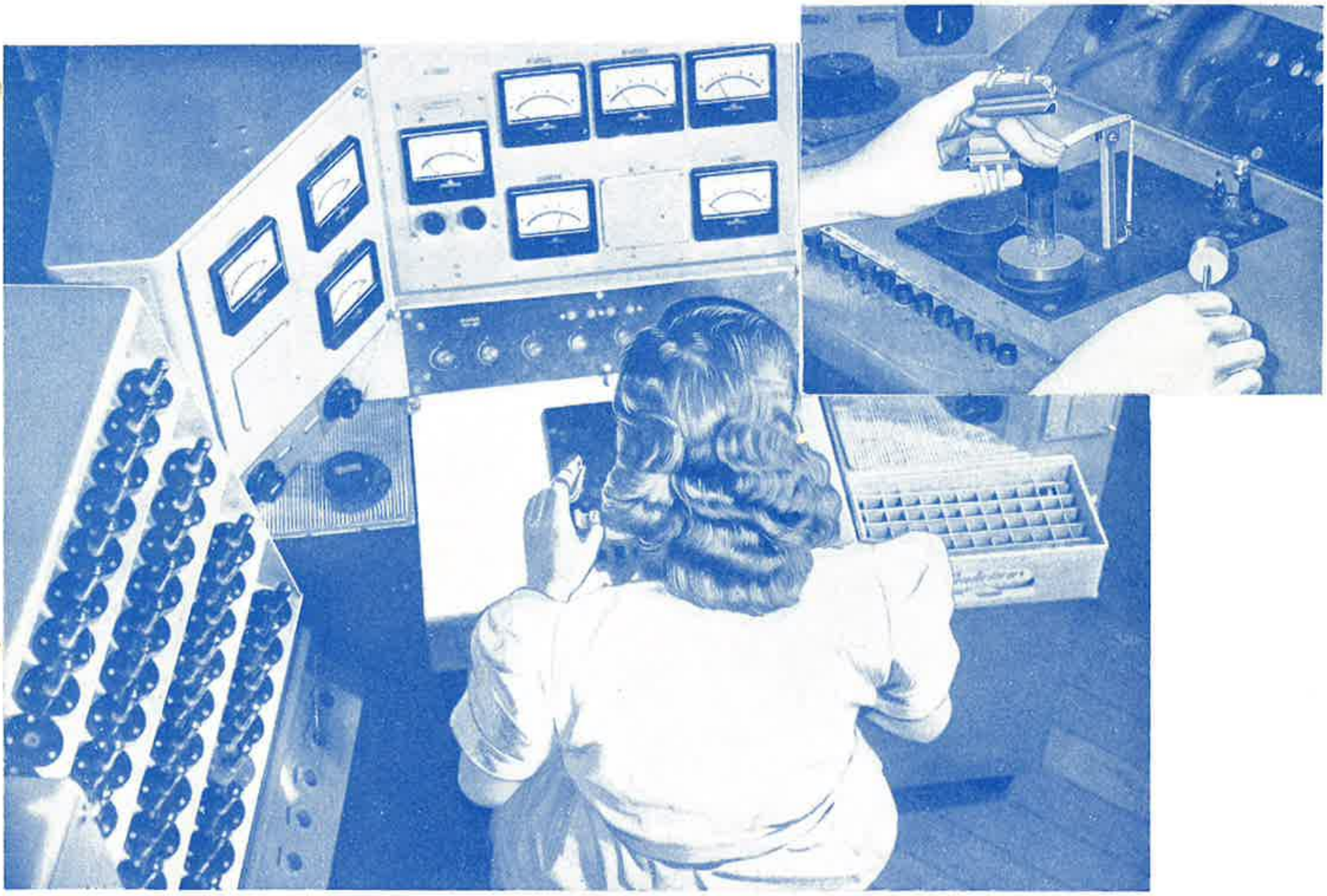
RADIOTRONICS

Volume 18 June, 1953 No. 6



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By the way —

The front cover this month shows the testing of a 6AQ5 miniature valve at the Amalgamated Wireless Valve Works at Ashfield, N.S.W.

The second edition of the Radiotron Valve Data Book is having a steady sale at a price of twelve shillings and sixpence, post free. This new edition contains revised data sheets on numerous receiving types together with additional sections covering transmitting tubes, phototubes and germanium crystals.

New Zealand subscribers should place their orders with Amalgamated Wireless (Australasia) Ltd., P.O. Box 830, Wellington, or with any branch of National Electrical and Engineering Co. Ltd.

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Editor:
Ian C. Hansen,
Member I.R.E. (U.S.A.)

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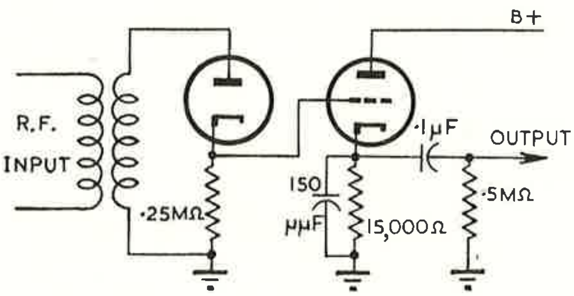
Full analysis in Jan. '56 issue, P.3.

A LOW-DISTORTION A-M DETECTOR

By F. Langford-Smith B.Sc., B.E.

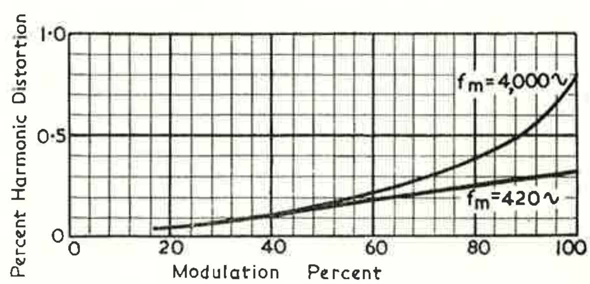
One of the difficulties with the design of low-distortion A-M receivers has been the distortion in the second detector which, at 100% modulation, is of the order of at least several percent with conventional diode or reflex detector circuits. Fortunately, in any normal type of receiver, the maximum percentage of modulation which the receiver is required to handle is always less than 100% even when a 100% modulated signal is applied to the aerial terminal — see Radiotron Designer's Handbook, 4th edition, page 1078. However, even with good design and with all these effects duly accounted for, the distortion in the second detector sets a limit to the reduction of distortion of a receiver or tuner.



The following description of a low-distortion A-M detector has been kindly made available by the inventors, Walter T. Selsted and Bob H. Smith. The development work was carried out in the Radiation Laboratory of the University of California.

* This is the circuit referred to on the last page of the December, 1952, issue of Radiotronics.

As will be seen from the circuit diagram, the detector is merely a conventional diode rectifier direct-coupled to a cathode follower which is in turn connected to an r-f filter to reduce the carrier signal output. Due to the fact that the load on the diode for normal A-M carrier frequencies is essential resistive, the normal effects of excessive shunting capacitance are eliminated.



Since the coupling to the cathode follower is direct, there is no effect of biasing currents which are normally developed in a diode loading circuit using coupling condensers. These two facts are responsible for the excellent performance of this circuit*.

The percentage total harmonic distortion, as shown by the curves, is only 0.3% at 100% modulation for a modulating frequency of 420 c/s, or 0.8% at 100% modulation for a modulating frequency of 4000 c/s. The frequency of the r-f input and the input voltage were not stated.

In a normal receiver it would be necessary to make separate provision for automatic volume control, requiring one additional signal diode.

CALCULATION OF IMPEDANCE WITH REACTANCE AND RESISTANCE IN SERIES

By F. Langford-Smith B.Sc., B.E.

This is an exact method, not requiring tables, and giving particularly high accuracy on any ordinary slide rule.

The ordinary formula for calculating Z with R and X in series is

$$Z = R \sec \Theta \qquad (1)$$

where Z = impedance
R = resistance
X = reactance
and Θ = phase angle = $\tan^{-1}(X/R)$

It may be proved that equation (1) is precisely the same as equation (2):

$$Z = R + X \tan(\Theta/2) \qquad (2)$$

The term $X \tan(\Theta/2)$ in equation (2) may be calculated by first determining Θ, dividing by 2 to obtain $\Theta/2$, finding $\tan(\Theta/2)$, and multiplying by X.

All these processes may be carried out with a slide rule. A normal slide rule is limited to 45° on the tangent scale, so that if X is greater than R it is convenient to interchange the values of X and R. This will not affect the value of Z, but the phase angle is the complement of the angle so determined. The accuracy is high because the "correction" $X \tan(\Theta/2)$ is always less than R.

The following procedure is an adaption to the slide rule, for which we are grateful to Dr. W. G. Baker, (Ionospheric Prediction Service.)

To perform this operation it may be necessary with some rules to turn the slider upside down, so that the T scale touches the D scale. The procedure is illustrated by examples.

Example 1: R = 4, X = 3.

1. To calculate Θ: Set the right-hand end of the T scale above 4 on the D scale. Set the cursor to 3 on the D scale and read the angle 36° 52′ on the T scale.
2. Move the right-hand end of the T scale under the cursor, and move the cursor to $\frac{36° 52'}{2} = 18° 26'$, reading correction 1.0 on the D scale.
3. Then Z = 4 + 1 = 5.

Example 2: R = 15, X = 8.

1. To calculate Θ: Set the *left*-hand end of the T scale to 1.5 on the D scale. (The use of the left-hand end multiplies by 10.) Set the cursor to 8 on the D scale and read the angle 28° 06′ on the T scale.
2. Move the right-hand end of the T scale under the cursor, and move the cursor to $\frac{28° 06'}{2} = 14° 03'$, reading "correction" 2.0 on the D scale.
3. Then Z = 15 + 2 = 17.

Example 3: If "correction" is less than 1

e.g. R = 2, X = 1.5.

1. Calculate Θ by the same procedure as for Example 1, giving angle 36° 52′.
2. Move the *left*-hand end of the T scale under the cursor, and move the cursor to 18° 26′, reading correction 0.5 on the D scale. (Note that use of the left-hand end of the T scale multiplies the correction by 10.)
3. Then Z = 2 + 0.5 = 2.5.

Reference to source of equation (2):
J. Bartels "Harmonic analysis of diurnal variations for single days" Terrestrial Magnetism and Atmospheric Electricity, Vol. 44 (1938), p. 145.

By K. Fowler and H. Lippert.

THE VIDEO DETECTOR AND AMPLIFIER

1. Basic design considerations.

The function of the video detector and amplifier is to process the signal at the output of the video i-f amplifier so that it can reproduce a picture of good detail and contrast when applied to the picture tube. Taking the detector and amplifier as a composite unit, it must be able to:

(1) demodulate the picture i-f carrier so as to pass on the video and synchronizing signals.

(2) Voltage amplify the demodulated signal so as to give the proper contrast to the black and white picture elements at the picture tube.

(3) Provide proper polarity of signal to the picture tube.

(4) Provide a linear frequency pass-band from almost zero to approximately 3 or 4 megacycles.

(5) Keep phase distortion at a minimum.

Of the above requisites, linear frequency response is the most important and hardest to obtain. Actually, frequency response is what principally distinguishes a video amplifier from a normal audio amplifier. However, if any of the other requirements are compromised, a loss in picture detail or an unintelligible picture will result.

2. Video detector.

At the video detector the picture i-f carrier is demodulated or detected, and the video signal and synchronizing pulses are passed on to the amplifier while the i-f is rejected. A typical diode detector as used in many television receivers is shown in Figure 7-1. This is called a series type and is similar to the detector found in the normal broadcast receiver except that the diode load resistance and filter capacitor are much lower in value than would be used for audio detection. This low value of resistance and capacity is required to maintain good frequency response out to 4 mc, as will be explained later.

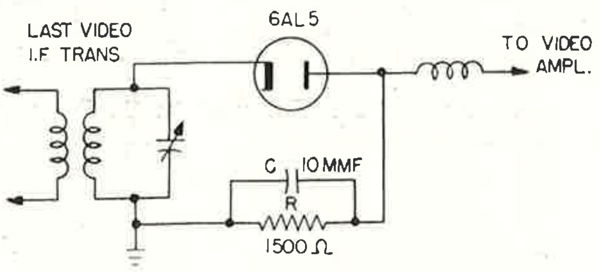


Fig. 7-1. Diode detector.

By courtesy of AGE, with acknowledgement to International General Electric Co. of U.S.A.

Since the video i-f signal frequency of 26.3 mc is only about five times the highest video modulation frequencies, there is not enough low-pass filtering at the load resistance and capacitor by themselves to discriminate against the i-f signal. Thus, some means must be provided to keep the i-f signal out of the video amplifier. This is accomplished quite simply by the use of a series choke coil placed between the output of the diode detector and the input to the video amplifier, as shown in Figure 7-1. This same series coil, if properly chosen, will provide video compensation in addition to keeping the i-f signal out of the video amplifier. Compensating circuits will be discussed in detail when we consider the video amplifier.

In the video detector circuit of Figure 7-1, C must be small enough that its reactance does not appreciably lower the effective value of the diode load impedance at the highest modulation frequency, and at the same time should be sufficiently large compared to R so as to hold the peak i-f charge of one cycle until the next cycle occurs. The value of C is usually about 10 $\mu\mu$f. The value of R is quite low in order to minimize the effect of shunt capacity at these higher frequencies and provide a fairly uniform response at all modulation frequencies. A typical value would be about 1500 ohms. The 6AL5 tube is widely used as a video detector.

3. Importance of video polarity.

As shown in Figure 7-2, two methods of connecting the diode load resistor are possible. In A, where the diode plate is above ground (the cathode is grounded through the i-f transformer), the video output increases in a negative direction as the carrier amplitude increases. In B, where the plate is grounded through the i-f transformer and the cathode is connected above ground, the reverse is true — the video output increases in a positive direction as the carrier amplitude is increased.

Now it will be recalled that in the RMA Standard Television Signal, black in the scene is represented by high-carrier amplitude while white in the scene is represented by low carrier amplitude. With the diode load resistor connected as in A, the output voltage across the diode load resistor will become more negative as the brightness in the scene decreases. If the diode load resistor is connected as in B, then the output voltage across R will become more positive as the brightness in the scene decreases. As brought out earlier, the polarity of the video signal applied between the grid and cathode of the picture tube must be such that an increase in carrier amplitude will cause the screen of the picture tube to become darker while a decrease in carrier amplitude causes the screen to become brighter. If the wrong polarity is used, a negative image will be produced. Therefore, if the control grid of the picture tube is connected to the plate side of the load resistor of the last video amplifier, the voltage at this end of the plate load resistor must become more negative with respect to ground with an increase in signal and less negative with respect to ground with a decrease in signal. On the other hand, if the cathode of the picture tube is connected to the plate side of the load resistor, as is the case in some GE model receivers, then the voltage at this end of the plate load resistor must become more positive with a decrease in signal, and vice-versa.

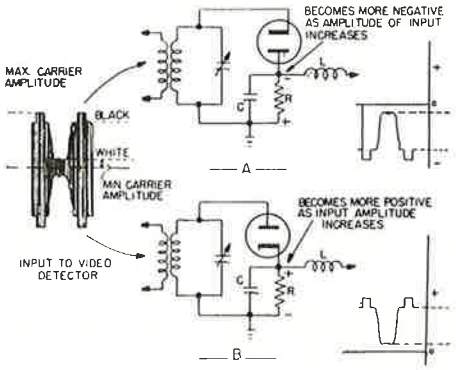


Fig. 7-2. Connection of diode load.

Now if the cathode of the picture tube connects to the plate load resistor of the video amplifier and if only one stage of video amplification is used as in the G. E. Model 801, then in order to provide the proper polarity of the video signal at the picture tube, the video detector must be connected with the diode plate above ground, as shown in (A) of Figure 7-2.

Likewise, if the signal is applied to the grid of the picture tube and the cathode held at a-c ground potential and if only one video amplifier stage is used, the video detector must be connected with cathode above ground as in (B) of Figure 7-2 in order to provide the proper polarity.

4. Shunt detector.

In some G. E. receivers, the last i-f transformer is designed so that the primary and secondary are a closely coupled, continuous winding. This places the secondary winding at B+ potential with the necessity of capacity coupling the secondary of the i-f transformer to the diode. Since the d-c return path cannot be through the i-f transformer secondary as shown in Figure 7-2, the diode load is placed in parallel with the diode as shown in Figure 7-3, and is called a shunt diode circuit. As shown in the illustration, it produces a negative-going signal across the load resistance R and is coupled to the

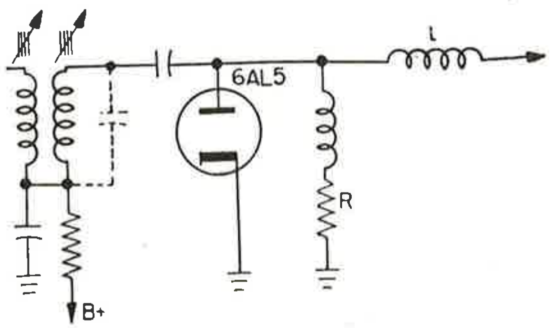


Fig. 7-3. Shunt diode circuit.

video amplifier through the i-f filter and series-peaking choke L. In this shunt diode circuit, the inter-electrode capacity of the tube provides suitable capacity for i-f filtering.

5. Video amplifier requirements.

To provide the gain, linear frequency response and good phase characteristics required to raise the detected signal to sufficient magnitude for operation of the picture tube, a compensated one or two stage resistance-coupled amplifier is employed. This is commonly designed into a single stage pentode voltage amplifier or a two stage triode amplifier. It will be noted that the video amplifier in general is similar to the normal audio amplifier with exception of the use of low plate resistor loads and compensating chokes in order to raise the high frequency response. Other considerations are the use of tubes having high mutual conductance and low input and output capacities; also, the preservation or restoration of the d-c component which is lost through the normal coupling devices. These problems will be discussed individually under separate headings in the following pages.

6. Plate Resistor choice.

An important factor in reducing the effect of tube input and output capacitance at the higher frequencies is the use of low values of plate resistors. This is illustrated in Figure 7-4. At (A) is shown a single stage RC coupled amplifier with the output capacitance C_o of one stage, the input capacitance C_i of the succeeding stage, and the distributed capacity C_d of the coupling network shown in dashed lines. At (B) is shown the equivalent circuit as far as high frequencies are concerned. At high frequencies, the reactance of the coupling capacitor C_c is very small and may be assumed to be a short circuit as far as the signal voltage is concerned and is, therefore, not shown in (B). The capacities C_i, C_o, and C_d are in parallel and may be combined into a single resultant capacity represented by C_s. Since normally R_g is so much greater than R_L in resistance value, its shunting effect is so small that it will be disregarded in order to simplify the following explanation. Thus the circuit under discussion in the following paragraphs will resolve itself into the circuit shown in (C) in Figure 7-4. The plate load impedance Z_o is then equal to the impedance of R_L and C_s in parallel and if the capacitive reactance of

C_s is low compared to R_L, then Z_o will be considerably less than it would otherwise be without the shunting effect of the capacitive reactance X_{cs}. In other words, at high frequencies

$$Z_o = \frac{R_L \times X_{cs}}{\sqrt{R^2_L + X^2_{cs}}}$$

At low frequencies, the value of X_{cs} will be very much greater than the value of R_L, and when a high impedance is connected in parallel with a low impedance, the impedance of the parallel combination is practically equal to that of the lower impedance. Therefore, at low frequencies the shunting effect of C_s can be neglected and the load impedance Z_o is then equal to R_L. If R_L is high compared to the value of X_{cs} at the highest frequency to be amplified, then the shunting effect of C_s will cause a resultant lower value of Z_o at the high frequencies.

The stage gain of a pentode is equal to the load impedance in its plate circuit multiplied by the Gm (mutual conductance) of the tube, thus u (stage gain) = $G_m \times Z_o$. From this, it is apparent that for uniform amplification at all frequencies it is necessary that the plate load impedance Z_o remain practically constant for all frequencies. To accomplish this, tubes having a low interelectrode capacitance are usually chosen in order to keep the shunt reactance X_{cs} high in comparison with R_L. Another factor is to make the value of R_L small so that its impedance to the highest frequency to be amplified is considerably less than the shunt impedance X_{cs}.

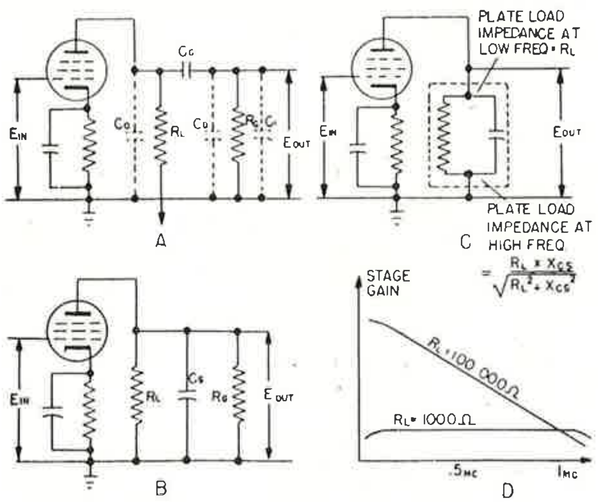


Fig. 7-4. RC-coupled amplifier.

To illustrate this further, suppose that the tube in Figure 7-4 has a G_m = 9000 micromhos = .009 mhos and that the resistor R_L = 100,000 ohms, and that the shunt capacitance C_s = 30 $\mu\mu$f. The gain of the amplifier is equal to $G_m \times Z_o$. At 100 cycles, Z_o will equal R_L since the reactance of the 30 $\mu\mu$fd. capacitor C_s at 100 cycles is approximately 50 megohms and it can be neglected.

Therefore, the gain of the amplifier at 100 cycles will be $G_m \times Z_o = .009 \times 100{,}000 = 900$. At 1 megacycle, the reactance of C_s is approximately 5000 ohms and must be considered in determining the value of Z_o. Therefore, Z_o will be equal to

$$\frac{R_L \times X_{cs}}{\sqrt{R^2_L + X^2_{cs}}} = \frac{100{,}000 \times 5000}{\sqrt{100{,}000^2 + 5000^2}}$$

is equal to approximately 5000 ohms. Therefore, the gain at 1 megacycle will be $G_m \times Z_o = .009 \times 5000 = 45$. In the preceding example the gain of the amplifier at 100 cycles is twenty times greater than that at 1 megacycle and the overall gain of the amplifier is far from being uniform where a high value of R_L is used. The response curve would look like curve 1 in Figure 7-4D.

Now suppose that the value of R_L in the preceding example is reduced to say 1000 ohms. The gain of the amplifier at 100 cycles will now be $G_m \times Z_o = .009 \times 1000 = 9$. The gain of the amplifier at 1 megacycle will be $G_m \times Z_o$ where

$$Z = \frac{R_L \times X_{cs}}{\sqrt{R^2_L + X^2_{cs}}} = \frac{1000 \times 5000}{\sqrt{1000^2 + 5000^2}} = 980 \text{ ohms,}$$

therefore, Stage Gain at 1 megacycle = $.009 \times 980 = 8.8$.

By using a low value of plate load resistance R_L, the gain of the amplifier is greatly reduced but its overall response from 100 cycles to 1 megacycle is practically flat, as shown by curve 2 in D of Figure 7-3, since its gain at 100 cycles is 9 and at 1 megacycle it is 8.8.

7. High frequency compensation.

In addition to using a low value plate load resistor, the range and gain of the amplifier over which uniform response is obtained can be improved considerably at the high frequencies by the use of high frequency compensation.

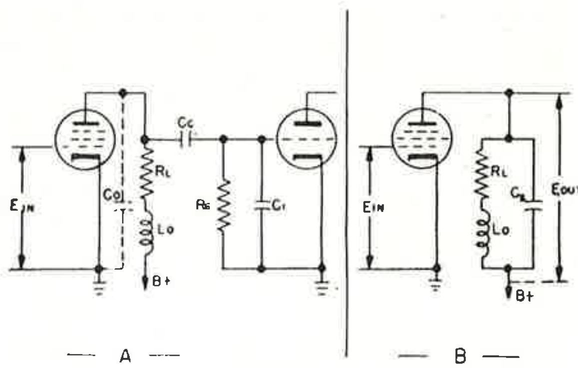


Fig. 7-5. Shunt compensation.

There are several methods used for accomplishing this, one of which is shown in (A) of Figure 7-5. In this circuit C_o is the output capacity of the first tube plus stray capacities, and C_i is the input capacity of the second tube plus stray capacities. R_g is the grid coupling resistor which should be much higher than R_L. R_L is the plate load resistor, and L is an inductance used for compensation.

This is known as the shunt peaking method of high frequency compensation and consists of a small inductance L in series with the plate load resistor R_L. At high frequencies, this inductance forms a broadly tuned shunt resonant circuit (hence the term "shunt peaking"), with the shunt capacity C_s as in the equivalent circuit at high frequency shown in (B) of Figure 7-5, and serves to maintain an essentially constant load impedance up to the highest frequency to be passed by the amplifier. The value of L is chosen so that the resonant frequency of the combination is somewhat higher than the highest frequency to be amplified. This method of compensation has practically no effect on the mid-range and low frequencies, and the equivalent circuit at low frequencies is shown in (B) of Figure 7-4.

The maximum value of R_L permissible for flat compensation is dependent on the total shunt capacity and the upper frequency limit of the amplifier. In practice, the plate load resistor R_L is usually made equal in value to the reactance of the total shunt capacity C_s at the highest frequency to be amplified. The reactance of the compensating coil L is usually made one-half the value of the reactance of C_s at the highest frequency to be amplified.

The relative gain of this type of network at the highest frequency to be passed is the same for the mid-frequency video range, where the gain $= G_m \times R_L$.

Another type of compensation network, known as series compensation, is shown in Figure 7-6. With this method the compensating inductance L_c is connected in series with the coupling capacitor C_c. This method gives approximately 50% more gain than for shunt compensation. The additional gain is due to the fact that the output capacity C_o is isolated from the input capacity C_i by the series inductance L_c, with the result that the load resistor R_L is shunted only by the capacitance C_o and the value of R_L may be proportionately increased, thereby increasing the gain of the amplifier without impairing its high frequency response.

Also, at high frequencies the inductance L_c forms a broadly tuned series resonant circuit with the input capacity C_1 and the voltage developed across

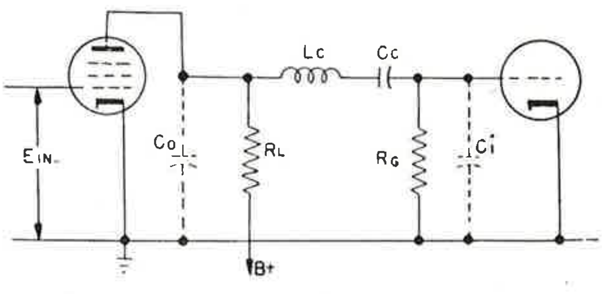


Fig. 7-6. Series compensation.

C_i tends to rise as the frequency approaches the resonant point of L_c and C_i, which keeps the gain up at the high frequencies. The point at which L_c and C_i resonate is somewhat higher in frequency than the highest frequency to be passed by the amplifier. To obtain maximum gain with this method, it is necessary that the effective ratio of C_i/C_o equal 2. The value of the plate load resistor R_L is usually 1.5 times the value of the reactance of the total shunt capacitance $C_1 + C_o$ at the highest frequency to be passed.

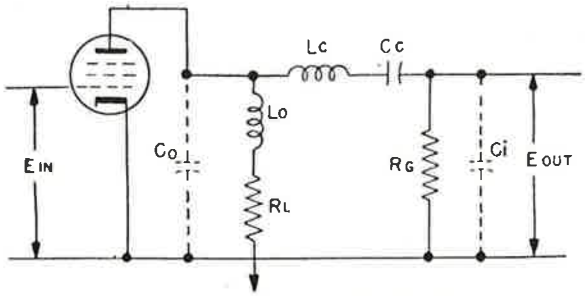


Fig. 7-7. Shunt-series compensation.

A circuit that gives approximately 80% more gain than the shunt peaking method is that shown in Figure 7-7 and is known as the combination or shunt-series compensation network. This circuit has the advantages of both methods. The plate load resistor R_L can be made somewhat larger than in either the simple shunt peaking method or the series peaking method and, therefore, results in increased gain.

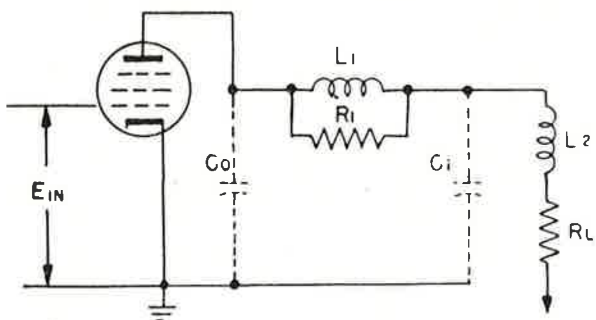


Fig. 7-8. H.F. compensating circuit.

A high gain video compensation circuit employing both shunt and series compensation and used in the G. E. Model 802 television receiver is shown in Figure 7-8. In this circuit, C_i and L_2 form a shunt resonant circuit while C_o and L_1 form a series resonant circuit. R_1 broadens the tuned circuit of C_o and L_1. The video signal that is applied to the picture tube is taken off at the mid-point of L_1 and L_2. By taking the picture signal off at this point, between L_1 and L_2, the loading effect of the input capacity is less than if the signal were taken off at the end as in Figure 7-8.

Values of L, C, and R employed in a typical circuit of this type are as follows:

$C_o = 9.5\ \mu\mu f.$ = tube output capacitance + all stray capacitances
$C_i = 19\ \mu\mu f.$ = picture tube or next tube input capacitance + all strays
$L_1 = 270$ uh.
$L_2 = 165$ uh.
$R_1 = 20{,}000$ ohms
$R_L = 3{,}500$ ohms.

8. Low frequency response.

To maintain background illumination having true values of lights and shadows over an appreciable length of time, it is necessary to have good low frequency response and also restore the d-c component to the video signal before application to the picture tube. This section will first discuss the problems involved in obtaining good low frequency response, leaving the discussion of the d-c component to a following section.

The low frequency response of the video amplifier is extended much lower than is normally required in audio amplifier design, the lower frequency limit being approximately 6 cycles per second. Since capacity coupling is used in the design of most video amplifiers, the choice of values of the coupling capacitor and the grid resistor will determine how good the low frequency response will be.

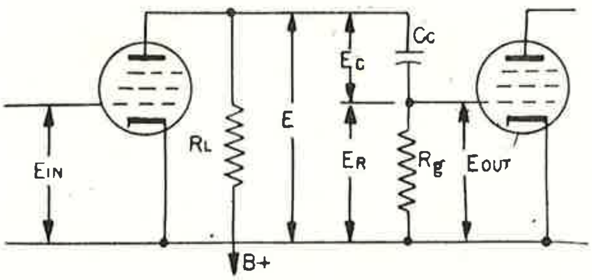


Fig. 7-9. Coupling capacitor in RC amplifier.

At low frequencies, the reactance of coupling capacitor is appreciable compared to the grid resistance, hence part of the video signal is lost in the capacitor, producing a decrease in the low frequency response. As shown in Figure 7-9, the coupling capacitor C_c and the grid resistor R_g form a voltage divider and as the frequency is decreased, reactance of C_c is increased, which results in less of the signal voltage developed across the plate load resistor R_L from appearing across R_g; that is, between grid and cathode of the following tube. The reactance of the coupling capacitor can be reduced by making the capacity larger and, thus, reduce the loss of signal voltage across the coupling capacitor, but the extent to which the value of the coupling capacitor can be increased is limited to a practical value of about 0.1 microfarad because the increase of the physical size of the capacitor will increase the stray capacitance and will affect the high frequency response. The value of the grid resistor R_g can be increased so that it will be very high compared to the reactance of the capacitor at low frequencies and, thus, a greater percentage of the signal voltage applied to the series combination will appear across R_g. However, R_g is limited in resistance value set by the manufacturer for the particular tube type used otherwise faulty operation may result.

In addition to the use of as large a coupling capacitor and grid resistor as is practical, other precautions must be taken so as not to impair the low frequency response. One of these sources of poor low frequency response may be due to the use of cathode resistor self bias in one of the video amplifier stages.

The signal variations appearing across this cathode resistor must be by-passed in order not to lose gain due to degeneration. In order to satisfactorily by-pass the signal variations across the cathode resistor, the reactance of the by-pass capacitor must be very low compared to the value of the cathode resistor that it by-passes. Its capacitance will, therefore, be quite large if it is to prevent degeneration at very low frequencies. Since the by-pass capacitor is usually made large enough so that at the lowest frequency to be passed, its reactance is lower than one-tenth the value of the cathode resistor; in practice the value of this by-pass may be well over a thousand microfarads. Even when a very large by-pass capacitor is used, a certain amount of degeneration takes place at the low frequencies. This effect can be eliminated by obtaining bias in other ways, such as by grid rectification.

Loss of gain at low frequencies is also caused by the impedance in the screen grid circuit. Usually, screen voltage is obtained through a series dropping resistor which is by-passed to cathode. At low frequencies, the reactance of the screen by-pass capacitor increases so that it no longer is an effective by-pass and allows signal voltage variations to appear between the screen grid and cathode which reduce the gain of the amplifier. However, this can be compensated for by the proper choice of screen by-pass capacitor in relation to the dynamic screen-to-cathode resistance of the tube.

9. Phase distortion.

An important consideration in the case of video amplifiers that has practically no importance in ordinary radio receivers is phase shift. At low frequencies, phase shift occurs due to the coupling capacitor C_c in conjunction with the grid resistor R_g, the cathode impedance when cathode bias is incorporated, and the impedance in the screen grid circuit. At high frequencies, phase shift is introduced by the stray capacity in the circuit.

If an amplifier has no phase distortion or time delay, then signals of all frequencies will pass through at the same rate. Thus, if a 1000 cycle modulation component and a 1.0 megacycle modulation component were transmitted with the peak amplitudes separated by — let us say — one-tenth of a microsecond, then after passing through the video amplifiers of the receiver they will still reach their peak amplitudes with the same separation of one-tenth of a microsecond when applied to the picture tube grid if there is no phase distortion present. This is a most desirable condition. However, if phase distortion is present, then even though both modulation components reach other peak amplitudes with a one-tenth of a microsecond separation when transmitted, they will not do so at the grid of the picture tube because one modulation component will take more time to pass through the amplifier than the other.

With phase distortion present in the amplifier, the picture on the screen will not appear sharp but will be blurred since the separation between the peak amplitudes of the various modulation components will not be the same as transmitted and the relative position of one modulation component with respect to the other will be displaced by the amount of time delay or phase shift in the amplifier. In addition to the distortion of the picture, phase shift in a video amplifier will change the waveform of the synchronizing pulses which are rectangular in shape and consist of a fundamental frequency plus a number of harmonic frequencies. In order to satisfactorily amplify a square wave or rectangular waveform, the peak amplitude of the fundamental and higher order harmonics must occur simultaneously in the output of the amplifier. However, if phase distortion is present, there will be a time delay between the fundamental and its higher order harmonics and they will not reach their peak amplitudes at the same instant. This causes the sharp edges of the rectangular pulse to become rounded off and may result in poor synchronization of the sweep generators.

As mentioned previously, low frequency phase shift is introduced by the coupling capacitor and grid resistor and will cause the high frequencies to lag the low frequencies. Referring to Figure 7-9, the coupling capacitor C_c and grid resistor R_g form a voltage divider across which the signal voltage is applied. At high frequencies the reactance of C_c is so small that it can be neglected and the series circuit of C_c and R_g becomes predominately resistive. Since the circuit is predominately resistive, the current through the series combination will be in phase with the applied signal voltage and, therefore, the voltage at the grid due to the IR drop across R_g will be in phase with the signal voltage across the combination. However, at low frequencies the reactance of the coupling capacitor is appreciable compared to the grid resistance, and the series circuit becomes reactive. Due to the capacitive reactance in the circuit, the current through the series combination will lead the applied signal voltage. Therefore, the voltage at the grid due to the IR drop across R_g will lead the signal voltage applied across the combination. Thus, the high frequencies will lag the low frequencies due to phase distortion introduced by the coupling capacitor C_c and grid resistor R_g. Thus again the choice of the coupling capacitor is important to the final pictorial perfection.

An ideal amplifier would have zero phase shift or time delay, but such is not the case in practice. If the time delay is uniform for all frequencies, then phase shift is not serious. For instance, if a 1 mc signal were shifted in phase by 90°, the time delay would be one-quarter microsecond. Also, a one-quarter microsecond delay would occur if a 100 kc signal were shifted in phase by 9°. If the phase shift is proportional to frequency, uniform time delay results for all picture components and a perfect picture is reproduced.

Fortunately, essentially uniform time delay over the entire range of video frequencies results when the video circuits are compensated for uniform response.

10. Direct current component.

The picture signal applied between grid and cathode of the picture tube must contain the direct current component of the picture signal. It will be recalled that the composite picture signal which modulates the carrier is a pulsating voltage which varies in one direction from a d-c reference level which establishes the peak level or voltage of the sync pulses, and, regardless of the character of the picture information being transmitted (whether it is all white, all black, or any other combination), the sync peaks always return to this d-c level. This d-c level is established at the transmitter and remains constant during transmission.

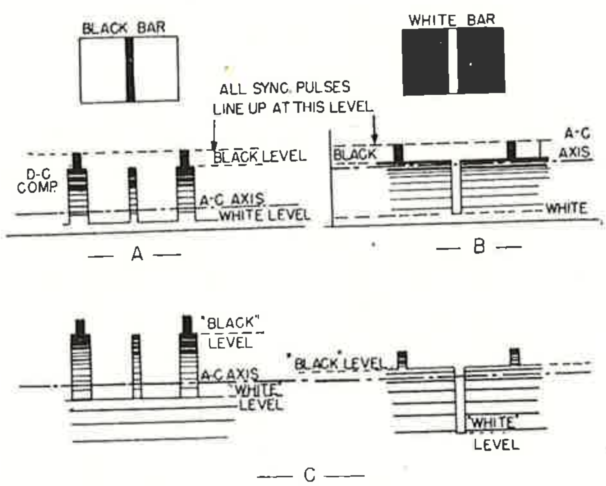


Fig. 7-10. Direct-current component.

The peak-to-peak variations of voltage which may be represented by an alternating current video waveform conveys only the instantaneous variations of the illumination of the screen and does not convey the average or background illumination. In order to obtain the condition where the synchronizing pulses and the black picture elements always occur at the same voltage level regardless of picture information, it is necessary that the effect of the a-c axis be removed and that the signal be displaced the correct amount below a reference, called the "d-c level", as established at the transmitter. As an example, see Figure 7-10; (A) represents variations of the picture signal for one line of a scene having a white field with a narrow black vertical bar; and (B) represents the other extreme of a picture having a black field with a narrow white vertical bar. It will be noted that in both cases the tips of the sync pulses line up at the d-c level established at the transmitter. By taking (A) and (B) and bringing them together on the same a-c axis as shown in (C), it will be noted that the height of the synchronizing pulses as well as the black level above the a-c axis are not the same for both picture presentations but vary according to the average amount of white in the scene being televised. This variation of the reference levels for the sync and black level would be the case if the video signal was passed through a capacitor coupling device, such as might be found in a video amplifier, without any device incorporated to restore the d-c level. Such a condition would cause a reproduction of a picture without the true levels of background illumination being represented under these conditions. This can be illustrated by Figure 7-11, which shows the light output versus video signal amplitude when the video is applied to the picture tube through a coupling capacitor so as to cause the d-c component to be lost. In (A) is represented a predominantly white scene wherein the Brightness Control at the receiver has been adjusted to give the proper gradation between black and white. When the scene is suddenly changed to one that is predominantly black, the shift in the a-c axis causes the video signal to change the illumination of the picture tube as shown in (B). This condition does not permit the black level from reaching cut-off bias on the picture tube with the result that black will reproduce as grey. Likewise, if the picture is adjusted when the scene is predominantly white, then the grays will reproduce as black when the scene shifts to the other extreme. The necessity of bringing the black level to a common basis irrespective of picture presentation is thus established.

Since the r-f and i-f portions of the receiver do not materially alter the modulation envelope, then the d-c component will be present upon detection just as it was sent out at the transmitter. However, any coupling device introduced in the circuit after detection will cause the loss of the d-c component. This makes it appear as though we cannot use the ordinary a-c coupled amplifier. Fortunately, there is a method whereby the d-c component can effectively be restored by means of a peak rectifier if connected in the circuit after the last coupling capacitor is used.

The restoration of the d-c component is exemplified by using a simple sine wave as shown in Figure 7-12. In (A) is shown the sine wave with a peak voltage value of 10 volts on each side of the axis. This may represent a sine wave at the grid of a tube which is capacity coupled to a preceding detector. At the dectector this sine wave would be all on one side of the zero axis. Now, suppose we insert a d-c potential of 10 volts and with the polarity shown in series with this sine wave, as in (B) of Figure 7-12. The sine wave will now vary in a negative direction below a reference level of zero

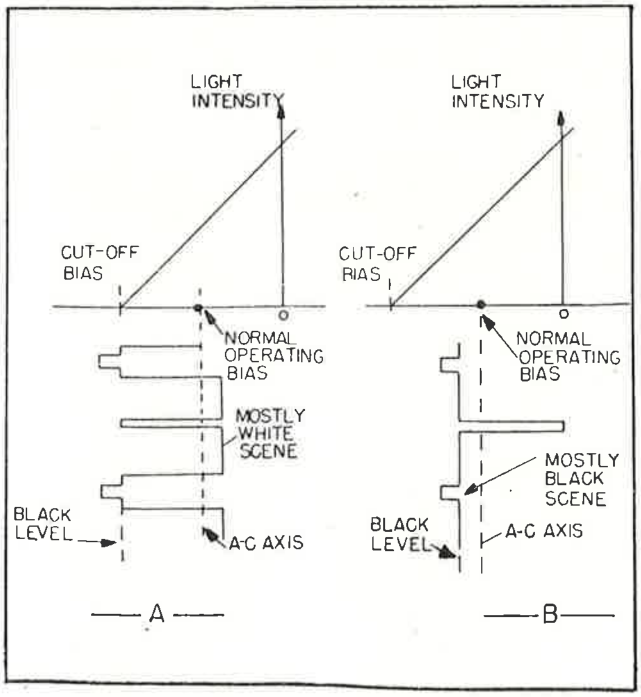


Fig. 7-11. DC component and light intensity.

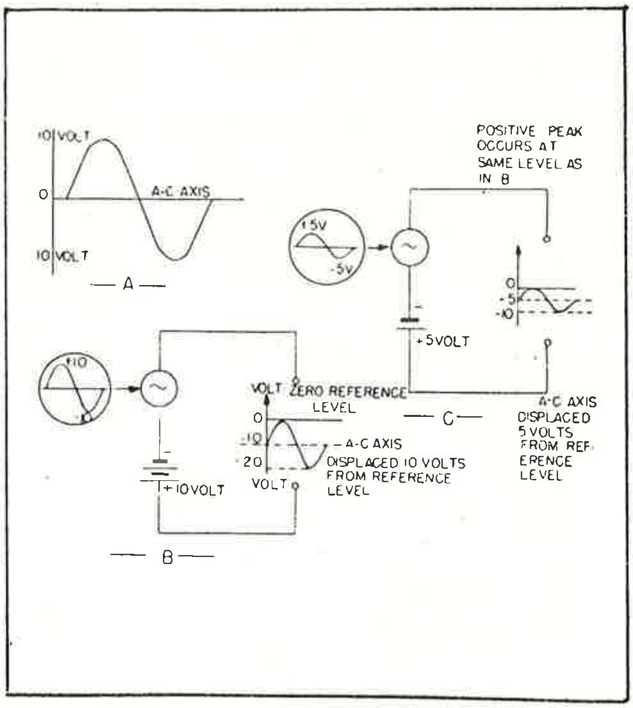


Fig. 7-12. DC restoration and sine wave.

volts with the positive peak of the sine wave just reaching this reference level, as shown by the curve to the right of (B). The d-c voltage placed in series with the sine wave is called its d-c component and is equal to the peak value of the sine wave above its a-c axis. Now, let us consider another sine wave which has only a 5 volt peak about its a-c axis and we desire to make its positive peak reach the same reference level, as in the case of the 10 volt peak sine wave. This can be accomplished by inserting a d-c potential in series with the 5 volt sine wave, as in (C) of Figure 7-12. In this case the d-c component is only 5 volts and displaces the a-c axis of the sine wave just enough to allow the positive peak to reach the zero reference level. In this way, by combining the a-c component of a signal with the correct value of d-c component, the positive peaks are lined up at a certain reference level regardless of the amplitude of the a-c component. The reference level at which the peaks line up need not be zero volts, as in the previous illustrations, but can be any value (either positive or negative). In effect, this is what we must do to the video waveshape so that the peaks of the sync pulses all line up to the same reference axis.

Instead of using a battery to restore the d-c component as shown in Figure 7-12, in a television receiver we make use of a diode connected as a peak rectifier to take the place of this battery. By referring back to Figure 7-12, it will be seen that when a d-c voltage is developed which is just equal to the amplitude of the synchronizing peaks from the a-c axis and then is added to the a-c signal voltage, that the effect of the d-c level will be restored and the top of each sync pulse will come to a common voltage level. Likewise, since the peaks of the sync pulses have a fixed amplitude relationship to the black level (blanking pulse level); irrespective of the placement of the a-c axis, it is apparent that if the sync pulse peaks are lined up to a common level, then the black level on each line will be automatically lined up.

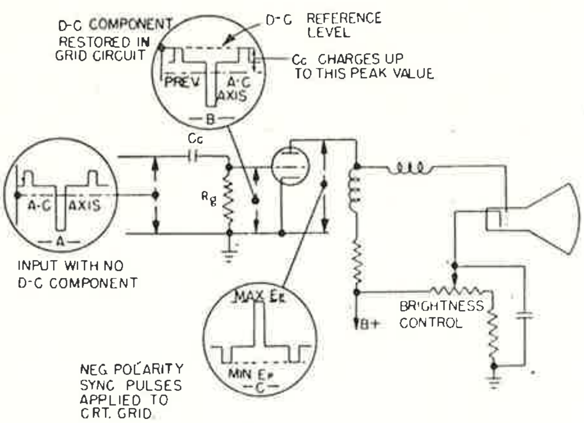


Fig. 7-13. Output stage of video amplifier.

Figure 7-13 shows the output stage of a video amplifier wherein d-c restoration takes place in the grid circuit of this output tube. This is a popular method of bringing this about since it does not require a separate diode section for d-c restoration. The requisites for operation of this circuit are that the sync pulse peaks have a positive-going polarity at the grid where the restoration takes place and that the plate of this last video amplifier be directly connected to the picture tube without the use of a coupling capacitor. This later requirement necessitates the operation of the picture tube grid and cathode elements at an elevated voltage (positive) with respect to ground.

The output stage is shown operated without any bias other than that furnished by grid rectification. With the application of the waveshape as shown in Figure 7-13 (A), the output tube draws grid current for the duration of the voltage waveshape above the a-c axis. This causes the grid side of the coupling capacitor to charge up negatively to the peak value represented by the peak amplitude voltage of the sync pulse above the a-c axis. The a-c voltage when superimposed on this peak bias causes the a-c axis to be displaced below the d-c reference level as shown in (B) of Figure 7-13. Irrespective of the picture make-up, this rectification in the grid circuit causes all signals impressed to line-up so that the peak value represented by the tips of the sync pulses always come to the same common axis. This is the condition which existed at the detect or before the d-c component was removed.

The value of the coupling capacitor C_c and grid resistor R_g in Figure 7-13 is somewhat critical in that the combination of R and C should not be such to cause the capacitor C_c to discharge very much when the signal is at its most negative value. In general, the time constant must be sufficiently long to maintain the bias substantially constant during the picture intervals between line synchronizing pulses, but should also be sufficiently short to allow the bias to vary with the change in the average brightness of the picture.

The picture tube grid is connected directly to the video amplifier plate so that the d-c component which has been restored in the grid circuit of the video amplifier will not again be lost; bias for the picture tube is obtained by connecting the cathode to a point of higher positive potential than the grid, as shown in Figure 7-13.

It will be noted from the foregoing that grid current flows during the peaks of the synchronizing pulses, thus maintaining them at approximately the zero bias point regardless of the position of the a-c axis, and that the black level which is always the same amount lower than the synchronizing peaks again occurs at the same relative level that it did in the video detector before it was lost. The signal appearing in the plate circuit will be reversed in phase with the synchronizing peaks, being the most negative portion of the signal as shown in Figure 7-13 (C). The brightness control is adjusted so that the bias on the picture tube is such that the black level of the signal just extinguishes the spot.

The resultant as it affects the light output of the picture tube is shown in Figure 7-14. With the black level clamped to a common axis, each scene will have the proper illumination gradations.

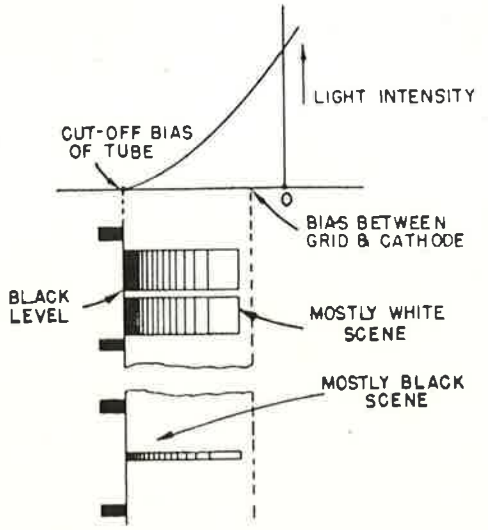


Fig. 7-14. Light intensity and black level.

Another common method of restoring the d-c component is to connect a diode as a peak rectifier at the picture tube grid or cathode element, depending upon which element the video signal is applied to. In Figure 7-15 is shown a single stage video amplifier which gives an output signal wherein the sync signal is positive-going. With this polarity, it is necessary to apply the video signal to the cathode. Since a coupling capacitor C_c is used between the video output stage and the picture tube, the d-c component is lost and must be restored. By connecting a diode as shown so that it acts as a peak rectifier for all signal above the a-c axis as shown in Figure 7-15 (A), the peaks of the sync pulses are all lined up to the same common axis or level as they appeared at the detector.

The operating point of the picture tube is established by the adjustment of the Brightness Control.

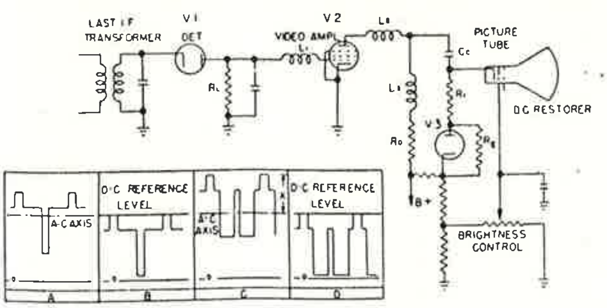


Fig. 7-15. Output stage of video amplifier and sync. pulses.

Any value of signal extending above the a-c axis as shown in Figure 7-15 (A) will cause the diode to conduct (plate becomes more positive than cathode) and establish a voltage across R_g which is equivalent to the difference between the a-c axis and the peak of the sync pulses as shown in Figure 7-15 (B). This voltage across R_g is such as to subtract from the voltage at the cathode end of this resistor so that the actual voltage appearing at the cathode of the picture tube at the peak of the sync pulse will be as shown in Figure 7-15 (B) instead of that in Figure 7-15 (A), which would be the case if no d-c restoration was present. If another example was taken showing a video signal which is predominantly white, the diode will rectify and charge C_c to a peak value represented by "X" in (C) and thus will cause the video signal to come to a peak value at the same axis, see (D), as in (B). In comparison, (A) and (C) which represent the same video signals without d-c restoration, it will be noted that the peak amplitude representing the tips of the sync pulses are not the same.

If the video were applied to the grid of the picture tube instead of the cathode circuit, the signal would have the opposite polarity. In this case, the connections to the diode would have to be reversed.

The resistor R_I does not enter into the proper operation of the d-c restorer; it is used to isolate the diode from the cathode of the picture so the diode capacity will not reduce the high frequency response of the video output.

11. Typical circuits.

Figure 7-15 shows a typical video detector and single stage video amplifier. Since a coupling capacitor C_c is used in the plate circuit of the amplifier tube V_2, a diode, V_3, is connected in the cathode circuit of the picture tube to restore the d-c component. This restorer action was described in the preceding paragraphs.

By the use of a high G_m pentode tube, only a single stage of video amplification is required to give the necessary video signal gain. A series diode is used with the plate connected above ground. The demodulated signal appearing across the diode load, R_L, will be negative-going. This signal is directly coupled to the grid of the video amplifier, V_2, through L_1, which serves to keep the video i-f signal out of the video amplifier and, also, acts as a series peaking coil to provide high frequency compensation.

Since the video signal in the plate of the amplifier tube is positive-going as shown in Figure 7-15 (A), the output tube is coupled through capacitor C_c to the cathode circuit of the picture tube. This produces the same effect as a video signal of opposite polarity applied to the grid circuit of the picture tube. High frequency compensation in the plate circuit of V_2 is obtained by the use of shunt and series peaking chokes. Choke L_3 in combination with the cathode-to-grid capacity of the picture, provides shunt peaking; while L_2 and the output capacity of V_2 provides series peaking.

The Brightness Control merely changes the amount of positive potential on the grid. As long as this potential is less than the positive voltage applied to the cathode circuit through the d-c restorer circuit components, the picture tube will be operating with a suitable bias.

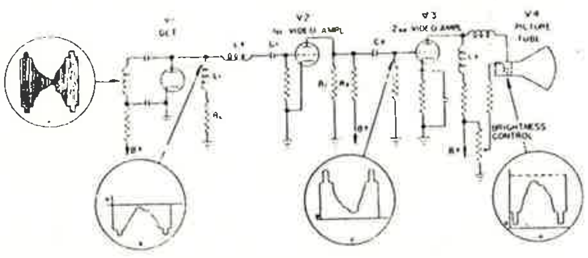


Fig. 7-16. Detector and video amplifier.

In Figure 7-16 is shown a two-stage triode coupled video amplifier which follows a shunt-connected diode. The waveshapes are shown for various parts of the circuit. The signal appearing across the detector load resistance R_L is negative-going as shown in (B). L_1 is a shunt peaking choke used to compensate for the capacity of the diode and circuit. L_2 is a series peaking choke used to couple V_1 to V_2 and is also used to prevent the video i-f from being passed on through the video amplifier. V_2 is an amplifier and limiter. The plate potential to V_2 is made relatively low by connecting its load resistance R_1 to ground and then bleeding down the B+ through R_2. The grid circuit of V_3, having a positive-going signal, is used to restore the d-c component lost by passage of the signal through the coupling capacitors C_1 and C_2. The operation of this restorer was described in the previous section 10.

Because of the restoration of the d-c component in the grid of V_3, the plate is direct-coupled to the grid of the picture tube through the series peaking choke L_4. With the application of the positive plate voltage of V_3 onto the grid of the picture tube, the cathode of the picture tube must be elevated at a B+ voltage to provide suitable bias on V_4. The amount of B+ or bias voltage is controlled by the Brightness Control connected in the cathode circuit of V_4.

In practice, a triode normally has a higher input capacity than a pentode. However, with the use of the miniature 12AT7 and 12AU7 dual triode tubes, their input and output capacity is far smaller than other type high G_m pentodes, such as the Type 6AC7, which is used in Figure 7-15. Thus the compensation required is not as great.

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Radiotron — 5718 is a medium-mu subminiature triode designed especially for use as an rf power amplifier and oscillator in UHF applications where dependable performance under shock and vibration is a prime consideration. It is capable of giving a useful power output of nearly one watt at 500 Mc/s. Operation with full input is permissible up to 1000 Mc/s.

Featured in the 5718 is high transconductance, a pure-tungsten heater to give long life under conditions of frequent on-off switching, and a compact design in which special attention has been given to structural details that provide increased mount strength to resist shock and vibration.

Because of its high transconductance, the 5718 is suitable for use in cathode-follower, multivibrator, and blocking-oscillator circuits. It is also useful as a resistance-coupled amplifier.

The "premium" quality of each 5718 is assured during manufacture by rigid controls and rigorous tests.

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Radiotron — 6211 is a new medium-mu twin triode of the 9-pin miniature type designed especially for frequency-divider circuits in electronic computers and other "on-off" control applications involving long periods of operation under cutoff conditions.

In such control service, the 6211 maintains its emission capabilities even after long periods of operation under cutoff conditions and, therefore, provides good consistency of plate current during its "on" cycles. Furthermore, balance of cutoff bias between the two units is closely controlled during manufacture. Production controls correlated with typical electronic computer operating conditions, and rigorous tests for shorts and leakage, insure long and dependable performance from the 6211.

The 6211 has separate terminals for each cathode to facilitate flexibility of circuit arrangement, and a mid-tapped heater to permit operation from either a 6.3-volt or 12.6-volt supply. The heater is made of pure tungsten to give long life under conditions of frequent on-off switching.

By K. Fowler and H. Lippert.

SWEEP GENERATORS

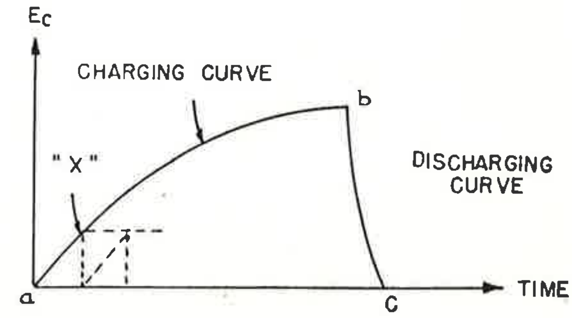
1. Introduction.

In order to reproduce the original scene televised at the transmitter, it is essential that the electron spot on the picture tube accurately follow the same path as the spot used in scanning the camera mosaic. As pointed out previously, the motion of the camera spot is the resultant of two sawtooth waves acting simultaneously on the electron beam, one moving it across the scene at a uniform speed at 15,750 times per second, and the other moving it downward at a uniform speed at 60 times per second. In the receiver it is necessary that we generate sawtooth waves with exactly the same timing and uniform repetition as those used at the camera tube.

2. Sawtooth oscillators.

Most sawtooth oscillators of sweep generators depend fundamentally on the charge and discharge of a capacitor to produce the wave, and there are innumerable circuit variations for charging and discharging the capacitor. When a capacitor and a resistor are connected in series to a source of potential as shown in Figure 8-1, the voltage across the capacitor varies with time as indicated. It will be noted that the charging cycle or rise in voltage across the capacitor is not linear but goes from a zero value toward a maximum value in the form of an exponential curve. The time required for the capacitor to charge to the maximum value depends on the time constant of the resistor-capacitor combination. The larger the value of the series resistor, the greater the time required to fully charge the capacitor since the resistor is in series with the capacitor and, therefore, limits the rate of current flow. Likewise, the larger the capacitor, the greater will be the time required for it to become fully charged. Now if S_1 is opened and S_2 is closed, the capacitor will discharge in much less time than that required for it to become charged, as indicated by the discharge portion of the curve b-c, since the series resistor is not in the circuit to limit the rate of current flow. It will be noted that the lower 10% of the charging curve is fairly linear and is suitable for use as a portion of a sawtooth wave.

The gradual rise in voltage from point "a" to point "b" represents the trace portion of the sawtooth wave, while the sudden drop from point "b" to "c" represents the retrace portion. It is necessary that for good reproduction of the picture that the trace portion of the sawtooth wave be linear, otherwise the picture will be crowded at one point, stretched out at another point, and in general appear very distorted. However, if a suitable switching arrangement is provided to abruptly discharge the capacitor at point "x" before it has time to acquire a full charge, then we can make use of the linear portion of the charging cycle of the capacitor and produce a sawtooth waveform the frequency of which depends upon the number of times per second that the capacitor is charged to point "x" and then discharged. Since the capacitor is only charged to approximately 10% of maximum charge in order to make use of the linear portion of the charging curve, the amplitude of the sawtooth waveform thus produced will be small and is, therefore, amplified before applying it to the deflection circuits of the picture tube, as will be discussed in the following chapter.



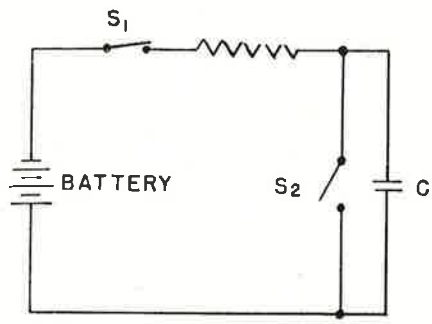


Fig. 8-1. Charging and discharging of a capacitor.

A sawtooth oscillator or sweep generator is simply a circuit which allows the capacitor to charge for a short time over the linear portion of the curve and then abruptly discharges it. This cycle is repeated over and over again to give continuous sawtooth waves of the proper frequency. Since it would not be practical to use ordinary switches to charge and discharge the capacitor, the switching is done electronically by a gas or high vacuum tube in practical sawtooth oscillators.

There are three general types of sawtooth oscillators:

(a) The Gas Tube Oscillator
(b) The Multivibrator
(c) The Blocking Oscillator

By courtesy of AGE, with acknowledgement to International General Electric Co. of U.S.A.

The sweep generators used in television receivers are free-running; that is, they do not require the application of an external pulse to keep them running. However, it is necessary to apply an external synchronizing pulse in order to accurately control the free-running speed of the oscillator.

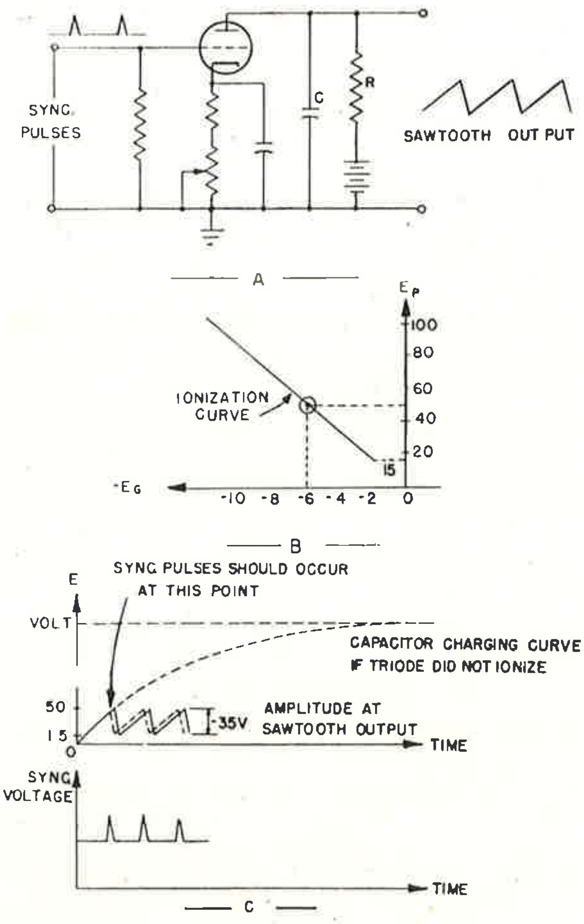


Fig. 8-2. Gas tube oscillator.

3. Gas tube oscillator.

One of the simplest types of sawtooth oscillators makes use of a gas-filled triode, such as the 884 or 885, as shown in (A) of Figure 8-2. Although the gas-filled type of tube is seldom used as a sawtooth generator in television receivers, it is widely used as a sawtooth generator in oscilloscopes and is the type that the majority of radio technicians are most familiar with. Therefore, it will be used as a starting point before considering the somewhat more complicated sawtooth generators that are usually used in television receivers.

Tubes of this type have a very low internal voltage drop when the gas is ionized and a very high one when not ionized. In other words, when the gas is not ionized, the internal resistance of the tube is very high and when the gas ionizes, the internal resistance of the tube becomes very low, thus providing a low resistance path across the charging capacitor C. This quickly removes the charge that accumulated on the condenser during the time when the gas was not ionized and the internal resistance of the tube was very high. The tube, therefore, acts as a switch, allowing the capacitor to charge up through R during the time when the gas is not ionized, and then discharging the capacitor through the tube when the gas in the tube becomes ionized. Ionization is caused by electrons in the region between the grid and plate striking the gas molecules with sufficient energy to transform some of the gas molecules into positive ions. When this occurs, the internal resistance of the tube becomes very low and the voltage drop in the tube is of the order of 15 volts. Whether the electrons have sufficient velocity to ionize the gas is of course a function of both the grid and plate voltages. (B) of Figure 8-2 shows typical control characteristic of a gas-filled triode or thyratron tube, an inspection of which shows that if the potential of the grid is –6 volts, the tube will not ionize until the voltage between plate and cathode is 50 volts, as indicated by the circle on the ionization curve. Since the capacitor C is connected between plate and ground with a –6 volt bias on the grid, the potential on capacitor C when plate voltage is first applied will gradually increase to 50 volts, as indicated by the curve in (C), Figure 8-2, at which time the gas in the tube will ionize and quickly discharge the capacitor down to 15 volts. When the potential between plate and cathode reaches 15 volts due to discharging the capacitor through the tube, ionization will cease as indicated by the ionization curve in (B).

Since the gas is no longer ionized, the tube presents a very high resistance path across the capacitor and the capacitor will again gradually charge up to 50 volts, at which time the gas again becomes ionized. The tube, therefore, presents a very low resistance across the capacitor, discharging it down to 15 volts, thus repeating the cycle as indicated by the output wave in (A), and we have sustained oscillations. The output waveshape is 35 volts peak-to-peak sawtooth wave, providing that the values of R, C, and the plate supply voltage E_b are such that only the straight or nearly straight portion of the capacitor's charging curve is used. The dashed curve in (C), Figure 8-2, is the capacitor charging curve if the tube had sufficient bias to prevent ionization even at the full plate supply potential. Since it is not desirable to operate over the non-linear portion of the capacitor charging wave, the bias is set at a level which will cause the gas in the tube to ionize and discharge the capacitor before it reaches the non-linear region of its charging curve.

The frequency of the sawtooth oscillator of Figure 8-2 is determined by the size of R, C, and the plate supply voltage, since these all affect the time required for C to charge from 15 to 50 volts. Also, the size or amplitude of the sawtooth waveform as well as its frequency are affected by the bias, inasmuch as the greater the bias, the greater will be the potential required between plate and cathode for ionization to occur and the greater will be the time required for the capacitor to charge up to this ionization potential. Therefore, if the bias is increased the amplitude of the sawtooth waveform will be greater and its frequency will be reduced. On the other hand, if the bias is reduced, the ionization potential will be reduced, thus decreasing the amplitude of the sawtooth waveform, and increasing its frequency since less time is required for the capacitor to charge up to this lower ionization potential.

With R, C, and the plate supply voltage E_b kept at a constant value, it is therefore apparent that if the bias of the gas tube is varied over a limited range by placing a variable resistor in the cathode circuit as indicated in Figure 8-2, it will act as a speed or frequency control and keep the free-running frequency of the oscillator within a certain limited range. The frequency of oscillation can also be controlled over a limited range by making the resistor R, through which C receives its charge, variable. If the value of R is increased, it will increase the time required for the capacitor to charge up to a given ionization potential and, therefore, decrease the number of times that the capacitor will charge and discharge per second, thus decreasing the frequency of oscillation. Likewise, if the value of R is decreased, the capacitor will charge up to a given ionization potential more rapidly, thus increasing the frequency of oscillation.

The oscillator of Figure 8-2 without the injection of synchronizing pulses, will normally operate over a range of frequencies as determined by the factors just mentioned and is said to be free-running. It does not require the injection of an external signal or pulse to make it produce a sawtooth waveform. However, it is absoultely essential that the frequency of the sweep generators in a television receiver be locked in or synchronized at exactly the same frequency as those at the television station. This is accomplished by providing synchronizing pulses in the transmitted signal which occur once each cycle and when properly applied to the sawtooth generator accurately controls its frequency so that it is not susceptible to variations in frequency due to slight temperature or voltage changes.

The sweep generators in a television receiver, regardless of type, are so designed that the free-running frequency of the vertical sawtooth generator can be varied slightly above and below 60 cps, while the free-running frequency of the horizontal sawtooth generator can be varied slightly above and below 15,750 cps. When the synchronizing pulses are applied, the vertical sweep generator runs at exactly 60 cps, while the horizontal sweep generator runs at exactly 15,750 cps and they will not vary due to slight variations in temperature or voltage.

4. Synchronization of gas tube oscillator.

Synchronization may be accomplished by impressing a positive sync pulse on the grid of the gas tube, which will cause the tube to ionize and discharge the capacitor before it otherwise would with the fixed bias. For instance, suppose that with a fixed bias of –6 volts, ionization occurs when the cathode-to-plate potential reaches 50 volts, and that as the charge on the capacitor gradually increases and nears this ionization level of 50 volts, a positive voltage is applied to the grid of the gas discharge tube. This positive voltage applied to the grid of the tube will make the grid less negative than it normally would be, and, therefore, lowers the ionization voltage required between plate and cathode to perhaps 48 volts, and the capacitor will discharge somewhat earlier than it would if no synchronizing voltage had been applied to the grid of the tube. Thus, by applying synchronizing pulses to the grid of the gas tube at regular intervals, the capacitor will be discharged at exactly the same level during every cycle and the frequency of the sawtooth waveform will be synchronized with the frequency at which these synchronizing pulses initiate the discharge of the capacitor. If the free-running frequency of the oscillator is such that the tube has already ionized when the synchronizing pulse is applied, the synchronizing pulse will be ineffective since the capacitor will already be discharging through the tube. For the synchronizing pulse to be effective, it must be applied during the charging cycle of the capacitor and at a point on the charging curve that is just below the normal ionization level of the tube when the oscillator is free-running. This condition is shown by the dashed sawtooth wave in (C) of Figure 8-2. In other words, for proper synchronization, the free-running frequency of the oscillator should be somewhat lower than the frequency of the synchronizing pulses. Therefore, normally the speed control of the oscillator would be set so that its free-running speed is somewhat less than 60 cps for the vertical oscillator and a little less than 15,750 cps for the horizontal oscillator.

5. High vacuum type tubes as sweep generators.

As mentioned earlier, the gas tube oscillator is seldom used in modern TV receivers because its stability is not sufficiently good for the great accuracy needed by television sweep circuits. Instead, the high vacuum type tube is usually used to perform the operation of charging and discharging the capacitor. As just described, a gas-filled tube utilizes a very simple circuit, inasmuch as it is only necessary to once start the ionization of the gas for it to remain ionized until the capacitor is discharged to the 15 volt level. Contrasted to this, the circuit required when a high vacuum tube is employed is considerably more complex. However, the increased stability provided through the use of a high vacuum tube makes it more desirable for use as a sweep generator in TV receivers.

6. The multivibrator sawtooth generator.

One common type of oscillator using high-vacuum tubes which may be used to generate a sawtooth waveshape, is the multivibrator or relaxation oscillator. There are many variations of this circuit which fundamentally is a two-stage resistance-coupled amplifier having common coupling or feedback between the two stages. Such an arrangement will oscillate because the 180° phase shift between the grid and plate of a vacuum tube causes the output of the second tube to supply to the first tube an input voltage that has the right phase to sustain oscillations. By placing a suitable charging capacitor from plate to ground at the output of the multivibrator, a sawtooth voltage waveshape may be derived. The usefulness of the multivibrator arises from the fact that the output waveshape may be made quite linear and also that the frequency of oscillations is readily controlled by an injection of a voltage pulse. The multivibrators used in TV receivers are free-running and do not require the injection of a pulse to produce continuous sawtooth waves. Pulses are used, however, to accurately control their frequency from an external source.

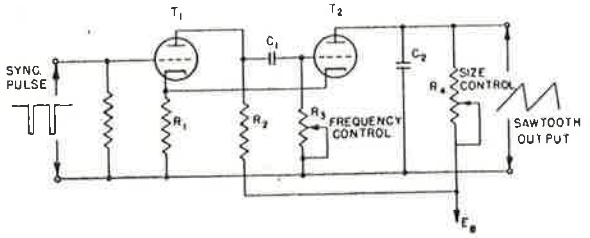


Fig. 8-3. Cathode coupled multivibrator.

7. Cathode coupled multivibrator.

A multivibrator which makes use of a common cathode resistor R_1 to provide feedback between the two tube sections is shown in Figure 8-3. Fundamentally, the two tubes are operated 180 degrees out of phase with each other; that is, tube T_1 conducts when T_2 is cut-off and vice versa. The circuit constants are chosen so that tube section T_2 is cut-off for a much longer period of time than T_1.

The tube section T_2 is operated as an electronic switch to perform the same function that S_1 and S_2 performed in Figure 8-1. When the tube T_2 is cut-off, it performs the operation of S_1 closed and S_2 opened in Figure 8-1, and when it is conducting, it operates similar to the switches in the opposite position; that is, S_1 open and S_2 closed. Therefore, the charging capacitor, C_2, across which the sawtooth voltage waveform is generated is connected in the plate circuit of T_2. With reference to plate current waveform (E) and the sawtooth voltage waveform of (F) in Figure 8-4, it will be noted that during the time that T_2 is not conducting, the charging capacitor, C_2, will gradually charge up through R_4 to form the trace portion of the sawtooth voltage wave. When this charge on C_2 has reached a certain value, T_2 will become conducting and rapidly discharge C_2 through the relatively low internal plate resistance of the tube, thus forming the retrace portion of the sawtooth waveshape. This results in the sawtooth waveform as shown in (F) of Figure 8-4. Because of the short time constant of the discharge cycle, it accounts for the very short conduction cycle necessary of T_2. The two-tube sections shown in Figure 8-3 are usually the triode sections of a dual triode tube, such as the Type 6SN7GT or 12AU7 tube. In order to understand the function of T_1 in conjunction with T_2, the following more detailed explanation of operation is given.

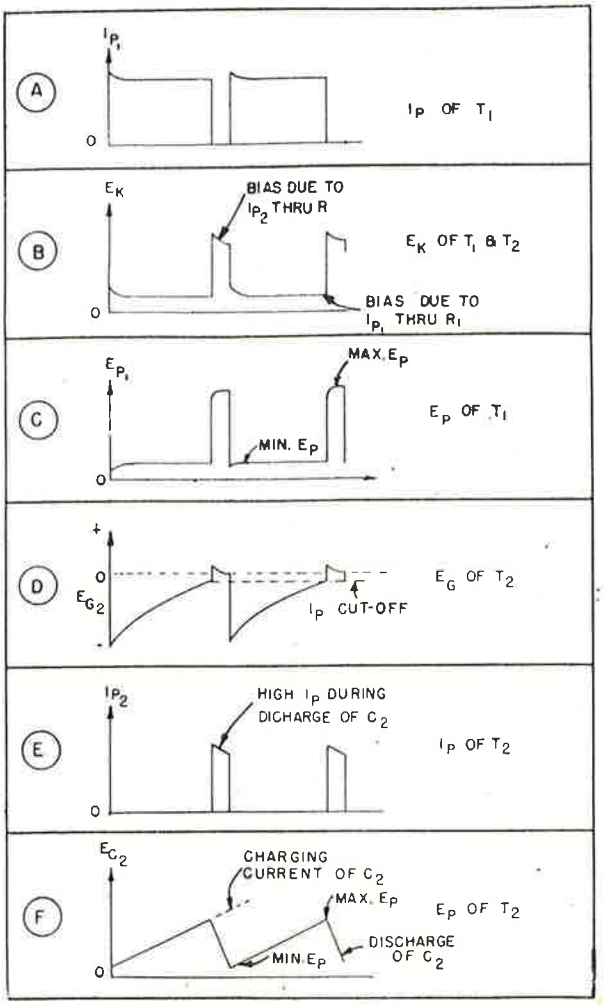


Fig. 8-4. Waveshapes in cathode coupled multivibrator.

To start the cycle of operation, it will be assumed that T_1 is conducting and T_2 is cut-off. During the time that T_2 is cut-off, the charging capacitor C_2 gradually charges up through R_4 to form the trace portion of the sawtooth wave. As the charge on C_2 gradually rises, the plate voltage of T_2 and the voltage E_c rises accordingly, as indicated by (F) of Figure 8-4.

At the same time that the plate voltage of T_2 is gradually rising, the grid bias on T_2 is gradually decreasing, as indicated in (D). In a short time the combination of the lower grid bias and the increased voltage on the plate of T_2 will start conduction of T_2. This is the point where the grid bias just reaches the dashed line marked I_p cut-off level. Grid voltages above this level will cause T_2 to conduct, while voltages below this level will cause plate current cut-off.

At the moment that T_2 starts to conduct, the plate current of T_1 begins to decrease, due to the increase in bias voltage developed across the common cathode resistor R_1 by the plate current of T_2 flowing through it. The decrease in plate current of T_1 causes the plate side of R_2 to become more positive and this positive potential is applied to the grid of T_2 through the coupling capacitor C_1 which further increases the plate current of T_2. This action is cumulative and almost instantaneous and causes a high value of plate current to flow which rapidly discharges C_2 through T_2, producing the retrace portion of the sawtooth wave, as indicated in (F). The high value of plate current through T_2 during retrace is indicated by (E) of Figure 8-4. This high value of T_2 plate current flowing through the common cathode resistor R_1 actually biases T_1 to cut-off during this time, as shown by (A) and (B). Since the positive voltage coupled to the grid of T_2 is much greater than the bias voltage developed across the cathode resistor R1 (due to the amplification provided by T_1), it more than overcomes the effect of this bias voltage as far as T_2 is concerned.

During the time that T_2 is conducting heavily, the bias voltage developed across R_1 will keep T_1 cut-off until the retrace is completed. Since T_1 is cut-off, its plate goes highly positive, causing the grid of T_2 to draw current, which places a high negative charge on the coupling capacitor C_1. This charge accumulates on C_1 during the time that the grid voltage of T_1 is actually positive, as indicated by the portion of the grid voltage above the zero bias line of (D). The grid capacitor, C_1, charges through R_1, the grid cathode circuit of T_2 and the plate resistor R_2. Electrons flow from cathode to grid and accumulate on the grid side of C_1, thus placing a negative charge on the grid side of C_1.

As soon as retrace is completed (capacitor C_2 discharged) the plate current of T_2 rapidly drops, (E) of Figure 8-4, which removes the high bias on T_1 (B), and allows T_1 to again draw plate current (A), thus removing the positive potential applied to the grid of T_2 (C). The grid of T_2 will now be biased beyond cut-off by the negative charge which was accumulated on the grid side of C_1 during the period of grid current flow. This is indicated in (D) of Figure 8-4. The cycle is thus completed and another cycle started with C_2 again gradually charging up through R_4 while T_2 is cut-off.

8. Frequency control.

The negative charge built up on C_1 during the retrace period must discharge or leak off before T_2 can again start to conduct. The discharge path is through the grid resistor R_3. It is therefore apparent that the time required for the negative charge on C_1 to leak through R_3 to the point where T_2 will again conduct will determine the frequency of oscillation or the number of sawtooth waveforms produced per second. The shorter the time required (the lower the RC time constant) the more frequently will T_2 conduct and the higher will be the frequency of oscillations. The longer the time required (the higher the product of R_3 and C_1) the less frequently will T_2 conduct and the lower will be the frequency of oscillation. By making the resistor variable, a convenient method of adjusting the free-running frequency of this multivibrator is had.

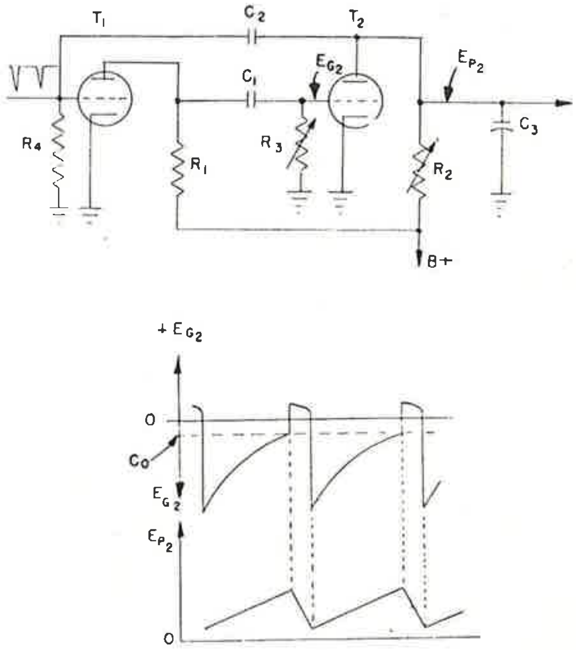


Fig. 8-5. Plate coupled multivibrator.

9. Plate coupled multivibrator.

Another version of the multivibrator that is quite popular is the circuit shown in Figure 8-5. The circuit is fundamentally the same as the one just described except that feedback occurs by means of plate coupling capacitors instead of a common cathode resistor as before.

Basically, the action which takes place is as follows:

Feedback action which occurs through the common coupling capacitors C_1 and C_2 allows T_1 to conduct while T_2 is cut-off and then allows T_2 to conduct while T_1 is cut-off. T_1 conducts most of the time, while T_2 conducts only during the short retrace period. Due to grid current flow when T_2 is driven into conduction, a large negative charge accumulates on the grid side of C_1 and must leak off through R_3 before T_2 can again conduct.

During the time that T_2 is cut-off, the charging capacitor C_3 gradually charges up through R_2 to form the trace portion of the sawtooth wave. By the time that the charge on C_3 reaches approximately 10% of its maximum value, the negative voltage on C_1 will have leaked off sufficiently to allow T_2 to conduct which rapidly discharges C_3 through T_2.

This action is repeated over and over again, at a frequency determined by the R-C time constant in the grid circuit of T_2. As in the case of the cathode coupled multivibrator, increasing this time constant decreases the frequency of oscillation and vice-versa.

10. Size (height or width) control.

By varying part of the resistance in series with the charging capacitor across which the sawtooth wave appears, the amount of charge accumulated on it during the trace period can be varied and hence the amplitude of the sawtooth wave can be varied. Increasing this resistance causes fewer electrons to flow into the charging capacitor in a given time (trace period) lowering the amount of charge and therefore reduces the size of the sawtooth wave. Decreasing this resistance of course has the opposite effect, increasing the size.

If the series resistance is held constant and the value of the charging capacitor is changed, it will also vary the amplitude of the sawtooth wave. Increasing the value of the charging capacitor reduces the size since more electrons are required to raise the level of charge to a given value. Likewise, decreasing the value of the charging capacitor increases the size of the sawtooth wave since the level of charge will rise at a faster rate. Although the amplitude or size of the sawtooth wave can be changed by varying either R or C, it is usually controlled by making part of the charging resistance variable over a limited range. This is called the Height or Width control, as the case may be, since it results in a change in the height or width of the picture.

11. Synchronization of the multivibrator.

The frequency of the multivibrator circuits shown in Figures 8-3 and 8-5 can be "locked in" or synchronized with an external signal by injecting synchronizing pulses at either the 1st or 2nd tube grid. If they are injected at the grid of T_1, they must be negative going or if they are injected on the grid of T2 they must be positive. Usually, a negative pulse is injected at the grid of T_1. Injecting a negative pulse on the grid of T_1 as shown in Figure 8-3 or 8-5 will cause a large positive pulse to appear in the plate circuit of T_1 and this, in turn, coupled to the grid of T_2 causese it to begin drawing plate current. This will cause the discharge cycle to occur somewhat before it normally would if no synchronizing pulse were applied to the grid of T_1.

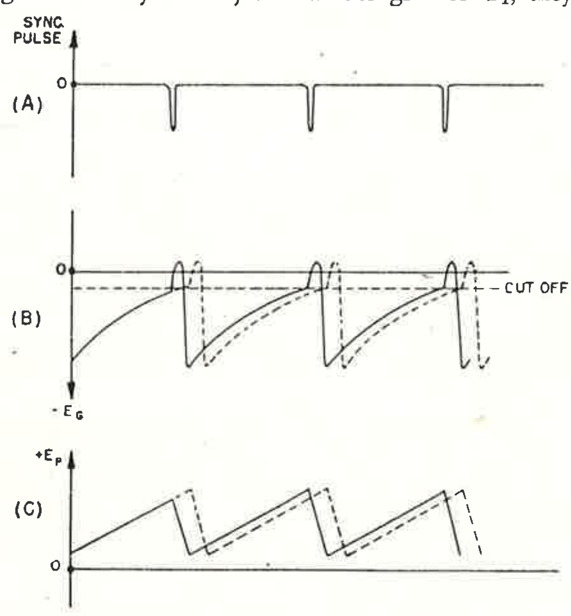


Fig. 8-6. Synchronization of multivibrator.

For proper synchronization, the negative pulse applied to the grid of T_1 should occur just before the grid voltage on T_2 reaches the conduction point so that the corresponding positive pulse appearing in the plate circuit of T_1 will drive T_2 into conduction slightly earlier than it would otherwise conduct. This condition is illustrated in Figure 8-6 — the solid line in (B) shows how the grid voltage on T_2 is made to go positive slightly ahead of time. The relationship between the sync pulse and the initiation of the retrace portion of the sawtooth wave is shown by (A) and (C) Figure 8-6. The dashed line represents several cycles of the sawtooth wave without the application of sync pulses (free-running condition) and the solid line represents the sawtooth wave when sync pulses are applied. It should be noted that the sync pulses cause retrace to occur slightly before its normal free-running time and, therefore, the synchronized speed or frequency will be slightly higher than the free-running frequency. As stated earlier, in connection with the gas tube type of sweep generator, the free-running speed of a sawtooth oscillator or sweep generator should be slightly lower than the frequency of the sync pulses for proper synchronization. Since retrace is started or initiated at exactly the same instant every cycle by the application of sync pulses on the grid of T_1, the frequency of the multivibrator will be locked in or synchronized with the transmitter.

Thus far, only direct synchronization has been considered, where the sync pulses are injected directly in the grid circuit of the sawtooth oscillator. This is the method used for synchronization of the vertical sweep generator (60 cps) in all GE TV receivers and for the horizontal sweep generator (15,750 cps) in the pre-war receivers. However, as mentioned in an earlier chapter, this method is not used for synchronization of the horizontal sweep generator in the post-war GE receivers. Instead, the frequency of the horizontal sweep generator is controlled indirectly by the transmitted sync pulses through the use of an automatic frequency control circuit (AFC). Since this type of circuit is rather complex and there are several versions of it, this method of synchronization will be discussed in detail in a separate chapter.

12. The blocking oscillator.

Another type of sawtooth oscillator, probably the simplest high vacuum circuit using one tube, is the blocking oscillator and is used in a number of GE receivers. The basic circuit together with associated waveforms is shown in Figure 8-7.

The transformer serves as a coupling between the plate and grid circuits and is so connected that an increase in plate current causes the grid to go positive, which further increases the plate current until finally the plate current can no longer increase because of plate current saturation.

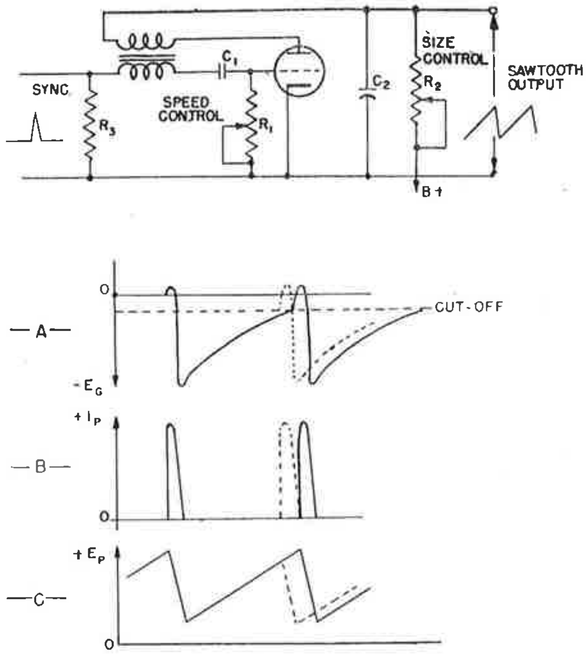


Fig. 8-7. The blocking oscillator

This causes the flux in the plate coil to start to collapse, which reverses the grid polarity, making it negative, reducing the plate current. As the plate current decreases, the grid is driven more and more negative until the plate current is cut-off.

In the description so far, it is apparent that the circuit performs quite similarly to a conventional oscillator having coupling between the plate and grid coils. However, there are several important differences to note.

First of all, the tube remains cut-off for a considerably longer time than it conducts, and during this cut-off period the charging capacitor C_2 is gradually charged up through R_2 to form the trace portion of the sawtooth waveform. The tube draws plate current for a very short period of time, during which C_2 is discharged to form the retrace portion of the sawtooth wave.

Secondly, the duration of the charging period or the interval during which the tube is cut-off is dependent upon the time constant of R_1 C_1 in the grid circuit. The frequency of oscillation is not dependent upon the values of L and C of the transformer, as in the case of the conventional oscillator, but rather upon the values of the grid leak resistor R_1 and capacitor C_1. However, the natural resonant frequency of the circuit as determined by the L and C values (transformer inductance and distributed capacity) bears an important relationship to the sawtooth output and must be considered in the design of the circuit.

In the first part of the cycle just outlined, the grid is driven positive causing grid current to flow. This places a negative charge on the grid side of C_1, the capacitor charging up through the cathode-grid circuit of the tube, the transformer secondary and R_3. The negative voltage thus developed biases the tube well beyond cut-off, as indicated in (A) of Figure 8-7, and must leak off through R_1 before the tube can again conduct. As mentioned previously, the charging capacitor C_2, across which the sawtooth waveform is developed, charges up through R_2 during the time that the tube is cut-off.

When the charge on the grid capacitor C_1 has leaked off sufficiently to permit the tube to conduct, a pulse of plate current occurs during the time that C_2 is rapidly discharged through the tube.

This pulse of plate current occurs for a very short period of time, less than $\frac{1}{2}$ cycle of the natural resonant frequency of the circuit as determined by the L and C values of the circuit.

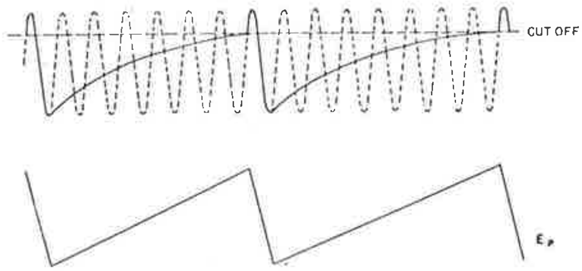


Fig. 8-8. Waveshapes in the blocking oscillator.

The blocking oscillator would oscillate at a much higher frequency (its natural resonant frequency) than that of the sawtooth wave output were it not for the fact that by the time $\frac{1}{2}$ cycle is completed, the plate voltage is very low due to the discharge of C_2 and the grid voltage quite negative due to the charge on C_1. This stops or blocks any further oscillation until the negative charge on C_1 leaks off to the point where the tube can again conduct. The sine waves in dashed lines on the grid voltage waveform shown in Figure 8-8 indicate how the grid voltage would vary at the natural resonant frequency of the circuit if it were not for the blocking action in the grid circuit due to the charge accumulated on C_1 during retrace.

It was mentioned earlier that the natural resonant frequency of the blocking oscillator bears an important relationship to the sawtooth waveform produced. This is due to the fact that the time required to discharge C_2 is equivalent to approximately one-half cycle of the natural resonant frequency of the circuit. Therefore, the retrace time is controlled by the L and C values of the circuit. If the natural resonant frequency is too low, then the retrace period will be too long and will not be completely blanked out by the blanking signal.

For example, suppose that the blocking oscillator is generating the sawtooth waveform for horizontal deflection. The frequency of the sawtooth output will be 15,750 cps when synchronized. The visible portion of the horizontal trace consumes approximately 53 miscroseconds, while the horizontal blanking period is approximately 10 microseconds in duration.

During the 10 microsecond blanking period the beam of the picture tube must be returned to the left edge of the screen and start moving toward the right edge at a linear rate. To accomplish this within the horizontal blanking time, only approximately 7 microseconds can be allowed for the actual retrace portion of the sawtooth wave.

Since approximately ½ cycle of the natural resonant frequency of the blocking oscillator must occur before retrace is completed (the discharge of C_2), it is necessary therefore that the natural resonant frequency be of a value such that ½ cycle occurs in slightly less than 7 microseconds. In other words, for horizontal retrace to occur within the 7 microsecond limit, one cycle of the natural resonant frequency must occur in slightly less than 14 microseconds. This would correspond to a frequency of approximately 82 kilocycles. Thus the inductance of the transformer and distributed capacity across it must resonate to at least 82 kc for proper retrace to occur (in the case of the horizontal sweep generator).

As in the previous sweep generators discussed, the size or amplitude of the sawtooth wave can be varied by making a portion of the series charging resistor variable. This is R_2 in Figure 8-7.

Also, as in the case of the multivibrator circuit, the free-running frequency or speed can be controlled by making part of the resistance variable in the grid circuit of the tube, which discharges the charging capacitor to produce the retrace. This is indicated by R_1 in Figure 8-7.

13. Synchronization of the blocking oscillator.

Synchronization may be accomplished by injecting a positive synchronizing pulse in the grid circuit, as shown in Figure 8-7.

If the pulse is of sufficient amplitude, it will drive the tube into conduction sooner than it would otherwise go into conduction by the leakage of the negative charge on C_1 through R_1. Here again, the free-running frequency of the oscillator should be somewhat lower than the frequency of the synchronizing pulses as indicated by the waveforms in Fig. 8-8.

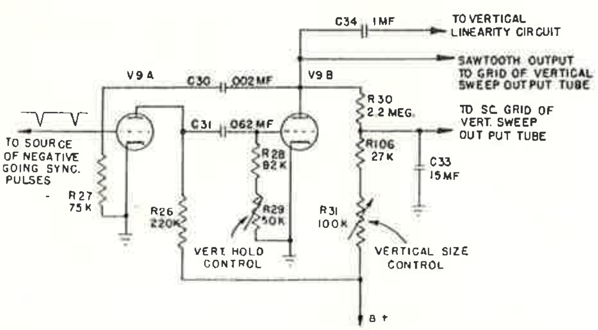


Fig. 8-9. Multivibrator as vertical sweep generator.

14. Typical sweep generator circuits.

Both the multivibrator and the blocking oscillator circuits can be used to produce sawtooth waveforms for horizontal sweep (15,750 cps) or for vertical sweep (60 cps).

A multivibrator used as a vertical sweep generator is shown with typical values in Figure 8-9. This is the circuit used in the GE Model 810 receiver and is of the plate coupled type.

This circuit is essentially the same as the plate coupled multivibrator discussed earlier except for several slight differences. One of these is that the charging capacitor, C_{34}, across which the sawtooth wave is developed, is not returned directly to ground but, instead, returns to ground through a linearity control. Also, the size control varies the voltage on the screen of the sweep output tube, as well as the charge on C_{34}. The screen of the sweep output tube is by-passed by C_{33}. The control of linearity and sweep output circuits will be discussed in the next chapter.

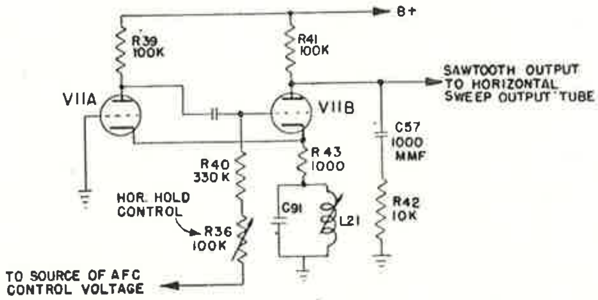


Fig. 8-10. Multivibrator as horizontal sweep generator.

Figure 8-10 illustrates a multivibrator circuit with typical values used as a horizontal sweep generator. This is the circuit used in the GE Model 801 receiver and is of the cathode coupled type. As far as the generation of the sawtooth waveform is concerned, this circuit is fundamentally the same as the cathode coupled circuit discussed previously. However, there are a number of differences. Direct synchronization is not used with this circuit and for that reason the horizontal frequency or hold control connects to a source of DC control voltage which provides automatic frequency control of the horizontal sweep. AFC control of the horizontal sweep will be covered in a later chapter. Resistor R_{43} provides common coupling. The tuned circuit consisting of L_{21} and C_{91} in the cathode circuit is tuned to 15,750 cps and is used to provide better stability in conjunction with the AFC circuit used. The charging capacitor, C_{57}, returns to ground through a 10K ohm resistor R_{42}, for reasons to be discussed in the next chapter. It will be noted that the series charging resistor R_{41} is not variable, since the horizontal size control (for reasons that will also be discussed in the next chapter) is in the sweep output circuit.

Figure 8-11 illustrates a blocking oscillator with typical values, used as the vertical sweep generator in the GE Model 901 receiver. Here again there are a number of differences, but the operation of the circuit is fundamentally the same as the blocking oscillator discussed earlier. The charging capacitor

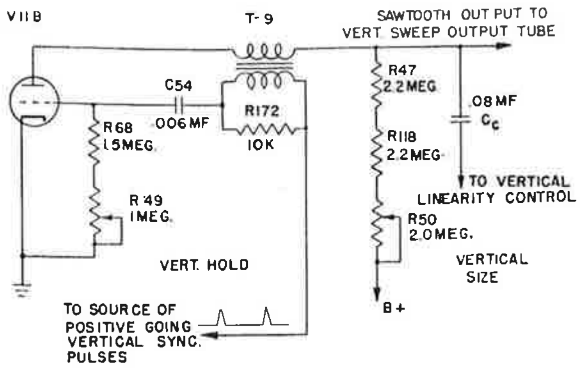


Fig. 8-11. Blocking oscillator as vertical sweep generator.

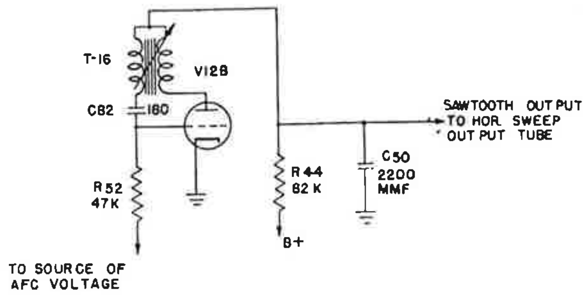


Fig. 8-12. Blocking oscillator as horizontal sweep generator.

C_c returns to ground through the vertical linearity control. The size control simply varies the amount of charge on C_c. Positive-going sync pulses are applied to the grid through the transformer secondary. The free running speed is controlled by R_{49} in the grid circuit. The resistor across the transformer-secondary provides damping to prevent transient oscillations from being set up.

Figure 8-12 illustrates a blocking oscillator used as a horizontal sweep generator and is the circuit used in the GE Model 810 receiver. The charging capacitor is C_{50} and the charging resistor R_{44} is not variable since the size control is in the sweep output circuit. The grid return connects to a source of AFC voltage. The usual speed control is not in the grid circuit in this case, but is a part of the AFC circuit which will be covered in detail in a later chapter. The transformer T_{16} is a continuous winding and is tuned by means of an iron core for optimum operation of the circuit.

New RCA Releases

NEW 27″ RECTANGULAR METAL-SHELL PICTURE TUBE

Radiotron — 27MP4 is a short, directly viewed, rectangular picture tube of the metal-shell type for use in deluxe television receivers. It has a screen size of $23\frac{7}{16}$″ x $18\frac{1}{8}$″ with slightly curved sides and rounded corners.

Capable of providing pictures having high brightness, excellent contrast, and good uniformity of focus over the whole picture area, the 27MP4 has a metal-backed, high-efficiency, white fluorescent screen on a relatively flat, spherical, high-quality faceplate made of frosted Filterglass. The frosting effectively suppresses reflections of bright objects regardless of their location with respect to the faceplate.

Employing wide-angle deflection (diagonal deflection angle of 90°) and magnetic focus, the 27MP4 is designed with a funnel-to-neck section having a shape which has been coordinated with deflecting-yoke design to provide high deflection sensitivity and low deflection distortion. Because the metal-shell construction permits the use of thinner glass in the funnel and accurate shaping of the funnel-to-neck section, maximum beam clearance is provided. As a result of this feature, the 27MP4 has short overall length — a length even shorter than that of the 21AP4. Furthermore, the picture can be positioned on the screen with less critical adjustment of the components on the neck.

Other features of the 27MP4 include its metal-shell construction which not only weighs about 30 percent less than a comparable all-glass tube but also possesses inherent strength to minimize the hazard of implosion; a higher-quality faceplate than is commonly used in all-glass tubes; a neck length of $7\frac{1}{2}$ inches; and an ion-trap gun.

MEDIUM-MU TRIODE — SHARP-CUTOFF PENTODE

Radiotron — 6U8 is a multi-unit tube of the 9-pin miniature type containing a medium-mu triode and a sharp-cutoff pentode in one envelope. It is intended for use as combined oscillator and mixer tube in vhf television and AM/FM receivers.