the intensity of the field. To determine the direction of the field, the two rings are rotated together.

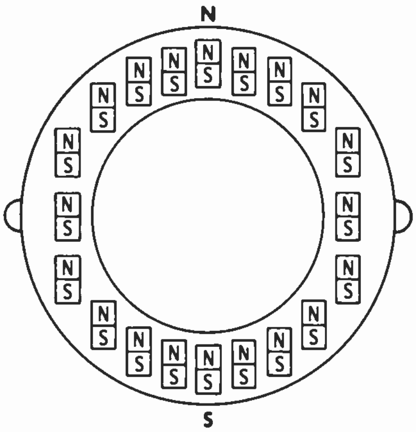


Fig. 6. A typical purity magnet.

In discussing purity, we have considered only the red beam and phosphor dots and, in fact, the blue and green guns are normally disabled while making purity adjustments. This procedure is acceptable since the blue and green beams are deflected the same amount by the purity magnet as was the red beam and, consequently, are automatically positioned at approximately the correct points on the deflection plane. The subsequent convergence adjustments position the points precisely to their correct locations.

Static Convergence

Fig. 7 is a distorted view of a color CRT which illustrates some of the geometrical aspects of convergence. The distance from the shadow mask to the phosphor screen has been increased while the distance from the tube neck to the shadow mask has been decreased so that the divergence of the beams after they pass through the mask may be seen. The plane depicted in Fig. 7 is horizontal, that is, the observer is looking down at the CRT and the center horizontal line of the CRT is being scanned by the electron beams. In actual practice, the plane of the convergence yoke is to the rear of the plane of the deflection yoke; however, for the purposes of this discussion the two planes are assumed to coincide. This greatly simplifies the explanation and, from a qualitative point of view, has no effect on the operation of the convergence system.

Consider the moment when the triad at the center of the screen is illuminated. The beam from the green gun crosses from left to right (looking from the guns toward the screen), the red beam crosses from right to left and the blue beam goes straight ahead. (Actually the red and green beams are aimed upwards slightly and the blue beam is slightly depressed, but this cannot be observed from the present point of observation). When the three beams actually behave as shown, the tube is correctly converged at the center. This is called static convergence, since there is no horizontal- or vertical-deflection field acting on the beams at this instant.

From the above, it may be inferred that static convergence consists simply of aiming all three guns at the center of the CRT screen. This is accomplished by adjusting the three permanent magnets which are part of the convergence yoke so that the electron beams are correctly aimed. Fig. 8 shows the red and blue positioning magnets and the blue lateral magnet, soon to be discussed. The positioning magnets

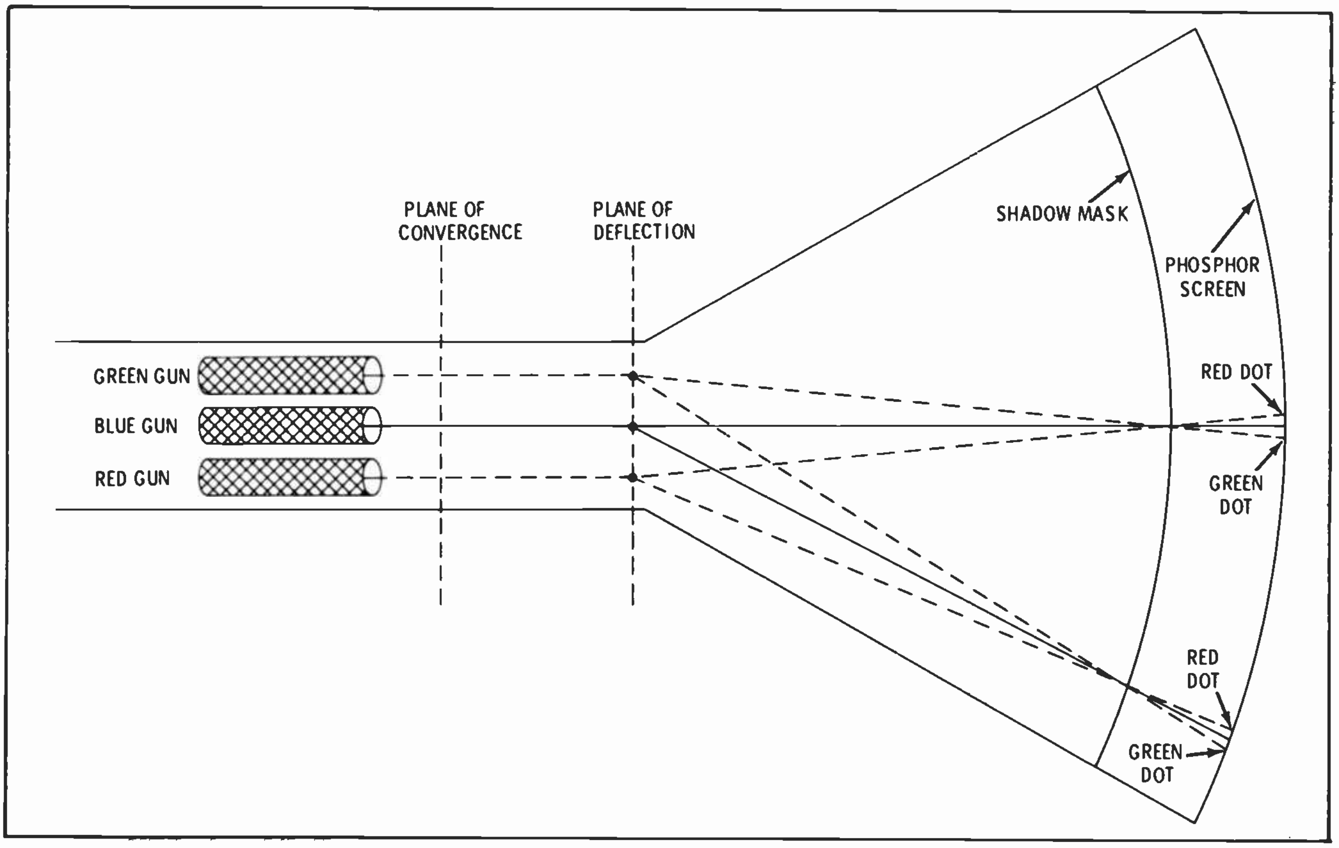


Fig. 7. Deflection geometry showing correct convergence.

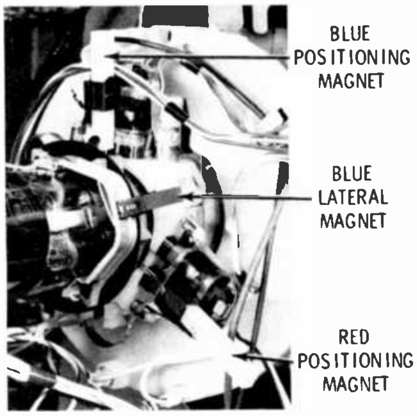


Fig. 8. Convergence components.

themselves are imbedded in a plastic sleeve, and these sleeves may be pushed in closer to the CRT neck to increase the effect of the fields. The sleeves may be rotated 180° in their holders to reverse the magnetic field if necessary. Of course, other magnet configurations are possible, but the operation is essentially the same.

Fig. 9 shows the effect of the positioning magnets on their respective beams. As the red and green postioning magnets are repositioned, the respective beams scan the screen in the direction shown by the arrows. Since the guns are aimed, roughly, at the center of the CRT during manufacture, the

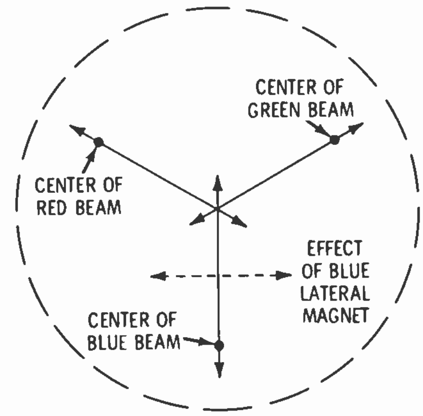


Fig. 9. CRT screen section shows effects of static-convergence magnets.

landing points of the red and green beams can be made to coincide at some point near the center of the screen. The blue positioning magnet causes the blue beam to fall at some point along a vertical line on the screen (also near the center), which point is determined at will by adjusting the positioning magnet. When the blue spot is directly alongside the spot illuminated by the red and green beams, the blue lateral magnet (also pointed out in Fig. 8) is adjusted to move the blue spot horizontally until all three spots coincide.

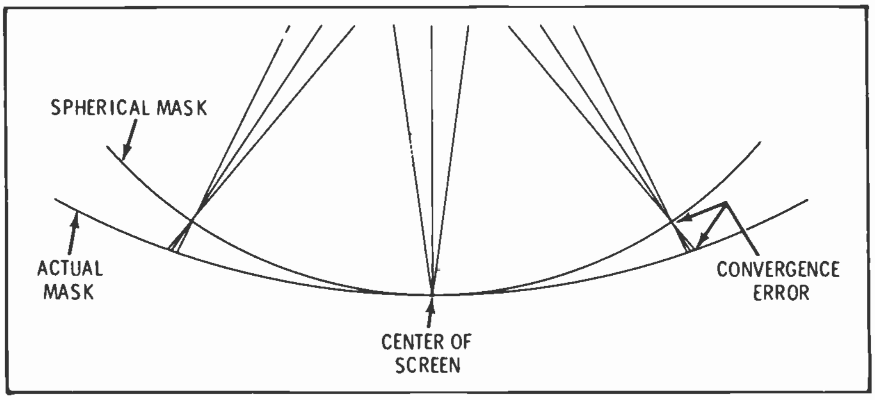


Fig. 10. Dynamic-convergence error due to non-spherical mask.

The blue lateral magnet is mounted on the CRT neck to the rear of the convergence yoke and is situated at right angles to the blue positioning magnet. Because of its position, the only one available, it has a slight effect on the red and green beams as well as on the blue beam. Accordingly, the position of all four magnets must be "roughed in" first and then retouched for best convergence.

Dynamic Convergence

Unlike static convergence, which may be accomplished by permanent magnets, dynamic convergence requires the generation of fairly complex magnetic fields. Before examining the circuitry required to produce these fields, let's define dynamic convergence and describe the magnetic fields which are required to accomplish it.

The major factor which affects dynamic convergence is the curvature (or rather, the lack of it) of the shadow mask and screen. If the screen were a sector of a sphere whose center was on the convergence plane, dynamic convergence would be relatively simple. However, the screen has a radius much greater than the distance to the deflection plane and, if nothing is done to correct it, a beam which is converged at the center of the screen will converge behind the mask at every other deflection angle. This is illustrated in Fig. 10. Accordingly, progressively less convergence field is required as the deflection field increases. Notice, too, that dynamic convergence is required for the vertical sweep as well as the horizontal sweep.

Fig. 11 shows the effect of no dynamic convergence on vertical and horizontal lines passing through the center of the raster. In considering the effects of convergence, these lines are of particular interest because the dynamic horizontal-convergence field is zero when these vertical lines are being produced, and the dynamic vertical-convergence field is zero when the center horizontal lines are being produced. The total dynamic-convergence field when any other point on the raster is being scanned is a sum of both the horizontal- and vertical-convergence fields.

Notice that three vertical, but only two horizontal, lines are

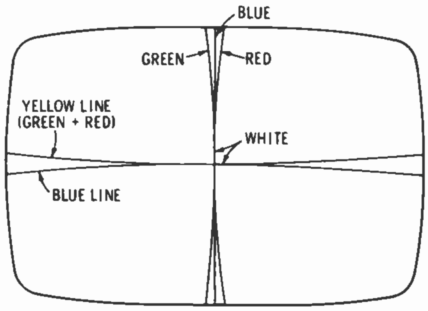


Fig. 11. Raster with dynamic-convergence circuits disabled.

produced. This is easily understood if one considers that each gun is in a separate vertical plane, but that the red and green guns are in the same horizontal plane. Thus the red and green beams track each other in a horizontal scanning line.

The curves produced on the raster by unconverged beams are parabolic in shape, and so a parabolic correction current through the convergence coils will straighten the lines. Although it is possible to dynamically converge all three beams with a single device, it is standard practice to use independent electromagnets for each gun and to

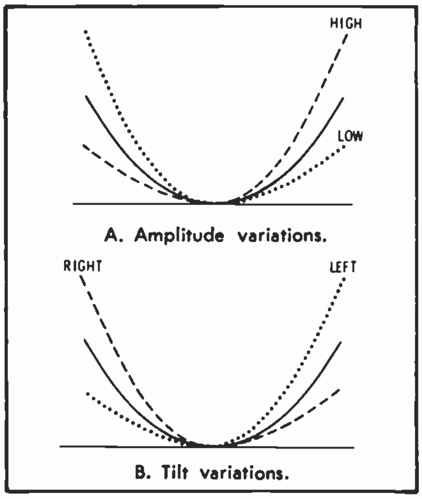


Fig. 12. Parabolic-current waveforms.

generate separate currents for each of them. By doing this, the amplitude and the tilt (angle of the axis of the parabola relative to the vertical axis of a graph) of the current in each convergence coil may be adjusted for the best possible appearance of the raster. Fig. 12 relates the terms "tilt" and "amplitude" to a parabola.

Referring to Fig. 8, the red and blue dynamic-convergence electromagnets may be seen immediately in front of the red and blue positioning magnets. Fig. 13 shows a single convergence electromagnet and illustrates the relative size of the horizontal- and vertical-convergence coils. Since the horizontal- and vertical-convergence currents have the same repetition rate as the horizontal- and vertical-deflection oscillators, the vertical-convergence coils must have many more turns

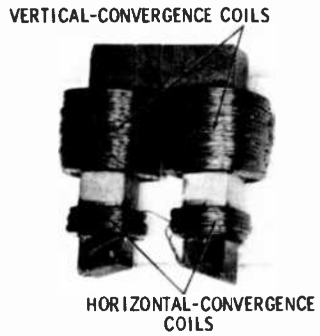


Fig. 13. Typical dynamic-convergence electromagnet.

to obtain a reasonable impedance. The circuitry used to develop the convergence currents will be discussed later in this article.

Pincushion Effect

Those who were in the TV servicing business a generation ago will recall when the effects of pincushion distortion first became a problem. In the earliest black-and-white sets, using a small deflection angle and a near-spherical screen with a fairly short radius, pincushion distortion was no problem. Then, as wider deflection angles and flatter screens became popular, pincushioning became objectionable. Fig. 14 shows a raster with severe pincushioning. Actually, pincushion distortion is aggravated by the use of cosine or cosine-squared yokes, necessary to maintain focus over the surface of the flat-screen, wide-deflection CRT.

In b-w sets, pincushion correction is fairly simple. By positioning small

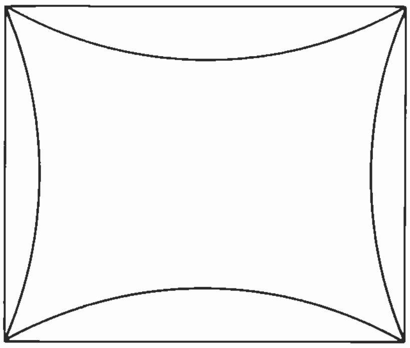


Fig. 14. Pincushion distortion.

permanent magnets around the CRT, just forward from the deflection yoke, the beam is caused to move outward as it crosses the horizontal and vertical center lines, straightening the scan lines on the CRT. Unfortunately, the use of pincushion magnets in a color set is not possible. As we all know, it is problem enough to get rid of magnetic fields around the CRT bulb without introducing any additional magnets. Secondly, since the three beams have different points of origin on the deflection plane, pincushion magnets would not affect the three beams equally, making convergence at the edges of the CRT very difficult.

Since the use of pincushion magnets is precluded, the correction field must be generated electromagnetically. Small correcting currents are generated in the pincushion circuit and added to the deflection currents in the yoke. A typical circuit will be examined under a separate head in this article.

Dynamic-Convergence Circuits

As pointed out previously, the currents through the dynamic-convergence electromagnets must be parabolic in shape and have controllable amplitude and tilt. This is true of both the horizontal and vertical dynamic-convergence currents; however, since the circuits used to generate these currents are different, they will be discussed separately.

Fig. 15 shows the vertical dynamic-convergence circuit used in the Admiral 2G1156-1 chassis. The sawtooth from the cathode of the vertical-output tube is fed through C6 and R199 to the top of a bleeder,

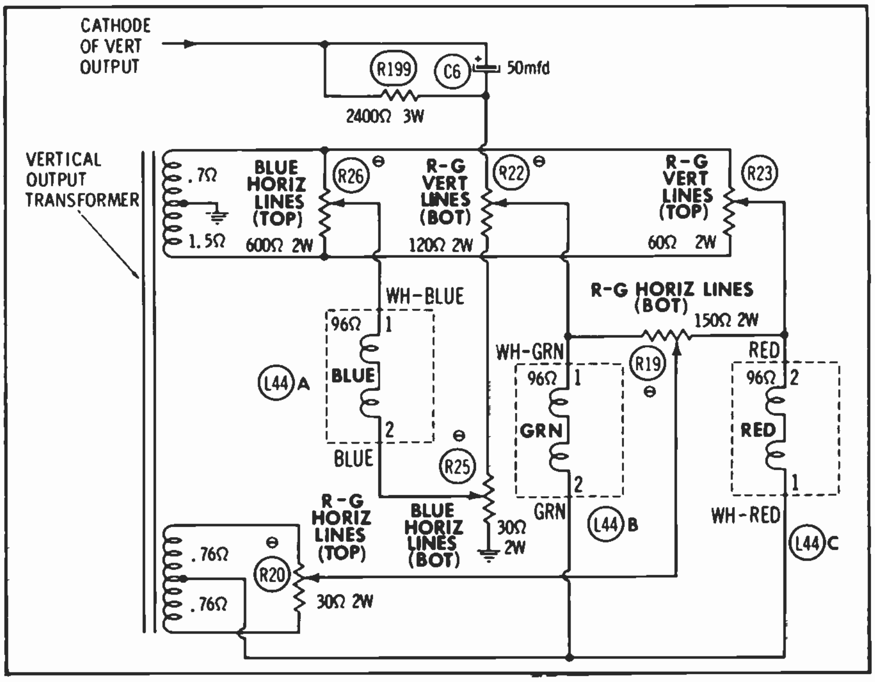


Fig. 15. Typical vertical-dynamic-convergence circuit.

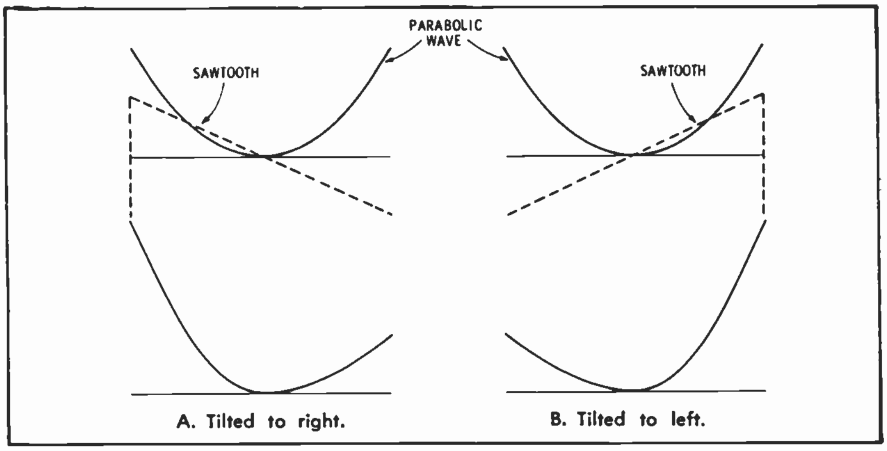


Fig. 16. Result of adding parabolic and sawtooth currents.

consisting of R22 and R25, and thence to ground. A portion of this voltage is picked off by the slider of R25 which connects to one end of the blue convergence coil, L44A. The opposite end of L44A is connected to the slider of R26 which, in turn, is across a winding of the vertical output transformer. Thus, the parabolic current waveform produced by the action of R199, C6, and L44A is modified by the sawtooth current from the transformer winding. This is illustrated in Fig. 16. Since the ends of the transformer are out of phase, either a positive-going or negative-going current may be picked off by the slider of R26; and the parabola may be tilted in either direction.

In considering the currents through L44B and C, we will first trace the current produced by the lower winding of the vertical-output transformer of Fig. 15. The amplitude and phase of voltage taken from this winding is determined by the setting of R20. The relative amplitudes of this voltage which are fed to L44B and C is determined by the postion of the slider of R19. Notice that as more current is caused to pass through one coil, less passes through the other; but that if current flow through one coil is from terminal 1 to terminal 2, it is from 2 to 1 in the other.

A second sawtooth is taken from the upper winding of the vertical-output transformer. This is tapped by R23 and fed through L44C, then L44B, to the slider of R22, and ultimately back to ground. Since the setting of R23 determines the phase of this sawtooth, current may be in either direction, but if it passes from 2 to 1 of L44C, it also passes from 2 to 1 of L44B. In either case this sawtooth modifies the tilt of the parabolic current supplied from R22. Depending on the positions of the various controls, R23, R19, R20, and R22, the amplitude and tilt of the parabolic currents through the red and green coils may be adjusted over a wide range.

Fig. 17 shows a typical horizontal-convergence circuit. Although the current waveforms are very similar to the ones developed in the vertical-convergence circuits (except for frequency), the circuitry is somewhat different. These circuit differences are present because a suitable sawtooth input is not available and also because the red and green guns are in the same horizontal plane and require approximately the same convergence current.

The input for red and green convergence is a 300-volt pulse from the horizontal-output transformer. This is shaped into a sawtooth by L1 and a parabolic current is caused to flow through the coils. The ganged potentiometers, R8A and B, are connected so that the current returns to ground through either of two windings. These windings are out of phase, and so the parabolic current may be tilted in either direction, depending on the setting of the potentiometers. Notice that the potentiometers are connected so that the currents in the two coils tilt in opposite directions. R9 changes the loading on the sawtooth at the top of C8 and has the effect of tilting the parabolic current in each of the two coils in the same direction.

Blue horizontal convergence is accomplished by a parabolic current developed by the network consisting of L2, L3, C6, and C7. A portion of the 30-volt pulse fed to R10 is fed to the network. L3 and C7 are tuned to 15,750 Hz while L2 and C6 are tuned to the second harmonic, 31.5 kHz. The combination of these impedances causes a parabolic current in the convergence coil. By tuning L2 above or below its center frequency, the current may be tilted in either direction at will.

Pincushion Circuit

To correct pincushion distortion at the top and bottom of the screen, it is necessary to decrease the vertical-deflection current in the yoke when the horizontal scan is at the

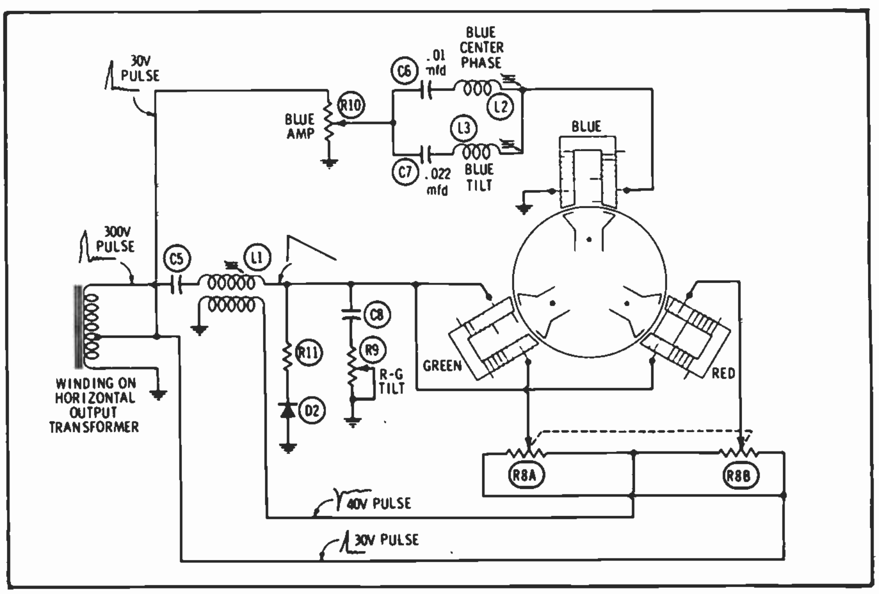


Fig. 17. Typical horizontal-dynamic-convergence circuit.

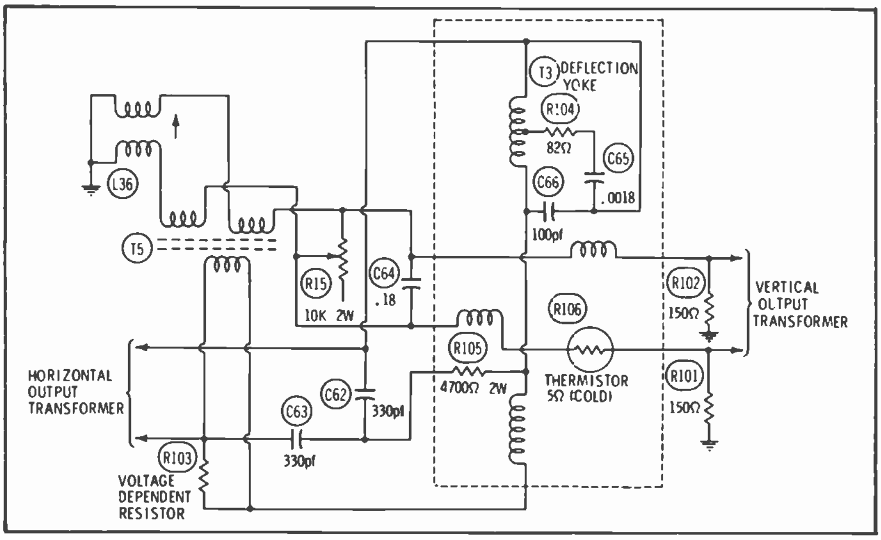


Fig. 18. Typical pincushion-correction circuit.

extreme right and left edges of the screen. Correcting pincushion distortion at the sides of the screen is accomplished by decreasing the amplitude of the horizontal deflection current when the vertical deflection is at its maximum upper and lower limits. The circuit in Fig. 18 shows one way of accomplishing this.

In Fig. 18, a portion of the horizontal deflection current flows through the bottom winding of T5, inducing a current in the two secondaries. The phase of this induced current is adjusted by the slug in L36 so that it opposes the vertical sweep current at the time when the horizontal scan is at the edges of the screen. Thus, the vertical deflection is reduced only at the edges of the raster. The amount of reduction of the vertical sweep is determined by the setting of R15.

Side pincushioning is also corrected by the same circuit. A portion of the vertical-deflection current passing through the two upper windings of T5 induces a voltage in the lower winding which opposes the horizontal deflection current. This opposing voltage must be present only when the top and bottom of the CRT are being scanned. This is accomplished by using a saturable core in T5. The core saturates when the vertical sweep is near the center of the raster and, consequently, there is only a small transfer of energy in the transformer. As the core comes out of saturation, more and more vertical-sweep energy is coupled to the horizontal-deflection system so that the scan lines are shortened at the top and bottom of the picture.

The Sony One-Gun Color Picture Tube

The Sony one-gun Trinitron* color picture tube often is confused with a different color CRT called the **chromatron,** which never got into a commercial set. But they're definitely not the same. Nor even much alike. To end any mixup, here's a thorough description of the Trinitron, how it works and what its advantages are.

For the most part, this little portable looks and operates much like any other color set.

That is, it does until you open it up and look for the adjustments. Then it's a different story. There are so few—only a half-dozen, including convergence. That's a far cry from the couple dozen a conventional picture tube needs.

*A Sony Trademark.

Color by Vertical Stripes

The first odd thing you notice about a Trinitron picture tube is its color phosphors. There are three primary colors—red, green, blue—just as in a regular color CRT. But these in the Trinitron are deposited on the tube face in vertical stripes, side by side. There's a red, then a green, then a blue; then another red, another green, another blue, and so on, as shown in Fig. 19.

The second unusual thing is the Trinitron shadow mask. It doesn't have holes like a conventional one. It has **slots.** It's not even called a shadow mask; it's called an **aperture grille.**

You can see its relationship to the phosphor stripes in Fig. 19. It's just behind the stripes, between them and the guns. The vertical slots line up with the green phosphor. Each slot is only as wide as one stripe. You can see from the sketch, the slots and stripes are both much narrower than the beam of electrons from the electron gun.

One Gun, Three Beams

To understand how the aperture grille works, you need to know what the electron gun of the Trinitron does.

Conventional color CRT's have three guns. The Trinitron has only one. But it puts out three beams. How that's done is illustrated in Fig. 20.

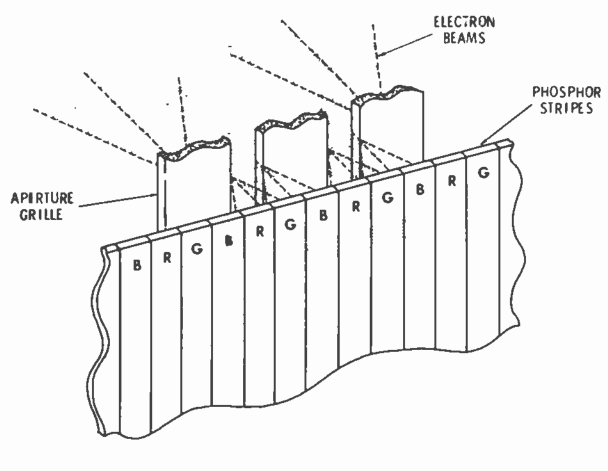


Fig. 19 Phosphors are deposited on screen in vertical stripes. Aperture grille is really a vertically slotted shadow mask.

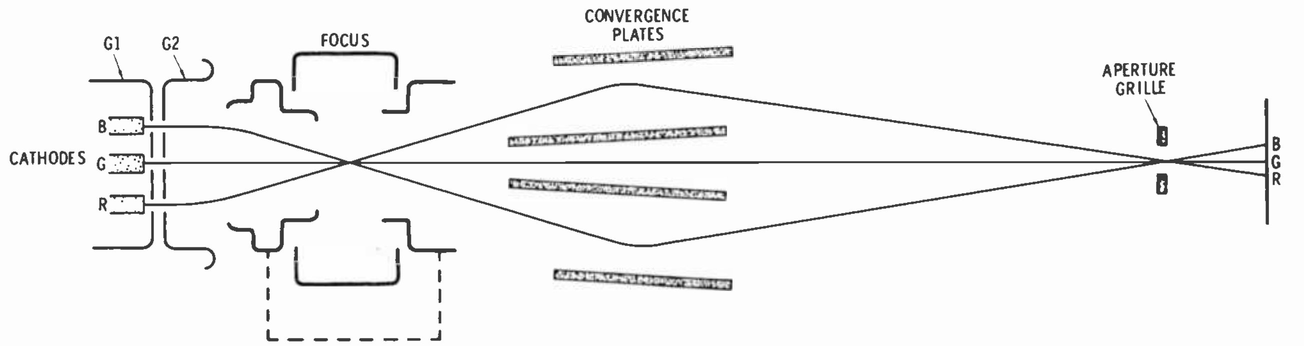


Fig. 20 Structure of Trinitron electron gun, which produces three beams. Simplicity makes surrounding circuits very simple.

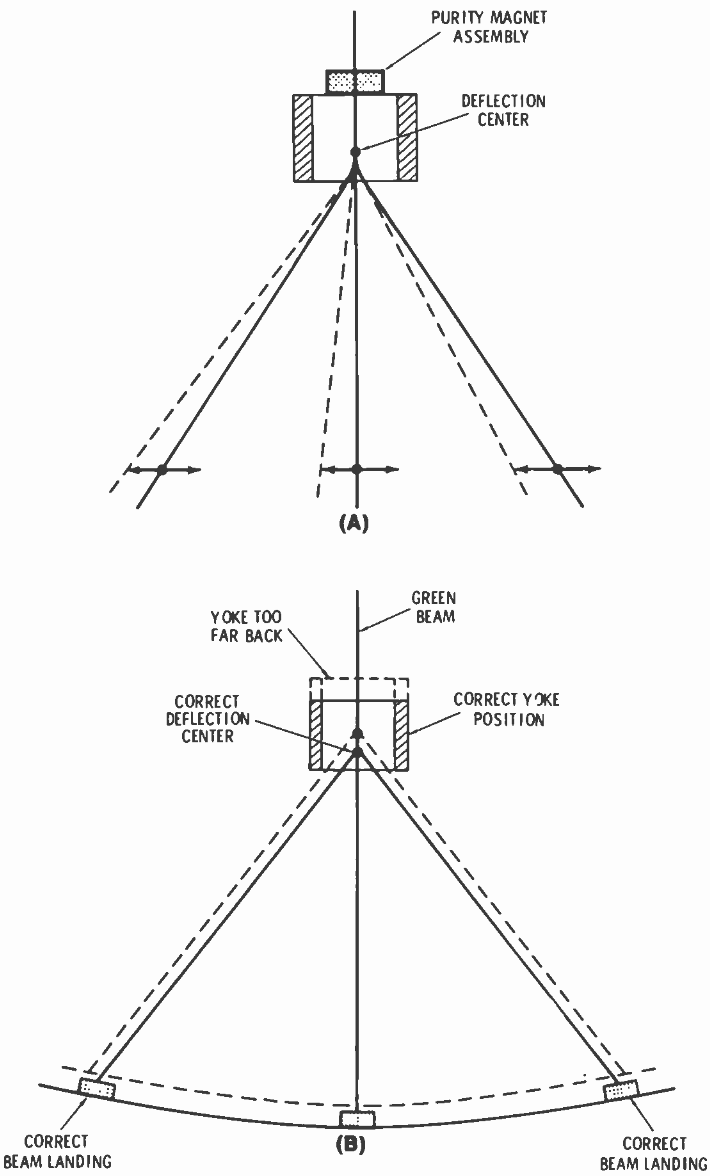


Fig. 21 Purity adjustment for center is with purity magnet; edge purity is adjusted by sliding yoke. Best to check green and red for purity.

There are three cathodes. Each one emits a bunch of electrons that are gathered roughly into a beam when they go through a hole in the control grid, or G1. The three cathodes are in line horizontally—side by side—and so are the three holes in G1. Green is in the middle.

Another grid, G2, is an accelerating anode. It's also called the **screen** grid. It speeds up the electrons in the three beams as they pass through the three holes in the G2 structure.

The beams, now with plenty of energy, enter the first of a set of focus elements. A DC field bends the blue and red beams toward the middle. At one spot inside this focus assembly, they cross over each other and the green beam. In the same process, each stream of electrons is concentrated into a tight, round beam. Green continues straight ahead. The red and blue beams angle off, aimed toward opposite edges of the screen.

Before they've gone very far, all three beams pass through an array of convergence plates. Those plates carry the same 19 KV supplied to the second anode. The static field and the shape of the plates bend the red and blue beams back toward each other and the green beam. If the voltage is exactly right, all three beams meet again (converge) right at the aperture grille.

One Color Per Beam

The aperture grille and its relationship to the deflection yoke are the key to color purity and easy convergence in the Trinitron.

First, look at Fig. 22 and how the aperture grille masks each of three beams. It allows each beam to hit only the phosphor it's supposed to excite.

Each beam covers at least two of the slots in the grille. The green beam, coming straight in, hits green phosphor stripes. The red beam comes into the slots at an angle set by the convergence plates. It strikes two red stripes, but can't hit blue or green—the grille prevents it. Likewise, the blue beam, coming through the grille slots from the opposite angle, can't hit red or green stripes—only blue.

Getting purity in the center is easy. But when the yoke sweeps the beams, they could easily land wrong. The yoke has to be in just the right position along the CRT neck so the point of deflection (called **deflection center**) is correct. Fig. 21 illustrates this.

A purity magnet (Fig. 21A) mainly affects how the beams land near the center of the screen. The arrows show how shifting the magnet moves beam relationships. If you're adjusting the purity magnet, you try for best purity in the center. Use a green raster.

Fig. 21B illustrates yoke positioning. If it's too far back on the neck, the deflection center is out of position. Out near the edges, the beams don't go through the slots like they're supposed to. You adjust for best purity all over the screen by sliding the yoke forward or backward on the neck of the CRT.

There's one more adjustment to assure that each beam lands on only its intended color. It's called the **Neck-Twist Coil.** Slight misalignment in the gun may keep the three beams from lining up horizontally. The effect is shown above the photo of the coil, in Fig. 23. The red or blue beam may spill over onto some green stripes. Turning the neck-twist coil corrects this. During adjustment, watch a red raster; this adjustment doesn't affect green.

Converging at the Edges

With purity as good as you can get it, turn your attention to keeping the beams going through the slots together all the way across the screen. The face of the Trinitron is only slightly convex. The angle of beam deflection inscribes a much

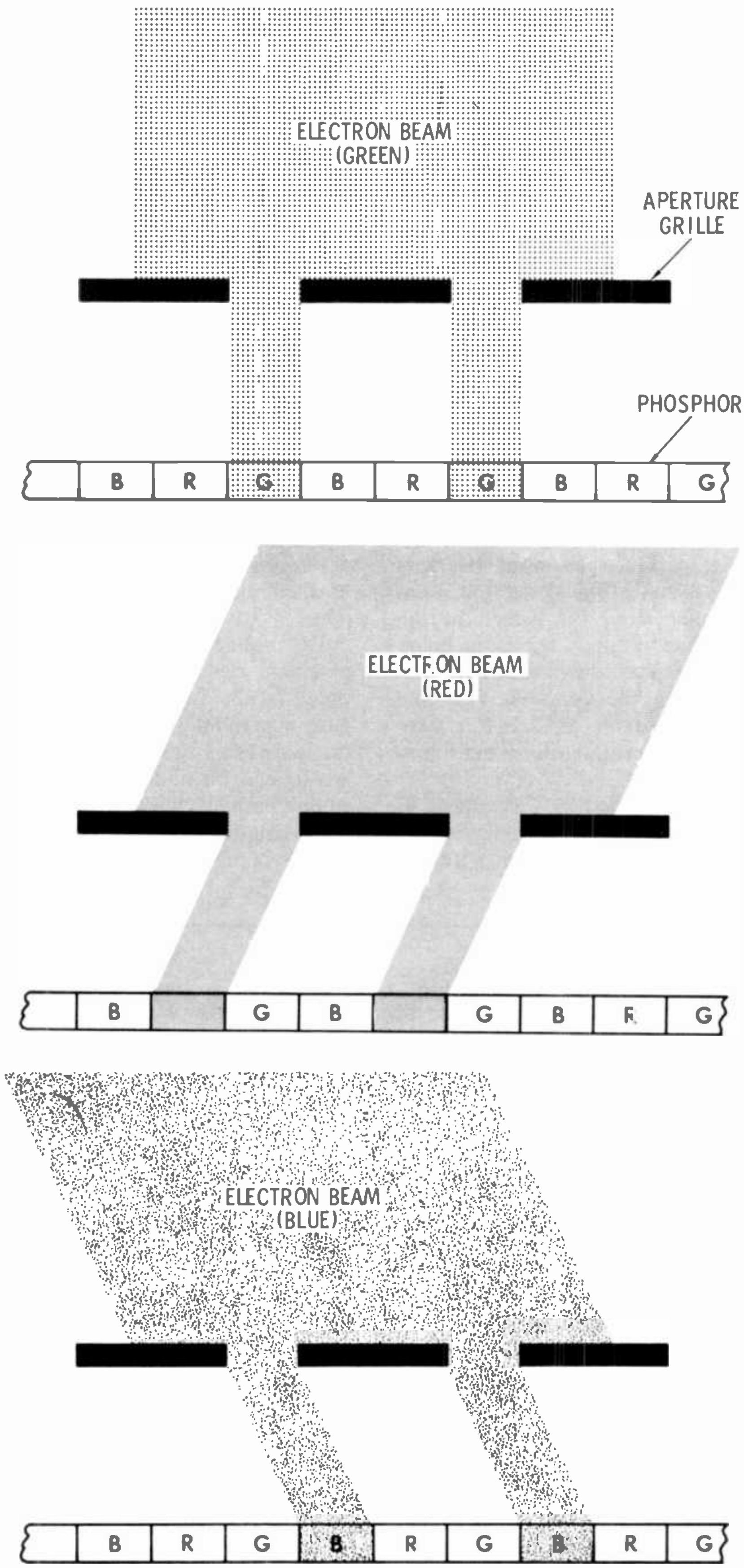
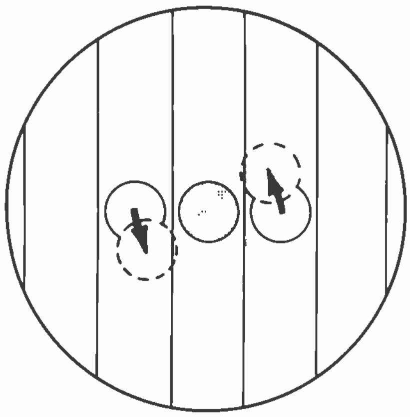
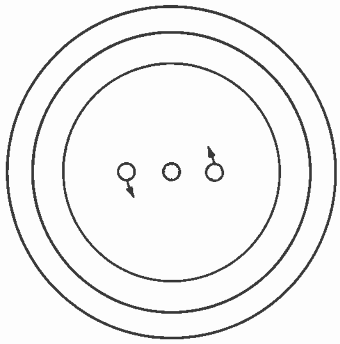


Fig. 22 Masking effect of aperture-grille slots. Angles of beams are adjusted so they illuminate only the phosphor for their own color.





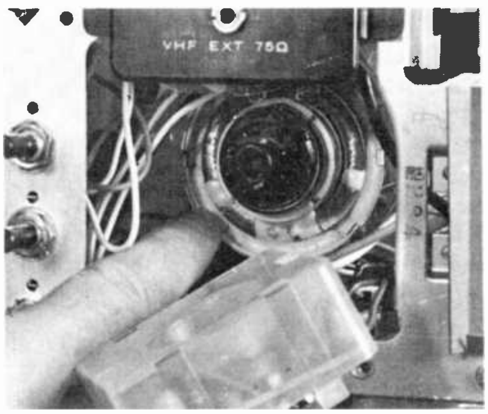


Fig. 23 Neck-twist coil is rotated to improve green purity. DC flowing in it keeps the three CRT beams lined up horizontally.

shorter radius.

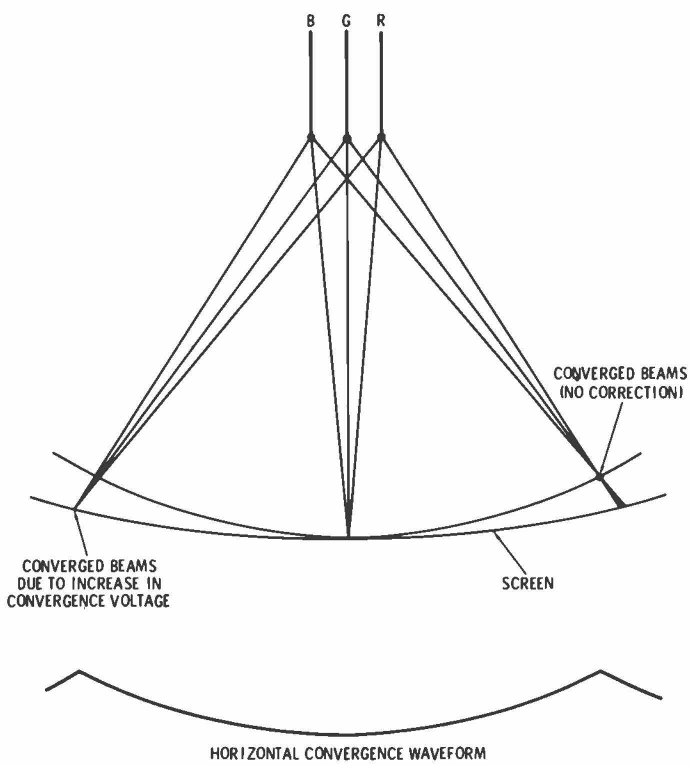
Fig. 24 illustrates what this does to the beams. The deflection point is further from the screen at the edges than at the center. If the beam crossover point stays the same distance from the gun, out near the edges the beams cross before they reach the aperture grille. That's misconvergence.

What's needed is something to lengthen their focal distance at the beginning of each line, let it return to normal as the beams sweep across center, and lengthen it again as they move on toward the right edge.

DC voltage on the convergence assembly determines the crossover point in the first place. So, logically, a parabolic voltage that's perfectly in step with the horizontal sweep can be added to it; the resulting voltage variation alters the focal length as needed.

The parabolic voltage waveform is shown in Fig. 24, below the sweep sketch. Applied to the convergence plates, the parabolic voltage adds to the 19-KV DC that's already there. Rising and falling, it stiffens the beams during the start and finish of each line—moving the focal point out and back to conform exactly with aperature-grille curvature.

That's all there is to horizontal convergence. The circuit is simple. A parabolic signal is taken from the horizontal-yoke return circuit



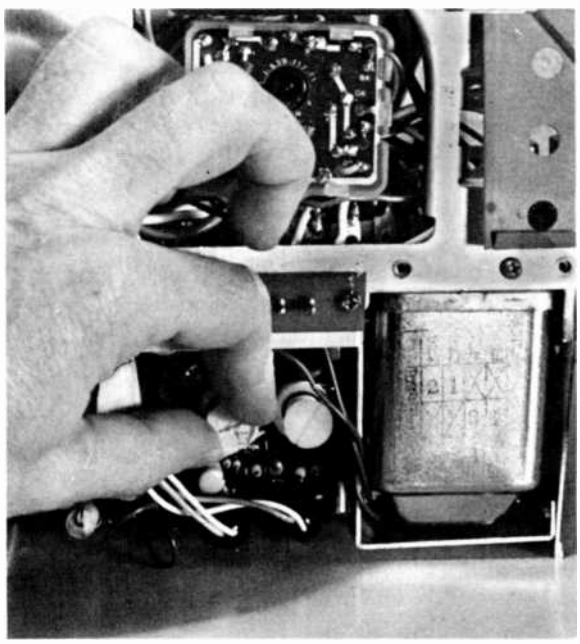


Fig. 24 Sketch shows how parabolic waveform added convergence plates lengthens focal point of beams as they sweep toward edges. The controls in the photo are the only convergence adjustments you normally will make.

and fed through a capacitor to the convergence plates.

Vertical convergence is even simpler. Slight vertical misconvergence doesn't show up anyway; if a beam lands a little high, it's still on the same color stripe so there's no visible error. Whatever slight correction is needed is designed into the vertical output stage and deflection coils.

There are only two convergence adjustments (Fig. 24). One is the **Vertical Static** adjustment. You make it first. Just put horizontal crosshatch lines on the screen and adjust the control to converge them. The other, the **Horizontal Dynamic** adjustment, primarily affects vertical crosshatch lines. It makes one end or the other of the parabolic signal higher in value, in case a little extra beam-stiffening is needed toward one side of the screen. There's also a capacitor you can change to improve convergence if a replacement Trinitron cannot be set up correctly.

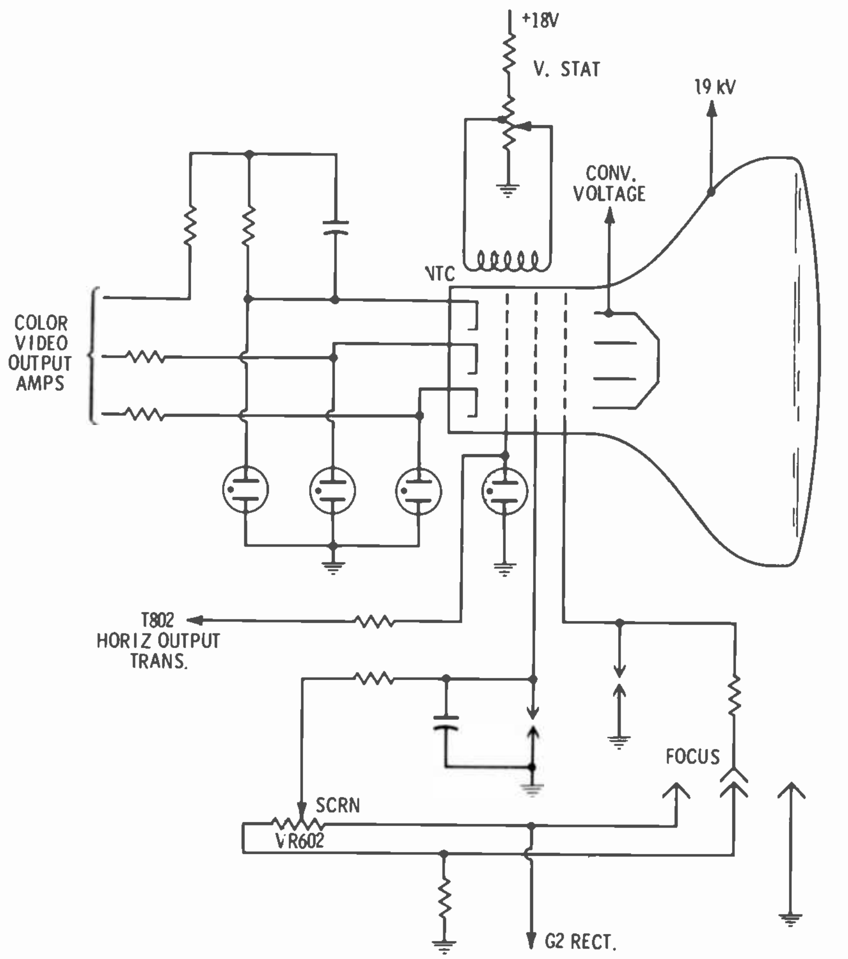
Powering the Trinitron

Supply circuits for the Trinitron are diagramed in Fig. 25. The photo shows the second-anode button and the 19-KV high-voltage connection.

As mentioned earlier, there are three cathodes. They are the driven elements in the Trinitron. Color video signals, to which the Y or brightness components have already been restored, are fed to the cathodes. They control how many electrons get into the beams, and thus determine average scene brightness and the mixture of primary colors that make a color scene. About 70 peak-to-peak volts of video are applied to them.

The control grid (G1) is common to all three beams. With the cathodes driven, G1 stays at ground (zero volts) potential most of the time. Average bias is set by whatever average DC is on the cathodes —about 100 volts DC. However, horizontal and vertical blanking pulses are fed to the grid, driving it deeply negative for those short intervals when they are present.

The efficiency of Trinitron phosphors is closely matched for all colors. Also, because of the common G1, transconductance for each beam is about the same. There's no need for complex gain adjustments,



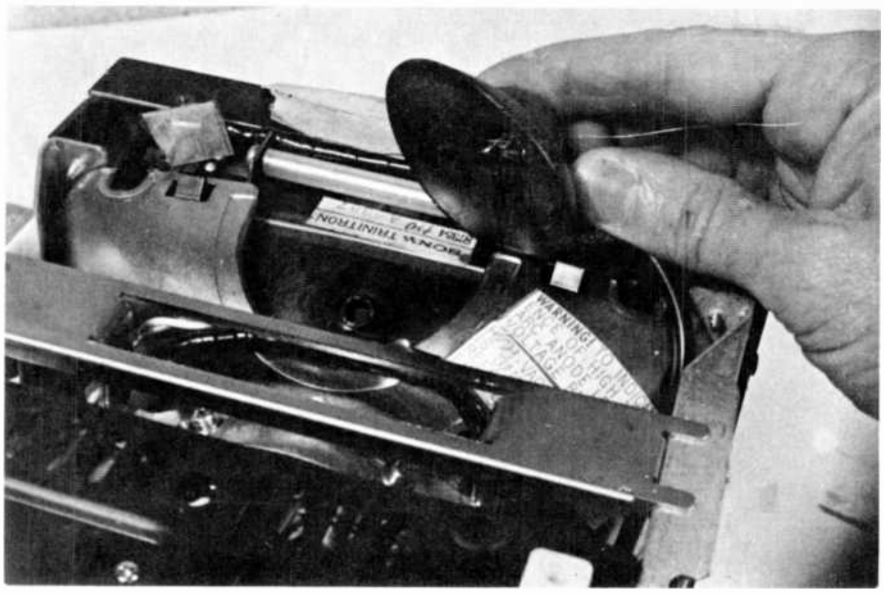


Fig. 25 Circuitry surrounding Trinitron picture tube stresses simplicity, compared with shadow-mask-tube circuits. Photo shows second-anode cap at top of tube instead of at side of bell as in conventional CRT's.

so·the screen (G2) is common to all beams, too. No gray-scale tracking is necessary. There's no series of screen adjustments—just one. It varies the DC voltage applied to G2—between 300 and 400 volts.

Focus voltage can be varied from zero (ground) potential to the full voltage that's available for G2. I have already indicated the voltage on the convergence plates; it's 19 KV, plus the parabolic dynamic-convergence signal.

A Step Toward Easier Servicing

The Trinitron is simpler than picture tubes of the three-gun shadow-mask variety. There's less to go bad inside and outside a Trinitron.

The Trinitron color portables I've seen in operation produce a satisfactory picture. Convergence was good, and colors seemed warm and true. Color pictures are bright enough to be viewed comfortably under strong flourescent lighting, but they don't seem any brighter than (if as bright as) pictures on the new black-surround color CRT's in some U.S. brands.

Whatever its other advantages, the Trinitron is unique and easy to set up.

Conclusion

The preceding chapter was written to explain how the geometry of the color CRT places certain requirements on the circuitry associated with it. After these requirements were pointed out, typical circuits which perform the various functions were described. Since specific circuits vary widely from one make of set to another, generalized setup procedures are almost pointless and no attempt has been made to formulate them. Through careful attention to the appropriate service data, and experience in making the various adjustments, you will evolve the techniques best suited to your specific requirements. ▲

Chapter 7

Solid-State Circuitry In Color Sets

Regardless of whether transistors or vacuum tubes are used in a color receiver, the functions of the several circuit systems remain unchanged. That is, a solid-state set still has a tuner, IF strip, horizontal and vertical deflection systems, chroma system, etc. Thus, the radical changes brought about by the adaptation of transistors to color-receiver circuitry occur in the circuits themselves, not in the overall design concept. There is one exception to this statement; the use of solid-state devices makes modular construction much more feasible than it was in vacuum-tube designs. A vacuum-tube color receiver of modular construction has been marketed (Setchell-Carlson), and at least one manufacturer is contemplating a solid-state color receiver which does not have modular construction. The relative merits of the two types of construction are not an appropriate subject of discussion here.

Review of Transistors

In presenting the following information, we assume that the technician has some prior knowledge of transistor circuitry. For this reason, we will not dwell on the theory of their operation. The physical laws which govern the operation of solid-state devices have been published many times and the interested reader may pursue the subject by studying any of a host of books. For our present purposes, we will consider most of the transistors used in a color receiver as "black boxes" which amplify.

It is worth noting that some of the characteristics of transistors dictate significant changes from conventional tube circuitry. In their most usual configuration, transistors exhibit very low input impedance. For this reason, resistors in the circuit are usually relatively small and capacitors have relatively large values.

Fig. 1 will help to illustrate this important point. Assuming that each circuit must operate at 1 kHz, we will calculate the value of C1 which will allow reasonably good energy transfer from the first stage (represented by a generator) to the second stage. As a general rule of thumb, the impedance of the coupling capacitor should not exceed 1/10 of the impedance of the input circuit of the following circuit. In the vacuum-tube circuit (Fig. 1A) the maximum impedance of C1 may be no greater than 50K ohms and the capacitance may be calculated by

$$C = \frac{1}{2\pi \quad F X_C}$$

$$= \frac{1}{6.28 \times 10^3 \times 5 \times 10^4}$$

$$= \frac{1}{3.14 \times 10^8} = 3.18 \times 10^{-9}$$

$$C = .0032 \text{ mfd}$$

To maintain the same impedance ratio between coupling capacitor and input, the value of C1 in the transistor circuit (Fig. 1B) may not exceed 47 ohms.

$$C = \frac{1}{2\pi \quad F X_C}$$

$$= \frac{1}{6.28 \times 10^3 \times 4.7 \times 10^1}$$

$$= \frac{1}{2.95 \times 10^5} = 3.39 \times 10^{-6}$$

$$C = 3.4 \text{ mfd}$$

Because large coupling capacitors are necessary, it is sometimes more practical to eliminate them entirely and couple successive stages directly. This is also done in vacuum-tube circuitry where response to very low frequencies is desirable, but each direct-coupled stage requires, roughly, at least 100 additional volts from the power supply, a serious limitation. Transistors operate from more moderate voltages than tubes, and several stages may be direct coupled even though the power supply produces only 100 volts or less. By using alternate NPN and PNP transistors in a direct-coupled cascade, an almost unlimited number of stages may be direct coupled without increasing the power-supply voltage.

While the use of direct-coupled transistor stages is desirable from the point of view of economy and design simplicity, it may introduce a troubleshooting problem. A direct-coupled amplifier will pass DC as well as any other frequency, and so any change in emitter-to-base DC potential of the first stage is amplified in each following stage. Thus, if a shift in bias of 0.1 volt should occur at the front end of a string having a gain of 100, a change in output of 10 volts would result.

In practical applications, a DC swing of 10 volts is nearly impossible because some transistor in the string probably would either saturate or cut off at a smaller shift in bias. The result is the same—the problem area is in front of the circuit where the symptoms are most likely to be detected.

DC feedback loops and current limiters are often used in transistor circuits to prevent the runaway condition just described. The circuit shown in Fig. 2 has such a feedback path. (The circuit is not in the Motorola receiver.) Assume that the positive base voltage of Q9 decreases for some reason. The conduction of Q9 will decrease, causing the base voltage of Q10 to rise. The increased conduction of the collector of Q10 causes an additional drop across R51, and the base of Q11 becomes less positive. This causes increased collector current in Q11 and the collector swings positive because of the increased drop across T1.

Without a feedback loop, a small decrease in bias at the base of Q9 might well cause such a large increase in collector current through Q11 that the transistor would be damaged. The feedback loop consists of R55, C19, and R54, connected between the collector of Q11 and the base of Q9. By virtue of this loop, as the collector voltage of Q11 tends to rise, a portion of this positive-going voltage is coupled back to the base of Q9 to increase its conduction. Thus, the action of the feedback loop is always to oppose any change in the DC operating potentials of the transistors. Note that C19 acts as a low-pass filter in the feedback loop. Because of it, only DC, or low-frequency AC, is fed back to Q9.

In our discussion about the DC feedback loop, we assumed that the potential at the base of Q9 changed for some unspecified reason. While idle conjecture about the cause of this shift is pointless, it is true that changes within Q9 itself, or variations among transistors in a production run, are a major source of the variations. Thus, we may conclude that DC stabilization is required for two reasons: the DC instability of direct-coupled amplifiers, and the inherent DC instability of transistors themselves.

In general, it is also true that the gain stability of transistors is poorer than it is for vacuum tubes. For this reason, an AC feedback loop also may be incorporated in an amplifier string. Referring again to Fig. 2, the AC feedback network consists of R53 and C49. A portion of the amplifier output is fed back to the input in phase opposition. Thus, if the gain of the overall circuit increases, so does the amplitude of the inverted signal fed to the input, and the gain is maintained constant. Note that this feedback circuit incorporates a high-pass filter so that it is insensitive to DC.

Another characteristic of transistors which is sometimes confusing to the technician is the manner in which they fail. Although a triode with a grid-to-plate short is quite rare, a base-to-collector short in a transistor is a distinct possibility. Furthermore, while a grid-plate short probably would be destructive to associated components, a base-collector short may cause no external damage. A transistor with a base-collector short may pass the signal (without amplifying it, of course) but there will be no inversion of the waveform.

Typically, vacuum tubes fail because of gradually decreasing cathode emission and lowered transconductance. This is unlike transistors which usually maintain a nearly constant gain throughout their lives; failures are due to shorts or excessive leakage.

Motorola 23TS-915 and 919 Chassis

Chroma Circuitry

Fig. 3 is a block diagram of the chroma circuits used in the Motorola 23TS-915 and 919 chassis.

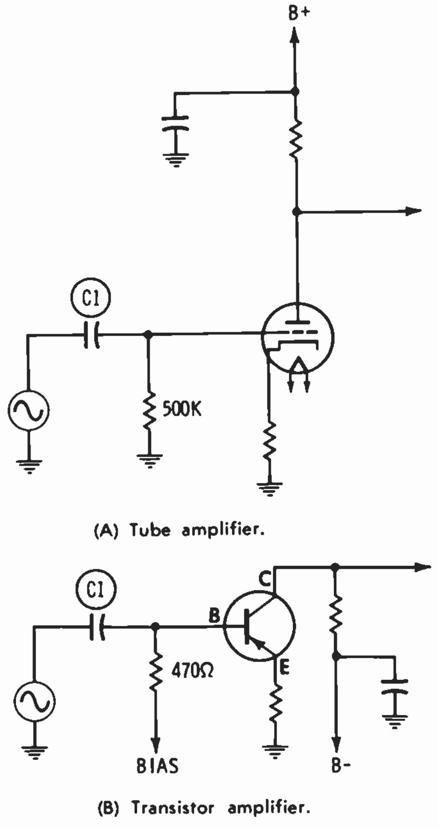


Fig. 1. Comparison of transistor and vacuum-tube amplifiers.

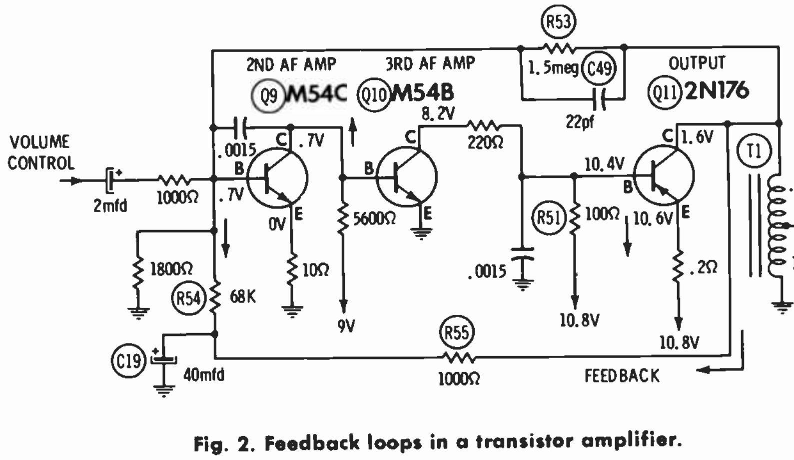


Fig. 2. Feedback loops in a transistor amplifier.

Although most of us are accustomed to only about half this number of blocks, the functions performed by this chroma circuit and one using vacuum tubes are essentially the same. With the exception of the demodulators and color amplifiers, which are rather novel, the entire chroma circuit is quite similar to a vacuum-tube design.

Reference Circuits

Following the same sequence that was used in Chapter 4 of this book, we shall examine the reference-signal circuits first. Video from the first video amplifier is fed to the first chroma-bandpass amplifier, which has two outputs. One of these outputs, consisting of the chroma sidebands and the color burst, is fed to the chroma-sync amplifier. The chroma-sync amplifier is the equivalent of the burst amplifier of conventional designs and is a coincidence gate. Also fed to the chroma-sync amplifier is an enabling pulse from the burst amplifier and pulse limiter. This pulse gates the sync amplifier on at the time when the color burst is present at its input, allowing it to pass. During the remainder of the scanning interval, the sync amplifier is gated off, removing the chroma sidebands from its output.

The color burst from the sync amplifier is developed across the 3.58-MHz crystal, causing it to ring from one burst to the next. The crystal is amplified by Q53 and the output of this stage is a CW signal which is rephased at the start of each horizontal scan.

The chroma-reference oscillator, Q54, is a Colpitts oscillator which free runs during black-and-white reception. However, if a color burst is present, the output from the crystal amplifier phase locks the oscillator, causing its output to be in phase with the color burst.

The chroma-reference phase inverter is actually a splitter which develops two out-of-phase signals from the single input. A potentiometer across these outputs, the hue control, selects the specific phase of reference signal required for correct hue of the picture.

Finally, the reference signal is amplified by Q56 and fed to the three demodulators. The phase of the signal from Q56 is correct for the green demodulator, and phase shifting networks develop the correct phase for the blue and red axes.

ACC and Color Killer

A portion of the output from the 3.58-MHz crystal amplifier is rectified and fed to the ACC amplifier. If the amplitude of the burst increases, the output of the crystal amplifier increases, and this, in turn, causes the bias developed in the ACC amplifier to increase. This decreases the gain of the first chroma bandpass amplifier to maintain a constant-amplitude output with a varying input. (See "Closed-Loop ACC" in Chapter 5 of this book.)

The output of the ACC amplifier also is supplied to the killer amplifier, Q43. In conjunction with Q44, the killer output stage, Q43, operates as an electronic switch. That is, the circuit is either cut off or saturated—there is no "in between." In the absence of a color burst, the

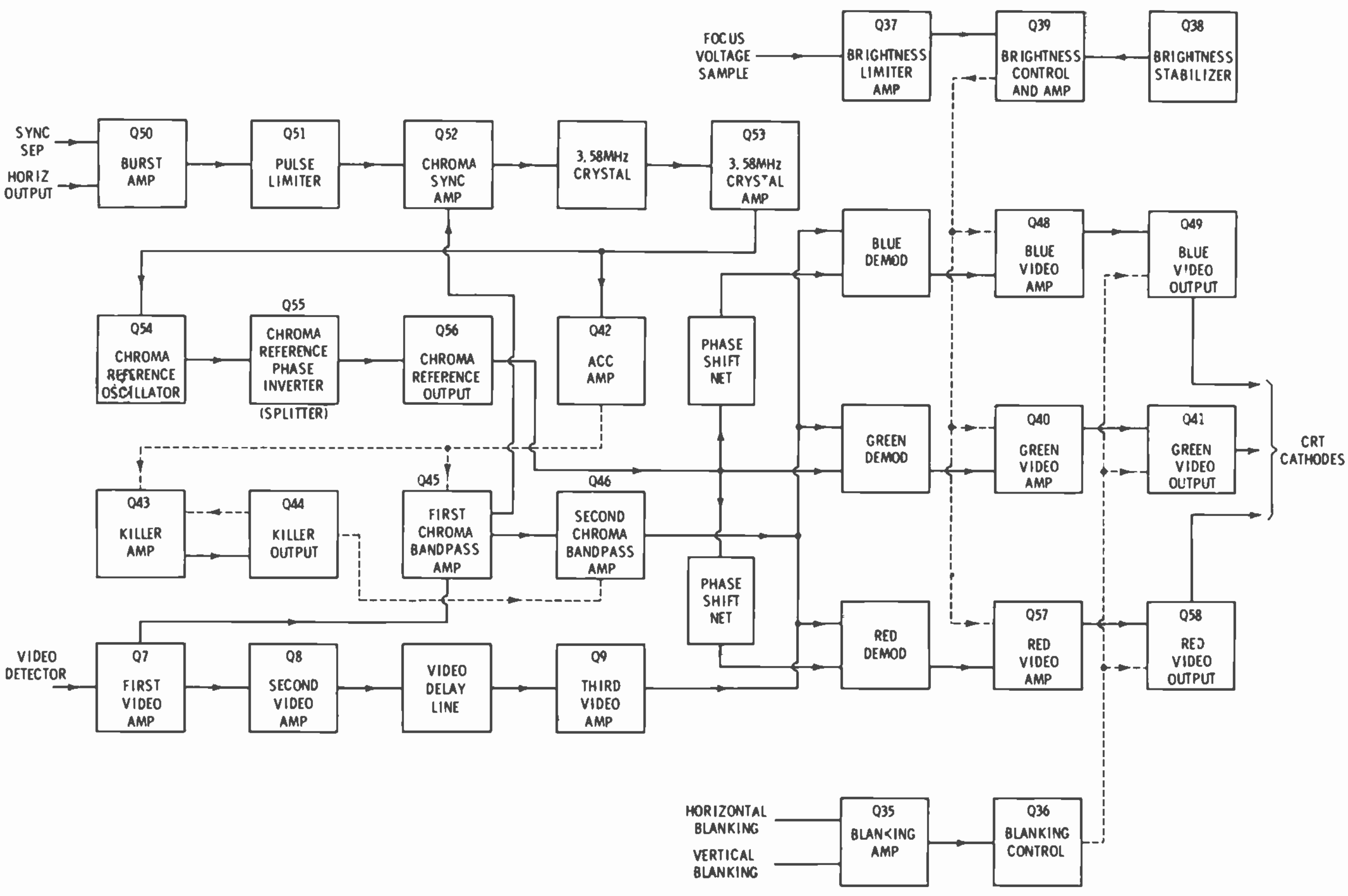


Fig. 3. Detailed block diagram of the Motorola solid-state chroma circuits.

killer cuts off the second chroma-bandpass amplifier; when a burst is present, the second chroma-bandpass amplifier is turned on. The color intensity control is located in the network between the killer and the second bandpass amplifier.

Bandpass Amplifiers

The two chroma bandpass amplifiers, Q45 and Q46, are straightforward in design. As stated above, the gain of the first stage is controlled by the ACC amplifier and the second stage is cut off during monochrome reception by the color-killer circuit. The color intensity control also is incorporated in the input of the second bandpass amplifier. The two stages are tuned to pass sidebands having frequencies up to 500 kHz above or below the burst frequency.

Chroma Demodulators

The diode chroma demodulators used in this receiver are similar in many respects to the ones used in some General Electric receivers. The essential difference is that a third input, the luminance (or Y) signal is also fed into the three Motorola demodulators.

Today, the majority of color receivers recombine the luminance signal and the three color-difference signals within the CRT. (See Chapter 4 of this book.) This practice was adopted because of its relative simplicity and economy, but it makes the ability of the receiver to track the gray scale dependent on the degree of nicety with which the three guns of the CRT can maintain identical transconductances. The concept of recombining the luminance and chrominance information outside the CRT allows more accurate adjustment of gray-scale tracking and the possibility of correcting for long-time variations in CRT parameters and also for changes in the external circuits. Time will tell whether or not this second-generation external matrix will fulfill expectations.

The three demodulators of Fig. 3 are identified simply by the color (not color-difference) signals which they produce. This is entirely proper, since color-difference signals never appear in this chassis; chrominance and luminance information are recombined (matrixed) in the demodulators. One of the demodulators will be examined in detail later in this article.

Color Amplifiers

Each of the color demodulators is followed by a pair of amplifiers which raise the signal level to an amplitude sufficient to drive the CRT cathodes. These amplifiers are direct coupled, allowing the brightness control to be connected to the emitters of the first stages of each amplifier string. Making the emitters of Q40, Q48, and Q57 more positive causes the collectors to swing in the same direction. This swing is inverted in the output stages, driving the CRT cathodes negative and increasing the brightness.

The emitters of Q41, Q49, and Q58 (NPN types) return to ground through the blanking-control transistors Q36. When Q36 is cut off during horizontal and vertical retrace, the video-output transistors also are cut off, driving the CRT cathodes positive into cutoff.

Brightness Circuit

The brightness circuit, Q37, Q38, and Q39, includes the manual brightness control and also an automatic brightness limiter (ABL). The purpose of the ABL circuit is to maintain constant CRT beam currents (for a given brightness-control setting) even though shifts in video-amplifier gain, line voltage, horizontal-output voltage, etc. might tend to change the brightness. Also, the ABL control allows the technician to preset the maximum brightness for optimum operation.

A sample of the focus voltage is used as a control voltage for the

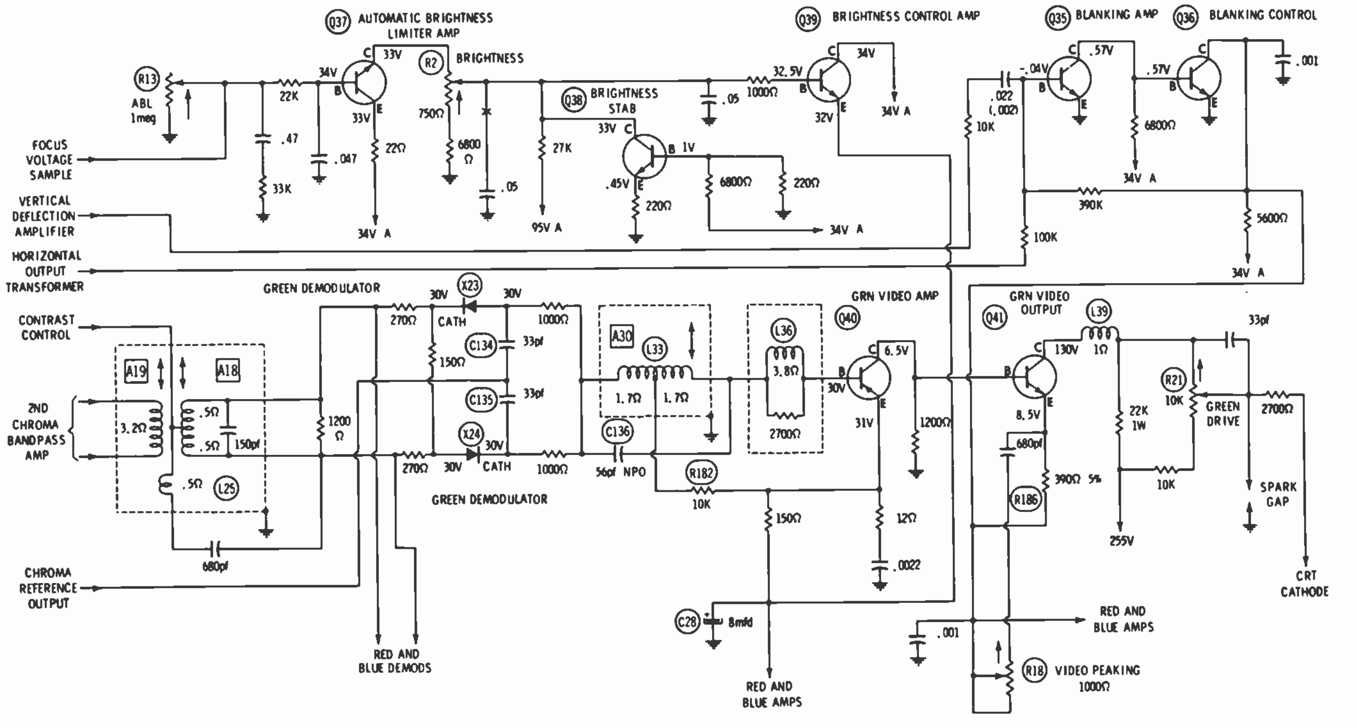


Fig. 4. Motorola demodulator and color amplifiers.

ABL. Any shift in brightness will shift the level of the control voltage. For example, if the CRT brightness decreases for some reason, the focus voltage will increase. A positive-going control voltage increases conduction of the video amplifiers, video outputs, and the CRT guns to increase brightness.

Blanking Circuit

Positive pulses derived from the vertical sweep amplifier and from the horizontal-output transformer are coupled to the blanking-control stage, Q36, by the blanking amplifier. These pulses cut off the video-output transistors during retrace.

To reduce the load on the horizontal-output circuitry, the horizontal retrace time of this chassis is longer than usual and the CRT is overscanned slightly. The width of the horizontal-blanking pulse is increased accordingly, and it is normal for a portion of the first color bar from a keyed-rainbow generator (yellow-orange) to be partially blanked and off the left side of the raster.

Chroma-Circuit Analysis

The three demodulators are practically identical, with the exception of the phase of the reference input and the value of the input attenuation. Since it serves no useful purpose to treat them separately, only the green demodulator is shown in Fig. 4.

During reception of a black-and-white signal, there are two inputs to the demodulator. A free-running, 3.58-MHz signal is fed to the junction of C134 and C135. Positive video from the contrast control, situated in the output circuit of the third video amplifier, is fed to the center of the secondary of L25. There is no input from the chroma-bandpass amplifier, since it is cut off by the color killer.

The reference signal produces no output at the base of Q40 because of the traps, L33 and C136, between the demodulator and the video amplifier. These same traps remove the 3.58-MHz ripple during color reception.

Positive video (a positive signal makes a black raster) passes through X24 and is developed across R182. The video passes through the trap and is duly amplified by the two video amplifiers and fed to the CRT cathode. Of course, the same things take place in each of the other demodulators and their amplifiers. Thus, the luminance signal is amplified and fed to the three CRT cathodes to produce a monochrome picture.

The operation of diode chroma demodulators is discussed in detail in Part 5 of this book.

To summarize, if the reference and chroma signals are in phase, a maximum output of one polarity is realized; if they are out of phase, a maximum output of the opposite polarity results; if they are 90° out of phase, the output is zero. In the demodulator of Fig. 4, in-phase signals produce a negative voltage at the base of Q40 and increase the CRT conduction.

voltage at the base of Q40 and increase the CRT conduction.

So far, we have considered the chrominance and luminance signals separately, but, since both of these signals are developed across the same resistor, R182, they are effectively added at this point. This sum (or difference) of the two signal voltages passes through the traps, which remove the 3.58-MHz ripple, to the base of Q40 and, ultimately, to the CRT.

Blanking Control

The blanking and brightness-control circuits also are shown in Fig. 4, not because the functions they perform are novel, but because the use of transistors results in unusual circuit configurations

The emitter currents of the three color-video output transistors must all flow through the blanking-control transistor, Q36. During scanning time, Q36 is forward biased and the complete circuit path of the collector current of Q41 is from ground, through Q36, R186 and Q41 to the CRT cathode and the 255-volt supply.

During retrace, Q35 is driven to saturation by the positive pulse on its base. This removes the forward bias from the base of Q36, cutting it off. This effectively opens the low-resistance path from ground to the emitters of the video-output transistors, cutting them off.

Automatic Brightness Limiter

The level of the sample of focus voltage taken from the low end of

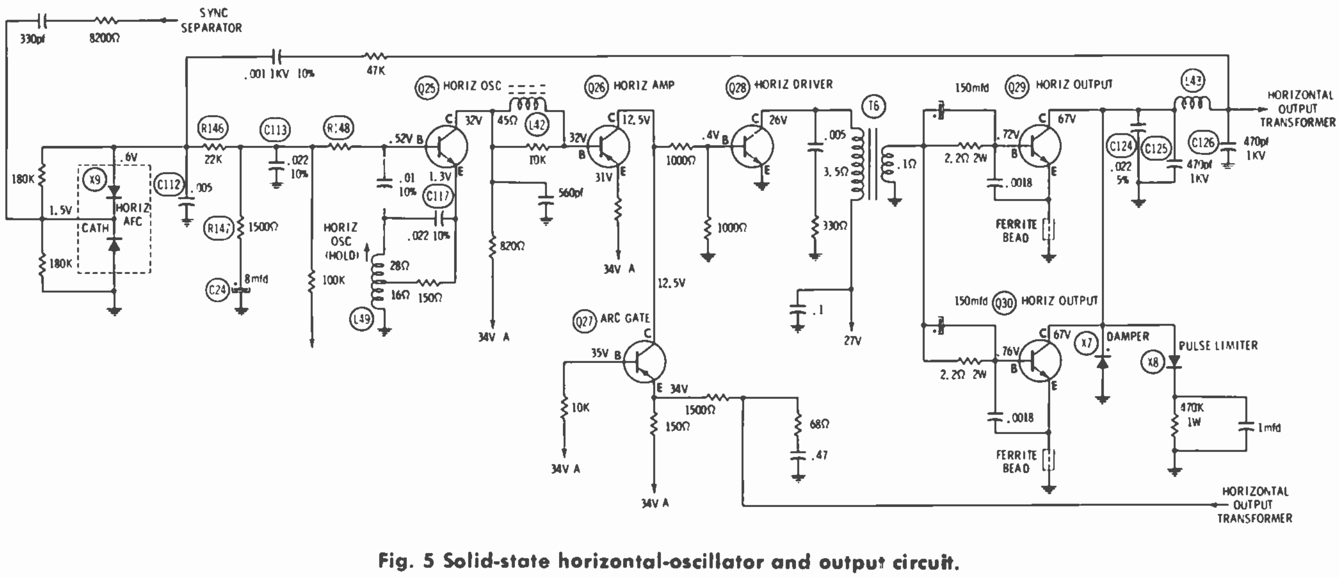


Fig. 5 Solid-state horizontal-oscillator and output circuit.

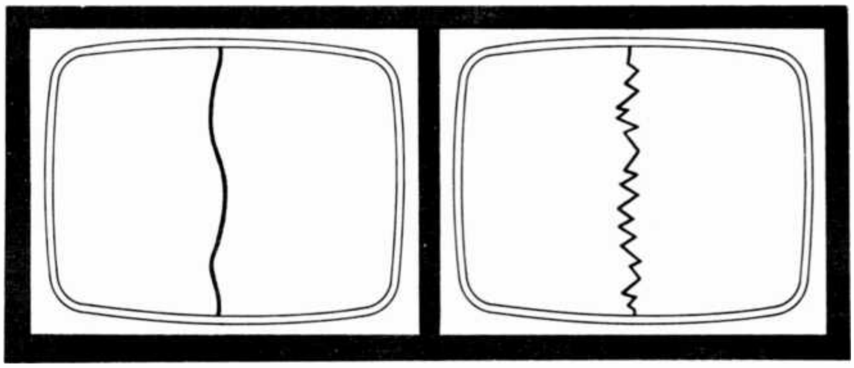
the CRT screen-control potentiometer indicates the total load on the high-voltage and focus-voltage supplies, and hence, the amount of CRT conduction, or brightness. The adjustment of R13, the ABL control, determines the size of the sample (control voltage) taken and acts somewhat as a coarse control of the brightness.

Q37 is connected as an emitter follower, roughly equivalent to a cathode-follower tube configuration, and the emitter voltage is dependent on the amplitude of the control voltage. Because of Q37, the voltage at the top of the brightness control cannot become more positive than the 34-volt supply at the collector of Q37. This limits maximum brightness of the CRT.

The voltage on the base of Q39 is determined by the setting of the control voltage. Q39 also is connected as an emitter follower, and the voltage on its base controls the bias on the video amplifiers, Q40, Q48, and Q57. Notice that the base voltage of Q39 is stabilized by Q38, which acts as a regulator for the circuit.

Analysis of Horizontal-Deflection Circuits

Space is not available for an analysis of all circuits in the Motorola solid-state chassis, so this discussion is necessarily limited to those of greatest significance. In addition to the chroma circuits already



(A) Hunting. (B) Jitter.

Fig. 6. Appearance of horiontal hunting and jitter

covered, the design of the horizontal deflection system is sufficiently different from designs using tubes to justify an analysis.

Horizontal Oscillator and AFC

The horizontal oscillator and the AFC circuit (Fig. 5) which controls its frequency are very similar to designs using tubes. The AFC circuit compares the phases of the horizontal sync pulse and the output of the horizontal-output transistors, Q29 and Q30, to develop a control voltage. This control voltage is integrated by R146, R147, C24, C112, and C113, and it is used at the base of Q25 to control the oscillator frequency.

The function of the integrating circuit between the AFC detector and the oscillator is not particularly mystifying, but it appears, from the number of letters we receive, that malfunctions in this circuit cause a great number of problems to our readers. The following comments apply to nearly all sets, vacuum-tube or solid-state.

Failures in the integrator fall into four general categories: (1) Loss of control voltage caused by R148 going open, for example. In this case, there is no horizontal sync. (2) Radical change in the DC level of the control voltage which causes a radical change in horizontal frequency, or may cut off the oscillator. (3) Too much integration of the control voltage. (4) Not enough integration of the control voltage.

The first two categories named above are generally not too difficult to diagnose, but the last two seem to cause many difficulties. If the control voltage is integrated too much, the response time of the system becomes too great. Therefore, the raster appears to slowly move back and forth across the CRT. The complete raster will not necessarily float the same amount, and so a

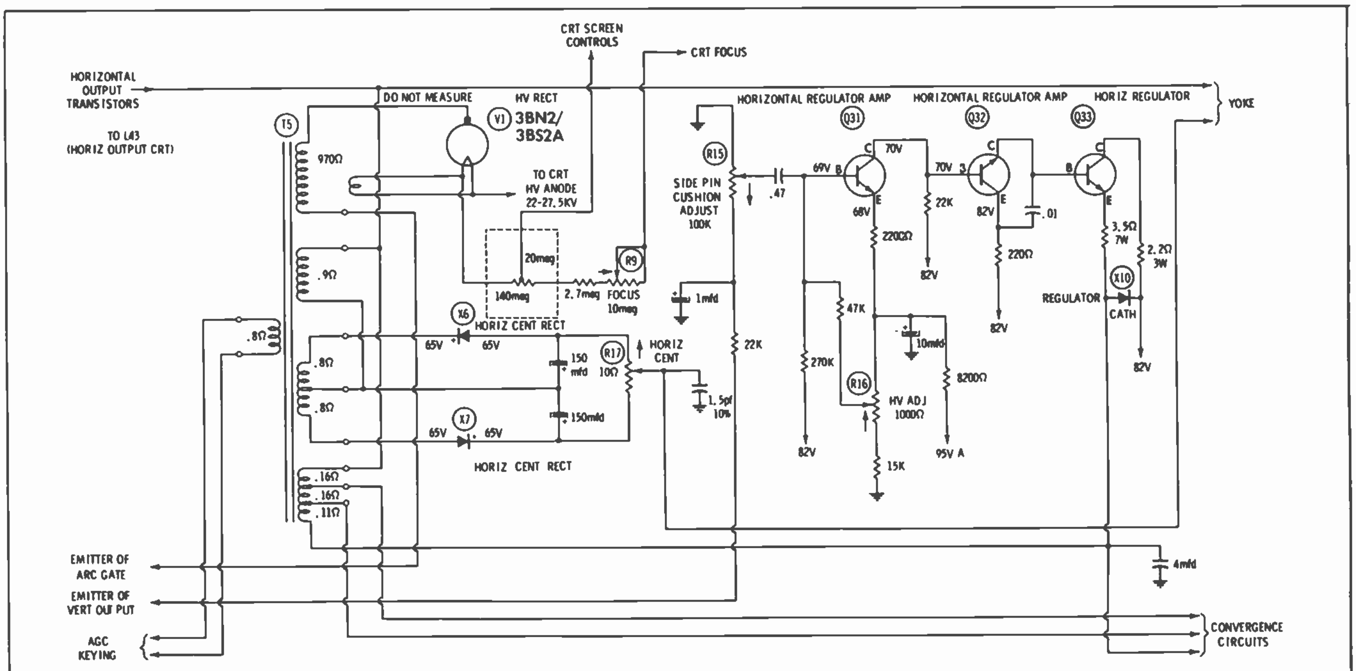


Fig. 7. Horizontal regulator circuit.

vertical line on the CRT may curve back and forth from top to bottom (see Fig. 6A). This is called horizontal hunting.

Too little integration causes the oscillator frequency to be overcorrected. With insufficient integration, the control voltage at the oscillator (or AFC tube) shifts slightly during each scan. This causes the scanning time of each horizontal line to be slightly different and a vertical line on the CRT appears ragged or broken as shown in Fig. 6B. This is called horizontal jitter. Jitter is usually caused by a decrease in value of C24 (or its counterpart in another set), hunting is usually caused by an increase in resistance of R146 or R147, or their counterparts.

Referring again to Fig. 5, the horizontal oscillator is a Hartley oscillator; L49 and C117 determine the frequency. When the top of the tank is negative, Q25 is cut off. As the top of the tank swings positive, the transistor begins conduction at some point on the sine wave. This point is determined by a combination of fixed bias and the control voltage from the AFC. The collector current of Q25 is a series of pulses and the waveform at the base of Q26 is approximately a square wave.

Amplifier, Driver and Output

The train of pulses is fed to the driver through the horizontal amplifier, and the output of the driver is coupled through T6. The phasing of T6 is such that the output transistors are turned on when the driver is cut off. Deflection of the trace from center to the right edge of the CRT occurs while Q29 and Q30 are conducting. At the instant that Q29 and Q30 are cut off, the sweep retraces, the damper begins conduction, and the left side of the raster is scanned.

The network consisting of C124, C125, C126, and L43 is a low-pass filter which prevents any high-frequency transients which may be generated in the amplifier, driver, or output stages from being coupled to the horizontal-output transformer. The pulse-limiter diode, X8, limits the amplitude of the collector pulses of Q29 and Q30 to protect them.

In case of a high-voltage short, a positive pulse is developed at the emitter of Q27, causing it to conduct. This clamps the base of Q28 to the base of Q27, causing it to conduct and cutting off the horizontal-output transistors until the arc clears.

Horizontal Regulator and Pincushion Circuit

The circuit which includes Q31, Q32, and Q33 (Fig. 7) performs the two functions of injecting a portion of the vertical deflection signal into the horizontal sweep system for pincushion correction, and it also regulates the high voltage. Both of these functions are accomplished by controlling the supply voltage for the horizontal-output transistors.

A parabolic voltage derived from the emitter of the vertical-output transistor is amplified by Q31 and Q32 and added to the supply for the horizontal output transistors. The supply voltage is increased when the vertical sweep is at the center of the tube, causing the horizontal scan to expand at this time to correct for pincushion effect. R15 picks off the amount of vertical-sweep voltage which is required for optimum correction.

At the same time, the relative values of the 82-volt supply and the 95-volt supply are compared in the emitter-base circuit of Q31. As CRT beam current increases, the load on the horizontal-deflection circuit increases and the output of the 82-volt power supply drops. This decreases the forward bias of Q31, causing the collector voltage to increase. This increase in collector voltage increases the conduction of Q32 and Q33, which tends to raise the supply voltage to the horizontal-output transistors, thereby regulating the high voltage.

Admiral's K10 Hybrid Color Chassis

To reduce the possibility of damage occuring in one section while working on another, Admiral has designed two etched-circuit boards: One contains the solid-state circuitry; the other, the vacuum tubes. These etched boards are separated into zones by heavy white lines. Each zone is identified by a large white letter designating the section.

- A—Video IF
- B—Sound, AGC, and video amplifiers
- C—Chroma amplifiers
- D—Chroma oscillator and color killer, color burst amplifier, and color demodulators
- E—Vertical
- F—Horizontal
- G—Convergence
- H—Miscellaneous chassis components
- U—UHF tuner
- V—VHF tuner

This coding system also makes it easier to find the components on the schematic. Resistor RC64 on the wiring diagram will be found in zone "C", with the number "64" written beside it on the board.

Circuit Description

Admiral chassis K10 is comprised of twenty-six transistors, twenty-four diodes, seven vacuum tubes and a CRT. The transistors are used for all signal-processing functions, while the vacuum tubes are used in circuits that have heavier power requirements.

Power Supply

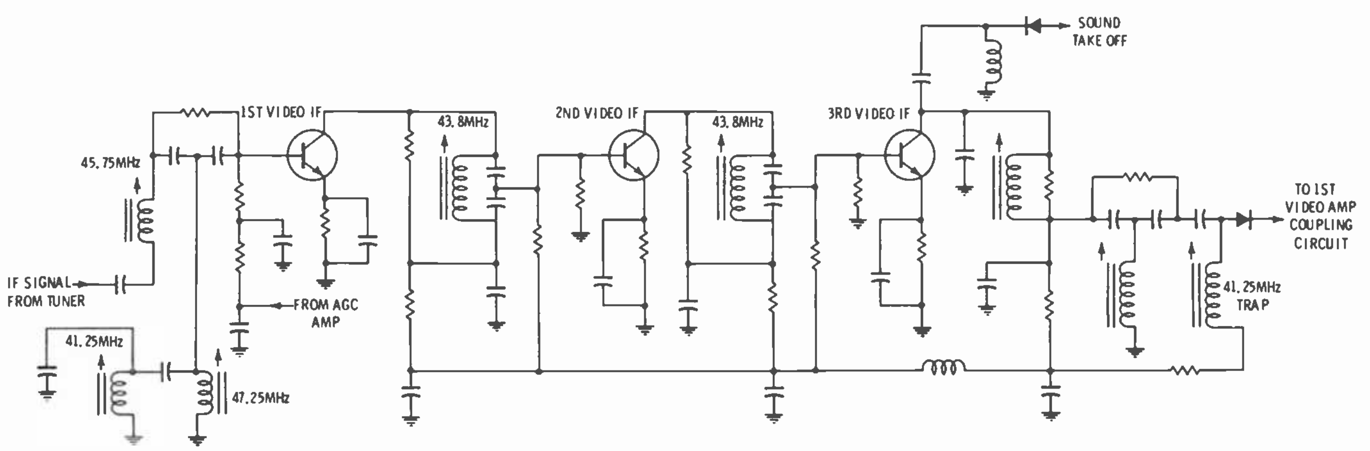
The power transformer is used only for the transistor circuits, although it does have a 6.3-volt AC tap to supply filament voltage to the CRT. A line choke is used to prevent any unwanted RF signals from entering the various circuits through the source voltage.

Video IF Strip (Fig. 8)

The outputs of the first and second video IF transistors are applied across tuned coils. These coils are shunted by a pair of capacitors to form a capacitance voltage divider. The input signal for the next stage is obtained from the junction of these capacitors. This configuration results in the proper impedance match for efficient signal transfer. The sound "take-off" is from the collector of the third video IF.

Sound Circuits

The sound section consists of the sound takeoff, sound IF, oscillator limiter, ratio detector, audio driver and audio output stages.



Oscillator Limiter

The oscillator limiter stage is self-oscillating, with the frequency determined by the incoming signal from the sound IF stage. This incoming sound signal serves as a "sync" signal for the oscillator limiter. Because of the dependency of the output of the oscillator on the incoming signal, the output of the oscillator does not change in amplitude and is, therefore, a very effective limiter stage. The output of the oscillator limiter is then applied to the ratio detector.

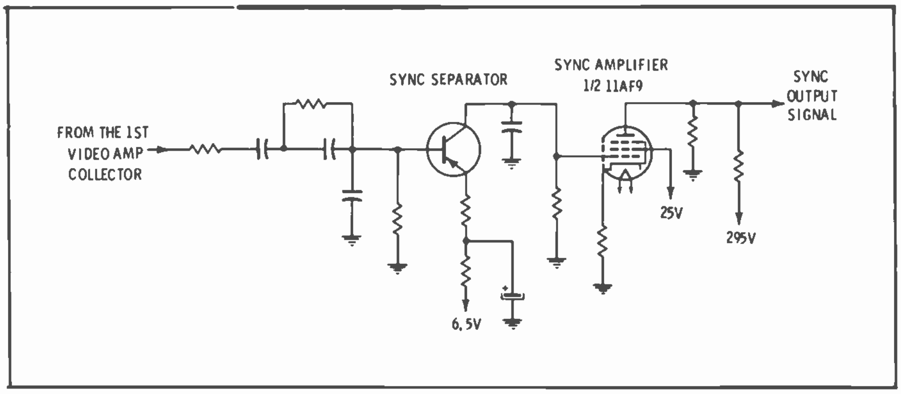


Fig. 9 Transistor and vacuum tube are mated in sync separator/amplifier section of K10.

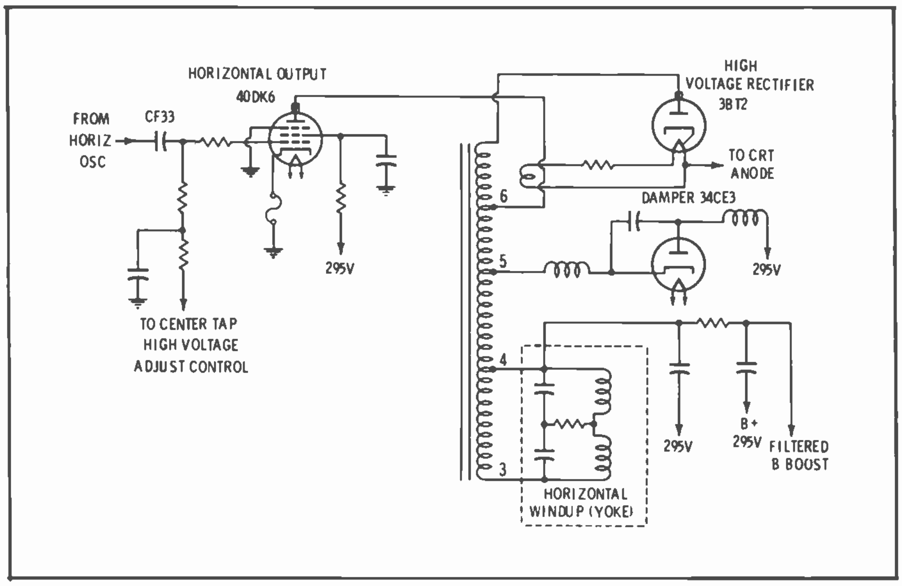


Fig. 10 Fuse in cathode circuit of horizontal output stage provides overload protection for horizontal output and flyback.

AF Sound Circuitry

The AF sound circuitry has two direct-coupled transistors driving an output transformer. A voltage dependent resistor (VDR) is connected across the primary of the audio output transformer to prevent damage to the receiver if it is operated with the speaker disconnected.

The output transistor is mounted on the end of the chassis, utilizing the chassis as a heat sink. A mica wafer isolates the collector (transistor case) from the chassis, resulting in a potential of 105 volts DC on the transistor case.

Sync and Sweep

A transistor is used as a sync separator with its output driving a vacuum-tube (11AF9) sync amplifier (Fig. 9). The sync output voltages are taken from the plate and applied to the vertical and horizontal circuits.

The horizontal phase detector and horizontal oscillator are similar to those used in previous Admiral models.

The horizontal output stage (Fig. 10) operates as a controlled switch. It conducts during half of the scan line, and is cut off the other half. When drive pulses are applied to the grid coupling capacitor (CF33), grid rectification produces an extremely

Fig. 8 Three-stage transistorized video IF employed in Admiral's K10 hybrid chassis.

high negative grid voltage.

When the output tube conducts, heavy current flows through the flyback windings. At the end of the line scan, the output tube ceases conduction, and the flyback field collapses, providing the flyback pulse. This pulse is stepped up by the high turns ratio of the high-voltage transformer windings. It is then rectified by the high-voltage regulator which provides the high CRT anode potential. As the collapsing flyback field shifts into its negative half cycle, the damper tube becomes forward biased, causing a strong damper current to flow. The yoke current will then increase and conduct in the same direction as the output tube. The damper causes the beam to sweep the remaining half line.

Simultaneously, the boost capacitor is charging up to boost level. The dampered flyback field will produce an output that is added to the

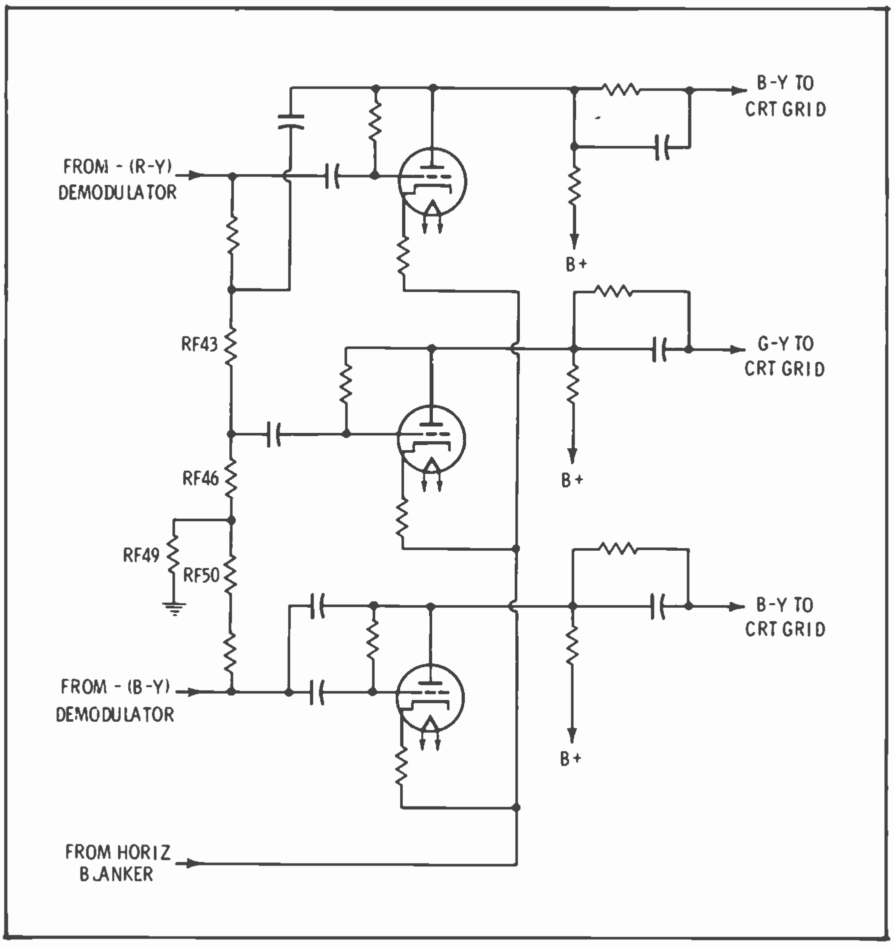


Fig. 11 Triple-triode 8AC10 performs color-difference amplification function in Admiral's hybrid chassis.

B+ potential. This boost voltage is filtered and used to supply the CRT screen voltage, vertical oscillator plate and vertical output bias requirements.

The horizontal output circuit is protected by a fuse in the cathode of the horizontal output tube. This circuit protection, which hasn't been used in a few years, saves the flyback. The HV regulation is provided by a "pulse feedback" circuit.

Chroma Oscillator

The reactance control stage provides (when necessary) a correction voltage to the 3.58-MHz subcarrier oscillator (Fig. 12). An N-channel junction field-effect transistor (FET) performs this function. Because FET's are susceptible to damage from static discharges and arcs from the CRT, care must be exercised

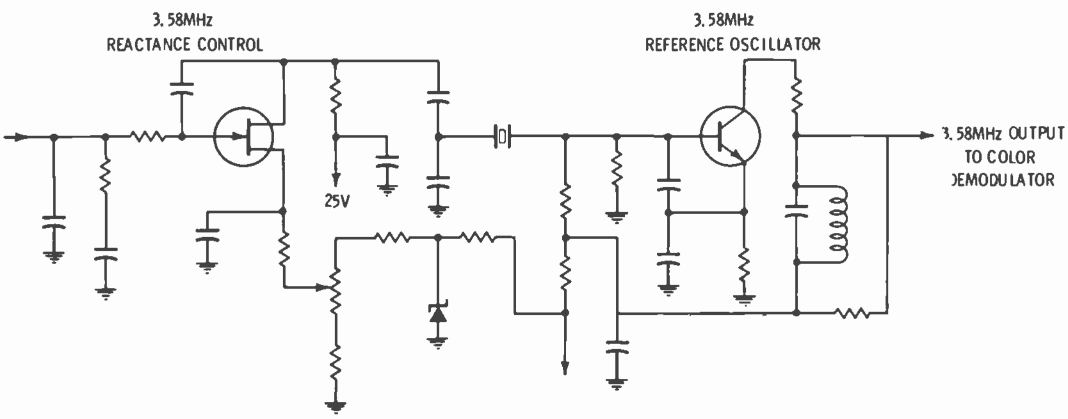


Fig. 12 Field-effect and conventional NPN transistors serve in the reactance control and chroma reference oscillator section of the K10's chroma circuitry.

when working in this area. One method of testing is to check for the presence of voltage on the gate of the FET when the 3.58-MHz oscillator is slightly off-frequency. Check operating potentials on the gate, drain and source, and compare the readings to the voltage values shown on the schematic.

As with most control circuits, the reactance control does not operate when the subcarrier oscillator remains on frequency. Varying the feedback correction voltage from the color phase detector or changing the setting of the reactance control potentiometer will initiate conduction in the FET. This changes the tuning, or phase, of the 3.58-MHz subcarrier oscillator.

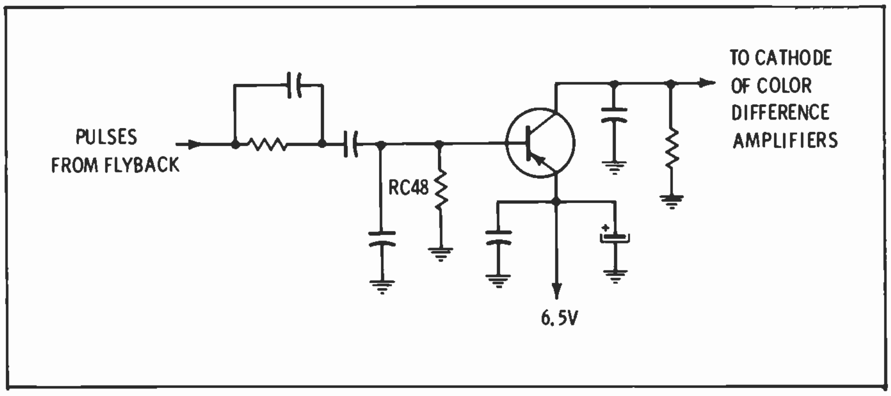


Fig. 13 Cutoff of transistorized horizontal blanker during horizontal retrace drives cathodes of color difference amplifiers more negative; difference amplifiers become saturated and their plate voltage decreases, biasing off CRT grids.

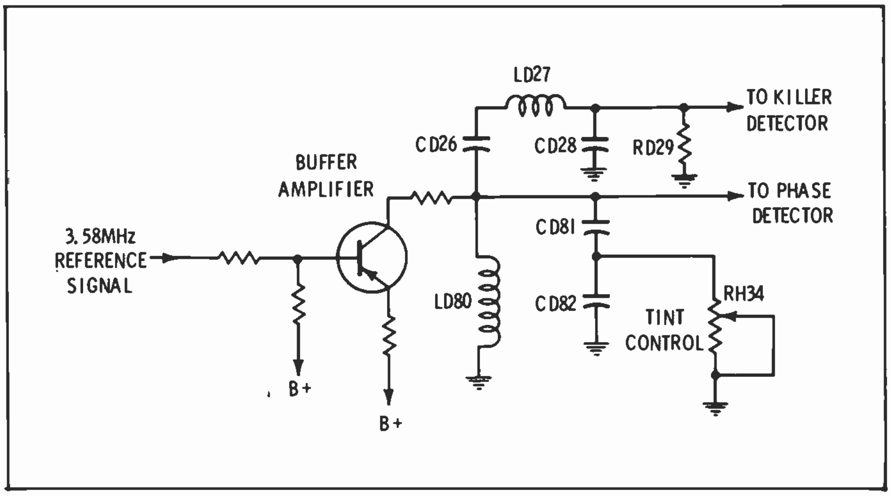


Fig. 14 Transistor buffer amplifier supplies 3.58-MHz reference signal to color-killer and phase detector diodes.

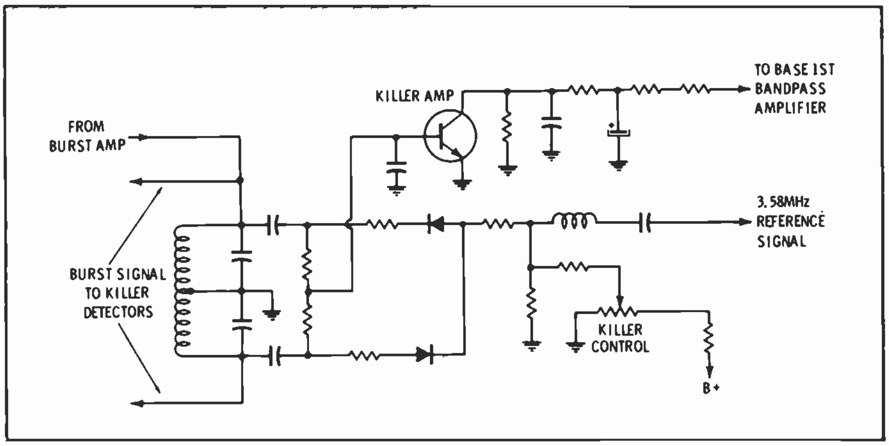


Fig. 15 Color-killer and associated detector circuitry.

Color Demodulation

Color detection is accomplished using —(R-Y) and —(B-Y) signals. The output of the demodulators is then inverted and amplified by the color difference amplifiers, and becomes R-Y and B-Y.

The demodulators are essentially electronic switches with the 3.58-MHz subcarrier signal directing their operation. The amount of conduction is determined by the amplitude of the chroma signal delivered to them by the bandpass amplifier.

The color demodulators used in the K10 chassis are PNP transistors. The chroma signal from the bandpass amplifier is applied to the emitters of the demodulators through a resistive voltage divider network. The switching signal (3.58 MHz) is fed to the base of each demodulator. Under these conditions, the negative-going peaks of the 3.58-MHz signal will cause conduction, and an amplified color difference signal will be present on the collectors of the demodulators.

To control the time of demodulator conduction, the 3.58-MHz signal is coupled directly to the —(R-Y) demodulator and through a phase shift network (90°) to the —(B-Y) demodulator. The 3.58-MHz signal present on the collectors of the demodulators is trapped out by a capacitance/inductance network.

Color Difference Amplifier (Fig. 11)

The color difference amplifiers are housed in a three-section triode tube, type 8AC10. The outputs of the color demodulators are fed through the difference amplifiers, and the signals are amplified, inverted and applied to the CRT red, blue and green grids. The —(R-Y) and —(B-Y) signals come directly from the demodulators, and the —(G-Y) signal is formed by a matrix circuit comprised of resistors RF43, RF46, RF49 and RF50.

Horizontal Blanker (Fig. 13)

To accomplish horizontal blanking, each color difference amplifier is biased off during horizontal retrace time. A negative blanking pulse is applied to each color-difference amplifier cathode, causing heavy conduction through the triodes, producing a high-amplitude pulse on the CRT grids. The color difference amplifier is driven into saturation, the plate voltage is decreased sharply, and the CRT is driven into cutoff. This action results in some signal detection, which provides a measure of DC restoration to the CRT.

The horizontal blanker stage cuts off the CRT during horizontal retrace time by driving the color difference amplifier into saturation. The PNP transistor acts as a switch at the horizontal rate.

During the horizontal scan time, the horizontal blanker conducts strongly, due to the biasing action of the base resistor, RC48. During the retrace interval, a positive pulse from the flyback transformer drives the base-emitter junction into a reverse-bias condition, and the transistor becomes an open switch. The collector voltage decreases to zero.

When the cathode bias voltage of the color difference amplifier is removed (horizontal blanker collector at zero volts), the color difference amplifiers are driven into saturation. The plate voltage decreases sharply, cutting off the CRT grids.

3.58-MHz Reference Feedback System (Fig. 14)

A buffer amplifier (3.58-MHz reference feedback amplifier) supplies a 3.58-MHz reference signal for the color-killer detector diodes and the phase detector diodes. The color demodulator diodes control the phase of the redeveloped chroma subcarrier.

The 3.58-MHz signal is coupled to the base of the buffer amplifier. The tint control, RH34, and capacitors CD81 and CD82, connected across inductance LD80, are connected in the collector circuitry and alter the phase of the output signal, controlling the color demodulation when the tint control is varied. Inductor LD27, capacitors CD26, CD28 and resistor RD29 restore the phase of the reference signal applied to the killer detector.

Color Killer System (Fig. 15)

The color-killer amplifier cuts off the first bandpass amplifier when no color information is present in the received signal. When no color information is present, the color-killer amplifier conducts heavily, driving the first bandpass amplifier to cutoff. This prevents color noise from contaminating the b-w picture.

When color information is present in the signal, the detected burst signal is applied to the base of tthe killer amplifier. When the killer amplifier is cut off, no killer bias is applied to the first bandpass amplifier and it is free to amplify the chroma signal.

When color information is not present, a positive voltage, derived from the color-killer control (connected to the 25-volt line), supplies forward bias to the killer amplifier through the color-killer detector diodes.

When color information is present, the color signal is detected and a negative voltage is developed to cancel the forward-bias voltage ▲

RCA's CTC40 Color Chassis

RF Amplifier (Fig. 16 & 17)

A dual-gate MOS field-effect transistor is used in this stage. Operation of this device (shown in Fig. 1) is as follows: A dual-gate FET functions in much the same manner as two vacuum tubes connected in a cascade configuration. Internal feedback capacitance, which tends to cause amplifier instability (oscillations), is defeated by the relatively low input impedance of the driven portion of the FET. Because of this characteristic of the FET, neutralization is not necessary at VHF frequencies. RF gain control is accomplished by reverse biasing of gate G2. This reverse gain-control voltage reduces the amplifier gain by reducing the drain current of both sections of the FET. Since this control voltage (reverse bias) is derived in combination with the chassis circuitry, it actually forms the RF AGC.

Circuit Operation (Fig. 17)

Signal from a 300-ohm antenna input is applied to an impedance-matching circuit on the cabinet back to match the 75-ohm input impedance of the receiver. This signal is fed to gate G1 of the FET through the high-pass filter network and the RF tuned circuitry. The RF AGC voltage is applied to gate G2 of the FET. This voltage may range from —5.0 volts on a very strong signal to a 6.7 volts on an extremely weak signal. The bias for gate G1 is comprised of a portion of the AGC voltage applied through resistor R1 and the voltage developed across the source resistor, R2. In this circuit the source and gate G1 voltages will coincide or vary with applied AGC, minimizing input capacitance variations. Printed inductances are utilized to couple the RF amplifier output signal to the mixer input circuit.

Mixer (Fig. 18)

The mixer stage is also connected in a cascade configuration, using two transistors. Transistor A operates in a common-emitter circuit and its output drives a common-base amplifier. As with the FET RF amplifier, the principal advantage to this type of circuit is its inherent stability.

The base bias network for transistor A is comprised of resistors R1, R2, R3 and R4. This bias voltage is maintained at a value designed to allow the most efficient mixing action. Transistor B base bias is derived from a biasing network composed of resistors R2, R3 and R4.

The output, or resultant IF signal, is coupled from the mixer by a circuit configuration called "low-side C." The IF output signal is developed across capacitor C1. This capacitor is connected from coil L1 of the tuned output circuit (L1 and C2) to ground. From this circuit configuration comes the term low-side C. This coupling arrangement tends to minimize the amount of oscillator energy that might be coupled into the IF circuit. At oscillator frequencies coil L1 acts as an RF choke and greatly reduces the amount of oscillator energy developed across C1.

Oscillator (Fig. 19)

The local oscillator is basically a

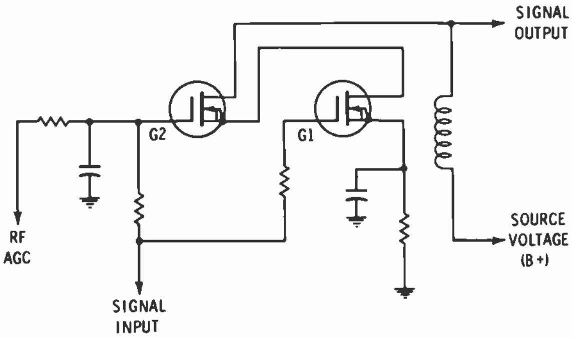


Fig. 16 Circuit diagram illustrating the basic operation of the dual-gate MOS FET employed in RF amplifier.

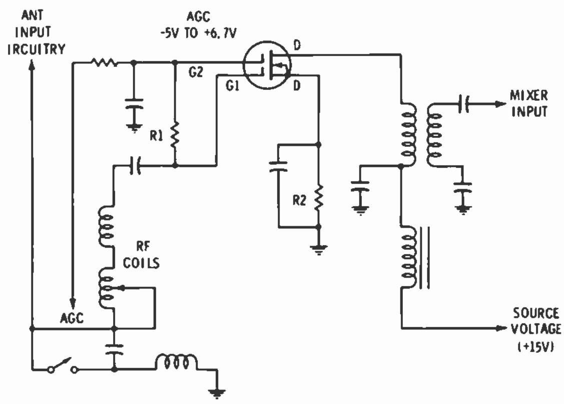


Fig. 17 Simplified schematic of RF amplifier employed in CTC40 chassis.

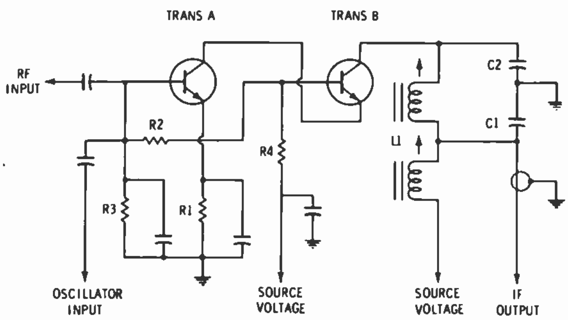


Fig. 18 Mixer stage employing two transistors in cascade arrangement.

Colpitts-type arrangement. Energy at the selected frequency of the oscillators is developed across the tank circuit. The inductance of the tank is composed of L1 (channel 13 adjustment) and the oscillator coil, which changes for each channel selected. The capacitance of the tank is supplied primarily by capacitor C1 and the AFT transistor. Oscillation is sustained by a capacitive voltage divider made up of the transistor internal capacitance existing between the emitter and collector (CCE) and the emitter and base (CEB). Capacitor C2 couples the oscillator transistor output to the tank circuit.

Automatic fine tuning is provided by transistor Q2. The internal capacitance of this transistor varies in proportion to the AFT control voltage applied to its collector and base terminals. This transistor then controls a portion of the oscillator tuning capacitance and, thus, the frequency output of the oscillator.

The KRK-132 UHF Tuner

The UHF tuner used in the CTC 40 chassis is a KRK-132 and has been used previously with several other RCA chassis. It contains no physical or electrical changes from those originally used.

Video IF Circuits

The CTC40 IF section contains three common-emitter amplifier stages (see Fig. 20) capable of supplying a maximum of 80 dB gain to frequencies within the limits of the IF bandpass. This IF bandpass is established through the proper tuning of eight tuned circuits located within the IF system. Alignment of these circuits is very similar to the alignment process of similar RCA circuits employed in tube-type receiver chassis.

Gain in the first and second IF stages is under AGC control and it is possible to reduce the overall gain of the IF section up to 70 dB under very strong signal conditions.

Link Coupling (Fig. 21)

The IF signal contained in the output of the mixer stage is link-coupled to the first IF amplifier. The link coupling circuit used in the CTC40 chassis is very similar to other recent RCA color chassis. The circuit is basically a double-tuned, or overcoupled, network consisting of the mixer collector coil, coupling capacitor, the first IF base transformer and two trap circuits with their associated capacitors. The link coupling components are essential in obtaining a good IF response curve. It is necessary that they be aligned and adjusted exactly. The correct curve is shown in Fig. 22. The manufacturer's specifications or other accurate service data should be consulted for the correct procedures to obtain this response.

First and Second Video IF Amplifiers

The first and second video IF amplifiers are identical. Both are common-emitter types employing identical input and output circuits.

The input coupling circuit to the base of each amplifier consists of a series resistance/capacitance combination. This coupling circuit provides DC blocking and highly efficient impedance transfer. Their output signals are developed identically across single-tuned circuits.

Bias for the first and second IF amplifiers is obtained as follows: The second IF amplifier receives base bias voltage from the output of the AGC inverter stage. This identical voltage (less the small drop across the second IF amplifier base-emitter junction) is applied to the first IF amplifier. In this manner both the first and second IF stages receive almost identical AGC control voltages. The emitters of both the first and second IF amplifiers are returned to ground through a 450-ohm, 7-watt wirewound resistor. The positive temperature coefficient properties of this resistor function to vary the bias on both transistors and compensate for long-term temperature-related gain changes. The emitters of both stages are bypassed by a capacitor.

Third Video IF Amplifier (Fig. 23)

The third video IF amplifier is a common-emitter configuration whose output is applied to the video detector, AFT, and sound circuits. The base bias is derived from a voltage network comprised of resistors R1 and R2. The source voltage for this network is taken from the collector resistor (R3) of the third IF amplifier transistor. DC negative feedback is obtained from this arrangement, which improves bias stability. Emitter bias is obtained, or determined, by resistor R4. This resistance is AC bypassed by capacitor C1. Due to the high gain (40 dB) of the third IF stage, neutralization is a necessity, and it is provided by feedback capacitor C2.

The overall frequency response, which is determined by the efficiency of the link circuit, the interstage tuned circuits and the third IF amplifier output, is illustrated in Fig. 23. Again, please refer to the manufacturer's data sheet or PHOTOFACT for information relating to alignment procedures.

Video Detector

The video detector circuit (Fig. 24) does not differ greatly from the circuits previously used in tube-type chassis. Harmonics of both the detected carriers and difference frequency signals are bypassed by C1. The 4.5-MHz difference frequency developed by mixing the picture and sound carriers is removed by a 4.5-MHz bridged-T trap composed of coil L1, capacitor C2 and resistor R1. L2 and L3 decrease harmonics developed by detector functions. The DC component of the detected video signal is retained by using the average DC level produced by the detector as the major portion of the first video amplifier base bias. The detected video signal is applied across the detector load impedance comprised of resistor R2 and peaking coil L4. This voltage is series-added to the comparatively small second detector pre-bias voltage derived from the 15-volt supply by a resistive divider network made up of resistors R3, R4 and R5. This voltage sets the initial bias level for the first video amplifier, and also provides a constant emitter bias for the gated AGC amplifier.

AGC

The purpose of any type of AGC is to maintain a constant video detector output level over a wide range of input RF signal levels. The changes or variations in video signal amplitudes are translated into DC voltages which are used to control the gain of the RF amplifier and the first and second IF amplifiers. in the CTC-40 chassis.

Video information from the first video amplifier is fed to the gated AGC amplifier to produce an output DC voltage that is proportional to the sync tip amplitude. This AGC output is filtered and applied simultaneously to the RF AGC clamp circuitry and the IF AGC inverter.

The RF AGC clamp circuit sets the requirements of the AGC voltage variations that can be applied to the RF amplifier. The RF amplifier operates under reverse AGC control. A more negative voltage (reverse bias) results in gain reduction, and a more positive voltage (less reverse bias) results in increased gain.

An AGC inverter stage is required

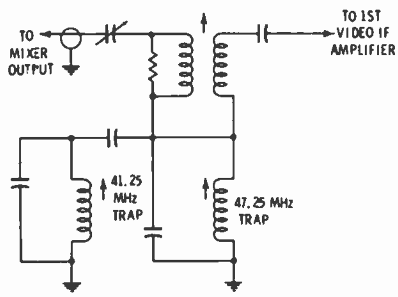


Fig. 21 Doubletuned link coupling is used between mixer output and video IF input.

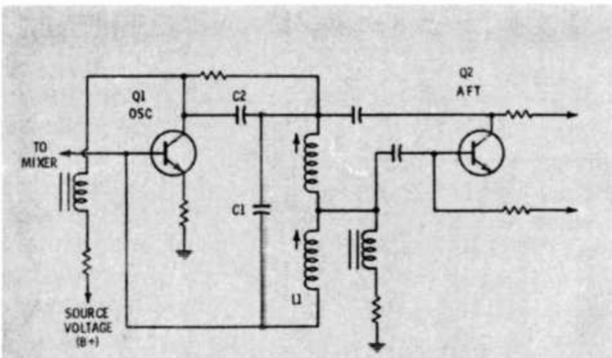


Fig. 19 Local oscillator and AFT circuitry employed in KRK-142 VHF tuner.

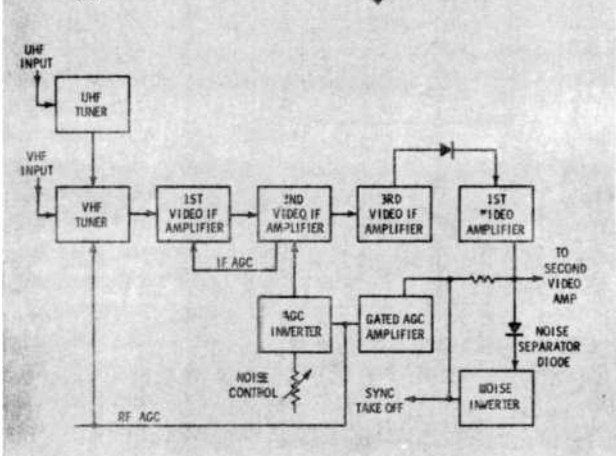


Fig. 20 Block diagram of tuner and video IF circuits employed in CTC-40 chassis.

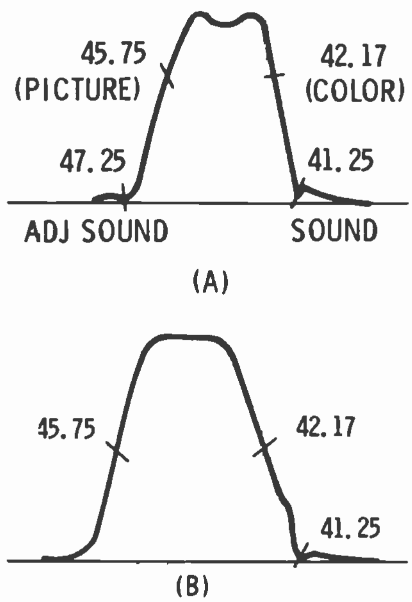


Fig. 22 Correct response curves of (A) link coupling circuit and (B) overall video IF circuit.

for the first and second video IF stages to satisfy their requirements for forward AGC control voltage. The AGC inverter base bias is made variable by the noise control, to establish the proper proportions of AGC voltage applied to the RF and IF stages.

Gated AGC Circuit Operation (Fig. 25)

A video signal which contains positive-going sync pulses is fed to the gated AGC amplifier base. (This signal level is proportional to the picture carrier strength.) The gated AGC amplifier is designed to conduct only during sync pulse time by the positive-going keying pulses coupled to the collector through capacitor C1. These pulses occur at the horizontal frequency rate and key the transistor simultaneously with the horizontal sync pulses contained in the video signal applied to the base. The bias on the transistor is such that the base-emitter junction can become forward biased only during the positive peaks of the sync pulses. This circuit keeps spurious noise to a minimum.

During conduction time, electron current flow is from the emitter to the collector, leaving a negative charge on capacitor C1. This negative charge becomes the AGC voltage and its value is directly proportional to the amount of amplifier conduction; and the amount of amplifier conduction is directly proportional to the peak positive amplitude of incoming sync pulses. The RC network, composed of R1 and C2, improves the overall stability of the circuit.

Service Switch

It is necessary to provide a blank raster to aid in picture tube setup. A positive voltage is applied to the AGC amplifier by operation of the service switch. When the service switch is actuated, a 30-volt potential (normally dropped across resistor R2) forward biases diode X1 and appears on the base of the AGC amplifier. This potential is sufficient to saturate the AGC amplifier and, consequently, produces a high negative AGC voltage. This voltage, in turn cuts off the RF amplifier, and the first and second video IF amplifiers. All video information is removed from the CRT, and a blank raster results. Diode X2, in the collector circuit of the AGC amplifier, prevents the developed negative AGC voltage from discharging back through the collector-base junction between keying pulses.

AGC Inverter (Fig. 26)

The first and second video IF stages require forward AGC control voltages; therefore, it is necessary to invert the AGC output before application to the IF circuits. This is the function of the AGC inverter stage illustrated schematically in Fig. 11. It is a common-emitter DC amplifier designed for a gain of approximately 0.15. Fractional gain is necessary to reduce the large voltage range of the AGC amplifier to within the bias base control limits of the IF amplifiers. The AGC inverter base bias voltage can be varied by the noise control. The noise control is used to set the proper proportions of AGC voltages applied to the RF and IF amplifiers throughout the AGC control range. The control is used basically to establish the point at which the AGC voltage starts reduction of RF amplifier gain, or, if you prefer, sets the RF AGC delay point. Changing the bias on the AGC inverter stage by changing the noise control setting varies the bias and the gain of the first and second video IF amplifiers. The gain of these two amplifiers sets the video signal level applied to the gated AGC amplifier, whose output determines the RF

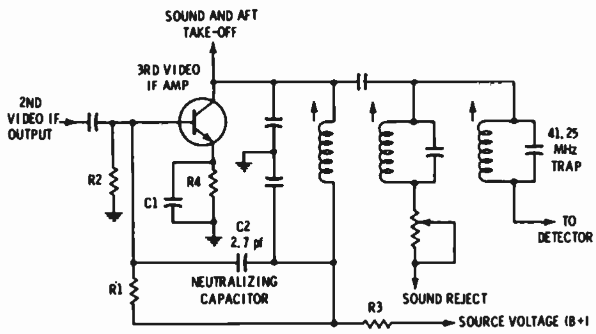


Fig. 23 Third video IF amplifier employed in CTC40 chassis.

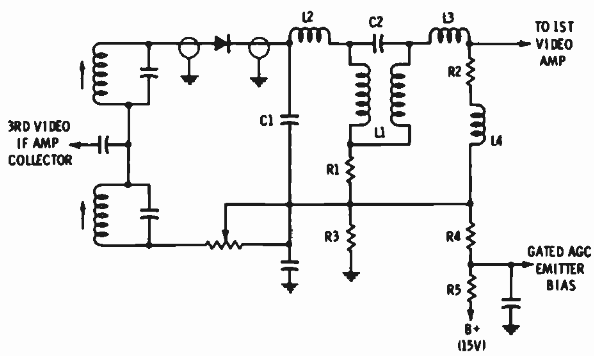


Fig. 24 Video detector circuit of CTC40 is similar to the design employed in previous tube-type chassis.

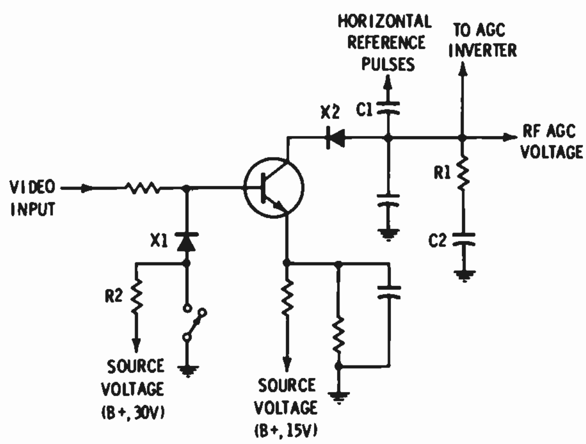


Fig. 25 Gated AGC system is keyed on by positive-going sync pulses applied to base and collector.

AGC voltage. The noise control should be adjusted while observing a noise-free signal and rotating the noise control in the opposite direction until the snow is gone.

Noise Inverter (Fig. 27)

The purpose of a noise inverter circuit is to prevent any spurious noise pulses that might be present in the video signal from interferring with the smooth operation of the AGC amplifier or upsetting the sync separator. The noise inverter minimizes the effects of any noise pulses by inverting and, thereby, cancelling the pulses before they are applied to the AGC and sync circuits.

The circuit operation of the noise inverter is as follows: A reverse bias is fed to the cathode of X1 from the constant source potential available at the emitter of the gated AGC amplifier. This potential sets the conduction threshold for the diode. Diode X1 conducts only during the interval of positive-going pulses. These pulses forward bias the diode and are applied to the base of the noise inverter through diode X1 and capacitor C1. The noise inverter transistor does NOT have a DC forward bias; therefore, conduction will occur only when an incoming positive pulse exceeds the base-emitter barrier junction potential of 0.6 volt. The positive-going sync pulses fed through diode X1 and capacitor C1 are only at approximately 0.2 to 0.3 volt amplitude and are insufficient to cause conduction. Only noise pulses in excess of 0.6 volt will trigger the noise inverter into conduction. When the noise inverter does conduct, the noise pulses that triggered the stage into conduction appear amplified and inverted in the collector circuit and cancel the noise pulses coupled to the collector from the first video IF amplifier. Resistor R1 is connected in series with diode X1 to limit the peak conduction rate. This is required to reduce charging of capacitor C1. If this capacitor is allowed to charge excessively, noise inverter action would be blocked until the capacitor discharged to its initial level.

Sync Separator (Fig. 28)

The sync pulses present at the collector of the noise inverter are applied to the sync separator ampli-

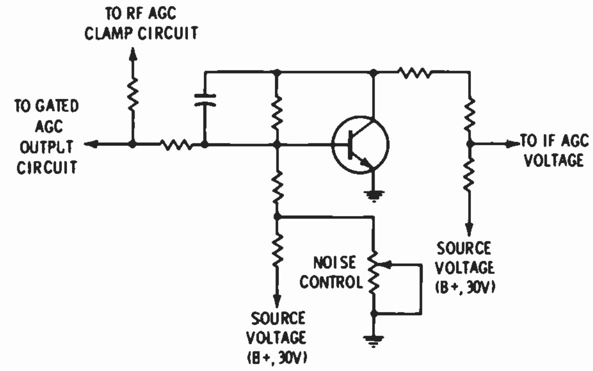


Fig. 26 AGC inverter provides forward AGC control of the 1st and 2nd video IF stages.

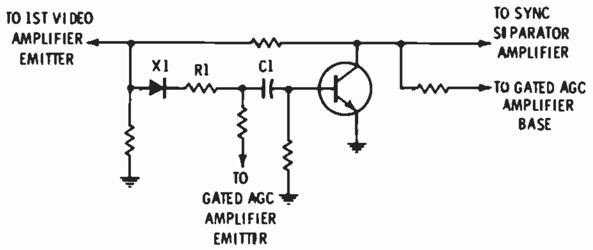


Fig. 27 Noise inverter prevents spurious noise signals from upsetting AGC and sync operation.

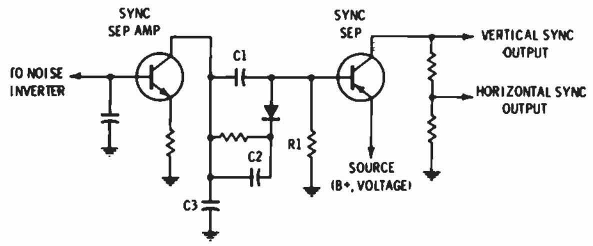


Fig. 28 Two-stage sync separator supplies positive-going sync pulses to horizontal and vertical sync circuits.

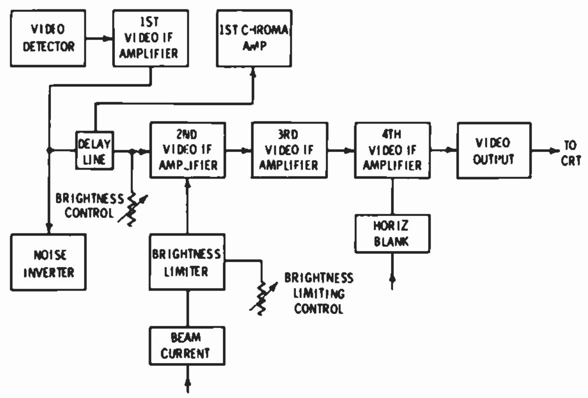


Fig. 29 Block diagram of video amplifier section employed in CTC-40 chassis.

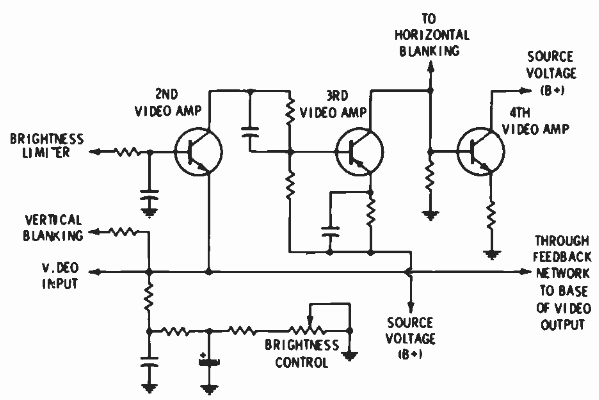


Fig. 30 Simplified schematic of the 2nd, 3rd and 4th video amplifier stages.

fier base. The output of the sync separator amplifier is of the correct polarity and amplitude to drive the sync separator.

The sync separator is a PNP common-emitter switch which is triggered into conduction by the negative-going sync pulses applied to the base through capacitor C1. Discharge path for coupling capacitor C1 is through resistor R1. The output of the sync separator is made up of positive-going sync pulses which are developed across a voltage divider network and applied to both horizontal and vertical deflection systems. The noise immunity features of the sync separator are enhanced during horizontal sync time through the use of .01-pf capacitor C1, which provides coupling between the stages. Capacitors C2 and C3 provide filtering for high video and chroma components of the incoming signal.

Video Amplifier Section (Fig. 29)

First Video Amplifier

The output of the video detector is effectively in series with the base bias of the first video amplifier. Base bias is developed by a voltage divider network. The first video amplifier is connected in an emitter-follower configuration. This circuit features a high input impedance to match the inherently high output impedance of the detector circuit. Output of the emitter-follower is developed across a 1000-ohm resistor in the emitter circuit. Additional circuit loading results from coupling to the following stages: The output circuit of the first video amplifier is connected to the first chroma circuit, the second video amplifier, the sync separator amplifier, the noise inverter and the AGC gate.

Delay Line

The signal output of the first video amplifier is coupled to the delay line. The delay line must be properly terminated to prevent ringing, faulty color registration, etc. The CTC40 delay line has a characteristic terminal impedance (input and output) of 680 ohms at video frequencies. A 560-ohm resistor, in combination with the first video amplifier output impedance, provides the delay line with the proper 680-ohm input terminal impedance.

The output terminal of the delay line is applied to the second video amplifier through a 680-ohm resistor. The second video amplifier stage is designed to exhibit an AC input impedance of zero ohms. Therefore, the output of the delay line is effectively coupled to AC ground through a resistor to properly satisfy termination impedance requirements.

Second, Third and Fourth Video Amplifiers

The simplified circuit configuration illustrated in Fig. 30 points up the relationship of the second, third and fourth video amplifiers. From a functional standpoint their operation is so similar that a brief discussion of the operation of each will suffice.

Second Video Amplifier

The second video amplifier utilizes a common-base configuration and is designed for an input impedance of zero ohms. This is accomplished through the use of a 10-mfd bypass capacitor. The stage functions as a power amplifier. Any fluctuations in the DC output of the first video amplifier are amplified throughout the range of video frequencies. This stage provides proper impedance matching between the delay line output and the third video amplifier input. Positive-going pulses at the vertical frequency rate are fed to the emitter to provide vertical retrace blanking.

The operating point of the second video amplifier varies with the setting of the brightness control. Any change in the brightness control results in a change of the operating point by changing the forward bias current. The lower the resistance of the brightness control, the greater the forward bias current. The result is a larger average current flow through the second video amplifier load resistance. This current flow is translated by the remaining video amplifiers as a reduction in CRT cathode bias and, consequently, an increase in brightness.

Third Video Amplifier

The third video amplifier employs a PNP transistor in a common-base configuration. The video signal is fed to the base through a 1000-ohm resistor. This resistance provides proper loading for the second video amplifier and impedance matching between the second and third video amplifiers. It also functions to prevent saturation of the third video amplifier in the event the second video amplifier develops a collector-emitter or emitter-ground short circuit. The output signal is developed across an 1800-ohm load resistor and direct coupled to the base of the fourth video amplifier.

Fourth Video Amplifier

The fourth video amplifier is connected as an emitter-follower. The output of the stage is developed across an 1800-ohm resistor and is direct coupled to the base of the video output transistor. Positive bias voltage applied to the collector is decoupled from the chassis 30-volts supply source by a filter network comprised of a 10-ohm resistor and an .01-pf capacitor. This decoupling network prevents feedback loops that could cause low-frequency smear, etc. Horizontal pulses, which occur simultaneously with the horizontal retrace interval, are fed to the base to accomplish horizontal retrace blanking. This circuit operation is as follows: Horizontal pulses originating at the high-voltage transformer are applied through a clamp transistor to an isolation diode. The isolation diode is reverse biased during scan time by a positive DC voltage developed at the collector of the clamp transistor. During this interval the blanking circuit is isolated from the fourth video amplifier to prevent loss of high-frequency components.

The negative-going horizontal pulses, fed to the isolation diode during retrace intervals, overcome the diode reverse bias and permit it to conduct. These negative-going pulses are present at the base of the fourth video amplifier and are of sufficient amplitude to affect cutoff. These pulses are applied to the CRT through the video output stage. This action causes picture tube cutoff, or a dark screen, during horizontal retrace time.

Brightness Limiter

A brightness limiter circuit is employed in the CTC40 chassis to hold the CRT beam current within proper limits. The drive potential of the horizontal deflection system is such that, with a high, non-limited brightness control adjustment, it is very possible to exceed the current capabilities of the CRT.

Brightness limiting action of the CTC40 functions to reduce the forward base bias voltage on the second video amplifier when the preset limit of CRT beam current is attained. The preset limit is 1600 micro-amperes (1.6 ma). Circuit action is as follows: The high-voltage transformer secondary winding is returned to B+ through the brightness limiter control. Therefore, all beam current drawn by the CRT must pass through the brightness limiter control. Connected between the low side of the brightness limiter control and ground is the brightness limiter transistor. The fixed base bias for this stage makes the voltage across it comparatively independent of the current through it, as long as it is conducting. This action is much like that of a zener diode; the zener voltage being determined by the resistive divider network in the limiter base circuit.

The current through the brightness limiter control has two parallel paths: one through the brightness limiter transistor, and the other through the CRT. If the brightness control is adjusted in such a manner that the CRT is cut off, the only path for current flow is through the brightness limiter control and the brightness limiter transistor. When the CRT is cut off, this current will be 1.6 ma, the desired CRT beam current limit. Should the brightness control be adjusted so that the CRT starts drawing current, part of the current will flow through the CRT and the remainder through the limiter circuitry, the total current flow remaining at 1.6 ma.

The constant voltage applied to the emitter of brightness limiter supplies a regulated bias voltage of approximately four volts to the base of the second video amplifier throughout the range of the brightness limiting system.

When the brightness control is set to the point where the CRT draws the total preset current of 1.6 ma, all of the current flowing through the brightness limiter control is beam current. Therefore, there is no current available to sustain conduction of the brightness limiter transistor. This results in a loss of the constant voltage applied to the base of the second video amplifier. If more current is demanded by the CRT, the voltage on the emitter of the brightness limiter transistor decreases, reducing the forward bias voltage on the second video amplifier. This action results in a decrease of average conduction in the second video amplifier, and a decrease in brightness and CRT beam current, holding beam current within the preset 1.6 ma limit.

Video Output (Fig. 31)

The video output circuitry is reminiscent of previously employed tube-type configurations. It consists of a common emitter amplifier whose input is DC coupled to the emitter of the fourth video IF amplifier, and whose output is DC coupled to the CRT.

The contrast control is used to vary the value of the series emitter resistance. The contrast control is AC bypassed by a 30-pf capacitor. This circuit action (varying the AC bypass of the contrast control) effectively controls AC degeneration with the end result of effective gain, or contrast, control.

Further control of the stage is provided by capacitor C1, which functions to reduce high-frequency degeneration and prevent changes in high-frequency response (peaking) at different contrast control settings. Inductance and capacitance components form a 3.58-MHz trap which functions to reduce the effects of interference resulting from the mixing of chroma signals and high-frequency video signals.

Output loading of the video output stage presents a familiar circuit configuration. Identical circuits have been employed in several previous RCA chassis.

Sound Section

An integrated circuit (IC) contains the bulk of the sound section. This IC performs the functions of sound IF amplifier, detector and audio driver. For circuit analysis purposes, the IC can be considered as made up of three parts, each section representing a specific circuit function. The first section functions to amplify the incoming 4.5-MHz IF signal to a useful level. The output of the first section (sound IF amplifier) is applied to the second, or FM detector, section. The FM detector removes the audio portion of the signal, which is then applied to the audio driver, the third section of the IC. The purpose of the audio driver is to raise the amplitude of the audio signal to the level required to drive the audio output stage.

Detailed Circuit Analysis (See Fig. 32)

The 4.5-MHz FM sound signal is generated in the conventional "mixing" method by diode X1. This signal is coupled by IF transformer L1 to the sound IF amplifier section of the IC. The output signal from the sound IF amplifier is applied to the phase shift transformer, T2, and, in turn, to the ratio detector diodes. The output of the ratio detector is the audio signal, which is capacitance-coupled to the volume and tone controls. The audio signal is then capacitance-coupled from the volume control to the audio driver section of the IC.

The audio driver section functions to provide the required current gain to raise the signal to a level sufficient to drive the audio output stage, Q1.

The audio output stage is a common-emitter, class A amplifier. The transistor is protected against high-amplitude voltage spikes by a 275-volt zener diode connected from collector to ground. DC stability is enhanced by a feedback network (R1, R2) connected from the emitter of the audio output stage to the input of the audio driver section. Capacitor C1 provides low-frequency compensation for this feedback network. Resistor R3, located

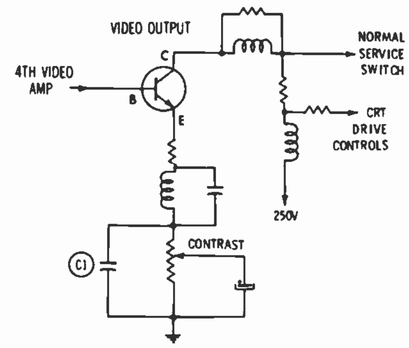


Fig. 31 DC coupling is employed in the input and output of the video output stage.

between the base and emitter of the audio output stage, functions to provide an additional load for the driver section. This minimizes the effects of output transistor leakage current.

Automatic Fine Tuning (AFT)

The fundamental AFT system is illustrated in Fig. 33. This system is basically the same as that previously used in the RCA CTC30 chassis. In this system, an integrated circuit (IC) discriminator/amplifier produces a differential DC voltage that is proportional to the applied IF frequency. This signal is then applied to a special "variable capacitance" transistor in the VHF tuner and a varicap diode in the UHF tuner that produce a correction voltage for application to the local oscillator.

Shown in Fig. 34 is a simplified schematic of the CTC40 AFT circuitry. The IC utilizes an internal, regulated power supply and does not require an external reference voltage for defeating the UHF AFT function. Automatic degeneration of the output amplifiers is such that it eliminates all AFT correction signals when the output terminals are shorted for AFT defeat action.

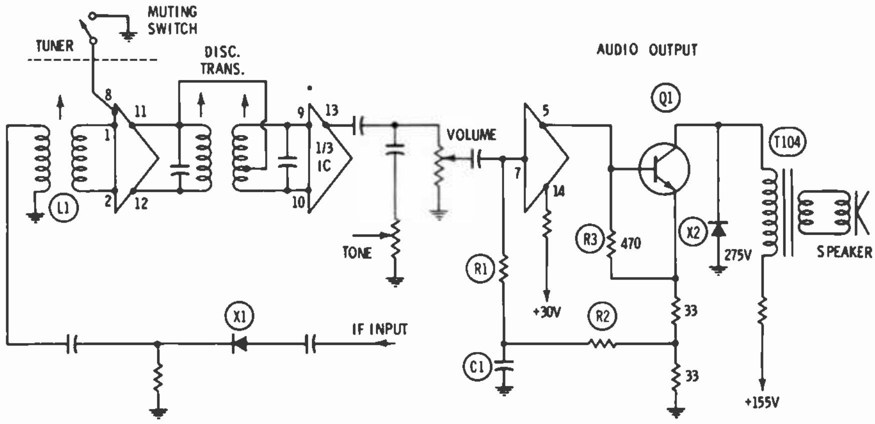
The AFT system is disabled during VHF channel change by the same method used to accomplish AFT defeat during fine tuning—shorting together of the AFT control-voltage outputs. This combination defeat action is initiated by a single switching mechanism built into the plastic housing located on the front of the VHF tuner shaft. The two 15-mfd electrolytic capacitors, C1 and C2, act to prevent undesired correction voltages generated during channel change time from affecting the local oscillator frequency. These capacitors also function to remove residual video information from the AFT output terminals, allowing only the undistorted DC correction voltage to reach the tuner.

AFT Operation

The IC AFT circuit is a type TA5360 that functions as follows:

A sample of the video IF output is applied to the AFT system through a coupling capacitor located in the collector circuit of the vidio IF amplifier. This signal is applied to a tuned input circuit comprised of L1 and C1. Coil L1 and capacitor C1 perform a dual role: They act as both an adjacent channel sound trap and as an IF frequency peaking circuit. Correct trap frequency is obtained automatically by peaking the input tuned circuit (L1, C1) at 46.1 MHz.

The output of the input tuned circuit is applied to the buffer amplifier section of the IC, the output of which appears across the primary windings of the discriminator trans-



• PART OF INTEGRATED CIRCUIT (IC)

Fig. 32 An IC functions as the sound IF amplifier, detector and audio driver in the CTC40 chassis.

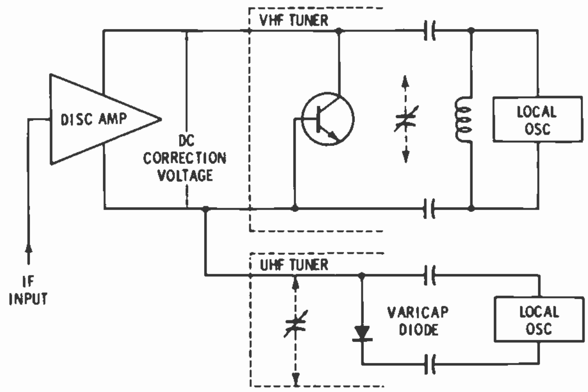
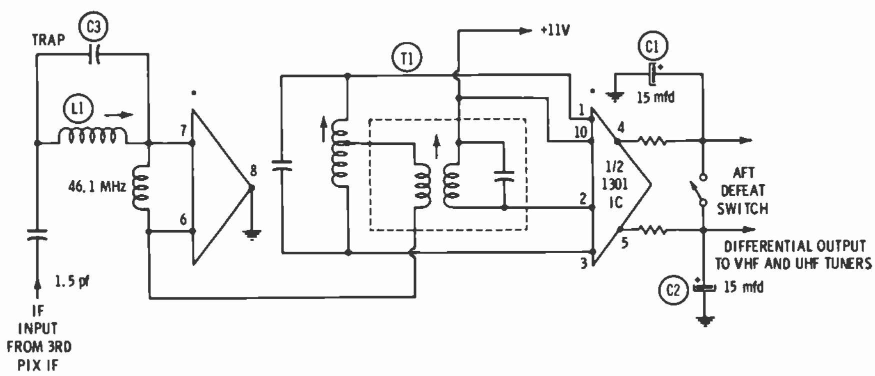


Fig. 33 Simplified illustrations of the CTC40 AFT system.



• PART OF INTEGRATED CIRCUIT (IC)

Fig. 34 Simplified schematic of the AFT circuitry.

former, T1. The discriminator primary is tuned to 46.1 MHz; the secondary winding is peaked at 45.75 MHz.

The discriminator transformer secondary windings feed the IC discriminator diodes. The output voltages of the diodes are applied to an amplifier that delivers a differential voltage output. This differential output contains two voltages, one appearing at each of the IC output terminals. The difference existing between these two voltages (differential) is indicative of the amount and direction the incoming IF signal deviates from the desired 45.75-MHz frequency. If the incoming IF signal is exactly 45.75 MHz, each output signal voltage will be exactly 6.5 volts, and no differential voltage will exist. When the incoming IF signal deviates from 45.75 MHz, one output voltage will increase, and the other will decrease an equal amount. The voltage at each output terminal will increase or decrease, depending on which direction the incoming signal deviates from 45.75 MHz. The maximum differential voltage produced by this circuit is +9 volts, well within the "pull-in" range of the AFT system.

Vertical Sweep

Basic System

The fundamental vertical sweep system employed in the CTC40 chassis is illustrated in Fig. 35. The integrator sweep circuit consists of a high-gain amplification system operating in conjunction with an integrating capacitor. Operation is as follows:

At the start of vertical trace, the integrating capacitor, C1, is charged from a voltage source. This capacitor charge causes the amplifier to supply yoke current, resulting in a voltage being developed across the feedback resistor, R1, which is coupled directly to the integrating capacitor. This feedback action maintains the amplifier input voltage at a constant level, producing a constant rate of voltage "build-up" across the integrating capacitor. The voltage developed across the feedback resistor is directly proportional to the yoke current; therefore, increase of the yoke current is constant, and a linear scan is produced.

The vertical sweep rate is determined by an electronic switch which discharges the integrating capacitor at a 60-Hz rate. Vertical sync pulses are applied to the switching transistor and determine the exact instant the switch is pulsed "on". This action synchronizes the vertical switching action with the transmitted vertical scanning interval. The "linearity clamping" transistor provides the initial charging current to the integrating capacitor.

Vertical Switch (Fig. 36)

The function of the vertical switch is to provide a discharge path for the integrating capacitor at the end of each vertical scan interval. This

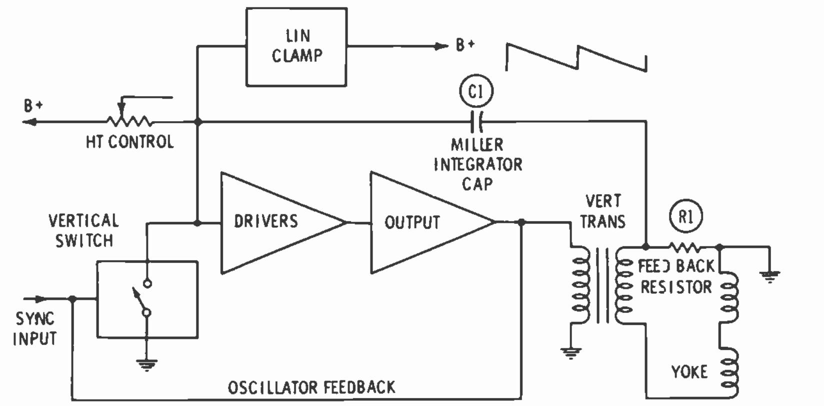


Fig. 35 Illustration of the fundamental sweep system employed in the CTC40.

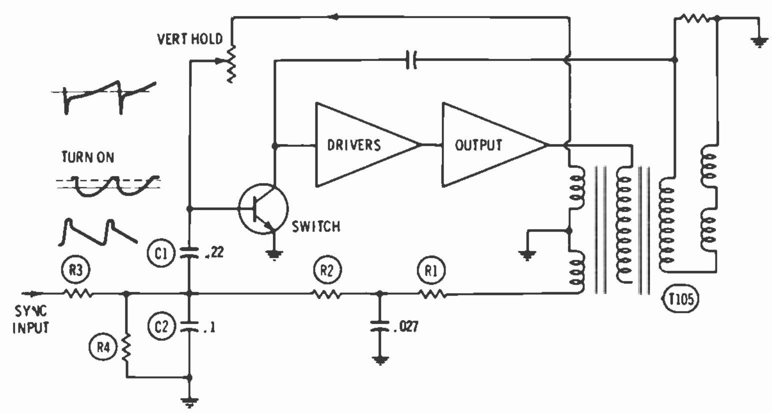


Fig. 36 Partial schematic of the vertical switch and associated circuitry.

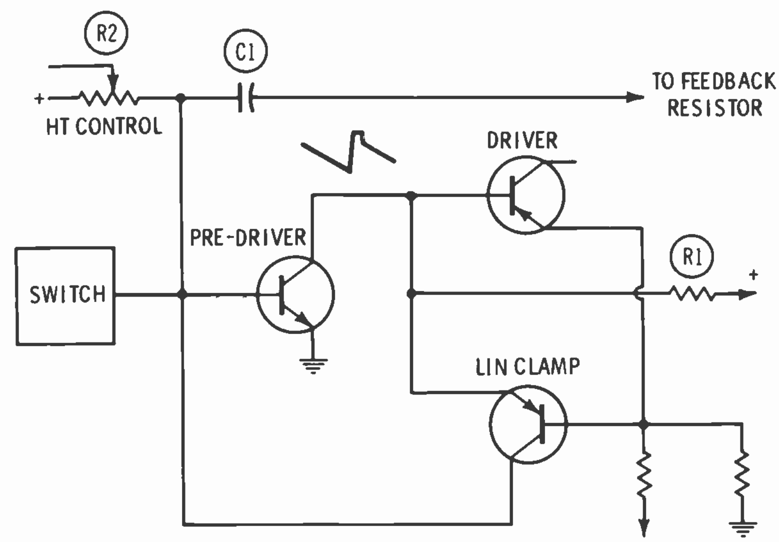


Fig. 37 Schematic showing basic circuitry of linearity clamper stage.

action causes beam retrace and prepares the circuit for the next vertical scan function. Operation of the vertical switch is made self-sustaining by the action of two feedback paths: One path, consisting of resistors R1 and R2 and capacitor C1, is applied to the base and provides the appropriate pulse to initiate "turn on". Vertical sync pulses, from the sync separator, are integrated by resistors R3 and R4 and capacitor C2 and add to the triggering waveshape. An additional feedback voltage is applied to the switch from the vertical output transformer via the vertical hold control. This additional voltage causes the switch base to pass rapidly through the "turn on" voltage potential. As a result, switch "turn-on" is extremely stable and comparatively immune from random noise pulses. The vertical hold control has some control of the "turn-on" point and, therefore, the frequency at which the circuit operates.

Linearity Clamp (Fig. 37)

Since it is necessary to provide a sufficient amount of initial charging current for the integrating capacitor, a special clamping circuit called the "linearity clamp," is utilized. Operation of this circuit is as follows:

The action of the vertical switch discharging capacitor, C1, also cuts off the predriver transistor. This produces a positive voltage on the collector of the predriver. This voltage is of sufficient amplitude to forward bias the linearity clamp transistor. The linearity clamper conducts; current flows through the transistor via R1 and the vertical switch. The vertical switch turns off after approximately 700 microseconds, and the linearity clamp then rapidly charges capacitor C1. As the charge rapidly builds up on capacitor C1, the predriver and driver stages start to conduct, causing the linearity clamp base-emitter junction to become reverse biased due to the voltage drop across the driver base—emitter junction. This circuit action cuts off the linearity clamp and originates vertical scan. Capacitor C1 continues to charge through the height control, R2, for the duration of scan time.

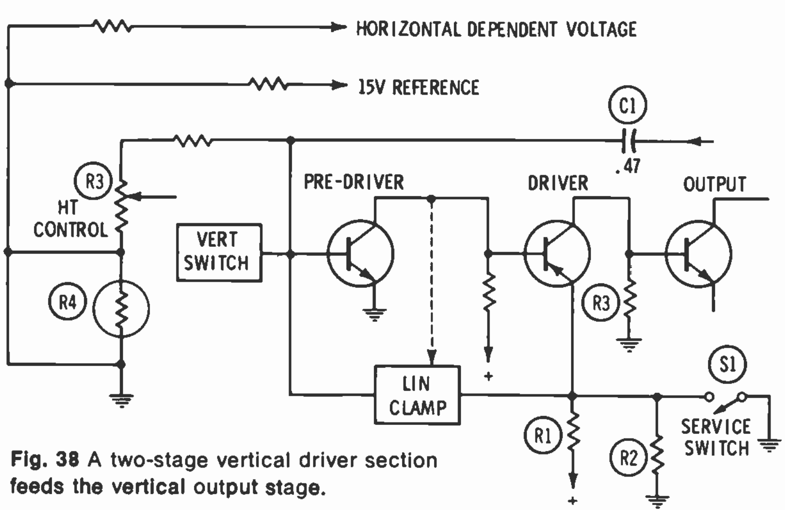


Fig. 38 A two-stage vertical driver section feeds the vertical output stage.

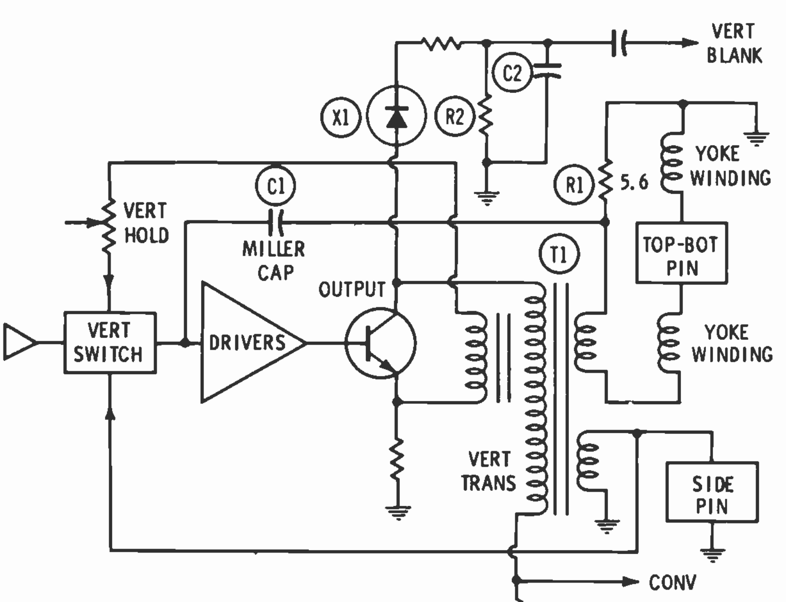


Fig. 39 Common-emitter amplifier is employed in vertical output circuit.

Vertical predriver and driver (Fig. 38)

The vertical driver section is comparatively more familiar. It consists of two stages: a predriver (NPN transistor operating as a common-emitter amplifier) directly coupled to a driver (PNP transistor operating as a common-emitter amplifier). Emitter supply voltage for the driver stage is obtained from a voltage divider network composed of R1 and R2. The driver collector load is comprised of R3 and the base-emitter junction resistance of the vertical output stage.

Provisions for picture tube setup are provided by switch S1, which functions to "short" the driver emitter to ground when actuated. The waveshape of the input signal to the predriver is determined by the charging action of the integrator capacitor, C1, which is charged through the height control, R3. The height control supply voltage is made relatively immune to temperature-induced variables by the action of thermistor R4. A degree of dynamic regulation for the circuit is provided by a signal from the horizontal deflection system. The insertion of this voltage tends to maintain a constant vertical sweep or height, regardless of horizontal scan and high-voltage fluctuations.

Vertical Output (Fig. 39)

The function of any vertical output circuit is to provide the power necessary to fulfill the vertical deflection requirement of the CRT beam. In the RCA CTC40 chassis the vertical output stage is a common-emitter amplifier with an input from the driver stages. Loading for the vertical output stage is provided by the vertical output transformer, T1, and the vertical convergence circuit.

The vertical output transformer is loaded by the vertical windings of the yoke, two feedback networks, and the pincushion correction circuit. Integrating capacitor C1 is connected to the output circuit by resistor R1, a 5.6-ohm feedback resistor in series with the secondary windings of the vertical output transformer and the vertical yoke windings. There are two feedback networks connected to the vertical switch transistor from the vertical output circuit; both of these networks perform waveshaping functions to provide stable, self-sustaining vertical switching. Diode X1, in conjunction with capacitor C2 and resistor R2, provides a protective clamping action for the vertical output transistor. Positive-going retrace voltage pulses cause diode X1 to conduct, effectively clamping the vertical output collector to the voltage existing across capacitor C2. A relatively slow discharge path is required for capacitor C2. This is provided by resistor R2. This discharge action sufficiently reduces the voltage across C2 during retrace time to insure the necessary voltage difference across diode X1 when retrace pulses occur. The pulses that appear across capacitor C2 during conduction of X1 are applied to the 2nd video IF stage to provide vertical retrace blanking.

Pincushion Correction (Top and Bottom) Fig. 40

Top and bottom pincushion correction in the CTC40 chassis is accomplished in a manner similar to methods used in previous RCA color chassis.

A signal voltage derived from the horizontal yoke circuit is coupled to transformer T1. This action energizes a circuit composed of capacitor C1 and coil L1, which is tuned to 15,750 Hz and is in series with the vertical yoke windings, L2 and L3. The resultant sine wave is added to the vertical yoke current waveshape in the proper phase and amplitude to effectively correct top and bottom pincushion distortion. A limited amount of control over the correcting sine wave phase and amplitude is provided by variable inductor, L1, and the damping resistance of R1.

Side Pincushion Correction (Fig. 41)

Side pincushion correction is accomplished by amplitude modulation (at a vertical rate) of the horizontal deflection current. This produces an increase in horizontal scanning width at the center of the raster with respect to the width at the top and bottom. This operation is made possible through the utilization of the saturable reactor circuit illustrated in Fig. 41.

A parabolic waveshape occurring at the vertical frequency is initiated by the action of the control winding of transformer T1, capacitor C1 and resistors R1 and R2. This waveform, coupled to the horizontal yoke

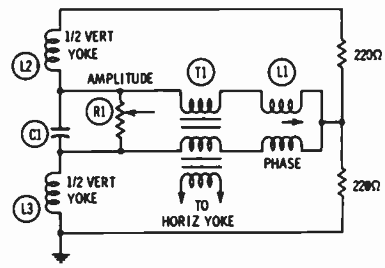


Fig. 40 Top and bottom pincushion correction circuitry.

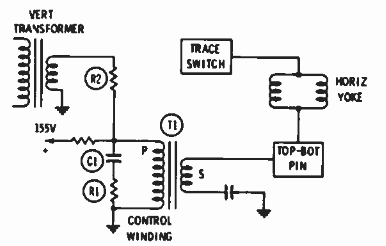


Fig. 41 Side pincushion correction is accomplished by amplitude modulation of the horizontal deflection current.

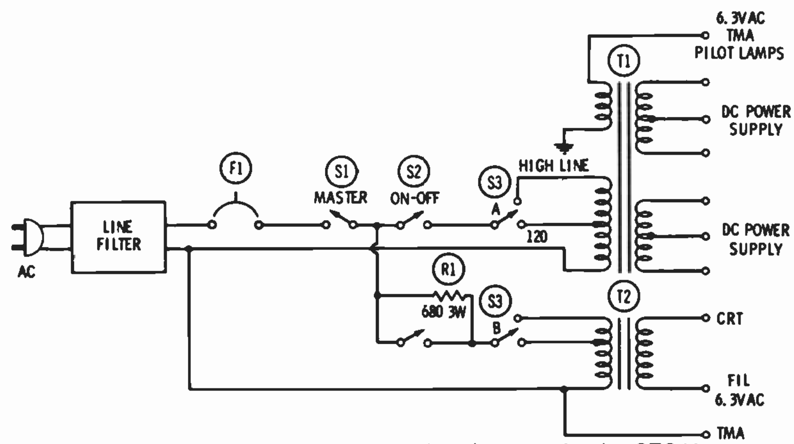


Fig. 42 Schematic of the switching circuitry that permits the CTC40 to take advantage of the "instant-on" characteristics of semiconductors.

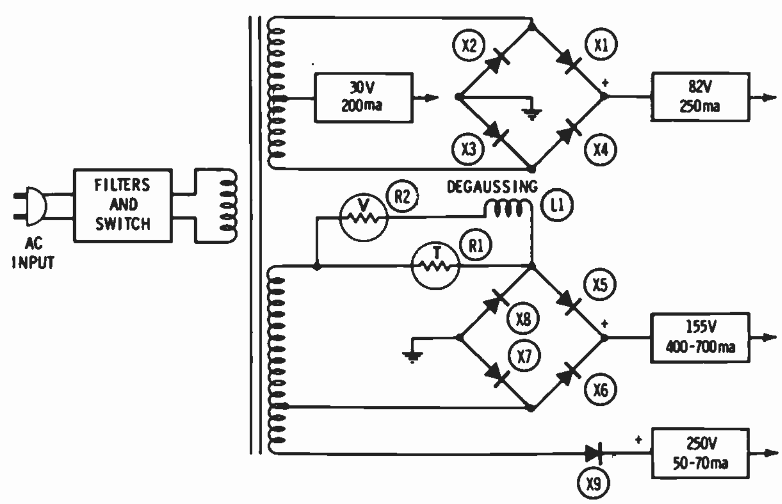


Fig. 43 Three separate rectifier circuits provide the CTC40 with four separate DC sources.

circuit by transformer T1, modulates the amplitude of the horizontal yoke scanning current, producing the proper change in raster width.

Power Supply

The CTC40 power supply provides four DC sources for general circuitry requirements and two AC power sources. The AC power sources are for the CRT filaments and pilot lamps.

Power supply switching circuits allow the CTC40 to take advantage of the "instant on" characteristics of solid-state devices. This switching circuitry is illustrated in Fig. 42.

Switching Circuit

AC power is applied through the line filter and circuit breaker to the master power switch, S1. The master power switch applies power through the "instant pic" switch, S2, to both the DC supply transformer, T1, and the CRT filament transformer, T2. However, when switch S2 is in the "off" position, reduced power is supplied to filament transformer T2 through resistor R1, a 680-ohm, 3 watt component. Using this method, the CRT filament is kept "warm" until full power is applied by closing switch S2. This design insures the full operation of the CTC40 within four to five seconds after turn on.

The master power switch, S1, is a rotary type switch located at the top of the auxiliary consumer-controls bracket. Switch S2 is a push-pull switch located at the top of the consumer-controls panel and is adjacent to the brightness control.

DC Supply Circuitry (Fig. 43)

The DC power supply provides four separate DC sources generated from three separate rectifier circuits. This is illustrated schematically in Fig. 43.

Rectifiers S1 through X4 are responsible for providing both the 82- and 30-volt sources. The 82-volt supply is derived from the full-wave bridge configuration of rectifiers X1 through X4. The transformer secondary winding that feeds this bridge circuit is centertapped and is used to feed two of the four rectifiers comprising the bridge network. This forms a full-wave, centertapped circuit, the output of which is 30 volts.

A second full-wave bridge circuit is comprised of rectifiers X5 through X8. The output of this circuit is the 155-volt source.

The automatic degaussing circuit is coupled to the secondary winding of T1, which feeds rectifiers X5 through X8. This circuit consists of thermistor R1, voltage dependent resistor R2 and degaussing coil L1. Operation of this circuit is the same as that of degaussing circuits previously employed in RCA color chassis.

The 250-volt DC source is obtained from the output of rectifier X9. During normal operation, the CTC40 chassis draws approximately 1.8 amperes of AC current at 120 volts AC input. The average DC current supplied by each leg of the power supply is as follows:

82-volt source—200 ma.
30-volt source—200 ma.
155-volt source—400-700 ma. (varies with beam current)
250-volt source—50-70 ma. (varies with beam current)

1st and 2nd Chroma Amplifiers (Fig. 44)

The chroma signal is applied to the base of the first chroma amplifier through a tuned circuit comprised of capacitors C1, C2 and inductor L1. This circuit is referred to as a "chroma take-off" or "chroma peaker" circuit. Resistors R1 and R2 broaden the bandpass to compensate for loss of chroma sidebands on the response curve of the peaker circuit. The ratio of values of resistors R1 and R2 also provide the proper input impedance for the first chroma amplifier.

Bias for the base of the first chroma amplifier stage is supplied and determined by the ACC amplifier. Output of the 1st chroma amplifier, Q1, is RC coupled to the base of the second chroma amplifier, Q2. The 2nd chroma amplifier stage is a straight-forward common-emitter circuit.

Bias for Q2 is obtained from a voltage divider network consisting of resistors R3 and R4. The output (collector) circuit consists of load inductor L2, capacitors C3 and C4, and the capacitance of the coupling cable. This combination of components forms a broadly tuned circuit with a response that falls within the limits of the desired chroma bandpass.

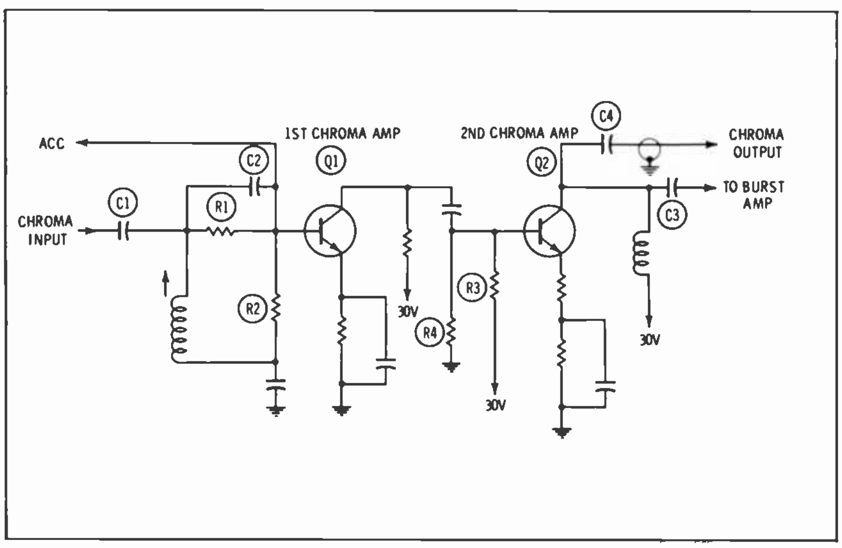


Fig. 44 First stage of two-stage chroma amplifier employed in CTC40 is ACC controlled.

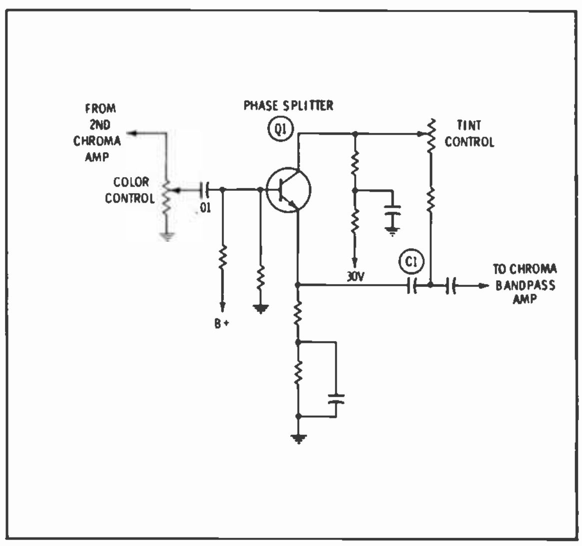


Fig. 45 Tint control in CTC40 is electrically located in collector circuit of phase splitter.

Phase Splitter (Fig. 45)

The phase splitter circuitry provides a means of varying the phase of the chroma signal, or, in effect, provides a method of varying the tint. Chroma signals are applied across the color control and capacitance-coupled to the base of the phase splitter. Output voltages are taken from both the collector and emitter and applied across a phase shifting network comprised of the tint control and capacitor C1.

The phase splitter and its associated circuitry are located on the customer controls bracket. The tint and color controls are consumer controls and extend through the front control panel.

Bandpass Amplifier (Fig. 46)

The output of the phase splitter is capacitance-coupled through C1 to the base of Q1, the bandpass amplifier. Q1 is connected as a common-emitter. Base bias for Q1 is obtained from the killer stage. The output load for Q1 is T1, the bandpass transformer.

A negative-going pulse is applied to the base of Q1 for burst blanking. This is done to prevent the color sync signal from being amplified by the bandpass amplifier.

A double-tuned bandpass transformer determines the exact range of chroma frequencies applied to the demodulators. The secondary winding of T1 is shunted and tuned by capacitor C2. Resistor R1 provides loading and aids in determining the "Q" of the circuit.

Burst Amplifier (Fig. 47)

A simplified version of the burst amplifier stage utilized in the CTC40 chassis is illustrated in Fig. 4. The color sync signal from the collector of the second chroma amplifier is applied to the base of the burst amplifier transistor through capacitor C1. A positive-going keying pulse (15 volts p-p) from the horizontal output transformer is applied to the base across resistor R1. The integrating characteristics of capacitor C1 and resistor R1 provides the required time delay for the keying pulse.

The burst amplifier transistor is keyed into conduction by the pulses from the horizontal output transformer, which arrive at the base of the transistor at the same time as the bursts of color sync signals.

The burst signal is amplified by the burst amplifier and appears across the burst transformer, T1. Loading for the burst transformer is provided by resistor R2.

During conduction of the burst amplifier, resistor R4 establishes the proper emitter operating point. Capacitor C3 functions as an AC bypass capacitor. Resistor R3 provides the required amount of emitter degeneration for proper amplifier stability with maximum voltage gain. Capacitor C2 provides the amount of feedback voltage necessary to cancel the effects of the internal feedback capacitance of the transistor.

The burst amplifier base-emitter bias is maintained below cut-off during scan time, or between burst keying pulses. This assures that only color sync signals are supplied to the AFPC detector. The required "scan-time bias" is developed by the discharge of capacitor C3, the emitter bypass capacitor, through emitter resistor R4. Emitter current flow, resulting from the application of the burst keying pulse, places a positive bias on the emitter. This reverse biases the transistor during scan time. Diode X1 functions to prevent the bias voltage from exceeding the reverse emitter-base breakdown voltage rating.

Automatic Frequency and Phase Control (AFPC) Detector (Fig. 48)

The purpose of the AFPC detector circuit is to develop a DC voltage that is proportional to the frequency and phase difference that exists between the applied color sync signal (burst) and the reference signal supplied by the 3.58-MHz oscillator in the receiver. Rigid control over the operation of the 3.58-MHz oscillator is a prerequisite for proper color demodulation. This is because the output of the 3.58-MHz oscillator is the reference, or standard, on which chroma demodulation is based.

The AFPC detector circuit in the CTC40 chassis is, in effect, a phase-sensitive discriminator. The burst signal is fed at equal amplitude and

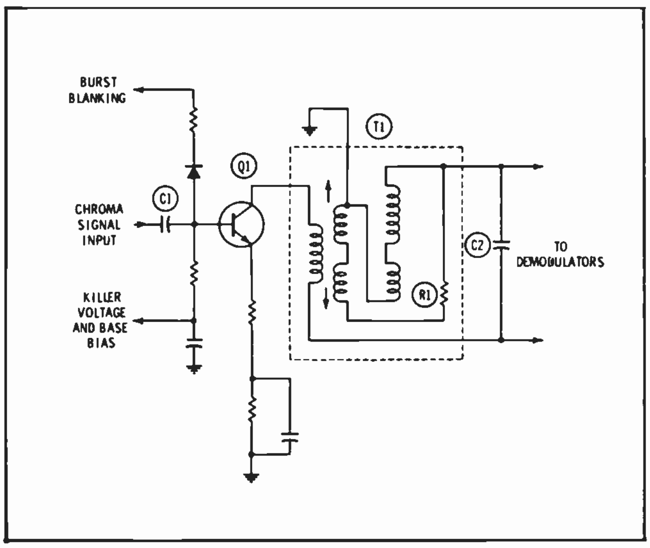


Fig. 46 Three signals or voltages are applied to base of bandpass amplifier: chroma input signal, bias from color killer and negative-going burst signal.

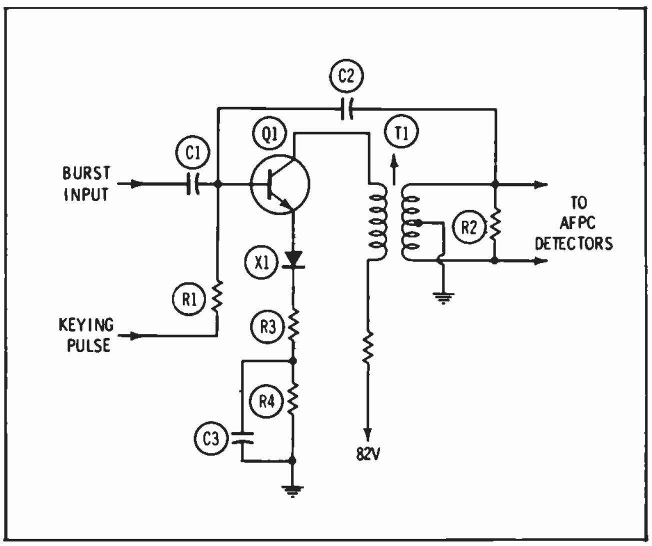


Fig. 47 Simplified schematic diagram of burst amplifier employed in CTC40.

opposite phases through capacitors C1 and C2 to diodes X1 and X2. A sample of the 3.58-MHz reference voltage is applied to the junction of the cathode of X1 and the anode of X2. When the reference voltage and the burst signal are in phase, the diodes will conduct in equal amounts but in opposite directions. The result is zero AFPC voltage.

If the reference voltage lags the incoming burst voltage, diode X2 conducts more than diode X1, causing an imbalance in current flow through resistors R1-R2, and a positive AFPC voltage will be developed. If the reference voltage leads the incoming burst signal, diode X1 conducts more than diode X2, again producing an imbalance in current flow through resistors R1-R2, but in this instance a negative AFPC voltage will be developed.

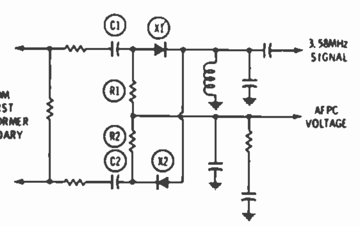


Fig. 48 Automatic frequency and phase control (AFPC) circuit develops DC voltage proportional to frequency and phase differences between color burst and 3.58-MHz reference signal.

3.58-MHz Reference Oscillator (Fig. 49)

The chroma reference oscillator is a modified Clapp-type circuit. Feedback is accomplished by R7, C3, and C2, which couple an inphase signal back to the base.

The operating frequency is determined by the 3.58-MHz crystal and the combined capacitance of capacitors C2, C3 and varactor X1. The varactor utilizes a specially constructed junction that enhances the normal voltage-dependent capacitance characteristics of a diode. The frequency of the oscillator can be varied over a very limited range by changing the voltage impressed across the varactor diode. Thus, the AFPC voltage, and the voltage determined by the divider network (AFPC adjust and R3), will vary the oscillator frequency a small amount. Capacitor C1 serves as a low-impedance ground return for the varactor and has no effect on the oscillator frequency.

The CW amplifier, Q2, operates into a high-Q, single-tuned transformer, T1, that develops a sine wave from the amplifier output current pulses. Capacitors C6 and C7 function as a capacitance voltage divider network that provides the 3.58-MHz reference level to the AFPC detector circuit. The transformer secondary couples the 3.58-MHz signal to the color demodulator stages.

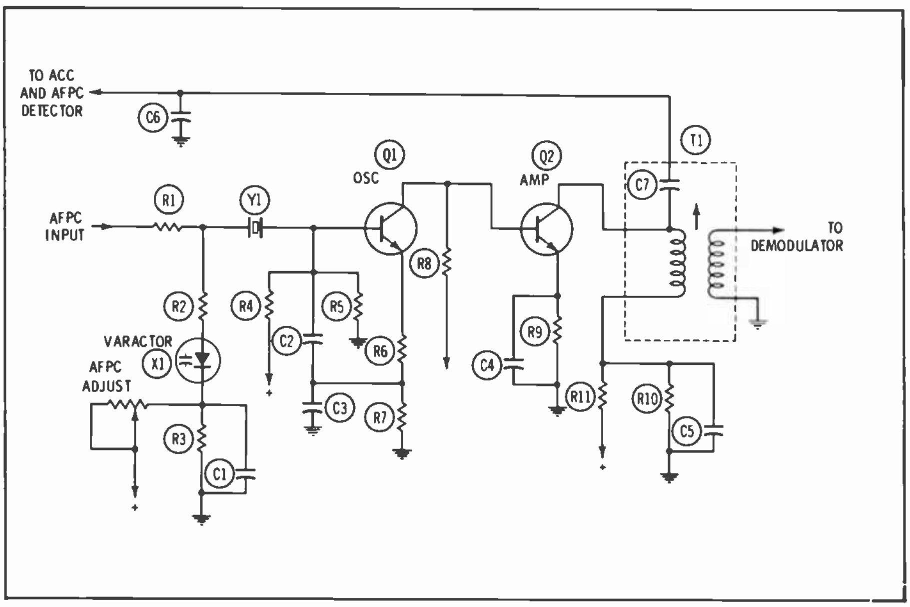


Fig. 49 Modified Clapp-type chroma reference oscillator utilizes regenerative feedback between emitter and base of Q1 via R7, C3 and C2.

Color-Killer Circuits (Fig. 50)

The primary purpose of the color killer system is to prevent spurious or extraneous color "noise" from being observed on the CRT during b-w reception. In the RCA CTC40 chassis this is accomplished by cutting off the bandpass amplifier stage.

Control voltage for the color killer is developed by the ACC detector circuitry and is applied to the killer amplifier. The killer amplifier controls the killer switch stage, which, in turn, switches the bandpass amplifier from a state of conduction to a state of non-conduction when a monochrome signal is received.

During periods of color transmissions, the killer switch stage is biased into saturation by the conduction of the killer amplifier. Saturation of the killer switch effectively clamps both its base and emitter elements to the potential of its collector. Since the killer switch is directly coupled to the base of the bandpass amplifier, its collector voltage (as determined by the divider network of R1-R2) determines the forward bias of the bandpass amplifier. This action controls the conduction of the bandpass amplifier which continues to conduct as long as there is a color signal being received.

During monochrome reception the absence of a color sync signal causes the ACC detector to develop a positive output voltage. This positive output signal biases off the killer amplifier and stops the forward biasing current to the killer switch. The killer switch is then cut-off, effectively "opening up" the bandpass amplifier forward bias circuit. With the bandpass amplifier "cut-off," no extraneous chroma information is fed to the color demodulators.

Chroma Demodulators (Fig. 51)

Three color demodulator circuits are employed in the CTC40 chassis, one for each color-difference signal. The use of a separate G-Y demodulator increases the bandwidth of the signal.

The chroma demodulator circuits are balanced dual-diode detectors. The output of each dual-diode circuit is proportional to both the phase and amplitude of the applied signal.

Two signals are applied to each demodulator circuit: a composite color signal from the bandpass amplifier and the reference signal from the 3.58-MHz oscillator circuit. The phase of the reference signal is shifted a specific amount with respect to the burst signal for each demodulator, extracting the appropriate color-difference signal from the input chroma signal. Circuit action is as follows:

The phase of the 3.58-MHz reference signal applied to the R-Y demodulator is shifted by capacitor C1 and inductor L1. This phase shift permits the R-Y demodulator output to be proportional to the amplitude of the R-Y component of the chroma signal. Phase shifting for the B-Y signal is accomplished by capacitor C2 and inductor L2. The 3.58-MHz reference signal is applied directly to the G-Y component of the chroma signal. Proper loading of the 3.58-MHz CW amplifier is provided by resistors R3 and R4.

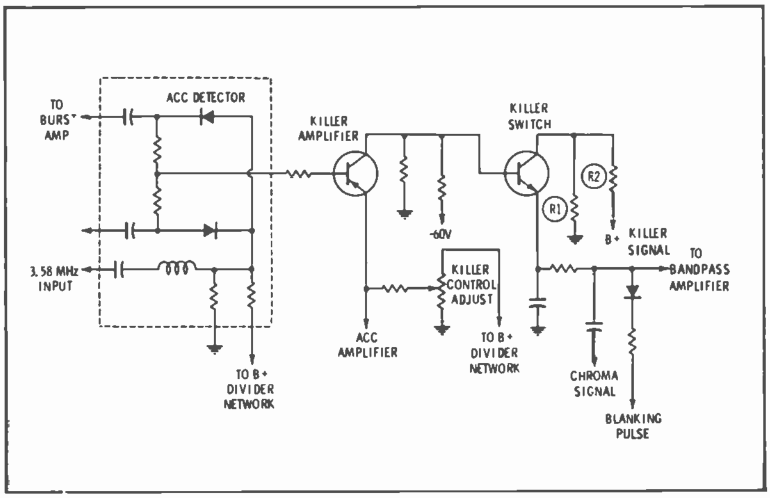


Fig. 50 CTC40 employs two-stage color killer section that controls conduction of the bandpass amplifier stage.

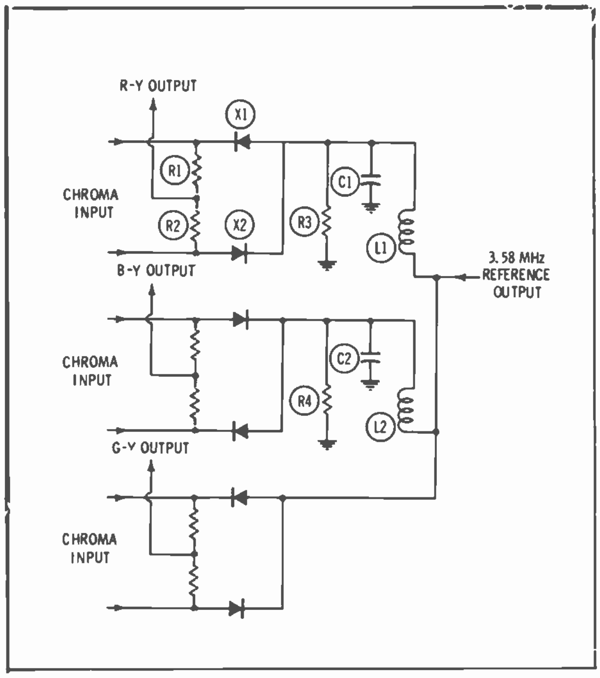


Fig. 51 Balanced dual-diode chroma demodulators produce outputs that are proportional to both the amplitude and phase of the applied chroma signal.

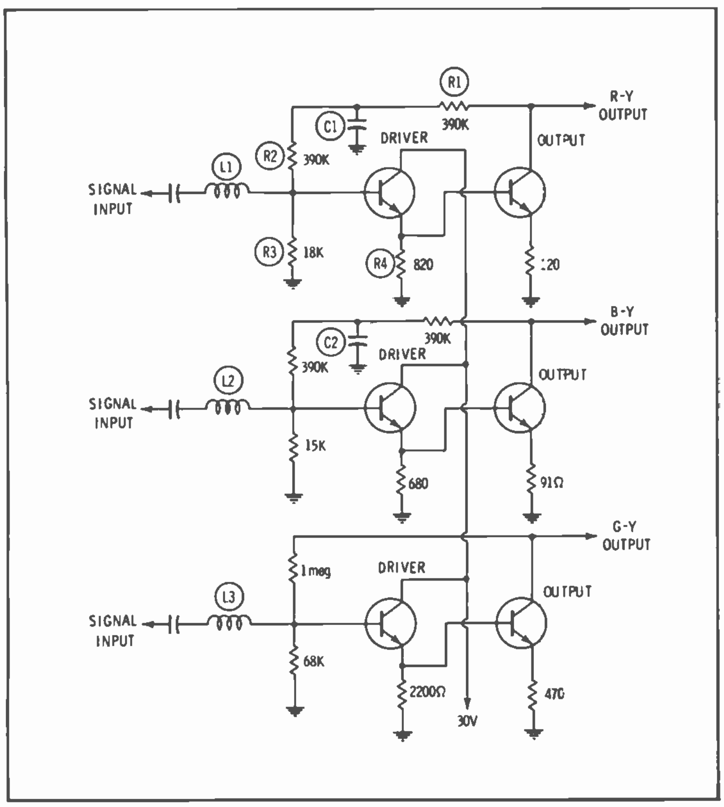


Fig. 52 Output of chroma demodulators is amplified by two-stage color difference amplifier. Driver stage is emitter follower for impedance matching between demodulators and output amplifiers.

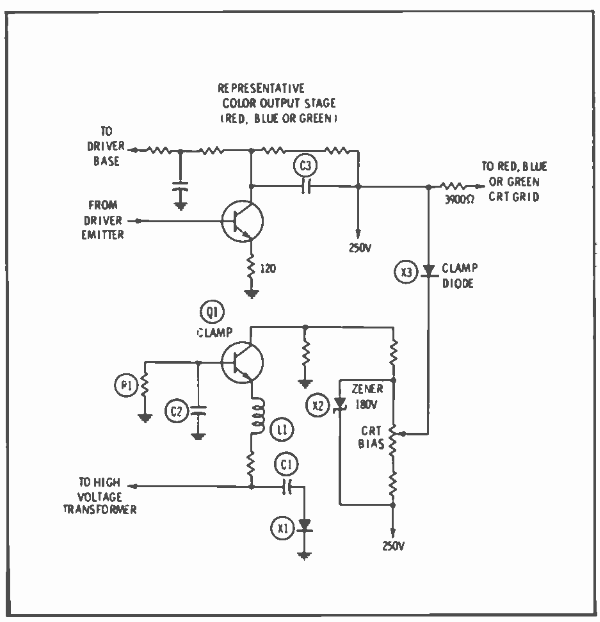


Fig. 53 Clamp diodes in grid circuits of CRT restores DC level of chroma signal lost in AC coupling between chroma demodulators, chroma amplifiers and CRT control grids.

Color Driver and Output Circuitry (Fig. 52)

All three color driver and output circuits are identical with the exception of certain component values. For purposes of operational analysis only one circuit, the R-Y driver and output, will be discussed. The actual theory and operation of the R-Y stage will be representative of all three circuits.

The driver circuit is connected as an emitter-follower to provide the proper impedance match between the comparatively high output impedance of the demodulator circuit and the relatively lower input impedance of the output stage. The 3.58-MHz ripple component is attenuated from the output of the demodulator circuit by a low-pass filter comprised of inductor L1 and the input impedance of the driver stage. Base bias for the driver stage is developed by the divider network comprised of resistors R1, R2 and R3. A better degree of stability is derived by connecting this network between the collector of the output stage and the base of the driver transistor. RF grounding is provided by capacitor C1, which reduces the effects of output-to-driver feedback capacitance. The gain of the output stage is a function of the emitter resistor R4 and the collector load resistance in the control grid circuitry of the CRT.

Clamp Circuitry (Fig. 53)

A certain amount of the chroma DC level is lost because of the AC coupling between the chroma demodulators, chroma amplifiers and CRT control grids. It is the function of the clamp circuit to restore this DC level.

A negative-going 35-volt pulse is coupled from a tap on the horizontal output transformer to the emitter of the clamp transistor, Q1. Diode X1 removes positive ripple between pulses, and inductor L1 suppresses any radiation present in the clamp circuitry. Capacitor C1 is used to sharpen the timing of the horizontal pulses. These negative-going pulses drive the clamp transistor into saturation. The resulting current flow through the base-emitter junction to ground through resistor R1 develops a voltage across resistor R1, and a negative charge on capacitor C2. The charge on C2 allows very sharp turn-off of the clamp transistor at the end of each pulse.

The amplified pulse that appears across the CRT bias control is clamped by zener diode X2, at a level equal to 180 volts below the B+ voltage of 250 volts. A portion of the pulse voltage is fed to a clamp diode, X3, located in each of the three CRT control grid circuits, resulting in diode conduction, which effectively clamps the CRT control grids to the bias pulse voltage (80 volts). This voltage charges the coupling capacitor, C3, located in the color amplifier output stages. **The DC level resulting from the average DC content of the chroma signal is added to this voltage.** Thus, a CRT operating point representing the DC level of chroma information is established.

Tracking

Tracking could be termed as the ability of the CRT to maintain gray-scale throughout the entire brightness range. Proper tracking is accomplished by the development of proper bias levels on the CRT elements, screen grids, control grid, and cathode. The required adjustment procedures of the CTC40 chassis is similar to the procedures used previously in RCA chassis CTC28 and CTC30.

Some automatic correction of the potentials on the CRT elements is provided to offset any AC line-voltage fluctuations. Variable voltages for tracking adjustments are derived from the cathode drive controls, the CRT bias adjustment and the screen grid controls. Any fluctuations in AC line voltage is reflected in the B+ potential, and this, in turn, is reflected in the CRT bias and drive voltages. The action of the zener clamp diode, X2 in Fig. 53, causes the CRT bias voltage to vary directly with changes in B+ potentials. This diode action assures that the CRT grids follow the cathodes during changes in zener B+ and, therefore, assures a constant CRT bias with varying line voltages. The CRT screen voltages are obtained from a regulated source and do not vary under line-voltage fluctuations.

Horizontal AFC and Oscillator

The horizontal AFC and oscillator circuitry is illustrated schematically in Fig. 54. The phase splitter stage supplies equal and opposite sync pulses to the familiar dual-diode phase detector. Incoming sync pulses are differentiated at the base of the phase-splitter transistor to reduce interference from the vertical sync pulses that are present in the output of the sync separator.

Output pulses from the collector and emitter of the phase splitter are coupled to the phase-detector diodes by capacitors C1 and C2.

A reference voltage taken from the high-voltage transformer is applied to the common diode junction through a waveshaping network. This network shapes the negative-going pulses from the high-voltage transformer into a sawtooth signal that is applied to the AFC circuit. The frequency of the pulses sampled from the high-voltage transformer is the same as that of the horizontal oscillator.

When the pulses from the high-voltage transformer and the incoming horizontal sync pulses occur at the exact same frequency, each diode is keyed into conduction by the sync pulses as the reference voltage passes though zero. The current through each diode will be equal, resulting in equal and opposite charges on capacitors C1 and C2. As these capacitors discharge through resistors R1 and R2, equal and opposite voltages are developed across the resistors. The voltage at their junction is zero (with respect to ground), and, consequently, the amount of correction voltage developed is zero.

If the horizontal oscillator is running at a frequency less than that of the incoming horizontal sync pulses, a change in the relative position of the reference voltage waveshape during the application of the sync pulses will result. The sync pulses will key the diodes into conduction during the positive portion of the retrace slope; diode X1 will conduct more strongly than diode X2; and the charge on capacitor C1 will become more positive, while the charge on capacitor C2 will become less negative.

Discharge action of these capacitors through resistors R1 and R2 will result in an imbalance of current flow through R1 and R2, and the voltage developed at their junction will go positive. This positive voltage is the correction voltage for the horizontal oscillator and will cause the oscillator to increase in frequency.

Should the oscillator be running at a frequency greater than the incoming sync pulses, a negative correction voltage will be developed at the junction of R1 and R2 as the result of circuit action similar but opposite to that described in the preceding paragraph. Application of the negative correction voltage to the oscillator will produce a decrease in the oscillator frequency.

The DC correction voltage present at the junction of R1 and R2 is fed to an AFC limiting and filtering circuit comprised of diodes X3 and X4, capacitors C3 and C4, and resistor R3. The function of this circuit is two-fold: The limiting diodes prevent the AFC correction voltage from exceeding —0.5V to 0.5 volts; and the filter network prevents the AFC output from being contaminated by unwanted frequencies, such as 60 Hz. (A 60-Hz signal present at this point would result in horizontal bending, twisting, etc.)

Horizontal Oscillator

A blocking oscillator circuit is employed as the horizontal oscillator in the CTC40 chassis. Basic circuit action is as follows:

Voltage pulses present on the collector of the horizontal oscillator transistor are transformer-coupled into the base circuit, driving the stage into cutoff. During the time that the oscillator transistor is cutoff, capacitor C5 discharges through the horizontal hold control circuitry to the "turn-on" potential of the oscillator. The oscillator conducts, a pulse appears at the collector and is coupled to the base, and the cycle repeats.

The settings or adjustments of the horizontal linearity and horizontal hold controls determine the discharge time of capacitor C5 or, in other words, the length of time the transistor remains cut off. In this manner the horizontal hold control determines the frequency of the horizontal oscillator.

This is accomplished by adding the correction voltage from the AFC circuit to the charge capacitor C5. This correction voltage, depending on its polarity and amplitude, will either add or subtract from the charge on capacitor C5, which, in turn, either increases or decreases the time required to discharge C5 to the turn-on potential of the oscillator. This circuit action alters the frequency and phase of the oscillator in accordance with the broadcast sync pulse.

Horizontal Oscillator B+ Source

A special 30-volt source is used to supply B+ to the horizontal oscillator. This is done to assure that the horizontal oscillator will be capable of supplying adequate drive to the horizontal output stage at the instant power is first applied to the receiver. The filter circuit of the normal +30-volt supply source requires too much time to reach full value; therefore, the special circuit, comprised of X5 (zener diode) and resistor R4, is used to develop the required 30 volts from the more lightly filtered 155-volt source. This circuit functions to reduce the time required for the oscillator output to reach its normal operating level.

Horizontal Output Circuitry

It is necessary to slightly alter the shape of the horizontal oscillator output waveform to minimize the possibility of pretriggering the commutator switch. This is accomplished by the waveshaping network composed of diode X5, capacitor C6 and resistors R5 and R6.

The voltage developed across the output winding of the horizontal blocking oscillator transformer is coupled to this waveshaping network. R6 and C6 function as a dif-

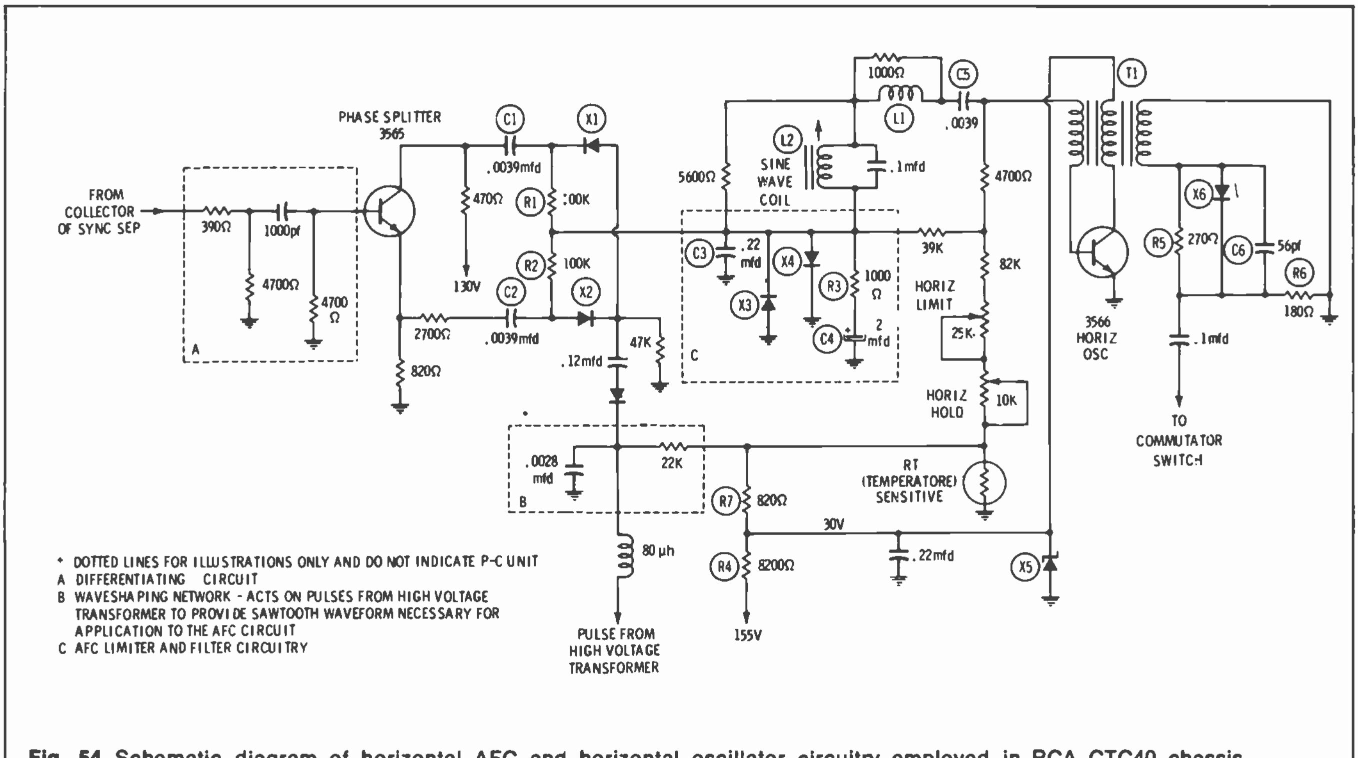


Fig. 54 Schematic diagram of horizontal AFC and horizontal oscillator circuitry employed in RCA CTC40 chassis.

ferentiating network, producing a positive voltage spike to turn on the commutator switch. The diode, X6, is reverse biased during the negative-going portion of the output voltage waveshape. This permits capacitor C6 to discharge through the parallel paths provided by R5 and R6. This discharging action holds the waveshape negative until the next positive pulse arrives. Thus, the commutator gate is held negative during trace-time, reducing commutator pretriggering.

Control of Temperature Induced Frequency Change

It is an inherent characteristic of transistors that their operation will vary with changes in ambient and internal temperature. A thermistor, RT, in conjunction with resistor R7, functions as a temperature-sensitive, voltage-divider network. As the temperature of the horizontal oscillator transistor changes, its operating frequency tends to change.

The same changes in temperature that affect the transistor also affect the thermister (RT). The transistor base-circuit voltage will be altered by the temperature-induced changes in the divider network comprised of RT and R7. This change in base-circuit voltage will be in a direction that will cancel out the effects of temperature on the transistor.

Horizontal Deflection

The RCA CTC40 chassis utilizes two silicon-controlled rectifiers (SRC's) and their associated components to generate the necessary

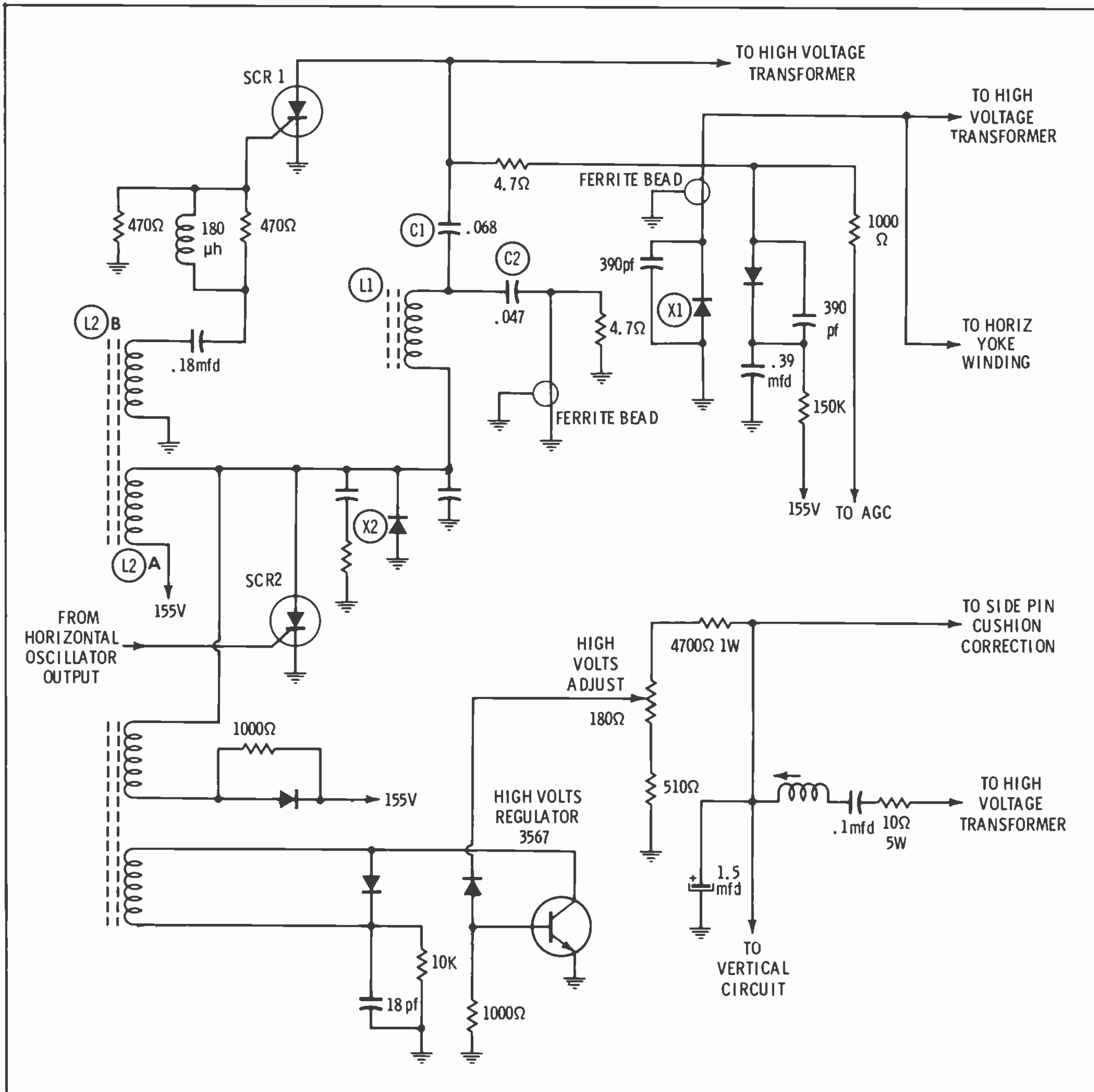


Fig. 55 Partial schematic diagram of horizontal output circuitry found in CTC40.

yoke current and fulfill high-voltage requirements.

The function of any horizontal deflection system used in television receivers utilizing electromagnetic deflection is to provide a linear flow of current through the yoke windings which, in turn, moves an electron beam from one side of the picture tube screen to the other in a linear sweep. This action is normally referred to as "trace", and the yoke current that caused the deflection is called "trace current."

Trace current must be in sync with the incoming TV signal. The yoke current must also provide a means of returning the CRT beam to the starting side of the CRT screen. The current that accomplishes this is referred to as retrace, or flyback, current.

Circuit Action

A partial schematic of the horizontal output circuit is shown in Fig. 55. Diode X1 and silicon-controlled rectifier SCR1 control the flow of current through the horizontal yoke windings during the CRT beam trace time. Diode X2 and silicon-controlled rectifier SCR2 control the flow of current through the horizontal yoke windings during retrace time.

Energy storage and timing properties are provided in the circuit by components L1, C1, C2 and Cy. Inductors L2A and L2B provide a charge path for L1 and C1, and a gating, or keying, signal to SCR1. The complete horizontal-deflection yoke-current cycle can be divided into a sequence of individual actions involving different modes of horizontal circuit operation. These actions are accomplished during discrete intervals of the horizontal deflection yoke current cycle.

During the first one half of CRT beam trace time, the current through the horizontal deflection coils decreases towards zero and flows through trace diode X1, resulting in a charge build-up on capacitor Cy. During this interval (first half of trace time), silicon controlled rectifier SR1, the trace SCR, is prepared for conduction by the application of the proper gate-voltage pulse. However, the SCR1 will not conduct until its anode/cathode junction is forward biased. This condition will be satisfied during the second half of the beam trace cycle.

At the end of the first half of trace, yoke current reaches zero, capacitor Cy starts discharging through the yoke inductance, and the current flow through the circuit reverses, reverse-biasing diode X1 and, simultaneously, forward biasing the SCR1. The capacitor discharges into the yoke inductance through SCR1 and the resulting yoke current completes the second half of trace.

When the second half of trace is concluded, the CRT beam has scanned across the entire width of the CRT screen. At this point, a pulse, derived from the horizontal oscillator circuit, keys the retrace SCR into conduction. This action releases the charge previously built up, or stored, on capacitor Cy, and current flows into the commutator circuit comprised of inductor L1 and capacitor C1.

Because of heavy forward current flow through the yoke circuit (SCR1, Ly and Cy), the net current resulting from the combined circuit actions of the commutating switch circuit and the yoke circuit continues to allow the trace rectifier, SCR1, to conduct.

At this point both rectifiers, SCR1 and SCR2, are conducting. However, the current flowing in the commutator circuit increases much more rapidly than the current flow in the yoke circuit. After an extremely short period of time (two to three microseconds) the net current flowing in SCR1 reverses, turning off SCR1 at the start of retrace.

Circuit conditions are now set to initiate retrace: Trace rectifier SCR1, along with diode X1, is cut off, and retrace rectifier SCR2 is conducting. The result is a series resonant circuit comprised of inductor L1, capacitor C1 and the horizontal yoke windings. (Capacitor CY is also in series with these components, but, because of its value, can be disregarded.)

The current through this circuit causes the CRT beam to retrace half way across the screen. At this point the current flow has decreased to zero. Current flow in the series resonant circuit now reverses, and retrace rectifier SCR2 ceases conduction because the current flow in the circuit is opposite the normal flow of forward current.

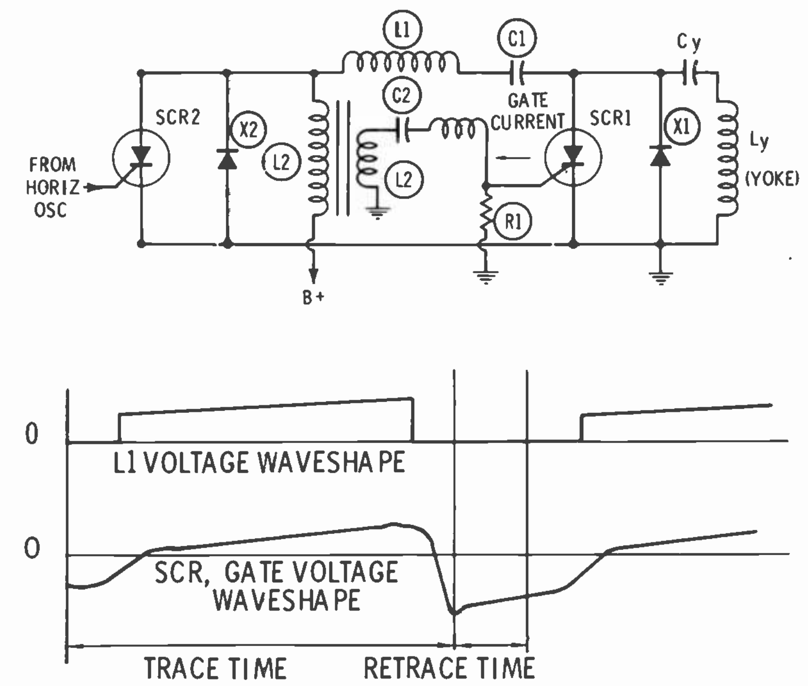


Fig. 56 Simplified schematic of horizontal output circuit, and diagram showing instantaneous voltages across L1 and SCR1 during trace and retrace times.

Diode X2 is now forward-biased by this reversal of current and starts conducting, supplying the energy for the remainder of retrace. The energy previously stored on capacitor C1 has been returned to the yoke inductance.

Retrace current flowing in the horizontal yoke winding returns the electron beam to its starting point. The time interval of yoke retrace current flow is made equal to the desired retrace time by selection of the proper values of components L1, C1 and Ly.

These components are selected to be resonant at a frequency which has a period equal to two times the retrace time interval. Therefore, the current flowing during one half cycle of circuit oscillations will accomplish the full retrace function.

After completion of one full cycle (trace and retrace), the circuit must be made ready for the next cycle. This includes restoring energy to the commutator circuit and resetting the trace rectifier, SCR1. Both of these functions are performed by utilizing circuitry which includes inductor L2.

During retrace, inductor L2 is connected between B+ and ground by the conduction of SCR2 and diode X2, respectively. When X2 ceases conduction, inductor L2 is removed from ground. A charge is built up on C1 from the B+ line through inductor L2. This charging process continues throughout the trace interval, until retrace begins. The charge on C1 serves to replenish energy to the yoke circuit during the retrace interval.

The voltage developed across inductor L2 during the charging of capacitor C1 is used to forward-bias the gate of SCR1. This sets up SCR1 and enables it to conduct upon receiving the proper signal. The voltage developed across inductor L2 is coupled to the gate of SCR1 via L2A, C2 and R1.

These components form a wave-shaping network that forms a pulse with the proper shape and amplitude to enable SCR1 to conduct when its anode/cathode junction is forward-biased. This will occur approximately mid-way through the trace interval.

This concludes our analysis of RCA's solid-state CTC40 color chassis. (Schematic diagrams used in this chapter courtesy of RCA.)

Zenith's 12B14C50 Color TV Chassis

The new Zenith 12B14C50 color TV chassis employs more transistors than any previous Zenith color chassis. It is equipped with 14 transistors and 12 tubes, and a fully transistorized "Y" channel.

The color and the "Y" signal are pre-mixed and applied to the cathodes of the CRT, permitting the CRT control grids to be returned to a fixed DC source.

As in an earlier color chassis,

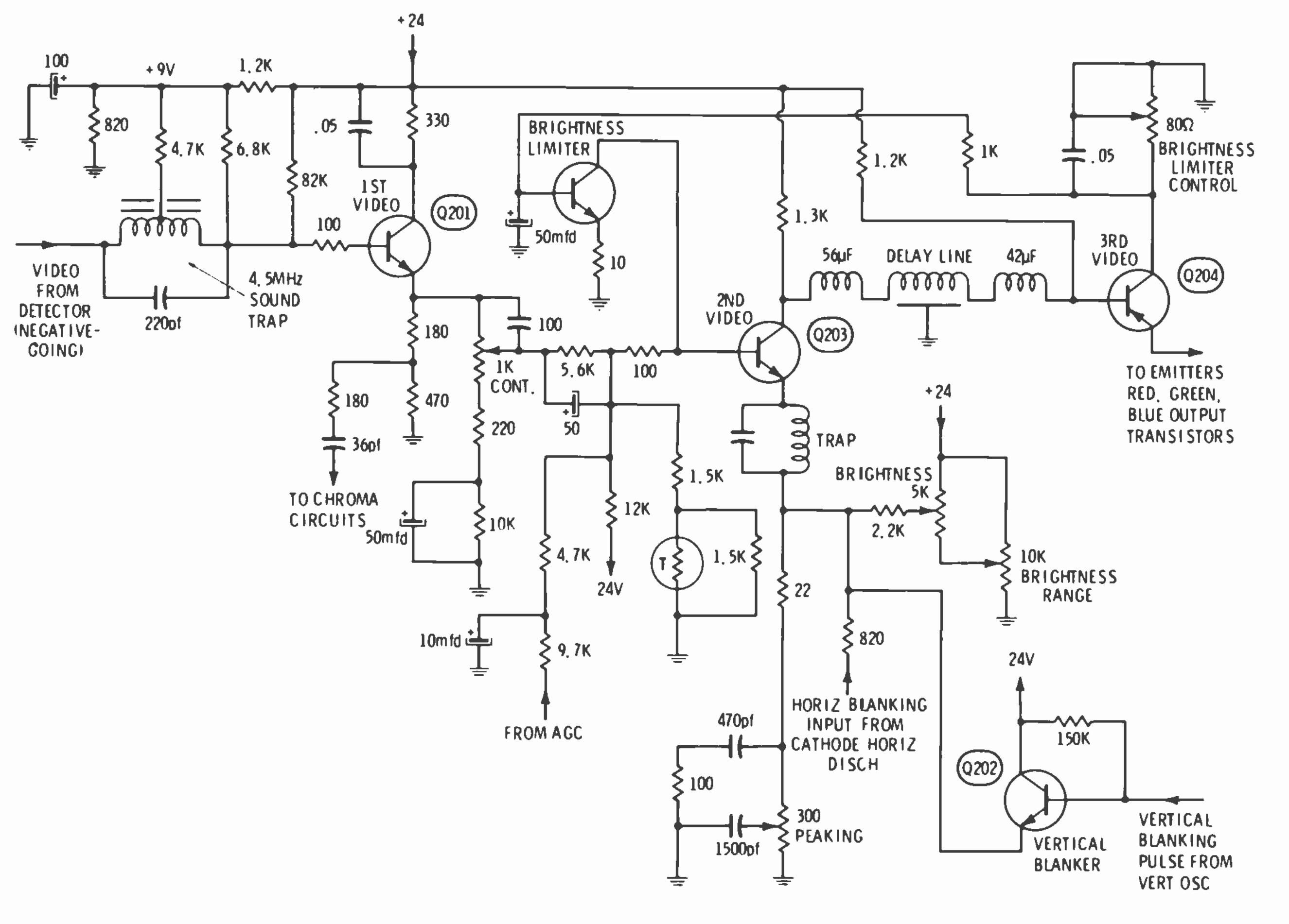


Fig. 57 Transistorized "Y" channel with brightness limiter and vertical blanking transistors is used in new Zenith chassis.

the 2nd color amplifier and the color demodulator integrated circuitry is contained in a plug-in module which can be removed and replaced from the top of the chassis.

Zenith uses a ganged contrast and color control, which they call a "color commander." To maintain the correct black level with a change in contrast, a special "black tracking" circuit increases or decreases the brightness slightly as the Contrast/Color control is rotated.

The chassis also is equipped with a simplified automatic degausser circuit, a new high-level noise clipper, a simplified focus circuit, a pincushion circuit using a saturable reactor, and an "automatic" tint control circuit.

The Video (Y) Amplifiers

The "Y" channel in the 12B14-C50 chassis is all transistorized, as shown in Fig. 57. There is an emit-follower 1st video amplifier, a common-emitter 2nd video amplifier, and an emitter-follower 3rd video amplifier. The 3rd video amplifier also serves as an impedance matcher and "modulator" to the emitters of the red, green, and blue output transistors.

The contrast control is in the emitter circuit of the 1st video; the chroma take-off also is in this circuit. The brightness and brightness range controls are in the emitter circuit of the 2nd video. Vertical and horizontal blanking also is inserted at the 2nd video emitter.

Stabilization from three sources is applied at the base of the 2nd video transistor, Q203.

1st Video Stage

The video signal from the detector diode is fed through peaking coils and a sound trap to the base of the 1st video, Q201. At first glance, the resistors in the base circuit of Q201 appear to comprise a rather complex DC biasing network. The reason for this many resistors, according to Zenith, is to provide a better AC match between the medium-high output impedance of the diode detector and the medium-low input impedance of the 1st video transistor. As far as the DC bias is concerned, the resistors are in parallel. The 4.7K-ohm resistor helps sharpen the tuning of the trap.

The contrast control circuit parallels the 180- and 470-ohm emitter resistors and varies the AC video signal level fed to the base of the 2nd video amplifier, Q203.

The "black-tracking" circuit operates because of the different AC and DC impedances of the contrast control circuit. For video (AC), the circuit offers an impedance of 1220 ohms (the 1k contrast control and the 220-ohm resistor in series with it, disregarding the 180- and 470-ohm parallel resistors) because AC is bypassed around the 10k resistor by a 50-mfd capacitor. For DC, however, the 50-mfd capacitor has no effect, so the circuit impedance is 11,220 ohms. Consequently, with the contrast control arm at the low end, the signal level is reduced about 5½ times, but the DC bias level (affecting the brightness) is reduced only about 1/11. In other words, lowering the contrast also slightly lowers the brightness, while increasing the contrast slightly increases the brightness, both actions keeping the CRT black level relatively constant.

Setting the Brightness Range and Brightness Limiter

To get full effectiveness from the Brightness Limiter circuit it must be adjusted in conjunction with the Brightness Range control as follows:

1. Turn the Brightness Limiter control fully clockwise.
2. Turn Brightness control fully clockwise.
3. Adjust Brightness Range until picture blooms 1 inch.
4. Adjust Brightness Limiter counterclockwise until picture blooms only ¼ inch.

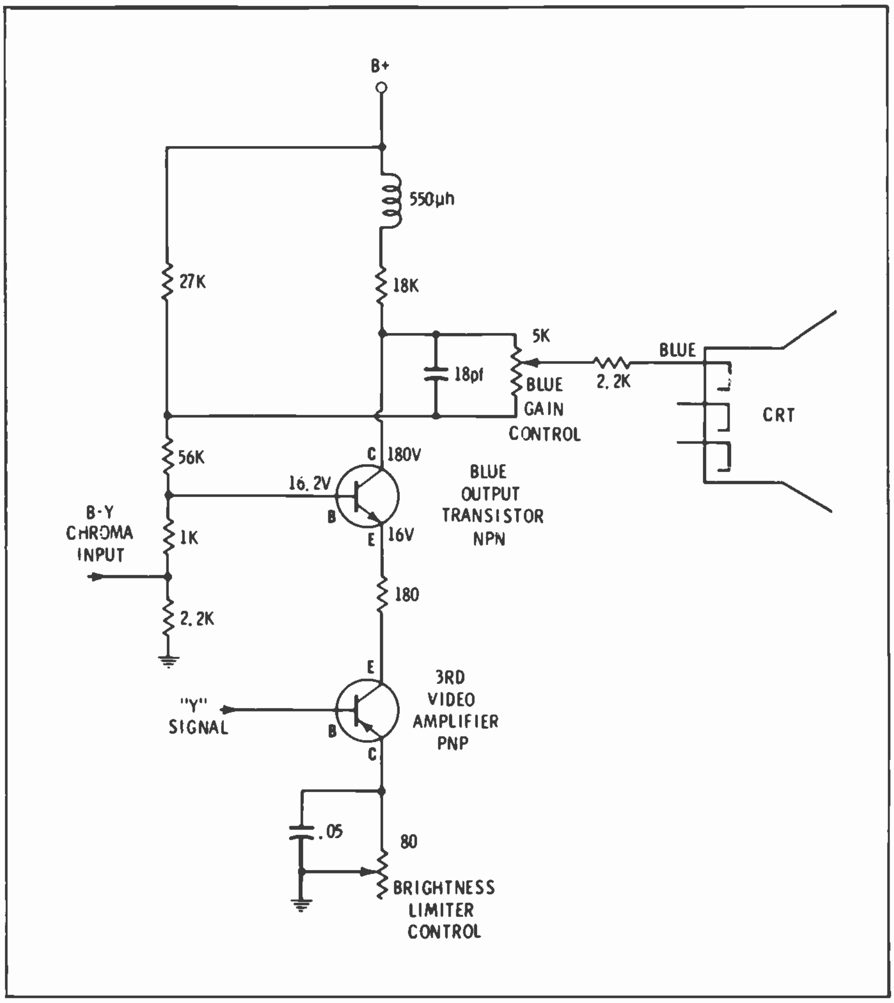


Fig. 58 Simplified schematic of pre-CRT mixing of chroma and "Y" signals.

The 2nd Video

With direct-coupling between video stages, which is used in most color sets, a change in bias in any stage causes a corresponding change in picture tube brightness. This is why brightness problems in color sets often occur in the video amplifiers, and why the brightness control and brightness range can be inserted at almost any convenient spot in the circuit following the chroma takeoff and contrast control.

But there are some disadvantages to the direct-coupled circuit. If, during warmup, the current flow through one or more of the transistors changes, the brightness also will change. To compensate for warmup change in this circuit, a negative-temperature-coefficient (decreases in resistance with increase in heat) thermistor is placed in the base bias circuit of the 2nd video stage. As the transistor heats it tends to draw more current, but the thermistor compensates by reducing the DC bias to the base, which, in turn, reduces the conduction of the transistor circuit.

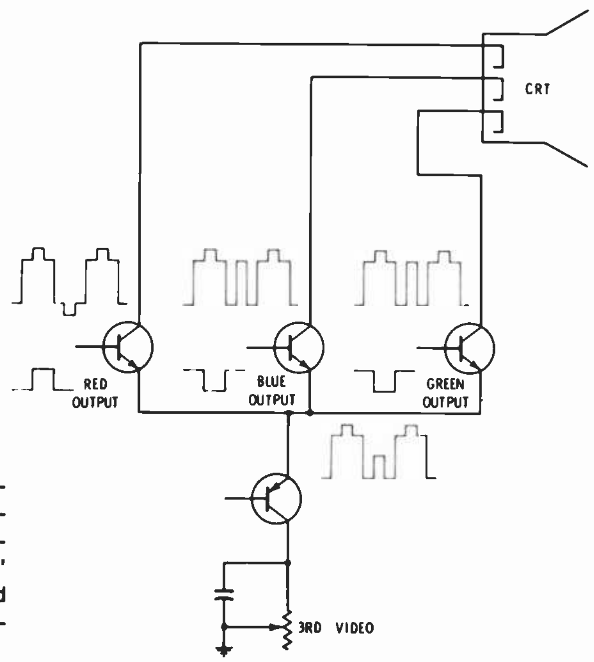


Fig. 59 Simplified circuit showing matrixing of chroma signals with the "Y" signal when a red color bar is being received.

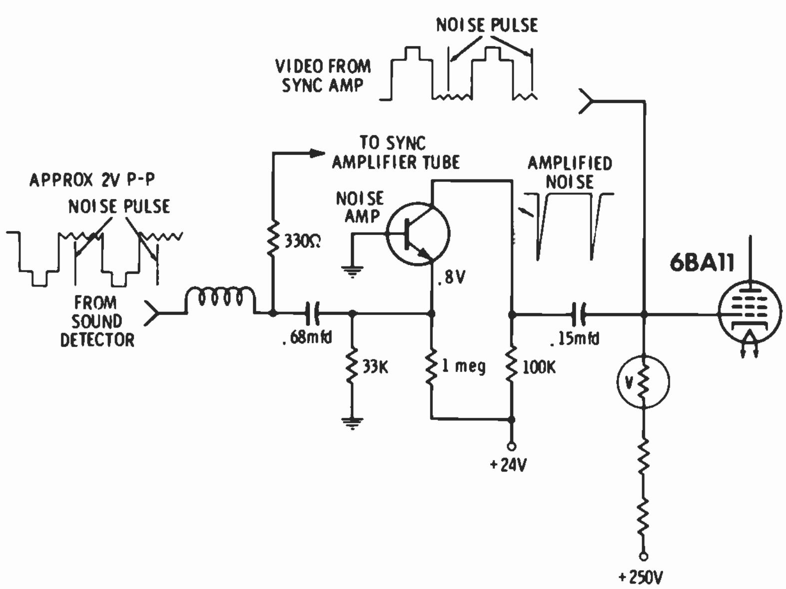


Fig. 60 Common-base, high-level noise amplifier used in new Zenith chassis. During normal reception, video input signals are not strong enough to override the bias of the noise amplifier transistor; however, a strong noise pulse will cause it to conduct and apply a strong negative pulse to the 6BA11 grid.

Another problem in direct-coupled circuits is that when weak signals are received the detector output drops below the normal level maintained by the AGC on stronger signals. This means that the brightness would increase on weak signals. To prevent this, the base of the 2nd video amplifier is tied through an RC network to the AGC line. Because the AGC line goes more negative when a weak signal is received, more negative voltage is applied to the base of the 2nd video, preventing a drastic change of brightness.

A third stabilizing signal is introduced at the base of the 2nd video by the brightness limiter transistor; this application will be discussed in detail later in this article.

Located in the emitter circuit of the 2nd video amplifier are another trap, a peaking control circuit, input for vertical and horizontal blanking signals, and the brightness and brightness range controls.

The peaking control functions by varying the frequency response of the video amplifier. When the arm of the peaking control is at the "hot" end, the higher video frequencies are bypassed around the major portion of the emitter resistance, producing maximum high-frequency gain and pictures with sharp outlines but more apparent noise and snow, especially in weak signal areas. With the peaking control arm at the "ground" end, the high-frequency gain is reduced and the outlines in the picture tend to smear, but noise and snow are less apparent. The peaking control is a "customer preference" control and operates on the same principle as a tone control in an audio circuit. "Normal" operation is with the control at about midrange.

The brightness and brightness range controls vary the emitter bias of the 2nd video and, in turn, this change in bias is projected through the remainder of the video stages to the cathodes of the CRT. Brightness is minimum when the arm of the brightness control is at the +24-volt side. For this reason, the brightness control can always vary the brightness to zero regardless of the setting of the brightness range control. The brightness range varies the amount of positive voltage on the "low" side of brightness control and sets the maximum level to which the customer can increase the brightness.

Horizontal blanking also is inserted at the emitter of Q203, through an 820-ohm resistor which is tied to the cathode of the horizontal discharge tube.

Vertical blanking is applied at the same point, but the vertical pulses, taken off the vertical oscillator-amplifier, are first fed to a vertical blanker transistor to reduce the loading and provide a better impedance match.

3rd Video Stage

The output of the 2nd video amplifier is fed through the delay line to the base of the 3rd video. There is an 80-ohm adjustable resistor in series between the 3rd video collector and ground. This is the brightness limiter control. The positive voltage developed across this control is filtered to remove video, and the resultant DC then is fed to the base of the NPN brightness limiter transistor. This positive voltage causes the transistor to conduct and supply a negative voltage to the base of the 2nd video amplifier.

The operation of this two-stage compensating circuit is as follows: If the current through the 3rd video transistor increases for any reason, there will be additional positive voltage applied to the base of the brightness limiter, which will increase the negative bias on the 2nd video, which, in turn, will reduce the bias on the 3rd video, almost completely cancelling an increase of current in the 3rd video.

The opposite condition occurs if the 3rd video transistor current should drop for any reason. In other words, this is an amplified stabilizing circuit that limits extreme changes in bias, thereby limiting undesirable brightness changes.

Pre-CRT Matrixing

Before increased use of transistors, almost all manufacturers mixed (matrixed) the video and chroma signals in the picture tube by applying the video to the CRT cathodes and the color to the CRT grids. This had the disadvantages of requiring high-level color amplifiers and the possibility of tracking problems in either or both the video and color amplifier circuits.

By pre-mixing the video (Y) and the color, the final video amplifiers can amplify both the video and the color, and the CRT grids can be returned to a fixed DC source, reducing the possibilities of tracking errors.

Fig. 58 shows how the video and one color signal are mixed in this chassis. The "Y" signal is fed into the 3rd video amplifier, which is in series with the emitter of the B-Y (blue) output transistor. This, in effect, mixes the two signals so that a composite of the two appears at the collector of the blue output transistor, where it is coupled through the blue gain control to the blue cathode of the CRT.

Fig. 59 shows, in simplified form, how this pre-CRT mixing occurs. Suppose a red bar is being transmitted. The bar will appear gray on a black-and-white set because the video signal is about "half way to black", as can be seen by the waveform at the emitter of the 3rd video amplifier. With no color input to any of the output transistors, there will be an exact replica of this signal (except amplified) at all three output collectors. (Remember,

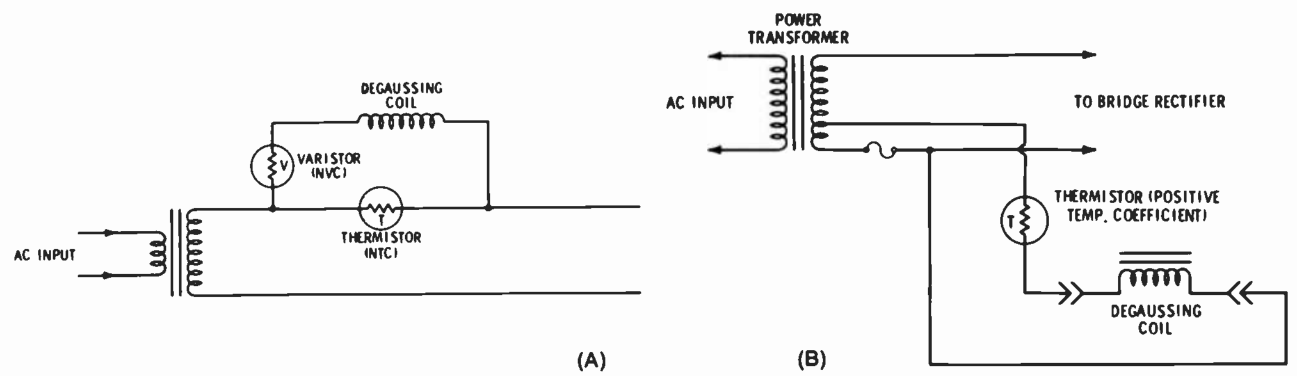


Fig. 61 (A) More conventional degaussing circuits have imployed at least two voltage and/or temperature sensitive resistors. Zenith automatic degaussing circuit (B) uses only one thermistor, and it has a positive temperature coefficient.

there is no phase inversion because the input is to the emitters.) But, with a red signal coming from the color demodulator, there is a positive pulse at the base of the red output transistor. This positive pulse causes the collector to go negative (phase is inverted between base and collector), making the red CRT cathode also go negative, which, in turn, increases the brightness of the red gun. At the same time the positive pulse is arriving at the red output transistor, the color demodulator is supplying negative pulses to the basis of the green and blue output transistors, which produces positive-going signals on their collectors and on the blue and green CRT cathodes, decreasing to zero the brightness of these two colors.

The same process occurs if a blue or green bar were transmitted, except that the corresponding output transistor would receive the "turn on" signal and the other two the "turn off" signal from the chroma demodulator.

The Noise Clipper

A new high-level noise canceller is used in this chassis. As shown in Fig. 60, the transistor is in a common-base circuit designed to feed high-level negative noise pulses to the 1st grid of the 6BA11 sync tube. During normal operation the video input to the noise amplifier transistor is insufficient to overcome the reverse bias between its base and emitter, and, consequently, there is no transistor output to the grid of the 6BA11. However, when a strong noise pulse occurs, it overrides the transistor reverse bias and appears as an amplified negative pulse at the grid of the sync tube. (There is no phase change between input and output of a common-base amplifier.) This amplified negative pulse cancels the effect of the positive-polarity noise pulse, which also is fed through normal channels to the other input grid of the 6BA11. This prevents the noise pulse from falsely triggering either sweep circuit.

The Automatic Degaussing Circuit

Most automatic degaussing circuits employed previously use at least two self-variable resistors, either varistors or thermistors or a combination of the two, such as the design shown in Fig. 61A. In this circuit both control units have negative coefficients.

When power is first applied to the set and the thermistor is "cold", its resistance is relatively high, about 120 ohms. Consequently, maximum voltage is developed across it. At the same time, the resistance of the varistor is at minimum, which permits maximum current flow through the degaussing coil.

As the set warms up, the resistance of the thermistor decreases, finally reaching about 2 or 3 ohms. The voltage drop across it also decreases, which, in turn, causes the resistance of the varistor to increase to maximum, cutting off almost all current flow through the degaussing coil.

The automatic degaussing circuit in Zenith's new chassis uses a low-voltage, high-current degausser and a single thermistor with a positive temperature coefficient. When the set first is turned on, the thermistor is cold, its resistance is at minimum, and about 6 amps of current flow through the degaussing coil. During the first few seconds of operation, the resistance of the thermistor increases to maximum and current through the degaussing coil decreases to about 30ma, which is low enough not to affect the picture tube, yet sufficiently high to keep the thermistor warm enough to maintain a high resistance.

The Focus Circuit

The focus circuit in the new Zenith chassis, shown in Fig. 62, is simply a voltage divider network connected directly to the 25KV high-voltage source for the CRT. By placing the control in the "tail", or "cold", end of the circuit, the voltage on the control is kept at a reasonable level, and the danger of arc-over or shock is minimized.

Pincushion Correction

Another change in the new Zenith chassis is the deletion of the pincushion amplifier tube. A saturable reactor is used instead—not a new idea, but one that Zenith has not used before.

The saturable reactor uses two separate windings on a special core material. One of the windings is in toroidal form (Fig. 63A), which has no air gap and, because it keeps all the lines of force within the core itself, needs no shielding—it does not radiate nor does it accept electromagnetic signals not directly induced into the core by another toroidal winding on the same core. However, because the core has no air gap, it does saturate rather easily with medium current. With low current in the coil, the coil has a high inductance and reactance. But as the coil current increases, the inductance and the reactance decrease, and once the core has saturated there is little reactance to current variations in the coil.

The second winding of this pincushion unit is wound in normal fashion on the same core as the first winding (Fig. 63B). Because the two windings are at a 90-degree angle to one another, there is no transfer

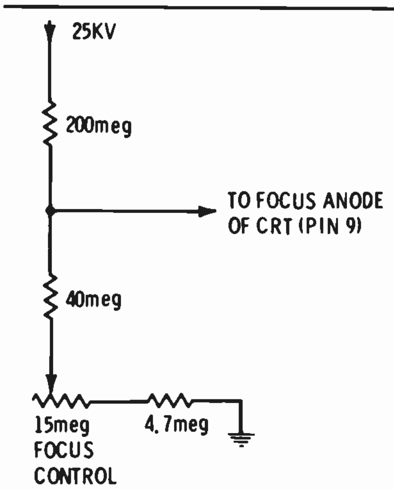


Fig. 62 Resistive-divider focus circuit eliminates the need for a focus rectifier.

of voltage between the two; however, because the core is common to both windings, the second winding can vary the reactance of the first winding by causing core saturation.

Utilizing this effect, the signal applied to the toroidal coil can be modulated by the signal applied to the second winding, but the second winding will not be modulated by the signal in the toroidal winding.

For correction of pincushion at the top and bottom of the screen, the vertical signal is applied to the toroidal coil, and the horizontal signal is applied to the normal coil. The vertical signal then is modulated by the horizontal signal, but the vertical signal does not affect the horizontal because, as far as the normally wound coil is concerned, its core has a large air gap and never saturates.

In this particular reactor it requires current flow in both the vertical and horizontal circuits to saturate the core.

Here's how it works: At the beginning of the vertical sweep (top of raster), the sawtooth of current is at its highest amplitude, and this current, plus the current of the top horizontal line, produces maximum saturation of the reactor, which produces minimum reactance and maximum vertical current flow. Because the current is maximum at the beginning and end of each horizontal line, there would be maximum vertical deflection at the ends of the lines—which is just opposite the effect needed. To correct for this, a secondary winding tunable around 15,750 Hz is provided. By adjusting the winding, the phase can be shifted to produce maximum vertical deflection at the center of the screen, instead of at the ends of the lines.

As the beam sweeps downward across the CRT, the vertical yoke current decreases at a linear rate, decreasing the saturation and increasing the reactance of the coil, so that no correction occurs at the center of the screen. As the current increases again to sweep the beam on down to the bottom, the saturation again occurs, producing maximum vertical sweep at the center of the raster.

"Automatic" Tint Control

By pulling out the tint control knob on this Zenith chassis the customer can have "automatic" tint. The control functions by shifting the demodulation angle from the normal 105 degrees to 132 degrees. This shifts the R-Y (red) more toward orange, and there is less apparent flesh tone change as the set is switched from channel to channel.

Fig. 64 shows how the automatic tint control circuit functions. The phase of the 3.58 oscillator injection to the color demodulator is changed by switching in a simple phase-shifting network so that the demodulation angle is changed by about 27 degrees. ▲

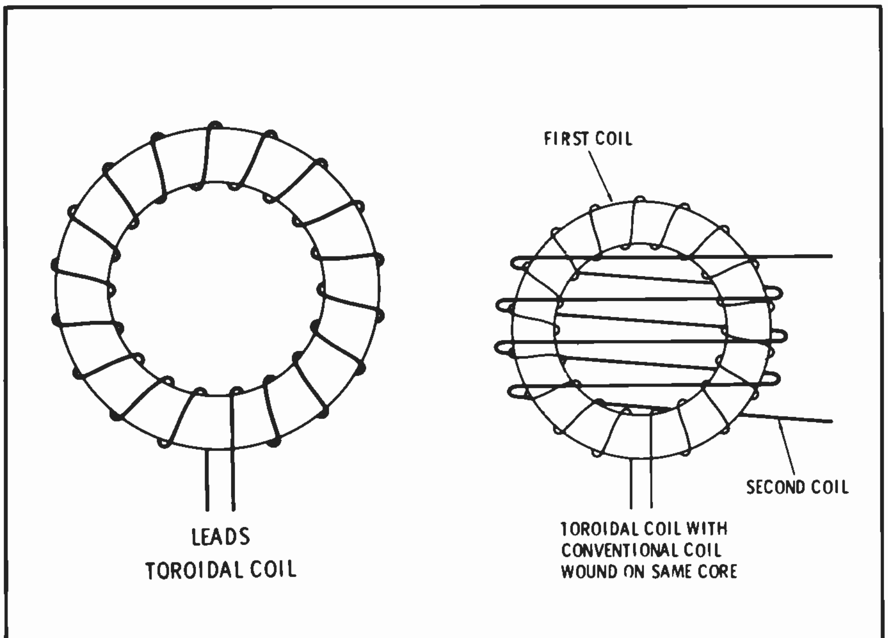


Fig. 63 (A) Toroidal coil. **(B)** Toroidal coil with a conventional coil wound on same core. There is no electromagnetic transfer of energy between the two coils, but current in the conventional coil can cause core saturation and a reduction in reactance of the toroidal coil.

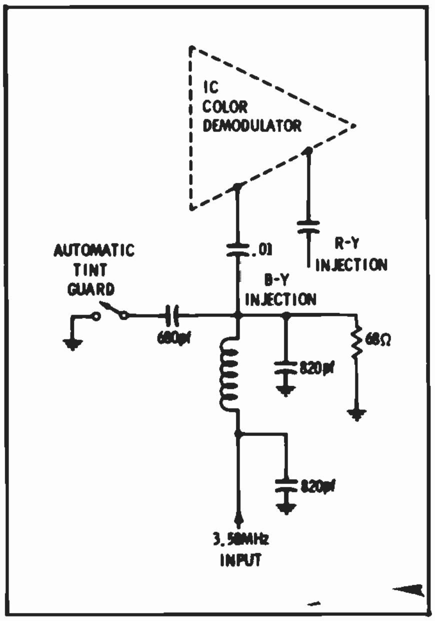


Fig. 64 Changing the 3.58 oscillator injection phase angle at the input of the color demodulator by about 27 degrees shifts the R-Y (red) signal toward orange, so there is less apparent flesh tone change when the set is switched from channel to channel.

Chapter 8

Automatic Tint Control Systems

Some Common Tint Problems

Many color TV manufacturers now have deluxe models featuring automatic fine-tuning (AFT) to accurately adjust the fine tuning electronically, and automatic chroma gain control (ACC) to hold the color saturation nearly constant. One annoyance still remains in the frequent changes of hue that are not caused by the receiver. You undoubtedly have seen the ghastly purple or sickly green faces that were produced when the program source was changed from live newscasts to filmed commercials, from old movies to taped commercials, or from one TV channel to another. This is a many-sided problem with no single source or cure. Some cases are the result of the different color rendition of various brands of slide or movie film, while others may be due to an unbalance of the three primary colors in the TV camera.

These broadcast shortcomings can be minimized, but not eliminated, by receiver adjustments. Color hue changes caused by variation in the phase of the transmitted burst signal usually can be restored by repeated use of the hue control on the receiver. However, this activity is often accompanied by a marked increase in the viewers blood pressure.

Some system to adjust the hue automatically is very desirable. At first thought, one solution would be to use a phase detector to electronically maintain a constant hue. Unfortunately, there is no standard to compare the burst phase against.

How Magnavox's ATC Works

One practical answer to this distressing hue problem is found in the Magnavox color receivers that use the T940 chassis. The manufacturer calls these models Total Automatic Color (TAC) since they have AFT, ACC and ATC (automatic tint control). Their ATC circuit changes all the yellow and red areas of the picture to an orange that is acceptable as skin color. Thus, skin color that ordinarily would look slightly green or purple will be rendered as orange so long as the ATC is switched on.

The ATC circuit board has no tubes, but uses diodes and transistors. It is mounted on the tuner mounting assembly and is completely shielded. A picture of the

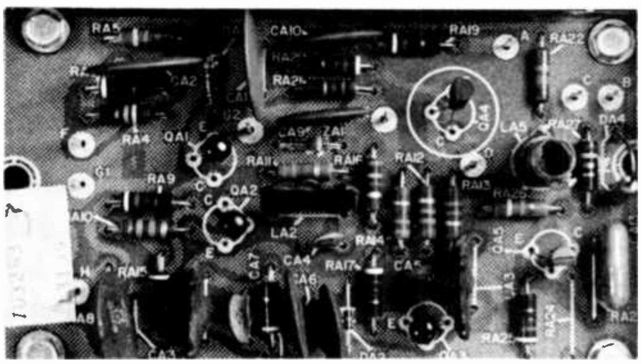


Fig. 1 (A) Circuit board and components for the Magnavox ATC circuit.

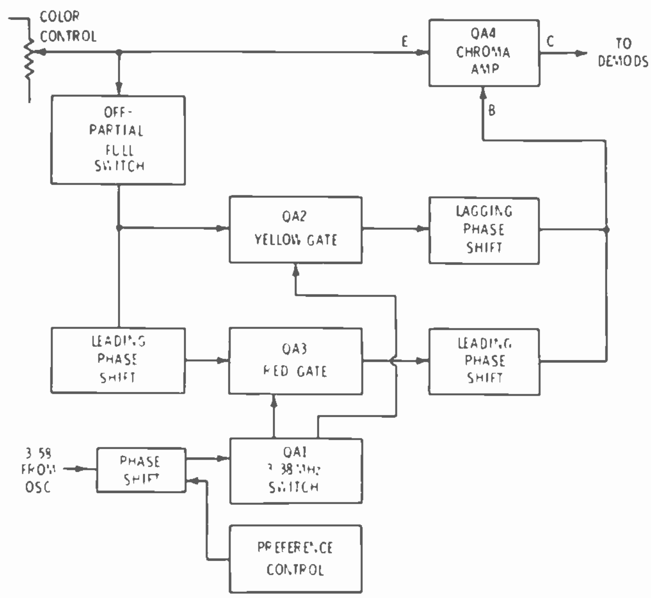


Fig. 1 (B) Block diagram of the ATC circuit.

component locations and a block diagram of the circuit are shown in Fig. 1.

The amount of ATC correction is selected manually by a three-position switch on the front panel which provides OFF (no correction), PARTIAL-correction and FULL-correction. The ATC circuit is inserted electronically between the moveable arm of the color control and the demodulators. There are two parallel paths for the chroma signal: one to the emitter of QA4, the chroma amplifier transistor, that produces an amplified, but not phase-inverted, signal at the collector. The phase correction path is through the Red and Yellow gate transistors, with their combined outputs going to the base of QA4. Since the gate transistors invert the signal, and the signal at the base of QA4 is inverted again before appearing at the collector, both of these signal paths are in-phase at the collector of QA4.

Fig. 2 illustrates a color wheel that shows the hues obtained when a chroma signal of the phases listed is compared (in a demodulator circuit) with the phase of the burst signal. Magnavox has chosen a chroma signal with a phase of 57° to give the desired reddish-orange skin color. The basic action of the ATC circuit is to change the yellow and red signal phases (those on either side of orange) to 57° without changing the phases of green, cyan and blue. This is the reason for keying off the gate transistors during the time the green, cyan and blue hues are displayed on the television screen.

The entire circuit is shown in Fig. 3. Both gate transistors are

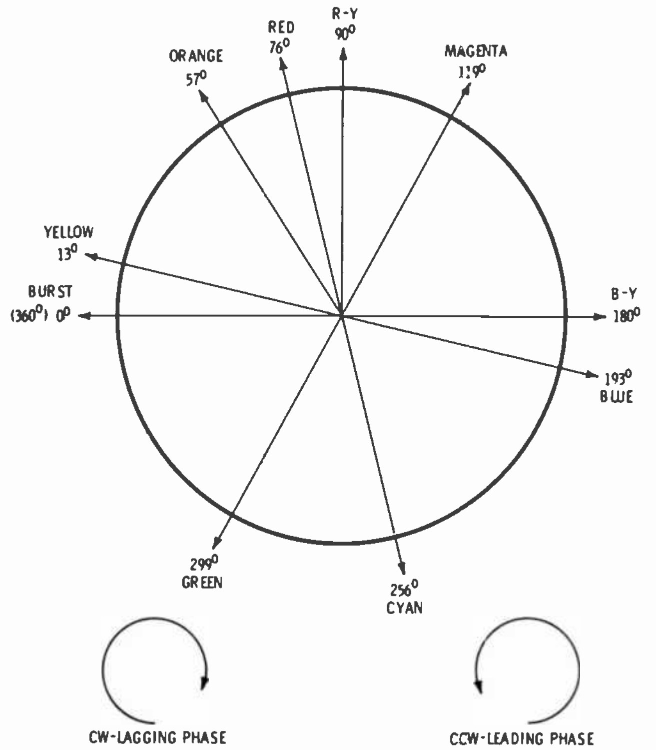


Fig. 2 A color wheel shows the phase (compared to burst) of the chroma signal necessary to obtain various hues.

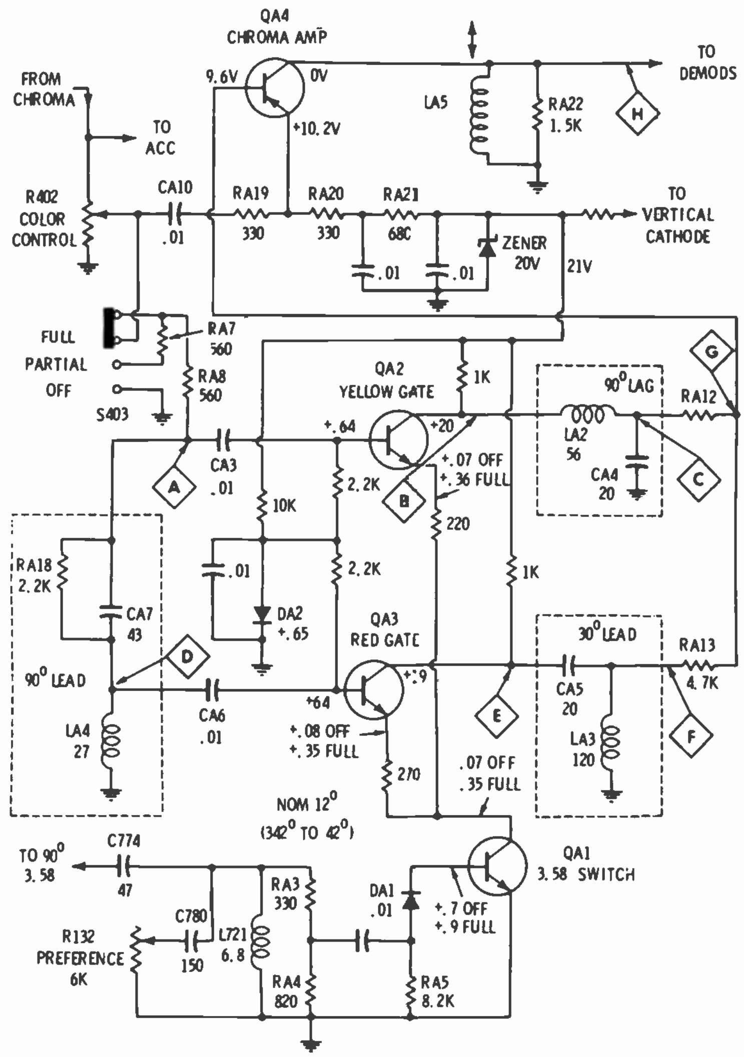
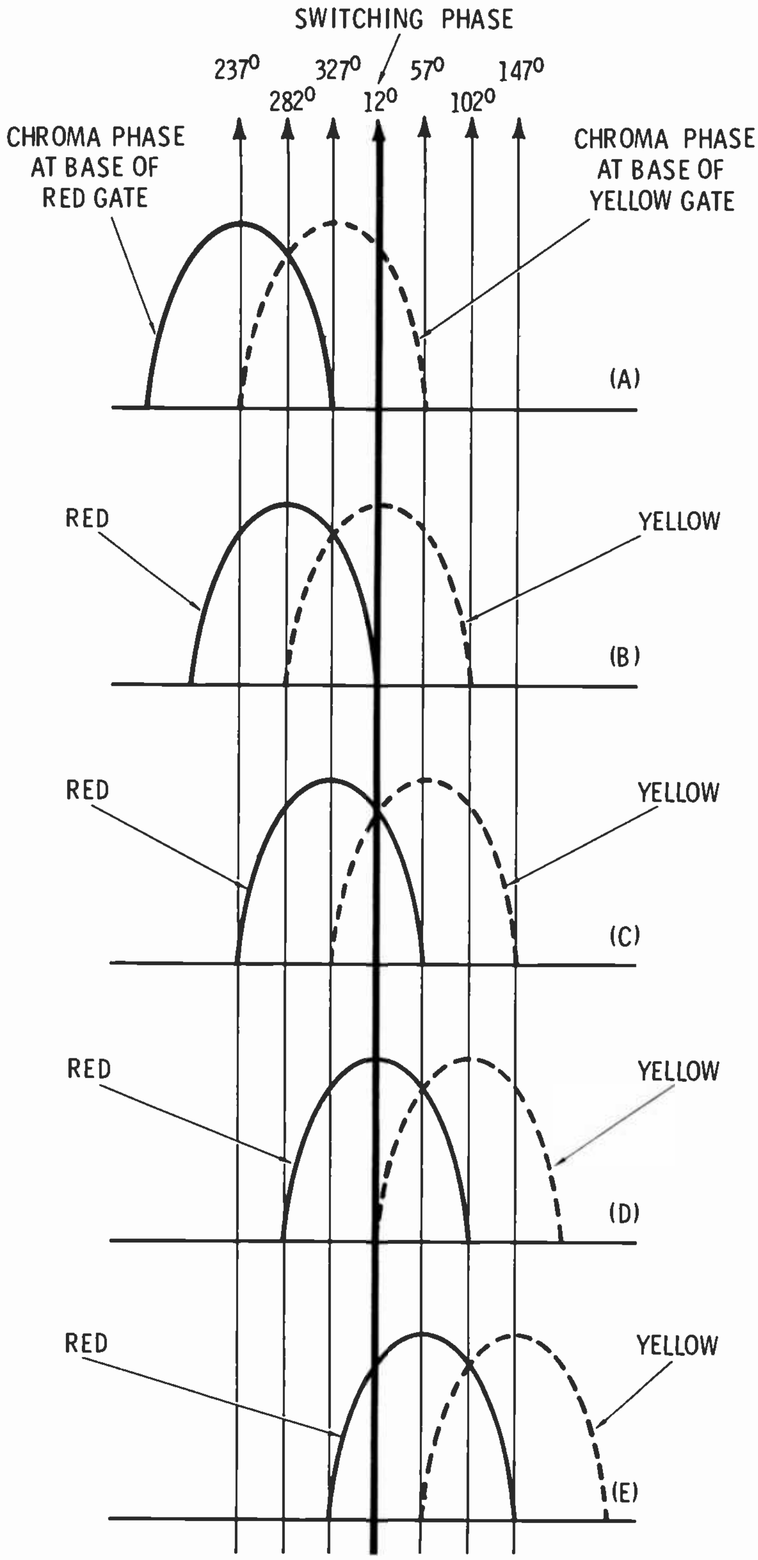


Fig. 3 The complete schematic of the Magnavox T940 ATC circuit.

non-conducting until two conditions are fulfilled. The gates are operated in Class "B", and, in the absence of an AC signal (chroma) at the bases, the forward bias is insufficient to allow any significant gain or collector current. Forward bias of just over .6 volt for the bases of both gates is developed across silicon diode DA2. Voltage from this source is temperature compensated, and any increase of base current in the silicon gate transistors resulting from a higher temperature will be reduced by a lower voltage drop across the diode.

In addition, the Red and Yellow gate emitters are returned to ground through the collector-emitter path of QA1, the switch transistor. The gates cannot conduct regardless of base



(A) Input phase is 327°, or greenish-yellow. There is a fairly large amplitude during the 12° sampling time at the base of the Yellow gate. This will give partial correction. The signal at the base of the Red gate is negative, so there will be no collector current.

(B) The input phase is 12° or yellow. Amplitude at the base of the Yellow gate is maximum, which will give full correction and make skin color orange. There is still no output from the Red gate since the base signal is zero.

(C) Input phase is 57°, or the desired orange skin color. Partial and equal signal voltages appear at each base. The resulting small corrections cancel out, leaving the output still at 57°.

(D) Input phase is 102°, or purple. The Yellow gate base signal is zero; therefore, there is no output. Input to the Red gate is at maximum and will give full correction to bring the skin color to orange.

(E) Input phase is 147°, or nearly blue. The voltage at the base of the Yellow gate is negative, so there will be no output. Enough signal appears at the base of the Red gate to give partial correction and bring the skin color back to bluish-red.

Fig. 4 The amplitude of the chroma sine wave at the base of each gate, during the time the switching transistor has full conduction, determines the collector current. Note that the ATC input signal phase is the same as the phase at the base of the Yellow gate.

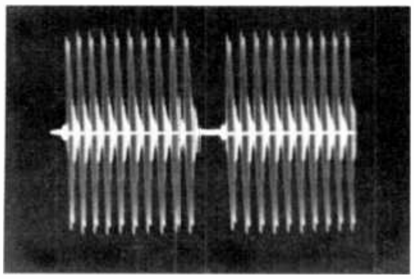
voltage until the switch transistor conducts. A 90° 3.58-MHz signal from the color oscillator transformer is shifted in phase by C774, C780 and L721. The preference control, R132, varies the phase angle 30° on either side of the nominal phase of 12°. This signal is rectified by diode DA1, and the resultant pulsating DC voltage of positive polarity is applied to the base of QA1, the switch transistor. The values are chosen so QA1 will draw current and be a virtual short circuit only during the very tip of the positive-going voltage applied to its base. During the rest of the cycle it is an open circuit, which disables the gates.

When the switching transistor is conducting and a positive chroma signal is present at the base of QA2, the Yellow gate, the base will be forward biased enough for collector current to flow and the transistor to amplify. The same conditions apply at the Red gate, QA3, except that the chroma signal applied to the base has a 90° leading phase compared to the phase at the yellow gate. Fig. 4 shows the positive-going halves of the chroma sine waves at the bases of the Red and Yellow gates during the time (12°) when the switch transistor is conducting. The output signal from each gate depends upon the instantaneous base voltage at the keying time, and **only** at this time. The more gate output signal, the more phase correction is possible.

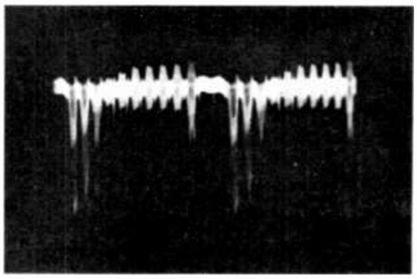
The gated signal at the collector of the Yellow gate is shifted by a 90° lag circuit consisting of LA2 and CA4. Similarly, the output of the Red gate is shifted to 30° (leading) by CA5 and LA3. The signal outputs from both gates are combined and applied to the base of QA4. After phase inversion this signal appears at the collector along with part of the original chroma signal from the input to the emitter. These two signals are never seen separately on a scope, for they add vectorially to become a sine wave of the resultant phase. From this point the phase-corrected chroma signal goes to the demodulators.

Fig. 5 shows the scope waveforms at the designated testpoints in the ATC cirucit when a gated color-bar generator is used as a signal source. The output tuning coil, LA5, is tuned by stray capacitance to form a low-Q circuit resonant to about 3.58 MHz. The function of the tuned circuit is to "ring" the clipped correction signal into a more symmetrical waveform.

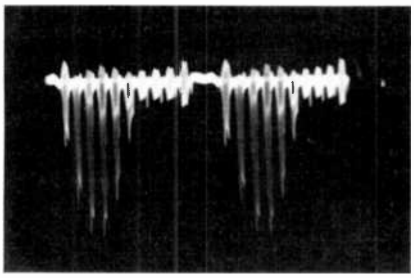
Vector diagrams are extremely important in helping us to fully understand the operation of this circuit. Fig. 6 shows the simplified schematic with phase shift components and testpoints, and vector diagrams showing the correction of a chroma signal of 30° leading phase (yellow skin hue) and one of 30° lagging phase (red skin hue). Here is how it works: With reference to Fig. 6B, the input signal phase is 27°. The Red gate has virtually no signal at its base during switching time (see Fig. 4), so it is non-conducting. The base of the Yellow gate has the same phase as the input (27°). After inversion in the transistor, the collector (testpoint B) phase is 207°. The 90° lagging circuit changes the phase to 297° at test points C and G of Fig. 6A. QA4 inverts the phase, so the phase at its collector (test point H) is 117°, which is added to the input phase of 27°, making a resultant signal of 57° for orange skin color. In the vector diagram in Fig. 6C, the 87° chroma signal at the Yellow gate is negative, making this gate inoperative. After the 90° lead change, the signal at the base (test point D) of the Red gate is 357°, and QA3 inverts the phase



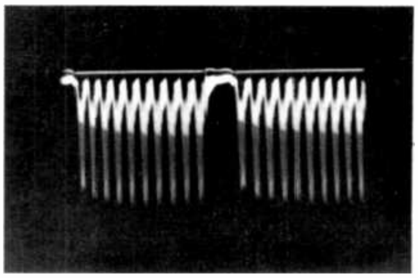
(A) The normal keyed-rainbow pulses at the input (test point A), also at the emitter of QA4, the chroma amplifier.



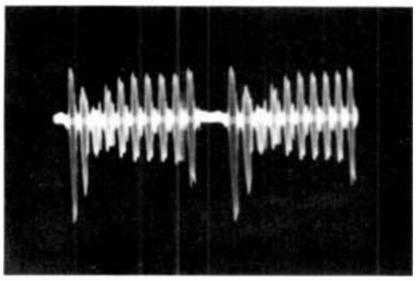
(B) Negative-going pulses at the collector of the Yellow gate (test point B). Maximum correction will be at color bar #1.



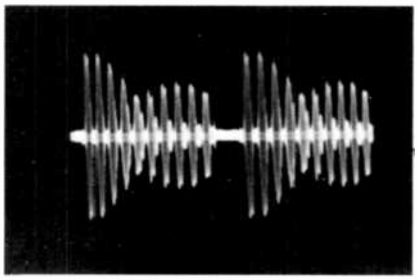
(C) Signal at the collector of the Red gate (test point E). Maximum correction will be at color bars 3 and 4.



(D) Negative-going "tails" on all ten bars at the collector of both the Red and Yellow gates is caused by a collector-emitter short in the switching transistor.



(E) Waveform at the base of QA4 (test point G) is the resultant of the output from the Red and Yellow gates.



(F) The signal at the collector of QA4 is the vectorial sum of the input signal (input at emitter) and a 180° inversion of the signal at the base (test point G). The larger amplitude of color bars 1, 2, and 3 is incidental, but it does make all the orange colors brighter.

Fig. 5 Chroma signal waveforms at each stage of the ATC circuit.

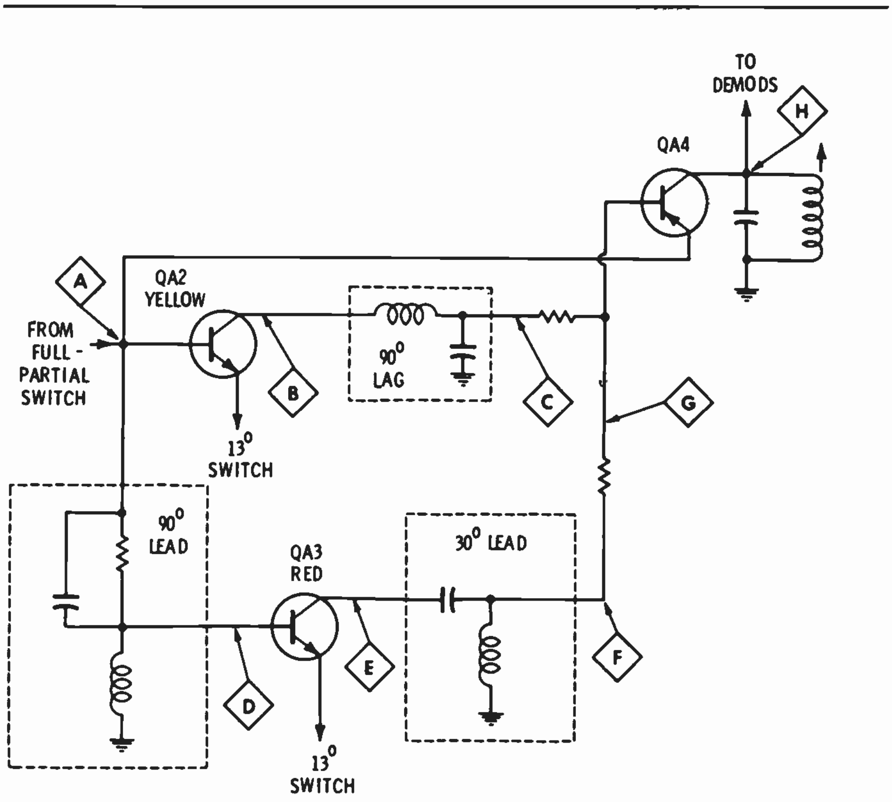
to 177° at the collector (test point E). The 30° leading network shifts the phase to 147° at test points F and G. After inversion by QA4, the phase at the collector (test point H) is 327°, which added to the input phase of 87° gives a resultant phase of 57° for orange skin color.

In the demodulators, phase differences become amplitude differences. The scope waveforms shown in Fig. 7 were taken from the picture tube grids, both with no ATC action and with full correction. In general, the pulses representing color bars become nearly identical for the first four bars. This statement is confirmed by color pictures of the color bar pattern on the screen of the picture tube. The large negative-going spike in each waveform is the horizontal blanking spike that comes from the –Y amplifiers whether any color is there or not.

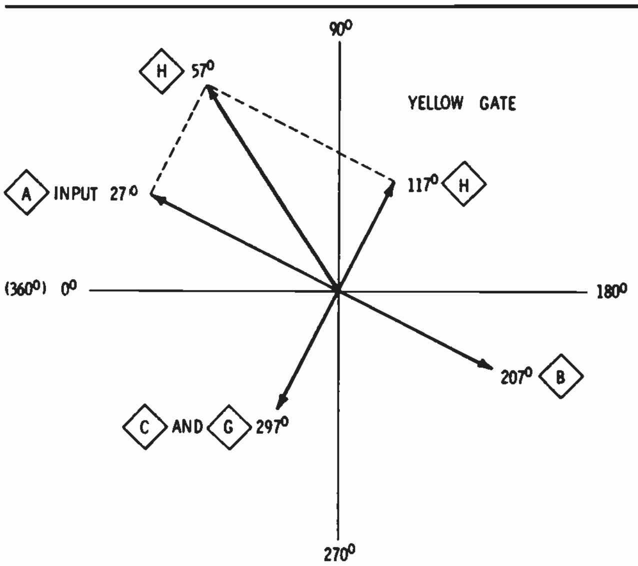
Vector patterns from the scope give the fastest and most accurate visualization of the ATC action, as shown in Fig. 8. With the ATC switch in the FULL correction position, the first three color bars have the same phase (57°), the fourth and tenth bars have some correction, and the other five bars are not affected.

Magnavox ACC

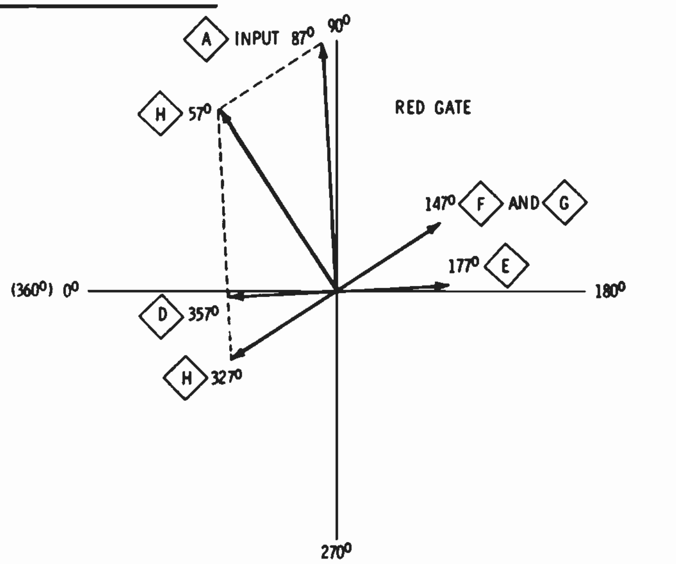
Another feature of the T940 Magnavox chassis is the ACC circuit which has a double action. As shown in the simplified schematic of Fig. 9, a total of four DC voltages are applied to the grid return of the chroma IF tube: 1) Negative voltage from the plate of the color killer is applied when the burst level is too low or missing. 2) Negative voltage from a killer detector diode is obtained through the 8.2-meg resistor. This voltage varies according to the amplitude of the burst signal. 3) A positive voltage applied through the 1.5-meg resistor from its own cathode cancels some of the negative voltage from the killer detector to cause more of the change in control voltage (from the burst) to reach the grid. 4) In addition to these conventional voltages, a variable positive voltage comes from QA5, the ACC transistor. Chroma voltage from the top of the color control is rectified by diode DA4, filtered by CA13 and applied to the base of QA5. The more positive the base voltage, the less positive the



(A) Simplified schematic with test points and phase shift components.

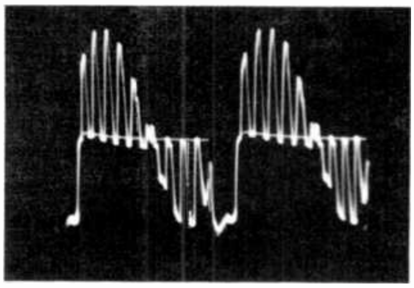


(B) Vectorgram for ATC correction of a 30° leading (27°) input chroma phase.

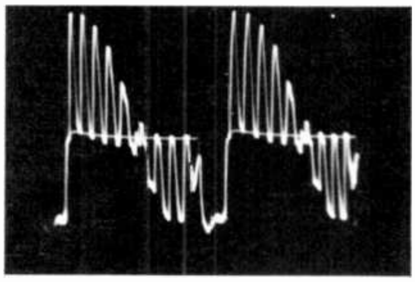


(C) Vectorgram for ATC correction of a 30° lagging (87°) input chroma phase.

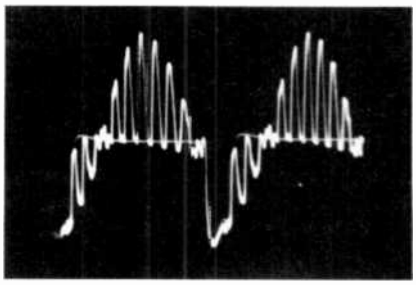
Fig. 8 Simplified schematic and vectorgrams for ATC correction.



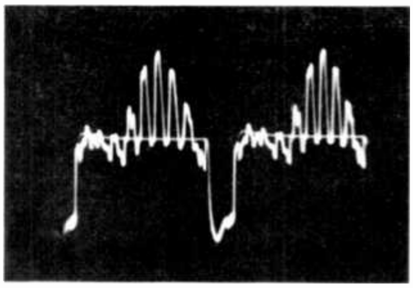
(A) Normal bar pattern at the red grid of the picture tube.



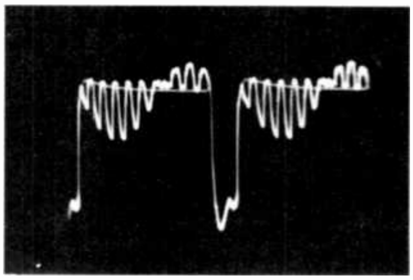
(B) Red grid with FULL ATC correction.



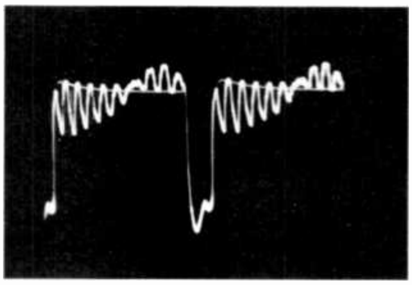
(C) Normal bar pattern at the blue grid of the picture tube.



(D) Blue grid with FULL ATC correction.



(E) Normal bar pattern at the green grid of the picture tube.



(F) Green grid with FULL ATC correction.

Fig. 7 Picture tube grid waveforms produced by a gated-rainbow pattern.

TABLE 1

DC voltage chart of the ACC circuit. The high chroma IF bias at 0% is due to color killer action. The last three lines are the chroma IB tube voltages without the additional bias from QA5.

ACC VOLTAGES

measuring point	(both ACC circuits working) chroma level at generator 0%	50%	100%	150%	200%
chroma IF cathode #7	3.6	9.2	7.0	6.0	5.6
chroma IF grid #2	-29	6.2	.5	-2.4	-3.5
chroma IF ACC bias	-32	-3	-6.5	-8.4	-9.1
QA5 collector	21	15	9	6.3	5
QA5 base	0	1.5	3.1	3.7	4.0
QA5 emitter	0	1.15	2.6	3.1	3.35
CR701A killer detector	-13	-28	-41	-46	-48
	(with QA5 base grounded)				
chroma IF cathode	3.5	9.2	8.6	8.0	7.7
chroma IF grid	-28	6.6	5.0	3.2	2.6
chroma IF ACC bias	-31	-2.6	-3.6	-4.8	-4.1

collector voltage, which is applied through a 1.5-meg resistor to the grid return of the chroma IF tube. This voltage also cancels out part of the negative voltage obtained from the killer detector, and since it is variable, makes the ACC voltage at the chroma IF tube more negative when the color is stronger.

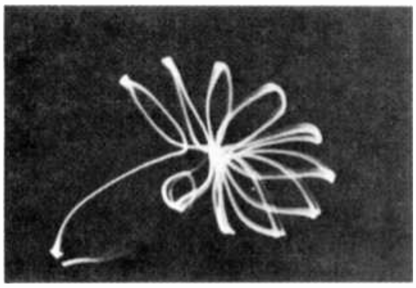
The DC voltage chart in Table 1 gives the important voltages in the ACC circuit. ACC voltages were obtained by comparing the chroma IF grid and cathode voltages measured to ground; this is the easiest way during troubleshooting. Slightly better accuracy can be obtained by measuring directly from cathode to the .047-mfd capacitor in the grid circuit. The difference in the voltage reading of the chroma IF ACC bias and the same reading with the base of QA5 grounded represents the added ACC gain correction obtained from the QA5 circuit. This extra control is very noticeable above chroma level of 75%.

You might think that obtaining part of the ACC from the amplitude of the chroma signal, rather than the burst amplitude alone, would defeat the natural color saturation in scenes having bright colors or others with little color. While there is some of this effect, it is overshadowed by the minimizing of another common problem: The many times a station will broadcast extremely strong or abnormally weak color without any corresponding change in the burst.

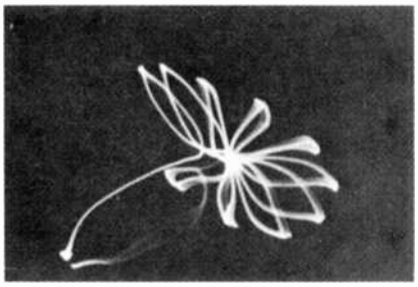
Troubleshooting the ATC Circuit

The first step in analyzing any ATC malfunction is to try the normal sequence of customer adjustments. Slide the ATC switch (on the front panel) to the OFF position, and with a color picture tuned in, adjust the tint control for normal skin color. If it is impossible to obtain good skin color, the ATC circuit is not at fault. If the picture has normal tint and saturation, QA4, the chroma amplifier on the ATC board is working and the circuit has B+ voltage. Weak or missing color can be caused by the color IF, video IF's, killer detector, 3.58-MHz oscillator, etc., the same as in any color TV. Connect a jumper wire from the top of the color control to the demodulator grid, pin 7; if the color improves, the ATC circuitry is at fault.

Slide the ATC switch to the FULL correction position. The preference control (with other controls on the back near the top) should change skin hues from greenish-yellow to magenta. If only magenta skin hues are seen, the Red gate may be defective; conversely, if the skin colors are greenish-yellow, the Yellow gate may not be working.

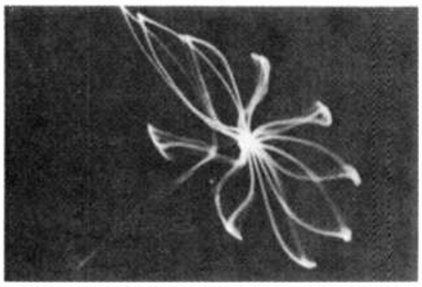


(A) Normal vector pattern with the third "petal" at 90°.

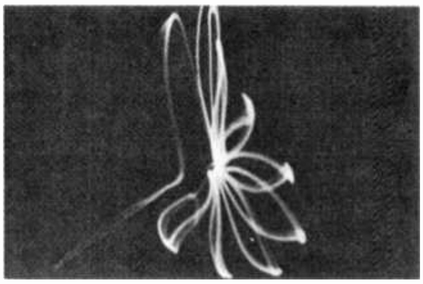


(B) ATC switch in PARTIAL position pulls the first and third petals nearer to bar two.

(C) With the ATC switch in the FULL position, the first three petals are all at 57° (orange).



(D) FULL correction with the preference control adjusted for yellow faces.



(E) FULL correction with the preference control adjusted for magenta faces.

Fig. 8 Vector waveforms of color bars.

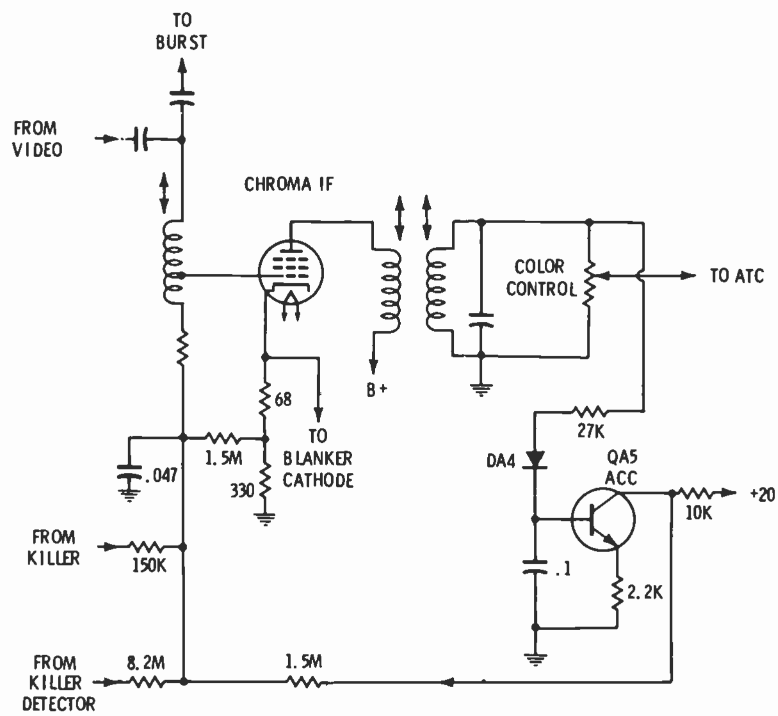


Fig. 9 Simplified schematic of the T940 ACC circuit. ACC is proportional to both burst level and chroma amplitude.

Defects in the gate circuits are best checked in the shop using voltage and waveform analysis to find the defective component. Use a gated-rainbow bar pattern and check for scope waveforms similar to those in Fig. 5.

Loss of the 3.58-MHz switching signal or an open or shorted QA1 switching transistor will eliminate any change in the color hues as the preference control is adjusted through its range. An open switch transistor will eliminate all gate action, and there will be no change in the color when either the preference or the ATC switch is adjusted. A shorted switch transistor will allow both gates to conduct at all times; the preference control will have no effect, but switching the ATC on FULL will brighten all ten color bars. A loss of 3.58-MHz signal to the switching transistor will give the same symptoms as an open transistor.

Conclusion

The Magnavox ATC circuit actually functions precisely as explained here. The action is strictly by phase changes (with a minor amplitude side-effect) and, therefore, is instantaneous in action without time lag, locking or registration effects. There are no adjustments to be made on the board, so it is not necessary to "tune-up" anything. When you are accustomed to the sight of one (never more than two) completely red bar on a gated-rainbow color display, it gives one a peculiar feeling to see four (sometimes nearly five) reddish-orange bars on the screen. Obviously, it is four times less critical of skin color than an uncorrected signal.

One small drawback is inherent in this type of phase correction: red becomes orange and greenish-yellow becomes orange regardless of whether these hues are applied to a face or some other object in the picture. This is the reason for the FULL and PARTIAL positions on the ATC switch. Usually the PARTIAL correction would be used where the variation in skin color is not too extreme. Any change in areas of the picture other than skin hues would be minimized.

All factors considered, this ATC circuit is a fascinating addition to modern color TV engineering.

RCA's Accu-Tint

The RCA **Accu-Tint** (A-T) system produces three basic changes when the A-T switch is turned to ON:

- The phase of the 3.58-MHz color subcarrier applied to the B-Y chroma demodulator circuit is changed so that it is nearer the phase of the subcarrier supplied to the R-Y demodulator.

- The output of the B-Y demodulator circuit is reduced about 33 percent.

- The screen color is changed from the normal blue-white to a brown-white, or sepia. (This "warming" of the screen color "temperature" occurs only when a colorcast is received (killer inoperative) and the A-T switch is ON.)

The result of these A-T actions is to increase the level of red and decrease the level of blue and green in the color picture.

Fig. 10 shows the demodulator and —Y amplifier circuits of the new RCA CTC39X chassis, which is similar to the CTC38X except for minor changes in the video amplifiers, and the addition of the A-T circuit.

B-Y Phase Change

The 3.58-MHz subcarrier for the R-Y demodulator is taken directly from the secondary of T703, while the phase of the 3.58-MHz carrier for the B-Y demodulator is made "leading" by a high-pass filter. When the A-T switch, S106, is in the OFF position, this high-pass filter consists of C732, L703 and R735. With this arrangement, the normal (A-T off) B-Y phase leads the R-Y phase by about 105 degrees.

When the A-T switch is turned to ON, L712 and R797 are added in series with L703 and R735. This reduces the lead of the B-Y phase to 90 degrees or less (RCA has not announced the exact figures). R-Y carrier phase is changed slightly by the difference in loading on the secondary of T703, but most of the shift is in the phase of the B-Y chroma subcarrier.

B-Y Amplitude At The CRT Blue Grid

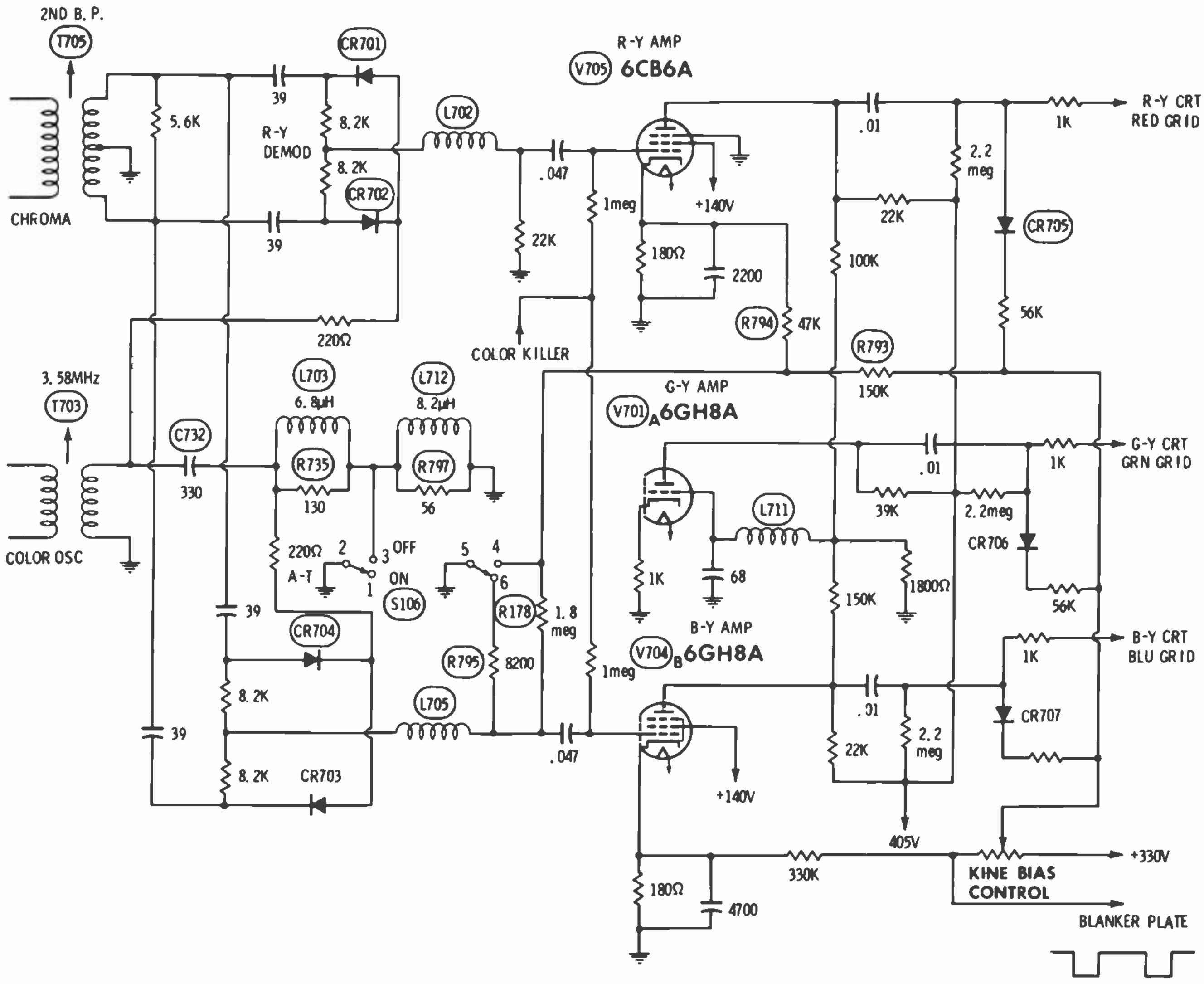


Fig. 10 Complete schematic of the demodulators and color-difference amplifiers in the new RCA CTC39X chassis. Notice especially the circuits connected to switch S106, the Accu-Tint OFF/ON switch.

Skin coloring that is acceptable to many people can be obtained with only an R-Y color signal, so logically the next step is to reduce the amount of blue and green in the color picture.

The second section of S106 performs a double function. One of these, when the A-T is ON, is to ground R795. This decreases the output of the B-Y demodulator so that the signal at the CRT blue grid is reduced from the normal 120 percent of red to about 80 percent of red.

R-Y amplitude is unchanged, but because G-Y is made from R-Y plus B-Y, it also will be reduced.

Change in CRT Screen Temperature

When the A-T is ON, the junction of R793, R794 and R178 no longer is grounded through terminal 4 of S106. The negative-going horizontal pulse originating at the center lug of the **kine bias** control is reduced by a voltage divider consisting of R793, R178 and R795, and then is applied through the .047-mfd coupling capacitor to the grid of V704B, the B-Y amplifier. After amplification and phase reversal (180 degrees) by the tube, it becomes a small, positive-going pulse at the plate, which is fed through the .01-mfd capacitor to the anode of diode CR707 and the blue grid of the CRT. On the cathode of CR707, a large negative-going pulse from the kine bias control is applied to the cathode of CR707 and turns it on, resetting the charge on the .01-mfd capacitor every cycle, and thus maintaining the CRT grid DC voltage constant regardless of the chroma waveform.

The positive-going pulse at the anode of CR707 and the negative-going pulse at the cathode **add** together to produce a larger pulse. Because the DC voltage on the CRT grid is "clamped" to the negative tip of the pulse, the DC voltage on the CRT control grid is made less positive (compared to the CRT cathode), and the current of the CRT blue gun is reduced.

A small pulse also is applied through R793 and R794 to the cathode of V705, the R-Y amplifier. Since the signal is applied to the cathode, this pulse is amplified without phase inversion, and the negative-going pulse of a few volts appears at the anode of diode CR705.

A large negative-going pulse from the kine bias control is present at the cathode of CR705. The true voltage across the diode is the difference between the pulse voltage on the anode and the pulse voltage on the cathode. The effect is the same as a decrease in the pulse from the kine bias control, and causes the red grid of the CRT to become more positive by a few volts, increasing the red gun current.

During normal operation, small samples from the plates of both the R-Y and B-Y amplifiers are matrixed to produce a —(G-Y) signal at the grid of the G-Y amplifier. Because of phase inversion across the tube, a G-Y signal is developed at the plate and applied to the green grid of the CRT. Voltages in the G-Y stage are not affected by the addition of the extra A-T pulse voltages at the R-Y and B-Y amplifier plates; because the polarity of one is positive-going and the other is negative-going and the amplitudes of the two signals are nearly equal, they cancel at the G-Y amplifier grid.

This method of using pulses to brighten the red, dim the blue, and leave green unchanged has another beneficial effect: When the color killer operates during b-w programs, the R-Y and B-Y amplifiers are biased to cutoff. Because of this, the A-T pulses cannot be amplified and, therefore, do not appear at the plates of the tubes or the anodes of the DC restorer diodes. Thus, the screen color is a normal blue-white, not sepia. (Of course, if the station leaves their color carrier on during a b-w commercial, or if the color killer malfunctions and does not operate normally during b-w programs, the screen color will be sepia if the A-T switch is in the ON position.)

Turning off the Accu-Tint circuit restores the B-Y demodulator phasing and amplitude, and restores the screen color to normal.

Conclusion

This RCA circuit prevents unwanted changes in skin coloration when the receiver is switched from channel to channel and/or when programing changes. Changes to the previous circuit are not extensive, and no critical or trouble-prone components have been used. Consequently, the circuit should be relatively service-free.

However, we must recognize that, despite excellent engineering and modern components, each basic type of automatic tint modifier is a stopgap emergency measure made desirable only by the continuing delay of the broadcasting industry in standardizing color hue. Also, it should be remembered that the circuits which most effectively eliminate green or purple faces also produce the most distortion of colors (except orange). ▲

Chapter 9

TV remote control systems

The only motor in RCA's remotely controlled CTC47 color TV chassis is the UHF tuning motor; remote control of VHF channel selection, the on/off function and adjustment of volume, tint and color saturation is accomplished electronically without motors.

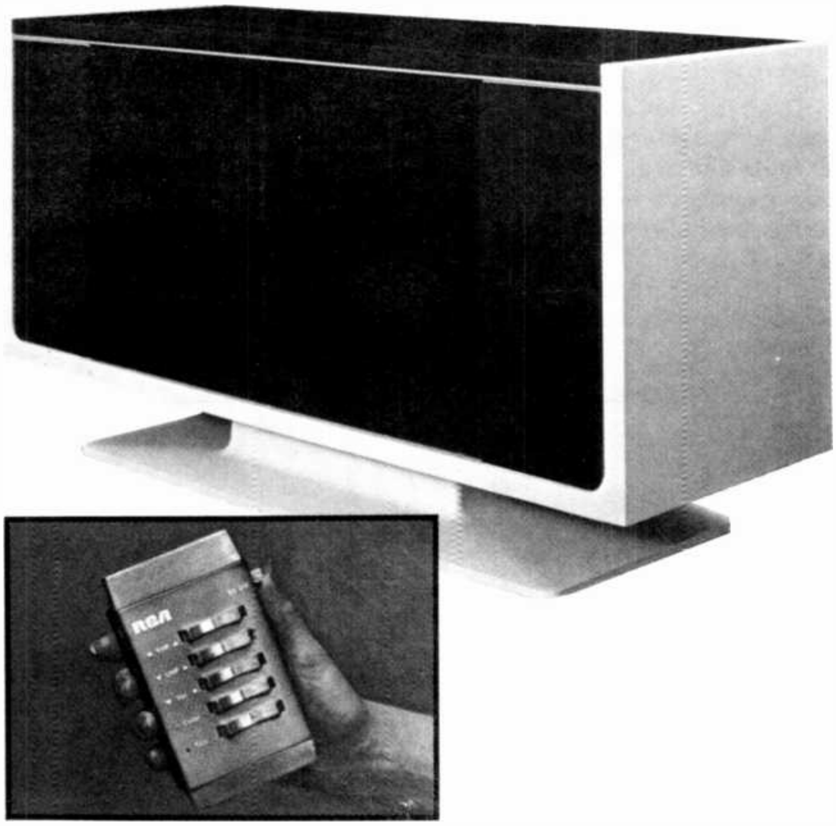
How the volume, tint and color saturation remote control circuits operate will be explained in this article, along with techniques for tracking down troubles in them.

How Functions Are Controlled

One of the basic differences between motor-controlled and motorless-controlled is the way in which volume, tint and color are made to vary. Because it is the most simple, let's start with volume control.

Fig. 1 illustrates the conventional method of controlling volume. The circuit itself is so simple that it requires no explanation; the point is that **something has to turn the shaft of the potentiometer.**

Now let's explore some alternate methods of controlling volume. If we were dealing with vacuum-tube circuits, we might use an audio-amplifier tube with remote-cutoff characteristics, and control its gain by varying the bias. This is similar to controlling the gain of an IF amplifier by varying the AGC bias applied to it. However, a couple of problems arise: First, remote-cutoff tubes tend to produce distortion



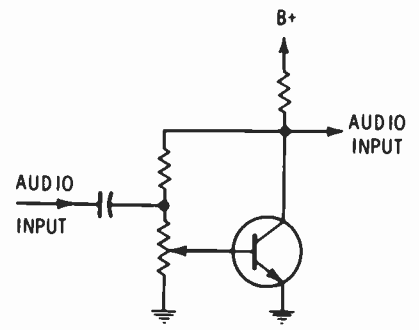


Fig. 1 Conventional volume control and audio amplifier

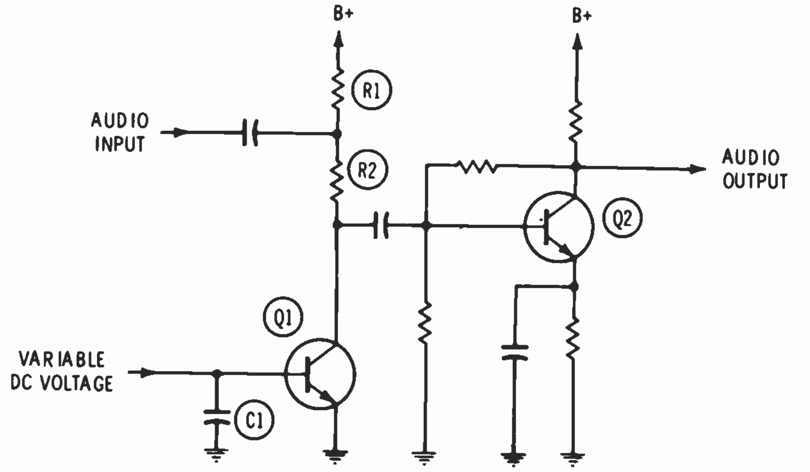


Fig. 2 Audio amplifier with DC volume control.

when they are used with fairly high-level signals, such as those from the sound FM detector of a TV receiver; and second, tubes are no longer used in most television audio sections.

Controlling the volume by varying the bias of an audio-amplifier transistor also can lead to distortion problems, just as in vacuum-tube amplifiers.

There is, however, another approach: The use of a transistor as a signal-shunting device. The fundamental circuit is shown in Fig. 2. If the base-bias voltage of Q1 is zero, Q1 appears as an open circuit from collector to ground, there is no attenuation of the signal passing to Q2, and the output volume is maximum. However, as Q1 is biased into conduction, the impedance from collector to ground decreases, reaching nearly zero when Q1 is saturated. As the collector impedance of Q1 decreases, more and more signal is dropped across R2, and the volume is progressively reduced. If the supply voltage for Q1 is large in comparison to the signal, the instantaneous changes in collector voltage, caused by the signal, will not have much effect on the impedance to ground, and significant distortion will not be produced.

This volume-control configuration actually has two advantages: One of these applies regardless of whether or not wireless remote control is anticipated. Because the audio signal itself does not need to be carried via wiring to the controlled stage, pick-up of hum and other spurious signals, which often occurs when a volume control is physically separated from the amplifiers, does not occur. Any stray AC voltages which might induce a spurious signal on the conductor leading to the base of Q1 will be shunted by C1, whose capacitance can be as large as necessary.

The second advantage is related to the first. Because the proposed remote-control system now needs to generate only a DC voltage, there is no need for a mechanical device.

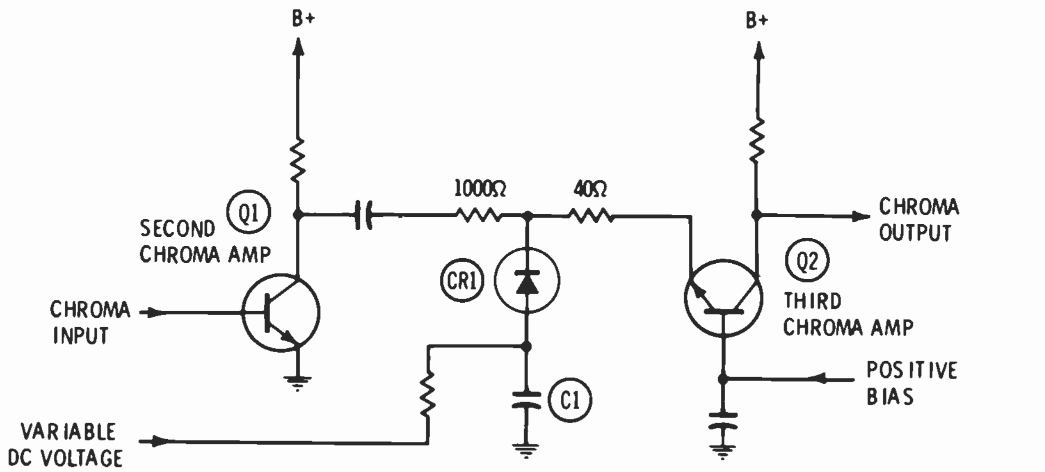


Fig. 3 Gain-controlled chroma amplifier using DC control voltage.

Obviously, the circuit in Fig. 2 could be adapted to control color level instead of volume, since the functions are essentially the same.

In the RCA CTC47, the volume-control circuit is similar to Fig. 2, but the active devices are contained in an integrated circuit (IC).

A simplified version of the color-control circuit appears in Fig. 3. Diode CR1 is the signal-shunt device in this circuit. As the variable bias is made more positive, the impedance of CR1 decreases, again shunting more signal to ground and allowing less to pass to the 3rd chroma amplifier. As with the volume control circuit, there is no signal on the control line and spurious signals which might be picked up are grounded by C1.

Remote control of tint is a bit more complicated because the phase of a signal must be controlled, rather than its amplitude. A convenient way of doing this is to split the reference signal into two signals having different phases, separately control their relative amplitudes, and then recombine them.

In Fig. 4, reference signals from the 3.58-MHz oscillator drive the bases of Q1 and Q2 in phase; however, the RL emitter impedance of Q1 and the RC emitter impedance of Q2 cause the phases of the collector signals to be about 90 degrees from the input phase, respectively. Assuming the transistors are biased equally, each will supply half of the signal energy to the output transformer, and the combination of these two signals will produce an output in phase with the inputs to Q1 and Q2. (If the transformer secondary leads were reversed, the output signal would be 180 degrees out of phase with the input to Q1 and Q2.)

Now, suppose the variable bias is made more positive. Q2 will amplify more than before, and the phase of the output signal approaches the phase present at the emitter of Q2. At the same time, the increased current through Q2 increases the voltage drop across R1, biasing Q1 toward cutoff. If the variable bias is made less positive, the gain of Q2 will decrease, the gain of Q1 will increase, and the phase of the output will shift in the opposite direction.

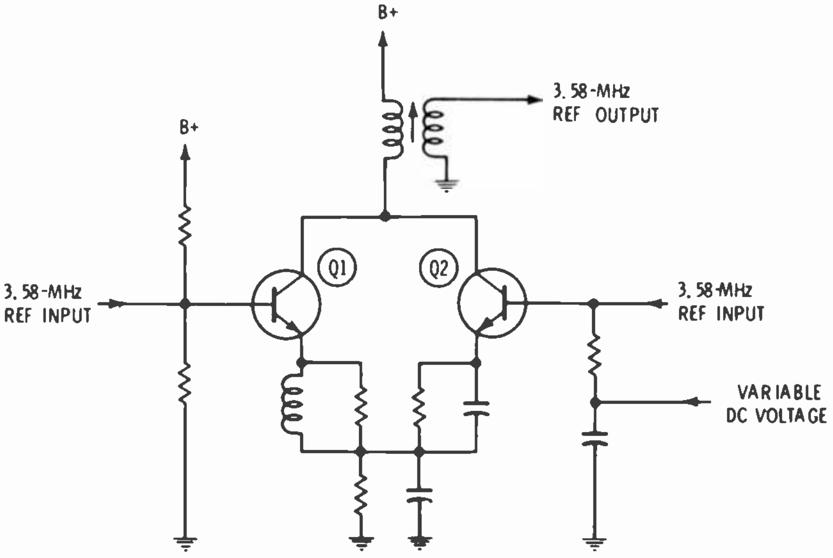


Fig. 4 DC-controlled tint circuit.

Memory Circuits

In a motor-driven remote-control system, the problem of memory does not arise, because once the motor comes to rest at the desired position, it will remain there. In a motorless system, the problem of memory becomes serious.

The circuit in Fig. 5, which has no memory capability, illustrates the point. Suppose the relay is closed by means of a hand transmitter and remote receiver. If R1 and C1 are very large, the voltage fed to the volume (or tint, or color) circuit will increase quite slowly, causing the volume to increase also. When the volume reaches a pleasant level, the operator releases the button and settles back to enjoy the program; however, C1 will gradually discharge through Q1 (of Fig. 2), and the volume again will increase back to maximum.

If discharge of C1 can be prevented, the volume will remain constant, once set, and memory will have been achieved. In Fig. 6, the basic configuration of a memory circuit is illustrated. In this circuit, the capacitance between gate and channel of the MOSFET, Q1, will charge towards B+ any time the contacts of K1 are closed. However, since the input capacitance of Q1 is about 5 pf, a charging resistance of several thousand megohms would be necessary to make the charging time long enough for a person to control it. To increase the charging time, C1 (about 1 mfd) is connected to ground.

Once the relay contacts are opened, there is no discharge path for C1 or the input capacitance of Q1, except through the leakage resistance of C1 and the MOSFET gate; this resistance can be made as high as several thousand megohms, and the time constant thus produced is in excess of several days, perhaps months. As we shall see later, this time can be extended even further. Q1 and Q2 serve as drain- and emitter-followers, respectively, to develop sufficient current-handling capacitance for the controlled circuit.

In addition to remote control, normally it is desirable to allow control of volume, tint and color at the receiver itself. One method of modifying the circuit of Fig. 6 to allow local control is to simply connect pushbutton switches across the "up" and "down" relay contacts. The use of remote controls which require that the button be depressed long enough for the function to change are acceptable, but people generally object to similar "time-consuming" controls located on the receiver itself. For some reason, a potentiometer type of control on the receiver is more pleasing. Also, as we shall see in the circuits described below, use of the potentiometer in the receiver improves the memory function.

In Fig. 7, C1 is returned to the wiper of a local-control potentiometer instead of ground; otherwise the circuit is the same as Fig. 6. However, in Fig. 7 the voltage on the gate of the MOSFET **will be almost exactly the same as the voltage at the potentiometer wiper.** This can be understood better if we consider C1 and the input capacitance as a voltage divider connected from the potentiometer to ground. Because the voltages across series capacitors are inversely proportional to their capacitances, and the value of C1 is about 200,000 times the input capacitance of Q1, essentially **all** the applied voltage is present on the MOSFET gate.

Assume that R1 in Fig. 7 has been set to produce normal volume, and +3 volts exists on the MOSFET gate. By actuating the remote "volume down" switch, the appropriate relay contact is closed and both C1 and the MOSFET input capacitance are charged to some higher voltage—in this example, 4 volts. Since the voltage across C1 is only 1 volt, instead of 4 volts as it would be in Fig. 6, the leakage current is one fourth as great, and memory is increased by a factor of four. (Since the gate resistance is almost infinite, the leakage current through it can be ignored.) Furthermore, even if C1 leaks significantly, the volume will return to the original level determined by R1 instead of to maximum, to which it would return in Fig. 6.

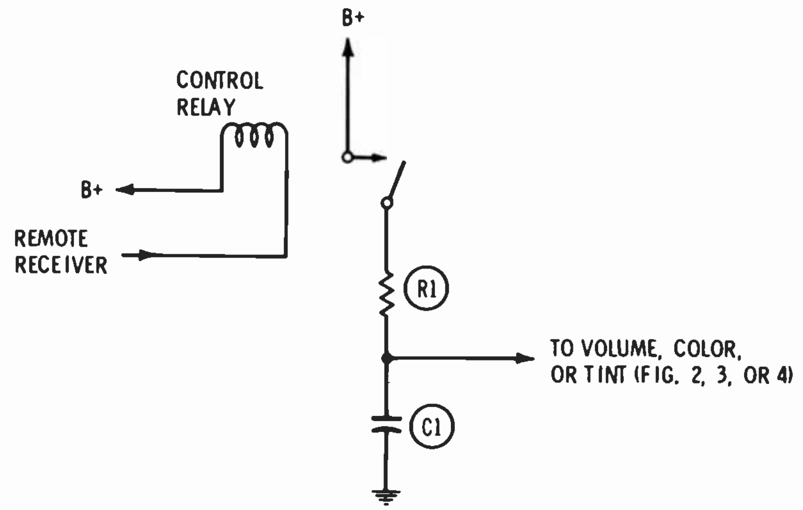


Fig. 5 Remote-control voltage source.

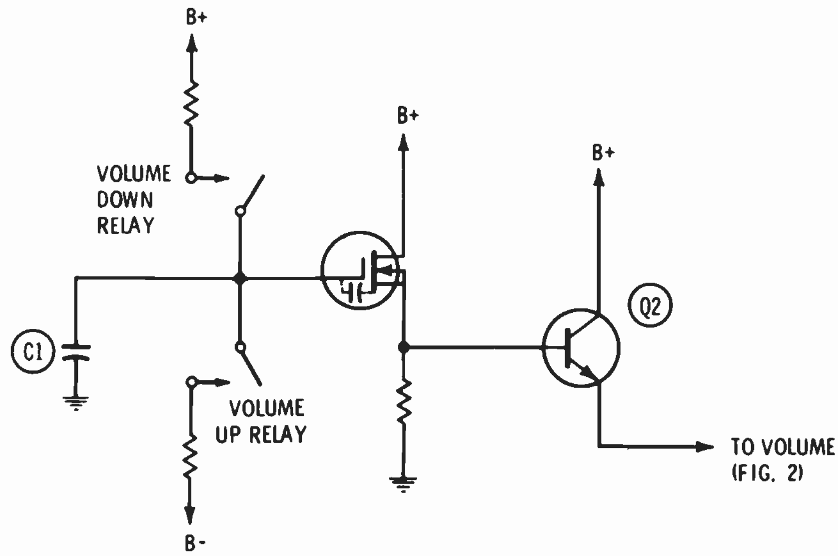


Fig. 6 Basic memory circuit and volume-control driver.

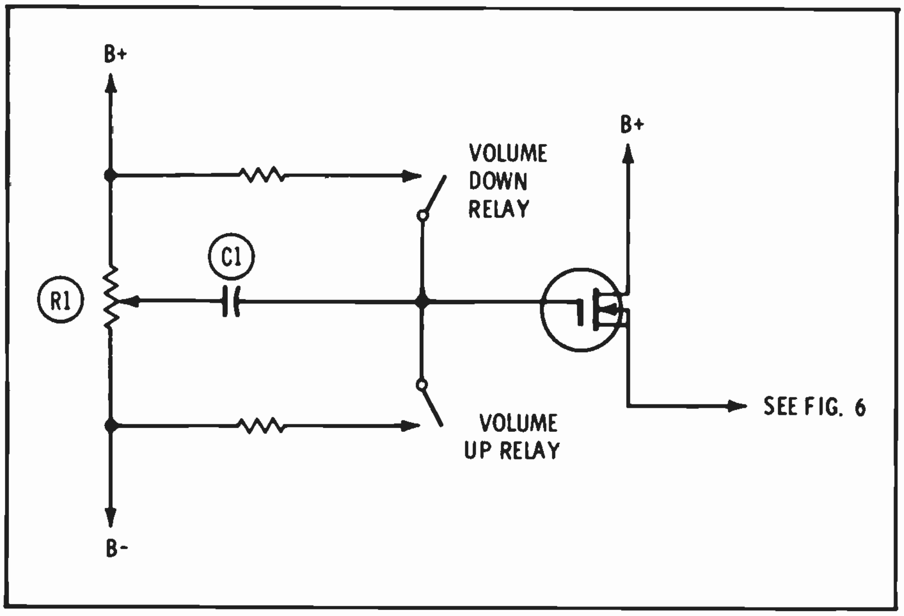


Fig. 7 Memory control circuit with local and remote inputs.

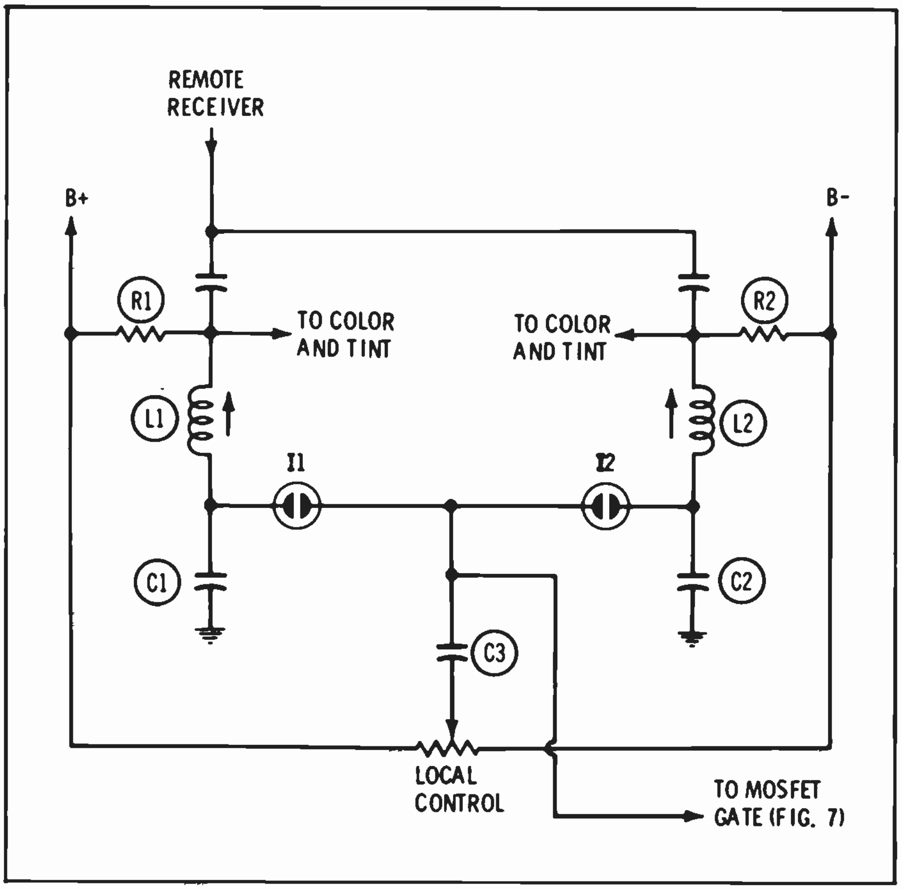


Fig. 8 Memory circuit input using electronic switches.

Electronic Switching

So far we have assumed relays were used to connect the voltage to the MOSFET. In the actual design, the relays have been replaced by electronic switches, as illustrated in Fig. 8.

When the "volume down" remote button on the transmitter is depressed, the transmitted frequency picked up by the remote receiver is the same as the resonant frequency of L1 and C1. At this frequency, the impedance of L1 and C1 is near zero and the current is large. The voltage across C1 is high, perhaps 200 volts, because it is equal to I (current) times Xc (capacitive reactance), and this ignites the neon lamp, I1. Once ignited, the resistance of I1 is low, and DC voltage from the B+ supply charges the MOSFET gate through R1, as already described.

One final circuit refinement is required, although it is not immediately apparent. If no means were provided to discharge C3 back to zero volts, enough voltage eventually would accumulate on C3, by virtue of the remote control, to make local control impossible. For example, suppose the local control originally had been set for a comfortable volume, and remote control subsequently was used to increase the volume. Then suppose local control were used to reduce volume, and remote control were used to raise it once more. After this had been repeated a number of times, the local-control potentiometer would be at its limit and volume still would be high.

To prevent this, the local-control potentiometer is constructed with a "delta switch", which closes momentarily each time the shaft is turned, even a few degrees. The delta switch energizes a relay whose contacts are connected across C3. Thus, each time the local control is used, all voltage resulting from remote control is cancelled, or discharged to the potentiometer wiper. With the addition of the delta switch, either remote or local control can be used repeatedly in any sequence without a "hang-up" resulting. ▲

Chapter 10

New in Color TV for 1970

General Trends

Innovations in solid-state circuitry highlight the 1970 TV designs, while very few changes are evident in tube-powered chassis. Hybrid receivers are numerous, with all the circuits transistorized except for vertical sweep, horizontal sweep, high voltage, video output and chroma −Y amplifiers.

Color TV continues to be the center of attention, with more color portables and more solid-state circuits. Field-effect transistors, spark gaps inside the CRT sockets, more plug-in boards or modules, and the beginnings of a trend to pre-CRT matrixing of chroma and video signals are just a few items of interest.

Emphasized in the manufacturers' service data are such safety precautions as high-voltage adjustments and the measurement of line-voltage leakage from exposed receiver parts to earth ground. High-voltage shunt regulators of the 6BK4 type are not used in many of the new receiver designs, as the manufacturers remain concerned about possible radiation hazards and more stringent government standards in the future.

Here are some of the most interesting 1970 features and circuits of the new color television receivers, with the manufacturers listed in alphabetical order:

Admiral

Only seven tubes, plus the picture tube, are used in the Admiral K10 chassis, a hybrid design found in their 12″, 14″ and 16″ portable color receivers. Horizontal sweep, vertical sweep, high-voltage, video output and −Y chroma stages employ tubes. All other functions utilize solid-state components.

Fig. 1 shows the schematic of the automatic degaussing circuit. Don't operate this chassis without a substitute load on the degaussing circuit; such a load can be a 5-ohm, 3-watt resistor, which is substituted for the coil during bench tests. Full degaussing is completed in a fraction of a second by the charging currents of filter capacitors CH8 and CH10A. The picture tube would be magnetized by the steady current drawn by the tubes after they heat and become conductive; therefore, the degaussing coil is shorted out before this time by a thermally operated switch whose heat is supplied by an internal element connected to the 6.3-volt winding on the power transformer.

A ratio detector is used for sound demodulation, and better sound limiting is accomplished by the final sound IF stage, which is designed to oscillate. The sound IF signal applied to the input of this stage acts as a sync signal to lock the frequency of oscillation. So long as there is enough sound IF signal to make the oscillator lock to it, the amplitude of the signal applied to the ratio detector will be constant.

The burst signal is usually taken off prior to the stage that is controlled by the color killer, because the burst must be passed regardless of the color control setting or the killer action. The Admiral K10 chassis is an exception to this usual design. Fig. 2 is a simplified schematic of the color killer and first color IF amplifier. When burst is present at the killer phase detector, there is zero voltage output from the detector to the base of Q16, which has no forward bias and does not conduct. The voltage at the collector is an amount determined by the voltage divider that supplies the base of Q13, the first color IF amplifier. Normal bias from this source is supplied to the base of Q13, which amplifies the chroma signal, including the burst.

During b-w reception, the output from the killer detector is about +0.6 volt. This is nearly normal forward bias and causes the killer amplifier, Q16, to draw collector current, which reduces the collector voltage to about 5 or 6 volts. The forward bias at the base of Q13 is reduced to about .2 volt (measured from emitter to base) and Q13 is cut off.

With the burst signal obtained from the collector of the stage (Q13) controlled by the color killer, it is apparent that without some other action the 1st color amplifier would remain cut off during color broadcasts and no burst would be passed to the color killer circuitry to trigger on the 1st color amplifier. (Note the closed-loop action described here). However, to prevent such a situation and to insure that the 1st color amplifier passes the burst signal, it is keyed on during burst time by a horizontal pulse. (Remember, the burst signal is positioned on the "back porch" of the horizontal blanking pulse.) Admiral calls this action "burst assurance" and it functions in the following manner: Before a pulse is applied to the anodes of diodes CRC19 and CRC32, both diodes are reverse-biased by the positive voltage on their cathodes, and are open circuits as a result. When the positive-going pulse at the anode of CRC19 exceeds the DC voltage at the cathode, the diode becomes a short circuit and allows the rest of the pulse to temporarily increase the forward bias of Q13 to the point where it conducts. If burst is being received at this time, it will be amplified. CRC32 is a pulse clipper that prevents the pulse from ever exceeding about 6.6 volts positive. When the pulse tries to rise above the 6.5 volts (plus a drop of about .1 volt across the diode), the diode is forward biased and connects the anode with the pulse to the +6.5-volt DC source. Therefore, Q13 is always normally biased at the time of burst, regardless of the color killer action.

Normal transistors do not perform well as reactance control devices. In the Admiral K10 chassis, 3.58-MHz oscillator frequency control is accomplished by using a field-effect transistor (FET) for a reactance control stage. The circuit, shown in Fig. 3, is nearly identical with ones that use tubes, except that the source voltage is varied to adjust the frequency instead of using a reactance coil.

Andrea

The Andrea VCX325 color TV chassis is patterned after the standard three-tube-IF design and has solid-state sound. A tuning eye (schematic in Fig. 4) is used to aid accurate fine tuning. The indicator shows when the picture carrier is tuned to 45.75 MHz.

Another rarity is an extra video circuit, evidently included to feed an external video tape recorder (VTR). This circuit, shown in Fig. 5, employs two emitter followers in cascade (Darlington), with no peaking coils or other compensation.

General Electric

A novel focus-tracking circuit (Fig.6) is a feature of the GE C-1 chassis. (This chassis is used in a hybrid 18" diagonal portable color receiver.) The tuner, AGC, sync, horizontal reactance, horizontal oscillator and horizontal discharge circuits are transistorized. Two transistor amplifiers and one blanker transistor are used in the video circuit, which has a tube-equipped output stage. The chroma section has one transistor, which is used as a 3.58-MHz buffer.

Most focus circuits add the B-boost to the rectified DC from the focus rectifier to provide the required focus voltage. For best focus, the high voltage and focus voltage should track together, with both increasing or decreasing in the same ratio. The C-1 chassis (see Fig. 6) has two 430K-ohm series resistors (for a total of 860K) common to the flyback voltages fed to both the focus rectifier and the high-voltage rectifier. Assume that the color picture tube draws one milliampere of current; this will cause 860 volts to drop across the resistors, which will reduce both the high voltage and the focus voltage by that amount. Thus, proper focus is maintained at all brightness levels.

Adjustment of the tint in GE's KE color chassis is accomplished by varying the DC reverse-bias on a varactor diode, which changes its internal capacitance. This change in capacitance changes the phase of

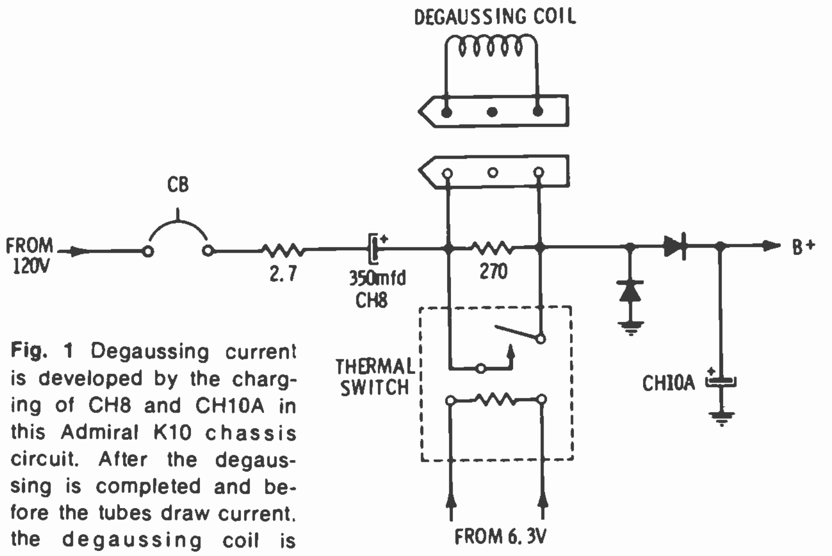


Fig. 1 Degaussing current is developed by the charging of CH8 and CH10A in this Admiral K10 chassis circuit. After the degaussing is completed and before the tubes draw current, the degaussing coil is shorted out by the contacts on the thermal switch, which has been heated by a resistive element connected to 6.3 volts AC.

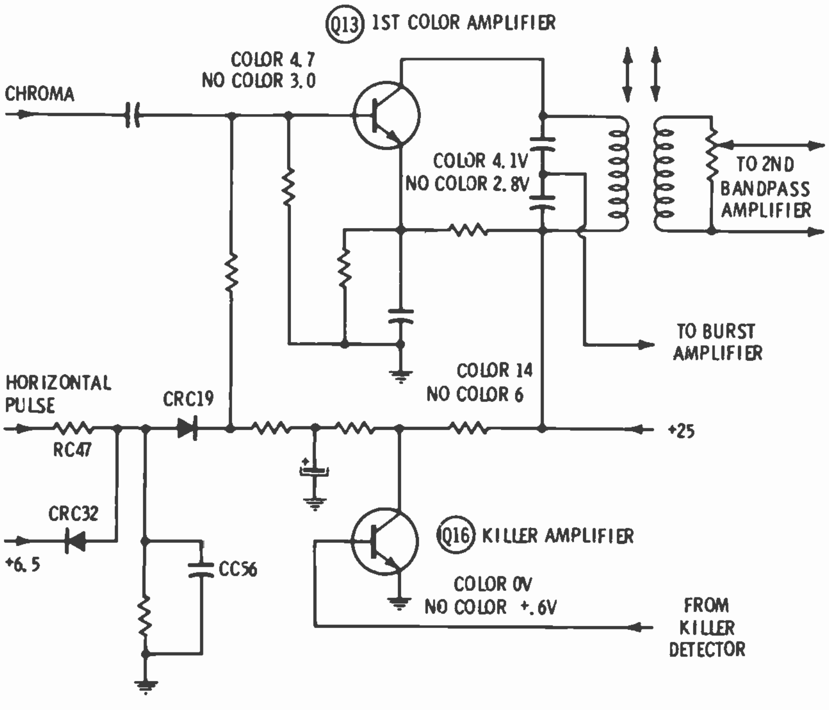


Fig. 2 During b-w reception, the Admiral K10 killer detector output is about +.6 volt, enough to make Q16 conduct and reduce the voltage on its collector to about 6 volts. This voltage is used as base supply voltage for Q13, which will have only .2 volt of forward bias and no gain. When color is received, the killer detector output voltage is nearly zero, Q16 has no bias, draws no current and the collector voltage is high (around 14 volts). This higher source voltage makes the base of Q13 about .6 volt more positive than its emitter, producing normal bias and gain. See the text for a description of the "burst assurance" action.

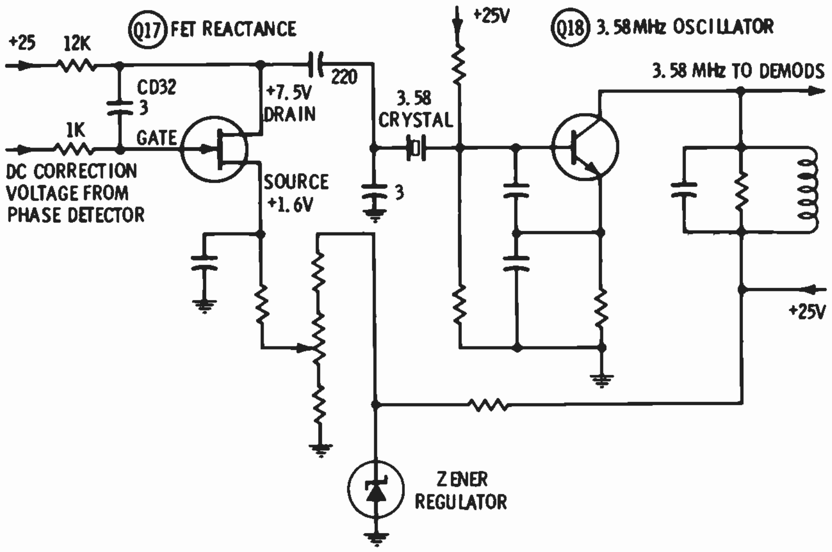


Fig. 3 A FET works just as well as a tube does in a reactance stage. The theory is the same, except a variable source voltage is used to set the basic oscillator frequency instead of the more conventional reactance coil.

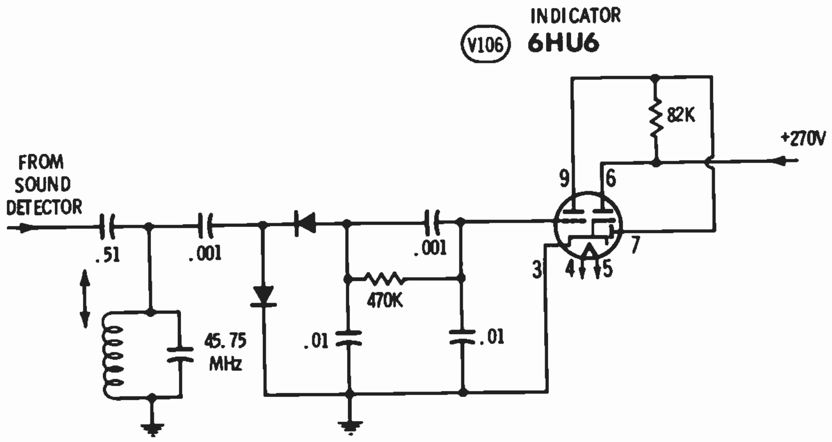


Fig. 4 This tuning-eye circuit is employed in the new Andrea color chassis.

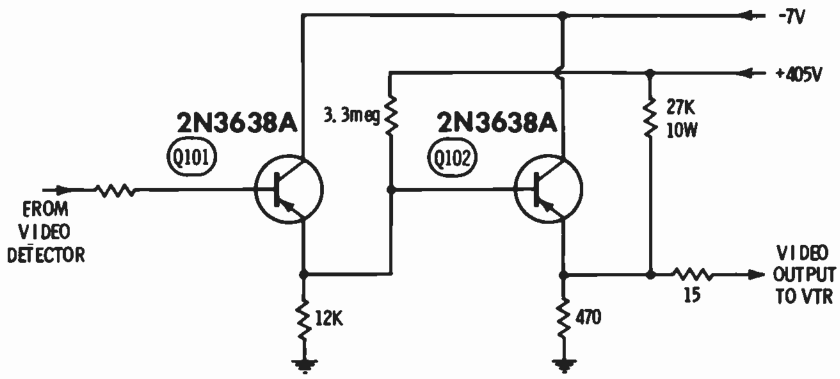


Fig. 5 Andrea has thought about the future and included a 75-ohm video output signal for use with a video tape recorder.

the 3.58-MHz carrier, which is obtained by ringing the 3.58-MHz crystal with the burst signal, as shown in Fig. 7.

The GE KE chassis, which is found in 23", 20" and some 12" color receivers, uses less solid-state circuitry than does the C-1 chassis; only two video stages, the blanking amplifier and the 3.58-MHz buffer are transistorized. High voltage is regulated by a 6LJ6 shunt regulator tube.

A separate negative power supply for the emitter of Q201, the first video amplifier, is provided so that the base can be direct-coupled to the negative-going video detector. The schematic is shown in Fig. 8. Just keep in mind that this circuit can be a source of hum which might be overlooked, and any decrease in the negative emitter voltage will make the picture darker, or eliminate the raster altogether.

Magnavox

New from Magnavox this year is the T940 color chassis which features TAC (Total Automatic Color). TAC consists of AFT (automatic fine tuning), pioneered by Magnavox in 1965, plus the completely new ATC (automatic tint control). These last two circuits were thoroughly discussed in the October '69 issue of ELECTRONIC SERVICING. Briefly, the principle of ATC is to change greenish-yellow and reddish-purple chroma phases into a 57-degree orange that is satisfactory as skin color. This is accomplished by overbiasing and gating two channels that have fixed amounts of phase shift in each, and combining this correction signal with the normal chroma signal just before it goes to the demodulators. Fig. 9 shows the complete schematic of the ATC circuit.

ACC voltages for gain reduction of the first chroma amplifier are taken from two different sources. One is from the killer detector, and is a conventional circuit (see Fig. 10). The other is from an additional DC voltage created by the rectification of the chroma signal itself. Control from the killer detector voltage is very good up to about 75% to 100% burst level; above that, its control is not effective. The control voltage from rectification of the chroma IF signal is very helpful above 100% burst level, and is especially effective where a station may transmit normal burst with color that is too strong.

Motorola

The new Motorola TS930 chassis is designed for 16" diagonal color portables. It is a hybrid design with very few tubes, and is identical (except in cabinet styling) to the Admiral K10 previously described. A rumor in the industry says that Motorola furnished transistors and other parts, while Admiral supplied the design and manufacturing.

The Quasar, Motorola's pioneering solid-state color receiver with the plug-in circuit boards, is manufactured in two different versions: The number of the familiar vertical chassis assembly that rolls out the front is TS915. The newer TS919 uses the same plug-in boards, but the horizontally mounted chassis slides out the rear for servicing.

All Quasars with the code letter "F" before the chassis number incorporate a new electronic voltage regulator for the 120-volt line input. As shown in the block diagram of Fig. 11, the filament transformer and the power transformer have 105-volt primaries. Between the transformers and one side of the line-voltage input are two resistors in series whose combined rating is 25 ohms at 100 watts of dissipation. A triac (bi-directional SCR) parallels these resistors and gives the effect of a variable voltage drop by shorting out the resistors for part of each cycle.

If the line voltage is 105 volts, the triac must conduct all the time so that the full voltage is applied to the transformers and no voltage is dropped across the resistors. With an input of 130 volts, the triac must be open at all times. The 25-ohm resistance develops 25 volts across itself, leaving the required 105 volts for the transformers. For line voltages between these extremes, the triac must be conducting for just part of each cycle. The lower the input voltage, the longer the triac conducts during each cycle.

The complete Motorola regulator schematic is shown in Fig. 12. The base of the regulator, transistor Q1Z, is supplied with a sample from the +95-volt power supply through a regulator control which sets the operating range. A low-pass filter eliminates most of the 120-Hz rip-

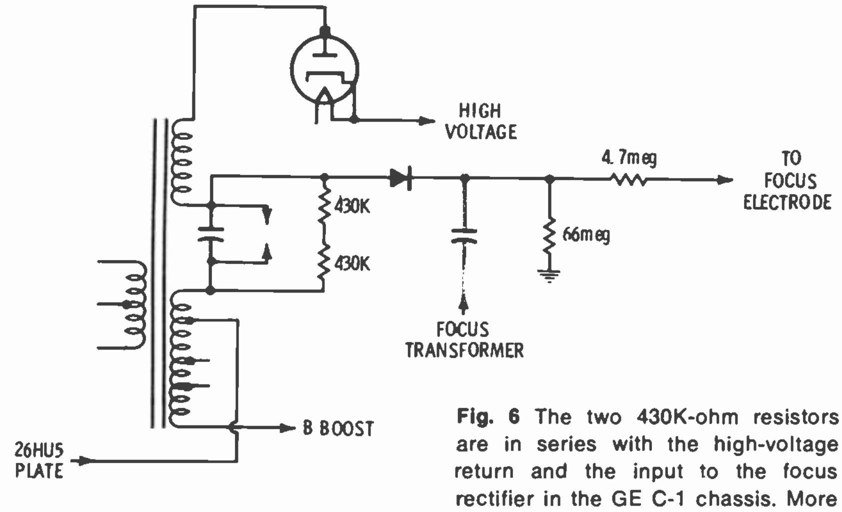


Fig. 6 The two 430K-ohm resistors are in series with the high-voltage return and the input to the focus rectifier in the GE C-1 chassis. More high-voltage current causes both the high voltage and focus voltage to drop and, thus, maintains good focus.

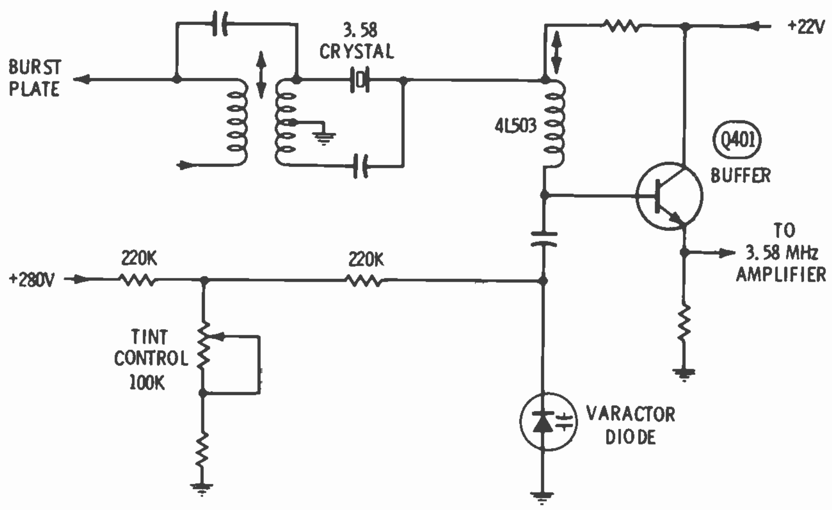
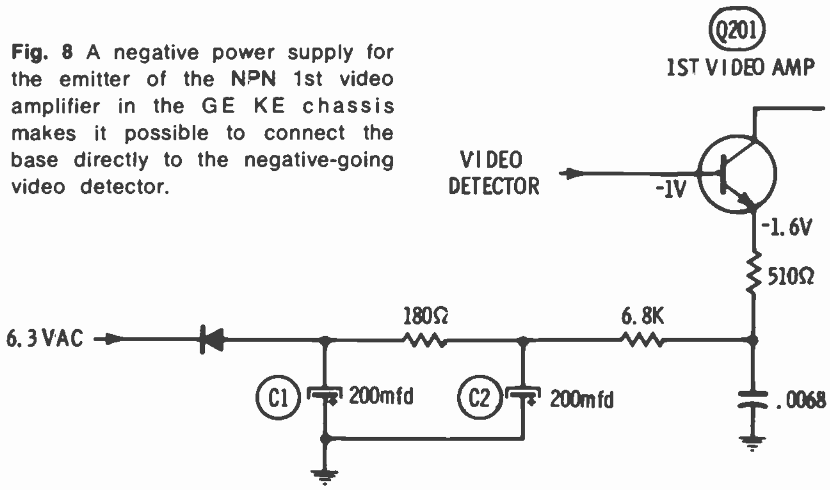


Fig. 7 Tint control action in the new GE color chassis is accomplished by changing the DC voltage on a varactor diode.

Fig. 8 A negative power supply for the emitter of the NPN 1st video amplifier in the GE KE chassis makes it possible to connect the base directly to the negative-going video detector.



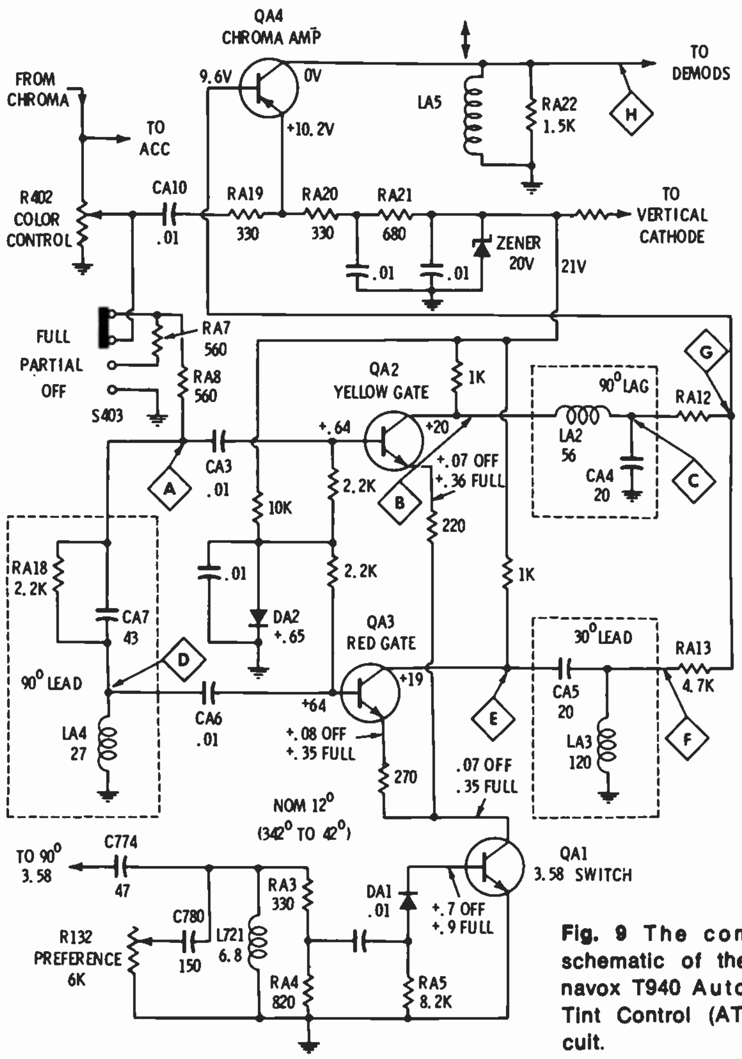


Fig. 9 The complete schematic of the Magnavox T940 Automatic Tint Control (ATC) circuit.

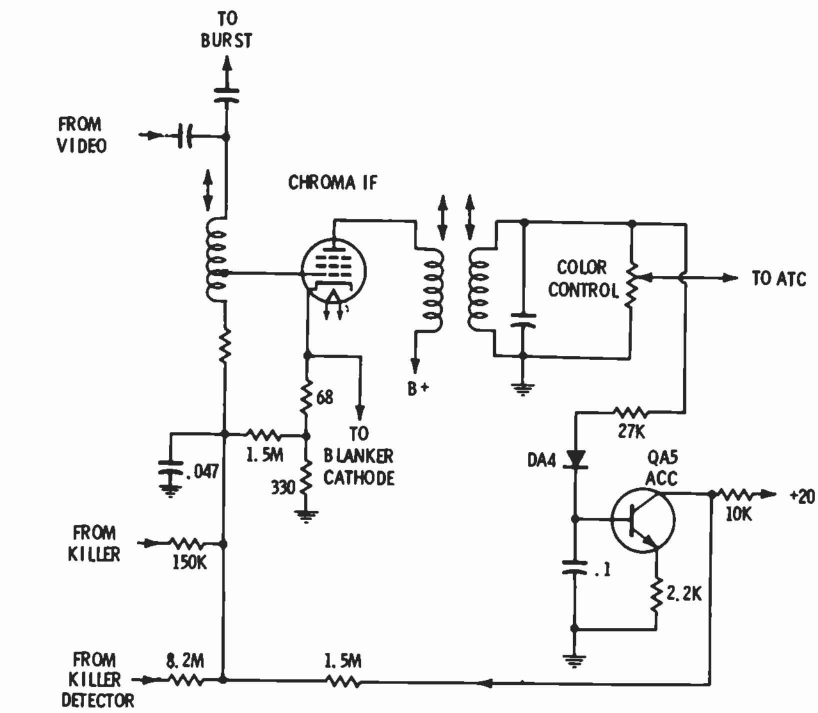


Fig. 10 Magnavox ACC has a double action which is especially helpful when the station broadcasts very strong color without excessive burst.

ple and slows down the response just enough to serve as an anti-hunt circuit. Voltage on the emitter is stabilized by a zener diode. Current from the emitter charges the .1-mfd capacitor, C3Z. When the voltage reaches the required level, a bi-directional switch (similar to two diodes back-to-back), E3Z, conducts somewhat like a zener and connects the capacitor to the transformer. The capacitor discharge current flowing through the transformer (T3Z) primary generates a sharp pulse in the secondary, which forces the triac (E1Z) into conduction. The triac will continue to conduct until its anode voltage drops to zero.

So far in our description, the firing of the triac is random, which would give very poor regulation. A synchronizer is needed to bleed the charge out of capacitor C3Z 120 times per second. This is accomplished by transistor Q2Z, which is reverse-biased and non-conductive until it is forward biased through C4Z by the positive-going tips of the parabolic waveforms from the rectified outputs of E4Z and E5Z.

Now, back to the regulator transistor. Assume the regulator control has been set correctly and the receiver is plugged into 120 volts AC. If the +95 volts decreases for any reason (such as increased drain on the supply or a reduction in line voltage), the forward bias on Q1Z is **increased.** This results in more emitter current, which charges C3Z faster, thus causing the triac to start conducting **sooner** in the cycle. Once fired, the triac stays on until the anode voltage goes to zero. With the triac conducting during **more** of each AC cycle, the voltage drops across the loss resistors are reduced, and the voltage applied to the transformers is increased. This, in turn, raises the +95-volt supply. These actions are all reversed if the +95-volt supply should increase.

Packard Bell

Integrated circuits (IC's) and field-effect transistors (FET's) are of special interest in the chroma circuit of the Packard Bell 98C-21 chassis. The "X" and "Z" chroma demodulators in this chassis are both dual-gate FET's, with the chroma applied to one gate and the 3.58-MHz signal applied to the other, as shown in Fig. 13. (Notice the similarity to circuits which use pentode tubes.)

An IC unit that is the equivalent of five transistors and two resistors is used as the 3.58-MHz oscillator crystal, and by power and brute force causes the oscillator to lock to the amplitude and phase of the burst. A separate phase detector supplies the control voltage for the color killer and ACC functions, as shown in Fig. 14.

Philco

The Philco 19FT60 chassis utilizes only seven tubes; all other active components are solid state. A new method of degaussing used in this chassis is shown in Fig. 15. Line voltage is supplied to the degaussing coil in series with a positive-temperature-coefficient resistor called a "posistor." When the receiver is first turned on, a large amount of AC flows through the low-resistance posistor and the coil. The current heats the posistor, and its resistance increases until it has shut off all significant degaussing

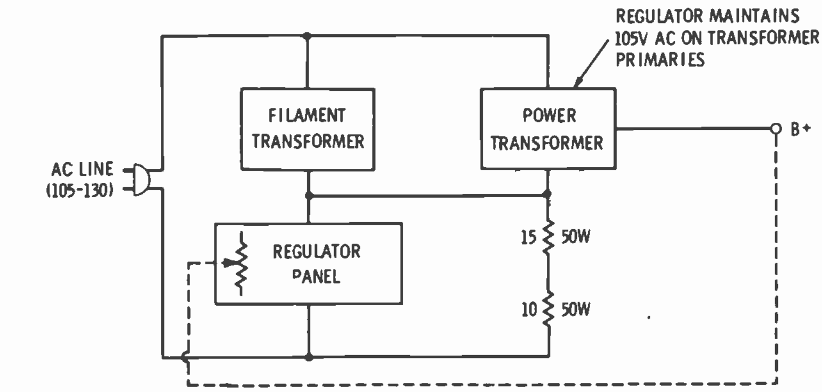


Fig. 11 All power supplies are regulated in the new Motorola Quasar color receivers.

action. Philco states that this system results in a stronger field at the start of degaussing.

Varactor diode control of an IC color oscillator is a noteworthy addition to the chroma channel (see Fig. 16). The positive feedback path from pin 7 of the IC back to pin 3 is through the 3.58-MHz crystal and the varactor diode. Any correction voltage from the phase detector changes the internal capacitance of the varactor diode and shifts the frequency or phase of the 3.58-MHz oscillator. To adjust the frequency, ground the phase-detector end of the 68K-ohm resistor and adjust coil L100 for zero beat with the station or a color-bar generator signal applied.

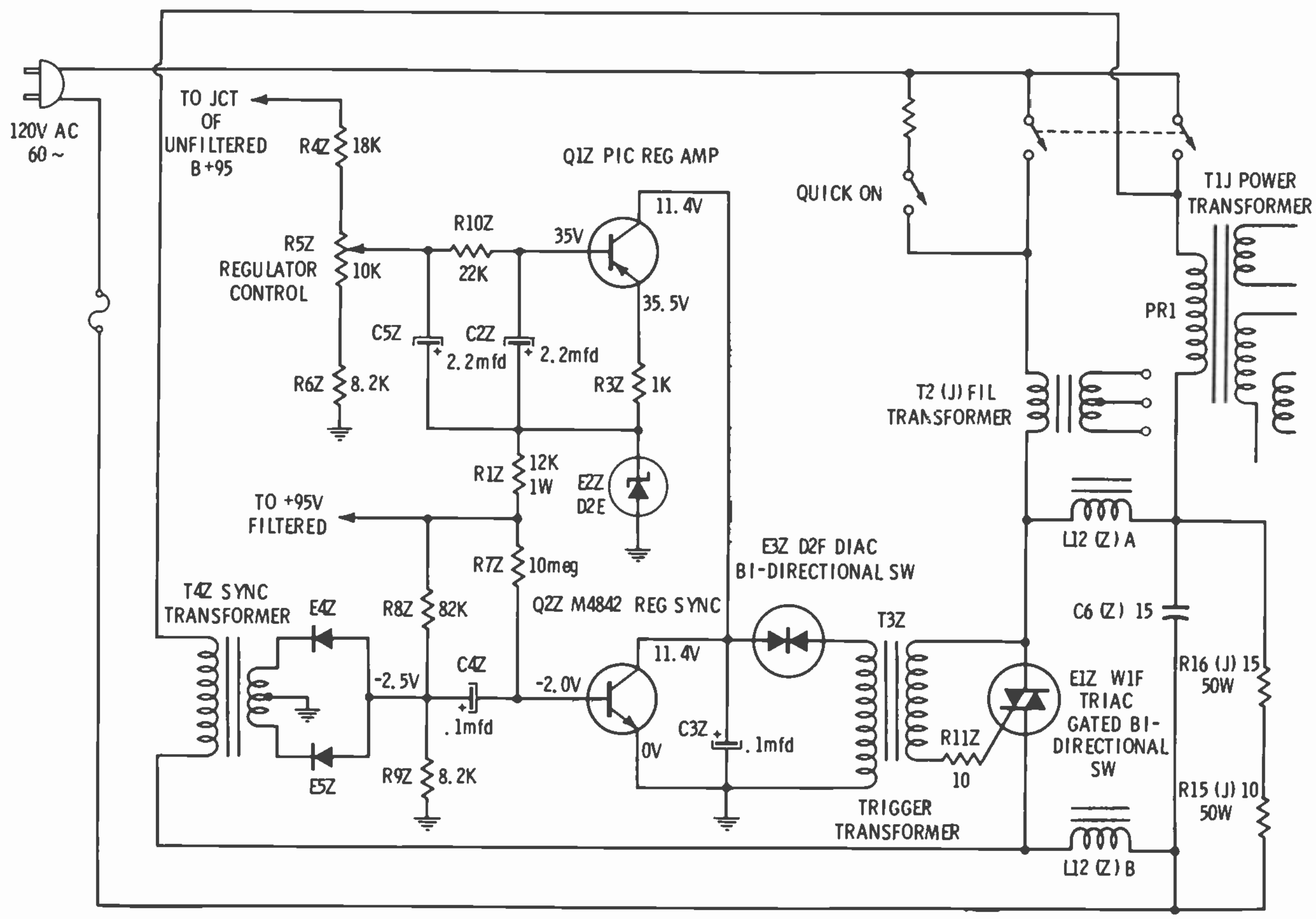


Fig. 12 A simplified schematic of the power supply regulation circuit in the Motorola Quasar chassis.

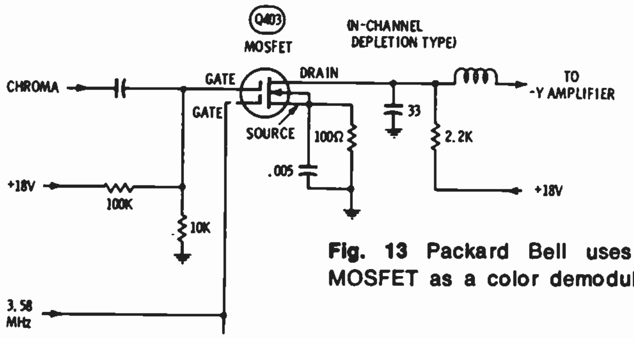


Fig. 13 Packard Bell uses a dual-gate MOSFET as a color demodulator.

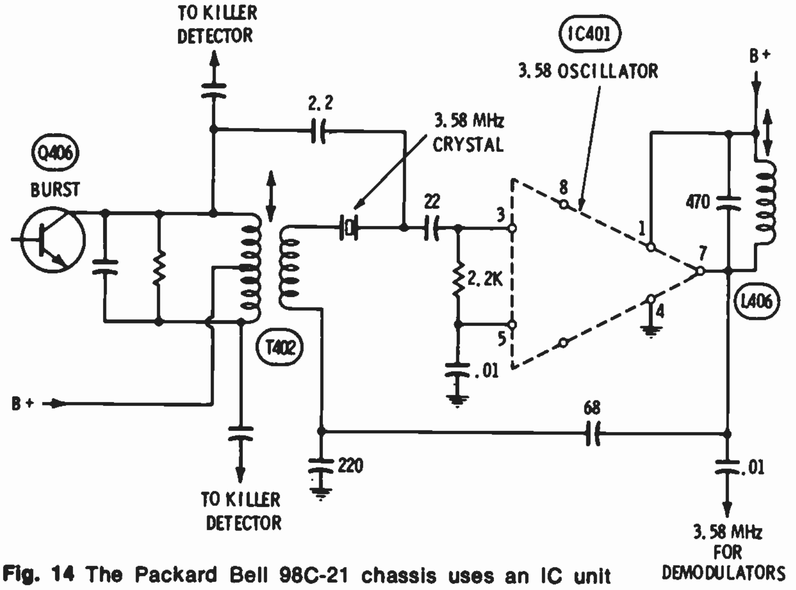
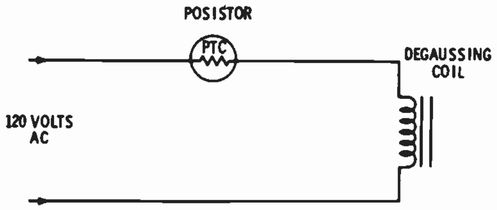


Fig. 14 The Packard Bell 98C-21 chassis uses an IC unit equivalent to 4 transistors and 2 resistors as the active element in the 3.58-MHz oscillator.

Fig. 15 A "posistor" is used to stop the degaussing action in the Philco 19FT60 chassis. The resistance increases along with temperature and squeezes off the degaussing current to the coil.



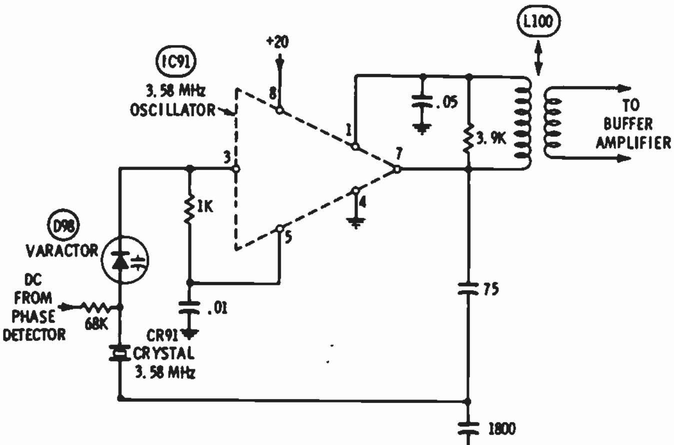


Fig. 16 Color locking in the Philco 19FT60 chassis is accomplished by a varactor diode wired between the crystal and the 3.58-MHz IC oscillator. Correction voltages from the phase detector change the internal capacitance of the varactor.

RCA

Several of RCA's current chassis are being continued, but there is a new CTC42X chassis used with 16" color kinescopes. The CTC42X is a hybrid design with 13 tubes (including 5 duals), 17 transistors, 2 IC's, 17 diodes, 2 zeners and 1 damper diode. The tuner, IF and chroma circuits are very similar to those in the CTC38, except that a 2DS4 is used as the RF amplifier in the tuner because of the series heater connections. High-voltage regulation is by AC pulse regulation exactly as is used in the CTC36 chassis. No high-voltage adjustment control is provided. A diode in the regulator cathode is a safety precaution; if the regulator draws no current, the diode is reverse biased and acts as an open circuit. The cathode voltage of the diode drops to zero and there is no plus voltage there to be fed back to the grid of the horizontal output tube. This makes the grid too negative, and the width and high voltage are both reduced until the regulator circuit is repaired.

The television industry has started to produce tuners without switches; in most designs, varactor diodes are used as variable capacitors by varying a DC voltage applied to them.

The RCA design does NOT function in that manner. In the CRC47, RCA has a switchless VHF tuner that is tuned by coils and stray-circuit capacitance, but the switch contacts have been replaced by switching diodes. Fig. 17 shows part of the antenna and RF tuned circuits in which the channels are selected by diodes.

Any diode is a voltage-controlled switch, regardless of the kind of circuit in which it is used. Assume that none of the channel selector inputs have voltage on them, so that all the diodes are reverse-biased and, therefore, are open circuits. If a more positive voltage (+16 volts) is applied to the anodes of CR2313 and CR2213 than is present on their cathodes, they become low-resistance short circuits. C13 acts as an AC ground, and the channel 13 coils are switched into the circuit. When the +16 volts is removed from the channel 13 diodes and applied to CR2311 and CR2211, channel 13, 12 and 11 coils will be bypassed to ground through the diodes. Channel 11 is then switched into the circuit. And so on, with the coils adding in series down to channel 2. The mixer and oscillator stages are tuned this same way by using voltage to key the diodes on or off.

This system would have one small advantage even if a regular switch were used to supply the keying voltage to the diodes: Dirty switch contacts would have no effect on the tuning until the voltage at the anode of the diode dropped below the voltage at the cathode. But there is much more to the system, and manual switches are NOT used. RCA's "The Two Thousand Technical Manual" uses 120 pages to explain the complete RCA system of electronic VHF tuning and motorless remote operation, which employs 78 transistors, 122 diodes, 4 FET's, 9 zeners and 6 IC's. The "Two Thousand" model is a prestige, limited-production version of RCA's well-known CTC40 Transvista chassis.

Before you read the manual, it would help if you studied some basic computer principals, because the operation of this tuner is based on binary mathematics and computer functions. For example, here is the binary code for the various channels:

channel 2	0000
channel 3	0001
channel 4	0010
channel 5	0011
channel 6	0100
channel 7	0101
channel 8	0110
channel 9	0111
channel 10	1000
channel 11	1001
channel 12	1010
channel 13	1011
channel 14	1100 (UHF)

A positive-polarity pulse is designated "1", and means "closed" time, "on" time or "yes" voltage. "O" designates alternate, or "off", half-cycles. According to this binary code for the channels, four sources and four gates are required to select the right channel.

Electronic scanning from one VHF channel to another is started by a "clock", which is merely a 330-Hz multivibrator oscillator. This is followed by three dividers (or counters). The clock and the counters have outputs of "1" or "0",

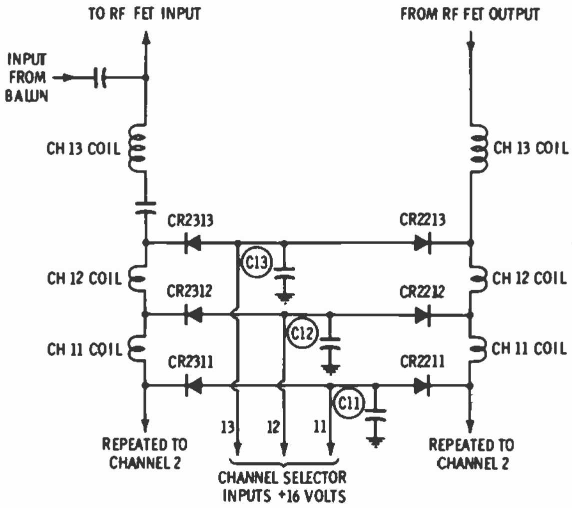


Fig. 17 Diodes replace switch contacts in the new RCA tuner.

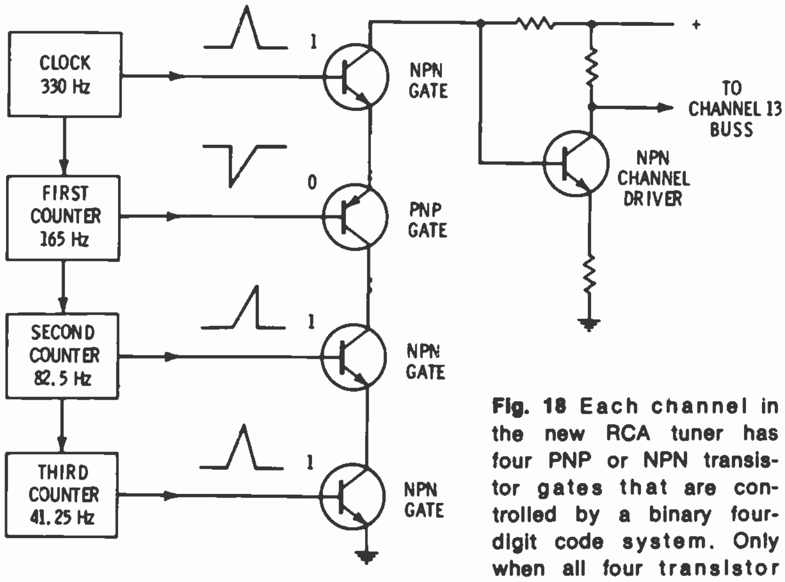
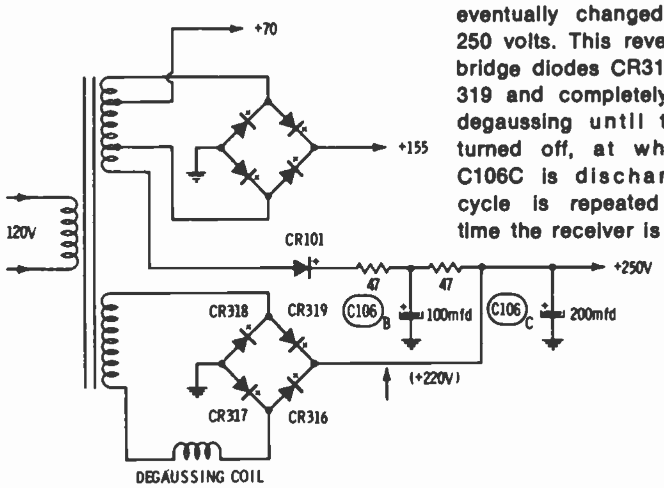


Fig. 18 Each channel in the new RCA tuner has four PNP or NPN transistor gates that are controlled by a binary four-digit code system. Only when all four transistor gates are conductive at the same time (proper pulse at the base of each) can the diodes for that channel be forward biased.

Fig. 19 In the RCA CTC47 chassis the picture tube is degaussed by a bridge rectifier (CR316, CR317, CR318 and CR319) charging C106C to about 220 volts. C106C is subsequently charged more slowly from CR101 through the two 47-ohm resistors and C106B until it is eventually changed to about 250 volts. This reverse biases bridge diodes CR316 and CR-319 and completely stops all degaussing until the set is turned off, at which time C106C is discharged. The cycle is repeated the next time the receiver is turned on.



and the outputs change during each cycle, thus making the combination for one channel, then the next, etc., until a channel is found with the programming switch set to stop. Fig. 18 shows the outputs from the clock and the counters going to both NPN and PNP polarity gates. An NPN transistor "closes" on a "1" and a PNP transistor closes on a "0". For example, when the gates are supplied with pulses that produce binary code 1011, all the gates close at the same time, current flows to reduce the base voltage of the channel driver, whose collector voltage rises and forward biases all the diodes used for switches on channel 13. Other channels have a different combination, but all work on the same principle as that given for channel 13.

Do you know what "interface" means? It is a word that has gained increased popularity in scientific and broadcasting circles during the last few years, and it means interconnected, or better yet, an interconnection of unmatched or unequal equipment. The new RCA tuning system must have interface between the gating, the read-out (channel indication) programming switch and the circuits that mute the picture and sound and disable the AFT during channel change. Remember, there is **no manual** channel selector.

The motorized UHF power tuning system can be directed up or down in frequency. There are no detents or manual tuning; the motor

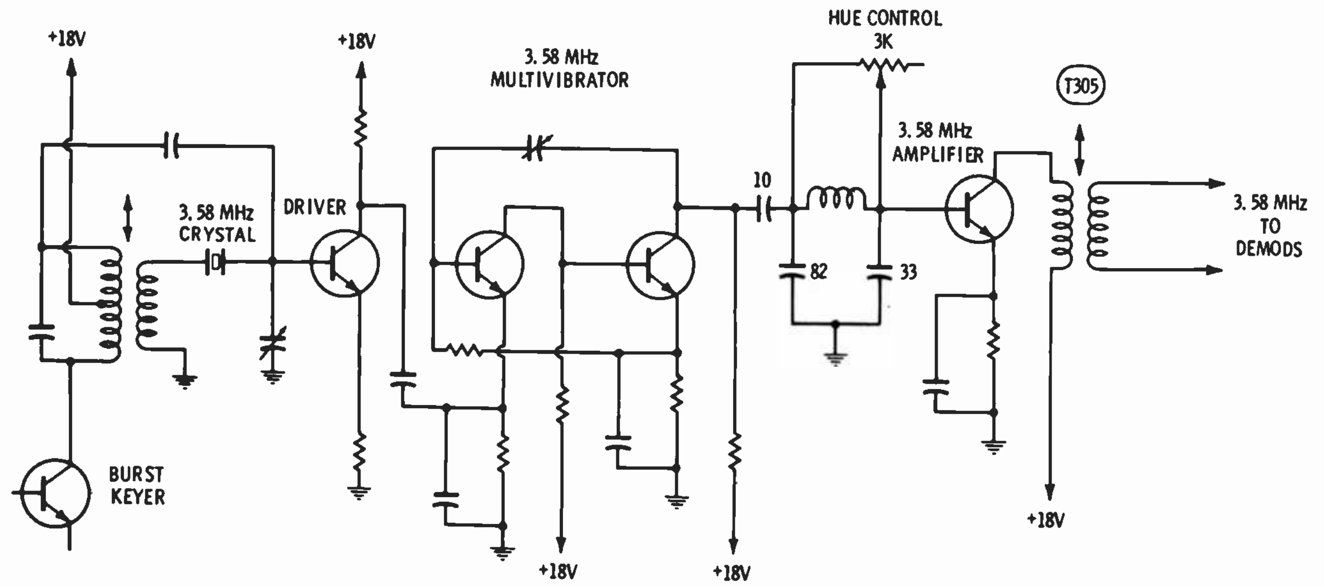


Fig. 20 The Sony 3.58-MHz multivibrator oscillator is synchronized at the correct frequency and phase by a signal produced by ringing a quartz crystal with the broadcast burst signal.

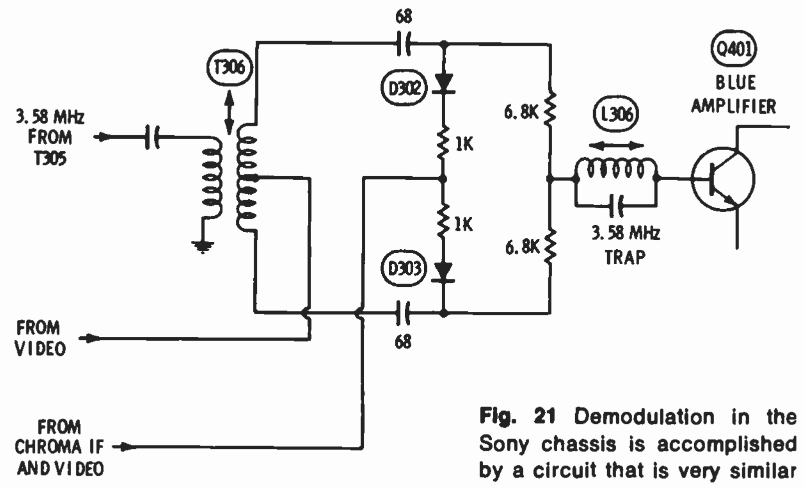


Fig. 21 Demodulation in the Sony chassis is accomplished by a circuit that is very similar to that of some GE chassis, except that the video is added to each demodulator.

keeps running until a signal of a certain pre-set minimum amplitude with horizontal sync is received. The motor then stops and the AFT pulls the signal into correct tuning.

Remote control is an integral part of the tuning assembly. It has only one motor and relay for the UHF function; all other active components are solid state. Control over volume, color and tint are by FET's, whose gate voltages are determined by voltages stored in "memory modules".

Less exotic circuit changes are also found in the CTC47, such as in the automatic degaussing system, the schematic of which is shown in Fig. 19. When the receiver is first turned on, C106C is charged to about +220 volts by a rapidly dwindling train of rectified full-wave pulses from the bridge rectifier, consisting of CR316, CR317, CR318 and CR319. This charging current passes through the degaussing coil and demagnetizes the picture tube in about 20 milliseconds. DC voltage from CR101 is also supposed to charge C106C, but it is delayed by the two 47-ohm resistors and C106B, the 100-mfd filter capacitor, and does not rise above +220 volts until degaussing is completed. When fully charged from CR101, C106C has +250 volts on it. This voltage reverse biases CR316 and CR319 (in the bridge), and absolutely no current comes through the bridge. When the receiver is turned off and C106C is discharged, the set can be immediately turned back on and full degaussing obtained. There is no thermistor to introduce a time delay.

Sony

A 12" Trinitron color picture tube is used in the Sony KV-1210U chassis, which employs 44 transistors, 35 diodes and 1 high-voltage rectifier tube. The Trinitron color picture tube has only one electron gun, although it does have three cathodes, to which are applied the b-w video and chroma signals. The

Fig. 22 The output to each cathode of the Sony Trinitron picture tube is a matrixed signal that includes both chroma and video. Drive and background controls are included in each of the three channels to make b-w tracking possible; there are no individual screen controls.

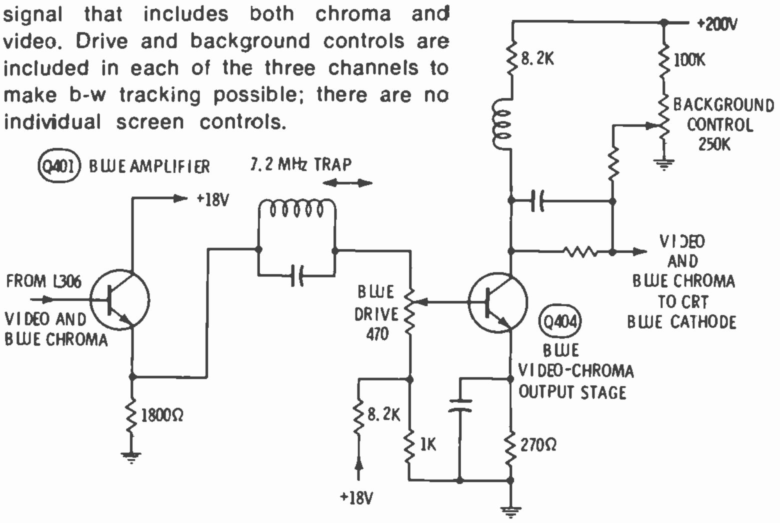
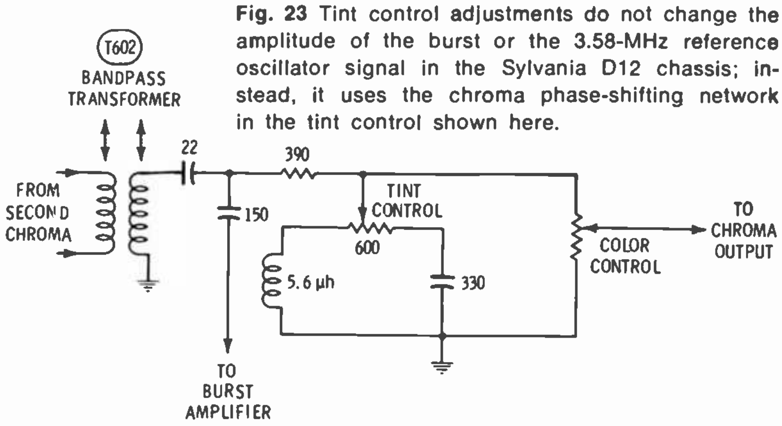


Fig. 23 Tint control adjustments do not change the amplitude of the burst or the 3.58-MHz reference oscillator signal in the Sylvania D12 chassis; instead, it uses the chroma phase-shifting network in the tint control shown here.



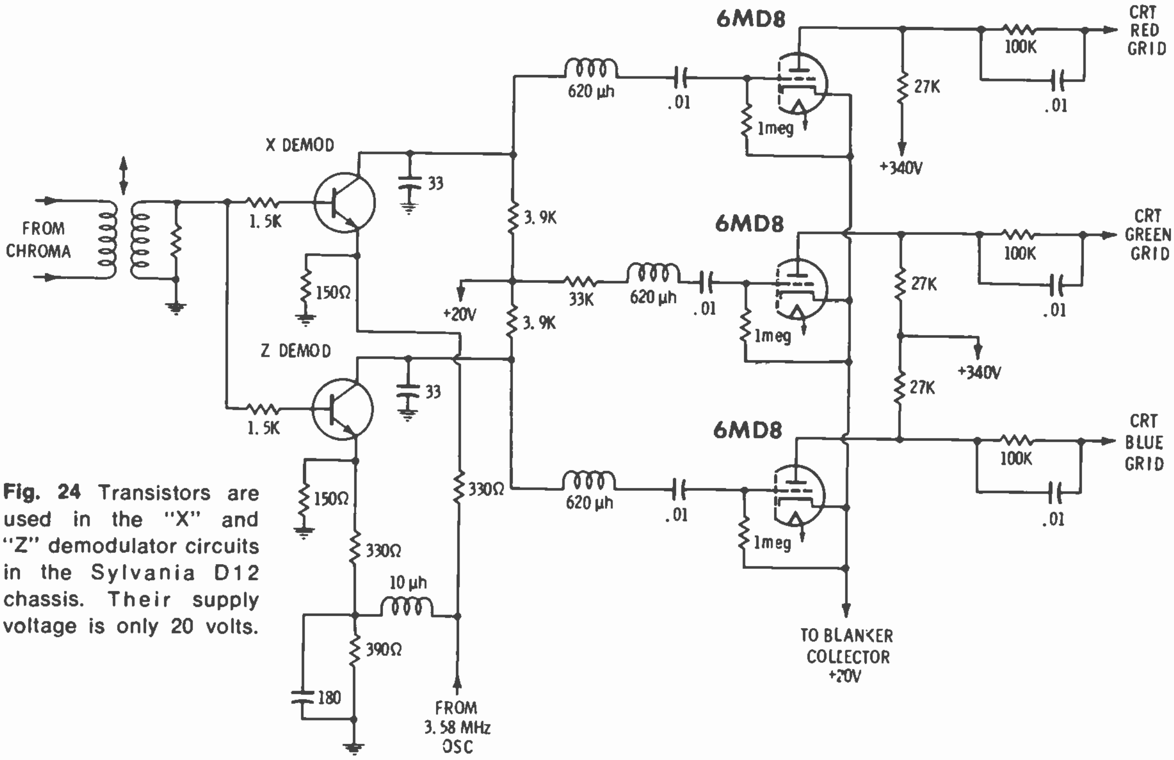


Fig. 24 Transistors are used in the "X" and "Z" demodulator circuits in the Sylvania D12 chassis. Their supply voltage is only 20 volts.

manufacturer claims twice the brightness of conventional three-gun tubes, and much simpler convergence adjustments.

Generation of the 3.58-MHz reference carrier also is accomplished in a different way in this chassis. The burst rings a 3.58-MHz crystal, and this nearly continuous signal is used to synchronize a 3.58-MHz multivibrator oscillator. The oscillator signal is fed through a hue control (see Fig. 20), an amplifier and a tuned transformer before it is applied to the three balanced diode demodulators.

One diode demodulator is used for each primary color. These demodulators, shown in Fig. 21, are similar to the ones used in some GE receivers, but do not have balancing controls. Matrixing is accomplished in the demodulator rather than at the picture tube; the video is applied to the junction of the two diodes and will go through either diode that is forward biased. Video and chroma both must be matrixed and brought to the CRT cathodes because there is only one control grid.

Fig. 22 shows more of the blue channel (the other two channels are nearly identical to this one), in which an emitter follower, Q401, drives the base of the power output stage through the adjustable control labeled "blue drive". There are no individual screen voltage adjustments because there is only one screen grid, so the three drive controls and the three background controls are used to obtain correct screen color and b-w tracking.

Sylvania

The Sylvania D12 (Gibraltar) chassis is another hybrid. It has 24 transistors, one IC and 9 tubes.

Generally speaking, tint-control circuits that tune a burst or 3.58-MHz reference signal transformer have one major drawback: A large amplitude change in the signal during tint adjustments.

Sylvania changes the tint in the D12 chassis by varying the phase

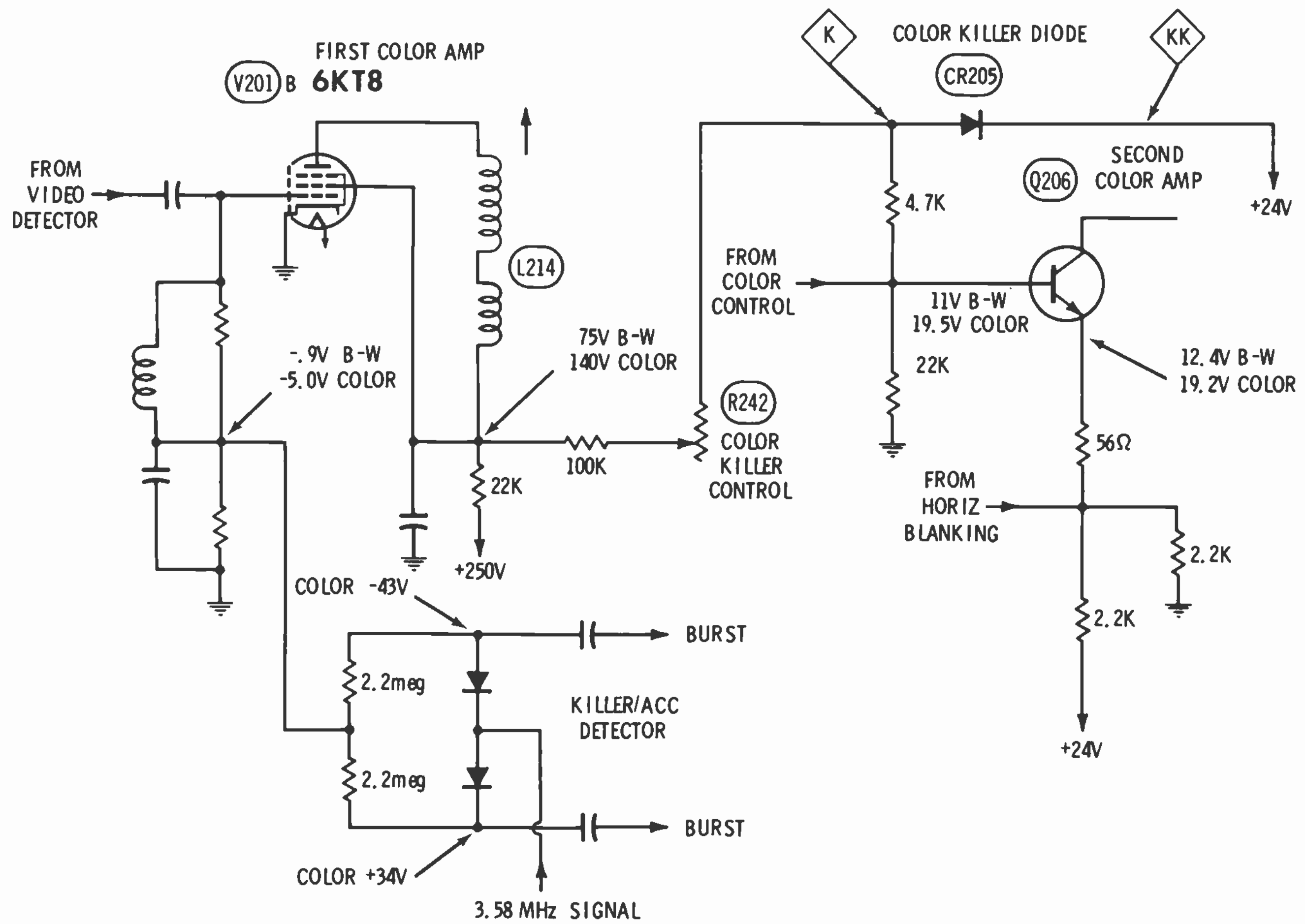


Fig. 25 Color-killer action in the Zenith 14A9C51 chassis is rather devious. ACC action changes the plate and screen voltages of V201B, the first color amplifier; these voltages, in turn, determine the bias on Q206, the second color amplifier. During b-w reception, the plate voltage on V201B is very low and the base of Q206 is less positive than its emitter; thus, the transistor is reverse-biased. With strong color tuned in, the plate voltage of V201B will be +140 volts or higher, and Q206 will have normal bias for good amplification. The color-killer diode is not essential for operation of this circuit, but is a refinement to prevent excessive forward bias on Q206. This is done by clamping the voltage at testpoint "K" to +24 volts whenever the voltage tries to exceed that limitation.

of the chroma signal between the bandpass transformer and the color control (Fig. 23). The 600-ohm tint control, in effect, switches in a capacitor or a choke to make the phase lag or lead, and a 390-ohm resistor isolates the variable part of the circuit from the burst.

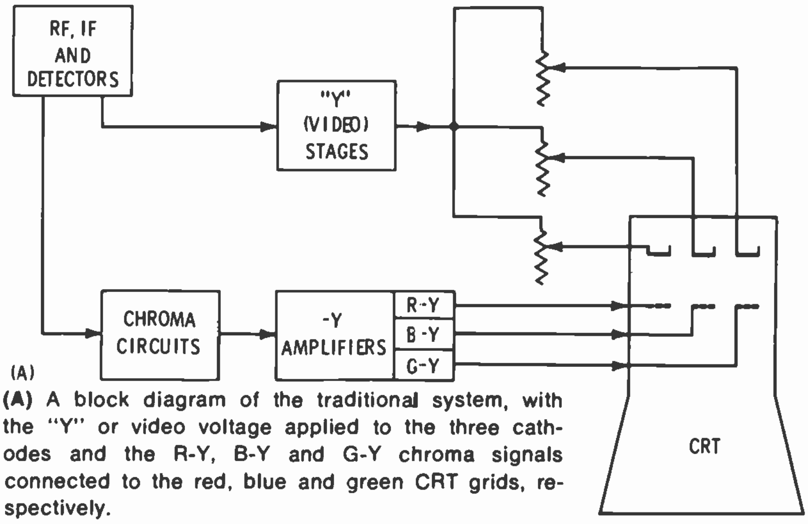
Transistors are used as demodulators in the D12 chassis, as shown in Fig. 24. Chroma is applied to the bases, while a 3.58-MHz reference signal with a 90-degree phase difference is supplied to the emitters. The demodulator collector circuits and the following –Y amplifier circuits are nearly identical to previous tube versions.

Zenith

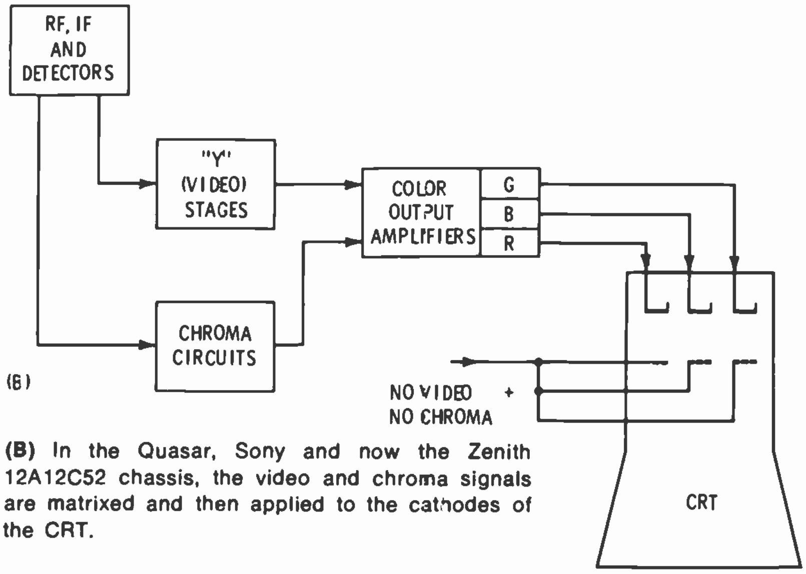
A different kind of color killer is used in the Zenith 14A9C51 chassis. The schematic of this circuit is shown in Fig. 25. Assume that the set is tuned to a b-w program. Q206, the second color amplifier, is biased to cut-off by the 12.4 volts applied to its emitter by the voltage-divider action of the two 2.2K-ohm resistors between the +24-volt source and ground. Its base voltage is only 11 volts because of the low source voltage of +75 volts at the screen of V201B. Thus, the transistor is 1.4 volts reverse biased. When color is tuned in, the killer/ACC detector has an output of several volts negative, which reduces the gain of V201B and causes the screen voltage to rise above +140 volts. This increases the base voltage of Q206 to about +19.5 volts; the resulting emitter raises the emitter voltage to about +19.2 volts, and the transistor amplifies. CR205, the killer diode, has two functions: One is to make sure that the anode voltage does not increase above +24 volts. If it does, the diode conducts and clamps the circuit to the +24 volts as protection and bias limiting for the second color IF amplifier transistor. The other function is to provide a convenient way to defeat the color killer so that the second color amplifier can conduct: short across the diode from test points "K" to "KK".

The first commercial color receiver ever placed on the market in 1954 had a very complex matrixing system to combine the chroma and video signals into three pure primary color signals that were fed separately to the three picture tube

Fig. 26 Matrixing of the b-w and chroma signals traditionally has been accomplished in the picture tube.



(A) A block diagram of the traditional system, with the "Y" or video voltage applied to the three cathodes and the R-Y, B-Y and G-Y chroma signals connected to the red, blue and green CRT grids, respectively.



(B) In the Quasar, Sony and now the Zenith 12A12C52 chassis, the video and chroma signals are matrixed and then applied to the cathodes of the CRT.

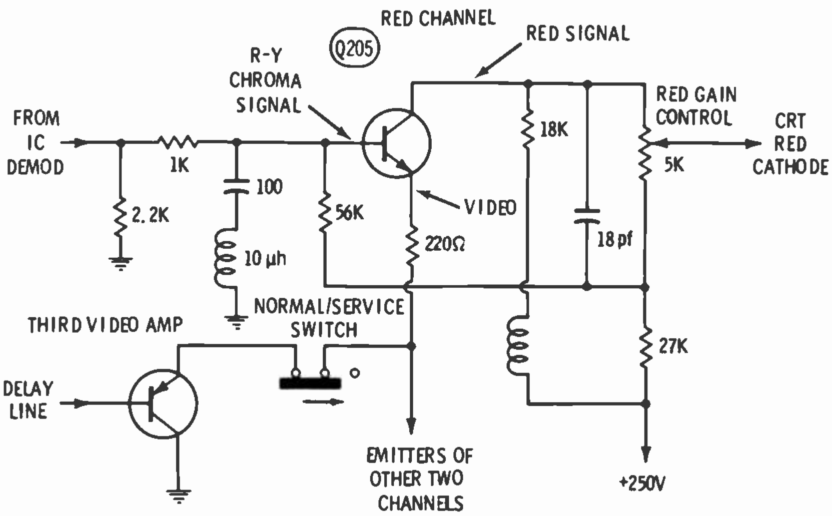


Fig. 27 In the red channel of the Zenith chassis, the chroma signal from the IC demodulator goes to the base of Q205, while the video signal goes to the emitter. The true red signal (not R-Y) is taken from the collector of Q205 and is routed through the red gain control to the red cathode of the CRT.

grids. This can be called pre-CRT matrixing. RCA used this system in the 1954 and 1955 model color receivers, then it was not used for years until Motorola updated it in the Quasar chassis.

Fig. 26A shows a block diagram of the traditional system of matrixing in the picture tube by feeding the video to the CRT cathodes and the chroma –Y signals to the three control grids. There apparently are no basic drawbacks in this system, except that four powerful amplifiers are necessary to completely modulate the CRT current. In solid-state circuits, this means that these amplifiers must operate with 160 to 200 volts on the collectors; consequently, elimination of the video output stage seems a good idea. Also, the new Trinitron color picture tube demands pre-CRT matrixing. Fig. 26B is a block diagram of a system in which the chroma/video matrixing is accomplished ahead of the picture tube.

The schematic of the red amplifier channel in the Zenith 12A12-C52 color chassis is shown in Fig. 27. The green and blue channels are similar to the red. From the IC, the R-Y signal is connected to the base of Q205. Video from the emitter of the third video amplifier is applied to the emitter of Q205. The true input to Q205 is between base and emitter, and the transistor sees the resultant voltage as though the video had been inverted in phase and applied to the base. The red signal on the collector includes both video and chroma and is applied to the CRT red cathode. The CRT grid has no signal on it, just a DC voltage to maintain normal brightness.

Conclusion

The preceding paragraphs have described briefly the most revolutionary designs we are aware of in the 1970 color chassis. For additional knowledge of the operation of these new circuits, we suggest that you obtain the manufacturers' literature about the circuits you are interested in and, whenever possible, attend manufacturers technical training sessions—in most cases such sessions are open to all electronic technicians and there is no charge. ▲

Chapter 11

New and Changed Circuitry In '71 Color TV

General trends in the designs of new color TV circuits for 1971 are almost identical with those reported in Chapter 10 of this book. These major trends are:

- Increased use of solid-state components in both hybrid and all-solid-state designs. Integrated circuits show a small increase in usage, but the increase is less than had been predicted by some sources.
- Solid-state high-voltage tripler and quadrupler rectifier assemblies, which include both the diodes and capacitors, are found in nearly half of the new chassis. Two such assemblies, employed in Zenith chassis, are shown in Fig. 1. High-voltage regulators of the 6BK4 type are used only in about 15 percent of the new chassis designs.
- Automatic tint control (ATC) circuits are in fashion now; most of the major manufacturers offer at least a few models equipped with one of the two basic types of ATC.
- Varicap- or varactor-tuned circuits in both VHF and UHF tuners have received much publicity, but, as yet, few new color receivers have this feature.
- The use of plug-in modules is increasing slowly, but the trend is well established and probably will eventually be adopted by most major manufacturers.

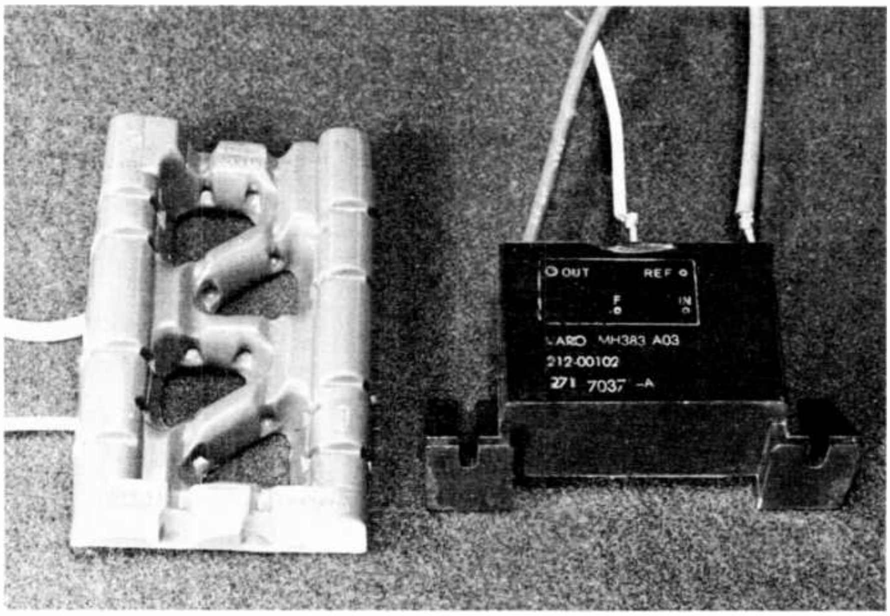


Fig. 1 Two of the several types of high-voltage rectifier and filter assemblies used in new Zenith color TV receivers.

Despite the undeniable fact that the major trends differ only in degree from those of last year, a varied assortment of completely new and modified versions of previously existing circuits are incorporated in the new color receivers. The new and most-changed circuits are analyzed for you in the following paragraphs.

Automatic Tint Controls

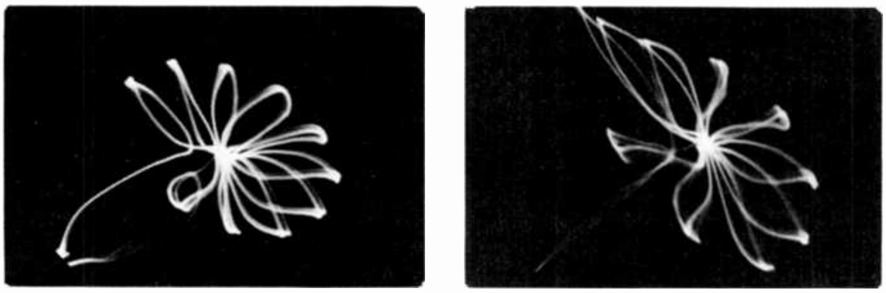
Automatic tint control (ATC) circuits are intended to offset the undesirable and frequent changes in tint, or hue, caused by variations in the broadcast signal. Because green or purple skin colors are the most displeasing flesh tints, all of the ATC circuits increase the amount of orange in the picture. Color-bar patterns, when the ATC is switched on, show as many as four orange bars instead of the usual one. Because some false hues will result from any ATC action that broadens the range of orange skin tones, only enough ATC should be used to keep the flesh tones relatively consistent and of an acceptable hue.

When servicing ATC-equipped receivers, remember that ATC action will change the null, or crossover, points of the color-bar signal viewed on a scope connected to the CRT, or the actual color-bar pattern viewed on the screen of the picture tube.

Magnavox and Admiral ATC

The ATC system in the new Magnavox T951 color TV chassis changes the phases of the yellow or red chroma signals to a 57-degree orange, which is satisfactory as skin color to most people. This is accomplished by gating on and off two transistorized channels, each of which has fixed amounts of phase shift. This correction signal then is combined with the normal chroma signal just before it is applied to the demodulators.

Vector patterns offer the fastest and most dramatic proof of this circuit action, as shown in Fig. 2. With the ATC switch in the FULL position, the first three petals are all at 57-degrees, and even petals 4 and 10 are moved nearer to 57-degrees. Petals 6, 7 and 8 are not affected at all; these are the hues in the cyan region. (Refer to page 26 of the October '69 issue of ELECTRONIC SERVICING for a more detailed explanation of this



(A) Normal vector pattern with the third petal at 90 degrees (red.)

(B) The first three petals all are at 57 degrees (orange) when the ATC switch is in the FULL position.

Fig. 2 Vector patterns illustrating Magnavox's ATC action.

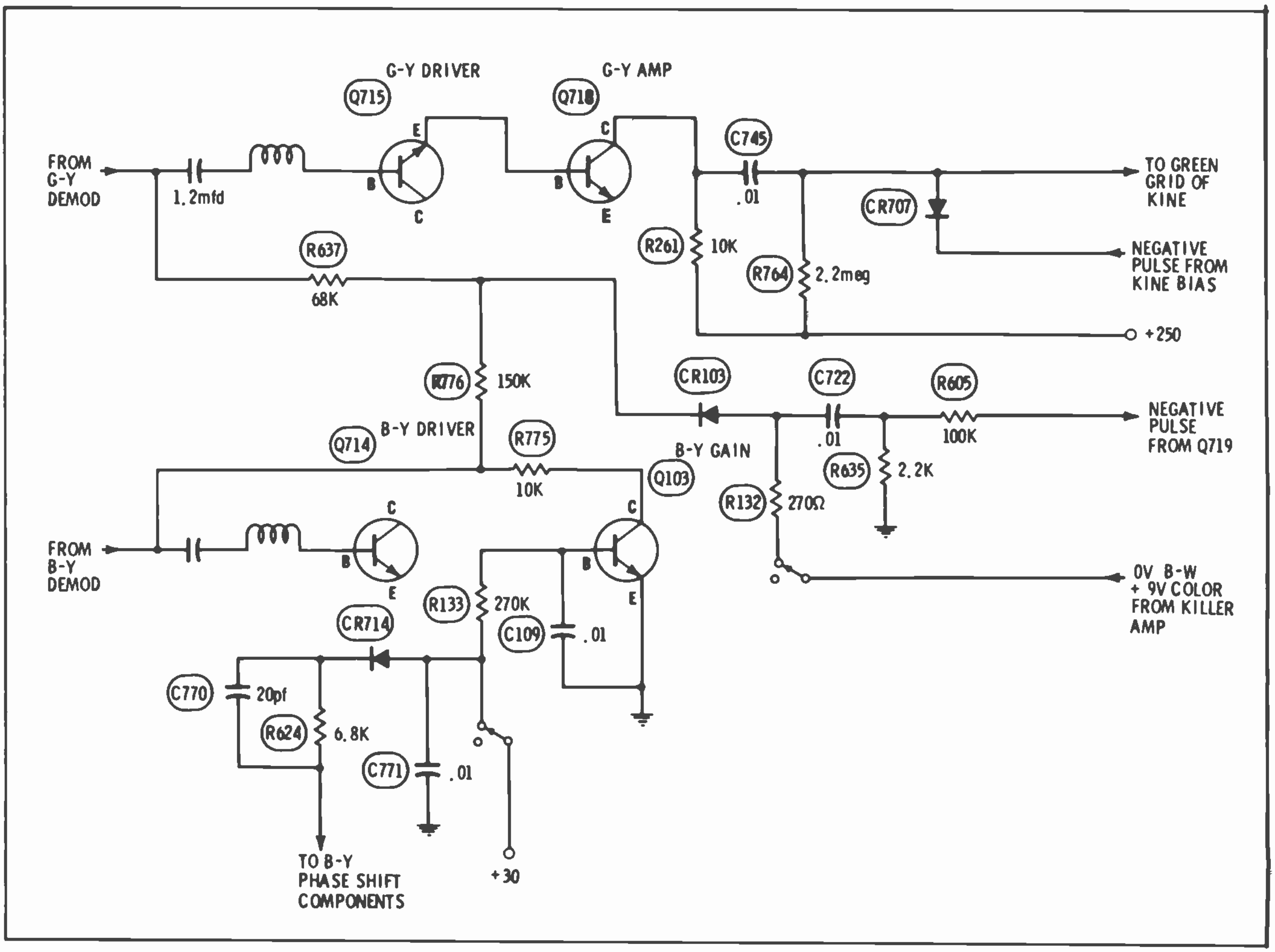


Fig. 3 The switching action of diodes and one transistor controls the ATC circuit of the RCA CTC44 chassis.

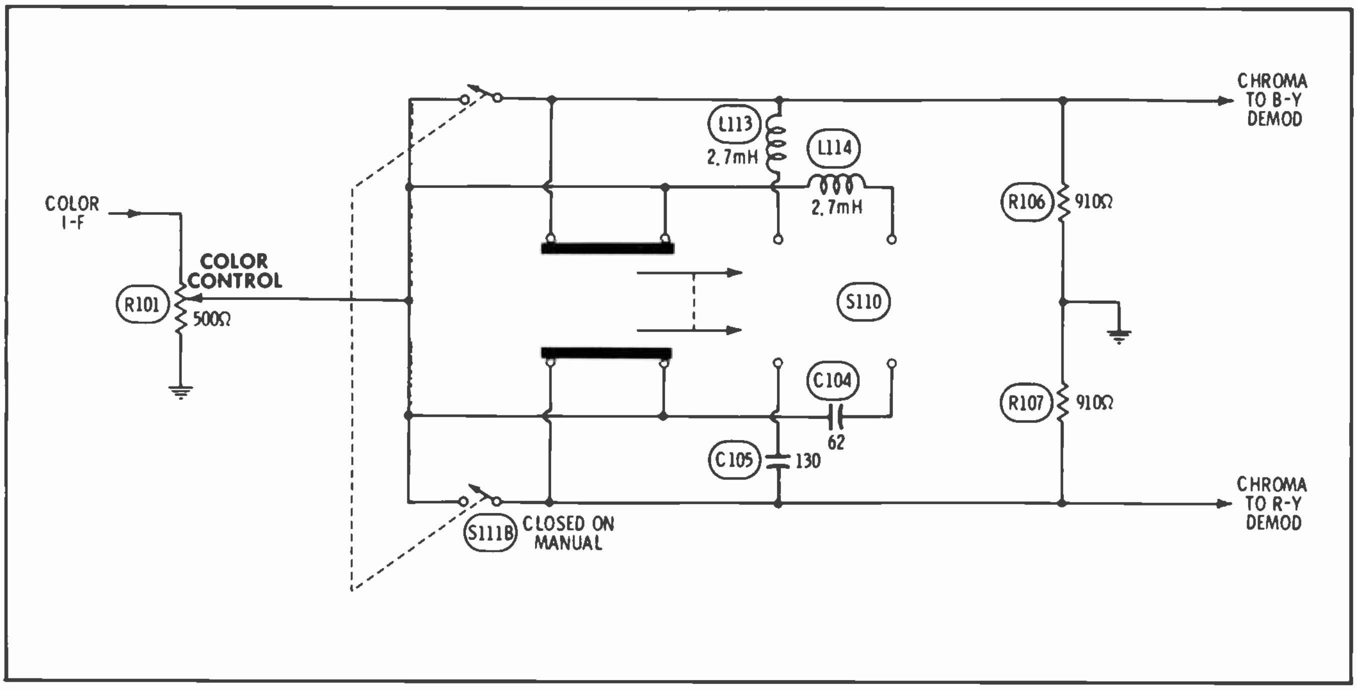


Fig. 4 Schematic of the Customatic Tint Lock circuit included in the General Electric KE-2 chassis

particular ATC system.)

Magnavox color TV chassis T950 employs an ATC circuit, the operation of which is based on the same principle as that used in their other models, but in which several significant changes have been incorporated.

Chroma signals from the color control are applied to the base of QA5. Output from the collector supplies the main 3.58-MHz signal to the demodulators, while the signal from the emitter goes through "Partial" and "Full" switching diodes (controlled by DC voltages) to the Red and Yellow gate transistors. It is important to notice that QA5 is the stage whose gain is eliminated by the color killer. A varicap (or varactor) diode is used in the phase-shifting network of the Preference Control. The diodes and varicap apparently are used to avoid signal radiation or degradation that could be caused by the long leads between the ATC board and the front panel, if direct switching were used.

Admiral's Color Monitor employs almost the same circuit as the one found in Magnavox color chassis T951, except a chroma driver (emitter-follower) is inserted before the "Partial" and "Full" switch and the gate transistors.

RCA Accu-Tint

An analysis of the version of the RCA Accu-Tint (A-T) system used in the CTC39X color chassis was presented in an article which begins on page 30 of the July '70 issue of ELECTRONIC SERVICING. As explained in that article, three basic changes occur in the chroma channel when the A-T switch is turned to the ON position. These changes are:

• The phase of the 3.58-MHz car-

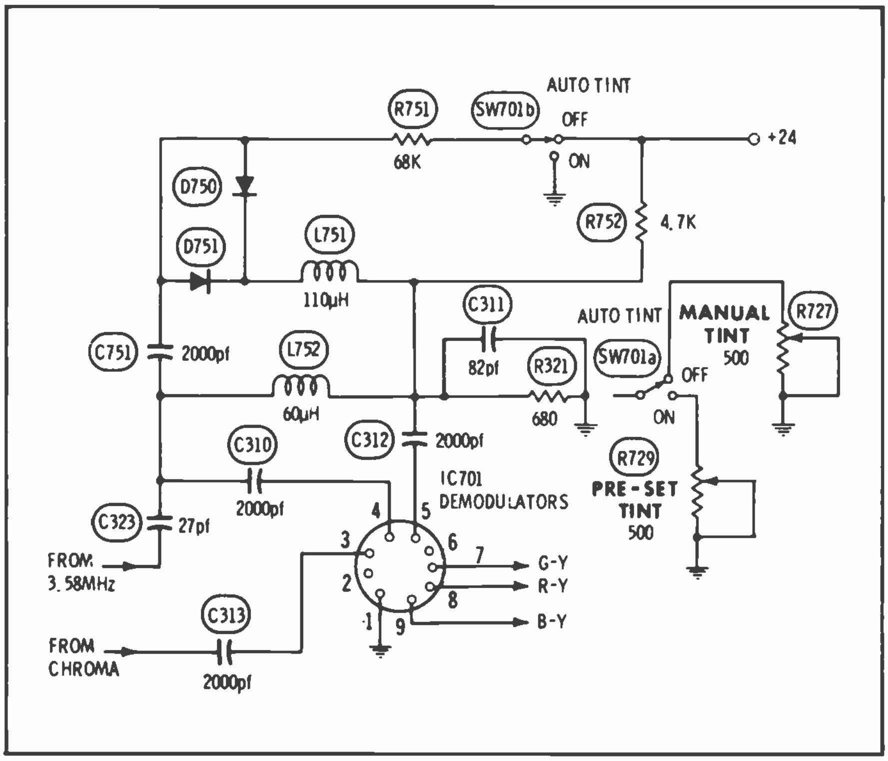


Fig. 5 The Auto Tint circuit designed into Electrohome's C9 chassis employs diode switching.

rier applied to the B-Y chroma demodulator is changed so that the output from this demodulator is separated from the R-Y signal by 124 degrees. Previously, beginning with their CTC16 chassis, RCA has used 105 degrees as normal phase difference between B-Y and R-Y signals. When a color-bar pattern is viewed on the screen of the TV and the Accu-Tint is switched to the ON position, the blue bars move about a half bar to the right (away from the red bars). This multiplies the number of color bars that are orange, or nearly orange.

- The output from the B-Y demodulator circuit is reduced about 33 percent.
- The hue, or gray-scale, of the b-w screen is changed from the normal blue-white to a brownwhite (sepia). This "warming" of the screen color "temperature" occurs only when the A-T switch is in the ON position during reception of a color signal (color killer inoperative).

These same changes occur in the chroma section of the RCA all-solid-state CTC44 chassis, but are the result of different ATC circuit actions. The schematic in Fig. 3 shows this part of the chroma circuit. When Accu-Tint switch S112 is turned to the ON position, +30 volts is applied to the anode of CR714, and, because its cathode is returned to ground through R624, L710 and the secondary of the tint transformer, it is forward biased and becomes a near-short circuit. This action connects C770 and R624 (through C771) in parallel with phase-shift components C758 and R778, which increases the phase shift of the 3.58-MHz signal that is applied to the B-Y demodulator.

The same +30 volts which is applied to the anode of CR714 also is applied, through R133, to the base of Q103. This forward-biases Q103, and the near-zero collector-to-emitter resistance grounds R775, reducing the output from the B-Y demodulator.

Warming of the screen color temperature is accomplished by adding opposing pulses to the grid clamp diodes connected to the green and blue control grids of the CRT. The +9 volts that appears at the collector of Q705 the killer amplifier transistor, when a color program is received is applied through S112 and R132 to the anode of CR103. This positive voltage forward biases CR103, which passes the clamping pulse (obtained from R-605, R635 and C722) on to R637 and R776. The negative pulses through R637 and R776 are amplified and inverted by the G-Y and B-Y amplifiers and finally applied to the anodes of CR707 and CR-708, which are connected to the green and blue grids of the CRT, respectively. The positive-going pulse on the green grid of the CRT and the anode of CR707 adds to the negative-going clamping pulse at the cathode of CR707 and drives the grid a few volts less positive than it would be without ATC action. Identical action takes place at CR708 and the blue grid of the CRT. Consequently, the brightness of both green and blue is reduced, while red remains the same.

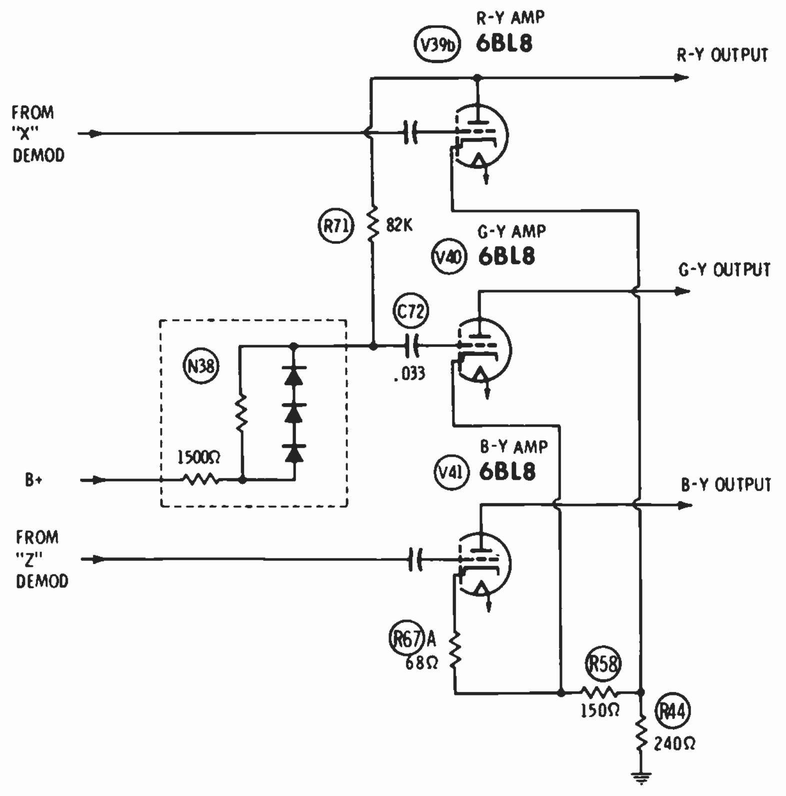


Fig. 6 A new type of matrixing is used in the production of the G-Y signal in Philco's 21ST90 chassis.

GE Customatic Tint Lock

Phases of the chroma signals applied to the B-Y and R-Y demodulators of the GE KE-2 color chassis are changed by the switch positions of the Customatic Tint Lock circuit. "Manual" and "Automatic" pushbuttons on the tilt-out control panel switch on the AFC and ATC functions in the "Automatic" position, and switch them off in the "Manual" position. Switching of the ATC action is shown in Fig. 4. In addition, S110 can be adjusted for no ATC action (110 degrees between R-Y and B-Y demodulators), moderate action (130 degrees) or maximum action (150 degrees).

Because of the weaker signal and the shorter length of lead, less radiation occurs when the chroma signal, instead of the 3.58-MHz carrier, is phase shifted by ATC adjustments.

Electrohome Auto Tint

Diode switching is used in the Electrohome hybrid chassis C9 to change the phase of the 3.58-MHz carrier applied to the integrated-circuit chroma demodulator, as shown in Fig. 5. The ON position of the Auto Tint switch (SW701, a front panel control) also switches in factory pre-set color and tint controls. These pre-set controls can be readjusted by a TV technician, if the customer is not pleased with the factory setting.

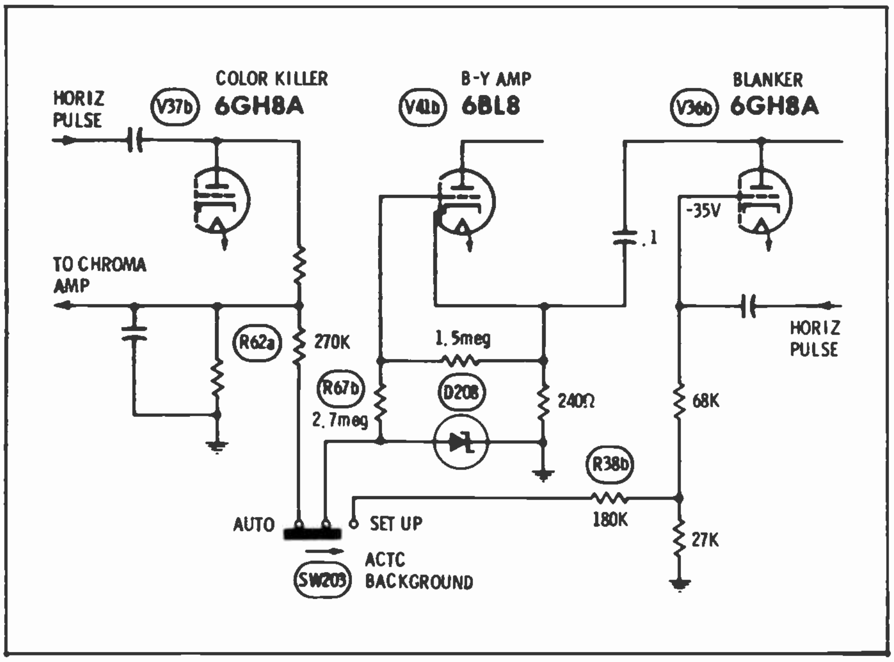


Fig. 7 Killer action in the Philco 21ST90 chassis during reception of color signals also changes the screen color to sepia.

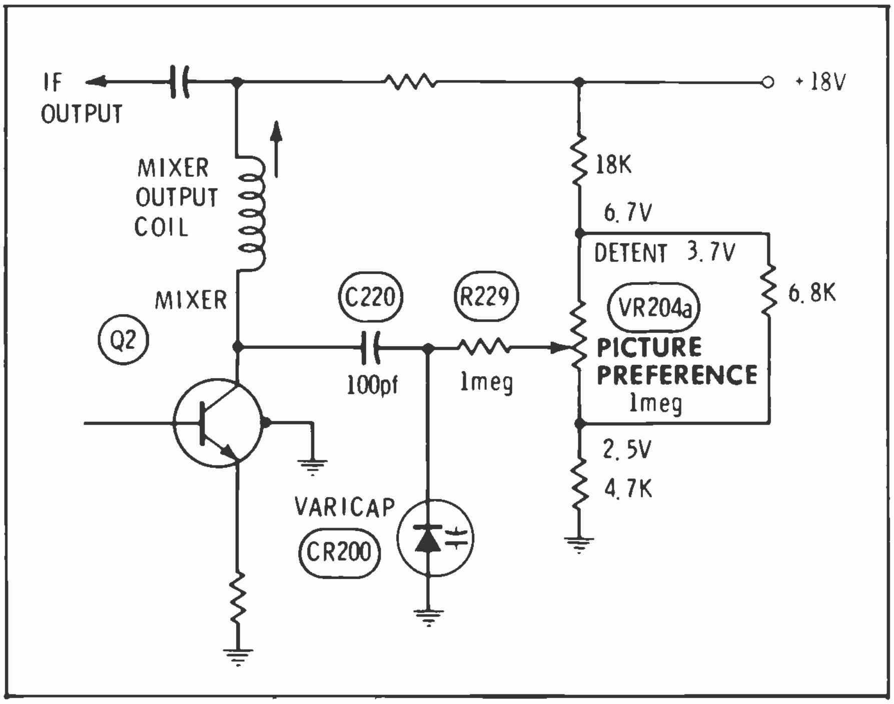


Fig. 8 A slight change of IF alignment by varicap tuning in the Philco 21ST90 chassis gives an effect similar to video peaking.

In the OFF position of SW701, the cathodes of D750 and D751 are more negative than the anodes; consequently, both diodes are a near-short, and L751 is connected in parallel with L752.

When SW701 is placed in the ON position, R751 is grounded and no DC voltage is applied to the anodes of the diodes. Because about +5 volts still is applied to the cathodes, the diodes are reverse biased, opening the circuit so that L752 is the only reactive element in the phase-shifting network. Consequently, the phase change is less than that produced with the switch in the OFF position.

Outputs from the R-Y and B-Y demodulators differ by 105 degrees in the OFF position of the Auto Tint switch and by 125 degrees in the ON position.

Philco Auto Tint

Philco does not have a switch or control for changing demodulator phasing, color balance or screen color of its new color receivers, but some of these features are built in.

Circuit features which produce actions that correspond roughly to a change in demodulator phasing are shown in Fig. 6. Most of the G-Y signal is developed in the common cathode circuit of the three tubes. R58 has been added to make the cyan hues less blue, and R67A slightly reduces the B-Y output. The G-Y amplifier needs additional signal from the R-Y stage to give proper matrixing, and this is obtained easily through R71. However, a better color reproduction of degraded picture information is produced if the R-Y signal fed through R71 is more negative than positive. The diodes and resistors in network N38 supply such a logarithmic voltage. Philco product information states that the interim quality of skin color is greenish; however when the tint control is readjusted to produce correct orange face hue, the green hues will have less blue contamination.

Philco also has an Automatic Color Temperature Control (ACTC) circuit which makes the raster color more sepia during reception of color signals. Fig. 7 shows the simplified schematic. Color-killer voltage is used to "identify" b-w or color programs for the circuit, and an ACTC set-up switch (SW203) is provided to switch in a standard voltage in place of the killer voltage, which might be undependable because of signal conditions during the adjustment.

In the SET-UP position of SW-203, a small amount of negative voltage, taken from the grid of the blinker tube, is supplied to the grid of the B-Y amplifier. To establish the desired b-w color temperature, the ACTC control is placed in the SET-UP position, and the screen color, or gray-scale, is adjusted in the usual way using the CRT drive and screen controls. Then the ACTC switch is returned to the AUTO position, in which the same amount of voltage established in the SET-UP position is supplied the grid of the B-Y amplifier by the color kilier when b-w programs are received. Thus, the b-w screen color is the usual blue-white.

Because no negative voltage is generated in the color-killer circuit during reception of color signals, the grid of the B-Y amplifier is less negative than during b-w reception or when the ACTC switch is in the SET-UP position. The less-negative grid causes the B-Y plate to become less positive, and because the blue grid of the CRT is coupled direct to the B-Y plate, less blue is seen in the raster and picture. Part of this same effect also is transferred to the green, because of the common cathode connections. Less blue plus slightly less green makes the raster more orange, to emphasize skin hues.

The next Philco feature to be discussed is not strictly an ATC circuit (see Fig. 8), yet it is intended to relax the rigid grip that automatic fine tuning has on the b-w picture quality. In the VHF tuner, a varicap diode is used to change the tuning of the mixer-output IF coil when the reverse voltage on the diode is changed by means of the Picture Preference Control (PPC), which is located on the front panel.

Color quality is changed very little by this feature because, as Philco states, the change in alignment is stopped before the color is degraded. A detent is provided, and the receiver has been factory aligned with the PPC control set at the detent. Visible effect on the picture quality is somewhat the same as that produced by a video peaking control, with normal sharpness obtained at the detent. However, it is possible that better vertical and horizonal locking might be obtained on weak or poor signals with the PPC control set to smear slightly the b-w detail, compared to normal locking obtained with a video peaking control.

Zenith Auto-Tint Guard

Several of the new Zenith models use the circuit shown in Fig. 9 to improve the varying skin hues produced by some color broadcasts. Operation is straightforward: The demodulation angle between R-Y and B-Y is increased when SW204 is closed and connects C262 in parallel with C1016.

Plug-in Circuits

The expanding use of plug-in circuit boards, panels or modules could prove to be one potential source of help in relieving the added burden imposed on TV technicians by miniaturization and the added complexity of solid-state designs.

Motorola Quasar removable panels and Zenith Duramodules (shown in Fig. 10) are well known and seem to have been well accepted.

RCA for some time has been testing the module concept in selected models of their b-w receivers. Recently, RCA announced their CTC49 all-solid-state portable color TV chassis, which uses eleven modules. The locations of these modules and other major parts are shown in Fig. 11. Fig. 12 shows the socket on board PW300 with its associated module (MAC) removed. Sound module PM22 and its socket can be seen in the background.

Pre-CRT matrixing for each color is accomplished in the three MAD modules. Fig. 13 shows both the original type, which uses conventional discrete components, and the newer version, which is an encapsulated "computer card" type that eventually will replace all the original versions of all the modules. RCA states that the latter type gradually will be phased into production during the next few months.

Warranty on the different brands of modules varies. Motorola still has a replacement warranty on the entire panel. Zenith warrants only the individual parts on the Dura-

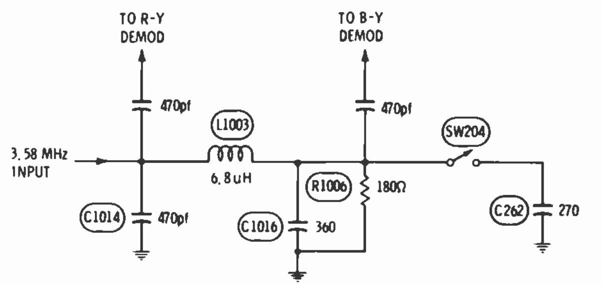


Fig. 9 The Automatic Tint Guard circuit in Zenith 40BC50 and 12B14C50 chassis adds a capacitor to the circuitry to alter demodulator phasing.

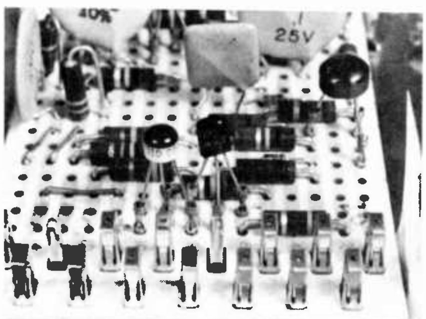


Fig. 10 Plug-in Duramodule used in Zenith's 4B25C19 chassis also features plug-in transistors.

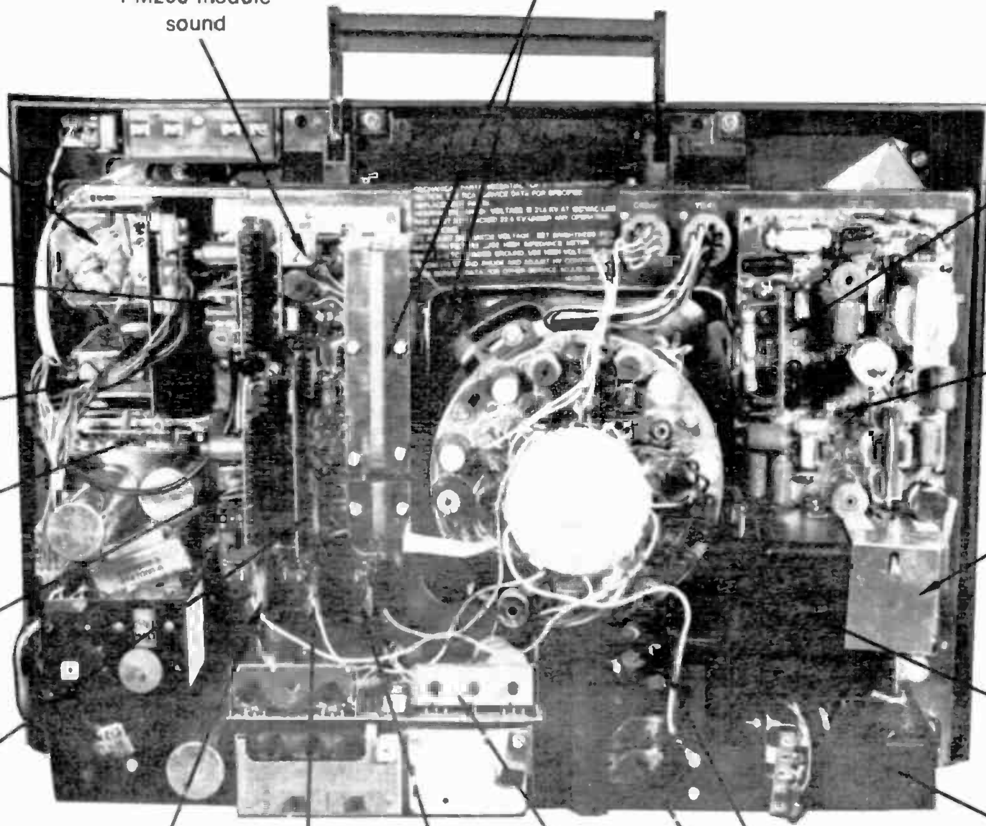


Fig. 11 Photo showing location of major components and the eleven modules that comprise the RCA CTC49 chassis.

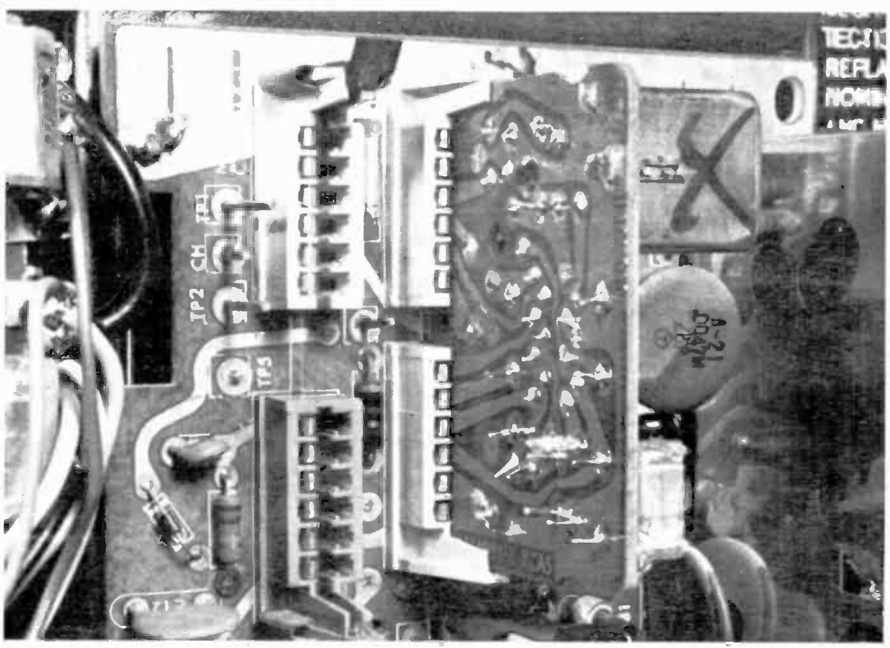


Fig. 12 Upper and lower sockets with module MAC removed from the RCA CTC49 chassis. Module PM200 and sockets are shown at right in background.

module (the transistors plug-in, as shown in Fig. 10). RCA provides a complete panel in exchange.

Motorola panels, Zenith Duramodules and the present type of RCA modules can be serviced by normal troubleshooting and repair techniques if the modules are out of warranty, or if repair seems desirable for any other reason. RCA encapsulated modules cannot be repaired with reasonable effort; therefore, replacement is necessary.

High Voltage and Automatic Brightness Control

Fewer tube-type, high-voltage rectifiers and regulators are being used because of the possibility of radiation if the high voltage becomes abnormally high.

Fail-safe circuits, tripler and quadrupler rectifier assemblies with capacitors and semi-conductor diodes in one unit, elimination of the high-voltage DC shunt regulator tube and a search for new methods of high-voltage regulation are the order of the day.

Automatic Brightness Control (ABL) circuits help high-voltage regulation indirectly by preventing excessive CRT current and, thus, any excessive decrease in high voltage (which, if severe enough, causes a noticeable amount of picture blooming).

Extra Admiral Diodes

Several chassis which regulate horizontal sweep and high voltage by changing the negative voltage applied to the grid of the horizontal output tube have added an extra diode in series or parallel with the one necessary for pulse rectification. The ultimate in such cautious design is the circuit in Fig. 14, an Admiral design that employs four diodes in series-parallel. Failure of any one diode (or in certain specific failures, up to three diodes) will not affect the regulation.

High-Voltage Regulation

Sweep/high-voltage regulation and ABL circuits employed in Hitachi CFA-450 and CSU-690 models are shown in Fig. 15. T703 is a saturable core reactor which changes impedance in accordance with the load on the high-voltage circuit. The Japanese-to-American translation of specifically how this circuit functions was too garbled to understand, and the actual physical construction of the reactor, which obviously is of an unusual design, was not ascertainable from the material supplied us. The Hitachi service data says that the impedance of T703 varies as the current through it changes. Thus, LX winding of T703 is in parallel with the primary of T704, the high-voltage transformer. Any reduction in the impedance of T704 because of increased load should be offset by an increase in the impedance of LX.

Understanding the operation of the ABL circuit is relatively easy if you imagine that R724 is connected to ground instead of +120 volts, and then realize that the voltage at the low end of the high-voltage transformer secondary is the source of negative voltage which varies in direct proportion to the average amount of rectified current passing through the tripler.

(Also of help in understanding this circuit is the following rule: "When rectified DC is obtained from a diode, the polarity of the voltage will be negative if it is taken from the anode, and positive if it is taken from the cathode.")

In the Hitachi ABL circuit, the negative voltage at the junction of C723 and R724 partially cancels the steady DC voltage through R724 from the +120-volt source. The more CRT current, the lower the positive voltage that is applied to the CRT grids. Since the cathodes of the CRT are operated at a fixed positive voltage, the CRT is

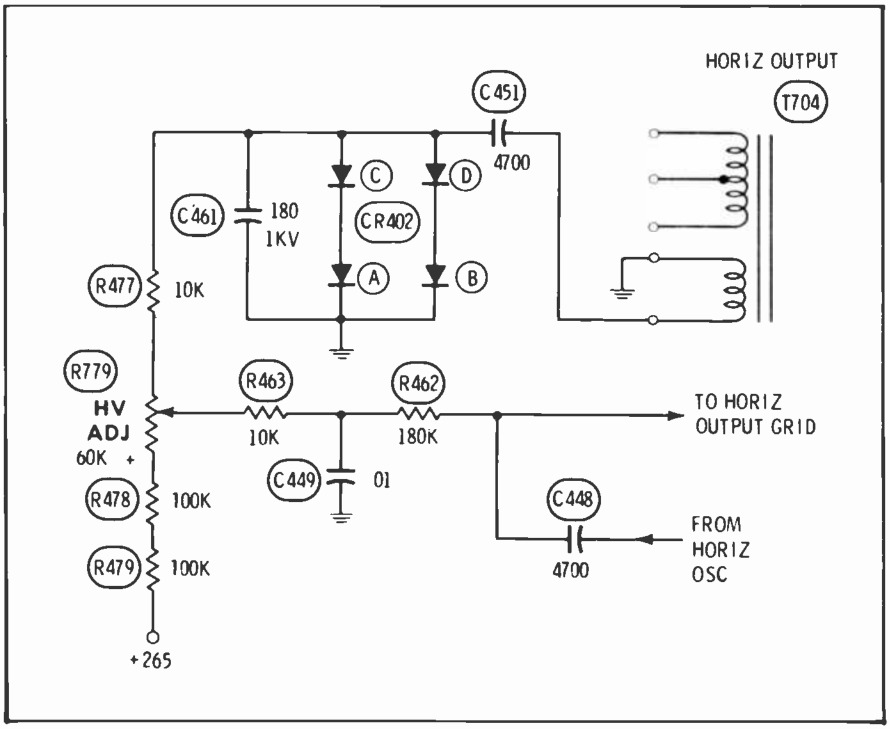


Fig. 14 To help insure adequate high-voltage regulation regardless of failure, four diodes are wired in series-parallel in the high-voltage regulator of Admiral's 19H10 chassis.

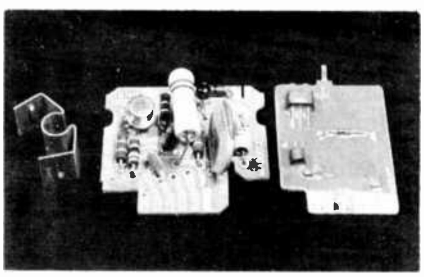


Fig. 13 Early-production MAD module and heat sink on the left, and a new type of encapsulated MAD module on the right.

biased more negatively with increased CRT current; therefore, the CRT is prevented from drawing an excessive amount of current.

Zenith Automatic Brightness Limiter

The ABL system in Zenith 40BC50 and 4B25C19 chassis does not monitor the actual high-voltage current; instead, the operation depends upon the amount of ripple in the focus voltage supply. The more high-voltage current drawn from the rectifier-filter assembly, the higher this ripple will be. Ripple (AC content) from the focus supply voltage (see Fig. 16) is coupled through C277 (inside the rectifier assembly, in some cases) and R338 to the base of Q204, which has no bias applied to it except for the ripple. If the ripple at the base is low, Q204 does not conduct and, consequently, the collector voltage, which is connected to a video amplifier stage, remains at almost the same value as the supply voltage.

When the ripple amplitude is high because of increased high-voltage current, Q204 is saturated by the large forward bias developed by the ripple, and the collector-to-emitter junction becomes a low-value resistance, which causes the collector voltage to drop, thereby increasing the negative bias on the CRT via the video amplifiers.

CRT current is limited to about 1.5 milliamperes by this ABL action.

Control of Color Saturation and Tint by Variable DC Voltages

Color and tint adjustments which are controlled indirectly by variable DC voltages instead of directly by manually adjusted controls are employed in '71 chassis of at least three manufacturers.

Such a system offers two distinct advantages: First, because the chroma IF and 3.58-MHz color subcarrier signals do not have to be routed over long wires to front-panel manual controls, radiation from these signals is minimized. Secondly, remote control by variable DC voltages does not require relays and/or motors, as does remote adjustment of manual controls.

Sylvania

One example of indirect control of color saturation and tint is found in Sylvania's all-solid-state E01

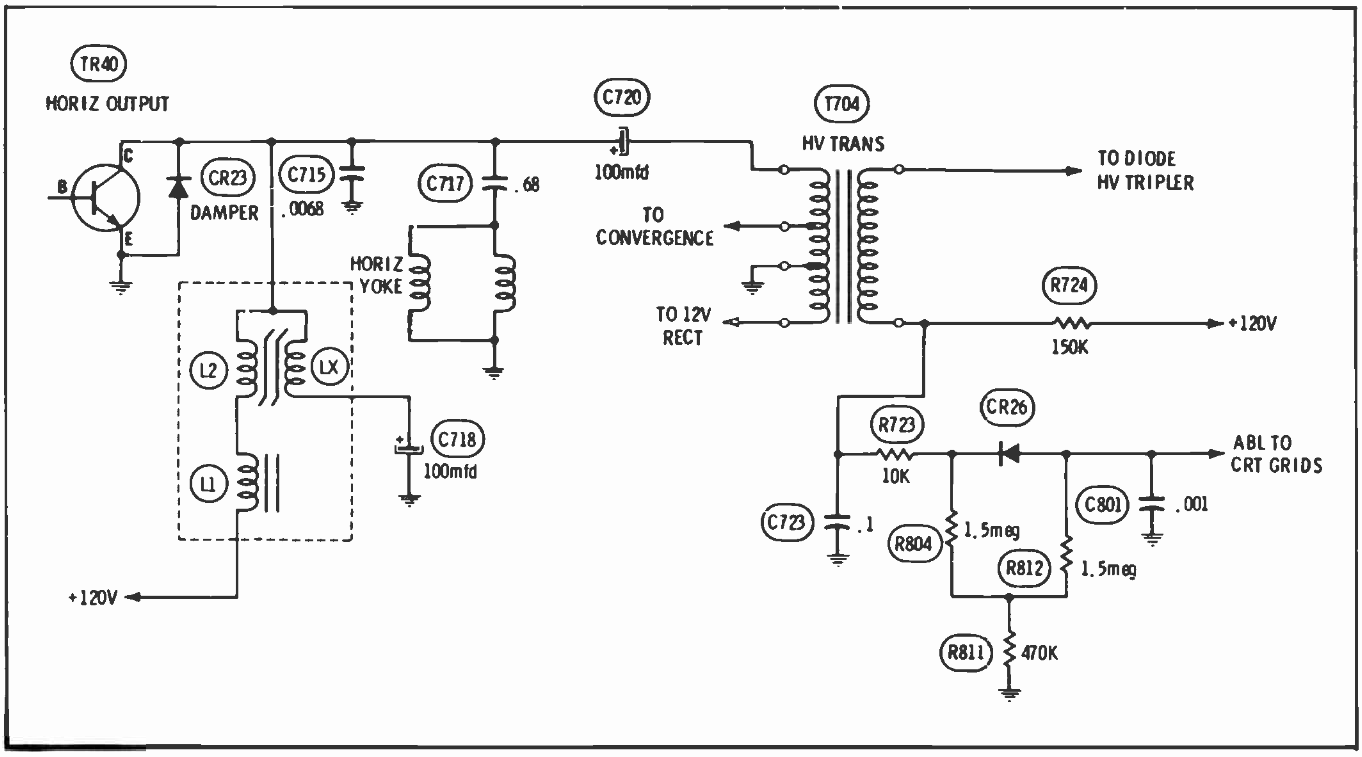


Fig. 15 Hitachi models CFA-450 and CSU-690 employ a saturable-core reactor as the high-voltage regulator. CRT current through the high-voltage tripler determines the amount of bias supplied the CRT by the automatic brightness limiter (ABL) circuit.

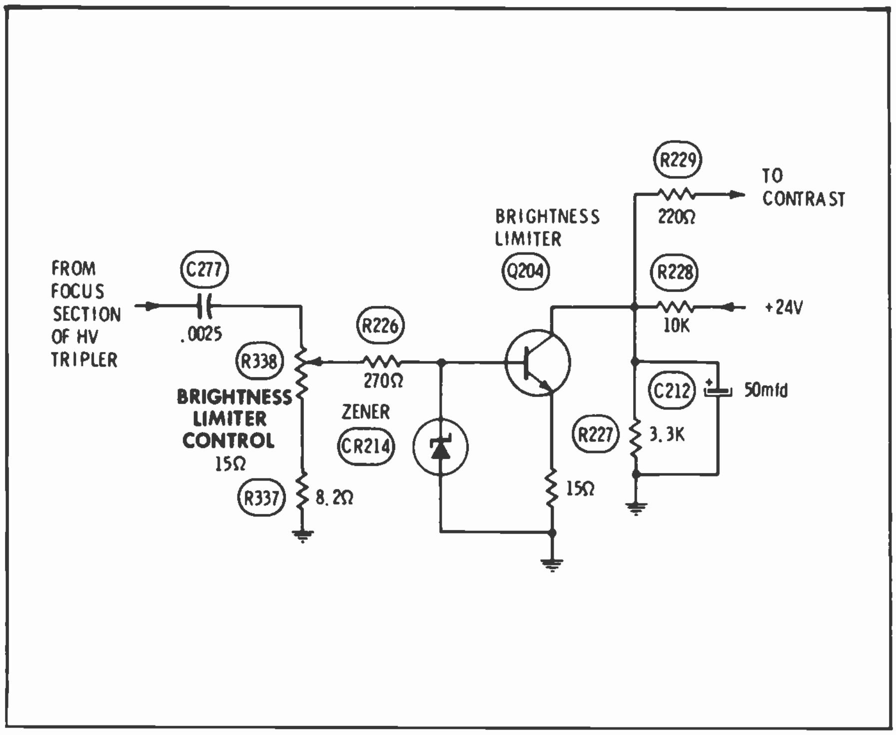


Fig. 16 The ABL circuit in Zenith's 40BC50 and 4B25C19 chassis senses the increased ripple amplitude of the focus supply voltage (resulting from excessive brightness). The change in ripple amplitude is used to control Q204, the collector voltage of which varies the conduction of the video amplifiers and, consequently, also varies the bias applied to the CRT. Q204 functions as both a rectifier and amplifier. The video amplifier voltage is directly coupled to the CRT, in the correct polarity to reduce the brightness level.

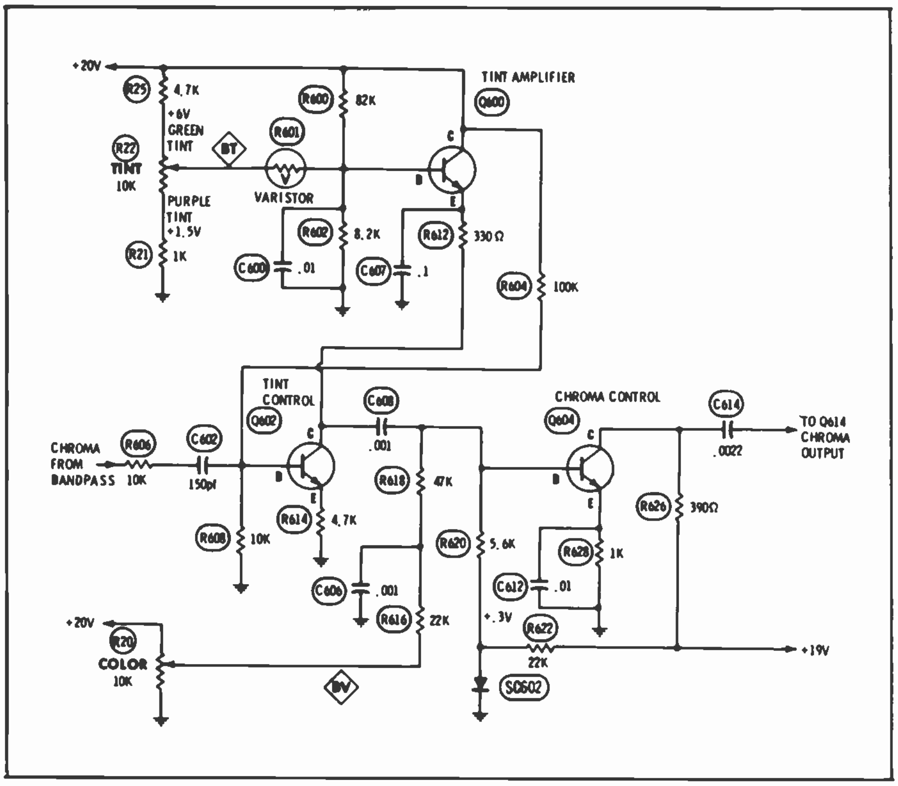


Fig. 17 Schematic diagram of the color and tint circuits in Sylvania's E01 color TV chassis. The internal capacitance of Q602 is the key to the operation of this tint-control system. Refer to the text for an explanation.

chassis. The associated circuitry is electrically located between the 2nd chroma amplifier and the chroma output stage, as shown in Fig. 17.

How the tint control circuit performs its function can be explained best by examining the results of adjusting tint control R22 to each of its extreme positions.

The actual change in the phase (tint) of the chroma signal is accomplished primarily by varying the conduction of transistor Q600. When R22 is repositioned CCW toward the position that produces greenish flesh tones, more forward bias (positive) is applied through varistor R601 to the base of Q600, increasing its conduction. The resultant increase in the emitter current of Q600 produces a larger voltage drop across emitter resistor R612, the "negative" end of which is connected to the collector of Q602. With a more positive voltage applied to its collector (relative to its base), and with a constant forward bias between its emitter and base, Q602 conducts and amplifies the chroma signal applied to its base.

Because of the inherent phase shift between the base and collector of a transistor, the amplified chroma signal at the collector of Q602 is 180 degrees out of phase with the input chroma signal at its base.

Thus, manual adjustment of the tint control to one of its extreme positions has produced a change in the DC voltages that control the conduction of the tint control transistor and, consequently, the phase of the chroma signal in relation to that of the reference subcarrier.

When tint control R22 is adjusted to the extreme CW position, which produces purple-tinted flesh tones, less forward bias is applied between the base and emitter of Q600, it conducts less, and the voltage at the "negative" end of the resistor in its emitter circuit becomes less positive. As a result of this action, the voltage applied to the collector of Q602 becomes sufficiently less positive to cut off conduction of this transistor.

Although none of the chroma signal applied to the base of Q602 is amplified when this transistor is biased off, some of the chroma signal does feed through the capacitance between the base and collector of the transistor. However, unlike the 180-degree phase shift resulting from normal transfer of an amplified signal from base to collector, the chroma signal fed through the base-to-collector capacitance is **not** shifted in phase relative to the chroma reference subcarrier generated by the receiver.

Using this system, the phase of the chroma signal at the collector of Q602 can be varied over a range of almost 180 degrees relative to the phase of the chroma reference subcarrier, the exact phase shift depending on the vectorial addition of 1) the phase of the chroma signal transferred to the collector of Q602 by transistor amplification (180 degrees shift) and 2) the phase of the chroma signal transferred to the collector via the base-to-collector capacitance (no phase shift).

Color control action is by the more direct method of reducing the normal forward bias of Q604 to obtain less transistor gain. Diode SC602 provides a temperature-controlled minimum forward bias for better stability.

RCA

A differential amplifier is employed in the tint circuit of the new all-solid-state RCA CTC44 chassis, as shown in Fig. 18. Transistors Q711 and Q712 appear, at first glance, to be in parallel; the bases are separated by a coupling capacitor, but the signal input is of the same amplitude and phase at both bases. Both collectors are connected together. The difference between the two amplifiers is their frequency responses, which are determined by the components in their emitter circuits.

R603, in the emitter circuit of Q711, is paralleled by C749. This RC combination reduces the high frequencies from emitter to ground; consequently, amplifier Q711 increases the amplitude of high-frequency signals more than that of lower-frequency signals (opposite to the response of the degeneration or feedback). Such a stage is the equivalent of a high-pass filter, which causes a leading phase characteristic.

The frequency response of Q712 is just the opposite that of Q711; R610 in the emitter of Q712 is paralleled by an inductance instead of a capacitance. Because the emitter circuit thus attenuates the low frequencies, the amplifier produces increased gain at low frequencies. This is the equivalent of a low-pass circuit, which causes a lagging phase shift.

The combined output at the collectors will have the same phase as the input signal, if the phase shift in each amplifier is equal and opposite and both amplifiers have the same gain. The opposing phase shifts are cancelled by vectoral addition. This is the desired condition when the hue control is in the center of its range.

Any difference in the amplitude of the output signal from one amplifier compared to the output of the other causes a change in the phase of the signal at the paralleled collectors. The forward bias of Q-711 is fixed, while the forward bias of Q712 can be changed by adjustment of tint control R123.

Rotation of the tint control in the direction that increases the forward bias of Q712 also increases the gain of Q712. At the same time, the increased emitter current of Q712 raises (by way of the common cathode resistor, R606) the emitter voltage of Q711, thus decreasing the gain of Q711. At the collectors, the phase of Q712 is dominant; the output signal has a lagging phase. The exact phase is determined by the difference in the gain of the two amplifiers, at least until Q711 is biased to cutoff or Q712 reaches satura-

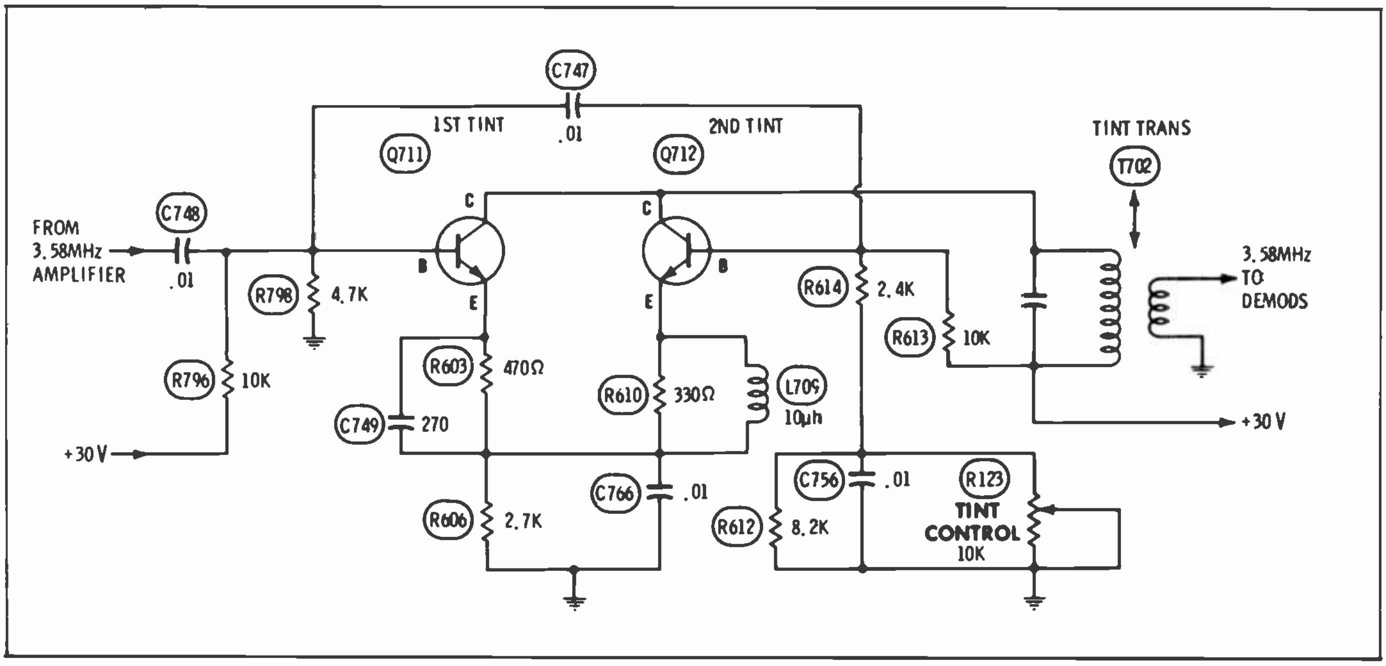


Fig. 18 The tint circuit of the RCA CTC44 chassis, shown here, is operated by voltage and, therefore, can be used in motorless remote-control applications.

tion, and this extremity is prevented by the resistor values in the base circuit of Q712.

Rotation of the tint control in the opposite direction reverses this action. The bias and gain of Q712 are reduced below that of Q711; thus, the output phase is leading.

Control of color saturation in the RCA CTC44 chassis is accomplished by a voltage-controlled variable voltage divider and by a change of bias which, in turn, changes the gain of the 3rd chroma IF amplifier transistor (see Fig. 19).

The arms of the voltage divider consist of R724 (1K ohms) and diode CR709, whose internal resistance decreases as its forward bias is increased. The resistance of CR-709 varies from 6.8K ohms at about .68 volt, to 750 ohms at about .75 volt, and to even lower values when supplied with higher forward bias voltages. The anode of CR709 is bypassed to ground. This enables the diode to function as the shunt leg of an AC voltage divider, in addition to being the source of a DC voltage that changes the emitter voltage (forward bias) of Q704 and, thus, its gain.

Transistor Q704 is operated as a grounded base type of amplifier, in which the input signal is applied to the emitter. When the color control is turned down (rotated CCW) maximum DC voltage is applied to the anode of CR709. The DC path from the cathode to ground is completed through R724 and L708; consequently, maximum voltage is dropped across CR709, and the internal resistance is at a minimum. This voltage divider action reduces the chroma signal (from C726/L708) before it is supplied to the emitter of Q704. Emitter DC voltage is at maximum because of the additional voltage entering through CR709; therefore, the forward bias and the gain of Q704 are both at minimum.

Clockwise rotation of the color control reduces the DC voltage applied to the anode of CR709, increasing the resistance of the diode (shunt arm of the voltage divider) and supplying more chroma signal to the emitter of Q704. Because the increase in the resistance of CR709 allows less DC voltage to reach the emitter of Q704, the emitter voltage decreases (forward bias is increased) and the transistor produces higher gain.

To avoid bandwidth and phase changes, the load across tuned circuit C726/L708 needs to be relatively constant during color saturation adjustments. This is accomplished as a result of the fact that while the resistance of CR709 is minimum when the color control is turned down, the resistance of the path through R742 and the emitter-base junction of Q704 is maximum. These conditions are reversed when the color control is turned up; thus, a fairly constant load is applied to the tuned circuit regardless of color control variations.

Other New or Changed Chroma Circuitry

General Electric—Color and Tint Control

After analysis of the preceding relatively complicated circuits, the manually-operated color and tint circuits of the General Electric N-1 portable color receiver (Fig. 20) seem even more simple than they are.

The color control action is conventional. The tint control is turned

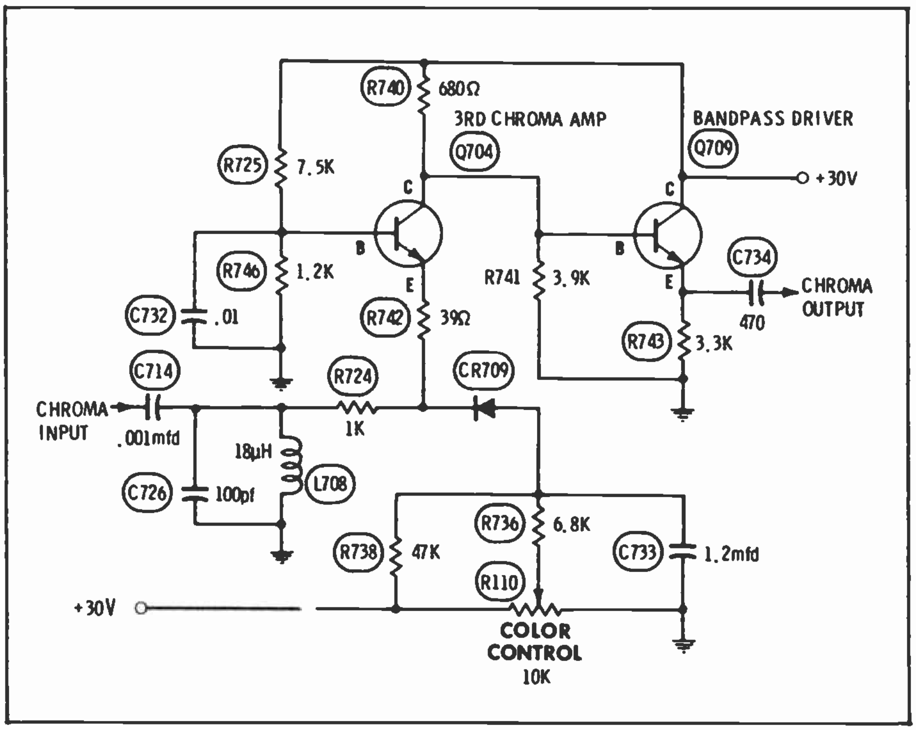


Fig. 19 The resistance of diode CR709 and the bias of Q704 are changed by one voltage to control color saturation in RCA's CTC44 solid-state color chassis.

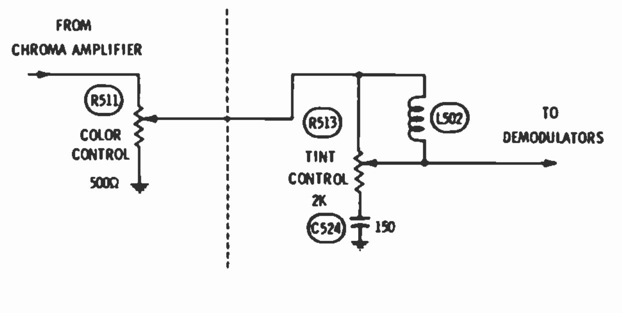


Fig. 20 Control of tint in GE's N-1 chassis is accomplished by phase shifting the chroma signal.

completely clockwise, L502 is shorted out and C524 has little effect because it is isolated by the 2K-ohm resistance of the control. Therefore, the phase of the chroma signal is unchanged.

Full CCW rotation of the tint control changes the circuit to a low-pass filter. The upper leg of the filter is L502 in parallel with R513, and the shunt leg is C524. This filter shifts the phase of the chroma signal by more than 100 degrees (lagging).

Sylvania—Chroma Sync Locking

A phase detector, DC amplifier and a varicap (varactor) diode are used in the Sylvania D16-2 color chassis to lock the 3.58-MHz crystal oscillator (Fig. 21).

The phase detector action is conventional except for the bias voltage for Q620, which is obtained **through** the phase detector.

Q620 actually consists of two cascaded emitter-follower transistors (Darlington pair) in one case. This type of operation provides very high input impedance. Error-correction voltage from the phase detector, plus forward bias, is applied to the base of Q620. Amplified error-correction voltage from the collector of Q602 is applied to the cathode of varicap diode SC608, whose anode is connected to a regulated voltage of about +7.6 volts, to establish reverse bias. The internal capacitance of SC608 changes when the voltage across it is changed. This capacitance, which is in series with the 3.58-MHz quartz crystal, changes the frequency of the color oscillator to accomplish and maintain chroma locking.

Electrohome—Pulse Clamping of CRT Grids

Pulse clamping of the DC voltages at the grids of the CRT in the Electrohome C7 chassis is provided by the circuit shown in Fig. 22. The two purposes of such a circuit are 1) to prevent variations in the plate voltages of the –Y amplifier tubes from changing the b-w screen color, and 2) to reset the DC voltage at each grid of the CRT during each horizontal cycle.

Both a negative-going pulse from the blanker tube and positive DC from the CRT bias control are present at the junction of R752, R755 and the cathode of D707. The CRT bias control supplies (through R755 and R752) a DC voltage of slightly more than the amount eventually needed at the grids of the CRT.

Capacitor C736 accepts and stores a charge because of the difference in DC voltages between the plate of the –Y amplifier and the grid of the CRT. Because the tip of the negative-going horizontal pulse at the cathode of diode D707 is less positive than the voltage at the anode (stored in C736), D707 conducts for the duration of the pulse and charges C736 to a new and lower voltage.

If a series of non-symmetrical chroma pulses at the grid of the CRT shifts the operating center to make the CRT grid more positive than it was during the previous blanker pulse, the next horizontal pulse will forward bias D707, whose conduction will lower the voltage on C736.

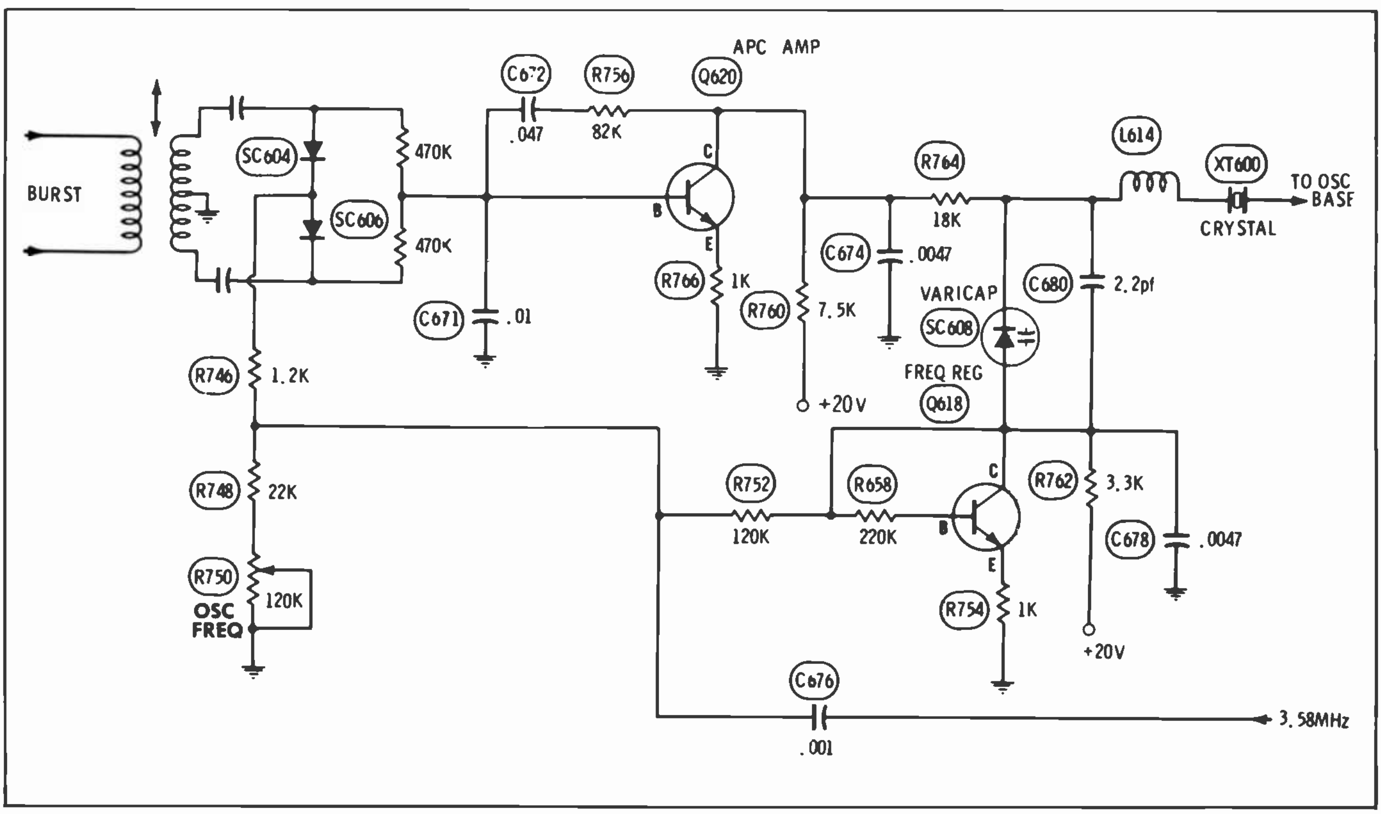


Fig. 21 Sylvania's D16-2 chassis uses a varicap diode to lock the color oscillator. A "Darlington pair" amplifier provides DC voltage gain between the phase detector and the varicap diode.

If the voltage at the grid of the CRT is not changed by the chroma signal, C736 is charged to a slightly higher voltage by the CRT bias voltage applied through R752. This slight increase is reduced by the next horizontal pulse.

Thus, undesirable changes in screen color are prevented by restoring the DC voltage on each grid of the CRT. As explained previously, this is accomplished during horizontal blanking, when no chroma information is present.

General Electric—Pulse Clamping of CRT Grids

A more elaborate system of pulse clamping the DC voltages at the grids of the CRT is employed in the General Electric KE-2 color chassis. Clipped and shaped sync pulses are used (Fig. 23) in place of pulses obtained from the horizontal sweep circuit.

About the same DC voltages are produced at the CRT grids when no station is received (no sync pulses) as are produced when station and sync pulses are present. Because the action is the same for all three grids of the CRT, only the effects on the blue grid will be explained here.

This pulse clamping action is easy to understand if two basic conditions are understood: 1) Transistor Q752 is an open circuit when there are no pulses present, and 2) the transistor is a near-short circuit when the sync pulse is present at the base.

When no signal is being received and, consequently, no sync pulses are present in the circuit, Q752 is open, and the grid voltage of the CRT is obtained through R762 from a +200-volt tap on the voltage divider comprised of R765, R766 and R767. With +325 volts applied to its cathode, diode CR752 is an open circuit, and the grid of the CRT floats at +200 volts because there is no chroma signal to cause it to drift. This action keeps the screen color and brightness stable.

When a signal is received, negative-going sync pulses are applied to C750, which is of sufficiently small capacitance to sharpen the pulses to a duration of 3.5 μs. The sharpened pulses then are applied to R759 and CR750, whose function is to clip the waveform so the amplitude does not exceed .8 volt PP. This shaped and clipped negative-going pulse at the base becomes, through normal transistor action, a larger positive-going pulse at the collector. Because the base-emitter junction of Q752 (NPN) has no forward bias, the transistor is an open circuit until the pulse from Q751 arrives through C752 and drives Q752 into complete saturation (a near-short circuit).

With Q752 a near-short circuit, the resistive values of the voltage divider are changed. R756, through Q752, is in parallel with R765 and R766, and the DC voltage at point 3 increases from the "no-pulse" voltage of +167 up to +190 volts. The same +190 volts also appears at point 1 (because Q752 is a near-short circuit). Between sync pulses, CR752 is biased off by the combination of +325 volts applied to its cathode and the +200 volts at its anode.

However, during the time the sync pulse is present the voltage on the cathode of CR752 drops to +190 volts, and the diode is biased on.

Normally, with CR752 conducting, the +200-volt charge on C740 would drop to about the +190-volt level present at the cathode of CR-752. However, a simultaneous 25-volt increase in the voltage applied to the anode of CR752 (via R762) attempts to increase the charge on C740. The result of these two opposing forces is an increase of the CRT grid voltage to about +200.06 volts.

When the pulse is gone, the CRT grid voltage slowly decreases to the original +200-volt level. The discharge of C740 from 200.06 volts to 200 volts is slowed by the combination of R762 and the storage action of C753.

If a chroma signal of non-symmetrical waveshape increases the average DC at the CRT grid above +200 volts, the pulse conduction path through CR752 and R769 is dominant over the charging path through R762, so the grid is restored by the next pulse back to the normal +200.06 volts. This action prevents any build-up of higher positive voltages on the grids of the CRT. Also the capacitive coupling prevents changes in the DC voltages at the plates of the -Y amplifiers from affecting the voltages on the grids of the CRT. Such changes are undesirable, because they would change the b-w screen color.

General Electric advises that any measurement of DC voltages on the grids of the CRT be made with a VTVM having an input resistance of no less than 100 megohms, because a lower input resistance interferes with clamp action. Similarly, a scope should not be connected to a grid of the CRT; instead, connect

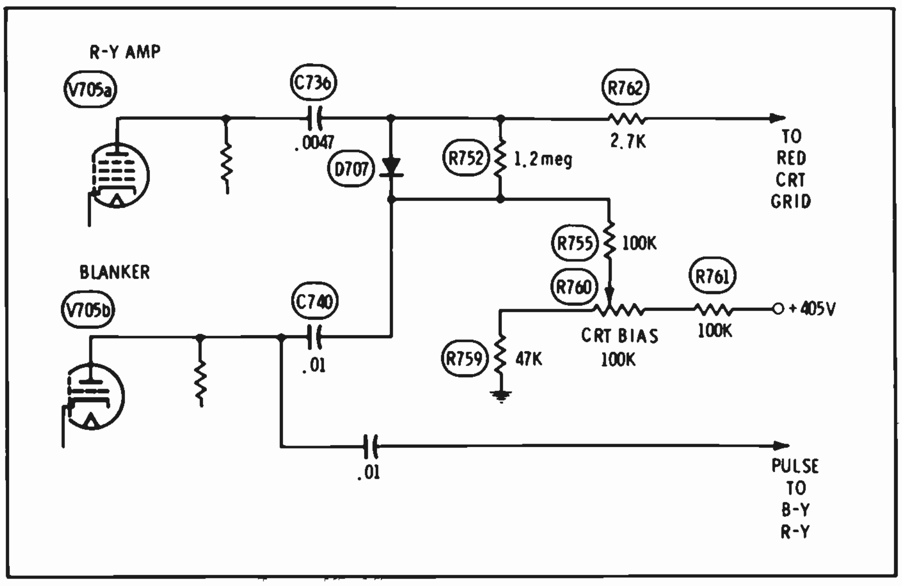


Fig. 22 Voltage clamping of the CRT grids in the Electrohome C7 chassis is by means of "pulsed diodes."

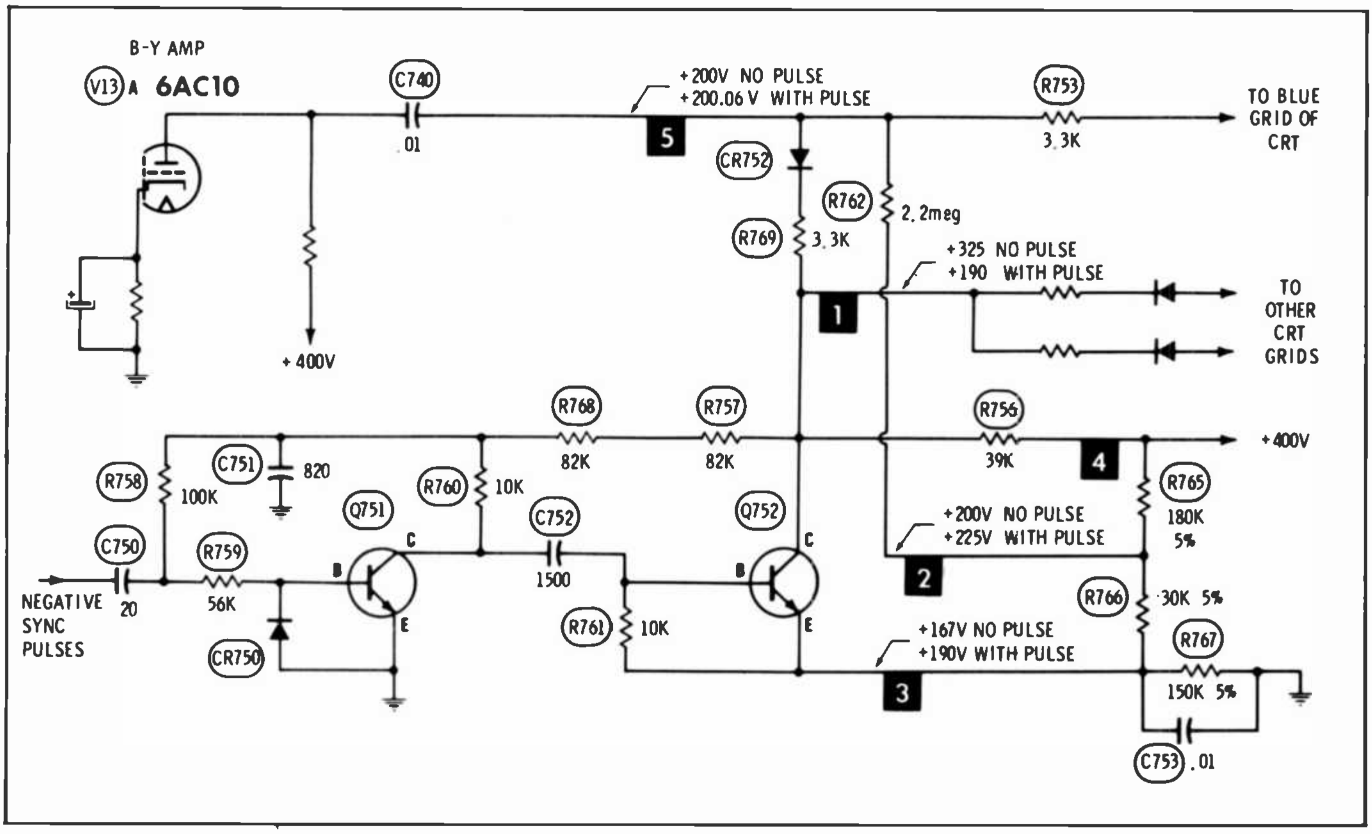


Fig. 23 The General Electric KE-2 chassis also clamps the DC voltages on the CRT grids by diodes controlled by pulses from the sync circuit.

it to the plate of the preceding –Y amplifier, where a small pulse should be found. It is advisable to observe the waveform both off channel and on a normal b-w program, to help you become familiar with the different waveforms thus obtained.

Remember, in any of these systems in which the grid circuits of the CRT are "pulsed", only a small percentage of the applied pulse ever reaches the grid of the CRT, or appears on a scope connected to the grid. The circuit action is identical to the peak-reading rectification of a pulse by a series rectifier circuit which has a large filter capacitor in parallel with the load. Only a small "ripple" remains. The capacitor connected from each –Y amplifier to the associated CRT grid performs the same function as the filter capacitor of the pulse rectifier.

A shorted clamp diode increases the DC voltage applied to its associated grid to about 275 volts, and increases the voltage to the other two grids to about 230 volts. The screen will be excessively bright over the area covered by the affected color, often to the point of blooming.

Panasonic—Auto-Color, Subcarrier Generation, and Pulse Clamping

Several new circuits found in Panasonic's CT98D color chassis are shown in the block diagram of Fig. 24.

When the Auto-Color circuit is switched on, pre-set color and tint controls are substituted for the conventional front-panel controls. In this chassis no provision is made for ATC by changing the demodulator phasing.

The "X" level control between the 2nd color amplifier and the diode-type "X" demodulator is provided to balance the red and blue chroma signals. The effect of a small amount of ATC is obtained if the "X" level control is adjusted to maximum.

"X" and "Z" transistor amplifier stages are used to increase the amplitude of these respective signals before matrixing and further amplification in the –Y amplifier tubes.

The 3.58-MHz subcarrier is generated by ringing the 3.58-MHz crystal with amplified burst. Loss of burst causes loss of the 3.58-MHz subcarrier, which, in turn, greatly weakens color "confetti" during b-w programs. However, because a color killer is used in this model to completely eliminate color snow when a color program is not being received, the absence of color confetti during a b-w program does not by itself indicate a loss of chroma subcarrier.

Pulse clamping of the DC voltages at the grids of the CRT by a method similar to that used by RCA and Electrohome is employed in the CT98D color chassis, with the exception that the resistor usually connected between the grid of the CRT and B+ in RCA and Electrohome chassis is connected between the grid of the CRT and the plate of the –Y amplifier in the Panasonic chassis.

RCA—Chroma Plug-In Modules

Pre-CRT matrixing of chroma and video signals, and pulse stabilization of b-w screen color temperature are functions of the MAD plug-in modules employed in RCA's CTC49 all-solid-state color chassis (Fig. 25). The early-production MAD module, which uses conventional components and a heat sink, and the newer encapsulated "computer card" MAD modules are shown Fig. 26.

Three identical MAD modules are used, one each for red, blue and green drive to the cathodes of the CRT. As shown in Fig. 9, video is applied to the emitter circuit of Q1 through the paralleled paths of R7, R6, C2 (for increased high-frequency response) and R336 to R335 (video drive control). There is no phase inversion of the video, because Q1 is a grounded-base amplifier to this signal.

The chroma –Y signal is applied to the base of Q1. Matrixing of the video and chroma signals is accomplished inside Q1. The output at the collector of Q1 is a complete red, blue or green color signal.

No signal is applied to the three grids of the CRT; only a DC voltage, which can be varied by the "kine bias" control, is present on the grids.

The collector of Q1 in each MAD module is coupled directly through a 3.3K-ohm resistor (inside the CRT socket) to the corresponding cathode of the CRT; therefore, stability of the DC voltages on the collectors of all three Q1 transistors will determine whether or not the b-w screen color remains unchanged. Stabilization of these collector voltages is accomplished by a pulse clamping circuit, which also supplies horizontal blanking.

A large positive-going horizontal pulse obtained from a winding on the high-voltage transformer is applied through C309; R340, R341 and R331 reduce the amplitude of

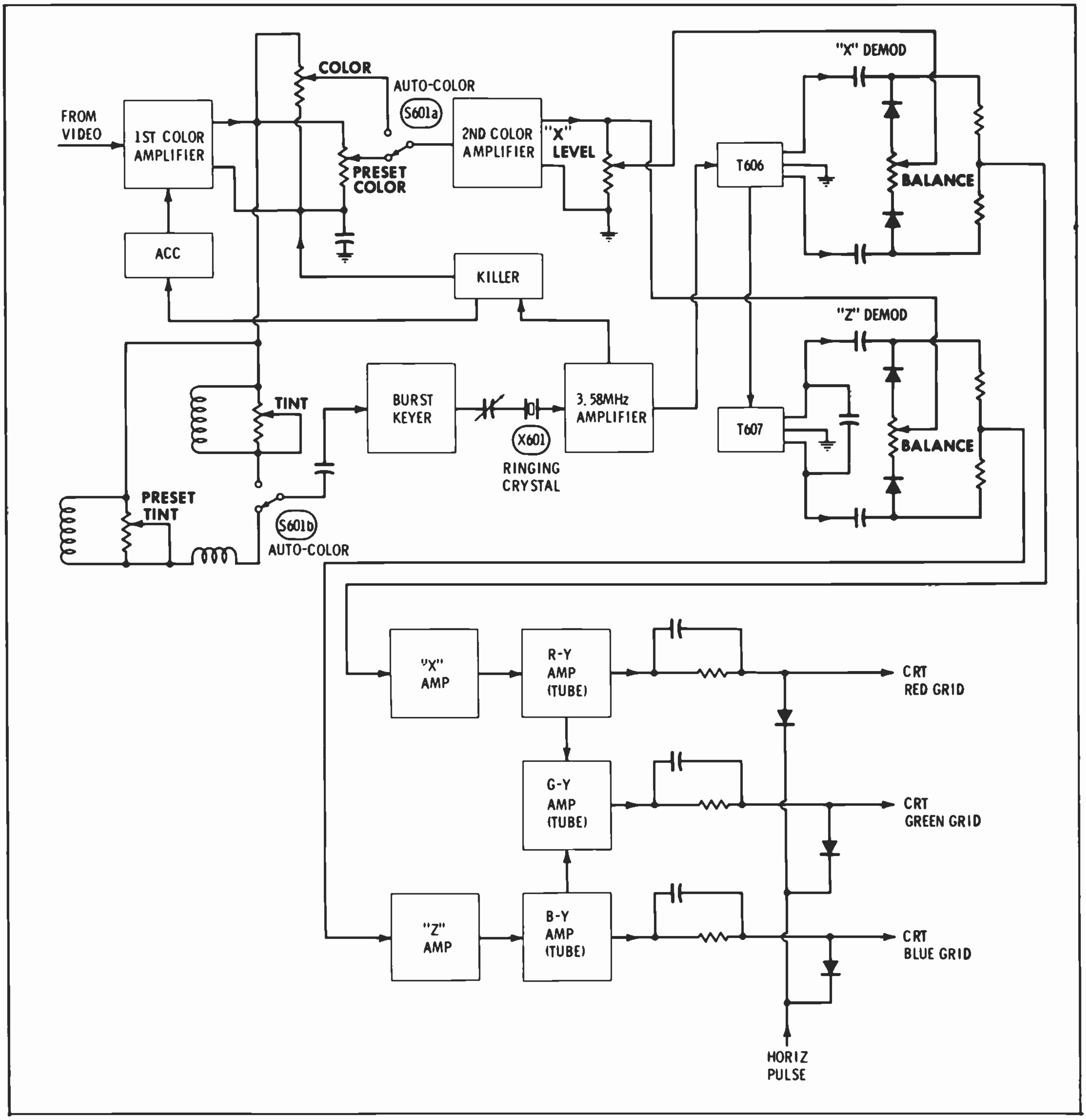


Fig. 24 A block diagram of the Panasonic CT98D chroma circuit shows pre-set color and tint controls, "X" balance control and a variation of the CRT grid clamp circuit.

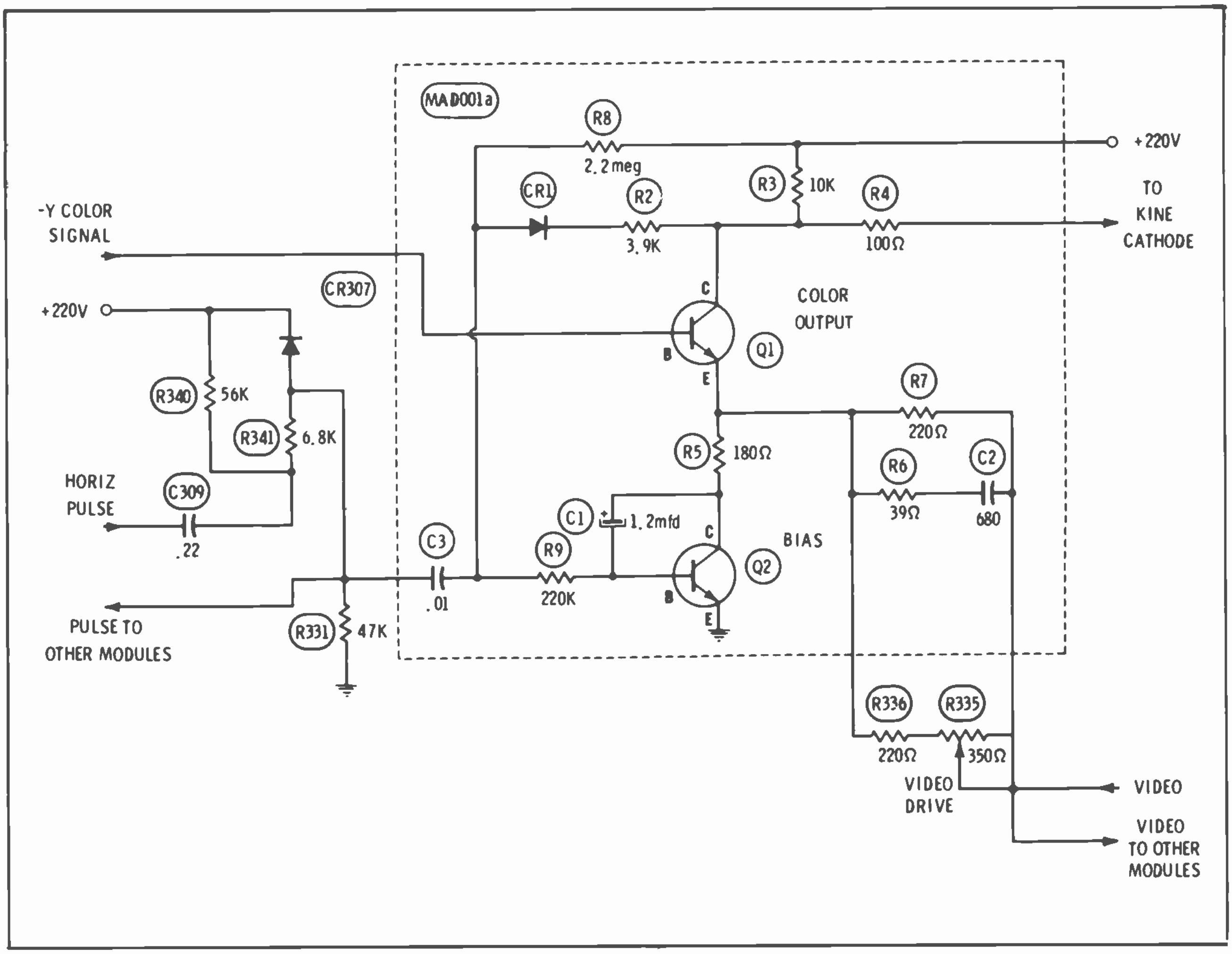


Fig. 25 Plug-in MAD module of RCA's CTC49 all-solid-state color portable receiver incorporates pre-CRT matrixing of video and chroma signals, plus pulse stabilization of the DC voltages applied to the CRT.

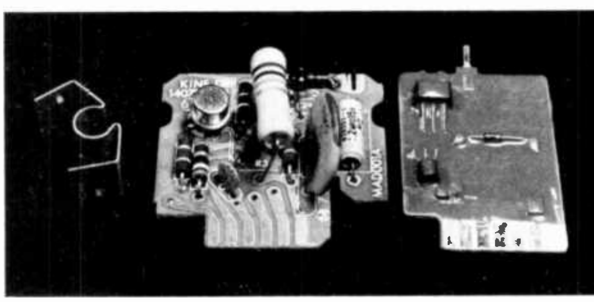


Fig. 26 An early production MAD module with heat sink is shown on the left, and the newer encapsulated version on the right.

the pulse and provide DC return paths. Diode CR307 "compares" the pulse to the +220-volt supply and clips off any portion of the pulse which exceeds 220 volts peak. The 220-volts (peak) pulse passes through C3 and diode CR1 (by forward biasing it into conduction), and, after attenuation to 160 volts peak by the voltage divider action of R2 and the C-E resistance of Q1, continues on to the cathode of the CRT and blanks off the associated gun. This blanking action is in addition to the horizontal blanking applied to a video amplifier.

Regulation of the DC voltage on the collector of Q1 is not quite as simple as the blanking function. During retrace time, when no color signal is present on the collector, the DC voltage at the collector of Q1 is compared to the blanking pulse. An error-correcting voltage developed by this comparison action is used as the variable forward bias for Q2. The variable voltage drop across the C-E junction of Q2 changes the forward bias of Q1; as a result, the DC voltage on the collector is maintained at the desired level. These complex actions are easier to understand if the explanation is divided into three segments.

First, imagine that R8 and C3 are removed, and that R2 and diode CR1 are removed and replaced by a resistor of the correct value to provide bias for Q2. If the ambient temperature increases, Q1 draws more current, the collector voltage decreases (less positive), less forward bias through the added resistor is applied to the base of Q2, and the collector voltage of Q2 increases, as does the emitter voltage of Q1. An increase in emitter voltage represents a decrease in forward bias; consequently, the transistor draws less current and the collector voltage is increased to the same voltage that was present there before the temperature change.

C1 slows the correction action to prevent overshoot or hunting.

It is obvious that a circuit which increases the forward bias of Q2 when the collector voltage of Q1 is increased also will stabilize that collector voltage at a level determined by the tolerance of the parts in the circuit. However, a standard is needed so that the collector voltage is stabilized at the desired voltage, and so that the DC voltages at the collectors of all three Q1 transistors will be nearly identical. This standard is furnished by the blanking pulse.

For the second segment, imagine that C3, diode CR1 and R2 are reinstalled according to the original wiring, but R3 is removed and the collector of Q1 is grounded. On the anode of diode CR1 will be found a non-varying negative voltage which is the result of shunt-type rectification of the horizontal blanking pulse by diode CR1.

Next, imagine that R2 is disconnected from the collector of Q2 and attached to a variable, positive DC supply. As the positive supply voltage is increased, the negative voltage at the anode of diode CR1 decreases. The reason for this action is simple, but important: Any positive voltage on the cathode of diode CR1 subtracts from the rectified (negative) voltage, because before rectification can occur more pulse is required to exceed the cathode voltage. Or to state this another way: Any positive voltage on the cathode of diode CR1 causes the same effect on the rectified voltage as would a reduction in the pulse amplitude.

The last analytical step assumes that the circuit is complete, as shown in Fig. 25. Resistor R8 supplies an unvarying positive voltage at a slightly higher level than the negative voltage from the anode of CR1, which changes when correction of the collector voltage of Q1 is required. This resultant voltage is the bias supply for Q2, and is supplied to the base of this transistor through R9. C1 filters out the horizontal pulses so that only DC voltages are applied to Q2. In addition, C1 retards the speed of any correction of the voltage at the collector of Q1.

Now the stabilizing loop is complete from the collector of Q1, through the pulse circuit, bias voltage of Q2, bias voltage of Q1 and back again to the collector of Q1.

Improved Serviceability and Product Safety

Motorola Quasar

The new Motorola TS934 color chassis (Fig. 27) is constructed with plug-in panels mounted in a roll-out drawer, in the distinctive Quasar tradition. But in addition, the entire front panel swivels, after a few screws are removed, to permit tuner repairs or contact cleaning without removal of the tuner.

Zenith

Zenith color TV receivers which feature the 4B25C19 chassis do not require a shielded high-voltage compartment because no high-voltage rectifier, regulator or focus rectifier tubes are employed. Instead, a solid-state, high-voltage tripler, housed in a sealed assembly (behind the high-voltage transformer in Fig. 28), furnishes both high voltage and focus voltage. Both the transformer and tripler are readily accessible for testing or replacement.

Clips and brackets prevent accidental removal of the medium- and large-size transistors during transport, as shown in Fig. 29.

General Electric

Symbols of components important to the prevention of fire and X radiation are printed over a gray background on General Electric's schematics, as shown in Fig. 30.

DC voltage sources and their destinations are marked with large symbols, also shown in Fig. 30.

GE's use of their unique in-line-guns color picture tube has been expanded to include larger screen sizes. Fig. 31 shows the relatively few convergence components necessary with this system, which permits simpler purity and convergence adjustments.

RCA

Vertical output transistors and horizontal sweep SCR's are readily accessible on the large heat sinks of the RCA CTC49 chassis (Fig. 32).

High voltage and focus voltage in the new RCA chassis are obtained from a solid-state quadrupler assembly (Fig. 33).

Pincushion Correction

Many pincushion correction cir-

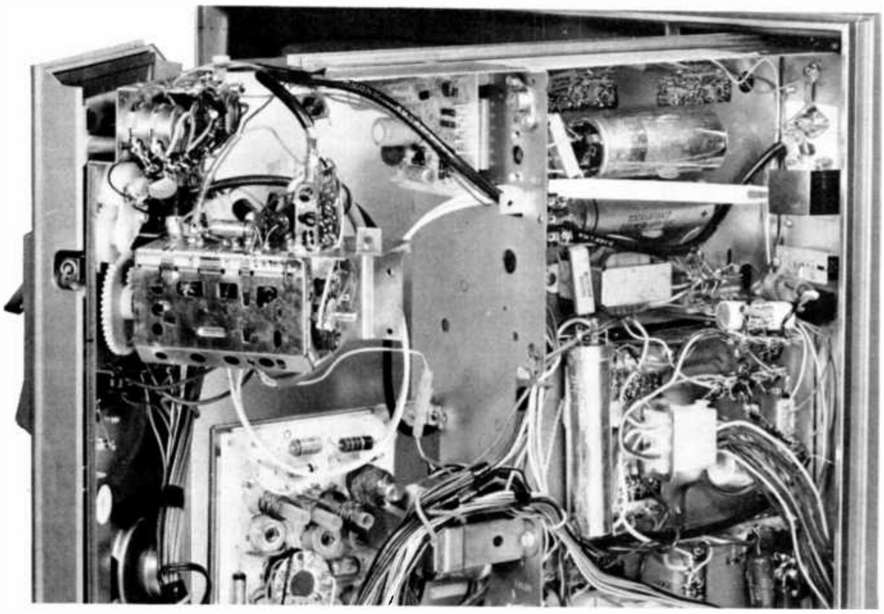


Fig. 27 The entire front panel of the Motorola TS934 swivel to permit access to the tuners.

Fig. 28 No high-voltage compartment is needed in the Zenith 4B25C19 chassis; consequently, the high-voltage transformer and the tripler assembly are easily accessible.

cuits have been changed. The trend is toward the use of special transformers and saturable reactors in circuits without tubes, transistors or diodes.

Packard Bell

The Packard Bell 98C32 color chassis employs the circuit shown in Fig. 34 to correct both vertical and horizontal pincushioning. The amplitude of horizontal pulses induced through pincushion modulator transformer T508 are corrected by the adjustment of R525. Resonant action of C513 and L502 shapes the horizontal pulses into near-sine waves, which are inserted in series with the vertical yoke coils. L502 also functions as a phase, or tilt, adjustment. Simultaneously, the sawtooth waveform of vertical yoke current changes the inductance of modulator transformer T508, reducing the width of the picture at the top and bottom (the primary winding of T508 is in series with the horizontal yoke deflection coils).

Zenith

The pincushion correction circuit of the Zenith 4B25C19 color chassis is similar in basic concept to that of the Packard Bell circuit just described. As shown in Fig. 35, horizontal pulses from the yoke circuit are routed through the two outside-leg windings of T203, thus inducing horizontal pulses into the center-leg winding. These pulses are changed into near-sine waves by the resonant circuit, comprised of L208 and C224/C246. L208 also functions as a phase, or tilt, adjustment. The sawtooth vertical yoke current flowing through the center-leg winding of T203 also changes the inductances of the two series-connected outside-leg windings. Because the outside-leg windings are in series with the horizontal yoke coils, the horizontal sweep width is reduced at the top and bottom of the picture.

Pincushion correction for the Zenith 40BC50 and 12B14C50 chassis is divided into separate top/bottom and side correction circuits. At first glance, the circuit in Fig. 36 appears to be nearly the same as that used by RCA for so many years. However, T208 is a toroidal type, whose ferrite core is doughnut shaped. The two toroidal windings in series with the vertical yoke coils are wound around the circular core in true toroidal fashion. The winding marked "outside" is wound around the entire core in the way a solenoid coil is wound. However, because the magnetic fields of the coils are at right angles, no AC signal would be induced from one type of coil into the other if two ceramic magnets were not added.

The two ceramic magnets are included with the toroidal core. Their magnetic flux is enough to nearly saturate the ferrite core. Current flowing through the outside winding also will saturate the core, and thus permit horizontal pulses to be transferred to the torodial windings. However, the vertical sawtooth current in the torodial windings is not induced into the outside winding.

The horizontal pulses obtained from the toroidal windings are filtered into a near-sine wave by the resonant action of C261 and T202, which also functions as the tilt adjustment.

Side pincushion correction in the Zenith 40BC50 chassis is accomplished by the circuit shown in Fig. 37. Vertical pulses from the vertical sweep circuit are filtered into a parabolic waveform and reduced in amplitude by the actions of CR212, R-313, C233, R312 (side pincushion control), R311 and C233 before they are applied to the base of transistor Q211. The amplified vertical parabolic waveform at the collector of Q211 is applied to the voltage regulator of the 128-volt supply, which furnishes the voltage for the horizontal-output stage.

Because the 128 volts applied to

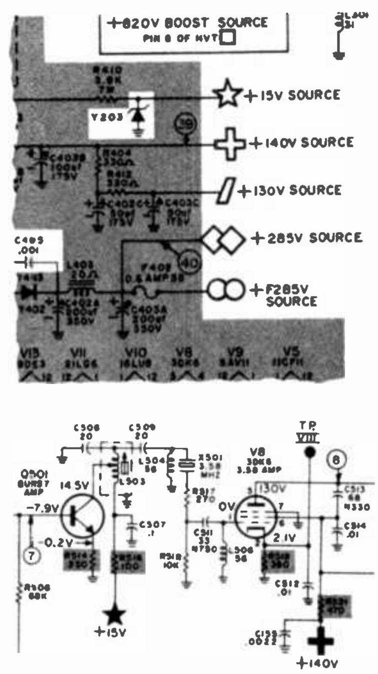


Fig. 30 Gray-shaded areas on GE schematics indicate parts that are important to product safety or the prevention of x-radiation. Each DC voltage source is shown by an outline of a distinctive symbol, and each connecting circuit is indicated by a filled-in black symbol of the same shape.

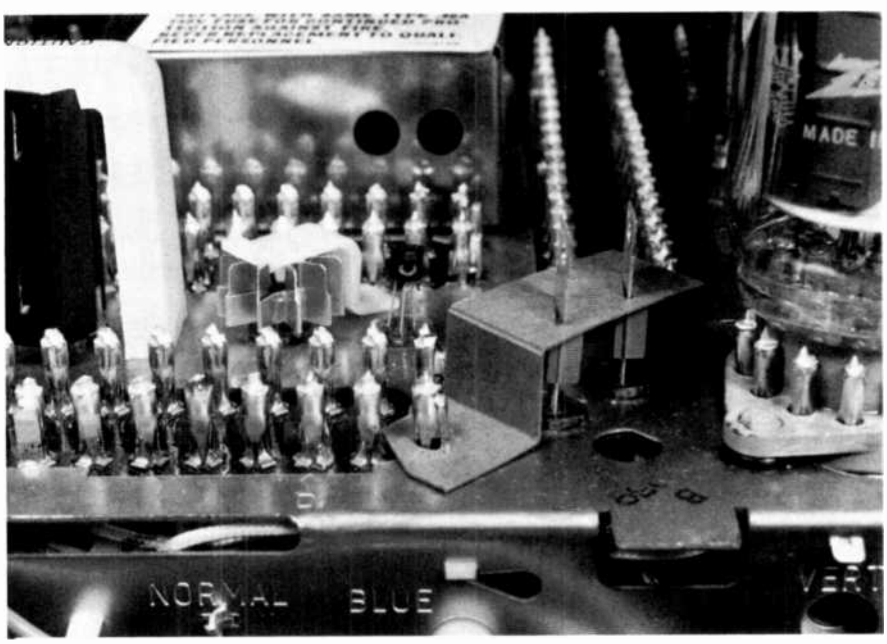


Fig. 29 Zenith provides clips and brackets to hold medium and large transistors in their sockets during shipping.

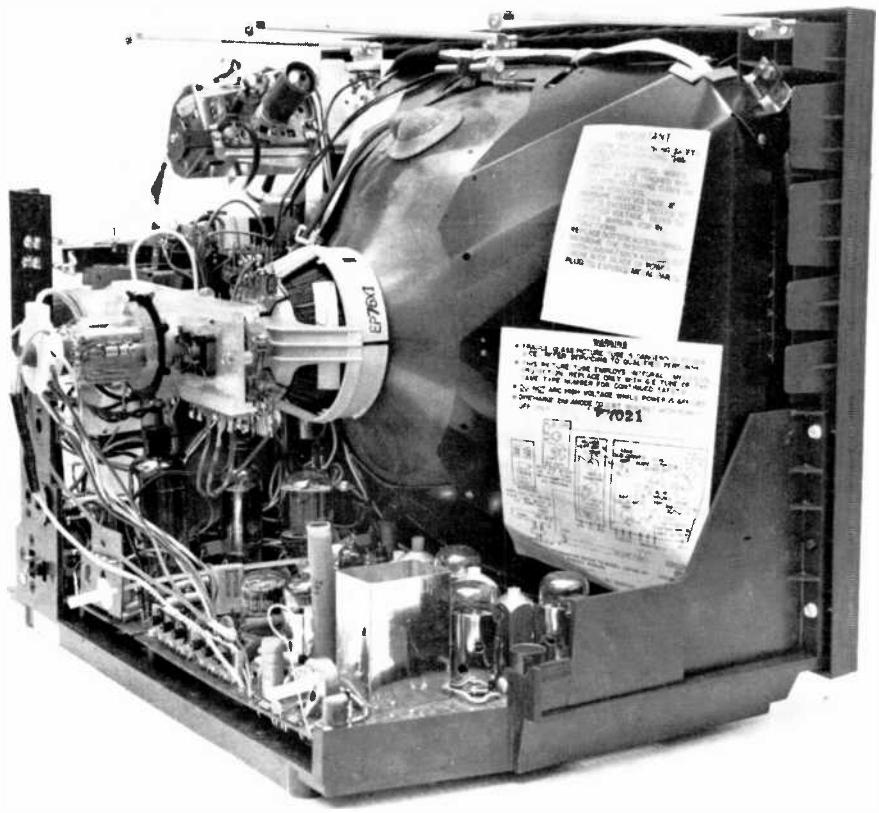


Fig. 31 There is no separate board for convergence controls in GE receivers employing the in-line-guns CRT; all adjustments are performed by changing connector taps and rotating magnets on the convergence assembly mounted on the neck of the CRT.

the horizontal output stage is reduced during the time the vertical sweep is scanning the top and bottom of the screen, the width of the raster at these points also is reduced.

Degaussing Circuits

Several of the new degaussing circuits use a bridge which becomes balanced when the positive-temperature coefficient resistor (posistor) reaches maximum resistance. This double action reduces the degaussing current to a very low value during normal receiver operation following degaussing.

JVC

The degaussing circuit of the JVC 7438 and 7408 color chassis, shown in Fig. 38, includes a double positive-temperature-coefficient thermistor, a tapped power transformer and a fixed resistor (which balances the bridge). Because both sections of R710 are cold and offer low resistance when the receiver is first turned on, almost the full 100 volts AC is applied to the degaussing coil. The resulting current through the coil heats both sections of R710, and the increased temperature causes the resistance of this unit to increase.

When the resistance of R710 reaches its maximum value the current through the degaussing coil will be minimum, but not quite zero. To completely stop the current through the coil, a voltage from the 110-volt tap of the power transformer is routed to the degaussing coil through R809 and thermistor R-710A, so that both ends of the degaussing coil are at 100 volts. With no voltage applied across the coil, its current stops completely, and the degaussing cycle is finished. However, current through R709 heats R710B as long as the receiver is turned on. The heat of R710B is physically transferred to R710A, to prevent R710A from cooling excessively during normal receiver operation, when no voltage is applied across R710A and the degaussing coil. When the receiver is turned off and the thermistor is allowed to cool, the degaussing cycle repeats.

Circuit actions of the degaussing circuit employed in the RCA CTC49 chassis (Fig. 39) are similar to those of the previously-described JVC design, except that a single positive-temperature-coefficient thermistor (posistor) is used. After the thermistor is stabilized at a high resistance, voltage from the power transformer, applied through R4, lowers the voltage at the R4/RT1 end of the degaussing coil until it is equal to that at the transformer end. Sufficient current flows through R4 and RT1 to keep RT1 hot enough to maintain a high value of resistance.

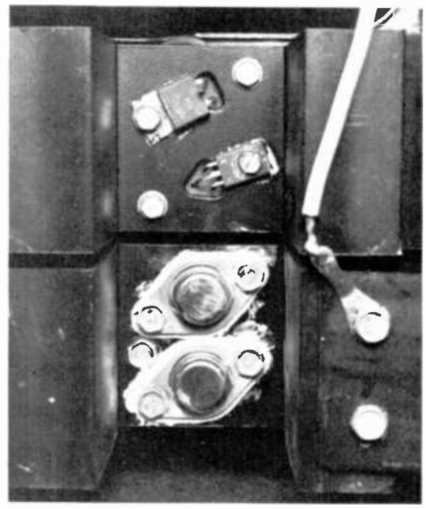


Fig. 32 The RCA CTC49 has the vertical output transistors and the horizontal sweep SCR's mounted on conveniently-located heat sinks.

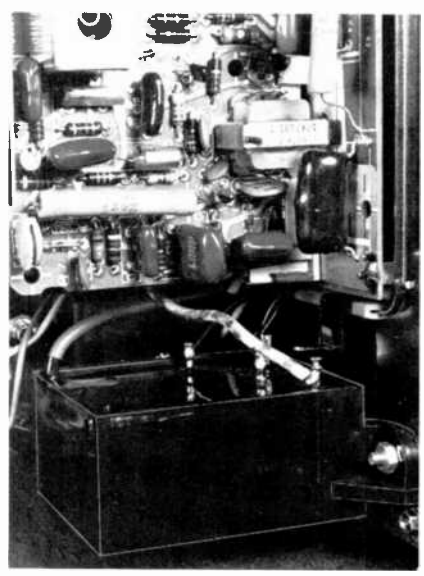


Fig. 33 Both high voltage and focus voltage in the RCA CTC49 are obtained from the solid-state quadrupler assembly.

Packard Bell

Degaussing of the CRT is not automatic in the Packard Bell 98C32 chassis, but is manually switched on by the TV viewer. This system is called "Instant Color Purity" (ICP). The advantage of this system is that the set operator can degauss the picture tube at any time while the set is operating, without waiting for thermistors to cool before the degaussing will re-cycle.

As shown in Fig. 40, the degaussing coils are in series with the AC supply to a bridge rectifier circuit which is used only for degaussing. The ICP pushbutton switches in C509, causing a large sine wave of 120-Hz current to pass through the degaussing coil. This current rapidly decays (like a damped wave train) as electrolytic capacitor C509 accepts a charge. Because the degaussing action is so quick, it is completed before the set operator can withdraw a finger from the pushbutton. A moving, swirling color display, when the button is activated, is proof of degaussing.

Chroma Circuits

Motorola

An integrated circuit (IC) in the Motorola TS929 chassis is employed as three demodulators, three color pre-amplifiers and three pre-CRT matrix circuits, as shown in Fig. 41.

The outputs of the matrix circuits are not –Y signals, but, instead, are pure color signals consisting of a pre-CRT mixture of video and –Y signals which, after more amplification, are ready for direct application to the appropriate cathodes of the picture tube.

Varactor diodes (varicaps) function as variable tuning capacitors in the Motrola TT651 UHF tuner (see Fig. 42). Positive DC voltage used to control the capacitance of the varactors is selected by the channel-selector switch from one of 13 potentiometers. This design enables all four of the varactors to be controlled by one voltage. The action is the same as a four-section variable capacitor which is activated by a common shaft.

Zenith

Varactor-controlled VHF and UHF tuners plus a unique system of pre-tuning and automatic switching are combined in Zenith's new "Varactor TV Tuning System" to enable viewers to select, with a single knob and without fine tuning,

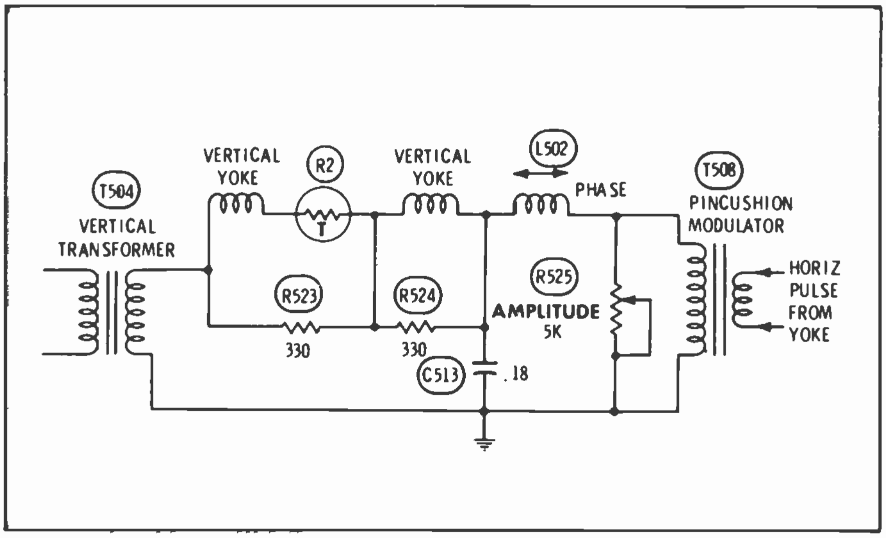


Fig. 34 The Packard Bell 98C32 color chassis uses this one circuit to correct both vertical and horizontal pincushioning.

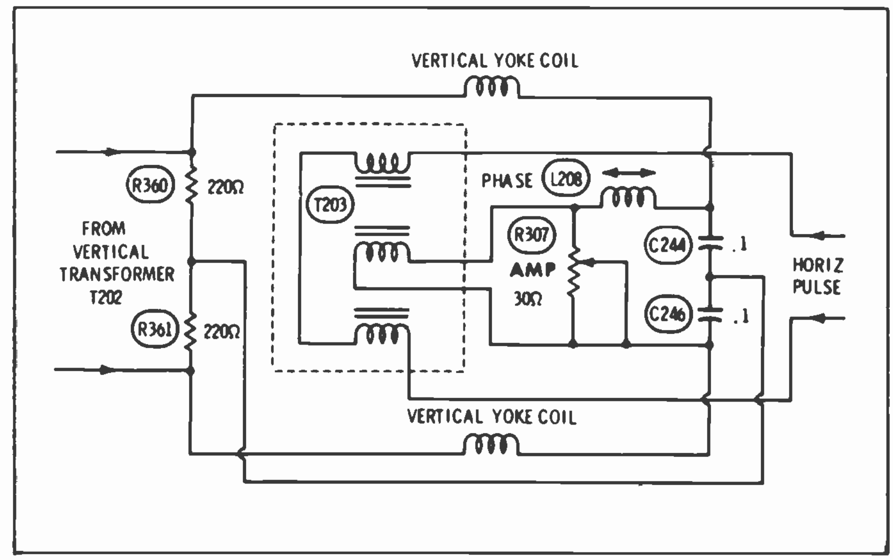


Fig. 35 Both vertical and horizontal pincushioning in the Zenith 4B25C19 chassis are corrected by this circuit.

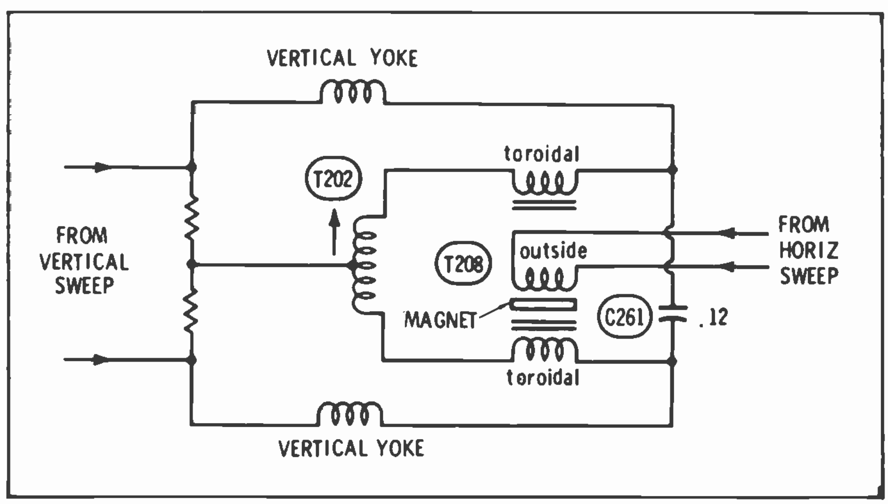


Fig. 36 New in the Zenith 40BC50 all-solid-state chassis is T208, which has a toroidal core, two ceramic magnets, two toroidal windings and one conventional winding. This combination provides a one-way signal transference. C261 and T202 filter the horizontal pulse into a near-sine wave for top/bottom pincushion correction.

any one of 14 pre-set VHF or UHF channels, or combination of VHF and UHF channels.

The new all-solid-state tuning system, a diagram of which is shown in Fig. 43, consists of:

- A channel-selector unit, in which all mechanical switching is accomplished;
- An all-electronic, FET-equipped VHF tuner whose frequency is changed by 1) changing the reverse bias applied across varicap diodes employed in its four tuned circuits and 2) triggering on and off five switching diodes which effectively insert or remove inductances from the tuned circuits so that the tuner can be tuned across both the low and high VHF bands with about the same amount of varicap control voltage.
- An all-electronic UHF tuner equipped with four varicap-tuned circuits, one of which is a transistor RF amplifier stage (itself, a radical departure from previous UHF tuner designs).
- A "nerve center" unit which is constructed on a Duramodule plug-in circuit board and which, in conjunction with the channel selector unit, controls the voltages necessary for proper operation of the system.

The channel selector unit, which, as mentioned previously, performs all mechanical switching, is equipped with a drum-type, detent channel-selector mechanism similar to that employed in some conventional VHF tuners. However, instead of coil strips, the Zenith channel selector mechanism is equipped with 14 vernier-type potentiometers, each of which corresponds to a desired channel.

When the channel selector is rotated to one of the 14 available positions, the full resistance of the potentiometer associated with that position is placed between a regulated +33 volts source and ground. At the same time, the slider of the potentiometer is connected to a line which is common to all varactors in either the VHF or UHF tuner. Consequently, depending upon the position of the slider, a portion of the varactor control voltage (+.5 to +28 volts) is applied as reverse bias across the varactors, changing the resonant frequency of the tuned circuits in which the varactors are used. (Changing the level of reverse bias applied across a varactor produces a corresponding change in its internal capacitance. Consequently, when the varactor is used as part of a tuned circuit, the resonant frequency of the circuit will be changed when the reverse bias applied to the varactor is changed.)

A bandswitch which effectively functions as a five-section, three-

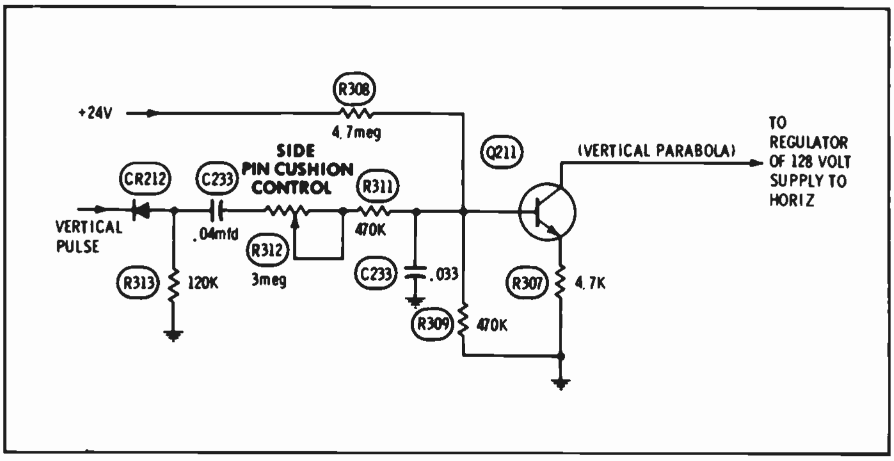


Fig. 37 Side pincushion correction in the Zenith 40BC50 chassis is accomplished by reducing the +128 volts supplied to the horizontal output stage during the time the sweep is at the top and bottom of the picture.

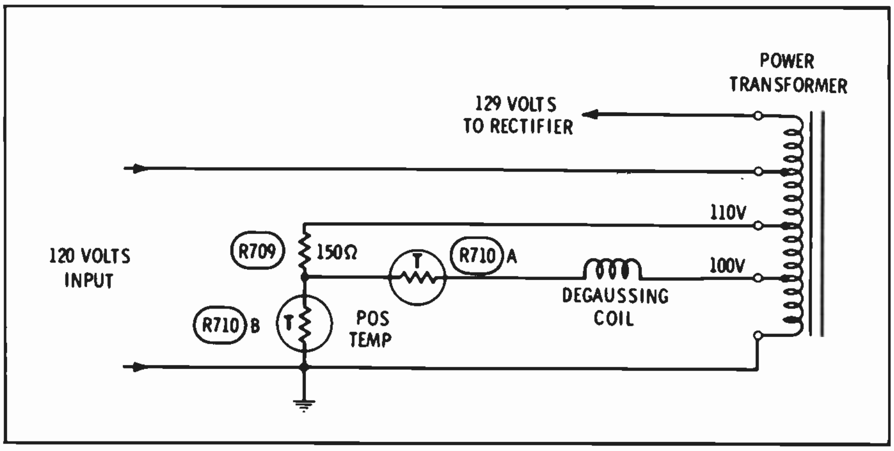


Fig. 38 The JVC 7438 and 7408 chassis employs a double thermistor plus a fixed resistor in a bridge circuit, to more completely shut off degaussing.

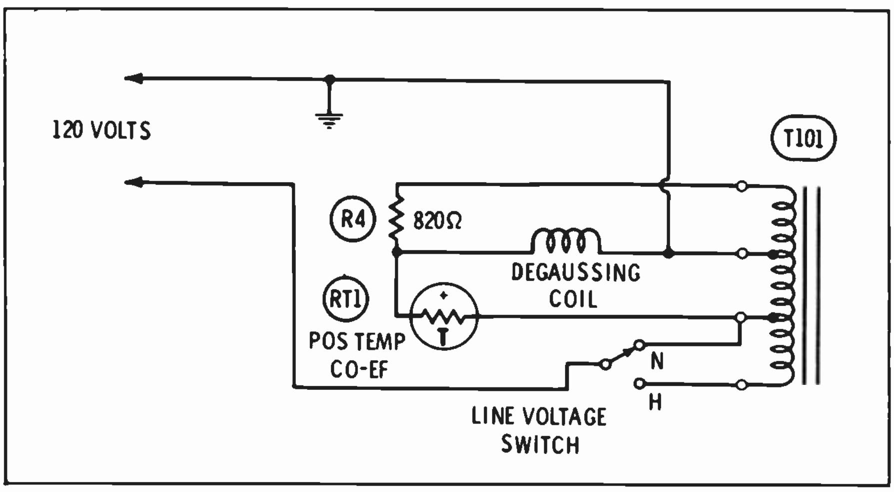


Fig. 39 A different bridge circuit is used in the deguassing circuit of the RCA CTC-49 chassis; a resistor and one posistor balance out the current flow through the degaussing coil.

position switch, and which is actuated by the tuning drum mechanism after initial setup, performs the circuit changes required for switching between low-band VHF (2-6), high-band VHF (7-13) and UHF (14-83)

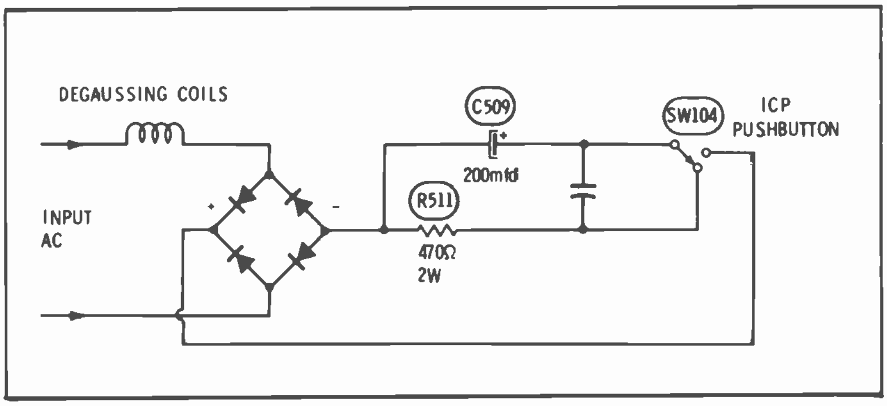


Fig. 40 "Instant Color Purity" in the Packard Bell 98C32 chassis is initiated by a manually operated control. Current through the degaussing coil decreases rapidly as capacitor C509 becomes charged.

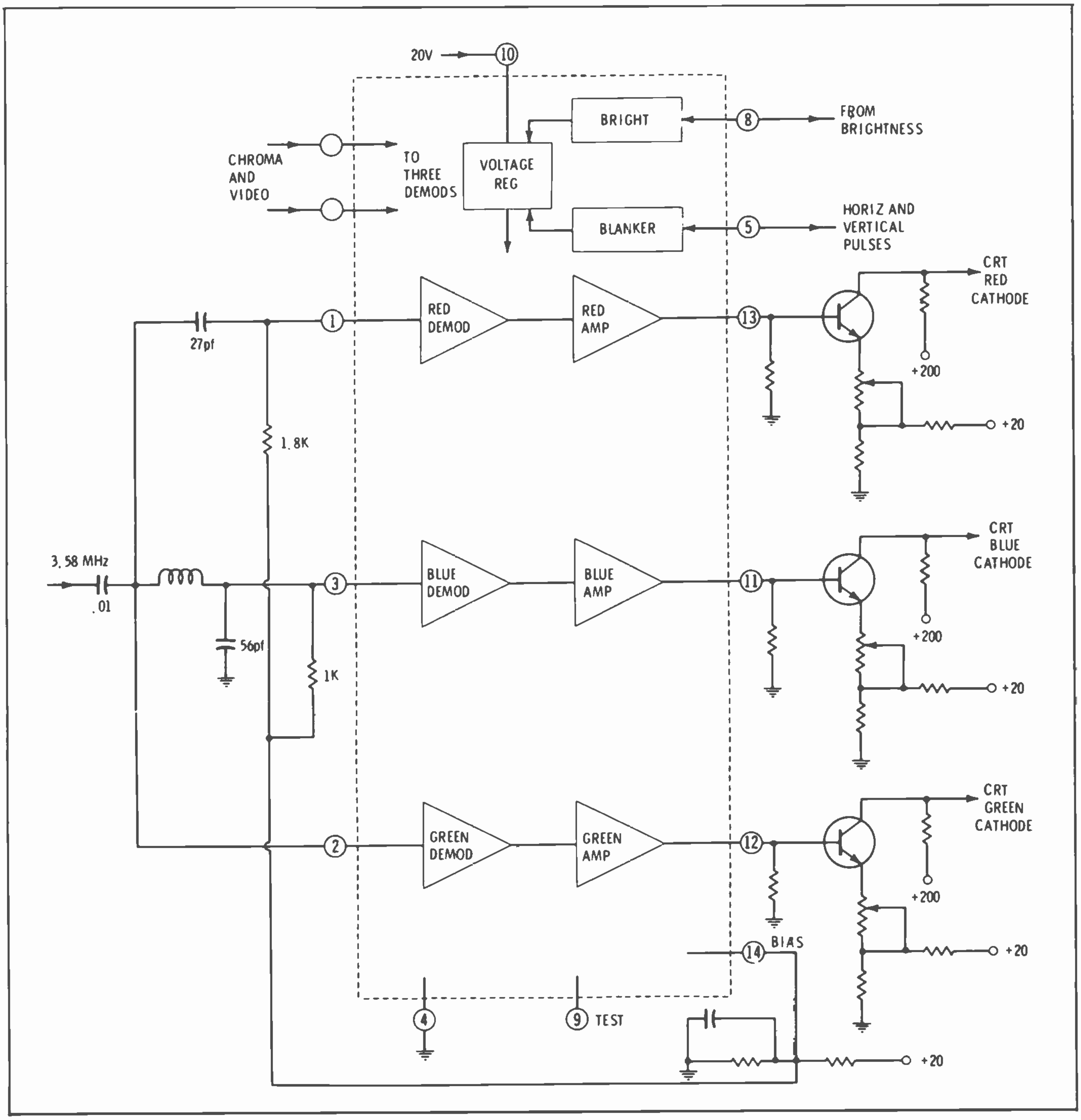


Fig. 41 The Motorola TS929 chassis employs an IC to perform demodulation, amplification and pre-CRT matrixing.

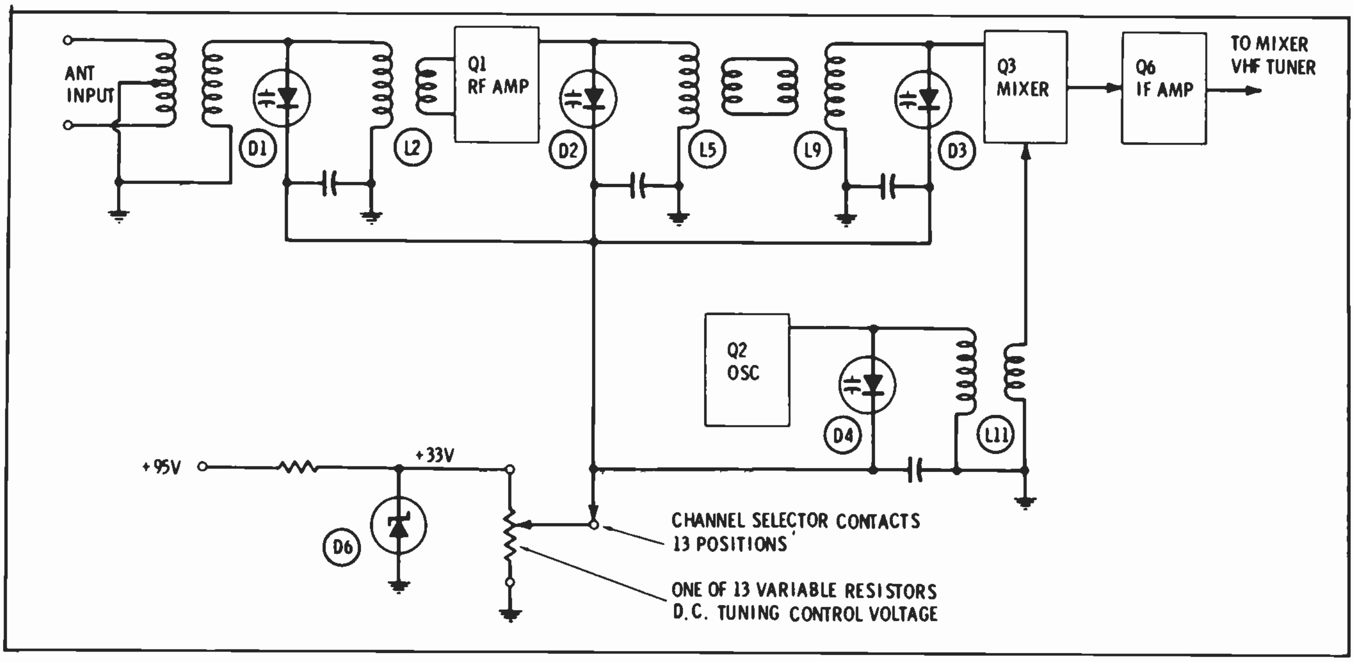


Fig. 42 Varactor diodes replace variable-tuning capacitors in the Motorola TT651 UHF tuner.

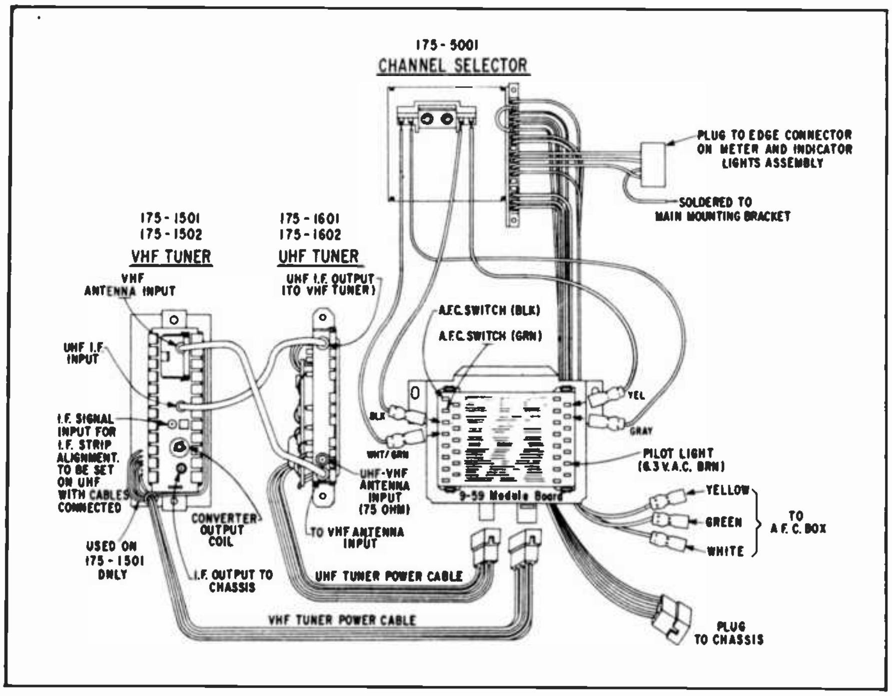


Fig. 43 Varactor-controlled tuning in both VHF and UHF tuners of the new Zenith Varactor Tuning system makes possible one-knob tuning of 14 pre-set VHF and UHF channels.

operation, each of which relates to one of the three switch positions. The five sections of the switch control B+, bandswitching, band-indicator lights, varactor and AFC tuning, and AGC switching, For example, in the low-band VHF position of the bandswitch:

- Section 1 switches B+ to the VHF tuner,
- Section 2 applies a reverse bias (negative) voltage to the five switch diodes in the VHF tuner. These diodes—X1, X2, X5, X6, and X10 in Fig. 44—effectively switch more inductance into their respective circuits for low-band VHF, operation and less inductance for high-band VHF operation.
- Section 3 applies filament voltage to the low-band indicator lamp.
- Section 4 performs no function when the bandswitch is in the low-band position. (In the high-band and UHF positions of the bandswitch more resistance is placed in parallel with the output resistor of the AFC discriminator, to limit the variation of the AFC correction voltage and thus maintain the same frequency pull-in range as provided on the low-band position.)
- Section 5 provides appropriate AGC voltage to the field-effect transistor (FET) employed as the RF amplifier in the VHF tuner. A conventional (bipolar) transistor is used for the same function in the UHF tuner. Consequently, because the techniques for controlling the gain of FET and bipolar transistors are different, the receiver must provide two separate AGC voltages. Thus the need for switching AGC voltages.

The design of the system also has permitted the inclusion of automatic frequency control (AFC) without the use of additional varactor diodes. AFC correction voltage is developed in the conventional manner by a discriminator circuit, except that the new AFC discriminator employed in this chassis provides a dual output, which is applied across a 6.8K-ohm resistor. When no correction is required or when the AFC circuit is manually defeated, the differential between the two outputs in zero and no voltage is dropped

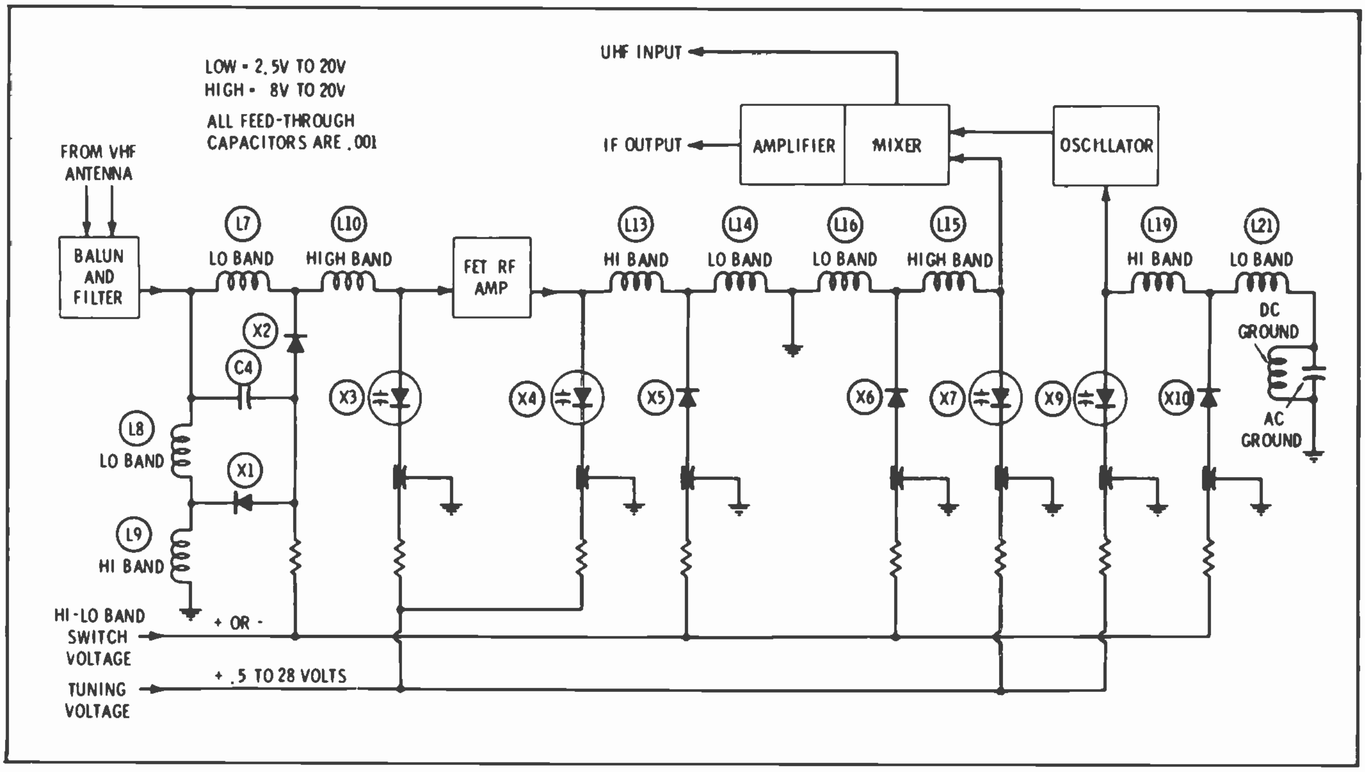


Fig. 44 A simplified schematic of the hi/lo band switching in the VHF tuner of the Zenith Varactor system. Diodes are used as voltage-controlled switches, to short out unneeded coils.

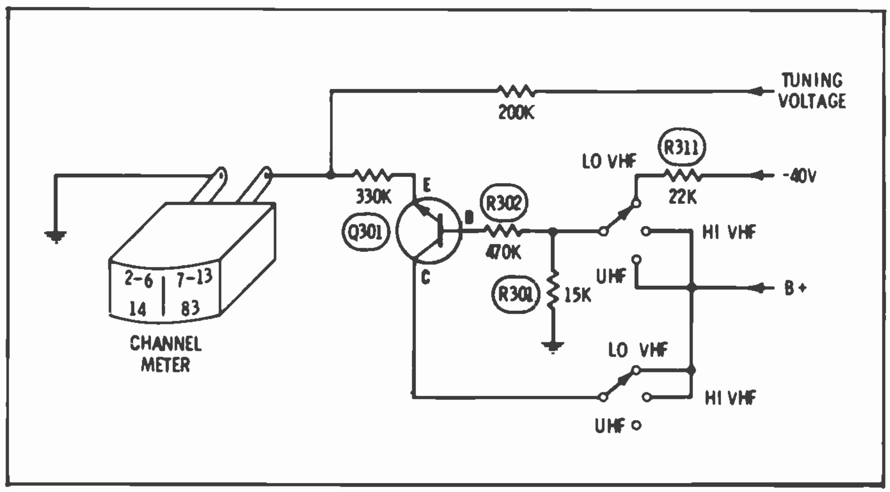


Fig. 45 A meter whose dial is calibrated to show the approximate location of the VHF and UHF channels is used during the initial set-up of each desired channel in the Zenith Varactor system.

across the AFC resistor. When correction is needed, the two outputs become unbalanced and a level of voltage proportionate to the required correction is dropped across the AFC resistor (maximum correction voltage is ± 1 volt).

AFC action is accomplished by routing the channel-selector voltage (from the slider of the potentiometers on the tuning drum in the channel-selector unit) through the AFC resistor. The voltage drop across the AFC resistor, which corresponds to the AFC correction needed, is added to or subtracted from the channel-selection voltage, which, in turn, is fed to the channel-selection varactors in the tuners.

Band indicator lights and a tuning meter are provided to assist the technician in the initial setup of the tuning system. These tuning aids are controlled by the bandswitch and the nerve center circuitry.

Because the range of varactor correction voltage is almost the same for the low-and-high-band VHF functions, additional voltage must be applied across the tuning meter during high-band operation, to swing the meter needle to midscale (channel 7 position) before channel setup of the high-band begins. This is accomplished by the tuning meter circuitry shown in Fig. 19.

During UHF and low-band VHF operation, transistor Q301 in Fig. 19 is cut off and only the varactor tuning voltage is applied across the tuning meter. When high-band operation is selected, transistor Q301 is biased on and the resultant additional current through the meter swings the needle to midscale. The needle then is positioned between the channel 7 and channel 13 positions, depending on the amount of varactor correction voltage applied to the meter.

Another unique feature of the tuning system involves the input circuitry to the tuners. A single antenna input balun, which can be used with either a balanced 300-ohm antenna line or an unbalanced 75-ohm coaxial line, feeds both VHF and UHF signals to the UHF tuner through a 75-ohm coaxial line. A low-pass filter in the UHF tuner then separates the VHF signal from the UHF and feeds the VHF signals to the VHF tuner through a 75-ohm cable.

A more detailed discussion of the operation and troubleshooting of

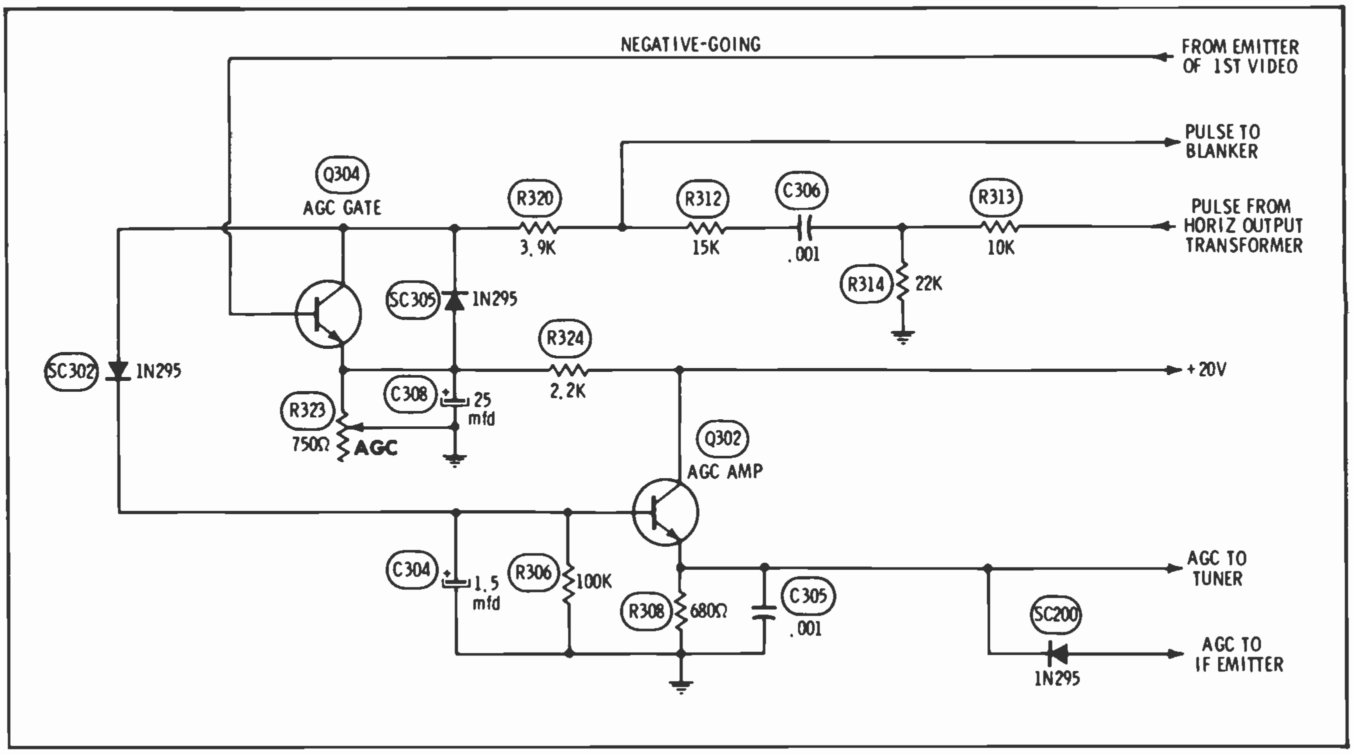


Fig. 46 The Sylvania D-16 color chassis AGC circuit employs a transistor (Q304) as a shunt resistor to vary the amplitude of the horizontal pulse, which is then rectified by diode SC302

Zenith's new "Varactor TV Tuning System" will appear in a near-future issue of ELECTRONIC SERVICING.

Miscellaneous Circuit Features

Sylvania AGC

Most AGC circuits, both tube and transistor, have a stage labelled "AGC gate" or "AGC keyer" which rectifies (with different efficiencies for different signal strengths) a horizontal pulse. The variable DC produced by this rectification is used as an AGC control voltage, or it varies the bias of a transistor whose output voltage supplies an AGC control voltage.

The AGC circuit of the Sylvania D-16 color chassis, shown in Fig. 46, is a departure from this traditional AGC system. Q304, although called the AGC gate, is actually the lower branch of a variable voltage divider that changes the amplitude of the pulse supplied to SC302, which performs the actual rectification. Positive DC voltage developed by the rectification is filtered by C304 and is applied to Q302 as forward bias. Because of the polarities involved, an increase in signal strength increases the forward bias on Q302, which in turn, increases (for saturation reduction of gain) the forward bias of the RF and IF amplifier transistors.

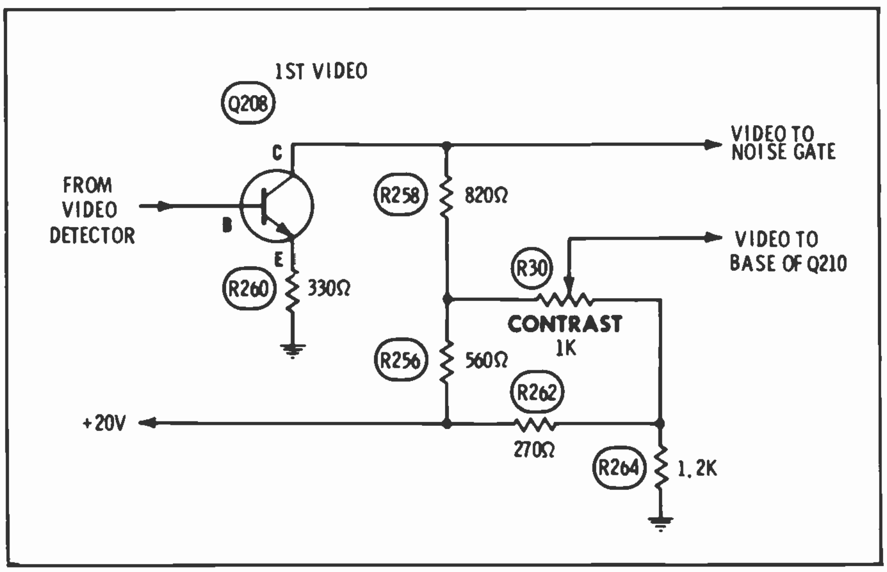


Fig. 47 DC voltages at each end of the contrast control in the Sylvania E01 chassis are nearly the same to prevent normal contrast adjustments from changing the DC voltages in the directly coupled video stages.

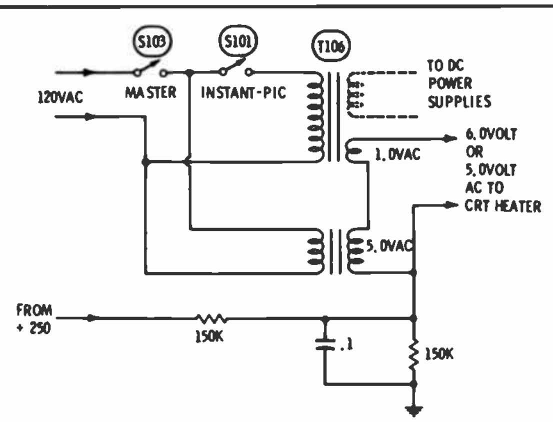


Fig. 48 Any receiver incorporating a fast-warmup picture circuit must operate the tubes at reduced heater voltage during periods when the set is turned off. The RCA CTC44 chassis obtains the 6 volts AC necessary for the CRT heater from two separate power transformers. When the receiver is turned off by the "Instant-Pic" switch, T107 continues to supply 5 volts to the heaters of the CRT.

The first action of the AGC is to lower the gain of the IF transistor, because of the voltage channeled through diode SC200. But when the voltage from the emitter of Q302 exceeds the positive bias on the IF transistor, diode SC200 is cut off. Further increases of the voltage at the emitter of Q302 are applied only to the RF transistor. Thus, application of AGC to the RF stage is delayed until a certain signal strength is exceeded. This prevents overload of the mixer stage in the tuner, and also establishes a good signal-to-noise ratio.

Sylvania Contrast

A different approach to contrast circuits is used in the Sylvania E01 color chassis (see Fig. 47). Part of the voltage from the collector of Q208 appears at the high side of R30, the contrast control, while the voltage divider consisting of R262 and R264 supplies a fixed DC voltage to the low end of the contrast control. Because both ends of the contrast control have approximately the same DC voltages, contrast changes, resulting from adjustment of the contrast control, have no effect on the bias of Q210, the following video amplifier transistor.

RCA Instant-Pic

The "Instant-Pic" circuit of the RCA CTC44 chassis (Fig. 48) utilizes two separate transformers to supply the heater voltage of the picture tube, the only tube used in receivers equipped with this chassis. With master switch S103 in the ON position, T107 supplies 5 volts AC to the heater of the CRT. Although the 1-volt winding of T106 is in series with this heater supply path, the primary of T106 is not energized, so the 1-volt winding functions as a very-low-value resistor.

When the "Instant-Pic" switch is in the ON position, T106 is energized, the output of the 1-volt winding is added in series with the 5 volts from T107, and the heater of the CRT has 6 volts applied.

A DC voltage from the voltage divider consisting of two 150K-ohm resistors is supplied to minimize the possibility of heater-to-cathode shorts inside the CRT. ▲