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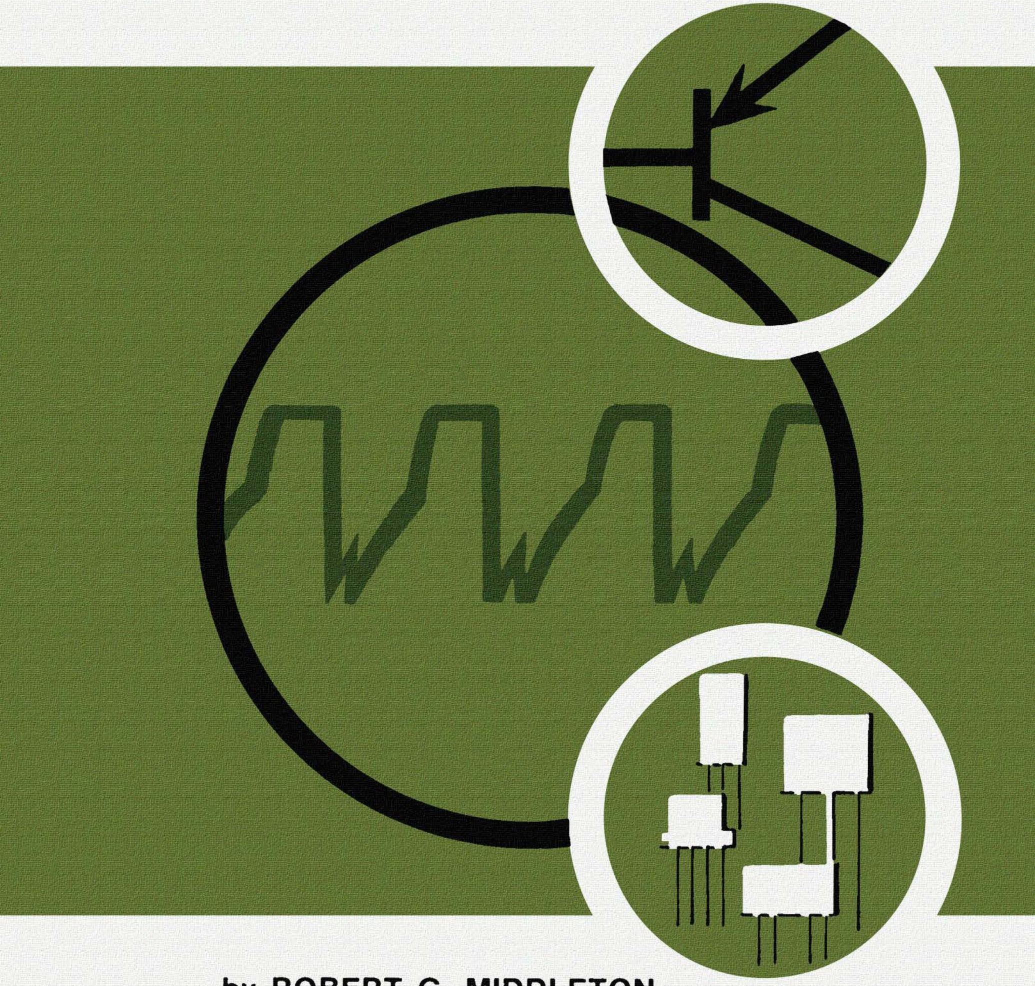
Using Scopes in
Transistor Circuits



by ROBERT G. MIDDLETON

Using Scopes in Transistor Circuits

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Preface

Scope application in transistor circuits is similar in various respects to application in vacuum-tube circuits. However, there are many unexpected situations encountered in transistor circuits because a transistor is a solid-state current-operated device, whereas a tube is an evacuated, voltage-operated device. A tube conducts current only by means of electrons, whereas a transistor conducts current by means of two charge carriers called electrons and holes. The electrons in a tube flow from cathode to plate because of electric field forces. In a transistor, electrons and holes flow partially because of electric fields (drift fields), and partially because of the diffusion effect. The diffusion effect can be compared with the expansion of hydrogen gas injected at the neck of a balloon. Because of mutual repulsion of the hydrogen molecules, the injected gas diffuses throughout the balloon.

In the base of a transistor, electrons or holes cannot be removed by a drift field. Charge carriers stored in the base region can be removed only by diffusion. Therefore, because of storage time, we encounter waveform distortions that are not found in vacuum-tube circuits. Furthermore, because a transistor is a current-operated device, we will discover that the input-current waveform to a transistor is distorted by an unexpected polarity reversal at the termination of the input drive pulse. In turn, waveform analysis in fast-acting transistor circuits is quite different from that in corresponding tube circuits. We will find that even semiconductor diodes produce unexpected output waveforms in high-speed switching circuits.

If we intend to use scopes intelligently in transistor circuits, we need to have a good understanding of transistor circuit action. Therefore, transistor circuit action is explained in this book wherever it has a direct bearing on waveform analysis. A wide range of technology has been covered, with emphasis on those configurations that are of key importance in the operation of most transistor equipment. Although the book has been written chiefly from the standpoint of the electronic technician, the depth of treatment is adequate to meet the needs of students in junior colleges and technical institutes. Mathematics has been held to a practical minimum, but some equations have been included at points where they contribute substantially to understanding of circuit action.

Many diagrams and waveforms are presented to show clearly the developments and conclusions of the text discussion. Review questions have been added at the end of each chapter so that the reader may check his progress. It is assumed that the reader is familiar with operation of service-type scopes, and that he has a basic knowledge of the principles involved in triggered-sweep scope operation. If the reader lacks this knowledge, he is advised to refer to basic scope-operation books. Whenever possible, it is advisable to check out the scope tests in practice. Facility in scope operation and waveform analysis can be obtained only by a combination of study and practical experience.

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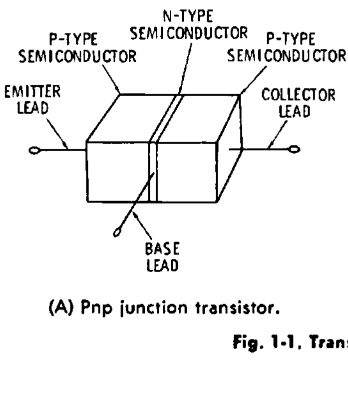
1

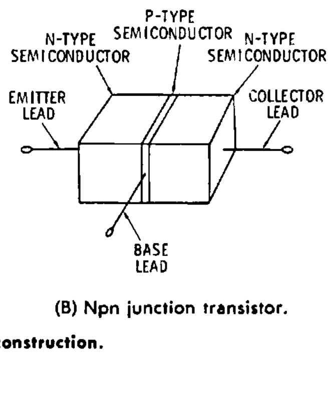
Semiconductor Principles and Waveform Analysis

Semiconductor devices are used for amplification, oscillation, modulation, demodulation, and other applications such as waveshaping. Fig. 1-1 shows the physical construction of pnp and npn junction transistors. Fig. 1-2 shows the general types of waveforms associated with amplification, oscillation, amplitude modulation, amplitude demodulation, frequency modulation, and frequency demodulation. A transistor may be used in a current, voltage, or power amplifier configuration. A signal of 1mA fed into the input of a transistor may appear at a 20-mA level in the output circuit.

As an oscillator, a transistor converts dc voltage into ac voltage. In suitable circuit arrangements, a transistor can provide amplitude modulation (variation in amplitude of an rf signal), or frequency modulation (variation in frequency of an rf signal). Demodulation of a-m and fm signals can be accomplished

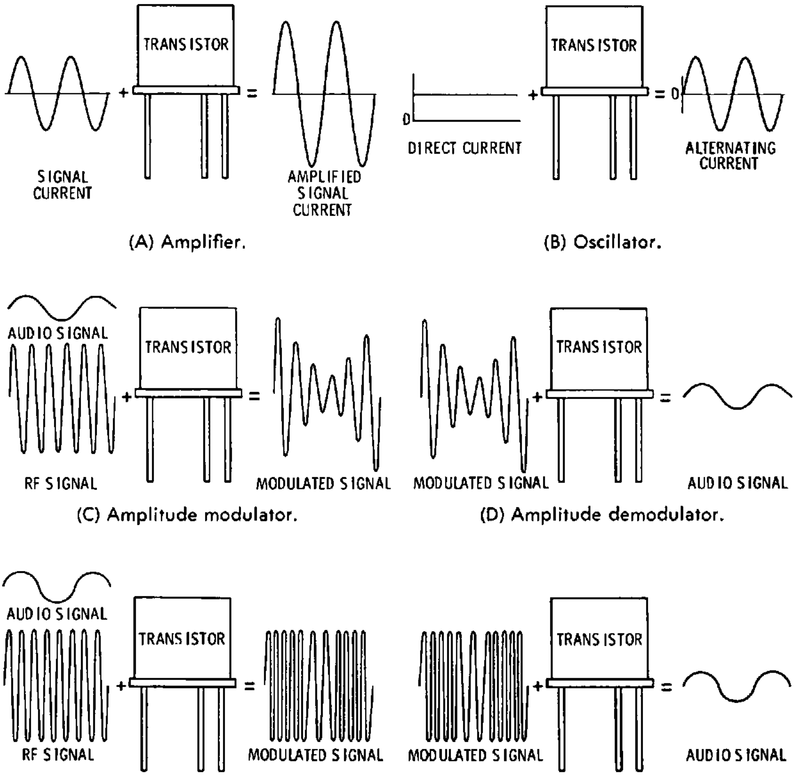


(A) Pnp junction transistor.



(B) Npn junction transistor.

Fig. 1-1. Transistor construction.



(A) Amplifier. (B) Oscillator.

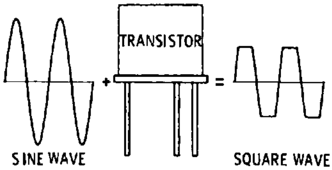
(C) Amplitude modulator. (D) Amplitude demodulator.

(E) Frequency modulator. (F) Frequency demodulator.

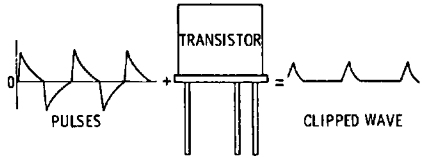
Fig. 1-2. Transistor applications.

by transistors in associated circuit configurations. A transistor may also be used to shape one waveform into another waveform. Fig. 1-3 illustrates the application of a transistor as a waveshaper to change a sine wave into a square wave. The operation of a transistor as a clipper is also shown. Basically, a transistor is an electronic valve that permits collector supply current in step with an input waveform. With a suitably chosen bias, collector current will be permitted only over a certain portion of the input signal cycle.

A transistor has two junctions, which can be operated as rectifiers. Whether a transistor amplifies, modulates, demodulates, or clips, depends on the emitter-junction bias. Fig. 1-4 shows the voltage-current characteristic of a pn junction. Diodes cannot amplify, unless they are of the tunnel-diode construction. This function of a diode is discussed subsequently.



(A) Squaring action.



(B) Clipping action.

Fig. 1-3. Transistor used to modify waveforms.

However, any diode operates as a modulator or demodulator if it is biased to a nonlinear interval of its voltage-current characteristic. The advantage of a transistor in modulation and demodulation action is the amplification, and consequently the stronger output signal, that is provided by the collector circuit. Diodes cannot oscillate, unless they are of the tunnel-diode construction. Both diodes and transistors can give squaring and clipping action, although a transistor provides a stronger output signal.

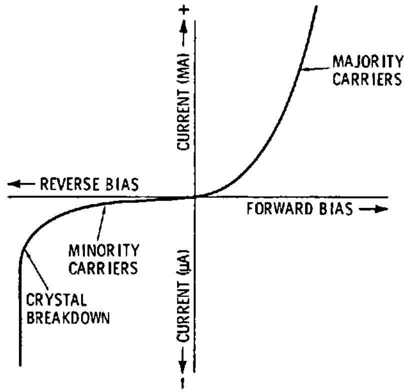


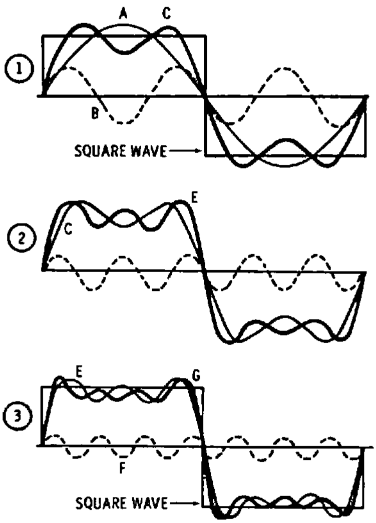
Fig. 1-4. Current-voltage curve for pn junction.

BASIC WAVEFORM ANALYSIS

There are certain principles of waveform analysis that we should understand at this point. Signal processing, in theory, could be described with reference to any fundamental waveform that we might choose. For example, we could choose a square wave as our fundamental waveform; however, this would be a poor choice, because it becomes very complicated to build up a sine wave, or pulse, or sawtooth wave, from a mixture of square waves. In practice we find only two waveforms that are suitable for use as fundamental waveforms. These are the sine wave and the exponential wave. The utility of these two fundamental waveforms stems from mathematical principles that we can neglect at this point. Instead, we will simply state that all waveforms can be regarded as built up from combinations of sine waves and/or exponential waves.

Let us consider the build-up of a square wave from sine waves, as shown in Fig. 1-5. In theory, an infinite number of harmonics would have to be combined with the fundamental sine wave to produce a perfect square wave. In practice, however, we find that about 20 harmonics suffice to give a reasonable facsimile of a square wave. This idea of square-wave composition is very useful because it can be used to explain how a square wave is modified when it passes through various kinds of circuits. For example, if a good square wave is applied to an amplifier that has limited bandwidth, we perceive that the higher-frequency harmonics cannot get through to the output of the amplifier. In turn, the rise time of the square wave is slowed down, in accordance with the highest harmonic that is passed by the amplifier.

This is such an important and basic consideration that it is advisable to explain some of the details that are involved. It is evident that wave C rises faster than wave A in Fig. 1-5. Similarly, wave E rises faster than wave C, and wave G rises faster



A: FUNDAMENTAL
B: 3D HARMONIC
C: FUNDAMENTAL PLUS 3D HARMONIC
D: 5TH HARMONIC
E: FUNDAMENTAL PLUS 3D AND 5TH HARMONICS
F: 7TH HARMONIC
G: FUNDAMENTAL PLUS 3D, 5TH, AND 7TH HARMONICS

Fig. 1-5. Synthesis of square wave from sine waves.

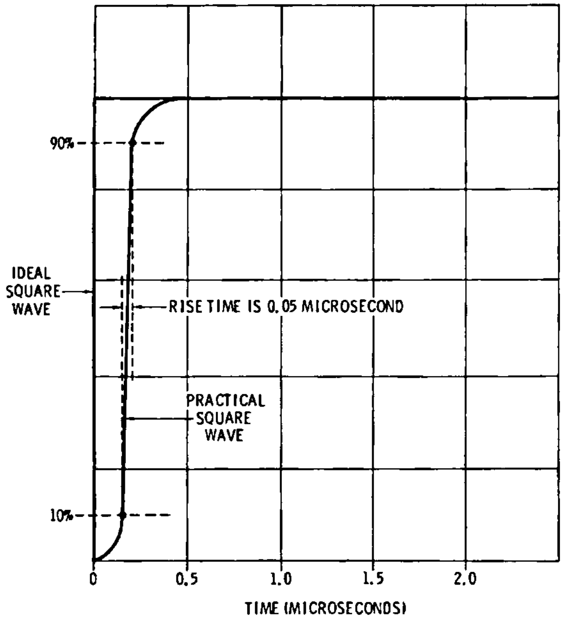


Fig. 1-6. Rise time (T_r) of square wave.

than wave E. In practice, we need to know how the rise time of an output square wave is related to the high-frequency cut-off point of an amplifier. This is a simple formula that is written:

$$T_r = \frac{1}{3f_c} \tag{1.1}$$

where,

T_r is the rise time of the output square wave,
f_c is the frequency at which the amplifier response is down 3 dB.

Fig. 1-6 illustrates the meaning of rise time. Rise time is measured between the 10-percent and 90-percent points on the leading edge of a square wave. Accurate measurement requires the use of a scope with calibrated and triggered sweeps. The waveform is greatly expanded on the triggered-sweep function so that its leading edge occupies a substantial portion of the horizontal interval. In turn, the 10-percent and 90-percent points are noted, and the rise time is read from the settings of the calibrated sweeps. The −3-dB point of an amplifier response curve occurs at the point where the output voltage falls to 0.707 of its maximum value. This is shown in Fig. 1-7. Since the out-

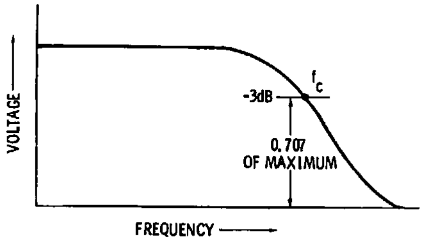


Fig. 1-7. High-frequency cutoff point f_c.

put power is reduced one half at the −3-dB point, the high-frequency cutoff point is also called the half-power point.

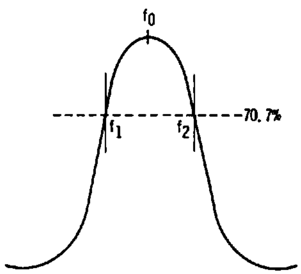
In spite of the utility of Fig. 1-5 in giving a general description of what causes reduced rise time, we must be on our guard to avoid absurd conclusions. In other words, the relations in Fig. 1-5 are not completely descriptive of amplifier action. For example, we might suppose from inspection of Fig. 1-5 that if a square wave is passed through an amplifier that has a narrow bandwidth, and that all harmonics above the 7th harmonic are removed, the reproduced waveform would have a "wavy" top. However, this is not so—the waveform will have a flat top. This example of square-wave analysis has given us an unexpected test result. Let us analyze the situation to see why our conclusion was incorrect.

When a square wave is applied to an amplifier, we do not actually apply a combination of sine waves. That is, we have merely stated thus far that a square wave could be built up or synthesized from a large number of sine waves. The fact of the matter is that a square wave is generated by switching a dc voltage on and off. There is more than one way of looking at a switched dc voltage. We can state that the waveform could be built up from a large number of sine waves. This is quite a different situation from what the generator actually does—it merely switches a dc voltage on and off. Therefore, we must ask what the amplifier response will be to a dc voltage that is suddenly applied and then as suddenly removed. This gets us into the relation between square waves and exponential waveforms; details must be reserved for subsequent discussion.

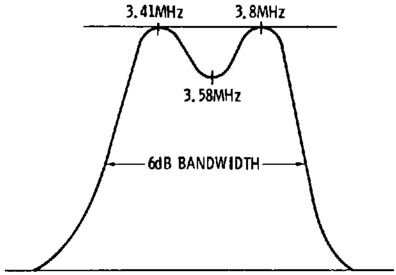
Fig. 1-7 shows the relation between a voltage value and a dB value. A tabulation of these relations is given in Table 1-1. Amplifier bandwidth is measured in two different ways. In the case of a radio receiver, the bandwidth is defined as the number of hertz between the −3-dB points on the frequency-response curve. On the other hand, in the case of a TV receiver, the bandwidth is defined as the number of hertz between the −6-dB or 50 percent-of-maximum voltage points. Examples

Table 1-1. DB expressed as power and voltage (or current) ratios

Power Ratio	Voltage Ratio	dB − ← + →	Voltage Ratio	Power Ratio
1.000	1.0000	0	1.000	1.000
.9772	.9886	.1	1.012	1.023
.9550	.9772	.2	1.023	1.047
.9333	.9661	.3	1.035	1.072
.9120	.9550	.4	1.047	1.096
.8913	.9441	.5	1.059	1.122
.8710	.9333	.6	1.072	1.148
.8511	.9226	.7	1.084	1.175
.8318	.9120	.8	1.096	1.202
.8128	.9016	.9	1.109	1.230
.7943	.8913	1.0	1.122	1.259
.6310	.7943	2.0	1.259	1.585
.5012	.7079	3.0	1.413	1.995
.3981	.6310	4.0	1.585	2.512
.3162	.5623	5.0	1.778	3.162
.2512	.5012	6.0	1.995	3.981
.1995	.4467	7.0	2.239	5.012
.1585	.3981	8.0	2.512	6.310
.1259	.3548	9.0	2.818	7.943
.10000	.3162	10.0	3.162	10.00
.07943	.2818	11.0	3.548	12.59
.06310	.2512	12.0	3.981	15.85
.05012	.2293	13.0	4.467	19.95
.03981	.1995	14.0	5.012	25.12
.03162	.1778	15.0	5.623	31.62
.02512	.1585	16.0	6.310	39.81
.01995	.1413	17.0	7.079	50.12
.01585	.1259	18.0	7.943	63.10
.01259	.1122	19.0	8.913	79.43
.01000	.1000	20.0	10.000	100.00
10^{-3}	3.162×10^{-2}	30.0	3.162×10	10^{3}
10^{-4}	10^{-2}	40.0	10^{2}	10^{4}
10^{-5}	3.162×10^{-3}	50.0	3.162×10^{2}	10^{5}
10^{-6}	10^{-3}	60.0	10^{3}	10^{6}
10^{-7}	3.162×10^{-4}	70.0	3.162×10^{3}	10^{7}
10^{-8}	10^{-4}	80.0	10^{4}	10^{8}
10^{-9}	3.162×10^{-5}	90.0	3.162×10^{4}	10^{9}
10^{-10}	10^{-5}	100.0	10^{5}	10^{10}



(A) In radio-receiver circuit.



(B) In TV-receiver circuit.

Fig. 1-8. Bandwidth concepts.

are shown in Fig. 1-8. In many cases, lab-type scopes have graticules with dB scales. To locate a −3-dB or a −6-dB point on a response curve, we adjust the waveform as depicted in Fig. 1-9. Then, dB points can be read directly from the scope screen.

Next, let us review the basic voltage of a sine wave, as shown in Fig. 1-10. Service-type VOM's read the rms value of a sine wave. The rms value is equal to 0.707 of the peak voltage (either positive or negative peak). In turn, the peak-to-peak voltage of a sine wave is equal to twice the peak voltage. It follows that the peak voltage is equal to 1.414 times the rms voltage, and that the rms voltage is equal to the peak voltage divided by 1.414. Again, the rms voltage of a sine wave is equal to its peak-to-peak voltage divided by 2.83. These are important voltage relations that are not always clearly understood by beginners.

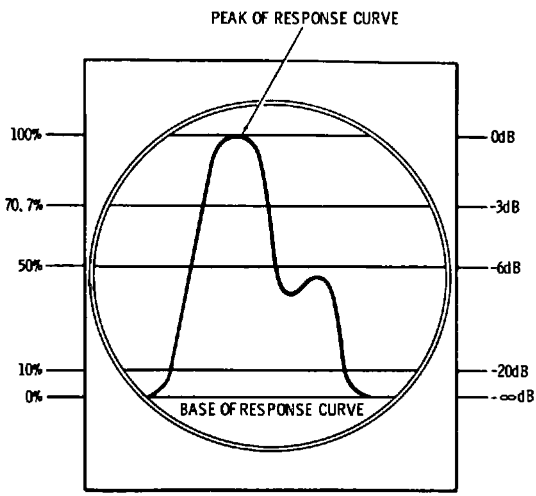


Fig. 1-9. Scope graticule calibrated in dB values.

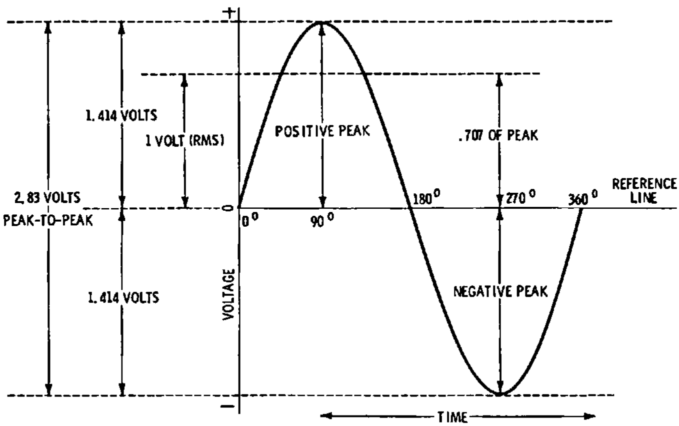
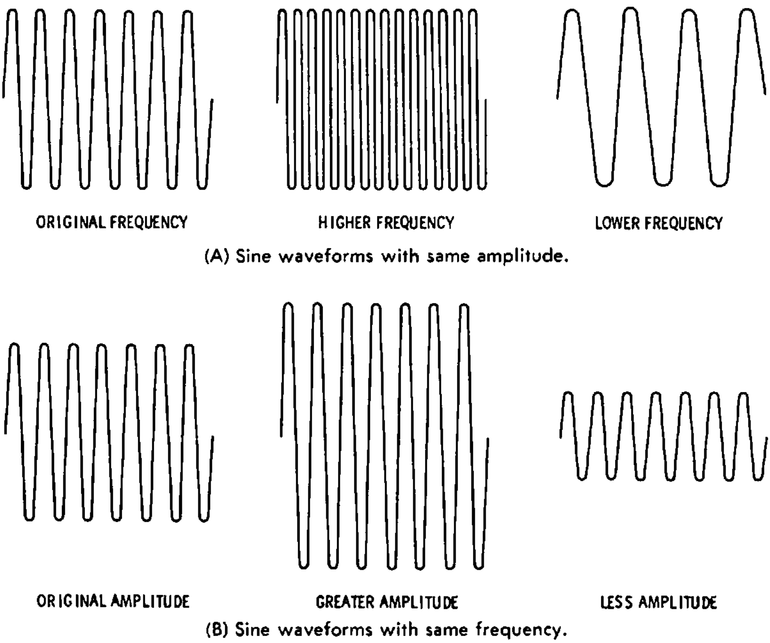


Fig. 1-10. Fundamental sine wave.

It is evident that the voltage of a sine wave is not related to its frequency. They are independent parameters. This fact is illustrated in Fig. 1-11. It must be emphasized that the relations among rms, peak, and peak-to-peak values that have been cited for a sine wave are not true for other waveforms, such as



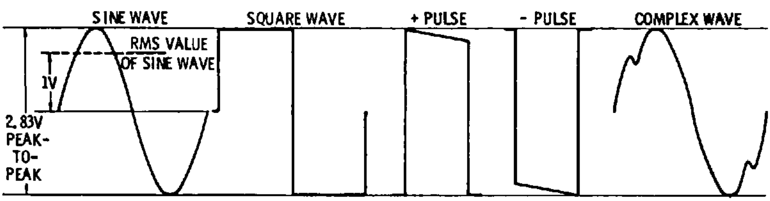
Courtesy Allied Radio Corp.

Fig. 1-11. Distinction between amplitude and frequency.

square waves. Therefore, it is customary to compare various waveform amplitudes only in terms of their peak-to-peak values. The peak-to-peak voltage is of chief concern in analysis of transistor circuit action, and we seldom investigate the rms voltage of a complex waveform. It follows that when we calibrate a scope with a VOM, we must use a sine-wave source, multiply the VOM reading by 2.83 to obtain its peak-to-peak value, and then calibrate the scope in terms of this peak-to-peak voltage. For example, if we start with a 6.3-volt sine-wave source, its peak-to-peak voltage is equal to nearly 18 volts. Therefore, the vertical deflection on the scope screen represents approximately 18 volts pk-pk.

After a scope has been calibrated in peak-to-peak voltage values, the peak-to-peak voltage of any complex waveform can be read directly from the screen. Fig. 1-12 illustrates this principle. Note that peak-to-peak values are equivalent to dc values; that is, we can calibrate a dc scope from a dc voltage source, and employ this calibration for measurement of peak-to-peak voltages. The sine wave depicted in Fig. 1-10 has a positive half-cycle and a negative half-cycle; similarly, the square wave has a positive half-cycle and a negative half-cycle. On the other hand, the "+ pulse" has a positive excursion only. Similarly, the "− pulse" has a negative excursion only. These terms are related to the dc component of a pulse, as explained subsequently. The complex wave has a positive half-cycle and a negative half-cycle.

In summary, an ac waveform (a waveform with both positive and negative excursions) has no polarity indication. For example, the sine wave, square wave, and complex wave depicted in Fig. 1-12 have no polarity indications. On the other hand, a dc waveform has a polarity indication; thus, the two pulse waveforms in Fig. 1-12 are marked positive and negative, respectively. These are called dc pulses because their total excursion is a single polarity. We will find that an ac pulse can be changed into a dc pulse, or vice versa, by suitable variation



Courtesy Allied Radio Corp.

Fig. 1-12. Different waveforms having same peak-to-peak voltage.

of the dc component. Accordingly, let us observe the distinction between an ac voltage, a pulsating dc voltage, and an ac voltage with a dc component, as depicted in Fig. 1-13.

An ac waveform has an average value of zero. This simply means that the area of the positive half-cycle is equal to the area of the negative half-cycle. In other words, if we apply an ac waveform to a dc voltmeter, the pointer does not deflect. However, a pulsating dc waveform has a dc component; this dc component exceeds the peak value of the ac component, and therefore, the waveform does not cross the 0-volt axis. In the example shown in Fig. 1-13, the pulsating dc waveform has a positive excursion only. The average value of a pulsating dc waveform is not zero; if we apply a pulsating dc waveform to a dc voltmeter, the pointer indicates the value of the dc component in the waveform. Finally, an ac waveform with a dc component crosses the 0-volt axis, because the dc component has a value that is less than the peak value of the ac component. An

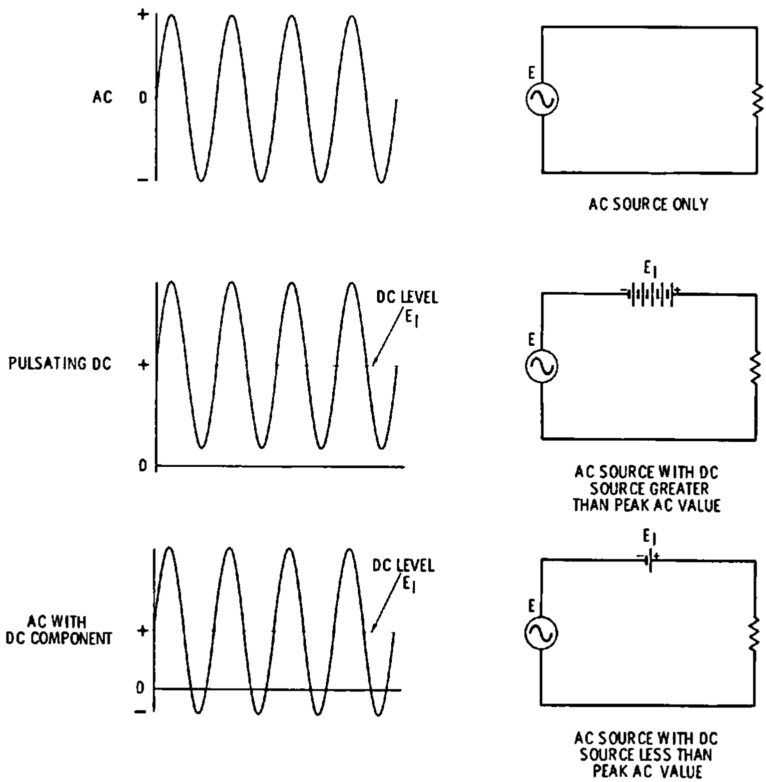


Fig. 1-13. Three basic combinations of ac and dc values.

ac waveform with a dc component has both positive and negative excursions. Its average value is not zero; if we apply an ac waveform with a dc component to a dc voltmeter, the pointer indicates the value of the dc component in the waveform.

With this understanding of the distinctions among ac waveforms, pulsating dc waveforms, and ac waveforms with dc components, let us consider the positive square wave depicted in Fig. 1-14. The bottom of the square wave touches the 0-volt axis, but does not cross the axis. Therefore, this is a pulsating dc waveform. We note that this positive square wave is formed by combining an ac square wave with a dc component that has a value equal to the peak voltage of the ac square wave. Note that when an ac square wave is changed into a dc square wave, the peak voltage of the waveform is doubled. That is, peak voltage is measured from the 0-volt axis. If we add a positive dc voltage to an ac waveform, the positive-peak voltage of the ac waveform is increased by the value of the dc component. Of course, this process decreases the negative peak voltage of the ac waveform by an amount equal to the value of the dc component.

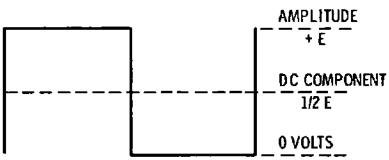


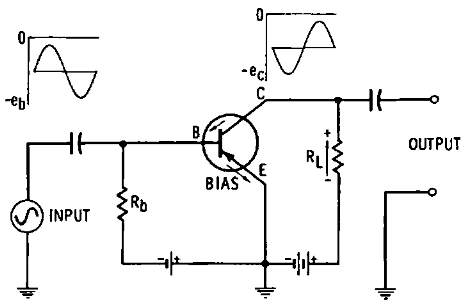
Fig. 1-14. Square wave with positive excursion only.

WAVEFORMS IN TRANSISTOR AMPLIFIER CIRCUITS

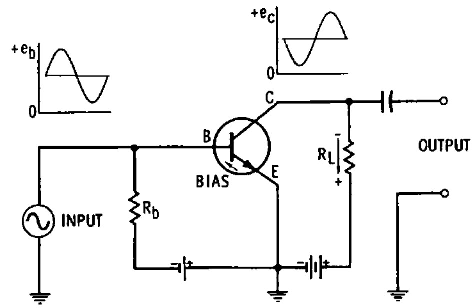
Most waveforms processed by transistor amplifiers are pulsating dc waveforms, as shown in Fig. 1-15. The reason for this occurrence is that dc bias voltages are applied to the base and collector of the transistor. Moreover, a transistor cuts off if it is reverse-biased. With reference to Fig. 1-15A, a negative dc bias is applied to the base of the transistor, and a negative dc voltage is applied to the collector. An ac signal voltage is coupled into the base of the transistor; here, the ac waveform combines with the dc bias voltage to produce a pulsating dc waveform. As seen in the inset, this waveform has a negative excursion only.

Next, in Fig. 1-15A, the pulsating dc signal that flows into the base diffuses into the collector circuit. Since the collector operates at a higher dc voltage than the base, the drop across R_L is much greater than the drop between base and emitter.

In other words, voltage amplification is obtained in the collector circuit. Of course, the collector output waveform is also a pulsating dc waveform; the output waveform has a negative excursion only. Note that the ac component of the output waveform is reversed in phase, compared with the ac component of the input waveform.



(A) Pnp grounded emitter.



(B) Npn grounded emitter.

Fig. 1-15. Pulsating dc waveforms in transistor amplifier circuits.

Next, if we use an npn transistor, as shown in Fig. 1-15B, the polarity of the pulsating dc waveforms is reversed. That is, the base of the transistor is biased positively, and a positive voltage is applied to the collector. An ac signal is coupled into the base of the transistor; here, the ac signal combines with the positive bias voltage to form a pulsating dc voltage. This waveform has a positive excursion only. The pulsating dc waveform diffuses through the base into the collector circuit, where it appears as an amplified pulsating dc waveform. Its ac component is reversed in phase, but the output waveform nevertheless has a positive excursion only.

It is important for us to note that a coupling capacitor removes the dc component of a pulsating dc waveform, as shown in Fig. 1-16. We observe that the input waveform has both an ac component and a dc component. However, the capacitor blocks the dc component; therefore, only the ac component appears on the output end of the capacitor. A transformer has the same action on a pulsating dc waveform; the transformer blocks the flow of dc, and when a pulsating dc waveform is applied to the primary, only the ac component appears at the secondary.

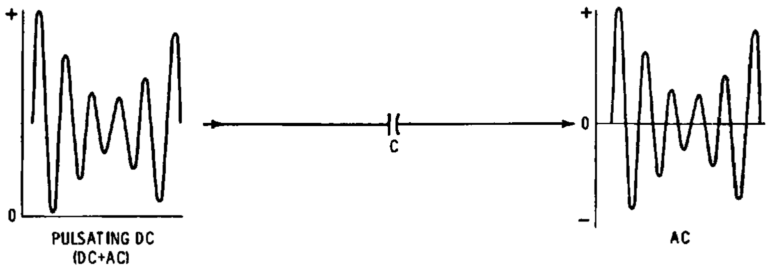


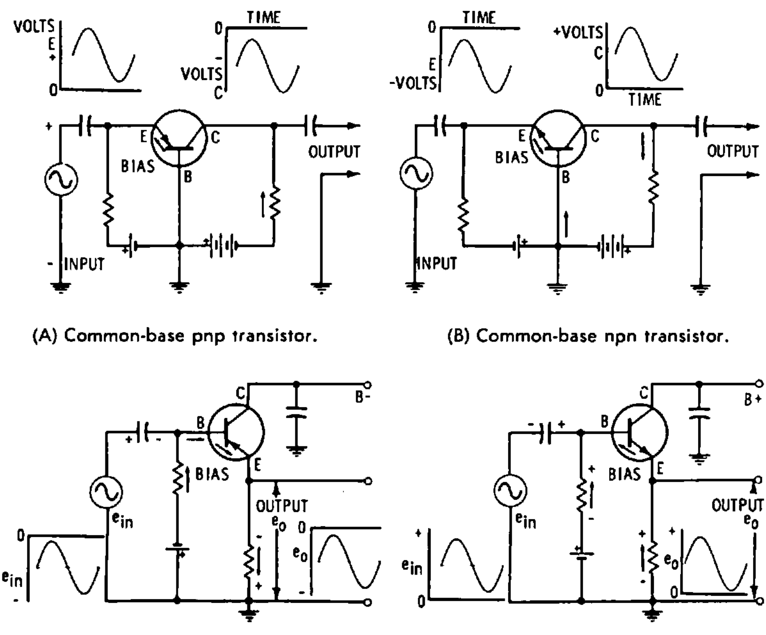
Fig. 1-16. Capacitor does not pass dc component.

Pulsating dc waveforms in the CB (common base) and CC (common collector) transistor amplifier circuits are seen in Fig. 1-17. Observe in Fig. 1-17A that the CB configuration does not reverse the signal phase from input to output. However, the emitter of the pnp transistor is biased positively, and a negative dc voltage is applied to the collector. In turn, the input waveform has negative polarity, but the output has positive polarity. Both are pulsating dc waveforms, but their polarities are reversed from input to output. If an npn transistor is used, as shown in Fig. 1-17B, the input waveform has positive polarity, but the output waveform has negative polarity. Both are pulsating dc waveforms.

In the CC transistor amplifier configuration shown in Fig. 1-17C, the output waveform has the same phase as the input waveform. The base is biased negatively, and the emitter is biased positively with respect to the base. Note carefully that current flow through the emitter load resistor produces a voltage drop that is negative with respect to ground. Therefore, the output waveform has negative polarity. In this arrangement, the input waveform is a negative pulsating dc voltage, and the output waveform is also a negative pulsating dc voltage. Next, with reference to Fig. 1-17D, the polarities are reversed because an npn transistor is used in the circuit. The input waveform is a positive pulsating dc voltage, and the output waveform is also a positive pulsating dc voltage.

If we use a dc scope to check the waveforms at the input and output terminals of a transistor amplifier, we observe a pattern such as illustrated in Fig. 1-18. Note that when no signal is applied to the vertical-input terminals of the scope, the beam rests at the 0-volt level. Next, when a pulsating dc signal is applied to the scope, the ac component of the waveform is displaced above the 0-volt level (assuming that the dc component is positive). The average value of the ac waveform rests at the dc component level in the pattern. If the scope has been calibrated, we can read the dc component voltage and the ac component pk-pk voltage directly from the screen.

The beginner should carefully note that the pattern shown in Fig. 1-18 is obtained only if a dc scope is utilized. If we employ an ac scope, the sine wave will not be displaced above the 0-volt level. Instead, the sine wave will appear centered on the 0-volt level. This is just another way of saying that an ac scope removes the dc component from a pulsating dc waveform. In turn, the dc component level coincides with the 0-volt level when an ac scope is used. Therefore, an ac scope cannot be used to measure the value of the dc component in a pulsating dc waveform.



(A) Common-base pnp transistor. (B) Common-base npn transistor.

(C) Common-collector pnp transistor. (D) Common-collector npn transistor.

Fig. 1-17. Waveforms in common-base and common-collector configurations.

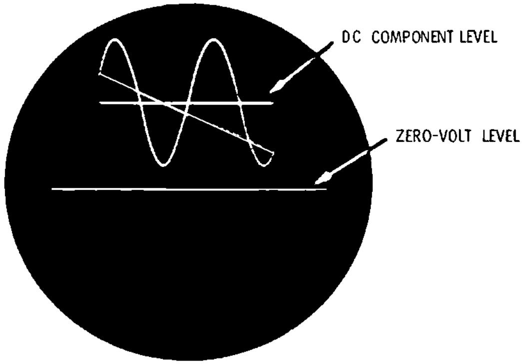


Fig. 1-18. Response of dc scope to ac voltage having dc component.

PULSE VOLTAGES

Some of the circuits in transistor TV receivers process pulse waveforms. Therefore, it is important for us to clearly understand the voltages that are specified in a pulse waveform. Fig. 1-19 shows the voltages in an ac pulse waveform. The positive-peak voltage is not equal to the negative-peak voltage. To find the peak-to-peak voltage of the pulse, we add the positive-peak and the negative-peak voltages. If this waveform is displayed on the screen of either a dc or an ac scope, the 0-volt axis of the pulse waveform coincides with the 0-volt level on the scope screen. In turn, if the scope is calibrated, we can read the values of the positive-peak voltage and of the negative-peak voltage directly from the screen.

Note carefully that the average of the ac pulse pictured in Fig. 1-19 is zero. This means that if the ac pulse is applied to a dc voltmeter, the pointer will not deflect. An average value of zero stems from the fact that the area enclosed by the positive excursion of the pulse waveform is exactly the same as the area enclosed by the negative excursion. Coulomb's law for quantity of electricity, or charge flow, is formulated:

$$Q = It = \frac{Et}{R} \qquad (1.2)$$

where,

Q is charge in coulombs,
t is time in seconds,
I is current in amperes,
E is potential in volts,
R is resistance in ohms.

Since the load resistance has a fixed value in a transistor amplifier circuit, the current value is proportional to the voltage value. Therefore, we recognize that voltage in Fig. 1-19 is proportional to current. In turn, the area enclosed by the positive excursion of the pulse waveform is proportional to the product of current and time. Similarly, the area enclosed by the negative excursion of the pulse waveform is proportional to the product of current and time. Formula (1.2) states that the product of current and time is equal to the quantity of electricity, or value of charge flow. In an ac pulse waveform, the positive and negative areas are always exactly equal. Consequently, there is just as much positive electricity in a pulse waveform as there is negative electricity, and the average value of the ac pulse waveform is necessarily zero.

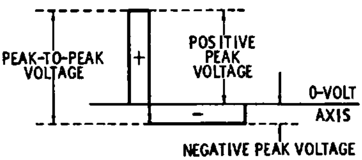


Fig. 1-19. Voltages in ac pulse waveform (positive and negative areas equal).

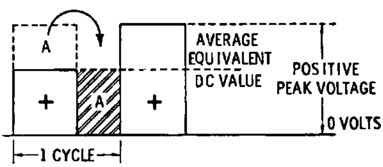


Fig. 1-20. Average value of dc pulse is equal to its dc component value.

Next, let us change the ac pulse shown in Fig. 1-19 into a dc pulse. This can be done by combining the ac pulse with a positive dc voltage that has a value equal to the negative peak voltage of the pulse waveform, thus obtaining a positive pulsating dc waveform. This waveform will have a positive peak voltage equal to the peak-to-peak voltage of the ac pulse. With reference to Fig. 1-20, it can be seen that the average value of a positive pulse is equal to its dc component value. That is, if we put the unshaded area A into the space indicated by the shaded area A, we have converted the dc pulse into a steady dc voltage. This means that if a dc pulse is applied to a dc voltmeter, the voltmeter will read the value of the dc component in the pulse.

Fig. 1-21 shows the meaning of the rise time of a pulse. Of course, a pulse has the same rise time, whether it is checked on a dc scope or an ac scope. In other words, the presence of a dc component in a pulse does not affect its rise time. If the pulse happens to have a dc component, the only difference between an ac scope display and a dc scope display is that the dc component will shift the waveform vertically on the screen of a dc scope. This results from the fact that application of a dc

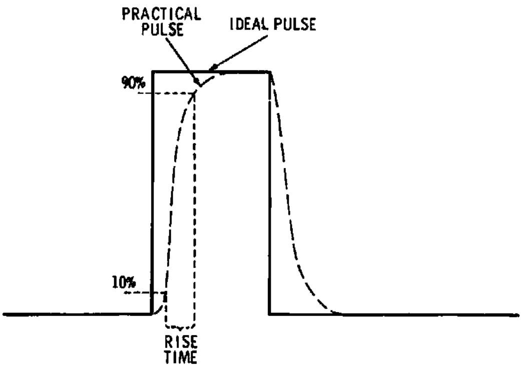
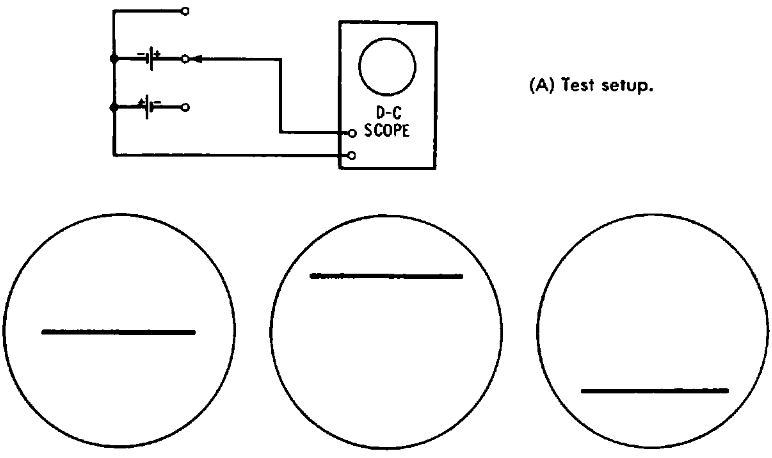


Fig. 1-21. Graphical definition of rise time.

voltage to a dc scope shifts the trace vertically, as shown in Fig. 1-22.



(A) Test setup.

(B) Zero volts applied. (C) Positive volts applied. (D) Negative volts applied.

Courtesy Allied Radio Corp.

Fig. 1-22. Application of dc voltages to dc scope.

RISE-TIME MEASUREMENT

Measurement of the rise time of an output pulse or square wave is a basic procedure in the analysis of various transistor amplifiers and various other transistor configurations. Therefore, we must know how to measure rise time. This is accomplished best by use of a triggered-sweep scope with calibrated time bases. A scope of this type used in pulse work is called a synchroscope. Fig. 1-23 is a block diagram for a typical syn-

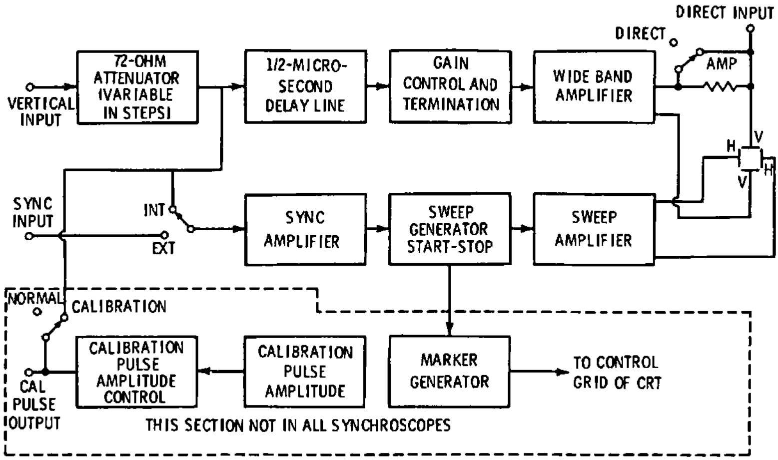
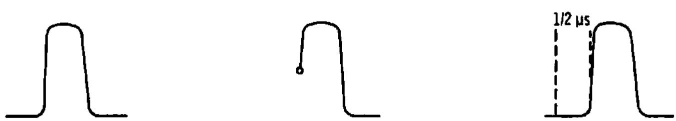


Fig. 1-23. Block diagram of typical synchroscope.

chroscope. Its central feature is a start-stop sweep generator; that is, the forward deflection interval does not start until the leading edge of a pulse (or other waveform) arrives via the sync amplifier. This feature facilitates the measurement of elapsed time.

Another important feature of the synchroscope depicted in Fig. 1-23 is the provision of a delay line in the vertical-amplifier channel. In this example, the delay line holds the incoming waveform for 1/2 microsecond before it is passed through to the vertical-deflecting plates. This provision is necessary in order that the entire leading edge of the waveform may be displayed on the scope screen—it takes almost 1/2 microsecond for the start-stop sweep circuit to "get started." Fig. 1-24 shows the effect of the delay line on an applied pulse. The input pulse is depicted at A; if no delay line is employed, the sweep circuit is slow in getting started, with the result that part of the leading edge is missing in the displayed waveform, as shown at B. However, when a delay line is used, the pulse waveform is held back for 1/2 microsecond in the vertical



(A) Input pulse. (B) Without delay. (C) With delay.

Fig. 1-24. Effect of delay on synchroscope trace.

channel, and the sweep circuit is given ample time to start. In turn, the complete leading edge of the pulse is displayed on the scope screen, as seen at C.

ACTION OF TRIGGERED-SWEEP CONTROLS

The time-base controls for a typical triggered-sweep scope are illustrated in Fig. 1-25. In usual operation, the Horizontal-Display switch is set to its Internal position; the horizontal amplifier is then driven by the sawtooth time base. Note that the Time-Base control is calibrated in microsecond, millisecond, and second intervals. A Variable control is also provided; the time base is uncalibrated when operating on the Variable function. In general, we operate the time base on one of its calibrated settings, so that we can measure rise time, or other elapsed-time interval in a waveform.

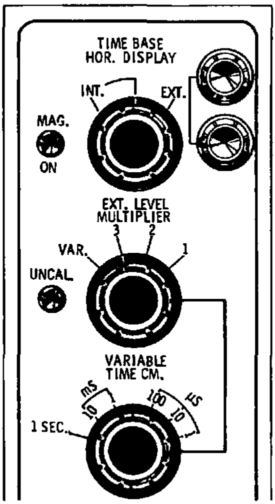
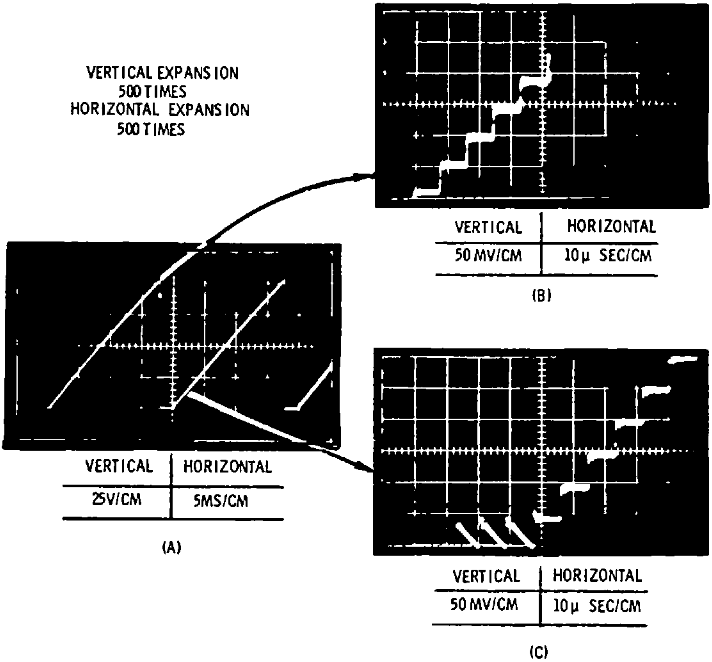


Fig. 1-25. Time-base controls of triggered-sweep scope.

Let us see how a waveform can be expanded for analysis of detail by operating the time base at high speed. In Fig. 1-26A, a combination sawtooth and stairstep waveform is shown as it appears when displayed at slow sweep speed. The steps in the waveform are invisible. However, when the vertical gain is advanced 500 times, and the sweep speed is likewise increased 500 times, the waveform detail appears clearly, as illustrated in Fig. 1-26B and C. Similarly, a pulse, square wave, or video signal can be expanded for analysis of detail.

The trigger controls of a typical triggered-sweep scope are shown in Fig. 1-27. Switch settings permit triggering to occur on either the positive or negative excursion of a waveform. Most waveforms are displayed in the ac trigger position. To the beginner, the dc trigger position might be misleading.



(A) Unexpanded waveform. (B) Expanded waveform, upper portion. (C) Expanded waveform, lower portion.

Fig. 1-26. "Stairstep" voltage waveform expanded 500 times.

Actually, the term "dc" in this case denotes that only the low frequencies of the signal are permitted to pass into the trigger section. This is a useful function for providing stable display and expansion of the color burst, for example. In the Automatic position, triggering occurs in a manner similar to the operation of a free-running scope. However, there is a basic difference in that synchronization is essentially automatic, and no sync-amplitude control is used.

When the trigger section is set to its Normal position, the Stability and Trigger-Level controls are operative. The Stability control must be operated over a suitable portion of its range, as depicted in Fig. 1-28. Typical response of the Stability control is as follows: At one extreme end of its range, we will usually obtain only a horizontal trace on the scope screen, as shown at A. Over the appropriate interval of control range, the desired pattern is displayed, as seen at B. At the

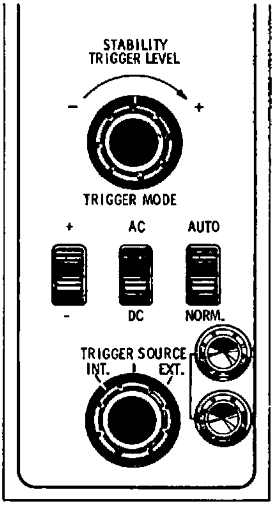
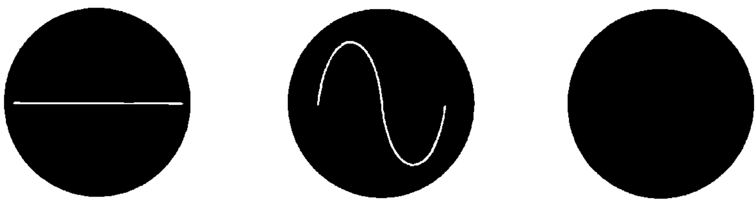


Fig. 1-27. Trigger controls of triggered-sweep scope.

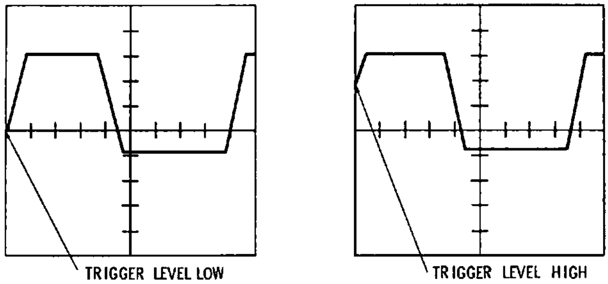
other extreme end of the control range, the screen often becomes blank, as shown at C. Suppose that we are using positive triggering. Then, the displayed waveform starts on its rising interval, as shown in Fig. 1-29A. By setting the Trigger-Level



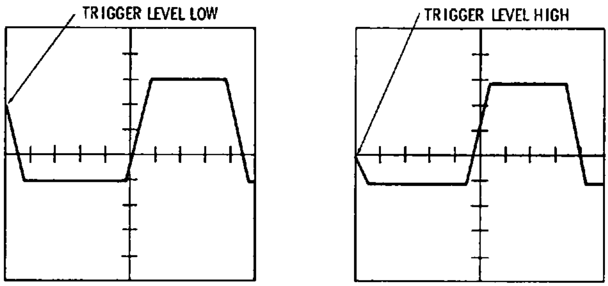
(A) Horizontal trace. (B) Desired pattern. (C) Blank screen.

Fig. 1-28. Stability-control action.

control suitably, we can trigger at any point along the rising interval. On the other hand, suppose that we are using negative triggering. Then, the displayed waveform starts on its falling interval, as shown in Fig. 1-29B. By setting the Trigger-Level control suitably, we can trigger at any point along the falling interval.



(A) Positive slope triggered (rising interval).



(B) Negative slope triggered (falling interval).

Fig. 1-29. Triggered scope waveforms.

REVIEW QUESTIONS

1. Name several common applications for transistors.
2. What fundamental waveforms are used for analysis of complex waveforms?
3. Distinguish between analysis and synthesis of a square wave.
4. How is rise time measured?
5. Define frequency-cutoff points for radio and TV receiver response curves.
6. Explain the relations of rms, peak, and peak-to-peak voltages in a sine wave.
7. How can dB values be measured directly on a scope screen?
8. Why is the average value of a sine wave equal to zero?
9. Describe a dc pulse.
10. Distinguish between ac, pulsating dc, and ac with a dc component.
11. Explain why transistors process pulsating dc waveforms.
12. How do pulsating dc waveforms differ in pnp and npn transistor configurations?

13. What is the effect of a coupling capacitor on a pulsating dc waveform?
14. Describe the display of a pulsating dc waveform on the screen of a dc scope.
15. How are positive-peak, negative-peak, and peak-to-peak values related in an ac pulse waveform?
16. Why is the average value of an ac pulse waveform equal to zero?
17. How can an ac pulse be changed into a dc pulse?
18. How is the average value of a dc pulse related to its dc component?
19. Briefly describe the plan of a triggered-sweep scope.
20. What is the function of a delay line in a triggered-sweep scope?
21. Name the controls associated with a calibrated time base.
22. Explain the action of a level control.
23. What is the effect of turning the stability control?
24. Give an example of an application in which dc triggering might be used.
25. In what units is a calibrated sweep control marked?

2

Waveforms in Transistor Oscillators

Many types of oscillators are utilized in transistor circuits. In the first analysis, these can be classified as sinusoidal and nonsinusoidal types. For examples, the local oscillator in a TV receiver is a sinusoidal oscillator. On the other hand, the vertical blocking oscillator is a nonsinusoidal oscillator. Subdivisions include free-running and triggered oscillators, monostable and bistable oscillators, stabilized oscillators, and other subdivisions relating to frequency characteristics. Some oscillators develop continuous-wave signals; others develop amplitude- or frequency-modulated signals. Sweep generators use fm oscillators, for example, to develop frequency-response curves as shown in Fig. 2-1. Let us consider the operating characteristics and waveforms of some basic transistor oscillators.

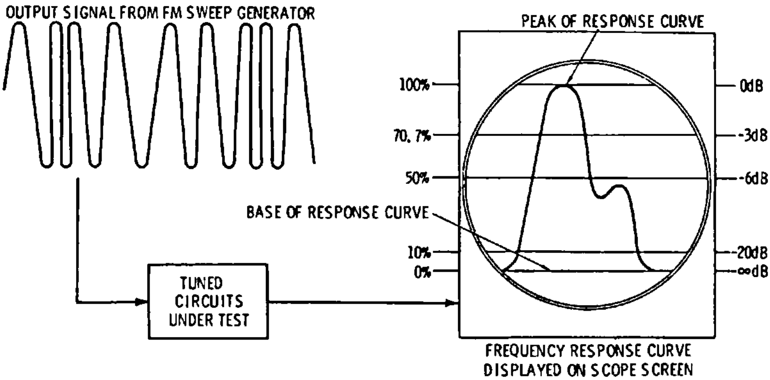
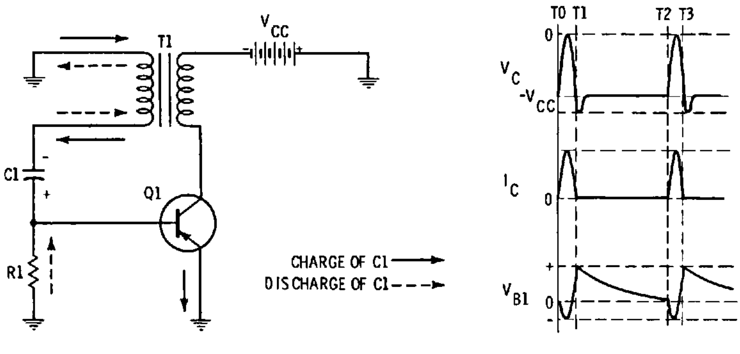


Fig. 2-1. Fm waveform used to develop a frequency response curve.

BLOCKING OSCILLATORS

A transistor blocking oscillator conducts for a short period of time and then is cut off (blocked) for a much longer period of time. A basic circuit for a transistor blocking oscillator using a pnp transistor is shown in Fig. 2-2. If an npn transistor were used in the circuit, the polarities of the collector supply battery V_{CC} would have to be reversed to maintain reverse bias across the collector-base junction. We will begin our analysis with the instant that the circuit is first energized. Current rises rapidly in the base circuit, due to the forward bias established across the base-emitter junction by the dc power supply V_{CC}. Simultaneously, the collector current I_C increases. This increasing collector current induces a negative voltage in the secondary winding of transformer T1.



(A) Schematic.

(B) Waveform.

Fig. 2-2. Transistor blocking oscillator with waveform.

The negative voltage that is induced in the secondary of T1 in Fig. 2-2A is applied at the base of the transistor, and increases the forward bias across the base-emitter junction. Meanwhile, capacitor C1 charges through the small forward resistance of the base-emitter junction; the transistor is quickly driven into collector saturation, whereupon this rapid regenerative action stops. At saturation, the collector current I_C becomes constant, and the collector voltage V_C falls to almost zero. Since no more voltage is induced in the secondary winding of T1, C1 stops charging, and starts to discharge through R1. Just before time T1 in Fig. 2B, the magnetic field in the secondary collapses, thereby inducing a voltage of opposite polarity that drives the base of Q1 positive. This reverse bias cuts the transistor off.

Cutoff corresponds to zero base and collector current in Fig. 2-2. At this time, the collector voltage V_C rises and makes the collector more negative than the V_{CC} supply for a short time following T1. This negative pulse is caused by the same collapse of the magnetic field that drives the base of Q1 positive. The transistor now remains cut off until the stored charge on C1 decays through R1. This occurs at time T2. As soon as there is base current it is amplified by Q1 and fed back regeneratively through T1 to the base. Therefore, Q1 is quickly driven into saturation, and the waveform cycle is repeated.

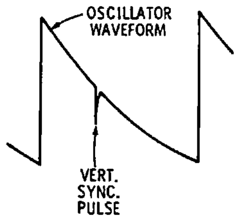


Fig. 2-3. Vertical sync pulse in TV receiver with slowly rolling picture.

Next, let us consider how the blocking oscillator in Fig. 2-2 can be synchronized. If negative-going sync pulses are coupled into the base circuit, they will ride on the V_{B1} waveform, and will extend downward toward the 0-volt level. It is evident that as a negative-going sync pulse approaches the 0-volt level in the time interval from T1 to T2, Q1 will be triggered into conduction somewhat sooner than if the sync pulse were absent. (See Fig. 2-3). Therefore, the blocking oscillator can be locked in frequency by the sync pulse. In normal operation, we adjust the value of R1 so that the oscillator runs a little slower than the repetition rate of the sync pulses. Accordingly, the incoming sync pulses speed up the repetition rate of the oscillator, compared with its free-running frequency.

Next, let us consider the free-running frequency of the blocking oscillator shown in Fig. 2-2. A capacitor charges and discharges through resistance according to an exponential waveform, as shown in Fig. 2-4. We measure charge and discharge intervals in *time constants*. One time constant is equal to the product, in seconds, of the capacitance and the resistance.

$$T = RC \tag{2.1}$$

where,

T denotes the time constant of the RC circuit in seconds,
R denotes the resistance in ohms,
C denotes the capacitance in farads.

Observe in Fig. 2-4 that a capacitor discharges to 37 percent of its initial voltage in one time constant. At the end of four or five time constants, the capacitor charge has decayed to nearly zero. Therefore, the free-running blocking oscillator in Fig. 2-2 has a repetition rate that is equal to approximately four or five time constants. Its exact repetition rate depends on the particular transistor, and how close its base voltage must approach zero to start a base-current. Next, with respect to the charging interval of C1, we recognize that as soon as there is base current, R1 is shunted by the low value of base-emitter resistance in Q1. This base-emitter resistance has a very low value, in the order of 35 ohms. Therefore, the charging time constant is very short, and C1 can be charged very rapidly.

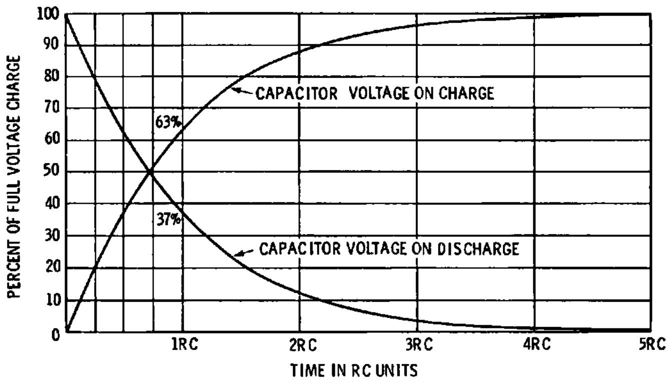


Fig. 2-4. Universal RC time-constant chart.

The exponential waveform chart shown in Fig. 2-4 is basic in analysis of nonsinusoidal oscillators. In other words, a blocking oscillator is basically an exponential waveform generator, as seen in the V_{B1} waveform of Fig. 2-2B. Therefore, it would be impractical to analyze the operation of the circuit on the basis of sine waves. To understand the circuit action, we must use exponential waveforms as our basic reference. Of course, the V_{B1} waveform in Fig. 2-2B is not a pure exponential; it begins and ends with pulses. That is, the base waveform is a combination exponential and pulse waveform. Combination waveforms are the rule, rather than the exception in transistor circuits.

Let us now compare the waveforms for transistor Q14 in Fig. 2-5 with the V_{B1} and V_C waveforms in Fig. 2-2B. At first glance, the waveforms seem to be different. However, the difference is chiefly in the scaling, rather than the waveshapes. Be-

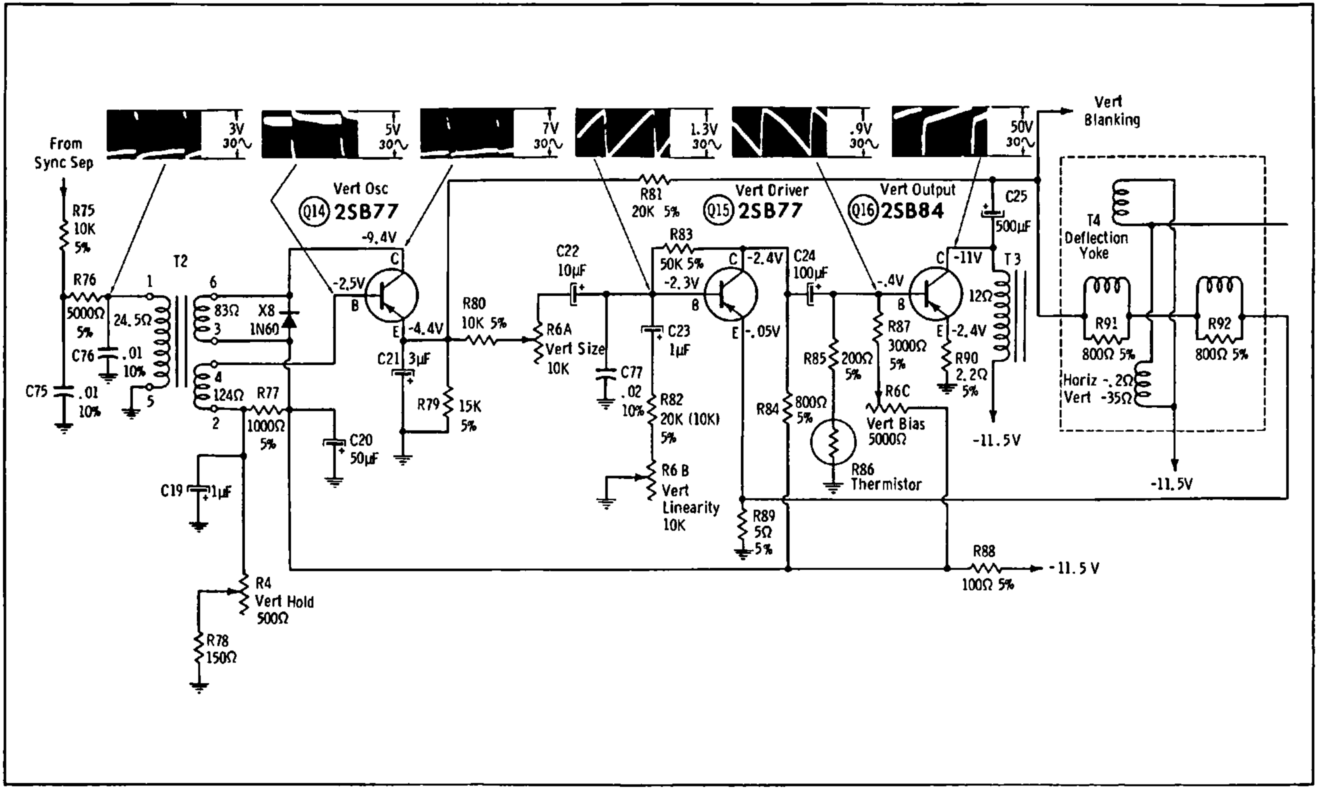


Fig. 2-5. Typical transistor vertical-sweep system.

cause of the additional circuit components, the base waveform of transistor Q14 has a large pulse component, and a less steeply sloping exponential component. Although there are superficial differences, the basic waveform is the same as that of V_{B1}. The collector waveform for Q14 is also basically similar to that of V_C. However, the positive peak of the waveform for Q14 is modified somewhat, due chiefly to the presence of diode X8.

Diodes are used in blocking-oscillator circuits to limit the peak voltage that is applied to the transistor by the collector waveform, and also to clip the sync pulses for stable triggering action. With reference to Fig. 2-6, resistor R5, in series with R4, operates as a damping control to set the amplitude of the output positive pulse. Diode X2 is connected across the primary of T1 to protect the transistor. That is, the collapsing magnetic field of T1 might exceed the breakdown voltage of the transistor used in the circuit. However, X2 is driven into forward conduction as the magnetic field collapses, and thereby limits the pulse amplitude across the primary. Additional reduction of amplitude can be obtained by adjustment of R5.

Resistor R3 serves two functions in this circuit. First, it provides emitter bias due to emitter current. As a transistor heats up, the emitter current tends to increase. If this tendency were not counteracted, the repetition rate would change. Since increased emitter current produces greater bias voltage across

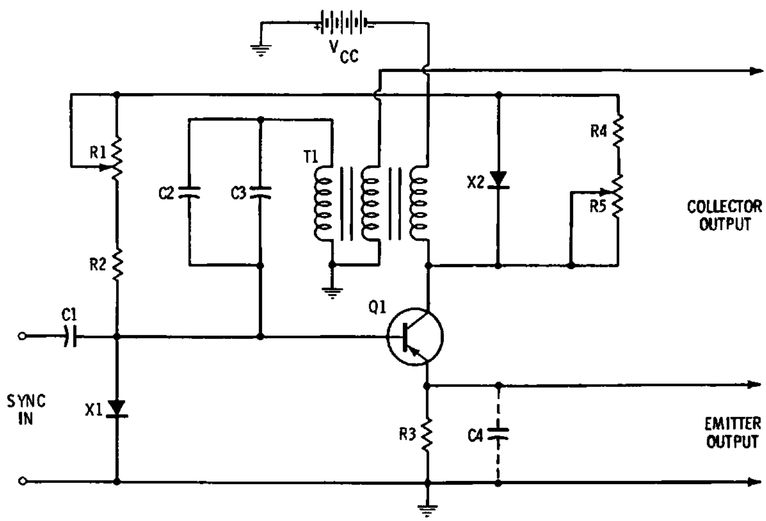


Fig. 2-6. Synchronized transistor blocking oscillator.

R3, the oscillator is self-stabilizing. Secondly, a pulse output signal can be taken from across R3. We often find a capacitor C4 used to bypass the emitter resistor. Since the capacitor tends to hold its charge between pulses, it improves stabilizing action. An emitter bypass capacitor also decreases the free-running frequency of the blocking oscillator, although this circuit action is merely incidental.

If C4 in Fig. 2-6 is chosen in a suitable value so that R3 is partially bypassed, another useful function can be served. That is, the capacitor partially discharges between pulses, with the result that an exponential waveform is produced by the emitter in accordance with the discharge waveform depicted in Fig. 2-4. Thereby, the emitter circuit operates as a waveshaping circuit, and produces a semisawtooth wave. Of course, this is not a true sawtooth waveform because it has an exponential curvature. However, additional waveshaping circuitry can be used to obtain a true sawtooth waveform.

The free-running frequency of the blocking oscillator in Fig. 2-6 is determined chiefly by capacitors C2 and C3, and resistors R1 and R2. That is, these capacitors and resistors have a certain time constant that determines when Q1 will come out of cutoff. Note also that R1 is not returned to ground as in Fig. 2-2, but, instead, to the voltage source V_{CC}. The reason for this connection is that V_{CC} causes Q1 to come to output cutoff earlier than if R1 were grounded. In turn, the exponential waveform at the base is not so greatly curved, and the trigger point is less affected by random noise voltages that might gain entry to the circuit.

Sync pulses are capacitively coupled to the base of Q1 in Fig. 2-6. Negative-going sync pulses are used, as previously explained. It is undesirable to have any positive noise pulses, or positive excursions of sync pulse applied to the base, because they would affect the time constant of the discharge circuit. Therefore, diode X1 is connected between base and emitter of Q1 to short-circuit any positive excursions that might be coupled into the oscillator circuit. Control R1 is adjusted to make the free-running frequency of the oscillator slightly lower than that of the incoming sync pulses. Output waveforms may be taken either from the tertiary winding on T1, or from the emitter of Q1, or both.

TRANSISTOR MULTIVIBRATORS

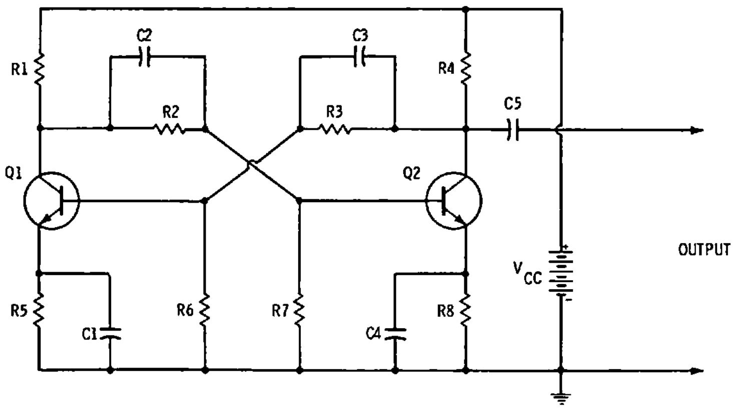
Transistor multivibrators are widely used in electronics technology. For example, many thousand multivibrators may be included in an electronic computer system. The free-running (astable) multivibrator is basically a two-stage amplifier with its output fed back to its input. RC circuitry is employed that has uniform response out to high frequencies. In turn, complex waveforms related to the basic square wave are produced. Fig. 2-7A shows a typical configuration for an astable transistor multivibrator. While one stage conducts, the other stage is cut off. At a certain point that depends on RC time constants, the stages reverse their conditions—the conducting stage suddenly cuts off, while the nonconducting stage suddenly starts to conduct. Thus, a form of square wave is generated. The emitter-coupled transistor multivibrator configuration employs feedback via its common-emitter resistor. On the other hand, a collector-coupled multivibrator employs coupling capacitors as shown in Fig. 2-7A.

The waveforms for the collector-coupled multivibrator arrangement are shown in Fig. 2-7B. The circuit comprises a two-stage, RC-coupled, common-emitter amplifier that uses npn transistors in this example. If pnp transistors were used, the collector supply battery V_{CC} would be reversed in polarity. Output from the first stage is coupled to the input of the second stage; also, output from the second stage is coupled to the input of the first stage. We recall that there is a 180-degree phase shift (polarity reversal) between the input and output of a CE configuration. This means that the output of each stage is in phase with the input to the other stage. Regenerative feedback provides the signals required for sustained oscillation.

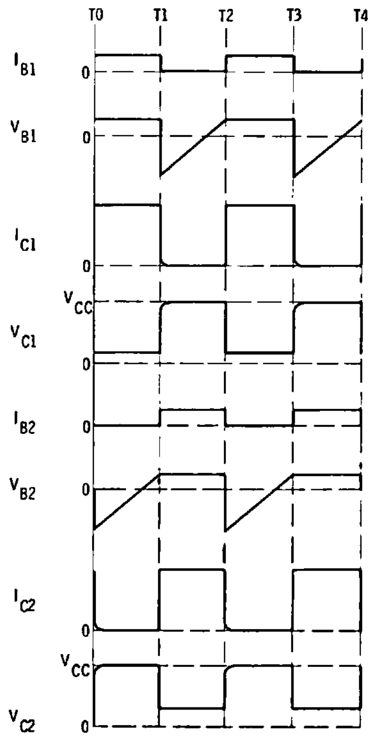
Let us assume in Fig. 2-7A that Q1 conducts more heavily than Q2 when operating voltages are applied. From time T0 to time T1 in Fig. 2-7B, collector current I_{C1} and collector voltage V_{C1} remain constant; capacitor C2 discharges through resistor R7 in accordance with the exponential waveform depicted in Fig. 2-4. As C2 discharges through R7, the voltage drop across R7, which is also the base voltage of Q2 (V_{B2}), decreases; this effectively decreases the reverse bias on Q2. This action continues until time T1, whereupon forward bias is re-established across the emitter-base junction of Q2, and Q2 starts to conduct.

As the collector current I_{C2} through Q2 increases, collector voltage V_{C2} becomes less positive from T1 to T2. This voltage, coupled through capacitor C3 to the base of Q1, drives the base more negative and causes a decrease in collector current I_{C1} through Q1.

Increased positive voltage at the collector of Q1 in Fig. 2-7A is coupled through C2 and appears across R7. Collector current



(A) Schematic.



(B) Waveforms.

Fig. 2-7. Transistor multivibrator.

I_{C2} through Q2 therefore increases. This process continues rapidly until Q1 is cut off. Transistor Q1 remains cut off and transistor Q2 conducts until C3 discharges sufficiently through R6 to decrease the reverse bias V_{B1} on the base of Q1 so that the operating cycle repeats following time T2. When checking waveforms with a scope, beginners are sometimes confused by the aspect of a waveform. That is, if the scope controls are varied, a square wave may appear in different scaling, as demonstrated by Fig. 2-8. Note that the second aspect is not a pulse, but is still a square wave, because the duty cycle is equal to 1.

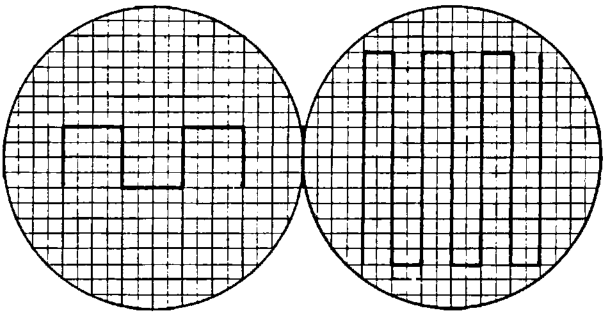


Fig. 2-8. Two aspects of square wave.

Resistors R1 and R4 in Fig. 2-7A are the collector load resistors for Q1 and Q2, respectively. Suitable bias for the base of Q1 is provided by a voltage-divider network comprising R3 and R6. Base bias voltage for Q2 is provided by R2 and R7. Resistors R5 and R8 are temperature stabilization resistors for Q1 and Q2. C1 and C4 are emitter bypass capacitors. The repetition rate of the multivibrator depends on the time constants of the RC-coupling networks, modified by the bias voltages which determine the point on the exponential decay curve that a transistor comes out of cutoff. The output signal is coupled through capacitor C5 to the load. The signal is basically a square wave, although it must be passed through a subsequent waveshaping circuit to obtain a good approximation to an ideal square wave. Output could also be taken from the collector of Q1.

Electronic switches are designed around astable multivibrators. A typical configuration is shown in Fig. 2-9. The multivibrator transistors are Q8 and Q9. Three repetition rates can be obtained by setting S1 to switch different values of coupling capacitors into the circuit. Amplifiers A and B pass input signal

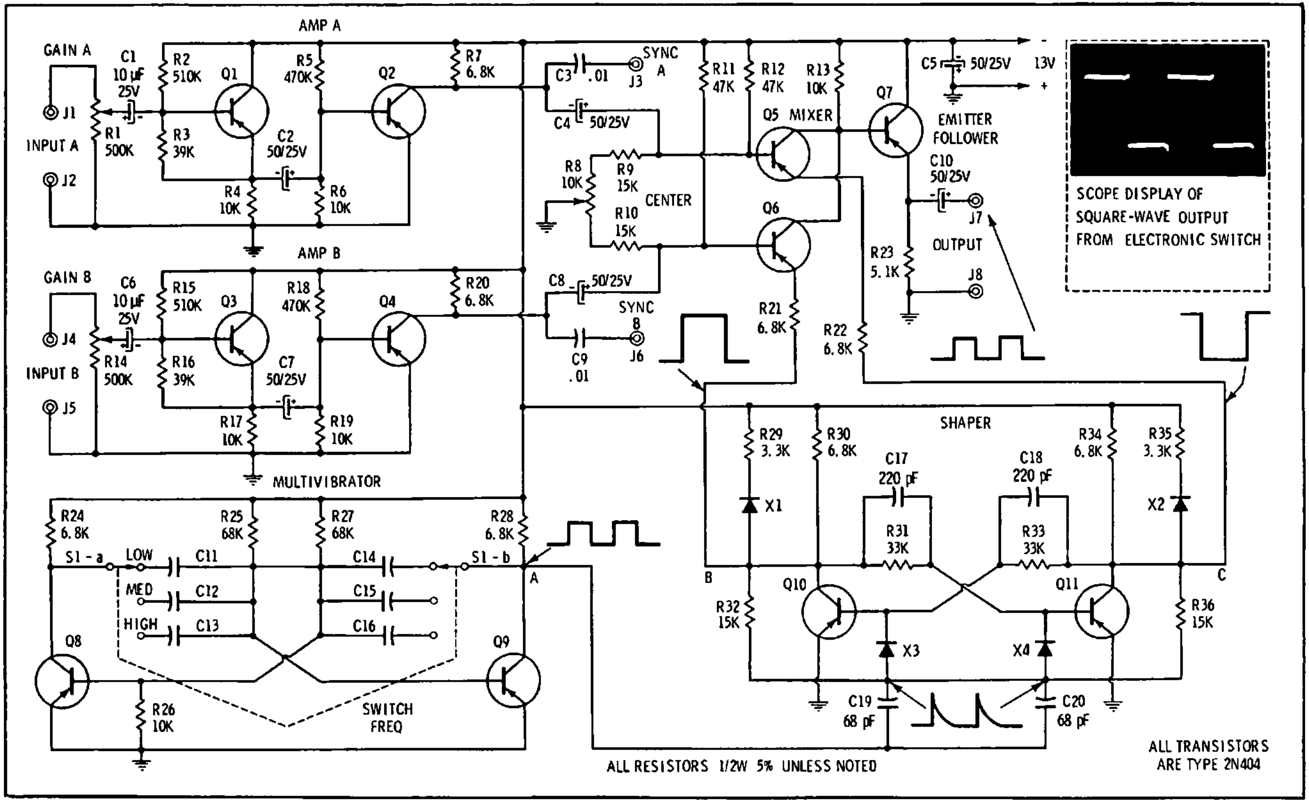
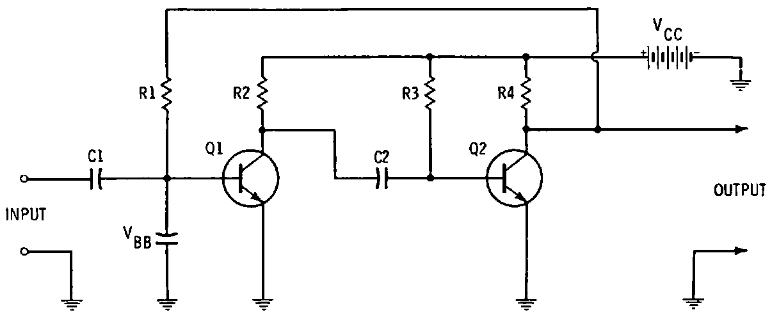


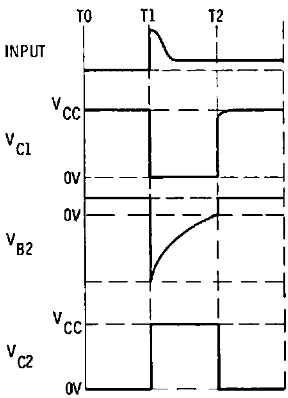
Fig. 2-9. Transistorized electronic switch.

alternately because the square wave causes Q5 and Q6 to conduct alternately. To obtain good switching action the output from the multivibrator is passed through the waveshaper comprising Q10 and Q11. In addition, the rise time of the square wave is improved. Signals at input A and input B are switched alternately into Q7, which operates in the common-emitter configuration and provides isolation between the mixer section and the scope input circuit. Details of waveshaping circuits are discussed subsequently.

Another important type of multivibrator is called the monostable, one-shot, single-shot, or single-swing configuration, as shown in Fig. 2-10A. When this configuration is triggered, it goes through a complete cycle of operation, and then returns to its original quiescent state. An external trigger pulse moves the operating point from the initial stable region. The time constant of the RC components holds the operating point in its new stable region for a determined period of time. Then, the operating point moves back into the original stable region.



(A) Schematic.



(B) Waveforms.

Fig. 2-10. Transistor one-shot (monostable) multivibrator.

When in its initial stable region, or quiescent condition, transistor Q2 is in saturation and Q1 is cut off. A positive trigger pulse of short duration applied to the base of Q1 causes the multivibrator to go through its active cycle. Then, the circuit rests in its quiescent condition until another trigger pulse arrives.

Battery V_{CC} supplies the necessary collector voltages for the transistors in Fig. 2-10A, and also supplies the forward bias for the emitter-base junction of Q2. Transistor Q2 is in saturation during the quiescent period; therefore, the collector voltage is practically zero. Fig. 2-10B shows the collector waveform V_{C2} during this time from T0 to T1. Reverse bias across the emitter-base junction provided by battery V_{CC} maintains Q1 at cutoff. Collector voltage V_{C1} is positive and equal to battery voltage V_{CC}. The collector waveform C_{C1} during this same time is shown in Fig. 2-10B. Capacitor C2 is charged to the value of battery voltage V_{CC} through the essentially short-circuited base-emitter junction of forward-biased Q2 and the collector load resistor R2.

If a positive input trigger pulse is applied at time T1 to the base of Q1 via C1, Q1 is driven into conduction. Collector voltage V_{C1} at time T1 decreases, or becomes less positive. C2 immediately starts to discharge through R3. Regenerative signal action from the collector of Q1 occurs via C2 to the base of Q2. A maximum negative voltage V_{B2} is applied to the base of Q2 at time T1. This places reverse bias across the emitter-base junction, and drives Q2 toward cutoff. As collector current through Q2 decreases, the collector voltage rises. This rise is coupled via R1 to the base of Q1, and increases its positive voltage. This regenerative action causes a rapid change of state in both transistors. It drives Q1 toward saturation and drives Q2 toward cutoff. When collector current ceases to flow at time T1, T_{C2} rises to equal V_{CC}.

From time T1 to T2 in Fig. 2-10B, C2 is discharging via R3 and the low saturation resistance of Q1. Capacitor C2 reaches 0 volts at T2, and Q2 then has the necessary reverse bias to keep the collector current cut off. Hence, at time T2, collector current starts to rise, drives Q2 into saturation, and brings V_{C2} to 0 volts. At this time, regenerative voltage from the collector of Q2 is coupled to the base of Q1 through R1 and falls to 0 volts, and V_{CC} once again provides reverse bias driving Q2 back into cutoff.

The collector voltage V_{C1} in Fig. 2-10 again equals V_{CC} at time T2, and capacitor C2 recharges through the emitter-base junction of Q2 and resistor R2. C2 continues to charge up to the collector supply voltage, when further circuit action ceases. If another trigger pulse is applied at the base of Q1, another cycle of operation ensues. Output is taken from the collector of transistor Q2. The collector waveform V_{C2} is basically a square wave, although its rise time can be improved by processing the waveform through suitable waveshaping circuitry. The duration of the output pulse from T1 to T2 is primarily determined by the time constant of R3 and C2 during discharge. Of course, the repetition rate of the output waveform is determined by the input trigger pulses.

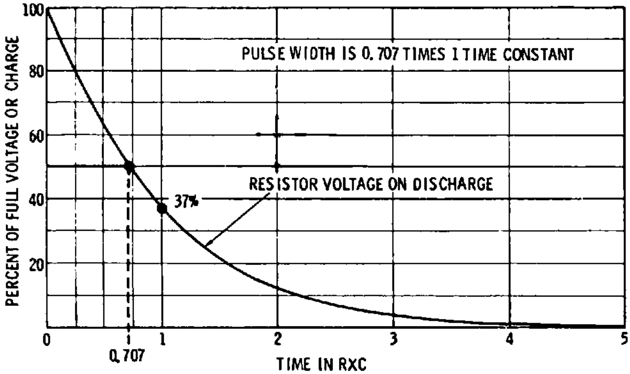
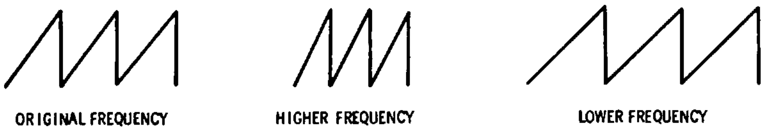


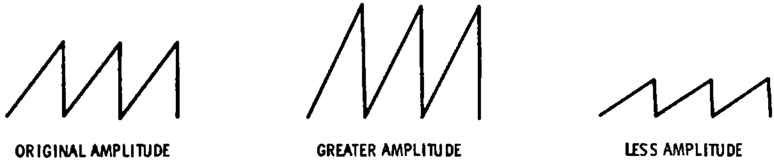
Fig. 2-11. Pulse width measurement.

Waveform V_{B2} in Fig. 2-10B is a semisawtooth, and may also be regarded as a pulse waveform. Its decay is essentially exponential. We measure pulse width between the 50 percent of maximum points on the leading and trailing edges, as indicated in Fig. 2-11. Waveform V_{B2} is usually called a pulse because it is not immediately followed by another cycle. In other words, a pulse is a waveform that occurs at separated time intervals. On the other hand, if the same waveform is immediately followed by another excursion, it is usually called a semisawtooth waveform. Fig. 2-12 illustrates sawtooth waveforms that have the same amplitude, but different frequencies, and sawtooth waveforms that have the same frequency, but different amplitudes.

Basically, frequency and repetition rate have the same meaning. However, we usually speak of the frequency of sine waves and the repetition rate of complex waves. Frequency is measured in hertz, and repetition rate can also be denoted in hertz. However, it is customary to speak of repetition rates in terms of pulses per second (pps). The term "pulses" in this connotation is very broad, and applies to almost any complex



(A) Sawtooth waveforms with same amplitude.



(B) Sawtooth waveforms with same frequency.

Courtesy Allied Radio Corp.

Fig. 2-12. Distinction between amplitude and frequency.

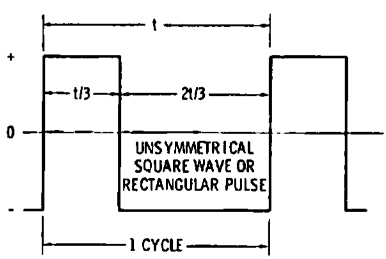
waveform. We will find that the distinction between pulses and square waves, for example, is not sharply drawn. Fig. 2-13 depicts a waveform that is called an unsymmetrical square wave, or a rectangular pulse. If the RC time constants are made unequal in Fig. 2-7A, the output waveform is not a square wave, but a rectangular waveform.

Exponential pulses are related to square waves as shown in Fig. 2-14. The voltage drop across the resistor is an exponential pulse, and is called a differentiated square wave. The complex waveform that is dropped across the capacitor is called an integrated waveform. These terms follow from mathematical analyses that are beyond the limits of our format. The RC circuit obeys Kirchhoff's voltage law:

$$E_1 = E_2 + E_3 \qquad (2.2)$$

The lower diagram in Fig. 2-14 shows how the addition of waveforms E_2 and E_3 results in a sum equal to waveform E_1. In this example, the time constant of the RC circuit is sufficiently short with respect to the repetition rate of the square-

Fig. 2-13. Rectangular waveform.



wave input that E_2 falls practically to zero before the next excursion starts. The peak-to-peak voltage of E_2 is practically double the peak-to-peak voltage of E_1; E_3 has the same peak-to-peak voltage as E_1. With reference to Fig. 2-4, we recognize that waveform E_3 has the same basic curvature as waveform E_2 "turned upside down." This fact follows from Formula (2.2).

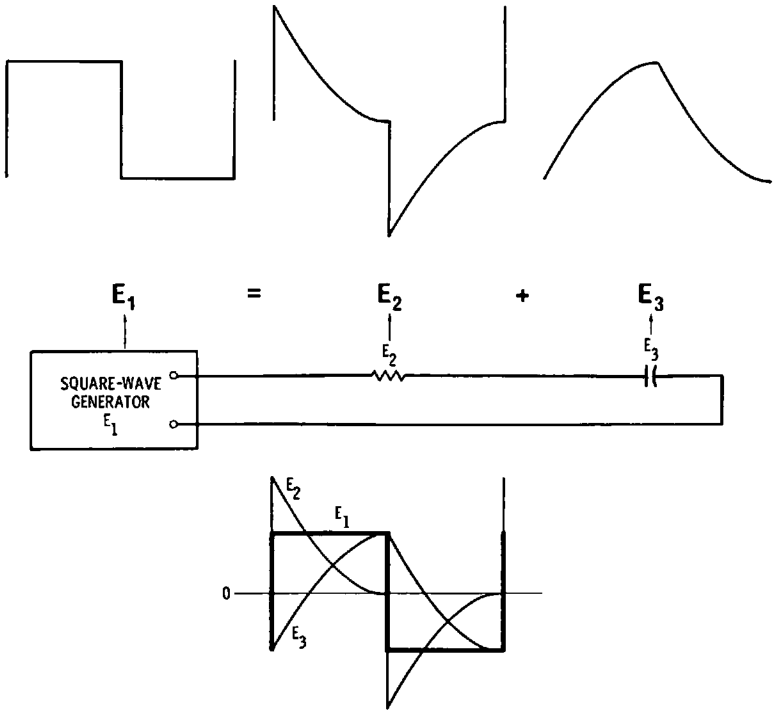


Fig. 2-14. Kirchhoff's voltage law applied to square-wave response of RC circuit.

Pulses can be synthesized, or built up, from sine waves, just as in the case of square waves. Fig. 2-15 depicts the build-up of a symmetrical ac pulse from a fundamental and its harmonics. Of course, a very large number of harmonics must be included to obtain a pulse waveform that has smooth leading and trailing edges. We will find that the harmonic frequencies in waveforms E_2 and E_3 in Fig. 2-14 are the same as the harmonic frequencies in a square wave. This is necessarily so because the square-wave harmonics flow into the RC circuit, and none of these harmonics are removed; neither are other harmonics added or generated by the linear RC circuit. There-

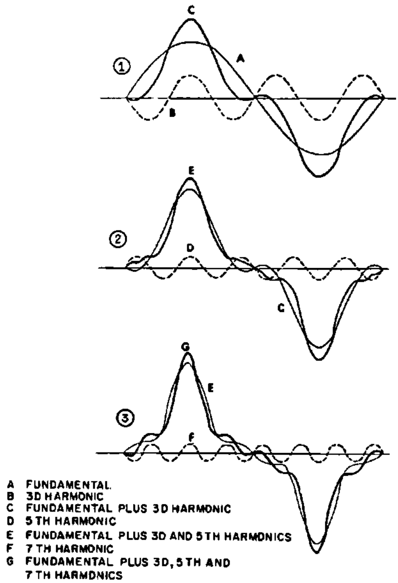


Fig. 2-15. Build-up of positive pulse followed by negative pulse.

fore, the only distinction between the harmonics in E_1 and in E_2 or E_3 consists in their comparative amplitudes. That is, the capacitor has decreased reactance for the higher harmonic frequencies and passes these frequencies at comparatively high amplitude.

This is just another way of saying that the drop across the resistor in Fig. 2-14 is a waveform that has harmonics with large amplitudes, compared to the input square wave—the capacitor reduces the amplitude of the fundamental current, and also the amplitudes of the lower harmonic currents. Therefore, the differentiated waveform contains comparatively large-amplitude harmonics. The opposite condition is found across the capacitor—the fundamental and low harmonic frequencies produce a comparatively large voltage drop across the capacitor.

The effect of varying the time constant of an RC circuit on the differentiated and integrated waveforms for an applied square wave is shown in Fig. 2-16. An RC coupling circuit is the same as a differentiating circuit except that it has a very long time constant—in excess of 5RC. Therefore, a coupling circuit does not noticeably differentiate a square wave. Fig. 2-16 also shows the effects of differentiation and integration of an applied sawtooth waveform, when the RC circuit has various time constants. When the time constant is in excess of 5RC, we can regard the configuration as a coupling circuit.

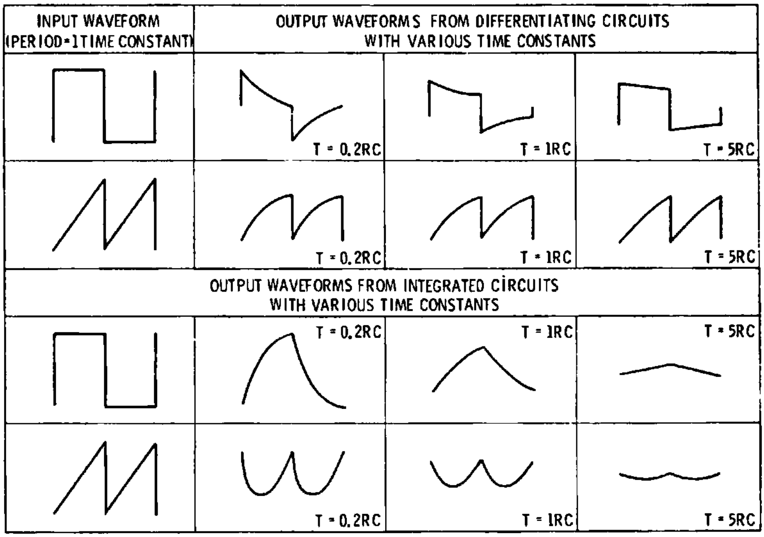


Fig. 2-16. Square-wave and sawtooth reproduction in series RC circuits with differing time constants.

REVIEW QUESTIONS

1. Name two basic classifications of transistor oscillators.
2. Define a blocking oscillator.
3. Describe the general nature of the waveforms produced in a blocking-oscillator circuit.
4. What determines the polarity chosen for a sync pulse in a blocking oscillator?
5. Explain the units of time used in a universal RC time-constant chart.
6. How much does an exponential waveform decay in one time constant?
7. Discuss the function of diodes utilized in a blocking-oscillator circuit.
8. Define a free-running (astable) multivibrator.
9. Describe the general nature of the waveforms produced in a multivibrator circuit.
10. Explain what is meant by the aspect of a square wave.
11. How does an electronic switch operate?
12. Define a one-shot multivibrator.
13. Describe the general nature of the waveforms produced in a one-shot multivibrator circuit.
14. How is pulse width measured?
15. Distinguish between a semi-square wave and an ideal square wave.

16. How can the shape of a semi-square wave be improved?
17. Explain the distinction between symmetrical and unsymmetrical square waves.
18. Describe the application of Kirchhoff's voltage law to a series RC circuit that is energized by a square wave.
19. How are the harmonics in a differentiated pulse related to the harmonics of a square wave?
20. Distinguish between a differentiating circuit and an RC coupling circuit.
21. How does the time constant of a differentiating circuit affect the output waveform?
22. Describe the effects of differentiation and integration on a sawtooth waveform.
23. Why is an exponential waveform sometimes called a semi-sawtooth waveform?
24. Describe a frequency-modulated wave.
25. State an application for a frequency-modulated waveform.

3

Other Transistor Oscillators and Waveforms

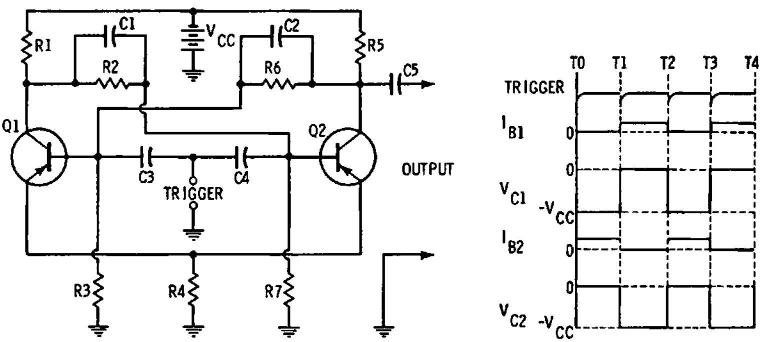
The distinction between oscillators and waveshapers is not sharply drawn. For example, a one-shot multivibrator is often called a monostable oscillator; it is also called a pulse regenerator. A pulse regenerator is used to produce a standard pulse waveform. For example, in an electronic computer, a pulse must pass through many circuits; as it progresses through various circuits, the pulse waveform becomes distorted. Since excessive distortion can result in erratic operation, a pulse that has undergone a certain amount of distortion may be fed to a one-shot multivibrator that restores the standard pulse shape and width. In this application, the one-shot multivibrator operates solely as a waveshaper. That is, whether we regard a configuration as an oscillator or as a waveshaper depends on its application.

Thus, the bistable (Eccles-Jordan) multivibrator shown in Fig. 3-1 may be regarded either as an oscillator or as a waveshaper. When bistable multivibrators are connected in cascade, they function as memory systems in electronic computers (see Fig. 3-2); in this application the bistable multivibrator is called a flip-flop (FF). If neon bulbs or other visual indicators are connected into the cascade circuit, binary readout of stored numbers is provided. Details on this are reserved for subsequent discussion. As a waveshaper, a bistable multivibrator may be triggered by the output from an audio oscillator. In turn, a square-wave output is provided that has the same repetition rate as that of the audio oscillator.

Let us consider the circuit action of the bistable multivibrator in Fig. 3-1A. It differs from a one-shot multivibrator primarily in that it requires two input triggers to complete one cycle of output. A flip-flop is in a stable state when either transistor is conducting while the other transistor is cut off. The transistor states are switched by application of a trigger pulse. When dc power is first applied, one transistor conducts, whereas the other transistor is cut off. The conducting transistor quickly saturates; the collector current of the other transistor is practically zero. Each transistor remains in this stable state each being steadily biased by the other, until a trigger pulse is applied. A negative pulse applied to the base of a pnp transistor in its conducting state has no switching action. In other words, the transistor is already conducting. On the other hand, the same negative trigger pulse applied to the base of the cut-off pnp transistor will change the state of the circuit.

The transistor that was cut off is driven briefly into conduction by the trigger pulse and feedback causes the other transistor to be driven into cutoff. Thus, the states are reversed, and the transistors remain in this stable state, each being steadily biased by the other. Note that a positive trigger pulse applied to the base of the conducting transistor will also change the state of the circuit. Basically, the method used to trigger an FF is determined by the polarity and magnitude of the available trigger pulse, and the desired repetition rate of the output waveform. Although the FF depicted in Fig. 3-1A uses a trigger pulse applied to the bases of both transistors, a trigger pulse of sufficient amplitude and proper polarity could also be applied to the collector of either transistor.

Let us assume that Q1 is conducting and Q2 is cut off in Fig. 3-1A. A negative trigger pulse applied to the base of each tran-

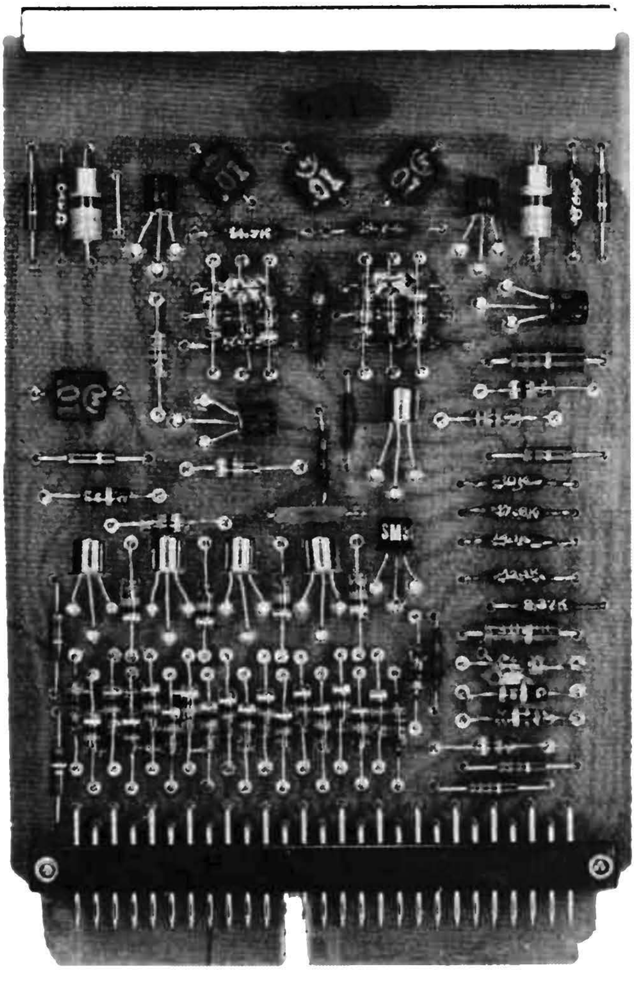


(A) Schematic.

(B) Waveforms.

Fig. 3-1. Transistor Eccles-Jordan multivibrator.

sistor at time T0 in Fig. 3-1B causes Q2 to conduct. The increasing collector current causes the collector voltage V_{C2} of Q2 to decrease (become less negative). This voltage change is coupled to the base of Q1, and reduces its forward bias. Collector current through Q1 begins to decrease, and collector voltage V_{C1} changes from zero to a negative value that approaches the battery voltage V_{CC}. This voltage change is cou-



Courtesy Computer Control Co., Inc.

Fig. 3-2. Transistor module for digital computer.

pled to the base of Q2, making the base more negative, and increasing the conduction of the transistor. Regenerative feedback continues until Q2 is saturated and Q1 is cut off.

Transistor Q2 in Fig. 3-1A continues conducting, and Q1 remains cut off. The transistors remain in this state until the next trigger pulse is applied at time T1; then Q1 conducts and Q2 is cut off. This completes one full cycle of the output square wave. The time constant of C1 and R2 and C2 and R6 basically determines the time from conduction to cutoff of Q1 and Q2. The capacitors provide rapid coupling of voltage changes, and the switching action is therefore rapid. The output waveform is coupled via C5 to the load. Note that output could be taken from Q2. Since an FF requires two input triggers to produce a complete output cycle, it operates as a frequency divider with a ratio of 2 to 1.

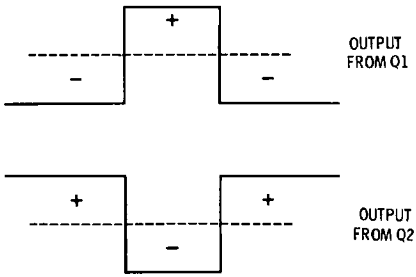


Fig. 3-3. Push-pull square-wave outputs.

Resistors R1 and R5 in Fig. 3-1A provide the dc collector loads for the transistors. The necessary forward bias for Q1 and Q2 is provided by a voltage-divider network comprising R3 and R6 for Q1, and R7 and R2 for Q2. R4 provides emitter temperature stabilization, and provides operating stability. C3 and C4 couple the input trigger pulse to the base of each transistor. Note that if outputs are taken from both Q1 and Q2, we obtain a push-pull output, as shown in Fig. 3-3. The rise time of the output waveform depends on the ratio of collector-load resistance to the stray capacitance in the load circuit. Therefore, faster rise time can be obtained by using low values of collector load resistance. However, the output amplitude tends to decrease with low values of collector load resistance because the transistor's internal resistance is not zero during collector saturation.

When a transistor is saturated, its internal resistance may be in the order of 35 ohms. If a 2,000-ohm collector-load resistance is used, the collector voltage drops to nearly zero during saturation. On the other hand, if we use a 75-ohm collector-load resistance, the collector voltage will be about one third of the supply voltage during saturation. In either case, the collector voltage will be practically equal to the supply voltage when the transistor is cut off, because its internal resistance is very high in this state.

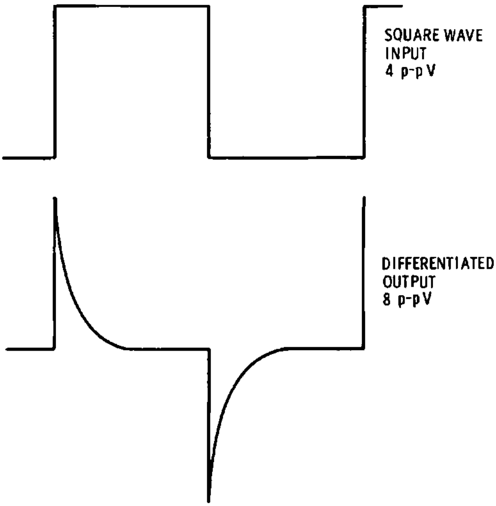


Fig. 3-4. Effect of differentiation on square wave.

RC CIRCUIT DIFFERENTIATION VERSUS MATHEMATICAL DIFFERENTIATION

Consider the case in which C5 in Fig. 3-1A feeds into a comparatively low value of resistance. Then, the square-wave output becomes differentiated into a narrow exponential pulse, as depicted in Fig. 3-4. If a very short time constant is used, the pulse becomes very narrow, and approaches an impulse. An impulse has zero width, as shown in Fig. 3-5, and represents the mathematical differentiation of a square wave. If an ideal square wave is applied to an ideal integrator, an output waveform is obtained that is called a back-to-back sawtooth wave. An approximate back-to-back sawtooth wave is obtained by RC integration of a square waveform.

Note in Fig. 3-5 that ideal differentiation of a sawtooth waveform produces an impulse waveform. Ideal integration of an ideal back-to-back sawtooth wave produces a parabolic waveform. In color-TV receivers, sawtooth waveforms are integrated by RC circuits to produce approximate parabolic

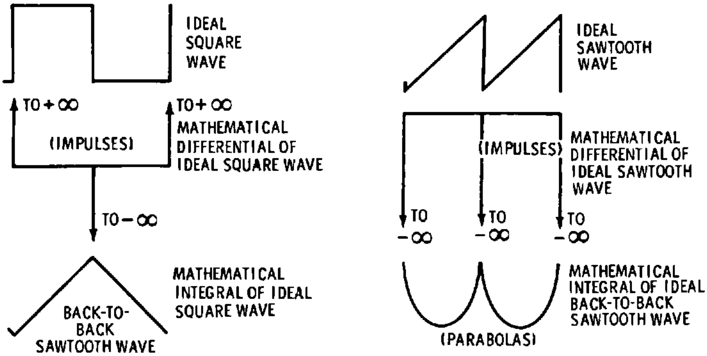


Fig. 3-5. True differentials and integrals of ideal square and sawtooth waveforms.

waveforms. This circuit action is employed in convergence circuitry. We know that when a square wave is differentiated, the fundamental and low-frequency harmonics are reduced in amplitude, compared with the high-frequency harmonics. Therefore, the harmonics are stronger in a pulse than in a square wave. Fig. 3-6 shows the comparative harmonic amplitudes for a square wave and a rectangular pulse. The narrower the pulse, the stronger the comparative amplitudes of the high-frequency harmonics.

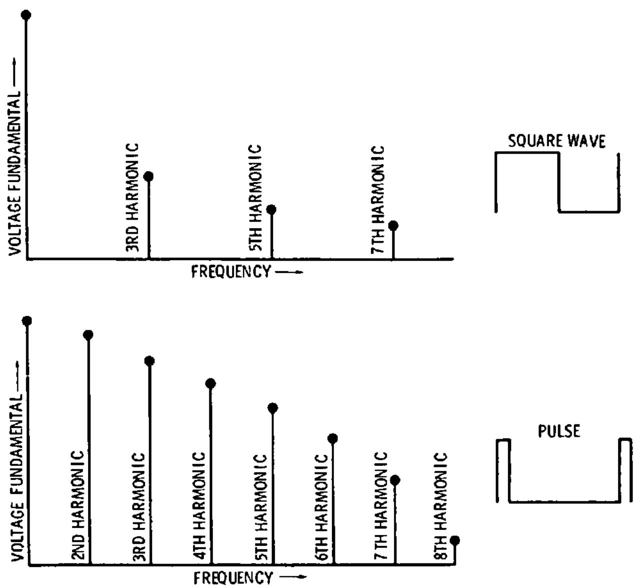


Fig. 3-6. Harmonic content of square waves and pulses.

In an impulse wave, we have a theoretical limiting situation in which the pulse width is zero, the pulse amplitude is infinite, and all the harmonics in the impulse have the same amplitude as the fundamental. Of course, a true impulse waveform cannot be produced in practice, although exponential pulses that approach an impulse waveform can be produced by using a differentiating circuit with a very short time constant. With reference to the ideal sawtooth waveform in Fig. 3-5, we note that a sawtooth waveform can be synthesized or built up from sine waves, as shown in Fig. 3-7. Mathematically speaking, this

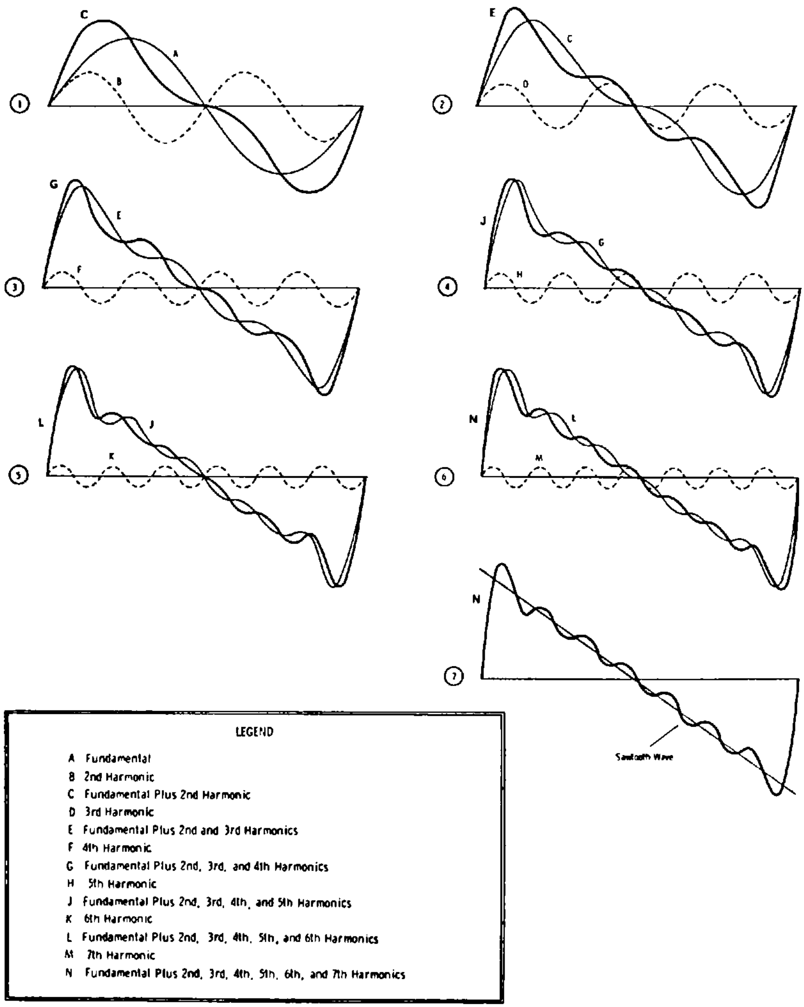


Fig. 3-7. Composition of sawtooth waveform.

synthesis represents the Fourier series for a sawtooth waveform. Although useful conclusions can be drawn from the synthesis viewpoint, we must be on guard to avoid misapplication of this concept. For example, if an ideal sawtooth waveform is integrated, the higher-frequency harmonics are removed. In turn, we might expect the output waveform to be "wavy"; however, the integrated output is a smoothly curved waveform.

Let us see why misapplication of the synthesis concept leads us to a false conclusion in the foregoing example. If an ideal sawtooth waveform were to be built up from sine waves, as depicted in Fig. 3-7, an infinite number of harmonics would be required. This corresponds to a circuit that has infinite bandwidth. In practice, however, we know that it is impossible for a circuit to have infinite bandwidth. Therefore, conclusions drawn from the synthesis concept may be in error. In an RC integrating circuit, exact analysis requires that we abandon the synthesis concept, and describe the circuit action on the basis of exponential waveforms. An ideal exponential waveform can exist in a circuit that has limited bandwidth. On the other hand, an infinite number of harmonics cannot exist in a circuit that has limited bandwidth.

EXPANSION OF RECTANGULAR WAVEFORMS

Although the output from a square-wave or pulse generator, or from the circuit in Fig. 3-1, might seem to have straight lines, sharp corners, and vertical sides, it is easy to show that such waveforms are actually combinations of various exponential waveforms. For example, if we expand any pulse waveform on the screen of a triggered-sweep scope, we observe the development illustrated in Fig. 3-8. At high sweep speeds, it becomes evident that the corners are not sharp, but instead are rounded; the leading edge of the waveform is not vertical, it slopes—the pulse has a certain rise time. In summary, the waveform actually consists of a certain combination of exponential waveforms.

A square wave or pulse produced by a generator has a certain rise time; a scope also has a certain rise time. If the rise time of the scope is very fast, compared to the rise time of the generator, we regard the rise time displayed on the scope screen as the rise time of the generator. On the other hand, if the rise time of the generator is very fast, compared with the rise time of the scope, we regard the rise time displayed on the scope screen as the rise time of the scope. Again, when the rise times of the generator and of the scope are in the same order

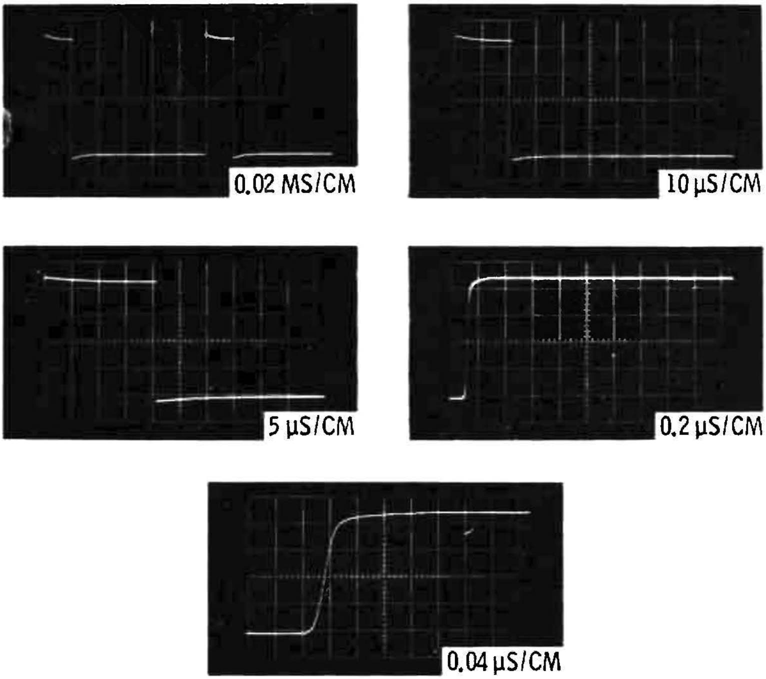


Fig. 3-8. Expansion of 20-microsecond pulse with increasing sweep speed.

of magnitude, the rise time displayed on the scope screen depends on both the generator rise time and the scope rise time. It can be shown that the rise time in this situation is formulated:

$$\text{Rise Time} = \sqrt{T_1^2 + T_2^2} \qquad (3.1)$$

where,

T_2 denotes the rise time of the scope's vertical amplifier, and T_1 denotes the rise time of the generator.

Fig. 3-9 illustrates the meaning of Formula (3.1). The rise time of the input square wave is T_1; the rise time of the scope's vertical amplifier is T_2. In turn, the displayed waveform has a slower rise time that is given by Formula (3.1). Practical

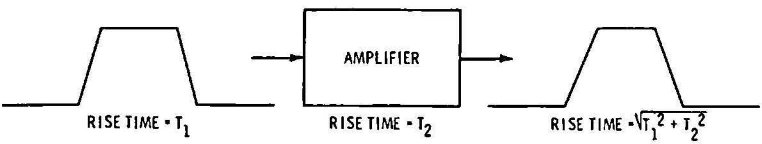


Fig. 3-9. How rise times combine.

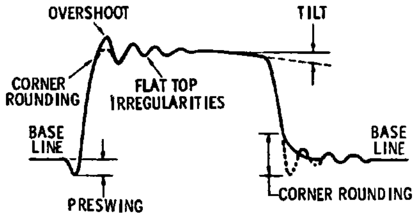


Fig. 3-10. Distortions commonly observed in square and pulse waveforms.

square waves and pulses often have various distortions, as depicted in Fig. 3-10. We may observe preshoot, overshoot, ringing, tilt, and corner rounding. In measuring rise time, preswing and/or overshoot are disregarded in locating the 10-percent and 90-percent points on the leading edge. That is, we assign a zero value to the base line, and a 100-percent value to the flat top. For example, consider the points depicted in Fig. 3-11; the flat top is preceded by overshoot and ringing. We regard the flat top as the 100-percent level, and locate the 10-percent and 90-percent points accordingly for measurement of rise time.

Let us briefly consider the circuit actions that produce the distortions pictured in Fig. 3-10. Preswing is not observed in simple circuits; however, it can be produced in delay lines, and also in certain types of video amplifiers, such as used in color-TV receivers. Corner rounding occurs because all circuits have residual inductance and capacitance, which prevent current from flowing instantaneously when a voltage is suddenly applied. Therefore, a small elapsed time is required for the circuit to change from its resting state and to enter the rise interval of the waveform. The relation of rise time to bandwidth was given in Formula (1.1), which we repeat here:

$$T_r = \frac{1}{3f_c} \tag{3.2}$$

Note that Formula (3.2) is not absolutely exact, but is sufficiently accurate for practical work. Overshoot is caused by leakage inductance and stray capacitance in a circuit or amplifier. That is, the leakage inductance of peaking coils in a video amplifier can be tuned by shunt stray capacitance to form a resonant circuit that "rings" when a square wave or pulse is applied. The first half-cycle of the ringing waveform appears as overshoot in the reproduced square-wave or pulse pattern. Subsequent excursions appear as a ringing interval that has the form of a damped sine wave. This means that the sine wave decays exponentially. Tilt is produced by a small amount of differentiation through RC-coupling circuit. Undershoot and ringing following the trailing edge of a waveform are produced by leakage inductance and stray capacitance.

Preshoot or undershoot, and overshoot are sometimes encountered also in transistor waveshaping circuits. In such a case, the "pips" result from a strong drive waveform applied to an RC-coupling circuit; the coupling circuit is associated with a semiconductor device that starts to differentiate the applied drive waveform, and then quickly settles down to a "flat top" response. In turn, the differentiated "pip" appears in the output waveform as an overshoot, undershoot, or preshoot, depending on the circuit configuration. Well-designed waveshaping circuits are free from "pips" in the output waveform. The damping time of a ringing waveform is measured as shown in Fig. 3-12. Damping time is measured in microseconds or milliseconds with a triggered-sweep scope that has calibrated sweeps.

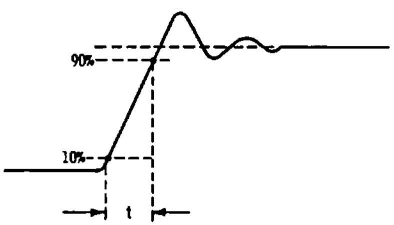


Fig. 3-11. Measurement of rise time t.

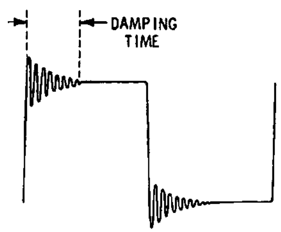
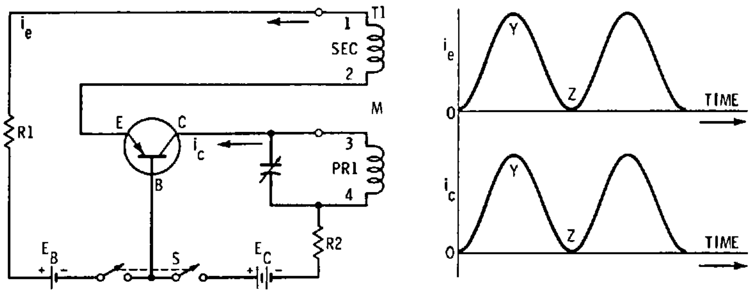


Fig. 3-12. Damping time of ringing waveform.

SINUSOIDAL TRANSISTOR OSCILLATORS

Sinusoidal transistor oscillators normally generate output waveforms that approach a sine waveshape. A well-designed oscillator of this type generates an almost ideal sine wave. A pnp transistor oscillator using a tickler coil for inductive feedback is shown in Fig. 3-13. The base is common to the input and output circuits. E_B biases the emitter-base input circuit for forward-current flow. V_C biases the collector-base junction in the reverse direction. Waveforms of emitter current i_E and collector current i_C are shown in Fig. 3-13B. Oscillation starts when switch S is closed. Emitter current i_E, and collector current i_C, both increase from O to Y, because of regenerative feedback in T1 from the 3-4 winding to the 1-2 winding.

At point Y in Fig. 3-13B, the transistor is saturated and no further increase in current flow can occur. Therefore, feed-



(A) Schematic. (B) Waveforms.

Fig. 3-13. Tickler feedback oscillator.

back ceases and i_E starts to decrease. In turn, collector current i_C starts to decrease. Feedback via T1 reverses the signal polarity and causes i_E to decrease to zero. At point Z, feedback again ceases. When the feedback voltage has driven the transistor to cutoff, the bias voltages return to their original values, and the oscillation cycle is repeated. The time between cutoff and saturation of the transistor is determined chiefly by the resonant frequency of the tank circuit. The frequency of the generated sine wave is stated approximately by the familiar resonant-frequency formula:

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{3.3}$$

The higher the Q value of the tank circuit, the more nearly does the oscillating frequency approach the value given by Formula (3.3). Also, the higher the Q value of the tank, the more nearly does the output waveform approach a true sine-wave shape. If the Q value is quite low, the output waveform will be a badly distorted sine wave. The Q value of the tank circuit is formulated:

$$Q = \frac{X_L}{R} \tag{3.4}$$

where,

X_L is the inductive reactance of the tank coil,
and R is its effective series resistance.

Inductive reactance, we recall, is formulated:

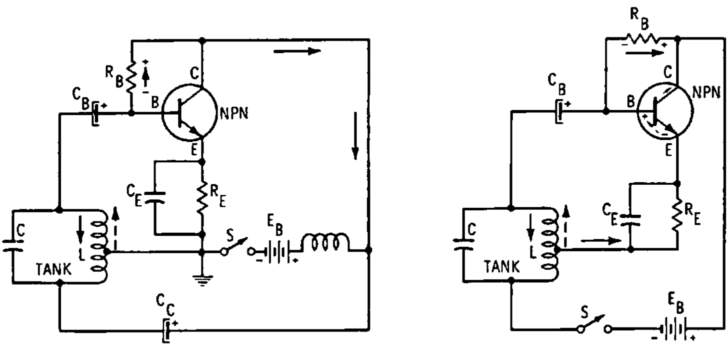
$$X_L = 2\pi f L \tag{3.5}$$

where,

X_L denotes reactance in ohms,
f denotes frequency in hertz,
and L denotes inductance in henrys.

Note that the effective resistance of L is not simply its winding resistance. That is, the coil has a high-frequency (rf) resistance that is greater than its dc resistance. Moreover, the effective resistance of the tank coil is increased by the shunt resistance of the transistor and the resistance that is coupled into the tank coil by the tickler coil. Therefore, calculation of the effective resistance of the tank coil is comparatively difficult, and is not developed in detail here.

In Fig. 3-14, an npn transistor is employed in a Hartley oscillator circuit. The tapped coil is returned to the common-emitter point. The collector circuit may be either shunt-fed, as in Fig. 3-14A, or series-fed, as in Fig. 3-14B. Resistor R_B limits the base-emitter bias current to a suitable value. L and C values of the tank circuit determine the frequency of oscillation. C_B prevents the tank coil from short-circuiting the base-emitter junction of the transistor. R_E limits the collector current to a safe value, and C_E eliminates negative feedback. V_B provides bias current for the input, and collector voltage for the output circuit.



(A) Shunt-fed collector.

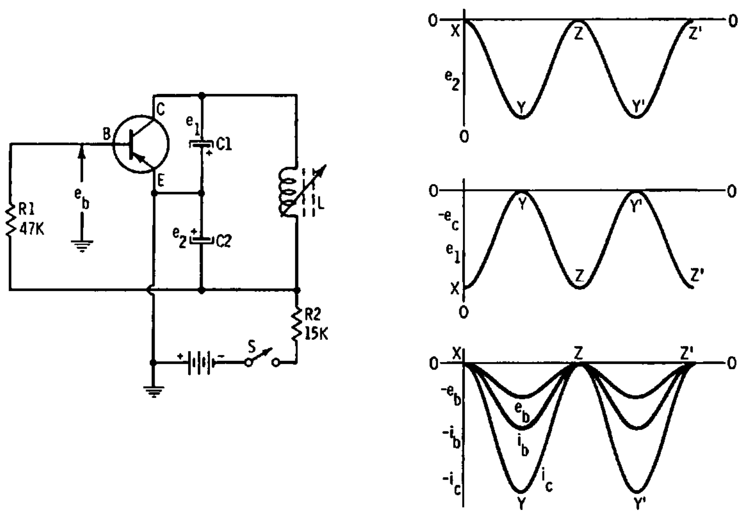
(B) Series-fed collector.

Fig. 3-14. Npn junction-transistor Hartley oscillator.

We observe that the tapped tank coil produces regenerative feedback in the same basic manner as a tickler coil. The oscillation frequency is somewhat lower than the value given by Formula (3.3) because of the loading that the transistor imposes on the tank circuit. If the Q value of the tank circuit is high, the output waveform approximates a true sine wave. On the other hand, low Q values are associated with a distorted sinewave output. It is also necessary to locate the tap point on the tank coil properly, to obtain a good sine waveform. When the shunt-fed version is used, the value of the rf choke must be suitably chosen; otherwise, the circuit may refuse to oscillate, or the output waveform may be badly distorted.

Fig. 3-15 shows a pnp transistor connected in a CE configuration as a Colpitts oscillator circuit. The capacitive voltage divider C1-C2 takes the place of a tapped coil in a Hartley arrangement. When switch S is closed, C1 and C2 charge with the indicated polarities. Base current i_B flows on the initial charge of C2; collector current flows through R2, and the collector voltage e_C decreases from X to Y, due to the voltage drop across R2. Capacitor C2 charges to about 3/4 of the battery voltage through R2; during this time, i_B and i_C increase. After this period of time, C2 starts to discharge through L into C1. The voltage across C1 is initially low at point Y; as the voltage across C2 falls, i_B and i_C both decrease to cutoff. Meanwhile e_C increases with the charge of C1 from Y to Z.

Next, C1 (Fig. 3-15) discharges into C2 through L. During the next portion of the cycle, from Z to Y, C2 acquires charge, i_B increases, and i_C starts to flow again. The collector voltage e_C drops, due to increasing voltage drop across R2. During this part of the cycle, the battery supplies current to C2 to meet the tank-circuit losses. Capacitor C2 then stops charging; e_B and i_B stop their increase at point Y′. The base-emitter bias reverts to its normal value; i_C decreases as C2 begins to discharge into C1 through L (points Y′ to Z′). The voltage e_2 across C2 de-



(A) Schematic. (B) Waveforms.

Fig. 3-15. Colpitts transistor oscillator.

creases; e_B, i_B, and i_C decrease from Y′ to Z′. Meanwhile, e_1 and e_C increase. As i_B decreases to cutoff, i_C decreases to zero, and C1 starts to discharge again into C2, thereby repeating the cycle at point Z′.

To develop a good sine waveform, the Q value of the tank must be high. That is, the transistor represents a nonlinear resistance in shunt with the tank coil, and a good sine wave (voltage waveform) can be dropped across the tank only if it has a dominant flywheel effect on the circuit action. In other words, the ratio of stored energy to dissipated energy in the tank circuit must be large. When the Q value is high, the energy that is withdrawn and dissipated in the nonlinear circuit resistance represents only a small fraction of the available energy surging back and forth between L and C1-C2 in Fig. 3-15. The optimum sine waveform also depends on the ratio of C1 to C2, which must be suitably chosen.

Output is commonly taken by coupling a load coil to L in Fig. 3-15. Tight coupling results in high output, at the expense of good waveform. That is, a load coil couples effective resistance into the tank circuit when it extracts energy from the oscillator. Therefore, light loading results in an improved waveform at reduced output power. Note also that a load coil couples reactance into the tank circuit, and changes its unloaded resonant frequency. A changing load results in frequency shift. An oscillator has maximum frequency stability when its loading is minimum. To isolate an oscillator from its load circuit, a buffer stage, such as a CE stage, can be inserted between the oscillator circuit and the load.

REVIEW QUESTIONS

1. Describe a pulse regenerator.
2. Explain the circuit action of a bistable multivibrator.
3. Give a typical application for a flip-flop circuit.
4. What is meant by the term "frequency-divider" circuit?
5. Distinguish between RC circuit differentiation and mathematical differentiation.
6. How does an impulse differ from a rectangular pulse?
7. What is the result of ideal integration of a square wave?
8. What is the result of ideal integration of a back-to-back sawtooth wave?
9. Why cannot a true impulse waveform be generated in practice?
10. How can an impulse waveform be approximated by a differentiating circuit?

11. Compare the number and amplitudes of harmonics in a pulse waveform with those in a square wave.
12. Explain how the rise time of an input square wave combines with the rise time of the vertical amplifier in a scope.
13. Define preswing, overshoot, ringing, and tilt.
14. What circuit actions can you describe that cause overshoot and/or ringing?
15. Describe the damping time of a ringing waveform.
16. Name three common types of sinusoidal transistor oscillators.
17. What is a tickler coil?
18. Why is a Hartley oscillator basically similar to a tickler feedback oscillator?
19. Define the Q value of a coil.
20. Define the inductive reactance of a coil.
21. What are the two basic types of Hartley oscillators?
22. Explain the basic configuration of a Colpitts oscillator.
23. How is oscillatory energy commonly extracted from the foregoing oscillators?
24. State the factors that can contribute to distortion of an oscillator waveform.
25. How can the frequency stability of an oscillator be maximized?

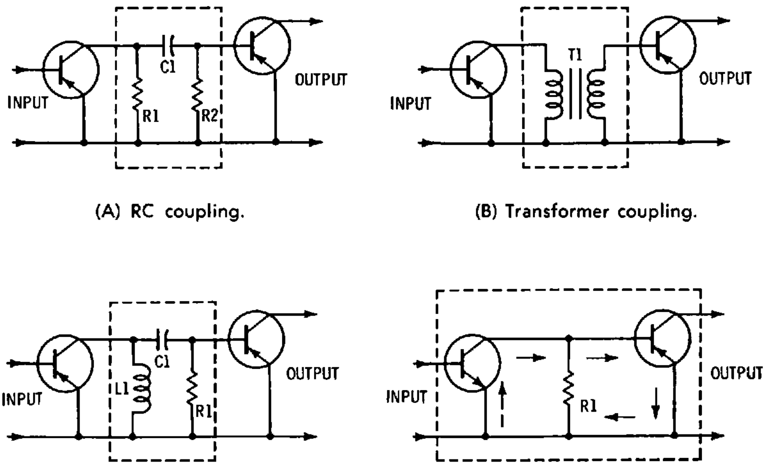
4

Transistor Amplifiers and Waveforms

Many types of transistor amplifiers are used in various applications. This chapter explains only the simpler types of amplifiers, and the waveforms associated with amplifier circuit action. RC coupling is used extensively with junction transistors. In Fig. 4-1A, R1, C1, and R2 form the RC network between the two transistor stages. C1 blocks dc and couples the ac signal; R2, the dc return resistor for the input element of the second stage, develops the signal applied to the input of the second stage. C1 prevents the dc voltage on the collector of the first stage from appearing at the input terminal of the second stage.

Note that the reactance of the coupling capacitor (which is in series with the low-valued input resistance of the following stage), must be small compared with the input resistance. Otherwise, an excessive amount of signal voltage will be dropped across the capacitor. A high value of capacitance is used because the input resistance of the following stage is low —usually lower than 1000 ohms. However, the physical size of the capacitor is comparatively small because of the low voltages in the circuit. The ohmic value of the dc return resistor is usually 7 to 15 times the value of the transistor's base input resistance of the second stage. This fairly high ratio of resistance is necessary to prevent shunting an objectionable amount of signal current around the input circuit of the second stage.

Since the reactance of the coupling capacitor increases as the signal frequency decreases, the very low frequencies are attenuated. The shunting effect of the collector-emitter capacitance of the first stage, and the base-emitter capacitance of the second stage limit the high-frequency response of the amplifier (assuming that the transistor is of a type that is capable of amplifying the high frequencies). Any transistor is rated for a beta cutoff frequency by the manufacturer, and may also be rated for an alpha cutoff frequency. In either case, the cutoff frequency of the transistor is determined by the frequency at which the response is down 3 dB, in an amplifier circuit that is capable of processing frequencies past the cutoff point of the transistor. It is found that the alpha cutoff frequency of a transistor is often higher than its beta cutoff frequency. Therefore, a transistor may be used in the CB configuration when very high frequencies are to be processed.



(A) RC coupling.

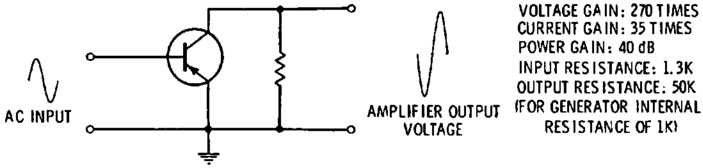
(B) Transformer coupling.

(C) Impedance coupling.

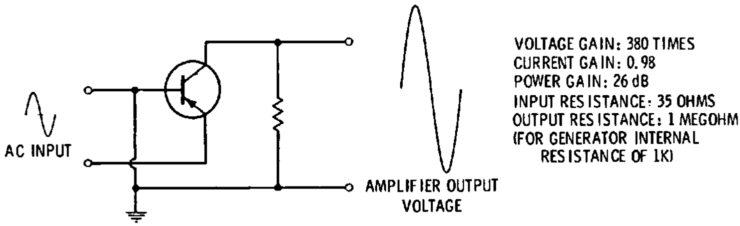
(D) Direct coupling.

Fig. 4-1. Interstage coupling networks.

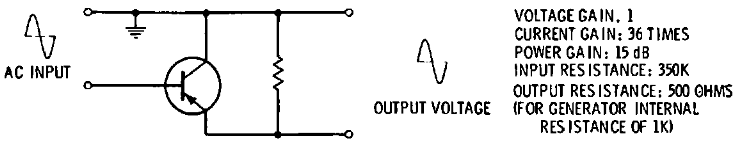
The efficiency of an RC-coupled amplifier is equal to the ratio of ac power output to dc power supplied to the stage. This efficiency value is rather low, due to the I^2R power loss in the collector load resistor. Other amplifier configurations provide considerably greater efficiency. RC coupling is used to a large extent in audio amplifiers, as in low-level low-noise preamplifiers, in driver amplifiers, and in high-level power amplifiers. The chief advantages of RC amplification are high gain, production economy, and compact construction. Because batteries such as are used in portable equipment have limited current capability, this type of equipment is generally limited to low-power operation. Fig. 4-2 shows a summary of the principal parameters of single-stage transistor amplifiers.



(A) Common emitter.



(B) Common base.



(C) Common collector.

Fig. 4-2. Transistor circuit parameters.

FREQUENCY RESPONSE OF RC-COUPLED AMPLIFIERS

Simple RC-coupled amplifiers have maximum gain over their midband region, with reduced gain at the low- and high-frequency ends of their response range. Low-frequency gain is limited chiefly by the increasing reactance of coupling capacitors at low frequencies. High-frequency gain is limited by the bypassing effects of stray circuit capacitance and junction capacitances of the transistors. Therefore, regardless of the particular R and C values, the frequency response of any RC-coupled amplifier is the same with reference to a universal RC frequency-response curve. For example, Fig. 4-3 shows the low-frequency response curve for a single RC-coupling circuit. The percent of maximum amplitude is directly proportional to frequency, resistance, and capacitance.

Note that the frequency-response curve in Fig. 4-3 is shown in semilog coordinates. The reason for this is that low-frequency sweep generators commonly have a semilog time base; therefore, the curve that is displayed on the scope screen has the aspect shown in Fig. 4-3. The only difference between a semilog time base and a linear time base is that the left-hand region of the pattern is expanded, or "stretched out." This is desirable when the cutoff characteristic of an amplifier is comparatively steep, because the operator can read voltage values more accurately. Of course, the same data are displayed, whether a scope employs a linear time base or a semilog time base.

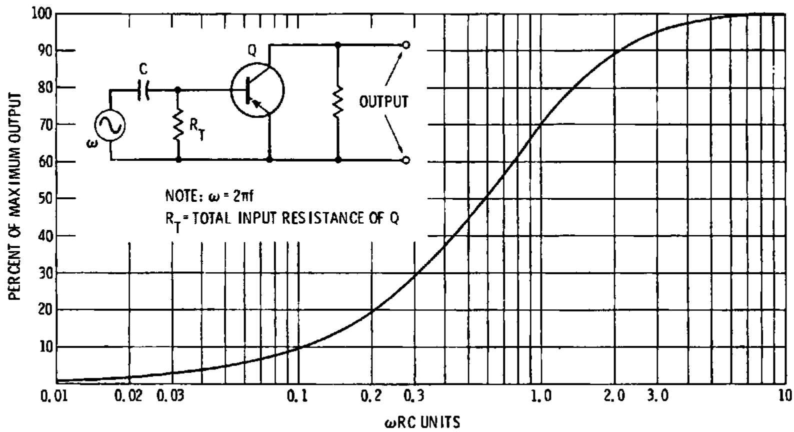


Fig. 4-3. Universal frequency-response curve for RC-coupling unit.

When two RC-coupled stages are connected in cascade, the output from the first stage is multiplied by the amplification of the second stage. In other words, the amplitude values depicted in Fig. 4-3 are squared by the second stage; this rule assumes that the same RC values are used in both the first and the second stages. Fig. 4-4 shows the low-frequency response curve for a two-stage RC-coupled amplifier. Note that Figs. 4-3 and 4-4 depict the low-frequency end of the frequency-response curves. The output rises to 100 percent through the midband region. At the high-frequency end of the frequency response, the output drops, due to stray capacitance and junction capacitance that shunts the collector load resistor. To measure the total output capacitance, we must use an impedance bridge.

If we know the value of the total output capacitance, and the output resistance (value of the collector load resistance in shunt with the transistor's collector-output resistance), we can find the frequency response at the high-frequency end of the range from the universal RC frequency-response curve given in Fig. 4-5. This is the high-frequency response curve for

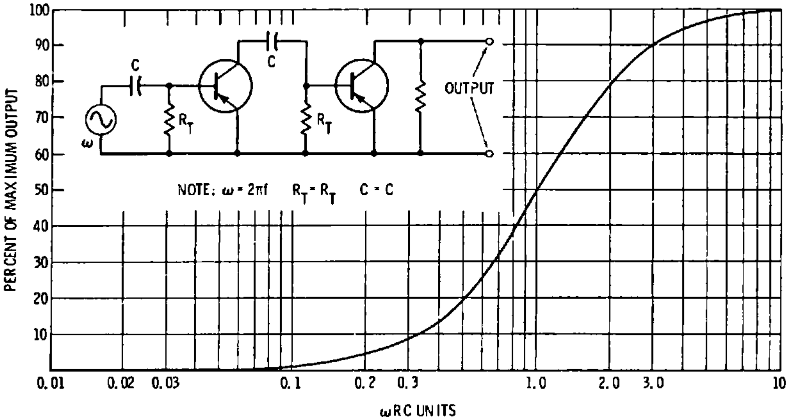


Fig. 4-4. Universal frequency-response curve for two-stage RC-coupled amplifier.

a single-stage amplifier. In most cases, we will not know the exact values of R_T and C_T. Therefore, we simply check the frequency response with a sweep generator and scope instead of calculating the frequency response. Nevertheless, it is instructive to know how the frequency response can be calculated. When two stages are connected in cascade, with the same type of transistors, and the same values of collector load resistance in each stage, the output from the first stage is multiplied by the amplification of the second stage. In other words, the output from a two-stage amplifier is equal to the square of the amplitudes depicted in Fig. 4-5. In turn, the two-stage output is

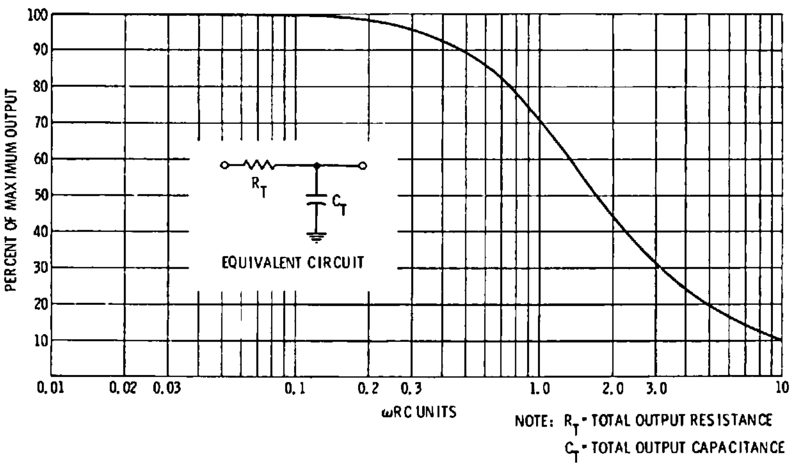


Fig. 4-5. Universal RC frequency-response curve for single-stage amplifier.

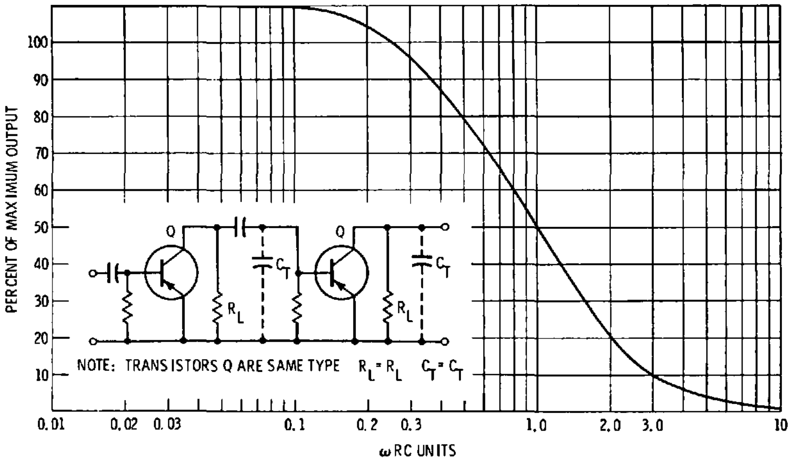


Fig. 4-6. Universal frequency-response curve for two-stage amplifier.

shown by the universal frequency-response curve given in Fig. 4-6.

Note in Fig. 4-6 that the R value in ωRC can be taken as equal to R_L, provided that the collector-output resistance and the base-input resistance of the transistor are comparatively high. In most amplifiers, the value of R_L is sufficiently high that we must calculate the value of R; in this case, the value of R is equal to the parallel combination of R_L, the base-input resistance of Q, and the collector-output resistance of Q. Of course, the total base-input resistance is equal to the parallel combination of the base return resistance and the input resistance looking into the base of the transistor. As before, C_T (Fig. 4-7) is the total shunt capacitance in the collector circuit, and is equal to the sum of the stray circuit capacitance, the collector junction capacitance, and the base junction capacitance.

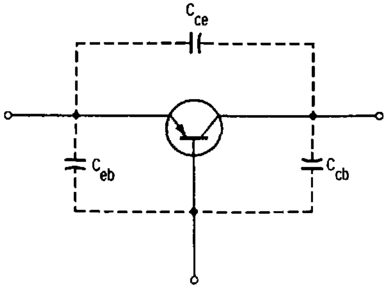
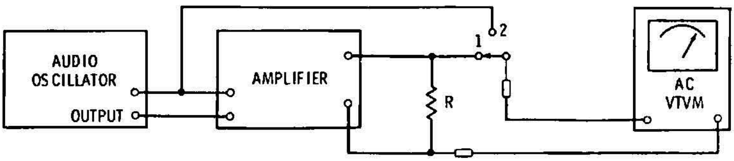


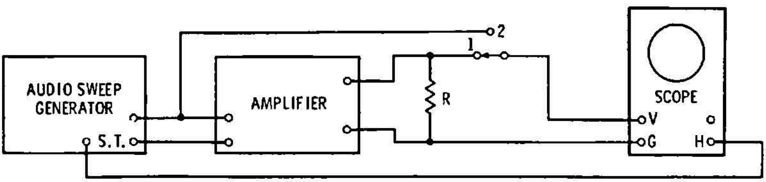
Fig. 4-7. Internal capacitances of a junction transistor.

Fig. 4-8 shows the frequency response of a typical audio amplifier from 20 Hz to 20 kHz. A low-frequency rise is present in this example, due to misadjustment of the low-frequency boost circuit.

The beginner should note carefully that the high-frequency response of an RC-coupled amplifier does not depend on the value of the coupling capacitor. In other words, a coupling capacitor is essentially a short-circuit at high frequencies of operation. Therefore, we could replace the coupling capacitors with short-circuits in Fig. 4-6, and the shape of the frequency-response curve would not be changed. On the other hand, the low-frequency response of an RC-coupled amplifier depends directly on the value of the coupling capacitor. That is, in Fig. 4-4, the low-frequency response is directly proportional to the time constant of C and R_T, where R_T is equal to the parallel combination of the base-return resistance and the input resistance looking into the base of the transistor.



(A) Test setup using audio oscillator and VTVM.



(B) Test setup using audio sweep generator and oscilloscope.

(C) Scope display of typical frequency-response pattern.

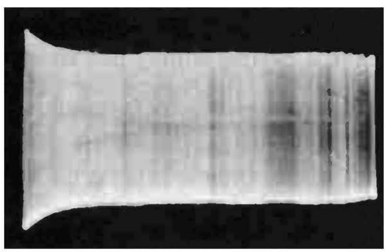
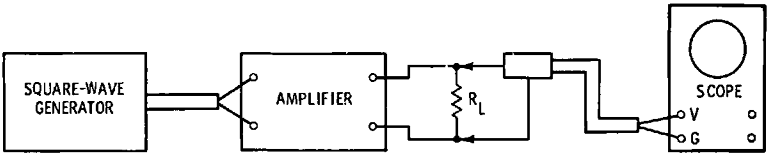


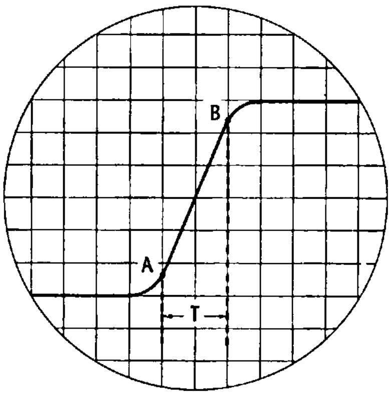
Fig. 4-8 Checking frequency response of audio amplifier.

SQUARE-WAVE RESPONSE OF RC-COUPLED AMPLIFIERS

The square-wave reproduction of a transistor amplifier is very informative. We can determine the frequency response, some phase-response characteristics, and the transient response by means of square-wave tests. For example, let us consider the measurement of high-frequency response. Fig. 4-9A depicts the test setup that is used. In any amplifier test procedure, it is essential to provide a load resistor R_L that has a resistance value equal to the rated speaker-load impedance. This



(A) Test setup.



(B) Waveform.

Fig. 4-9. Rise-time measurement.

load resistor must generally be of the power type, and should have a rating somewhat in excess of the maximum rated power output of the amplifier. The square-wave generator should have a rise time that is considerably faster than the rise time of the amplifier. Similarly, the scope should have a comparable rise time, plus a triggered-sweep function and calibrated time base.

As in frequency-response tests, it is desirable to check the square-wave response at maximum rated power output. If a calibrated scope is used, we check the power output in Fig. 4-9 as follows: Measure the peak voltage of the display square wave on the scope screen. Then, the output power is formulated:

$$P_o = \frac{E_p^2}{R_L} \text{ watts} \tag{4.1}$$

Note that the peak voltage is equivalent to a dc voltage. Since the power dissipated in R_L is equal to E^2/R_L with respect to dc, it follows that Formula (4.1) gives the output power supplied by the amplifier. Next, the scope controls are set to expand the leading edge of the reproduced square wave, as depicted in Fig. 4-9B. The high-frequency cutoff point is related to the rise time from A to B according to the formula:

$$f_{co} = \frac{0.35}{T_r} \tag{4.2}$$

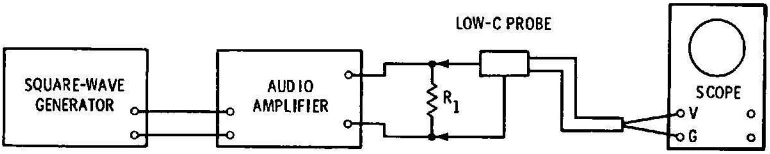
If the rise time T_r is measured in milliseconds, f_{co} will be given in kHz. Although Formula (4.2) is approximate, it will be found to check quite closely with a measurement of f_{co} with an audio-oscillator or an audio sweep-generator test. A square-wave test is much faster than an audio-oscillator test; it is also more practical than an audio sweep-generator test, since this type of instrument is usually available only in laboratories.

Next, let us see how the low-frequency response of an amplifier is measured in a square-wave test. The test setup is shown in Fig. 4-10A. The square-wave generator need not have particularly fast rise, and an ordinary scope can be used, provided that its vertical amplifier has good 60-Hz square-wave response. We reduce the repetition rate of the square-wave signal until the reproduced square wave displays appreciable tilt, as illustrated in Fig. 4-10B. Then, we calculate the percentage of tilt according to the formula:

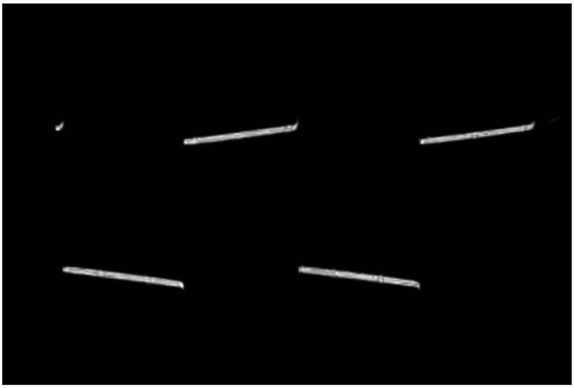
$$\%\ \text{Tilt} = \frac{E_2 - E_1}{E_2} \times 100 \tag{4.3}$$

The terms E_2 and E_1 in Formula (4.3) are determined simply by counting squares on the scope screen, to measure the relative amplitudes depicted in Fig. 4-10C. It is often sufficient merely to measure the percentage of tilt and to note the corresponding repetition rate (frequency) of the square wave. Amplifier manufacturers often rate their products for maximum tilt at a reference square-wave frequency, such as 20 Hz. If we wish to calculate the cutoff frequency, we employ the following formula:

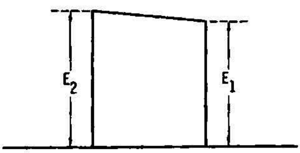
$$f_{co} = \frac{2f\,(E_2 - E_1)}{3\,(E_2 + E_1)} \tag{4.4}$$



(A) Test setup.



(B) Waveform.



(C) Measurement of tilt.

Fig. 4-10. Square-wave tilt at low repetition rates.

The low-frequency cutoff point f_{co} is, of course, the frequency at which the amplifier response is down 3 dB. In Formula (4.4), f denotes the frequency of the square-wave signal, and E_2 and E_1 are the amplitudes depicted in Fig. 4-10C. Again, this is an approximate formula, but it will be found to check quite well with an audio-oscillator test of an audio sweep-generator test. Formula (4.4) is based on the relation of frequency response to phase shift. In other words, at the center frequency of an amplifier, the output is maximum, and the phase shift is zero. On the other hand, at low frequencies, the amplifier output decreases and a leading signal current is drawn, as depicted in Fig. 4-11. Note that f denotes the center frequency, and f_1 denotes any lower frequency that we might choose. We observe that when f_1 is 0.01 of f, the phase shift is 85 degrees.

Phase shift at low frequencies produces tilt in a reproduced square wave. If the phase shift is not too great, no curvature is visible in the tilted top of the displayed square wave. However, when a large amount of phase shift is present, the accompanying low-frequency attenuation produces noticeable curvature in the tilted top, and we say that the amplifier is differentiating the square wave. In tilt tests, we usually work with percentages of tilt in the order of 10 to 15 percent. We will find that an amplifier develops appreciable phase shift before the frequency-response curve has dropped off substantially. Therefore, tilt tests provide easily made measurements of amplifier characteristics.

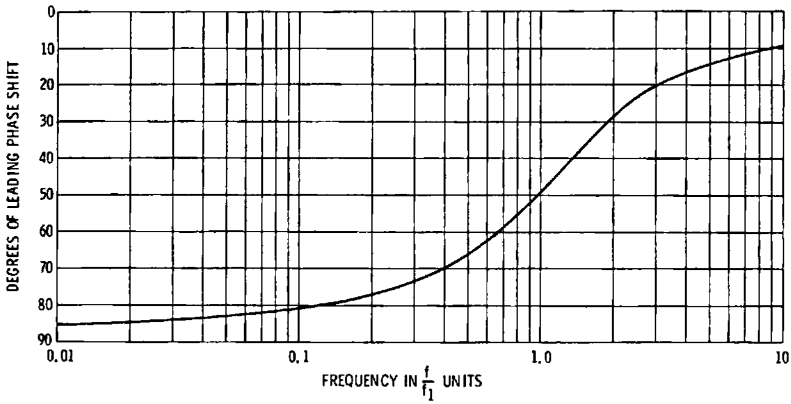
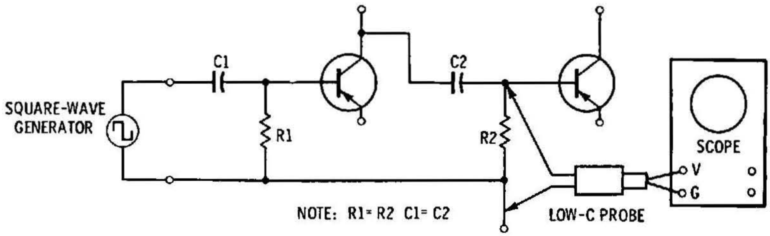


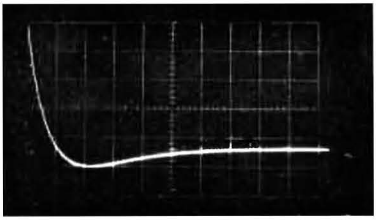
Fig. 4-11. Universal phase-shift chart for RC-coupled amplifier at low frequencies.

When the repetition rate of the square-wave generator is slowed down to the point that a large amount of tilt is developed, the response of a single-stage transistor amplifier is the same as that of an RC differentiating circuit. In the case of a two-stage amplifier, the square-wave response is as illustrated in Fig. 4-12. As previously explained for frequency-response curves, the time constant of the coupling circuit is equal to RC, wherein C is the value of the coupling capacitor, and R is equal to the equivalent resistance of R1 and the base input resistance of the transistor in Fig. 4-12A. Since the resistance calculation is not always easy, it is more practical to make a test of the low-frequency square-wave response.

At high square-wave frequencies, the leading edge of the reproduced square wave can be expanded as shown in Fig. 4-13. If the scope has a calibrated time base, we can measure the



(A) Test setup.



(B) Low-frequency square-wave response.

Fig. 4-12. Square-wave test of two-stage RC-coupled amplifier.

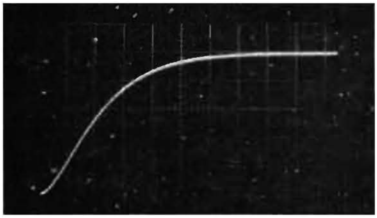
rise time of the amplifier. In turn, the high-frequency cutoff point of the amplifier is formulated:

$$f_{co} = \frac{0.35}{T_r} \qquad (4.5)$$

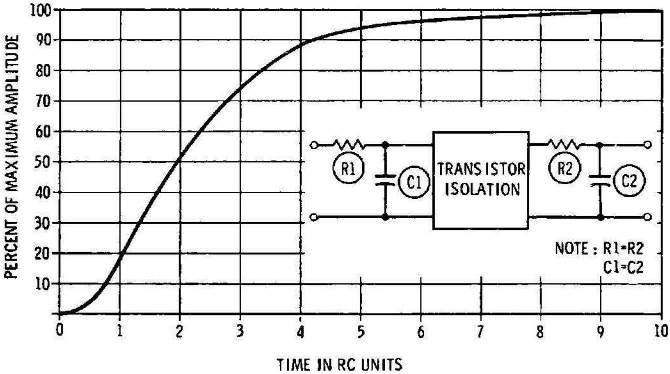
where,

f_{co} denotes the frequency at which the response is down 3 dB, and T_r is the rise time of the amplifier.

Formula (4.5) can be applied to an amplifier with any number of stages. Of course, the shape of the reproduced square wave changes somewhat when stages are cascaded. That is, a single-stage transistor amplifier has the same high-frequency square-wave response as that of an RC-integrating circuit. In the case of a two-stage amplifier, the leading edge has a somewhat different shape, as seen in Fig. 4-13. As a greater number of stages are cascaded, the delay time of the amplifier increases. The delay time is defined as the time required for the reproduced square wave to rise to 50 percent of its maximum value, as depicted in Fig. 4-14. A single-stage amplifier has a delay time that is equal to 0.7 of an RC unit. On the other hand, a two-stage amplifier has a delay time of 2 RC units. A three-stage amplifier has a delay time of approximately 4.5 RC units. These figures are for symmetrical amplifiers (amplifiers with identical stages).



(A) Universal RC time-constant chart.



(B) High-frequency square-wave response.

Fig. 4-13. High-frequency response of a two-stage RC-coupled amplifier.

A transistor provides considerable isolation between its input and output circuits. As previously explained, this isolation is not perfect because there is a certain amount of transfer impedance from input to output, and vice versa. Some transistors, such as FET types, have negligible transfer impedance. However, regardless of the transistor type, the universal RC time-constant charts that have been discussed are sufficiently accurate for all practical purposes. Note in Fig. 4-13 that the values of R and C in the time constant are not easy to calculate. The value of R is equal to the equivalent resistance of the collector load resistance, the collector output resistance, the base

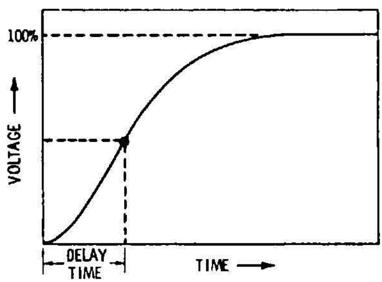


Fig. 4-14. Measurement of delay time.

resistance, and the base input resistance. Similarly, the value of C is equal to the equivalent capacitance of the stray circuit capacitance, the collector output capacitance, and the base input capacitance.

The delay time of an audio amplifier is of no consequence, because it does not affect the quality of sound reproduction noticeably. However, when we consider video amplifiers we will find that delay times are of great importance in obtaining good color picture reproduction. Therefore, it is helpful to know the meaning of delay time and how to measure it. When delay times must be equalized in an amplifier system, a suitable delay line is connected in series with the amplifier that has the shortest delay time.

LOW-FREQUENCY BOOST CIRCUIT

A transistor amplifier contains coupling capacitors, bypass capacitors, and decoupling capacitors. In turn, the system tends to differentiate low-frequency square waves. When good low-frequency square-wave response is desired, a low-frequency boost circuit is often used, as shown in Fig. 4-15A. The boost circuit comprises a boost resistor R1 connected in series with a load resistor, and a boost capacitor C1 shunted across R1. If the values of R1 and C1 are properly selected, the low-frequency square-wave response of the amplifier can be greatly improved. However, perfect compensation cannot be obtained, and there is a practical limit to the amount that the low-frequency response can be extended.

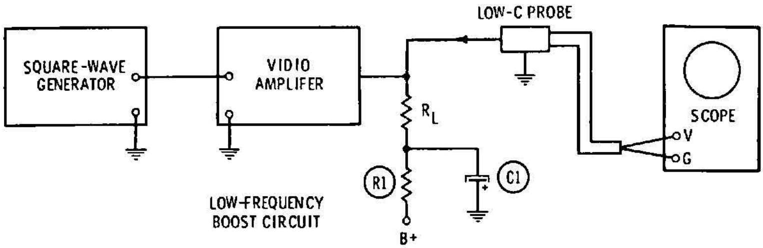
With reference to Fig. 4-15B, the photographs illustrate reproduced 60-Hz square waves. We observe that when no low-frequency boost circuit is used, the square wave is considerably differentiated. When various values of R1 and C1 are connected into the boost circuit, both the curvature and tilt of the reproduced square wave are changed. The top of the square wave may be either convex or concave, and the tilt may be either uphill or downhill. In this example, the best 60-Hz square-wave response is obtained when R1 is 15k and C1 is 2μF. Note that the compensation is not perfect, and there is a small amount of overshoot.

Some basic square-wave distortions are depicted in Fig. 4-16. Let us consider the circuit characteristics that are associated with these distortions:

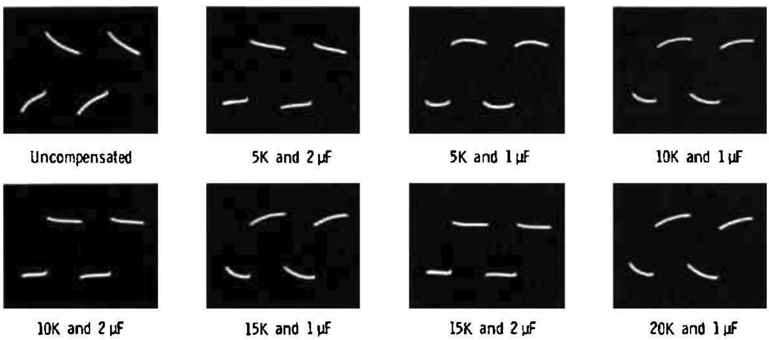
1. Downhill tilt; phase is leading at low frequencies. This means that the low-frequency components of the square

wave lead the high-frequency components. A corresponding phase characteristic is depicted in Fig. 4-14.

2. Concave top; fundamental frequency attenuated. This means that there is low-frequency attenuation without phase shift. Certain R and C values in a low-frequency boost circuit can produce this type of distortion.



(A) Test setup.



(B) Waveforms obtained with various values of R1 and C1.

Fig. 4-15. Square wave test with low-frequency boost circuit.

3. Concave top with downhill tilt; low-frequency attenuation with leading phase shift. This situation is very common, and is encountered in familiar RC differentiating circuits.
4. Convex top; fundamental frequency boosted. This means that there is low-frequency boost without phase shift. An example is seen in Fig. 4-15, when R1 is 5k and C1 is $1 \mu F$.
5. Rounded corners. This means that there is high-frequency attenuation present without phase shift. We observe this type of square-wave distortion in an RC-coupled amplifier that has a slow drop-off at high frequencies.
6. Diagonal corner rounding. This means that high-frequency attenuation is present with significant phase shift. This type of distortion is observed in RC-coupled ampli-

fiers that have a rather abrupt drop-off at high frequencies.

7. Overshoot (may be accompanied by ringing). This means that the amplifier has a rising response at high-frequencies, a very abrupt drop-off at high frequencies, or both. Both conditions are associated with a rapidly changing phase characteristic.

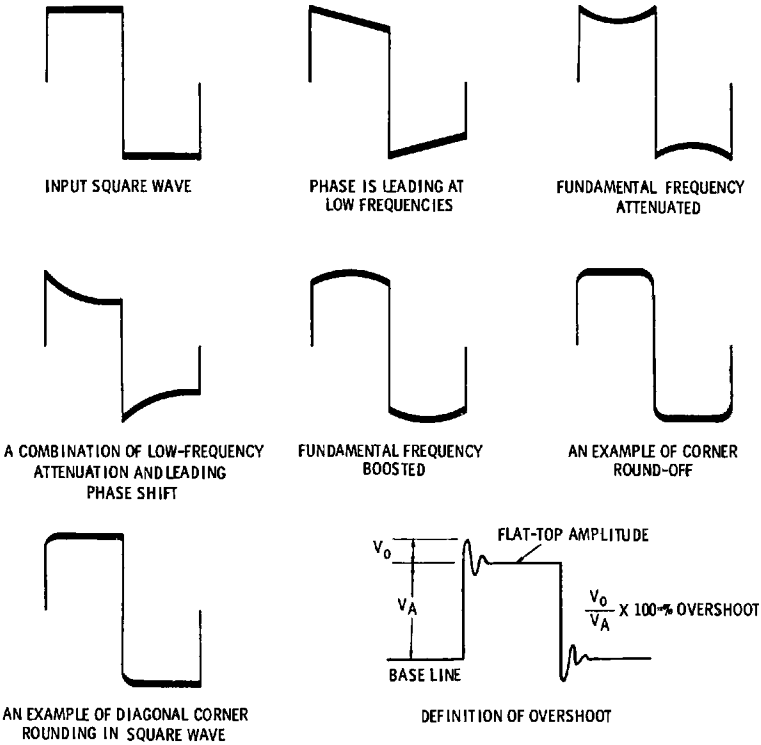


Fig. 4-16. Some basic square-wave distortions.

APPROXIMATE MEASUREMENT OF RISE TIME

Sometimes we need to make a measurement of rise time, although a scope with calibrated sweeps is unavailable. An approximate measurement of rise time can be made with a simple differentiating circuit, as described in Fig. 4-17. The scope must have ample bandwidth, but this is the only requirement. In this example, the rise time of the output waveform from a square-wave generator is being checked. To make the rise-time measurement, we vary the value of R or C in the differentiating circuit to obtain a differentiated pulse that has 65 percent of the amplitude of the undifferentiated square wave. Then, the time constant of the differentiating circuit is approximately equal to the rise time of the square wave.

This method of rise-time measurement is based on the fact that since the square wave has a certain rise time, capacitor C starts to discharge through R before it is fully charged. Therefore, when the RC time constant is short, the applied square wave cannot charge C to the peak voltage of the square wave. It can be shown that when the RC time constant of the differentiating circuit is equal to the rise time of the square wave, the pulse output will have approximately 65 percent of the square-wave amplitude. For accurate measurements, it is necessary to minimize the effect of stray circuit capacitance. If R has a comparatively low value, such as 75 ohms, the effect of stray capacitance will be negligible.

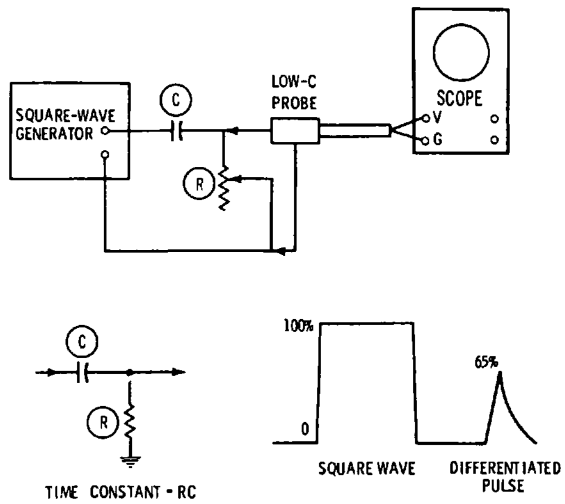


Fig. 4-17. Rise-time measurement with differentiating circuit.

REVIEW QUESTIONS

1. Describe the chief features of an RC-coupled amplifier.
2. Explain why the total base-input resistance of a transistor is less than the value of the base-return resistor.
3. Define the efficiency of an RC-coupled amplifier.
4. Why does the response curve drop off at low frequencies? At high frequencies?

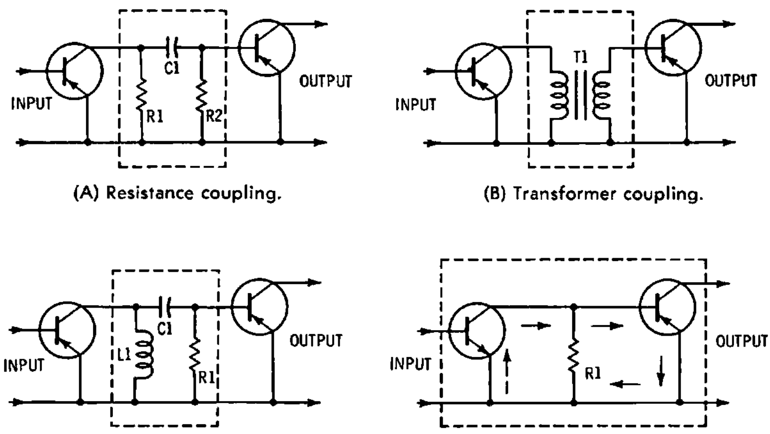
5. How is the low-frequency cutoff point related to square-wave tilt?
6. How is the high-frequency cutoff point related to square-wave rise time?
7. State two ways of checking the frequency-response curve of an amplifier.
8. Explain how amplifier output power is calculated with respect to a square-wave signal.
9. Why does phase shift produce square-wave tilt?
10. Define delay time.
11. What is the delay time of a single-stage RC-coupled amplifier?
12. What is the delay time of a two-stage symmetrical RC-coupled amplifier?
13. What is the approximate delay time of a three-stage symmetrical RC-coupled amplifier?
14. Describe the function of a low-frequency boost circuit.
15. State the cause of downhill tilt in a reproduced square wave.
16. When a concave top is observed in a reproduced square wave, what circuit condition could be the cause?
17. Name a simple circuit that produces a concave top and downhill tilt in a square wave.
18. What circuit condition could produce a convex top in a square wave?
19. State an amplifier characteristic that produces rounded corners in a square wave.
20. Discuss the cause of diagonal corner rounding in a square wave.
21. Describe an amplifier characteristic that can produce overshoot in a square wave.
22. What basic square-wave distortions may be produced by misadjustment of a low-frequency boost circuit? Explain.
23. How can rise time be approximately measured with a differentiating circuit?
24. Why is the peak voltage of a square wave attenuated by a differentiating circuit that has a sufficiently short time constant?
25. How can stray capacitance effects be minimized in Problem 23?

5

Transformer-Coupled, Impedance-Coupled, and Direct-Coupled Amplifiers

Interstage coupling with a transformer is depicted in Fig. 5-1B. The primary winding of T1 (including the reflected ac load from the secondary) is the collector load impedance of the first stage. The secondary winding applies the ac signal to the base and also serves as the base dc return path. This very low resistance in the base path aids in temperature stabilization of the dc operating point. If an emitter bias resistor is included, the current stability factor will be very good. Because there is no collector load resistor to dissipate I^2R power, the power efficiency of a transformer-coupled amplifier approaches the theoretical maximum available gain (MAG) of 50 percent. For this reason, transformer-coupled amplifiers are used extensively in portable radio receivers and other electronic devices that employ battery power.

Transformers facilitate the matching of the output impedance of the driver transistor to the base-input impedance of the driven transistor. Matched impedances are required to realize the maximum available gain of the amplifier. The frequency response of a transformer-coupled stage is not as good as that of a well-designed RC-coupled stage. That is, the shunt reactance of the primary winding at low frequencies drops off, and causes the low-frequency response to drop off. At high frequencies, the response is reduced by the collector junction capacitance, stray capacitance, and leakage reactance between primary and secondary windings.



(A) Resistance coupling.

(B) Transformer coupling.

(C) Impedance coupling.

(D) Direct coupling.

Fig. 5-1. Interstage coupling networks.

Fig. 5-2 shows the basic equivalent circuit for an untuned transformer, such as an audio-frequency transformer. In an ideal transformer, the entire magnetic field of the primary would pass through the secondary; that is, all of the transformer inductance would be in the form of mutual inductance. However, it is impossible to design a transformer that has 100 percent coupling between primary and secondary. Inevitably, some lines of magnetic flux from the primary escape into surrounding space and do not cut the secondary turns. Again, some of the primary flux lines cut only a fraction of the secondary turns. Because of this lack of perfect coupling, leakage reactances $L_p - L_m$ and $L_s - L_m$ are effectively connected in series with the mutual inductance. There is, in turn, an IX drop across these leakage reactances, which reduces the output voltage from the transformer when it is loaded by the base of a transistor.

We will find that a practical transformer also has distributed capacitance. In other words, $L_p - L_m$, L_m, and $L_s - L_m$ in Fig. 5-2 are shunted by distributed capacitance. This capacitance consists of the sum of the capacitances from each turn

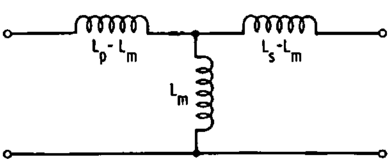


Fig. 5-2. Basic equivalent circuit for untuned transformer.

to the next. The result is that although the transformer is designed as an untuned transformer, it turns out in practice that it is a broadly tuned transformer. The result of resonance in an "untuned" transformer produces characteristic frequency-response curves that may have a rising response at high frequencies, as illustrated in Fig. 5-3. This rising response is often helpful in obtaining the required bandwidth—from 60 to 7500 Hz in this example. On the other hand, nonuniform frequency response produces distortion. Therefore, audio transformer design is generally a compromise between wide-band response and tolerable frequency distortion.

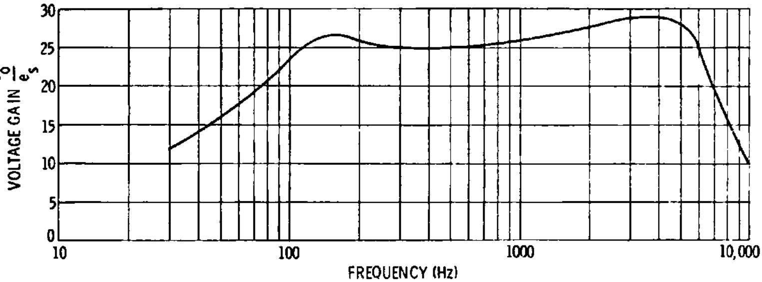
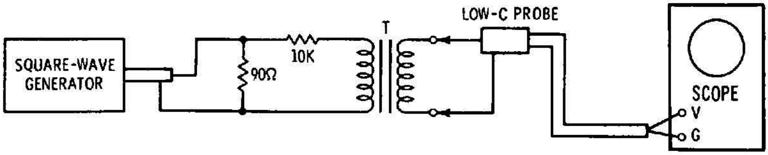


Fig. 5-3. Voltage gain curve for transformer-coupled voltage amplifier.

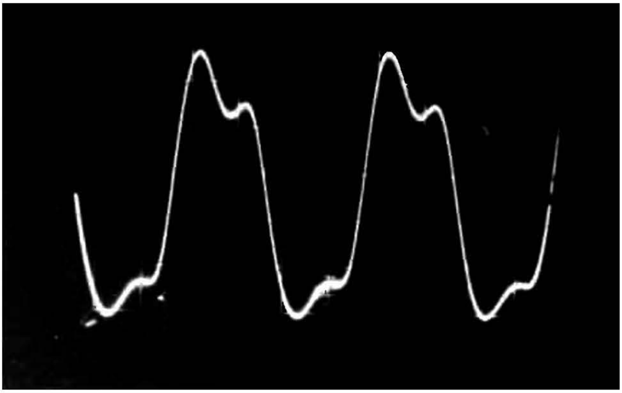
Beginners should note that rising frequency response can be controlled by resistance loading. To put it another way, if a sufficiently low resistance is connected across the secondary, the effects of distributed capacitance are "swamped out," and a comparatively flat frequency response can be obtained. However, this expedient also causes bandwidth reduction. We will find that an output transformer does not have a rising frequency response because a speaker has a sufficiently low impedance that the secondary is loaded heavily and the effects of distributed capacitance are swamped out.

TRANSIENT RESPONSE OF UNTUNED TRANSFORMER

The square-wave response of an audio transformer depends greatly on the loading of primary and secondary. Shunt capacitance shifts the resonant frequency of the leakage reactance and distributed capacitance. Shunt resistance tends to damp out ringing, and reduces overshoot. A typical test setup is shown in Fig. 5-4A. The primary is damped by a resistance of approximately 10,000 ohms; the secondary is shunted by a capacitance of 10 pF. Considerable square-wave distortion is



(A) Test setup.



(B) Response to 2.5-kHz square wave.

Fig. 5-4. Square-wave test of audio transformer.

produced, as illustrated in Fig. 5-4B. If resistance is shunted across the secondary, the top of the square wave can be flattened. For example, a 100,000-ohm resistor connected across the secondary makes the reproduced square wave essentially flat in this example. However, a ripple then appears along the top, due to residual ringing at a comparatively high frequency.

If the primary in Fig. 5-4A is damped by a higher value of resistance, the reproduced square wave becomes noticeably integrated. On the other hand, small values of primary damping resistance correspond to increased ringing amplitude. If another type of transformer is used in the same test setup, quite different transient response is observed. Therefore, a coupling transformer should be designed for the particular circuit impedances present in the amplifier. A transformer that provides good frequency and transient response in one amplifier may produce excessive frequency and transient distortion in another amplifier.

Fig. 5-5 shows a single-ended driver transistor transformer-coupled to a pair of push-pull output transistors. Transformer coupling is also used from the push-pull stage to the speaker. A step-down interstage transformer is used to match the comparatively high collector impedance of Q6 to the low base impedances of Q7 and Q8. Each transistor is forward-biased; Q6 operates essentially in class A, while Q7 and Q8 operate in class AB, due to increased signal-current drive to the push-pull stage. It is impractical to bias Q7 and Q8 to collector-current cutoff, and thereby operate the transistors in class B. Crossover distortion would be excessive, due to the highly nonlinear base characteristic in the vicinity of cutoff. Fig. 5-6 illustrates the waveform that results from crossover distortion.

Another difficulty in operating transistors in class B with transformer coupling is the large change in base impedance

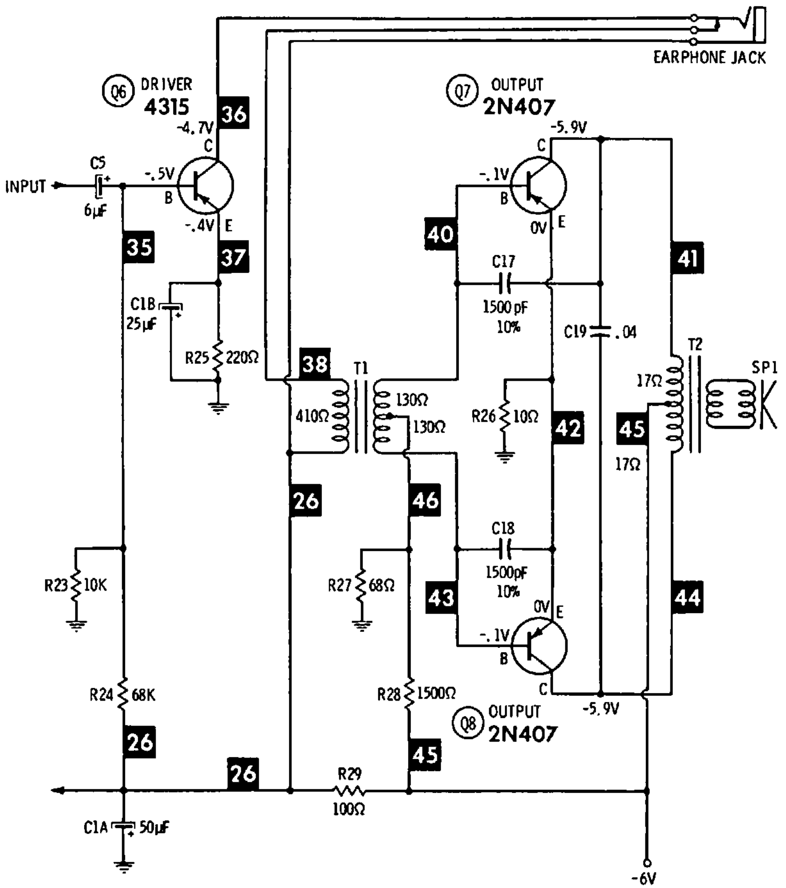


Fig. 5-5. Transformer-coupled audio amplifier.

from small-signal to large-signal conditions. If the secondary of T1 in Fig. 5-5 is unloaded for very small signal levels, lightly loaded for medium signal levels, and heavily loaded for strong signal levels, it becomes very difficult to design the transformer for reasonable fidelity. Because of the foregoing problems, it is more practical to sacrifice some efficiency for improved fidelity by operating the push-pull stage in class AB. Although class-AB operation produces second-harmonic distortion from each of the push-pull transistors, this harmonic distortion is cancelled out. Third-harmonic distortion is minimized by use of negative feedback via capacitors C17 and C18. Capacitor C19 provides capacitive loading for the primary of T2 to optimize fidelity.

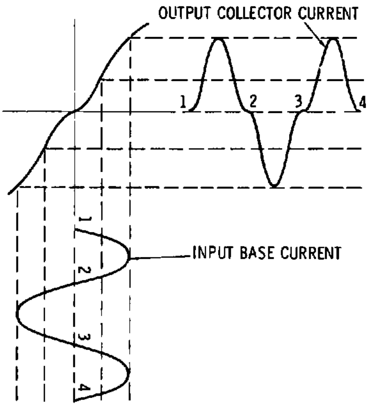
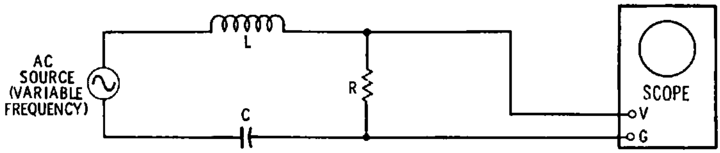


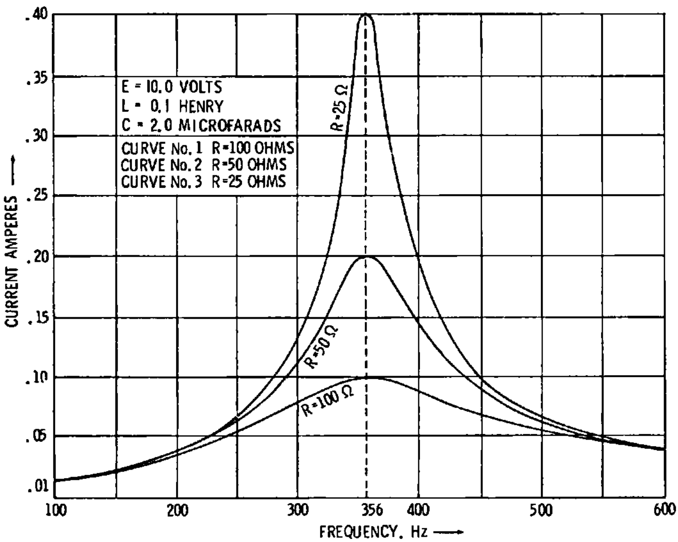
Fig. 5-6. Dynamic transfer characteristic curve of class-B, push-pull amplifier.

In the example of Fig. 5-5, negative feedback increases at higher audio frequencies, due to decreasing capacitive reactance. Thereby, rising high-frequency response is eliminated, and improved bandwidth is obtained. Of course, this improved frequency response is obtained at the expense of gain. R26 is a common-emitter resistor that provides temperature stabilization. If the output transistors should heat up, their collector currents would increase. This increased current flow would return through R26, and would produce a corresponding voltage drop that would reduce the base-emitter bias voltages on Q7 and Q8. In this manner the operating point is stabilized.

A scope is the most useful instrument for tracing a signal through an audio amplifier, and for localizing objectionable distortion. For example, if one of the transistors in the push-pull stage of Fig. 5-5 has excessive collector leakage, its gain is reduced. In turn, the waveform across one half of the primary winding on T2 has less amplitude than the waveform across the other half of the winding. That is, the stage develops second-harmonic distortion under this condition. An audio oscillator should be used in troubleshooting an amplifier because it provides a steady signal with a true sine waveform.



(A) Test setup.



(B) Response curves.

Fig. 5-7. Testing response of LCR series-resonant circuit.

TUNED TRANSFORMER

Tuned transformers are used in all transistor radio and TV receivers, and in many other types of electronic equipment. Tuned-transformer action is based on the resonant response of series and parallel LCR circuits that have mutual inductance. A series-resonant circuit has frequency-response curves such as those in Fig. 5-7. In this test setup, the scope displays the current variation vs. frequency. The resonant frequency of the circuit is formulated:

$$f_o = \frac{1}{2\pi\sqrt{LC}} \qquad (5.1)$$

Since capacitive and inductive reactance cancel at resonance, the current flow is given by:

$$I = \frac{E}{R} \qquad (5.2)$$

In turn, the bandwidth is given by the number of hertz between the −3-dB, or half-power points:

$$BW = f_2 - f_1 \qquad (5.3)$$

where,

f_2 denotes the frequency at the higher half-power point (70.7 percent of maximum voltage point),
f_1 denotes the frequency at the lower half-power point.

The Q value of the series-resonant circuit is equal to the inductive reactance divided by the resistance of the circuit:

$$Q = \frac{X_L}{R} \qquad (5.4)$$

$$X_L = 2\pi fL \qquad (5.5)$$

Note that the Q value varies with frequency; at resonance, we write:

$$X_{Lo} = 2\pi f_o L \qquad (5.6)$$

$$Q_o = \frac{X_{Lo}}{R} \qquad (5.7)$$

The bandwidth of a series-resonant LCR circuit is given to a practical approximation by:

$$BW = \frac{f_o}{Q_o} \qquad (5.8)$$

Observe, also, that R is usually greater than the dc resistance in the circuit, being equal to the rf resistance; R_{ac} increases with frequency. Therefore, a coil in a series-resonant circuit usually has a maximum Q value at some particular operating frequency.

A parallel-resonant circuit has frequency-response curves such as those in Fig. 5-8. The resonant frequency is given approximately by Formula (5.1). The impedance of a parallel-resonant circuit is approximately equal to:

$$Z_0 = \frac{L}{RC} \tag{5.9}$$

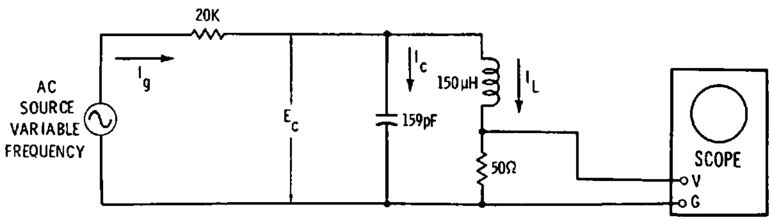
where,

Z_0 denotes the impedance at resonance in ohms,
L denotes the inductance in henrys,
C denotes the capacitance in farads,
and R denotes the effective ac resistance of the circuit.

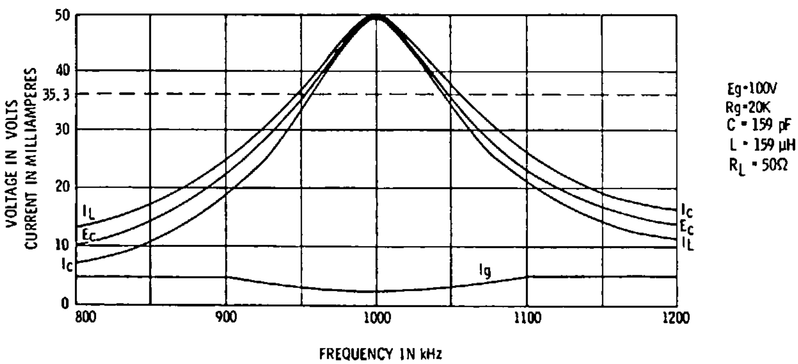
Response curves for inductor current, capacitor current, and capacitor voltage are approximately the same. The generator current has an "upside-down" response curve that is basically similar to the other response curves. In Fig. 5-8, the 20,000-ohm resistor represents the source resistance, such as the collector output resistance of a transistor. At resonance, the circulating current between L and C is equal to Q times the generator current I_g. That is:

$$I_C = I_L = QI_g \tag{5.10}$$

This is called the current amplification of the LCR circuit. It is analogous to the voltage amplification of a series-resonant



(A) Test setup.



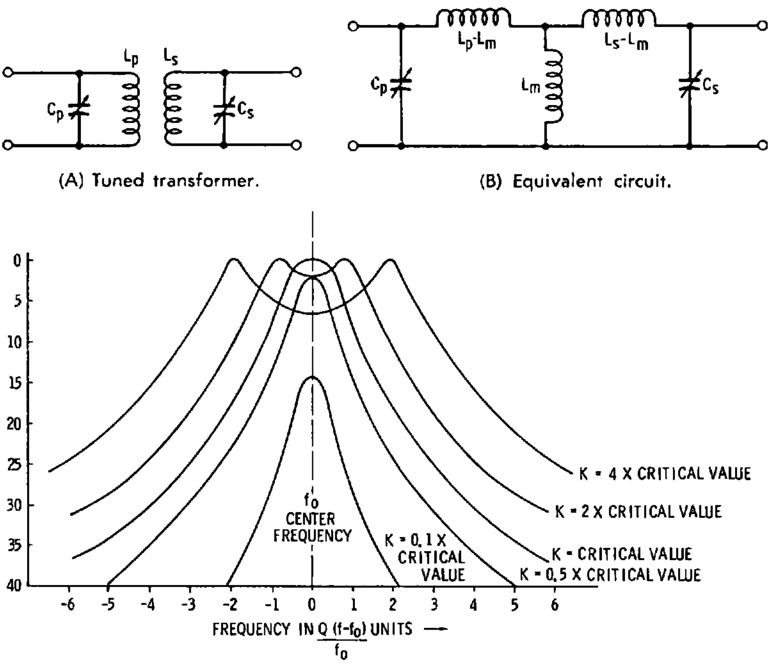
(B) Response curves.

Fig. 5-8. Testing response of LCR parallel-resonant circuit.

circuit; in a series LCR circuit, the generator voltage is stepped up Q times by the inductor and by the capacitor (Fig. 5-7):

$$E_C = E_L = QE_g \qquad (5.11)$$

In other words, resonant circuits, in the absence of transistors, can produce either voltage amplification or current amplification, but not both. A transistor can provide voltage amplification and current amplification simultaneously. This is because a transistor is an active device, whereas a resonant circuit is a passive device. There can be no power gain in a passive device.

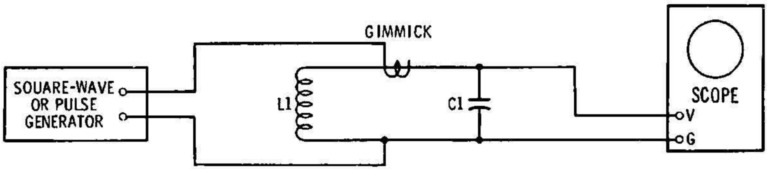


(C) Frequency-response curves for various coefficients of coupling (k).

Fig. 5-9. Bandwidth in two tuned, coupled circuits.

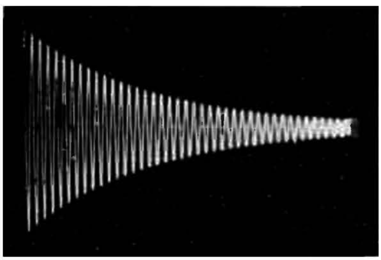
Tuned transformers (Fig. 5-9) comprise a parallel-resonant circuit and a series-resonant circuit with mutual inductance. The primary operates as a parallel-resonant circuit; the secondary operates as a series-resonant circuit. Since the primary and secondary are coupled, the series-resonant circuit reflects an impedance into the parallel-resonant circuit that varies with frequency. A preliminary equivalent circuit for a tuned transformer is shown in Fig. 5-9B. If the primary and secondary have the same Q values (as is usually the case), and are tuned to the same center frequency, the output voltage from the secondary displays frequency-response curves such as shown in Fig. 5-9C. The primary or secondary alone has a ringing waveform in a square-wave test as shown in Fig. 5-10. A tuned transformer has a ringing waveform such as that illustrated in Fig. 5-11.

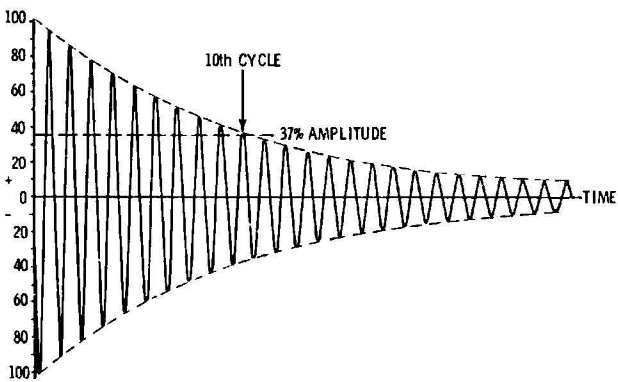
Note in Fig. 5-9 that the bandwidth of the frequency-response curve depends on the coefficient of coupling. In ordinary tuned transformers, the primary and secondary have the same



(A) Test setup.

(B) Waveform.





(C) Q equals 10π, or 31.4, in this example.

Fig. 5-10. Ringing test of primary or secondary of transformer.

inductance and the same Q values, and are tuned to the same center frequency. In turn, the maximum possible value of mutual inductance L_m is formulated:

$$L_{m(max)} = L_p = L_s \qquad (5.12)$$

The coefficient of coupling is the ratio of actual mutual inductance L_m to the value of $L_{m(max)}$:

$$k = \frac{L_m}{L_p} = \frac{L_m}{L_s} \qquad (5.13)$$

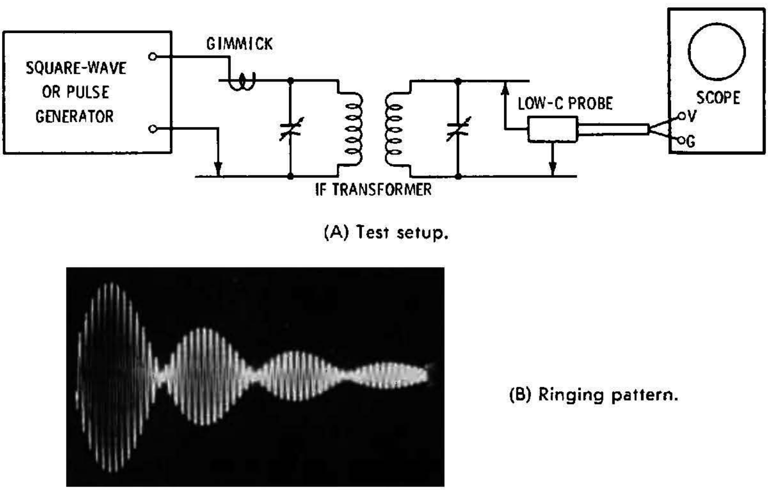
At critical coupling, the frequency-response curve has a single peak. At any greater value of coupling, the response curve has two peaks. The value of critical coupling is formulated:

$$k_c = \frac{1}{Q_p} = \frac{1}{Q_s} \qquad (5.14)$$

Critical coupling depends on the value of ac resistance. At critical coupling, it can be shown that:

$$\omega L_m = R_p = R_s \qquad (5.15)$$

It follows from Fig. 5-9B that a tuned transformer has two resonant frequencies, which correspond to the double humps in Fig. 5-9C. That is, the signal current branches in Fig. 5-9B, and the branch currents have different resonant frequencies.



(A) Test setup.

(B) Ringing pattern.

Fig. 5-11. Ringing test of tuned transformer.

The hump frequencies are formulated:

$$f_h = \frac{f_c}{\sqrt{1 \pm k}} \qquad (5.16)$$

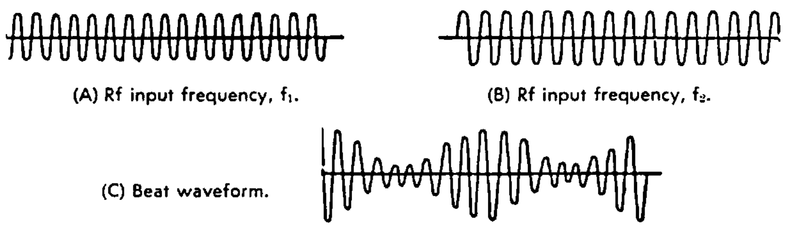
where,

f_h denotes the hump frequencies,
f_c denotes the center frequency,
and k denotes the coefficient of coupling. (The $\pm$ signs for k give the two answers.)

Since there are two hump frequencies present, these frequencies beat in a square-wave test, as depicted in Fig. 5-12. If the primary and secondary are tuned to the same center frequency, the beat waveform goes periodically through zero beat, as illustrated in Fig. 5-11. The beat waveform decays exponentially because of the I^2R losses in the primary and secondary. The frequency of the beat envelope is given by Formula (5.16). Therefore, a high-Q tuned transformer has a large number of cycles between zero-beat intervals. The ringing oscillation in Fig. 5-11 is equal to the average of the hump frequencies (it is the center frequency):

$$f_r = \frac{f_1 + f_2}{2} \qquad (5.17)$$

Since a tuned transformer always has two hump frequencies, the question arises concerning the display of a single peak at critical coupling in Fig. 5-9. The answer is that the two hump frequencies are so close together in this situation that they seem to merge and form a single peak on the scope screen.



(A) Rf input frequency, f_1.

(B) Rf input frequency, f_2.

(C) Beat waveform.

Fig. 5-12. Formation of beat waveform.

IMPEDANCE-COUPLED AMPLIFIERS

Transistor impedance-coupled amplifiers are widely used in TV receivers and other electronic equipment. A simple impedance-coupled amplifier is diagrammed in Fig. 5-1C. The collector load is an impedance, because the inductor has winding resistance; the inductor is also shunted by R1 and the base-input resistance of the second transistor. Analysis of this amplifier is basically the same as that of an RC-coupled amplifier, except that the collector load impedance is formulated:

$$Z = \sqrt{R^2 + X_L^2} \qquad (5.18)$$

where,

Z is the load impedance in ohms,
R is the effective resistance in ohms,
and X_L is the inductive reactance in ohms.

$$X_L = 2\pi fL \qquad (5.5)$$

Fig. 5-13 shows the configuration for a typical transistor video amplifier. This is an impedance-coupled amplifier. The impedance loads comprise inductors (peaking coils) L211, L212, L214, and L213. We call L211 and L212 shunt peaking coils. L213 and L214 are called series peaking coils. These peaking coils provide high-frequency compensation, and flat frequency response from 60 Hz to approximately 3.5 MHz. A video-frequency response curve is illustrated in Fig. 5-14.

Fig. 5-15A shows a simple resistance load, as used in a conventional RC-coupled amplifier. In B, L1 provides shunt peaking. In C, L2 provides series peaking. Of course, series peaking can be used with shunt peaking as in D. It is impractical to design a wide-band amplifier that does not have some form of peaking, because the resistive load must be so small that little gain is obtained. On the other hand, if a shunt peaking coil is used, the load resistor can have a fairly high value, and good gain is obtained over a wide frequency range. If we use series peaking instead of shunt peaking, 50 percent more gain can be obtained. Again, if we use both series and shunt peaking, we can obtain about 80 percent more gain than with shunt peaking alone. In Fig. 5-14, the amplifier is slightly over-compensated, since the response tends to rise at high frequencies.

Fig. 5-16 shows frequency-response curves for various values of shunt peaking. The gain at low frequencies is determined by the value of the load resistor. High-frequency gain is determined by the peaking inductance. Note that when the load is resistive only, the frequency response in this example starts to drop off at 700 kHz, but the use of a suitable shunt peaking coil maintains flat response out to 4 MHz. A slight rise occurs at high frequencies. If a peaking coil is used that overcompensates the load, the response is 100 percent at 5.5 MHz, but there is almost 20 percent rise at 4 MHz. This is an excessive amount of frequency distortion.

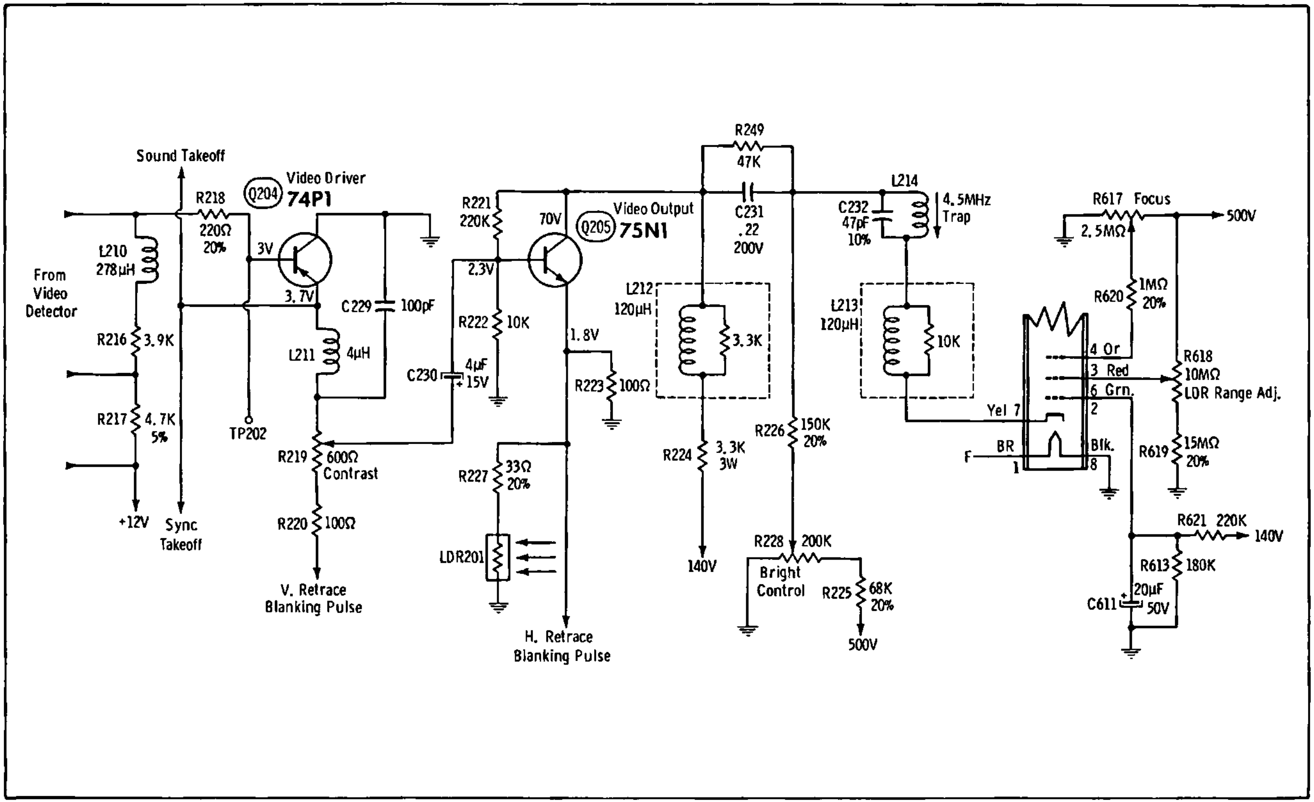


Fig. 5-13. Transistor video-amplifier circuit.

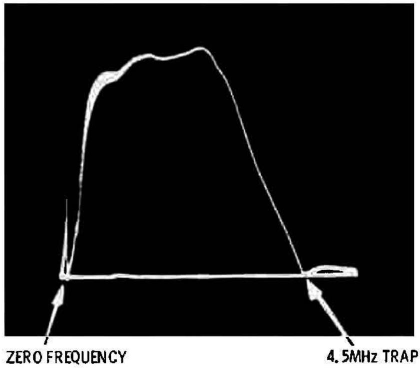
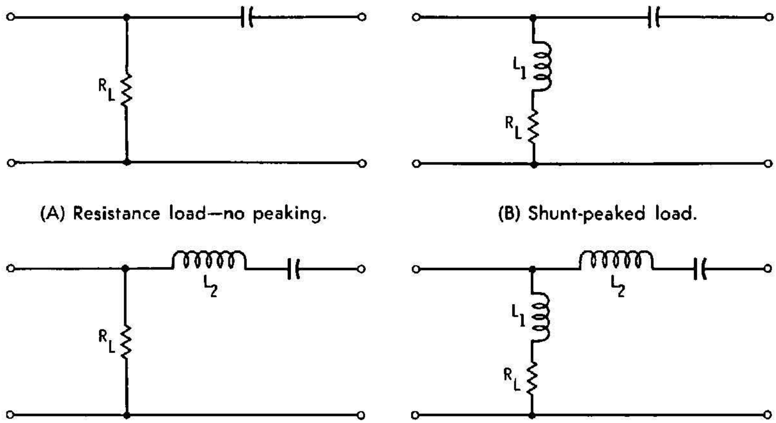


Fig. 5-14. Video-frequency response curve.

When both series and shunt peaking are used, flat response can be obtained out to 4 MHz with a larger value of load resistor. In turn, higher gain is obtained. When both series and shunt peaking are used, the dropoff is much more abrupt past 4 MHz. This results in poorer square-wave response, because abrupt dropoff is associated with a very nonlinear phase char-



(A) Resistance load—no peaking.

(B) Shunt-peaked load.

(C) Series-peaked load.

(D) Series- and shunt-peaked load.

Fig. 5-15. Peaking circuits in RC-coupled amplifier.

acteristic. In scopes with transistor vertical amplifiers, series peaking only may be utilized, since series peaking produces less overshoot than shunt peaking in square-wave reproduction. A scope that employs both series and shunt peaking has higher gain for a given number of transistors, but in turn has poorer square-wave response. Fig. 5-17 illustrates the 100-kHz square-wave response of a series- and shunt-peaked video amplifier in a color-TV receiver that has a phase-equalizing cir-

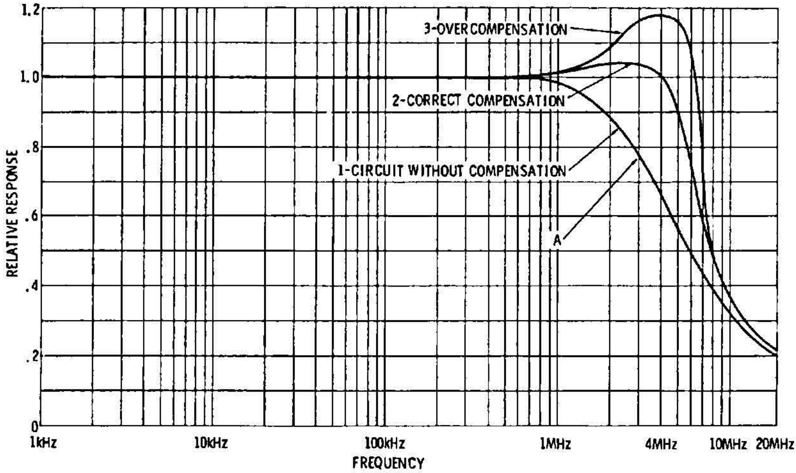


Fig. 5-16. Frequency-response curves with shunt peaking.

cuit to develop symmetrical square-wave response. Overshoot and preshoot are approximately equal in this arrangement.

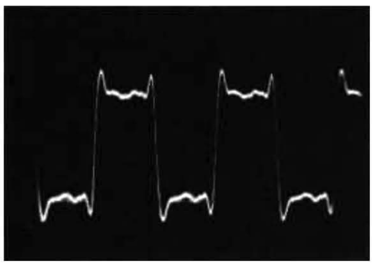


Fig. 5-17. Preshoot and overshoot in the 100-kHz square-wave response of a color-TV video amplifier.

DIRECT-COUPLED AMPLIFIERS

Direct-coupled amplifiers are extensively used in transistor equipment. A basic direct-coupled amplifier is shown in Fig. 5-1D. This type of amplifier is used for amplification of dc signals and very–low-frequency signals, as well as high-frequency signals. Direct-coupled amplifiers are attractive because they dispense with coupling capacitors or transformers. On the other hand, temperature stabilization is a greater problem than in ac coupled amplifiers. Various circuit means are utilized to avoid dc drift. In the foregoing circuit, an npn transistor is connected directly to a pnp transistor. Thereby, bias polarities are correctly provided. If the collector current in the first stage is greater than the base current in the second stage, a collector load resistor R1 must be included.

Fig. 5-18 depicts a three-stage direct-coupled amplifier. Npn transistors are used, and are forward-biased for class-A operation by means of a voltage-divider network. To provide good bias stabilization, R7 operates as a negative-feedback emitter resistor in the output stage. In addition, dc feedback is provided from the emitter of Q3 to the base of Q1 via R6. Suppose that the base bias on Q1 tends to drift more positive; in turn, the base bias on Q2 becomes less positive, as does the base bias on Q3. Therefore less emitter current is drawn by Q3, and its emitter voltage becomes less positive. Accordingly, the base bias on Q1 is prevented from drifting via feedback through R6.

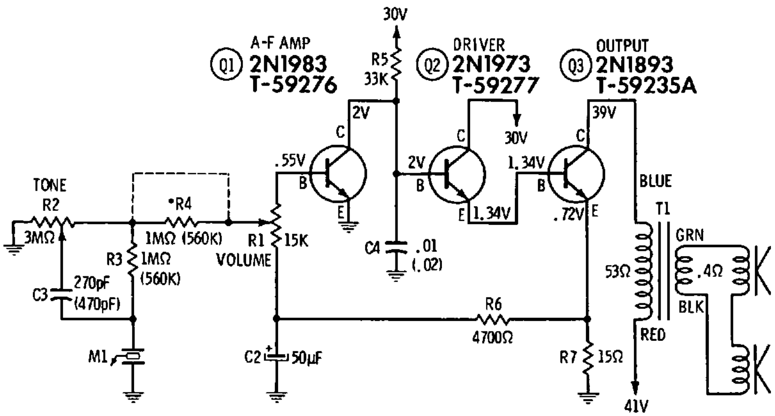


Fig. 5-18. Three-stage direct-coupled amplifier.

Q1 in Fig. 5-18 operates in the CE configuration, as does Q3. Q2 operates as an emitter follower. Additional bias stabilization is provided in the emitter circuit of Q2; R7 and the base-emitter junction resistance of Q3 serve as a current-feedback emitter resistance for Q2. Q2 also serves as an electronic impedance-matching device between Q1 and Q3, since the collector-output resistance of Q1 is high, and the base-input resistance of Q3 is comparatively low. Transformer coupling is used between Q3 and the speakers to obtain a good impedance match. Since the frequency response of a direct-coupled amplifier is very good, the system in Fig. 5-18 has a frequency response that is basically the response of T1. Square-wave response also is chiefly determined by the characteristics of T1.

REVIEW QUESTIONS

1. What is the theoretical maximum available gain of a transformer-coupled amplifier?
2. State an advantage and a disadvantage of transformer coupling.
3. Define leakage reactance.
4. Why does an output transformer not have a rising frequency response?
5. How does a rising frequency response affect square-wave reproduction?
6. Describe crossover distortion as displayed on a scope screen.
7. How is crossover distortion minimized in a push-pull amplifier?
8. Explain how the Q of a coil affects the bandwidth of an LCR resonant circuit.
9. Define the voltage amplification of a series-resonant LCR circuit.
10. Define the current amplification of a parallel-resonant LCR circuit.
11. Describe what is meant by the coupling coefficient of a tuned transformer.
12. Define critical coupling.
13. Explain how the hump frequencies of an overcoupled transformer are calculated.
14. Discuss the characteristics of an impedance-coupled amplifier.
15. How is series peaking distinguished from shunt peaking?
16. With other things being equal, does series or shunt peaking provide higher gain?
17. Compare the gain provided by series-shunt peaking with that of shunt peaking alone.
18. How do peaking coils affect the square-wave response of an amplifier?
19. Why is a slight rise at high frequencies often tolerated in a video amplifier?
20. What is the disadvantage of substantial overpeaking?
21. Describe the general features of a direct-coupled amplifier.
22. What is the basic problem that must be contended with in direct-coupled configurations?
23. Name an advantage of a direct-coupled amplifier.
24. How can bias stabilization be obtained in a direct-coupled amplifier?

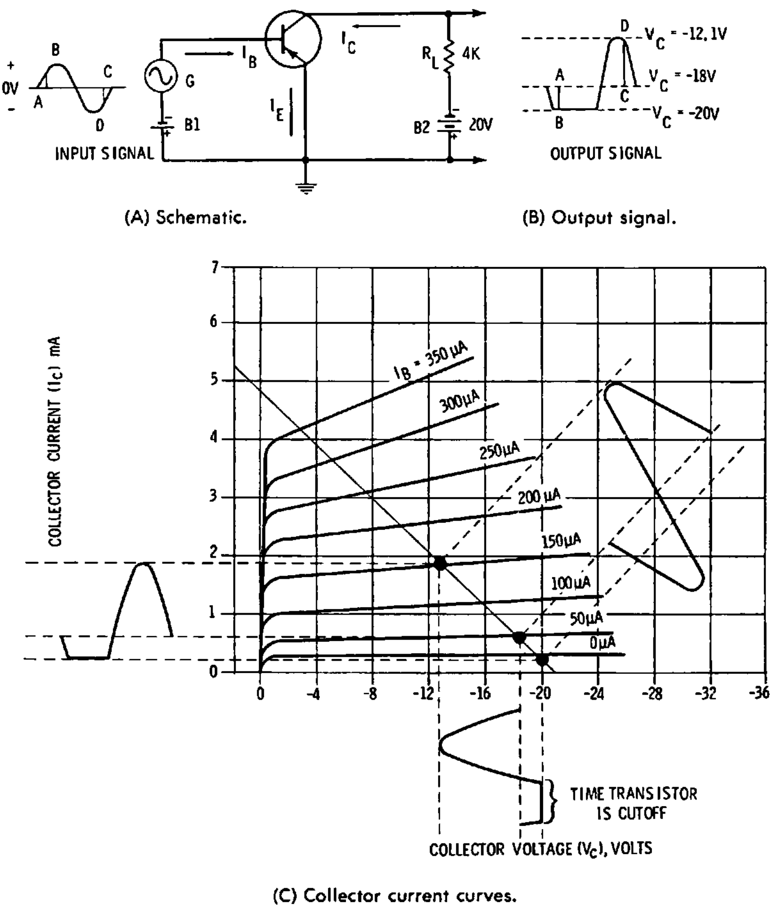
6

Waveforms in Transistor Waveshaping Circuits

A very wide range of waveshaping circuits is used in transistor equipment. Only the more basic and common types of waveshapers can be covered in this chapter. The most fundamental arrangement is called the limiter circuit. Limiter circuits are classified either as cutoff limiters, saturation limiters, or combination saturation-and-cutoff limiters. These circuits are related to the transistor amplifier circuits that have been previously discussed. Fig. 6-1A shows a typical transistor limiter configuration. To simplify the explanation of circuit action, the flow of minority carriers across the reverse-biased base-collector junction may be ignored.

Let us consider the first alternation of the input signal voltage in Fig. 6-1, which opposes the forward bias produced by the base-emitter battery B1. This effectively decreases the bias voltage, and therefore decreases the emitter current. Both the collector and base currents are decreased by corresponding amounts. The decreased collector current flowing through the load resistor decreases the voltage drop across the resistor. As the current and voltage drop across the load resistor continue to decrease, the collector voltage V_C approaches the collector-emitter bias voltage (−20 volts in this example). Since the operating point of the circuit has been placed on the load line at a point that will cause the collector current to reach zero with the input signal amplitude as shown, the transistor will be cut off during a portion of the alternation. This circuit action is called cutoff limiting. When the amplitude of the input signal is AB, the signal drives the circuit beyond cutoff and no current flows. The output in this example is at −20 volts.

Next, let us consider the second alternation in Fig. 6-1. The input voltage aids the forward bias produced by the base-emitter battery B1. As a result, the forward voltage is increased, thereby increasing the emitter current. Both the collector and base currents are increased by corresponding amounts. Increased collector current flowing through the load resistor increases the voltage drop across the resistor. Thus, the collector voltage V_C rises (becomes less negative) to −12.1 volts. This effect is seen in the output waveform at B. As the input signal starts to decrease (CD), the output waveform begins to go more negative. During the entire negative alternation, the input signal aids the forward bias. The output signal is positive-



(A) Schematic.

(B) Output signal.

(C) Collector current curves.

Fig. 6-1. Cutoff limiting using pnp transistor.

going, although, as shown, the entire output signal is negative due to the applied −20 volts.

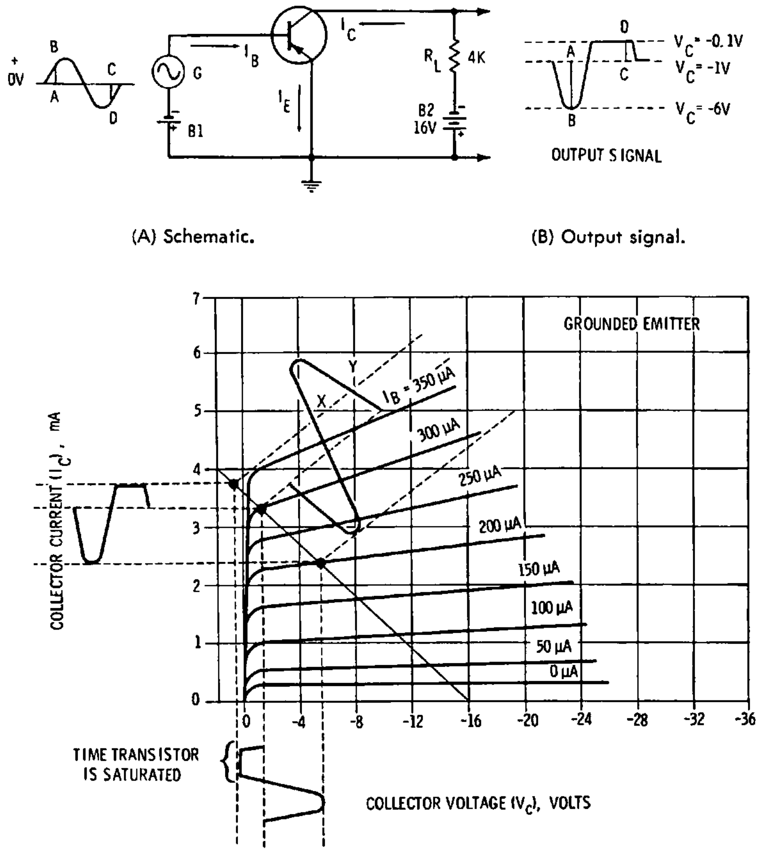
The extent of cutoff limiting in the signal is determined chiefly by the forward-bias value. However, the type of transistor, as well as the input circuit resistance and input-signal amplitude are also determining factors that affect the amount of cutoff limiting. A decrease in the forward-bias voltage shifts the operating point to a lower value of quiescent base current, thereby increasing the amount of cutoff limiting. Increasing the forward bias shifts the operating point to a higher value of quiescent base current, thereby decreasing the amount of cutoff limiting. If the input signal in Fig. 6-1 should be reduced to a very small amplitude, the limiter would operate as a CE amplifier.

SATURATION LIMITING

The circuit in Fig. 6-2 uses a pnp transistor connected in a CE amplifier configuration, and biased for saturation limiting. During the first (positive) alternation, the input signal voltage opposes the forward bias supplied by the base-emitter battery B1. The resultant forward voltage is decreased, thereby decreasing the emitter current. Both the collector and base currents are decreased by corresponding amounts. The decreased collector current flowing through load resistor R_L produces a smaller voltage drop across R_L. Collector voltage V_C changes from a no-signal value of −1 volt to −6 volts as shown in Fig. 6-2B. This may be determined from the characteristic curve of the limiter circuit shown in Fig. 6-2C. For the entire half-cycle that the input goes positive, the output is negative-going. No limiting action occurs over this half-cycle.

During the second alternation in Fig. 6-2, the input signal voltage aids the forward bias produced by the emitter-base battery. The resultant forward voltage is increased, thereby increasing the emitter current. Both the collector and base currents are increased by corresponding amounts. The operating point of the circuit in Fig. 6-2C has been established on the load line above the point for class-A operation. The increase in amplitude of the input signal between X and Y does not increase the collector current, which is at saturation. Increased collector current flowing through load resistor R_L increases the voltage drop across R_L. At saturation, a steady voltage of about 15.9 volts opposes the battery voltage B2.

Note that the output voltage levels off at −0.1 volt in Fig. 6-2. This may be determined from the characteristic curve of the limiter circuit shown in Fig. 6-2C. Its effect, seen on the output waveform, is called saturation limiting. Increasing the forward bias shifts the operating point to a higher value of quiescent base current, thereby increasing the amount of saturation limiting. Decreasing the forward bias shifts the operating



(A) Schematic.

(B) Output signal.

(C) Collector current curves.

Fig. 6-2. Saturation limiting using pnp transistor.

point to a lower value of base current, thereby decreasing the amount of saturation limiting. Other factors determining the amount of saturation limiting are the type of transistor, the input circuit resistance, and the amplitude of the input signal. Saturation limiting is employed less commonly than cutoff limiting because the transistor is operated near its maximum rated power dissipation, which complicates the design problem.

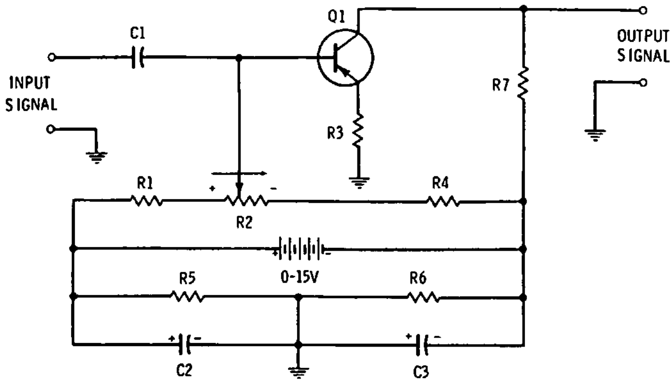


Fig. 6-3. Transistor triode limiter.

CUTOFF AND SATURATION LIMITING

Fig. 6-3 depicts a conventional CE amplifier circuit that can be adjusted for either cutoff or saturation limiting. When R2 is adjusted for minimum forward bias, the circuit operates as a cutoff limiter; if R2 is adjusted for maximum forward bias, the circuit operates as a saturation limiter. Again, if R2 is set to the midpoint of its range, conventional amplifier action is obtained with a small or moderate input signal amplitude. A large input signal amplitude produces both cutoff and saturation limiting, as shown in Fig. 6-4. The saturation level is indicated at (1), the quiescent level at (2), and the cutoff level at (3). Thus, a sine wave is shaped into a semisquare wave. This arrangement has the same disadvantage as noted previously for a simple saturation limiter. Therefore, it is more common practice to use a pair of cutoff limiters connected in cascade to provide both positive and negative peak limiting.

When a square wave is to be formed by limiting a sine wave, successive amplification and clipping must be utilized to obtain fast rise. Because numerous transistors and associated components are required, it is usually preferred to generate square

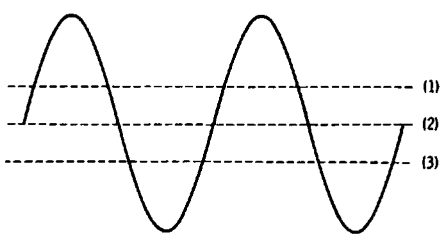


Fig. 6-4. Clipping of both positive and negative peaks.

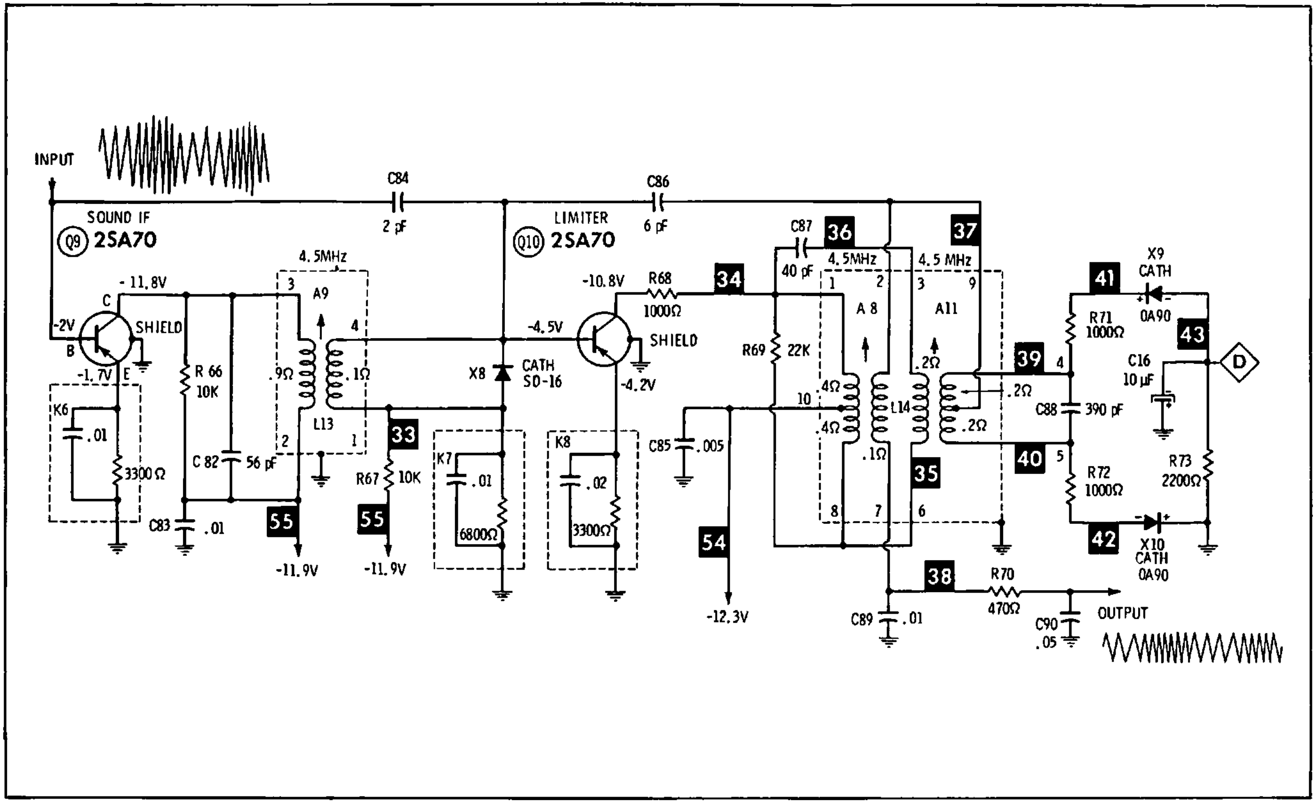


Fig. 6-5. Fm limiter in TV receiver.

waves directly by means of multivibrators, as explained previously. However, limiters serve a unique purpose, for example, in fm radio receivers, and in the sound sections of TV receivers. Fig. 6-5 shows a limiter arrangement for a TV receiver. This is a diode-transistor cascade limiter. Diode X8 conducts on negative peaks, and thereby limits the negative excursion of the signal by short-circuiting the input circuit of the transistor. Positive peaks are applied to the base of transistor Q10, thereby providing cutoff limiting. Subsequent limiting action is provided by the ratio detector.

TRANSISTOR CLIPPER CIRCUITS

We have seen that a limiter rejects the positive peak, negative peak, or both positive and negative peaks of a waveform. On the other hand, a clipper passes either the positive peak or the negative peak, and rejects the remaining portion of the waveform. For example, Fig. 6-6 depicts a negative-peak clipper. The base-emitter junction of the transistor is normally zero-biased. However, when an input signal is applied, the base draws current on negative half-cycles. This causes the coupling capacitor to charge and produce positive signal-developed bias on the base. Therefore, the transistor is reverse-biased with signal present. Between negative input-signal peaks, some of this bias leaks to ground. The amount of decay depends on the RC time constant of the base circuit.

It follows that the input signal in Fig. 6-6 will produce sufficient base conduction on each negative peak to replace the charge decay. In turn, the amount of clipped-peak waveform that appears in the collector circuit depends on the RC time constant of the base circuit. The extent of clipping also depends on the input signal level. At very small input levels, almost one-half cycle is passed into the collector circuit; however, at high signal levels, only a small portion of the peak

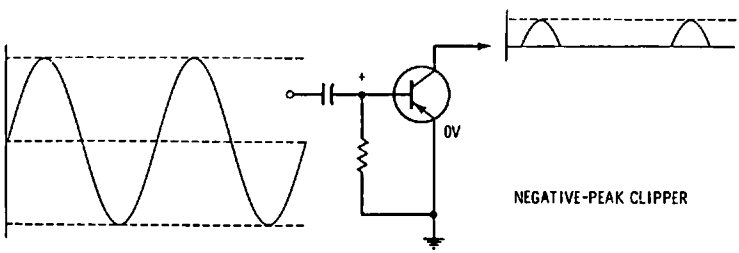


Fig. 6-6. Negative-peak clipper.

drive waveform passes into the collector. This basic arrangement is widely used in sync-separator configurations, as shown in Fig. 6-7. Note that the negative sync tips are stripped from the video signal, amplified, and passed into the collector load.

The normal dc voltages in the absence of signal in Fig. 6-7 provide forward bias of 0.2 volt on Q13. An average video signal applies 1 volt pk-pk drive to the base of the transistor. Signal-developed bias reverse-biases the base-emitter junction, and only the negative sync tips can drive the transistor into conduction. Because the video-signal level varies from one channel to another, it is desirable to have bias regulation to maintain the clipping point at the black level of the video signal. This is the function of diode X38. When the signal level is high, the diode conducts more, or its internal resistance decreases. In turn, the emitter bias voltage becomes less negative, and part of the signal-developed bias is effectively cancelled. Thereby, the clipping point of the transistor tends to "follow" the black level in the video signal.

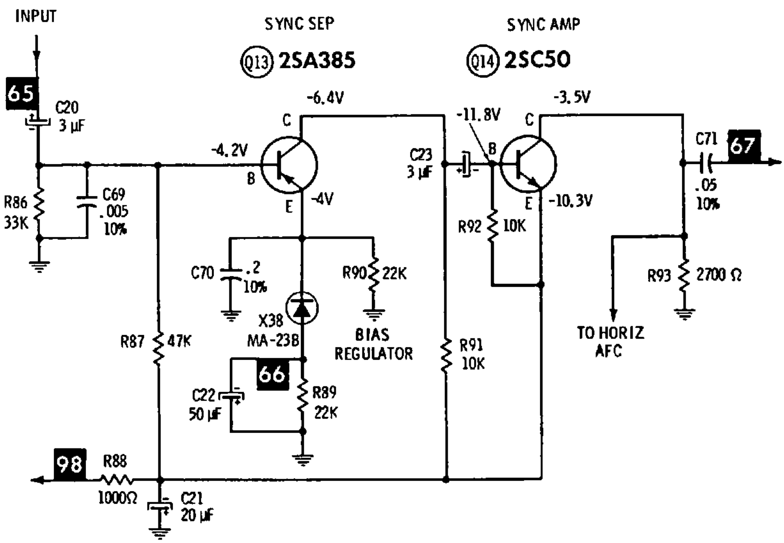


Fig. 6-7. Peak clipper used as sync separator.

TRANSISTOR SWITCH WAVESHAPERS

Pulse and switching waveshapers are used in television, radar, telemetering, pulse-code, and computer equipment. In this chapter, we are concerned chiefly with the basic features

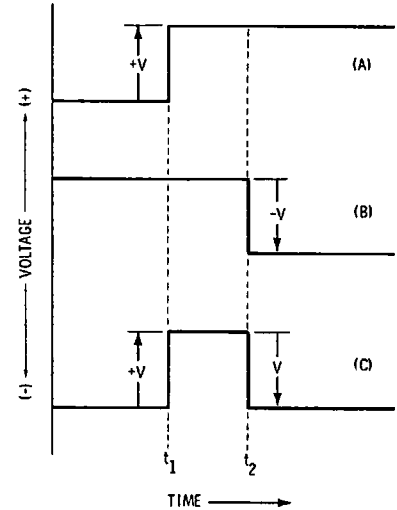


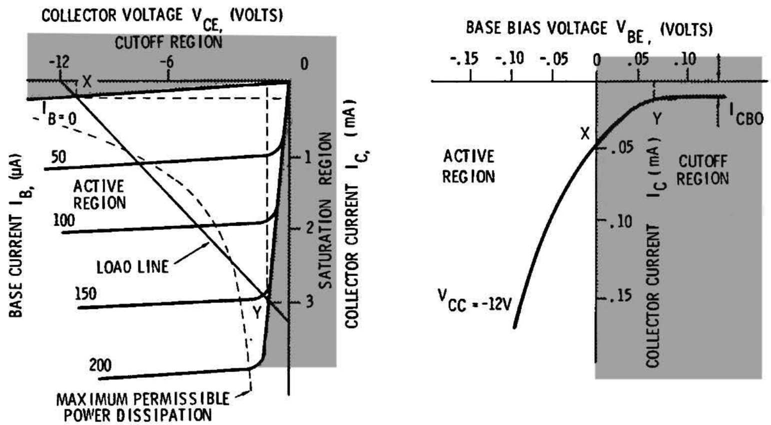
Fig. 6-8. Unit step waveforms, showing formation of a pulse.

of transistor switch waveshapers. Pulse and switching circuits are normally characterized by large-signal, or nonlinear operation of a transistor. Circuit operation usually entails a driving pulse, which might be generated by a blocking oscillator, for example. The circuit develops a discontinuous change in voltage level or current level. That is, the input waveform (trigger pulses) produces output signals that have large and sudden changes in voltage or current. We will find that the output waveform may differ considerably from the input waveform. A switching circuit has two states: the transistor is either cut off, or it is in collector saturation.

Trigger-pulse waveforms are illustrated in Fig. 6-8. For simplicity, these are shown as ideal waveforms with zero rise and fall times. A voltage that undergoes a sudden change in amplitude from one level to another is called a unit step voltage. In pulse and switching-circuit application, when the unit step voltage is the applied signal, it is commonly of sufficient amplitude to cause the circuit to change from a conduction state to a cutoff state, or vice versa. In turn, a unit step voltage in the output section of the circuit results from the change of state.

In Fig. 6-8, a positive unit step voltage waveform is shown at A. At time t_1, the voltage level is increased positively by an amplitude V. This voltage level does not necessarily have to increase from zero to its maximum value. If the initial voltage was at a negative potential, and then changed to zero, note that a positive unit step voltage would be produced. Fig. 6-8B depicts a negative unit step voltage. At time t_2, the voltage level is decreased by an amplitude V. In this example, the change in level could be from a high positive potential to a lower positive potential. Fig. 6-8C depicts a square or rectangular pulse waveform. At time t_1, the voltage level is increased by an amplitude V. Between t_1 and t_2, a new constant voltage level is established. At time t_2, the voltage level is decreased by an amplitude V. Thus, a square pulse can be described as formed by two unit step voltages, one of which is positive and the other negative.

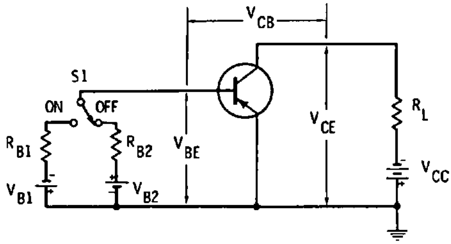
When a transistor is operated in the switching mode by a pulse-input waveform, it may be regarded basically as an overdriven amplifier, being driven alternately from cutoff into collector saturation, and vice versa. A transistor switch is analogous to a relay. Various circuit configurations have associated advantages and disadvantages. The CE output characteristics of a typical pnp transistor are shown in Fig. 6-9A. These characteristics are arranged in three regions: cutoff, active, and saturation. An arbitrarily chosen load line and the maximum permissible power dissipation curve are also shown. We consider the cutoff and saturation regions to be the stable or quiescent regions of operation. A transistor is considered to be in the *off* (nonconducting) or *on* (conducting) state when it is operated in the cutoff or saturation regions, respectively. The third region of operation, called the active region, is the unstable (transient) region through which operation of the transistor passes while changing from the *off* state to the *on* state.



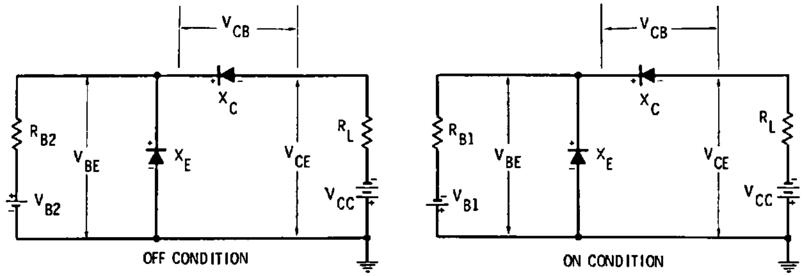
(A) Output characteristics.

(B) Effect of base bias on collector current.

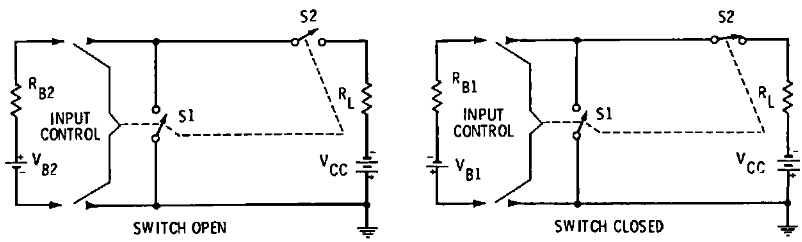
Fig. 6-9. Transistor characteristic curves, switching application.



(A) Conventional schematic.



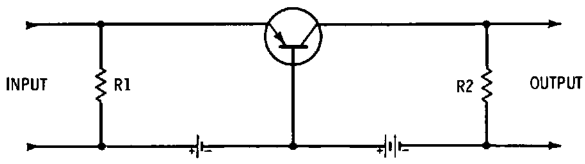
(B) Diode equivalent circuit.



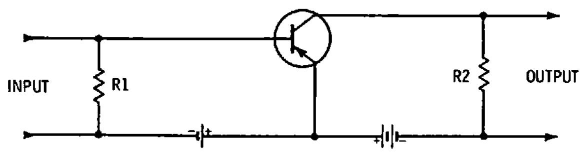
(C) Switch equivalent of circuit.

Fig. 6-10. Transistor switching circuits.

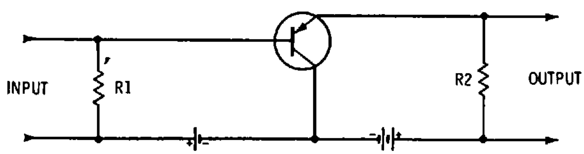
The effect of base-bias voltage V_{BE} on collector current I_C in the cutoff and active regions is shown in Fig. 6-9B. A typical transistor switching circuit is shown in Fig. 6-10A. Switch S1 represents a pulse waveform source, such as a blocking oscillator. It controls the polarity and amount of base current from battery V_{B1} or V_{B2}. Resistors R_{B1} and R_{B2} are current-limiting resistors. The emitter-base and collector-base diode and switch equivalent circuits representing the *off* and *on* (dc) conditions of the transistor switching circuit are shown in Fig. 6-10B and C. Note that although a CE configuration is depicted in Fig. 6-10, we may employ any of the three basic configurations shown in Fig. 6-11. These various circuits have advantages and disadvantages in switching applications, as explained subsequently.



(A) Common base (CB).



(B) Common emitter (CE).



(C) Common collector (CC).

Fig. 6-11. Basic amplifier configurations.

The cutoff region in Fig. 6-9A includes the area above the zero base-current curve ($I_B = 0$). Ideally, with no base current flow, there would be zero collector current. That is, the collector voltage would equal the battery voltage V_{CC} in Fig. 6-10A. However, at point X on the load line in Fig. 6-9A, a small amount of collector current flows. (See the horizontal projection to the collector-current axis.) This is more clearly indicated in Fig. 6-9B, where, at zero base-bias voltage (point X), the collector current I_C equals approximately 0.05 ma. This is the reverse-bias collector current for the CE configuration. Note that the application of a small reverse base-bias voltage V_{BE} (approximately 0.075 volt), reduces the value of reverse-bias collector to the value of I_{CBO}. We recall that the significance of I_{CBO} is as shown in Fig. 6-12.

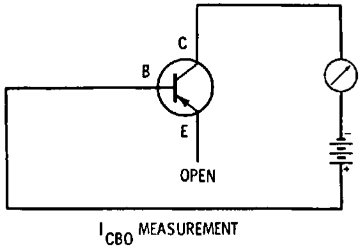


Fig. 6-12. I_{CBO} **measurement.**

Collector voltage V_{CE} in Fig. 6-9A is indicated by the vertical projection from point Y to the collector-voltage axis. This value is equal to the difference in magnitude between the battery voltage (12 volts in this example) and the voltage drop produced by reverse-bias collector-current flow through load resistor R_L in Fig. 6-10A. Normal quiescent conditions for a transistor switch in this region require that both the emitter-base junction and collector-base junction be reverse-biased. With S1 in its *off* position, the emitter-base junction is biased by V_{B2} via R_{B2}. This is comparable to application of a positive unit-step voltage. The collector-base junction is reverse-biased by V_{CC} via R_L; the transistor is in its *off* state. We will find that a switching circuit is designed to produce very fast transition between its *off* and *on* states.

Both limiting and switching circuits are commonly used in cascade for horizontal deflection of a picture tube. For example, Fig. 6-13 shows a limiter stage that is driven by the output

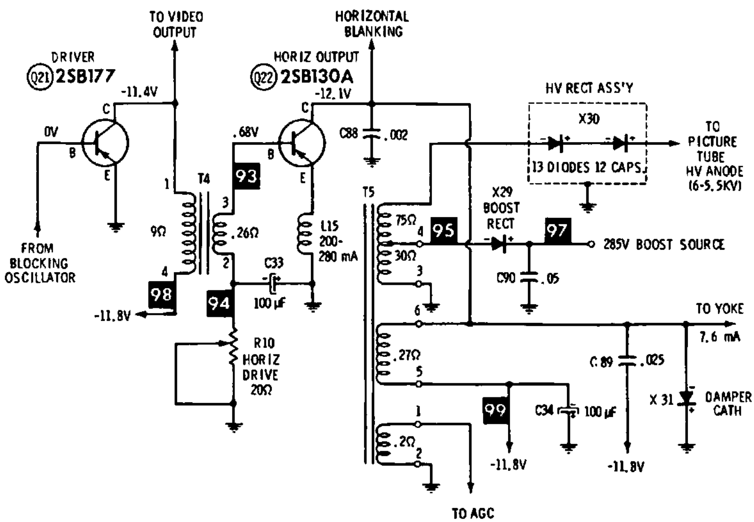


Fig. 6-13. Limiter and switch waveshaper circuit for horizontal-output stage.

from a blocking oscillator. The output waveform is limited to form a rectangular pulse. This pulse drives a power-output transistor connected in a switching arrangement. Note that the output transistor Q22 is reverse-biased; this is a signal-developed bias produced by base-current flow on drive peaks that charge C33. The chief requirement is that Q22 be driven with great rapidity from its *off* state to its *on* state, and vice versa.

In the diode equivalent circuit of the transistor in Fig. 6-10B, diodes X_E and X_C represent the emitter-base and collector-base junctions, respectively. Diode X_E is reverse-biased by V_{BE}; diode X_C is reverse-biased by V_{CB}. Ideally, there is no current flow through R_L, and V_{CE} equals V_{CC}. The circuit thus represents an open switch, as depicted in Fig. 6-10C. That is, the initially applied bias causes S1 and S2 to open the output circuit. The active linear region shown in Fig. 6-9A is the only region that provides conventional amplifier action. In the linear region, the collector-base junction is reverse-biased, and the emitter-base junction is forward-biased. Transient response of the output signal is essentially determined by the transistor characteristics in this region. In a switching circuit, this is called the transition region.

Operation of S1 in Fig. 6-10A to its *on* position is comparable to the application of a negative unit-step voltage. Forward bias is established by V_{B1}, via R_{B1}, and I_B and I_C become transitory, moving from point X in Fig. 6-9A along the load line to point Y. Here, collector current reaches saturation; the signal passes through this region rapidly. Note that load line A passes through the maximum power-dissipation curve. This is permissible because the excursion of collector current through the "forbidden region" is very rapid, so that the average power dissipation is low through the forbidden region and does not damage the transistor.

Some horizontal-deflection systems include an emitter follower that serves as an impedance-matching device, as shown in Fig. 6-14. It provides improved power-transfer efficiency, although an extra transistor is required. Good efficiency is desirable to minimize battery drain. Note that in the saturation region shown in Fig. 6-9A, an increase in base current does not cause an appreciable increase in collector current I_C. At point Y on the load line, the transistor is in its saturation region. I_C, measured by the horizontal projection from Y, is at a maximum, and V_{CE}, which is measured by vertical projection from Y, is at its minimum value. This value of collector voltage is called V_{sat}. Excessive or "deep" saturation is undesirable,

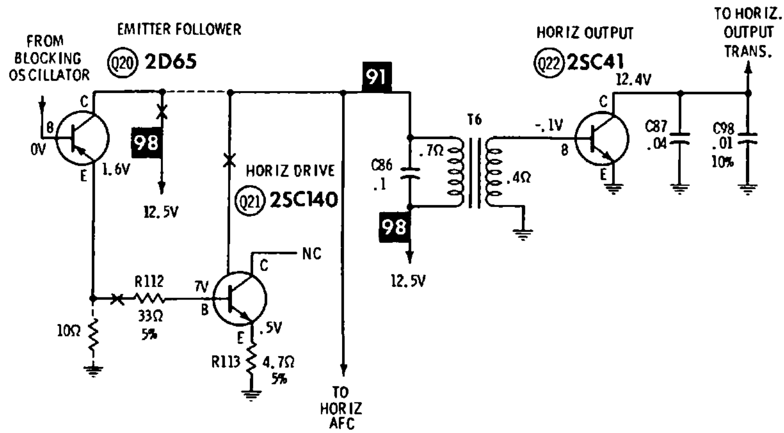


Fig. 6-14. Emitter follower used as an impedance-matching device.

because it tends to distort the output waveform, as explained subsequently in greater detail. The saturation region is also called the bottomed region of the transistor.

In Fig. 6-10A, when I_C reaches its limited value, the transistor saturates and is in its *on* state. In the diode equivalent cir-

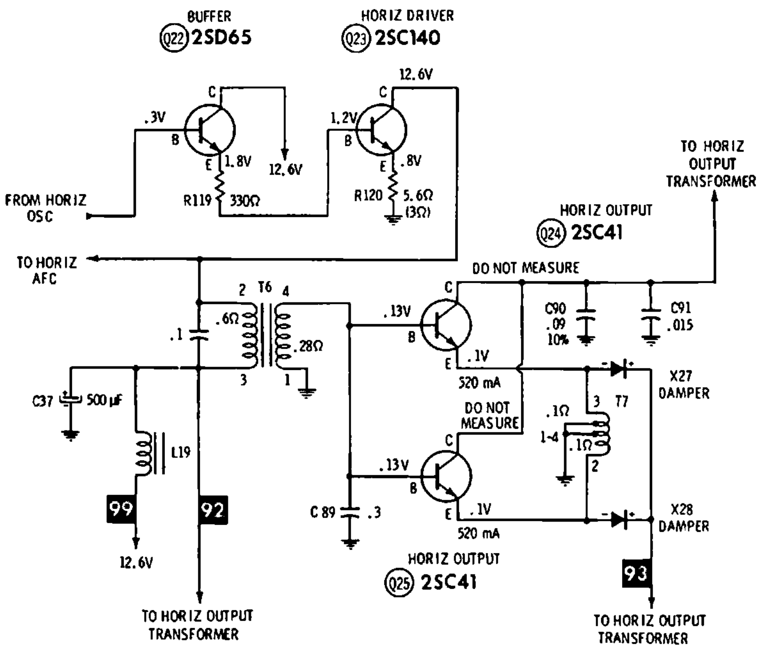


Fig. 6-15. Paralleled transistors provide double power output.

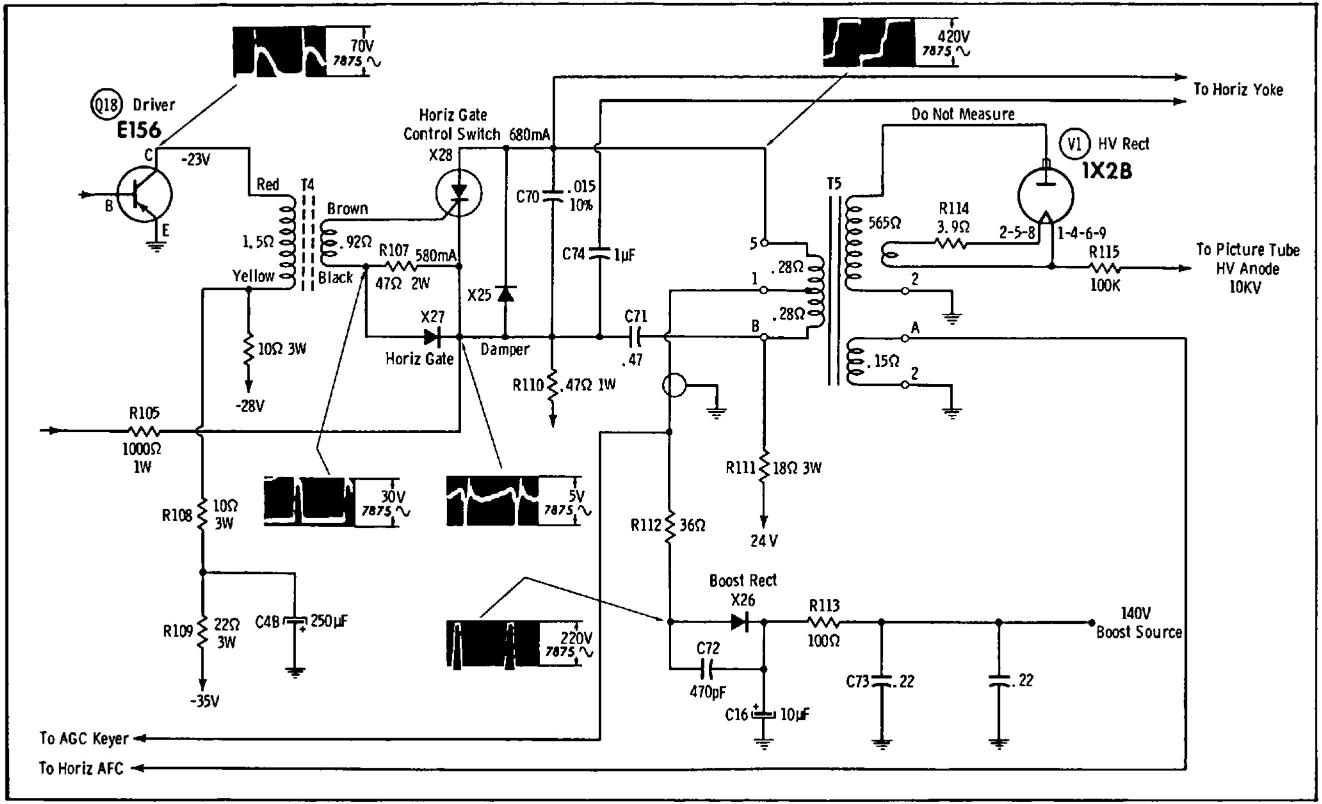


Fig. 6-16. Silicon controlled rectifier used instead of horizontal-output transistor.

cuit of Fig. 6-10B, X_E and X_C are forward-biased. X_E is forward-biased by V_{BE}. The saturation voltage V_{CB} drops to a smaller negative value than V_{BE}; the difference of these voltages produces the forward bias V_{CB} on X_C. That is, the equivalent switch circuit is closed. S1 and S2 close the circuit for V_{CC} via R_L. The circuit is switched from the *on* state to the *off* state in a similar manner. Small input voltage or current pulses can control large output voltage or current pulses. If greater power output is required than can be provided by a single transistor, two transistors are often paralleled, as depicted in Fig. 6-15.

Another method of obtaining greater output power is to use a silicon controlled rectifier instead of a horizontal-output transistor, as shown at X28 in Fig. 6-16. An SCR is analogous to a thyratron, as a transistor is analogous to a triode tube. That is, an SCR can be gated into conduction, but will not thereafter cease conduction until the output voltage swings to zero. That is, the input gating pulse cannot bring the SCR out of conduction. It is often more economical to use an SCR as a gated control switch than two or more transistors connected in parallel.

We also find horizontal-output transistors connected in the CC configuration, as shown in Fig. 6-17. All three of the basic configurations shown in Fig. 6-11 may be operated in the switching mode. The CB configuration operates as a series switch. The collector output current is a large percentage of the emitter input current. For practical purposes, we may consider these two currents to be equal. The transient response of the CB arrangement is better than that of the other basic circuits. However, impedance matching may present a problem because the input resistance is low and the output resistance is high.

The CE configuration operates as a shunt switch. That is, the base input current merely controls the collector output current. Transient response is poorer than that of the CB arrangement. However, it has high power gain, and impedance matching usually imposes fewer problems. Next, the CC configuration also operates as a shunt switch. Its voltage gain is practically unity. The CC switch is easily driven because of its high input resistance. Its low output impedance is well adapted to loads that draw heavy ac currents. A CC switch is usually driven near, but not into its saturation region because both input and output resistances change greatly in the saturation region, which causes difficulties in avoiding output waveform distortion.

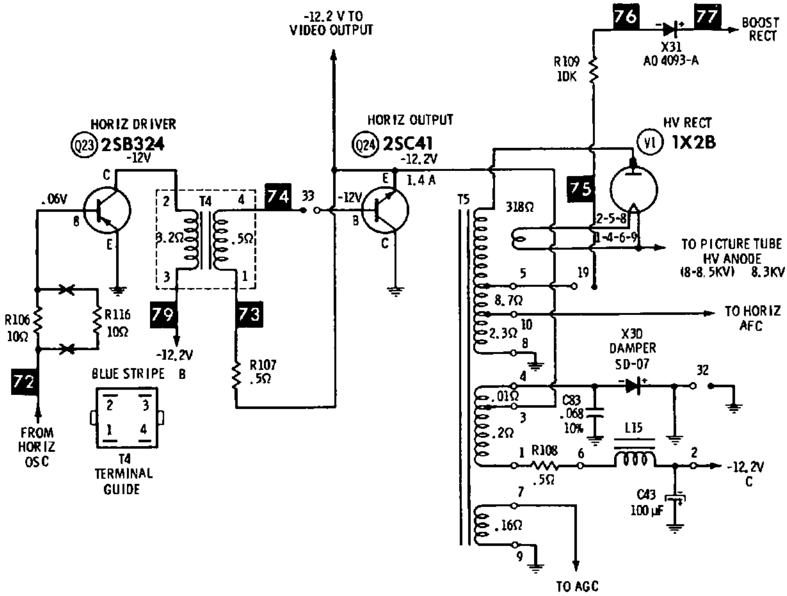


Fig. 6-17. Horizontal-output transistor connected in CC configuration.

REVIEW QUESTIONS

1. Name the three basic types of limiters.
2. How is cutoff limiting accomplished?
3. Explain how saturation limiting is provided by a transistor.
4. Describe combination cutoff and saturation limiting action.
5. How can a semiconductor diode and a transistor provide limiting of both signal peaks?
6. Discuss the operation of a negative-peak clipper that uses signal-developed bias.
7. Distinguish between a limiter circuit and a clipper circuit.
8. Define a unit step voltage waveform.
9. What is the basic characteristic of a transistor switch waveshaper?
10. Explain the action of trigger pulses in a switching circuit.
11. Define the cutoff, active, and saturation regions of a transistor.
12. Which are the stable regions of operation?
13. Identify the unstable region of a switching transistor.
14. Describe the basic I_{CBO} measurement.
15. Define the *on* and *off* states of a switching transistor.

16. Why is a limiter often used to drive a switching transistor?
17. What is the advantage of driving a switching transistor with an emitter follower?
18. Explain why a pair of switching transistors may be operated in parallel.
19. Describe the function of a silicon controlled rectifier.
20. What advantage does an SCR offer in a horizontal-output system?
21. Name the three basic transistor switching configurations.
22. State the chief features of a CB switching circuit.
23. What are the basic characteristics of a CC switching configuration?
24. Compare a CE switch with a CC switch arrangement.
25. Why is a CC switch ordinarily driven short of saturation?

7

Transistor Black-and-White TV Circuits and Waveforms

Several types of waveforms come into analysis in transistor black-and-white TV circuits. The most prominent type is the self-generated waveform; with reference to Fig. 7-1, such waveforms are found in the uhf tuner, vhf tuner, vertical-oscillator, and horizontal-oscillator sections. These are the source sections of self-generated waveforms, from which the waveforms branch out through waveshaping circuits into many driven sections. In other words, we are concerned with the ac voltage distribution of the receiver in the same way that we are concerned with the dc voltage distribution. Many of the source waveforms, such as those generated by the vertical and horizontal oscillators and processed by waveshapers, can be observed directly with the aid of a scope. On the other hand, waveforms generated by the uhf and vhf oscillators have frequencies that are beyond the capabilities of ordinary service scopes, and these waveforms must be analyzed indirectly. Therefore, we subclassify self-generated waveforms into the directly observable types, and the indirectly observable types.

Next, let us note another basic type of waveform that is almost as prominent as the self-generated waveform. This class of waveform is the applied type, which is subclassified into the signal and power types. With reference to Fig. 7-1, such waveforms are found throughout the signal channels and the sync channel, and in the low-voltage power supply (provided the transistor receiver is operated from a 60-Hz line). Applied waveforms are, in turn, subclassified into normal-operation and test-signal types. For example, the composite modulated-rf video signal supplied by an antenna is a normal-operation

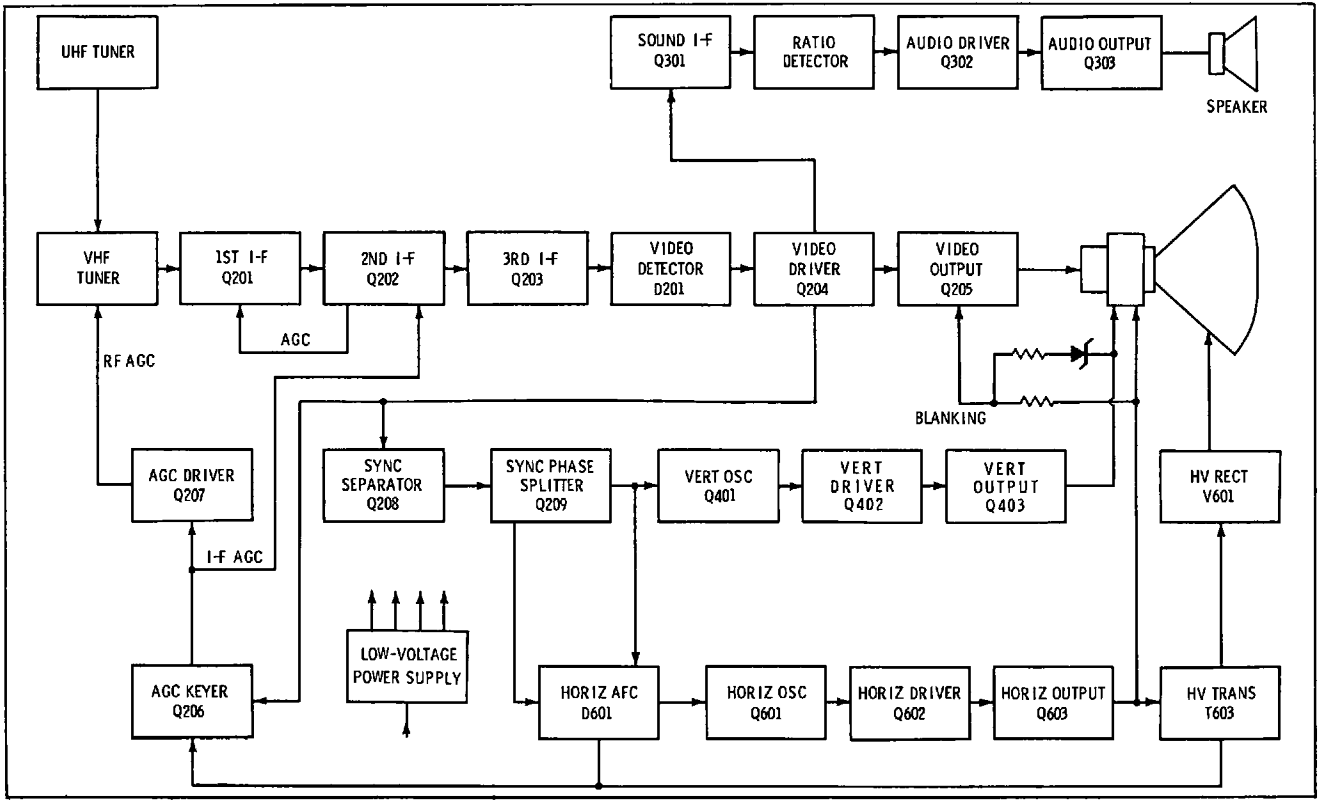


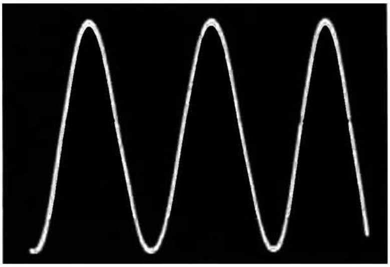
Fig. 7-1. Block diagram of transistor television receiver.

waveform. Signals supplied by signal generators, audio oscillators, sweep-and-marker generators, square-wave generators, and TV analyzers are test-signal types of waveforms.

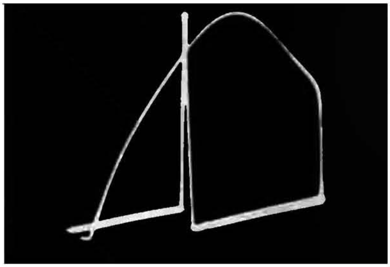
Waveforms are also classified into generic and derived types of waveforms. For example, the output from a vhf oscillator is a generic waveform, which means that it is the actual waveform generated by the transistor circuit. On the other hand, the beat waveform that is produced by mixing the vhf oscillator output with the output from a signal generator is a derived waveform. This means that we invoke a process that operates on the generic waveform and develops a new waveform that is related to the generic waveform, but is not the same. For example, if we apply an fm sweep signal to the input terminals of a vhf tuner, the signal beats with the vhf-oscillator waveform and develops a new waveform called a frequency-response curve. A frequency-response curve is a derived waveform.

Generic waveforms are illustrated in Fig. 7-2A and B. Two generic waveforms are also shown in Fig. 7-2C. On the other hand, the waveform to Q22 is a shaped waveform. In the first analysis, we merely note the waveform amplitude by measuring its peak-to-peak voltage. Preliminary signal-tracing procedures involve only a check for presence or absence of various waveforms. More detailed analyses include measurement of frequency or repetition rate, and observation of waveform distortions. There are many classifications of waveform distortion, the more basic types of which are explained subsequently. Every effect has its cause, and if you know how to analyze the effect, you can proceed without hesitation to its cause. Note that waveform displays can be completely misleading unless good practices are observed. For example, circuit loading can be a problem in some situations.

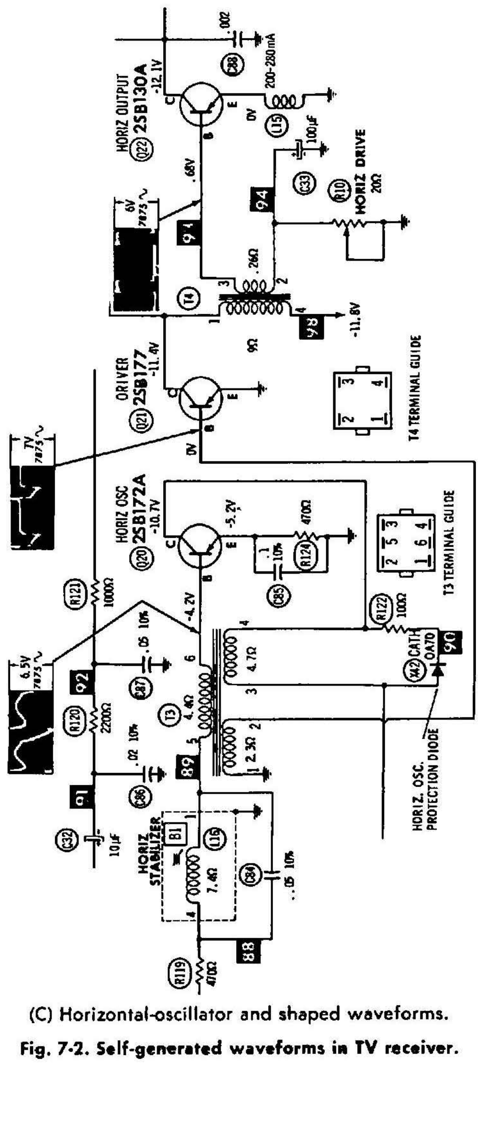
Fig. 7-3 illustrates normal and distorted waveforms of the applied test-signal type. Note that the antenna supplies a composite modulated-rf video signal, which is a normal-operation waveform. This waveform is immediately converted into a derived waveform by beating against the local-oscillator output. In turn, the derived waveform is a composite modulated-rf video signal with a beat envelope. This derived waveform is then heterodyned by the mixer transistor which operates as a nonlinear device to produce the i-f signal. This i-f signal is a composite modulated-i-f video waveform which has a lower carrier frequency than the input waveform. Next, the i-f waveform is processed by a semiconductor diode that operates as a nonlinear device to develop the envelope waveform as depicted in Fig. 7-4.



(A) Local-oscillator waveform.

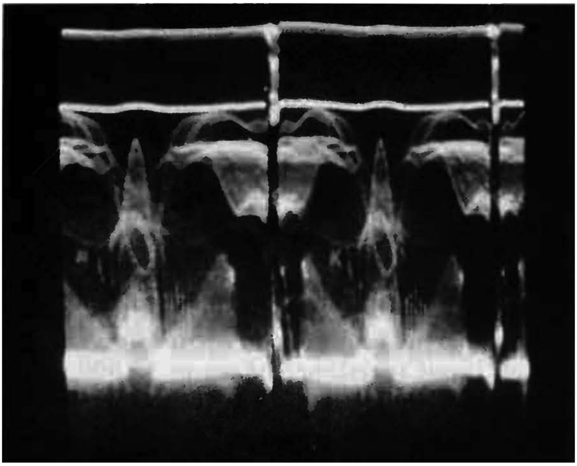


(B) Vertical-oscillator waveform.

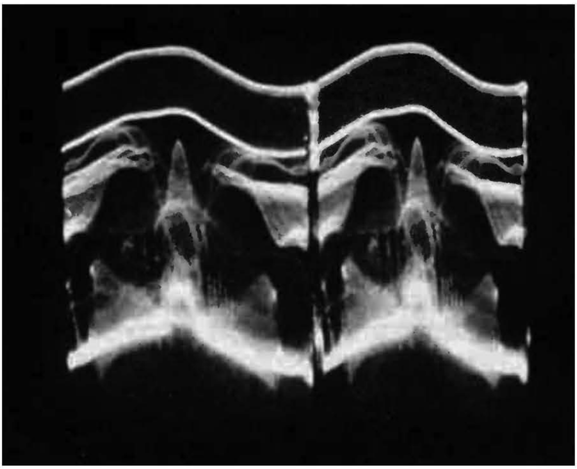


(C) Horizontal-oscillator and shaped waveforms.

Fig. 7-2. Self-generated waveforms in TV receiver.



(A) Normal video-signal waveform.



(B) Presence of 60-Hz hum voltage.

Fig. 7-3. Normal and distorted signal waveforms.

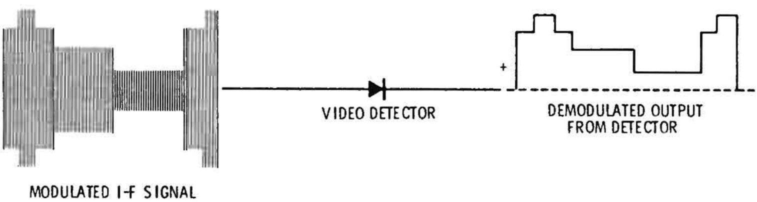
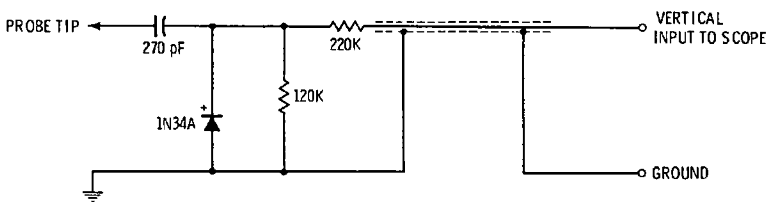


Fig. 7-4. Processing of video i-f signal to develop envelope waveform.

SIGNAL-TRACING PROCEDURES

Signal tracing generally refers to checking the course of an applied composite video signal through the picture, sound, and sync channels. On the other hand, waveform checking generally refers to the analysis of self-generated waveforms. These two areas of troubleshooting overlap because a horizontal-afc waveform, for example, comprises both applied and self-generated components. Nevertheless, the general distinction is a useful one. In the i-f amplifier, a demodulator probe must be used with the scope to check the progress of an i-f signal. A typical demodulator probe is depicted in Fig. 7-5. Note that the probe has an input resistance of about 15,000 ohms and an input capacitance of approximately 2.5 pF when an i-f signal is applied. This is a comparatively high input impedance, compared with the internal impedance and the load impedances of the i-f transistor (see Fig. 7-6). Therefore, signal tracing in the i-f amplifier usually involves no serious loading problem.

Waveforms processed by a demodulator probe are usually distorted, due to the limited envelope response of the probe. Therefore, it is helpful to operate the scope on 30-Hz deflection, so that the vertical interval and low-frequency compo-



(A) Schematic of probe.

FREQUENCY RESPONSE CHARACTERISITCS:	
RF CARRIER RANGE	500 kHz to 250 MHz
MODULATED-SIGNAL RANGE	30 to 5000 Hz
INPUT CAPACITANCE (APPROX.)	2.25 pF
EQUIVALENT INPUT RESISTANCE (APPROX.):	
AT 500 kHz	25,000 OHMS
1 MHz	23,000 OHMS
5 MHz	21,000 OHMS
10 MHz	18,000 OHMS
50 MHz	10,000 OHMS
100 MHz	5000 OHMS
150 MHz	4500 OHMS
200 MHz	2500 OHMS
MAXIMUM INPUT:	
AC VOLTAGE	20 RMS VOLTS 28 PEAK VOLTS

(B) Specifications.

Fig. 7-5. Typical demodulator probe and specifications.

nents of the video waveform are displayed. We are chiefly interested in the presence or absence of signal. Peak-to-peak voltage measurements are unreliable because the input capacitance of a demodulator probe tends to detune the circuit across which it is applied. This detuning effect will occasionally throw an i-f amplifier into oscillation, if the amplifier is substantially regenerative. For example, if one of the neutralizing capacitors in Fig. 7-6 were open, there is a possibility of oscillation when the probe is applied at associated terminals. Oscillation in an i-f amplifier blocks signal passage, and can be confused with other defects such as short-circuits. Note that if oscillation occurs, the dc voltage output from the picture detector increases greatly.

A demodulator probe is also used to signal-trace the intercarrier sound i-f signal from the sound-takeoff point in the video amplifier to the fm detector. Fig. 7-7 shows the configuration of a typical intercarrier sound i-f system. Since the bandwidth of a sound i-f amplifier is comparatively narrow, the collector load impedances are comparatively high. Therefore, a demodulator probe loads the tuned circuits and detunes them from their 4.5-MHz center frequency; however, the presence or absence of signal at each stage is clearly indicated in the test. Note that an intercarrier sound signal must be present. A TV station signal is suitable, or a TV analyzer signal that has a 4.5-MHz tone signal included with the video signal. The tone signal is displayed as a sine wave on the scope screen.

An rf sweep generator provides a suitable test for signal tracing through the picture channel to the output of the video amplifier. However, an rf sweep signal applied to the antenna-input terminals of the receiver will not provide an intercarrier-sound signal. If a sweep generator or an a-m generator is to be used for signal-tracing the intercarrier-sound channel, the signal must be applied at the output of the picture detector, and the generator must be tuned to a center frequency of 4.5 MHz. Note in passing that service-type a-m generators usually have more or less incidental fm. This means that if the generator is tuned to 4.5 MHz, and is amplitude-modulated, the output signal will contain incidental fm in addition to amplitude modulation. Greater percentages of amplitude modulation tend to produce greater deviation of the carrier.

In the video-amplifier section of the receiver, a low-capacitance probe is used with the scope in signal-tracing procedures. Fig. 7-8 depicts the configuration of a typical video amplifier. The first stage operates as an emitter follower, and produces no voltage gain. However, it provides the needed current gain

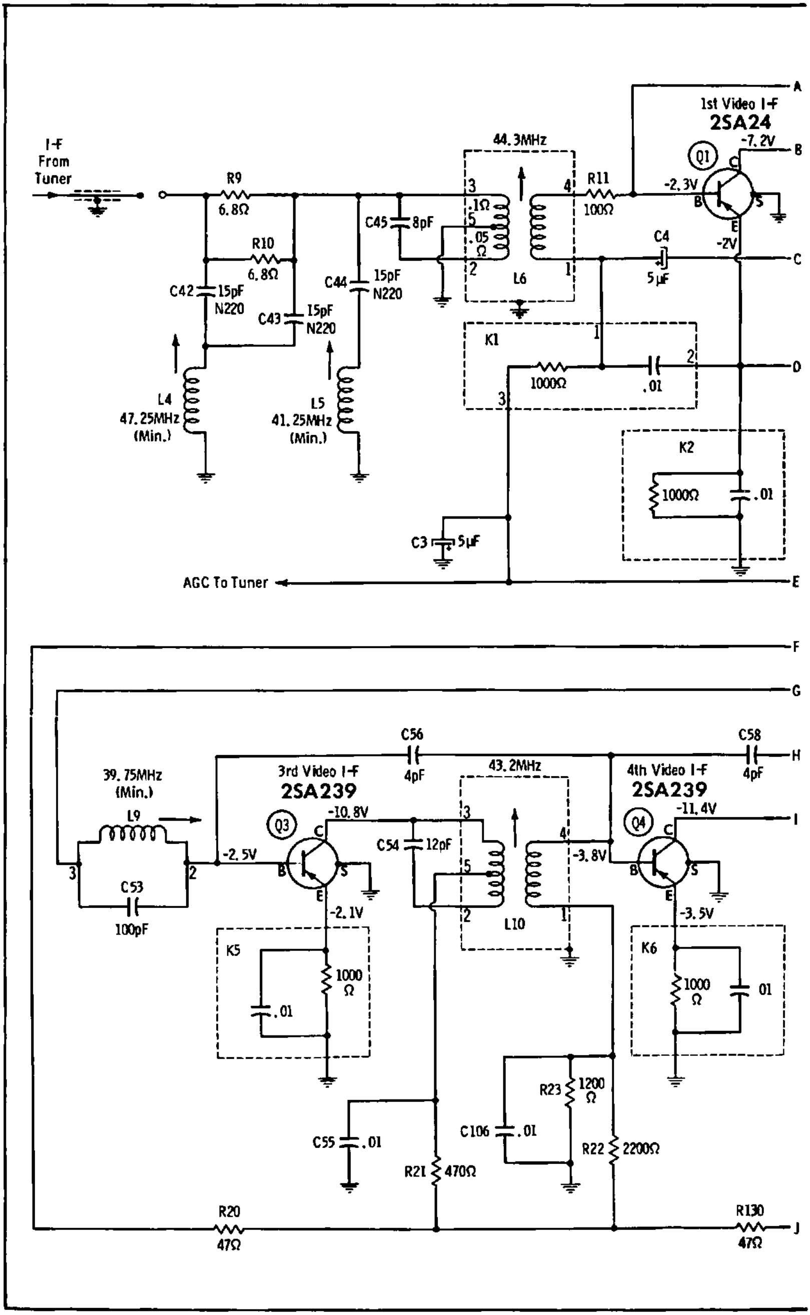
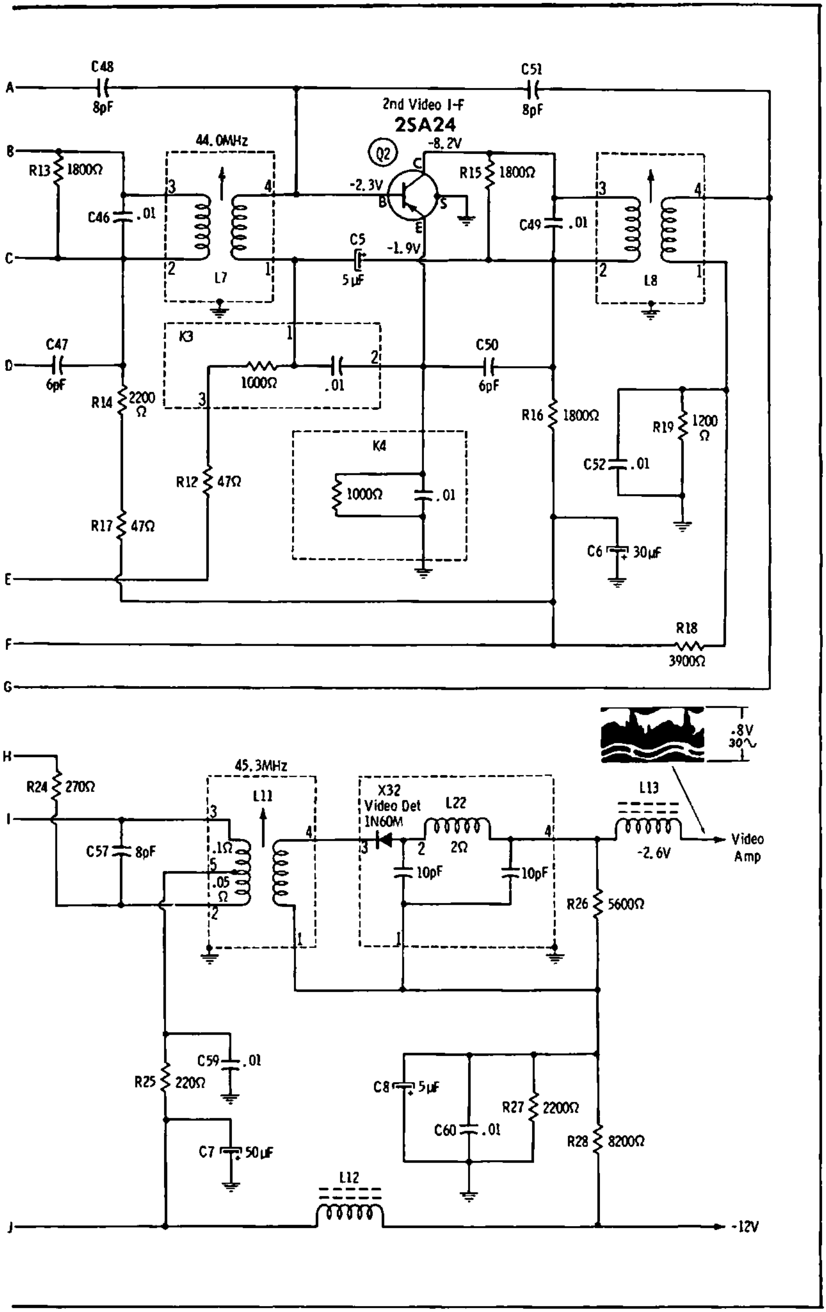


Fig. 7-6. Typical video



i-f amplifier strip.

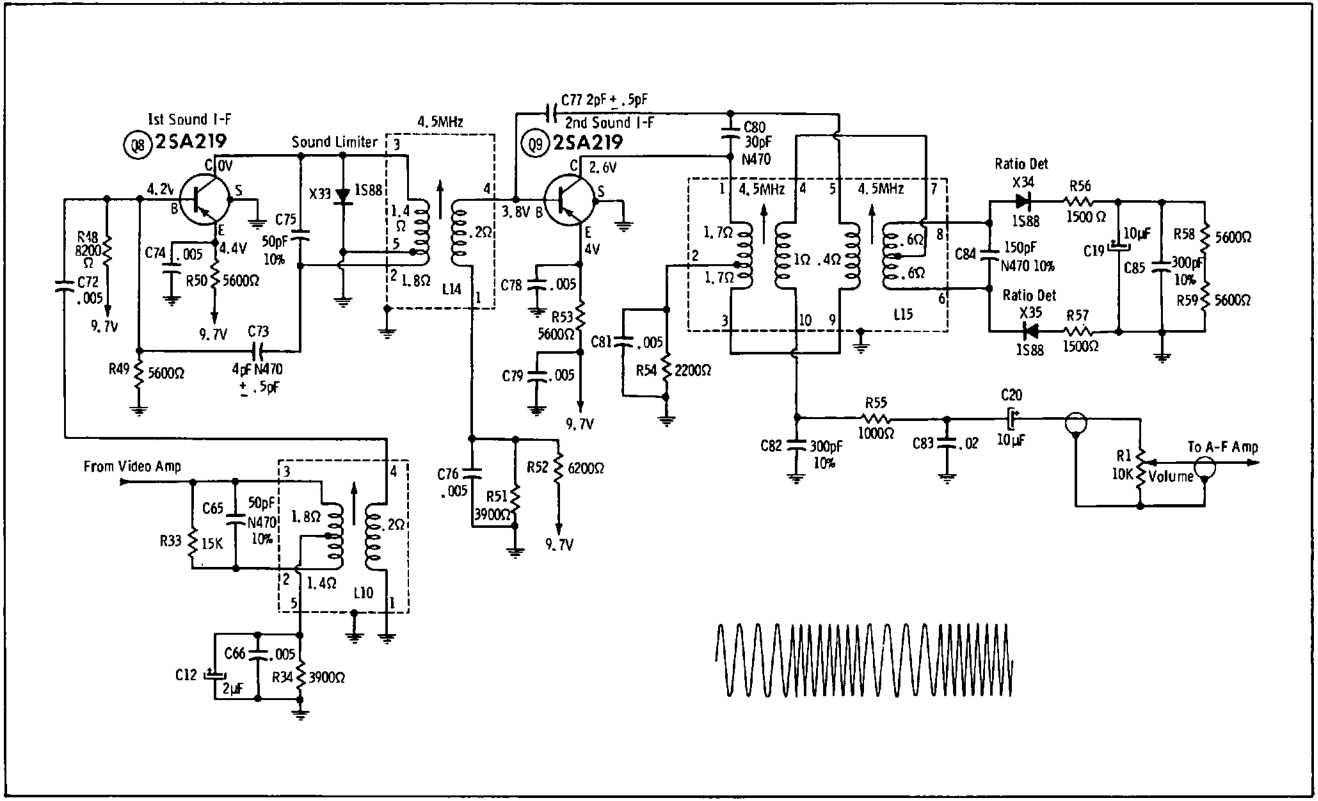


Fig. 7-7. Typical intercarrier sound i-f system.

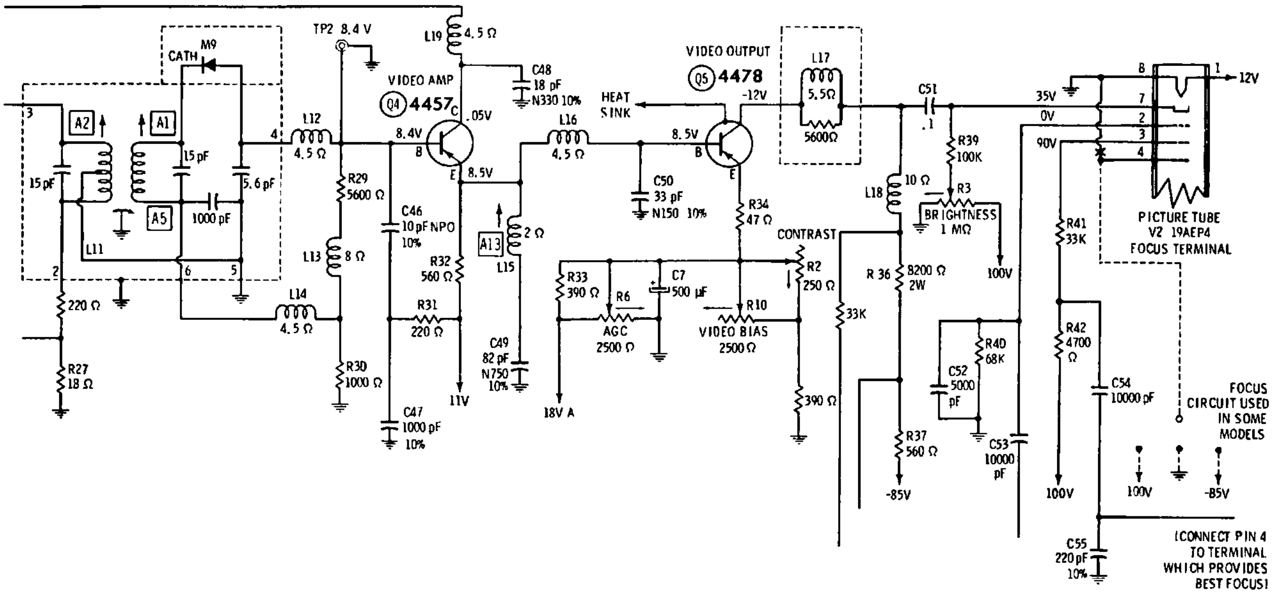


Fig. 7-8. Video-amplifier configuration.

for driving the video-output transistor. The total voltage gain of the system is 125 times in this example. A signal-tracing test not only shows the stage gain, but also shows distortion that may be occurring. For example, if a transistor is clipping the camera signal, the scope display appears as illustrated in Fig. 7-9. Poor high-frequency response produces rounding of the sync pulses. However, analysis of frequency response can be made to better advantage with a video-frequency sweep generator, or a square-wave generator.

A low-capacitance probe is also used to trace signals through the audio amplifier and the sync system. A TV station signal can be used, although generator signals are preferable because they are steady. If a TV analyzer is available that provides a

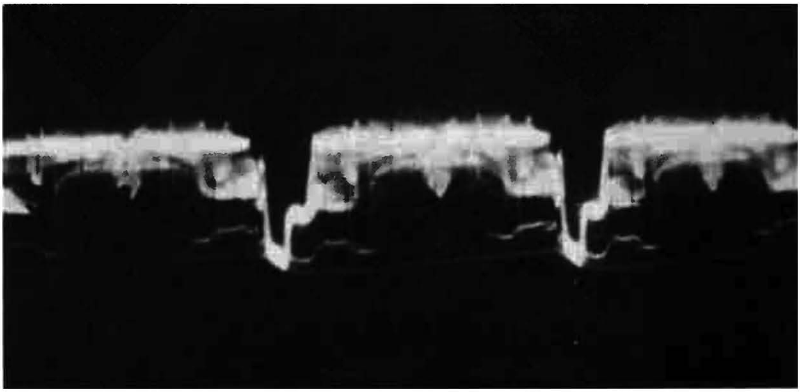


Fig. 7-9. Clipping of camera signal.

4.5-MHz tone signal, a sine waveform will normally be observed in the audio section. Otherwise, the audio output from an a-m generator, or from an audio oscillator, can be applied at the input of the audio section. To check the sync system, any pattern generator can be used. For example, a test-pattern generator, white-dot or crosshatch generator, or even a color-bar generator is suitable. It is desirable to use a generator that provides well-shaped sync pulses, so that a meaningful comparison can be made with the waveforms specified in receiver service data.

Fig. 7-10 shows a typical horizontal and vertical sync system. The sync-separator transistor Q12 is reverse-biased, and clips the sync tips from the composite video signal. This stage provides a voltage gain, and drives sync-phase inverter Q13, which provides double-ended output. R71 and C67 form an integrating circuit that accepts the vertical-sync pulses and re-

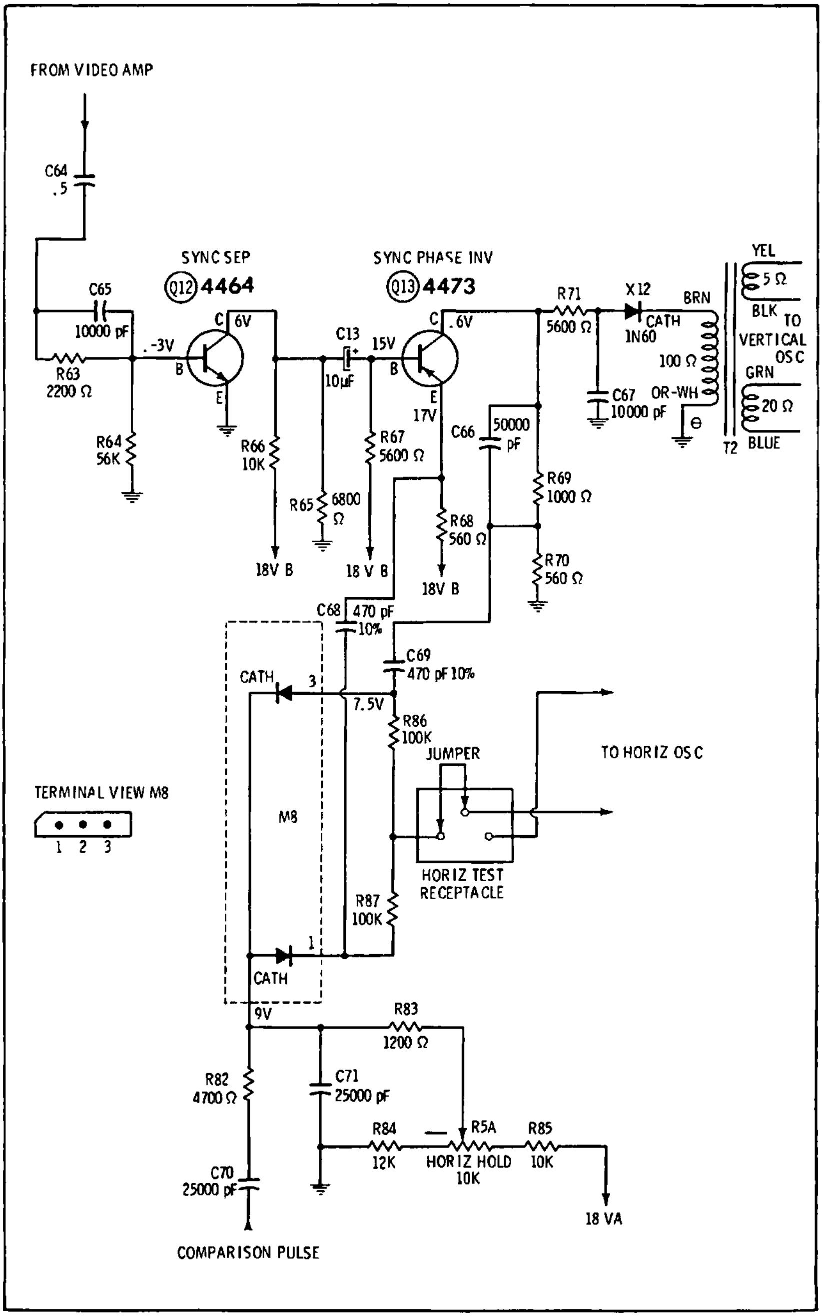


Fig. 7-10. Typical horizontal and vertical sync system.

jects the horizontal-sync pulses. To avoid confusion due to kickback from the vertical oscillator, the oscillator transistor is removed from its socket during the waveform check. The integrator normally introduces an insertion loss of 37 percent. Diode X12 is used to minimize reverse coupling from the vertical oscillator into the sync phase-inverter stage.

The foregoing waveforms are signal waveforms. However, the waveforms associated with M8 are combination waveforms; they comprise horizontal sync pulses and comparison pulses from the flyback section that have been shaped in sawtooth waves. Both the amplitudes and the shapes of these waveforms are of practical concern. In case of sync trouble, signal-tracing procedures are very helpful to localize the defective stage. Then, dc voltage and resistance measurements are generally utilized to close in on the defective component. Due to circuit interaction, trouble symptoms are not always clear. For example, if one diode in the sync phase-comparison circuit has a poor front-to-back ratio, all three waveforms will be distorted. Therefore, it is usually necessary to check the front-to-back ratios of both diodes.

WAVEFORM ANALYSIS

The waveforms in a TV receiver system are not sharply distinguished from signals, as noted previously. However, it is helpful to classify those ac voltages that are self-generated as waveforms. From this practical viewpoint, we find waveforms in the rf tuner, vertical and horizontal sections, and in a related way in the power supply system. The 117-volt 60-Hz line applies a sine waveform to the power supply, and this sine waveform is not classified as a signal. Instead, it is an applied power waveform. Again, with reference to Fig. 7-2B, the three ac voltages that are illustrated are called waveforms, and are not described as signals.

Analysis of the drive waveform to Q22 in Fig. 7-2C entails a measurement of its peak-to-peak voltage, repetition rate, pulse width, waveshape, and its rise and fall times. A scope calibrated in peak-to-peak voltage values is utilized to measure the amplitude of the waveform. Its repetition rate can usually be taken for granted because the symptoms of horizontal sync loss are generally prominent. However, in the case of a dark screen, measurement of repetition rate is necessary. That is, the amplitude and shape of the waveform might be normal, but if its repetition rate is greatly abnormal or subnormal, the horizontal-output system develops deteriorated circuit action.

If we know that the repetition rate of the drive waveform to Q22 in Fig. 7-2C is normal, we need not measure the pulse width directly. Instead, we can merely note the duty cycle and compare it with the duty cycle specified in the receiver service data. Direct measurement of pulse width requires a scope that has a calibrated time base. The waveshape can be evaluated with an ordinary service-type scope, provided that the scope has ample bandwidth to accommodate the higher harmonics in the waveform. The horizontal-output system will not operate normally if the rise and fall times of the drive waveform are too slow. Measurement of rise or fall time requires a scope that has triggered and calibrated sweeps. Since only a minority of service shops have this type of scope available, rise and fall times are generally omitted from receiver service data. That is, routine service work does not provide for this aspect of waveform analysis.

Key waveforms for a vertical-sweep system are illustrated in Fig. 7-11. Waveform analysis entails a measurement of peak-to-peak voltage, repetition rate, and waveshape. Rise and fall times are of comparatively little significance in this case, and are considered only under the classification of waveshape. The amplitude of a vertical-system waveform is of basic concern. Repetition rate has an important effect on amplitude, but we need not measure the waveform repetition rate unless the screen is dark. As long as the pattern is in vertical sync lock, the repetition rate will be correct. Waveshape is also of basic significance. Distortion of any type indicates a circuit defect that should be localized and identified. In the analysis of waveshape, the linearity of the drive waveform to Q18 is of basic concern. It is closely associated with vertical linearity in the picture. If the specified waveshape cannot be matched by adjustment of the vertical-linearity control, we know that a component defect is present.

Of course, there is a possibility of reflected trouble in the system of Fig. 7-11. For example, if there is an open damping resistor, a ringing interval will be generated in the output waveform. Due to feedback via T3, this ringing interval will also appear in the drive waveform to Q18. Therefore, waveform analysis necessitates the consideration of reflected trouble possibilities; otherwise, we can waste time looking for the defective component in the wrong places. Interpretation of distorted waveforms requires both study and practical experience. If we have a good understanding of circuit action, waveform analysis becomes much less mysterious than it seems to the beginner.

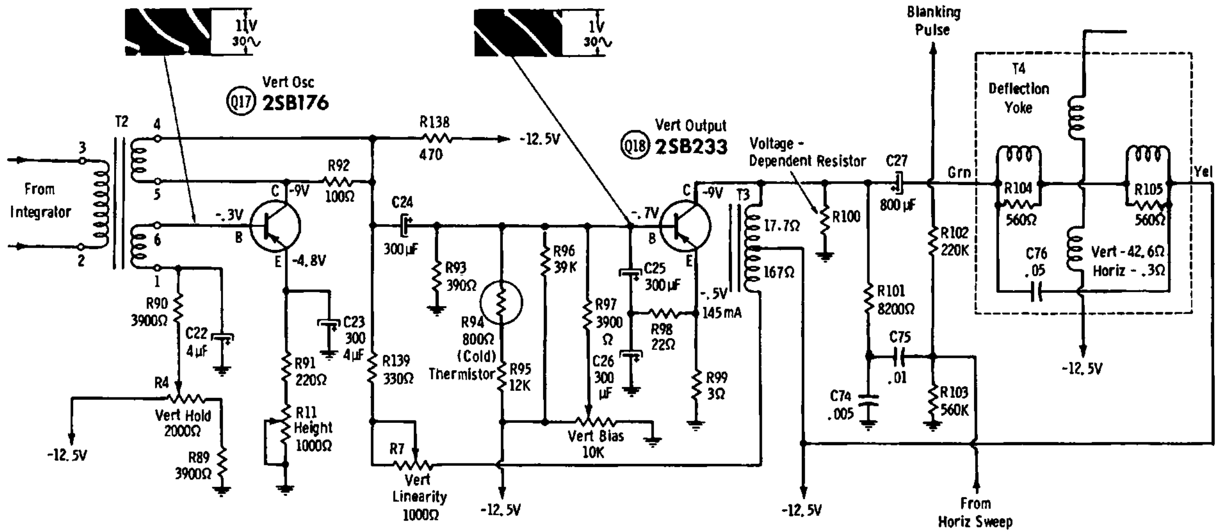


Fig. 7-11. Two-stage transistor vertical-sweep system.

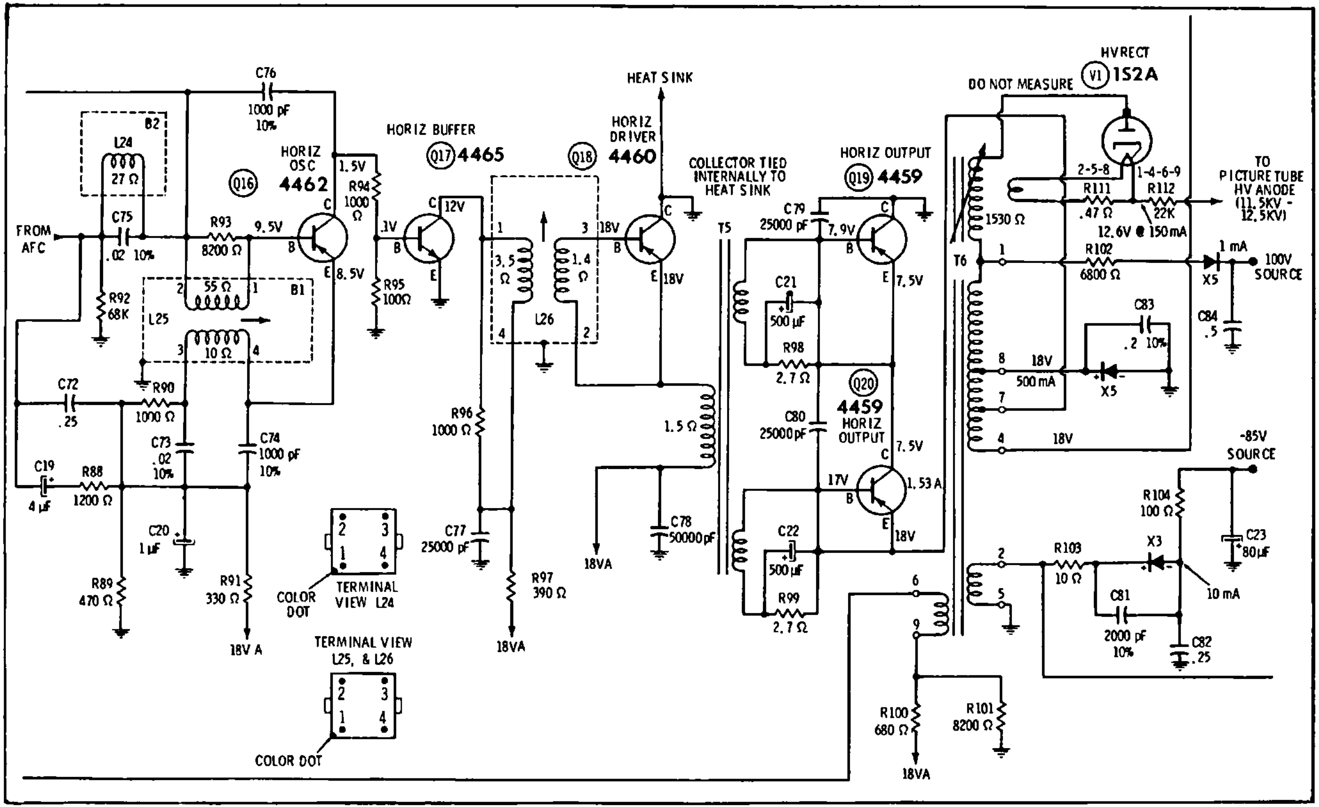


Fig. 7-12. Horizontal-output system.

Circuit interaction is also encountered in the horizontal-output system. For example, with reference to Fig. 7-12, a few shorted turns in T6 can change the amplitudes and shapes of many of the waveforms. The horizontal-hold control should first be adjusted so that the horizontal oscillator free-wheels at approximately 15,750 Hz. This permits us to determine whether the drive waveform to Q17 is correct. Since the transistor operates as a buffer, component defects in the output system will not affect the drive waveform under this condition of test. Leaky or open capacitors can produce waveform distortions that are easily confused with transformer defects. Since capacitors are the more common troublemakers, we check out the capacitors first in case of doubt.

Next, let us briefly consider a keyed-agc circuit, such as the one depicted in Fig. 7-13. We normally find a video waveform at the base of the transistor, such as that illustrated in Fig. 7-3A. A flyback pulse is introduced from the sweep section, which becomes shaped into a sawtooth waveform. This pulse is the collector-supply voltage, and the pulse is normally coincident with the horizontal-sync pulse applied to the base. Since the emitter is reverse-biased, Q7 can conduct only when the flyback pulse is present. The amount of pulse conduction depends on the amplitude of the horizontal-sync pulse. Since the sync pulse is amplified during the conduction interval, the collector waveform displays a combination of the shaped compar-

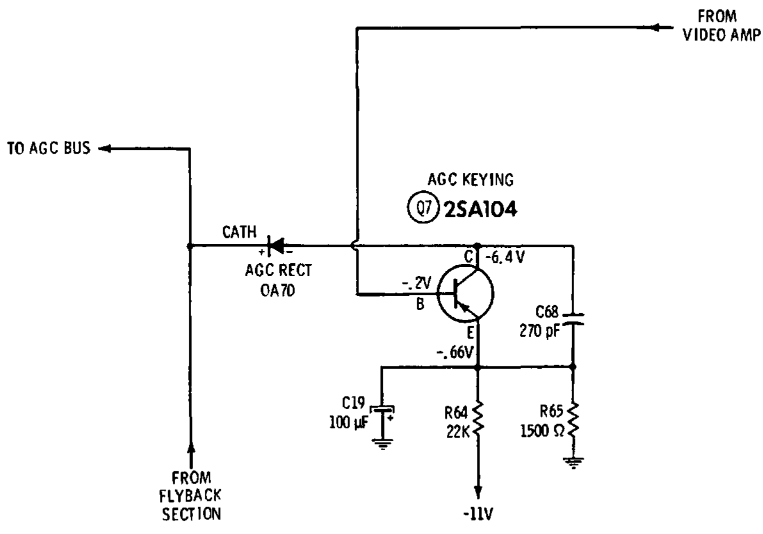


Fig. 7-13. Keyed-agc circuit.

ison pulse and the horizontal-sync pulse. Therefore, the picture must be in horizontal sync in order to make a valid waveform check.

Fig. 7-14 depicts a transistor power supply. The waveform of interest in this case is the ripple waveform on the −11.8-volt line. Its amplitude should not exceed the specified value—0.5-volt pk-pk in this example. An excessive amplitude points to a defective filter capacitor in most cases, although a leaky feedback-amplifier or current-regulator transistor can also cause excessive ripple amplitude. Rectifier bridge defects show up as subnormal dc voltage output. It is difficult to analyze the waveforms in the bridge-rectifier circuit unless a scope with balanced vertical input is available. Few service shops utilize this type of scope. Therefore, most component defects are localized by means of voltage and resistance measurements. It is usually necessary to disconnect one end of a rectifier for test.

FREQUENCY-RESPONSE CURVES

A comprehensive coverage of sweep- and marker-generator application cannot be included in this book. Instead, only the basic principles of response-curve analysis will be covered. Interested readers may refer to specialized alignment handbooks for procedure information and data. Of course, because of circuit variations, the receiver service data should always be consulted when a particular receiver is under test. Fig. 7-15 shows the configuration of a typical transistor tuner. A normal frequency-response curve is illustrated in Fig. 7-16A. We are concerned with the hump frequencies and the waveform amplitude. The hump frequencies are measured with a marker generator. Waveform amplitude can be evaluated only on the basis of experience with the sweep generator and scope that are used in the alignment procedure.

Fig. 7-16B shows how the picture-carrier and sound-carrier frequencies normally fall on the humps of the rf response curve. The humps may not be at exactly the same amplitude, but large discrepancies should be avoided. That is, a reasonably uniform 4-MHz band should be provided for the video-frequency spectrum, and the audio-frequency spectrum should not be excessively attenuated or boosted. Excessive tilt in the top of the rf response curve is associated with impaired picture quality. Because there are various active channels to contend with, compromise alignment is often necessary, but we try to approximate the specified frequency-response curve for each active channel.

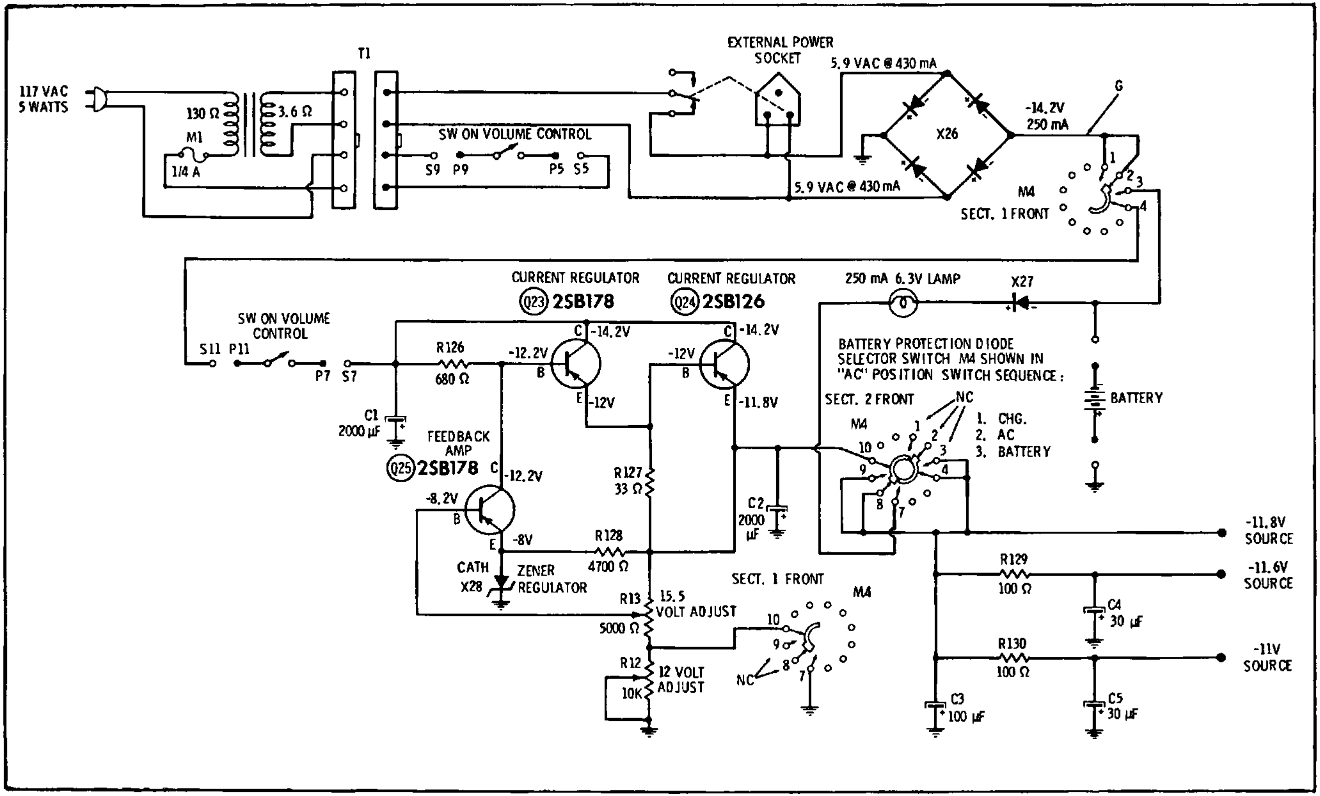


Fig. 7-14. Transistor power supply.

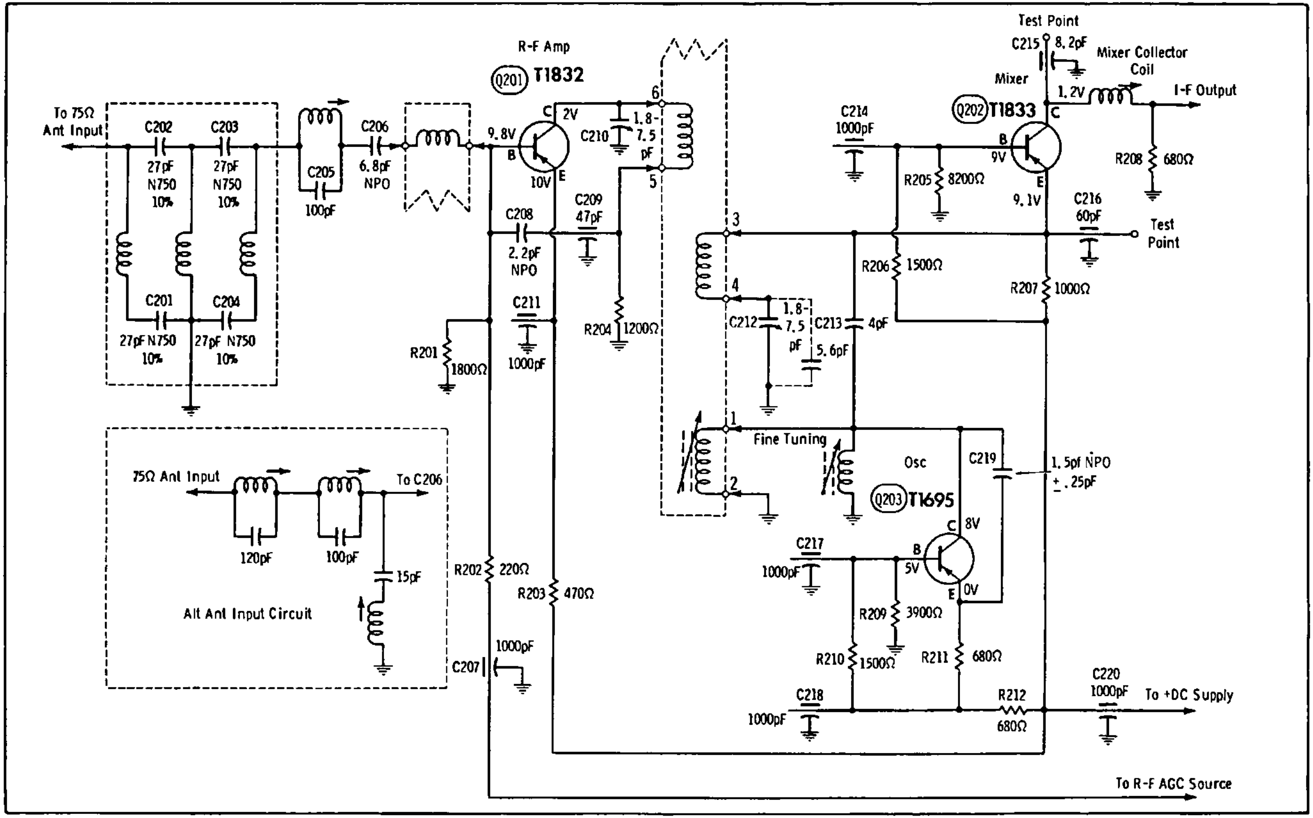
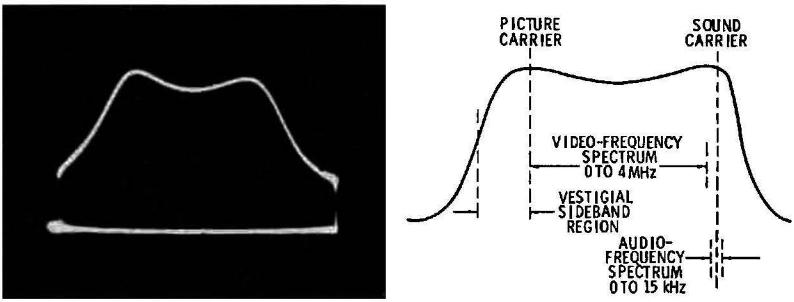


Fig. 7-15. Schematic of typical turret tuner.

A typical i-f amplifier strip is depicted in Fig. 7-6. Most of the receiver gain and selectivity is developed in the i-f system. Therefore, i-f alignment is a critical consideration in regard to picture quality. Alignment procedure and the specified i-f response curve for a typical transistor TV receiver are shown in Fig. 7-17. The chief points in curve analysis are the frequencies at the 50 percent-of-peak points, and the trap frequency. These are 26.75, 23.75, and 22.25 MHz, respectively, in this example. A specified curve also depicts the permissible sag at the midband frequency (25 MHz). The bandwidth in this example is 3 MHz. Reduced bandwidth is objectionable because it results in loss of picture detail. Increased bandwidth is desirable, if it can be obtained without developing objectionable sag.



(A) Waveform. (B) Signal distribution.

Fig. 7-16. Rf frequency response curve.

Careful attention should be directed to the sound traps. Unless the traps are adjusted correctly, sound interference is likely to be visible in the picture. Then, if the fine-tuning control is misadjusted to minimize sound interference, the picture quality will be impaired. Note that if a sound trap is mistuned and falls in the video-signal passband, either the bandwidth of the response curve will be reduced, or a "suck-out" will appear in the curve. Both distortions are objectionable from the standpoint of picture quality. If the specified response curve cannot be reasonably approximated, a defect will be found in the i-f system. Capacitors are the most common culprits.

A transistor video-amplifier configuration is shown in Fig. 7-8. The video-amplifier response curve can be checked with a video-frequency sweep generator and a scope. Fig. 7-18 illustrates a typical video-frequency response curve. Curve analysis concerns chiefly the bandwidth and the setting of the 4.5-MHz trap. Bandwidth between the 50 percent-of-peak points should be at least as great as that of the i-f amplifier; otherwise, the

Use an isolation transformer and maintain voltage at 117 volts. Allow a 20-minute warm-up period for the receiver and test equipment.
Suggested Alignment Tools: A1 thru A10 .... GENERAL CEMENT #8868, 8987, 9089 ... WALSCO #2531-X, 2541, 2587
Mixer Plate Coil .. GENERAL CEMENT #8284 .... WALSCO #2584

VIDEO IF ALIGNMENT

Connect the synchronized sweep voltage from the sweep generator to the horizontal input of the oscilloscope for horizontal deflection.
Use only enough generator output to provide a usable indication. Note: Response may vary slightly from those shown.
Connect a variable bias supply to the IF AGC line (point ⟨A⟩) and adjust to obtain a response curve which shows no indication of overload.
Disable Oscillator section of Mixer-Osc. Set the Channel Selector to any non-interfering channel.

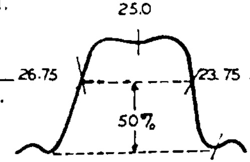
	INDICATOR	GENERATOR COUPLING	SWEEP GENERATOR FREQUENCY	MARKER GENERATOR FREQUENCY	ADJUST	REMARKS
1.	Connect DC probe of a VTVM thru a 47K resistor to point ⟨B⟩. Common to ground.	Connect high side to Point ⟨U⟩, low side to ground.		22.25MHz	A1	Adjust for MINIMUM.
2.	Connect DC probe of a VTVM thru a 47K resistor to point ⟨B⟩. Common to ground.	Connect high side to Point ⟨U⟩, low side to ground.		25.8MHz 23.5MHz 26.6MHz 23.8MHz	A2 A3 A4 A5	Adjust for maximum.  FIG. 1
3.	Connect vertical input of a scope to point ⟨B⟩. Low side to ground	Connect high side to Point ⟨U⟩, low side to ground.	25MHz (10MHz Sweep)	22.25MHz 23.75MHz 25.0 MHz 26.75MHz	Mixer Plate Coil	Adjust for maximum gain and symmetry of response with markers as shown in Figure 1. In order to obtain a proper response, it may be necessary to slightly retouch A2, A3, A4 and A5.

Fig. 7-17. Typical i-f alignment procedure.

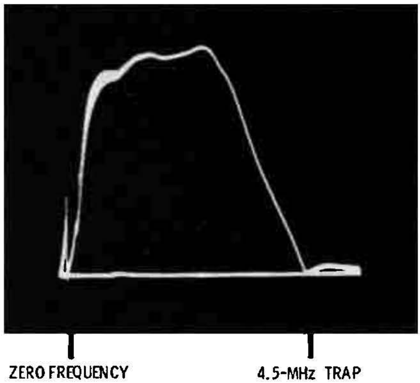


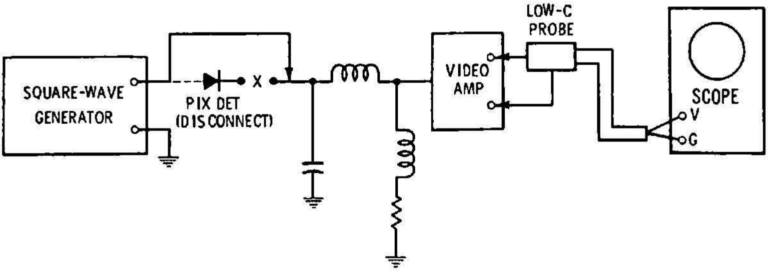
Fig. 7-18. Typical video-amplifier response curve.

picture detail will be attenuated. Absorption markers are generally preferred to beat markers in this procedure, because spurious markers are often produced by beat harmonics, and can be very troublesome. Absorption markers do not produce spurious markers.

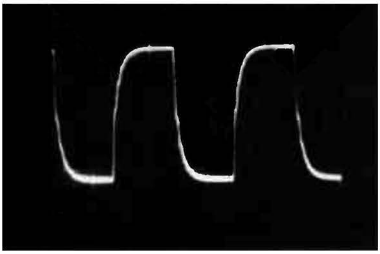
SQUARE-WAVE TESTS

Since many service shops do not have video-frequency sweep generators available, a square-wave test is often preferred. A square-wave test is made as shown in Fig. 7-19. It is standard practice to use a 100-kHz square wave, although other repetition rates are also useful in waveform analysis. The shape of the reproduced square wave is chiefly dependent on the peaking-coil inductances and damping-resistor values. However, the values of the load resistors also affect square-wave response. The square-wave response shown in Fig. 7-19B is typical of economy-type receivers. Deluxe receivers have somewhat better transient response.

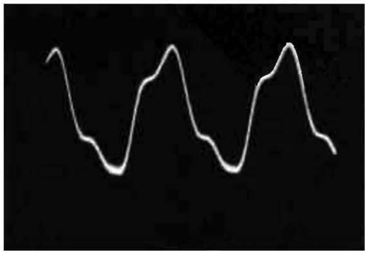
Objectionable waveform distortions include excessive overshoot, which may be accompanied by ringing. An overshoot of 10 percent can be tolerated. Diagonal corner rounding is seen in Fig. 7-19B; excessive corner rounding points to out-of-tolerance peaking coils, and should be corrected. Overshoot and ringing are associated with excessive gain at high frequencies, which in turn is associated with load resistors that are too small in value. Open damping resistors can also cause overshoot and ringing. If a scope is available that has a triggered and calibrated time base, the rise time of the reproduced square wave can be measured. A video amplifier that has a bandwidth of 4 MHz has a rise time of approximately 0.08μs.



(A) Test setup.



(B) 100-kHz waveform.



(C) 1-MHz waveform.

Fig. 7-19. Checking square-wave response of a video amplifier.

In any square-wave test, it should be remembered that the scope must have better response than the video amplifier under test. Otherwise, distortions introduced by the scope will be falsely charged to the video amplifier. On the other hand, in frequency-response tests, it is necessary that the scope merely have good 60-Hz square-wave response. This is because the sweep signal that is used in a frequency-response test is demodulated, and the wave envelope has a 60-Hz repetition rate.

REVIEW QUESTIONS

1. Name the sections in a TV receiver that generate waveforms.
2. Why is it impractical to display the output waveform from a uhf oscillator on a scope screen?
3. How is the video i-f signal processed by the picture detector?
4. Describe a signal-tracing procedure.
5. Why must a demodulator probe be used in signal-tracing an i-f amplifier?
6. Explain how a video amplifier may clip the camera signal.

7. What is meant by an intercarrier sound signal?
8. Describe incidental frequency modulation.
9. Why must a low-capacitance probe be used in checking various waveforms?
10. Give an example of circuit interaction.
11. Explain the operation of a signal-biased sync clipper.
12. State some characteristics that are entailed in waveform analysis.
13. Give an example in which rise time is a critical characteristic.
14. When would the repetition rate of a waveform be of concern?
15. Discuss a requirement for waveform linearity.
16. Why are waveforms with excessively high voltages avoided in scope tests?
17. Describe the key waveform in an agc circuit.
18. What is the chief waveform of interest in a power-supply system?
19. Explain what is meant by a frequency-response curve.
20. Which sections of a TV receiver are checked for frequency response?
21. What characteristics do we look for in an rf response curve?
22. What characteristics do we look for in an i-f response curve?
23. Name two ways in which a video amplifier can be checked.
24. Discuss the requirements in adjusting a video amplifier for good frequency response.
25. Explain how component defects affect the square-wave response of a video amplifier.

8

Transistor Color-TV Circuits and Waveforms

A color-TV receiver contains all the circuit sections found in a black-and-white receiver, plus the chroma processing, color sync, and convergence circuitry. Fig. 8-1 shows the block diagram for a typical color-TV receiver. As far as the duplicated black-and-white sections are concerned, there are certain minor distinctions to be noted, as follows:

1. Tuner frequency-response curves are generally held to somewhat tighter specifications than in monochrome receivers.
2. Similarly, i-f response curves are held to comparatively tight specifications.
3. A delay line is included between the video-amplifier and video-output stages. This delay line normally introduces negligible waveform distortion, and delays the video signal for approximately .9μs.
4. The video-amplifier system often provides phase compensation and provides better square-wave response than its monochrome counterpart.
5. A 3.58-MHz color-subcarrier trap is included in the video amplifier (trap is not shown in Fig. 8-1).
6. Separate intercarrier-sound and picture detectors may be employed.

Since waveform analysis is basically the same in the monochrome sections of both color and black-and-white receivers, the reader is referred to receiver service data for these sections. This chapter explains waveform analysis in the chroma

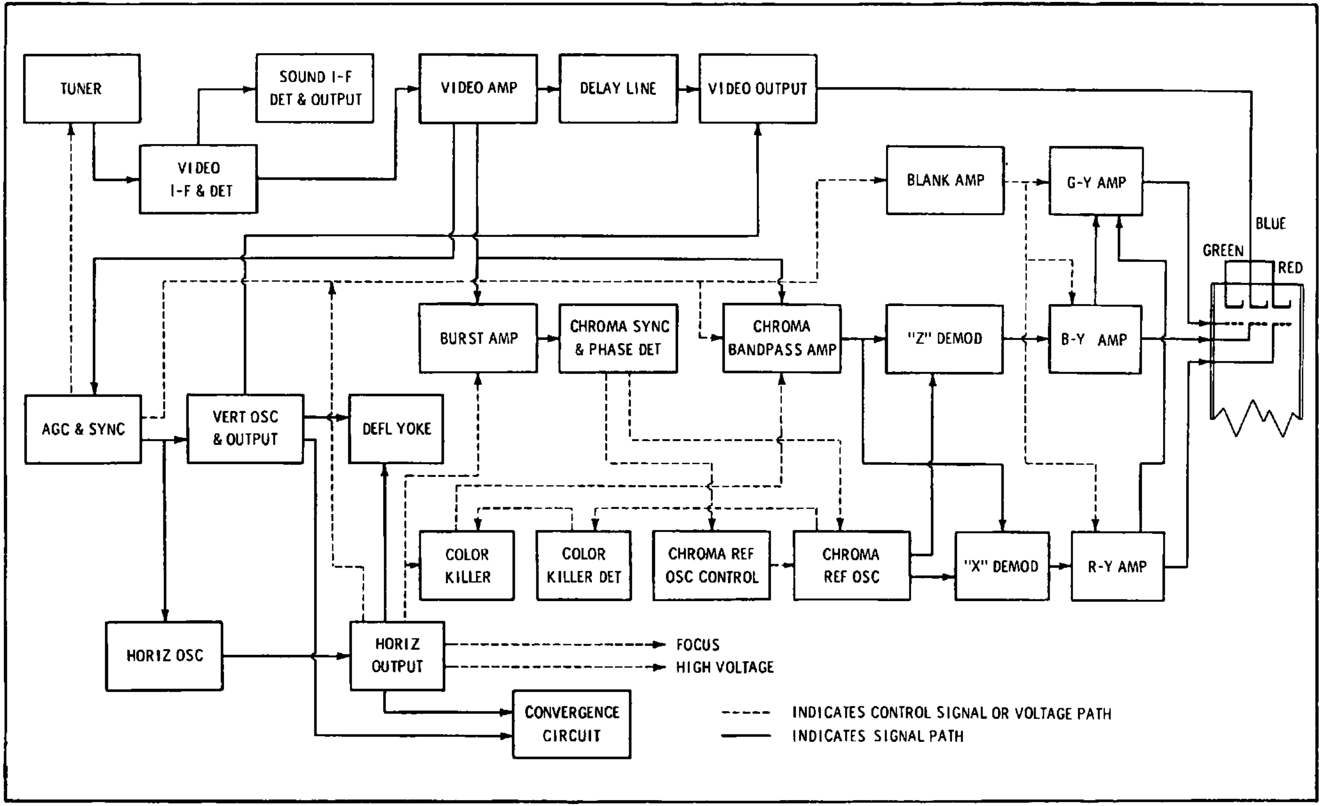


Fig. 8-1. Block diagram of color-television receiver.

processing and color-sync sections of transistor color-TV receivers. We will find that the basic chroma waveforms are quite different from those that have been discussed previously, although the general principles of scope application are the same. It is essential to understand the circuit actions in the chroma sections to interpret the waveforms effectively.

GENERAL SURVEY OF THE COLOR CIRCUITRY

Fig. 8-2 shows a typical chroma signal-processing and color-sync configuration. The color signal is first applied to a chroma bandpass amplifier, which is also called a color i-f amplifier. This chroma section has the function of effectively separating the chroma signal from the black-and-white signal, as explained in greater detail later. From the color i-f amplifier, the chroma signal is fed to the three chroma demodulators. These are combination phase and amplitude demodulators, which operate as product detectors; note that a 3.58-MHz color-subcarrier signal is fed into the chroma demodulators from the color oscillator. The chroma demodulators also function as matrices in this example, by mixing the demodulated chroma signal with the monochrome signal.

Other subsections include the color killer, which operates as an electronic switch. In the absence of a chroma signal (during reception of a monochrome signal), the color killer disables the chroma circuitry, so that colored snow (confetti) cannot be displayed on the picture-tube screen. Another subsection is an automatic chroma-control (ACC) circuit that operates as an agc system for the chroma signal. Still another subsection is the 3.58-MHz color oscillator, which contains an associated color-sync function. Its purpose is to reconstitute the color subcarrier signal, and to maintain its frequency and phase in step with the color subcarrier that is suppressed at the color-TV transmitter.

CHROMA WAVEFORM FUNDAMENTALS

The complete color signal is a multiplexed waveform in which the chroma information is encoded on a 3.58-MHz color subcarrier. This subcarrier itself is suppressed at the color-TV transmitter, and only the chroma sidebands are transmitted. This is done to minimize crosstalk between the chroma information and the monochrome information. As noted previously, this color subcarrier (a 3.58-MHz sine wave) is reinserted by the color receiver. A complete color signal with its encoded

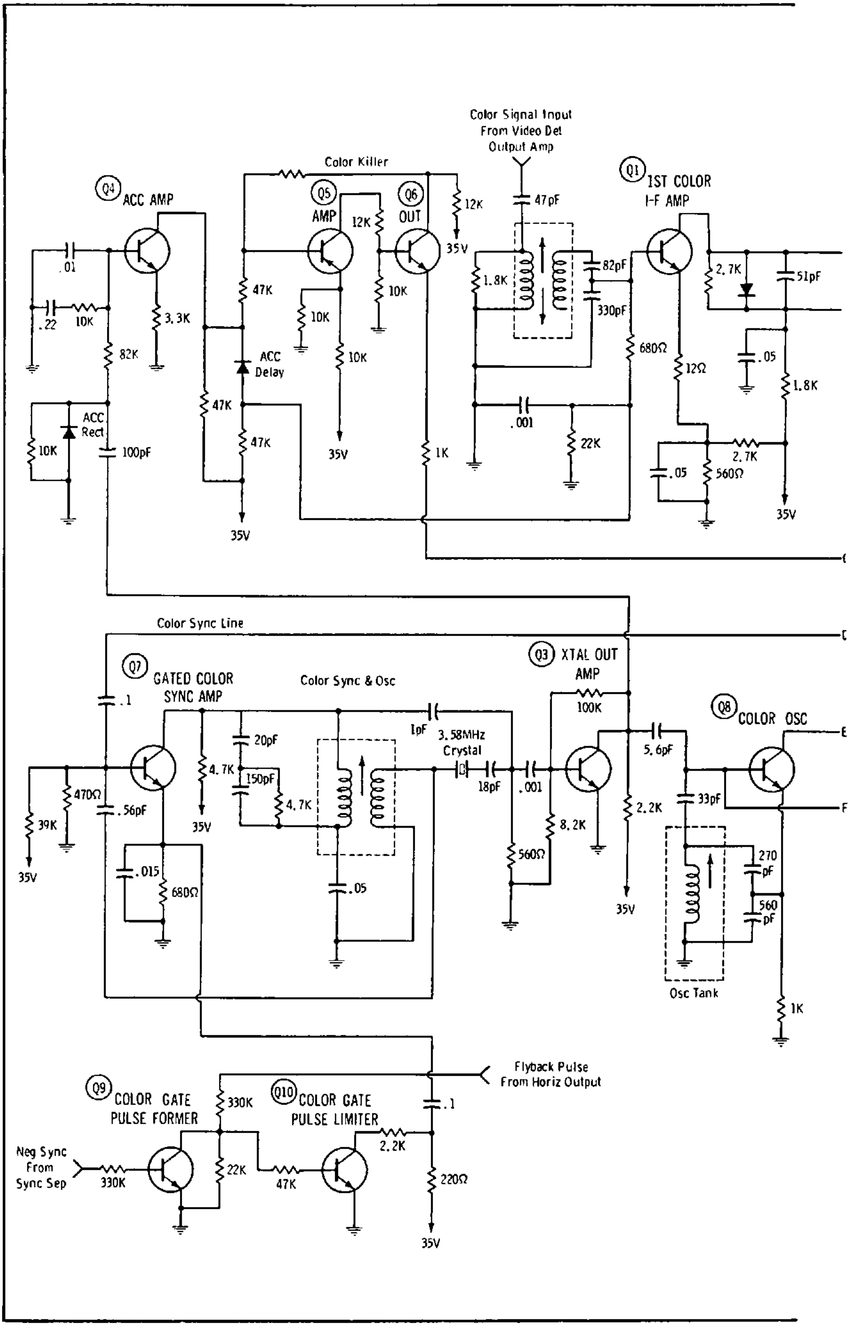
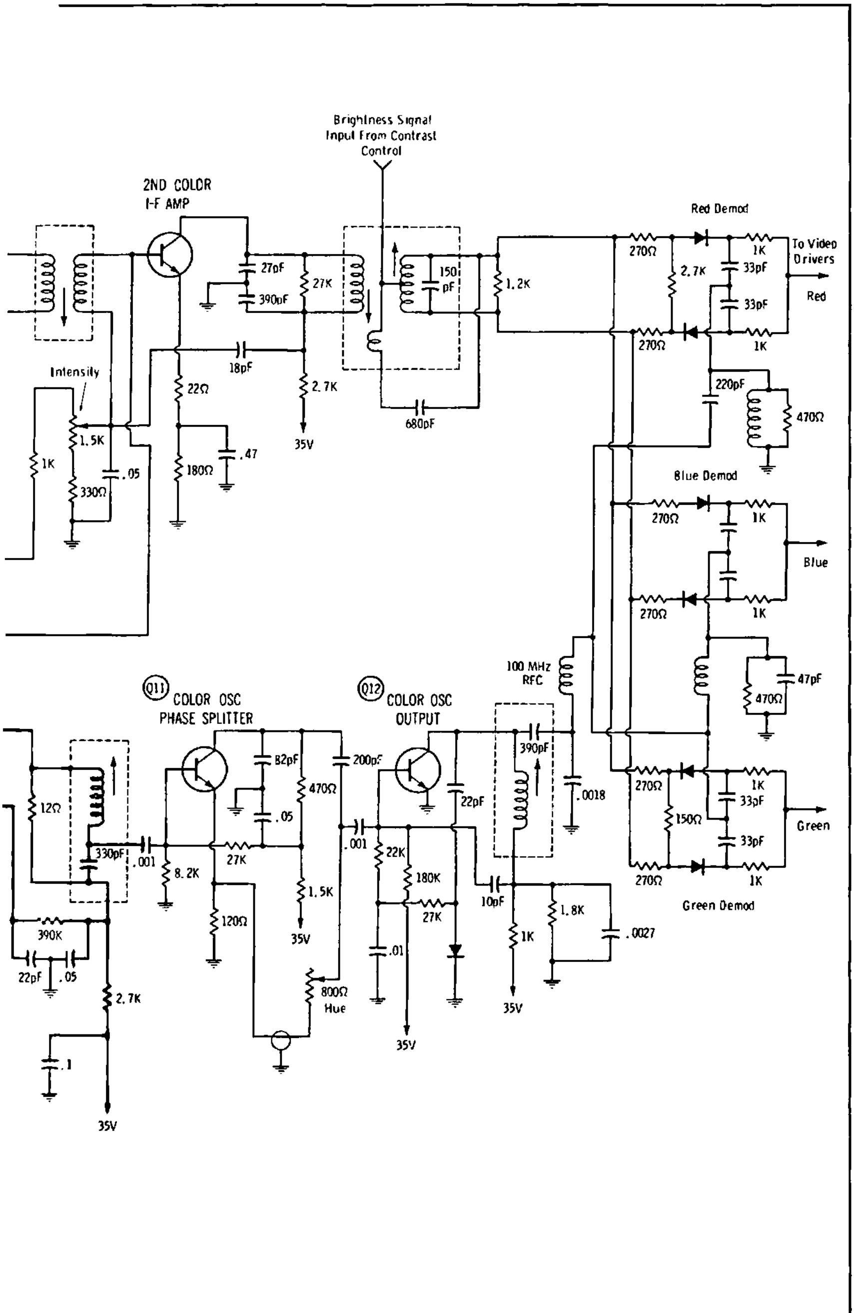
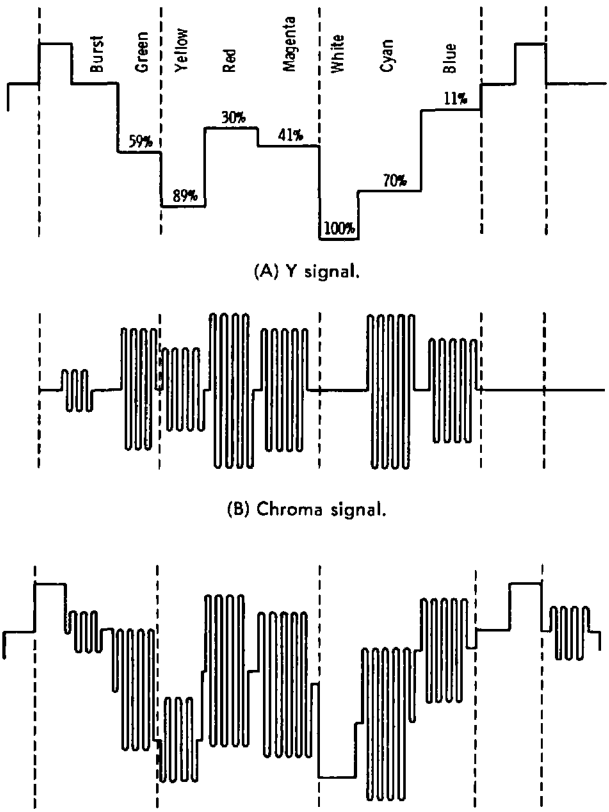


Fig. 8-2. Chroma signal processing



and color-sync configuration.



(A) Y signal.

(B) Chroma signal.

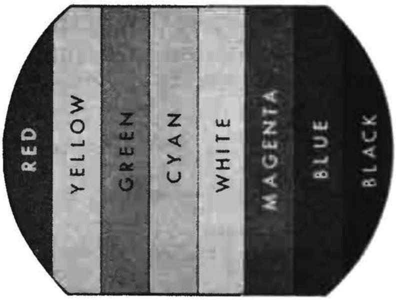
(C) Saturated color-bar signal.

Fig. 8-3. Elements of NTSC color bars.

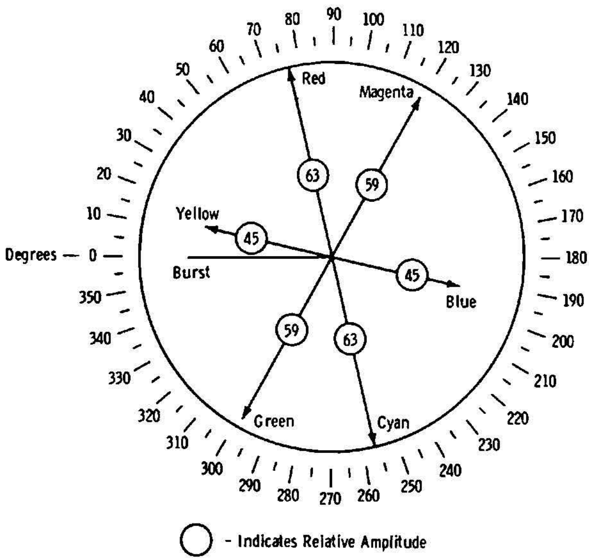
chroma components is depicted in Fig. 8-3. In this example, the chroma information comprises color bars, consisting of the primary colors, complementary colors, and white. Color sync is accomplished by means of the color burst, which appears on the back porch of the horizontal sync pulse.

Note that the chroma signal is characterized by phase and amplitude variations, as depicted in Fig. 8-4. A color bar displayed on the picture-tube screen has the properties of brightness, hue, and saturation. Brightness is determined by the amplitude of the Y signal (Fig. 8-3). Hue is determined by the phase of the chroma signal; saturation is determined by the amplitude of the chroma signal. The Y signal is the same as the monochrome camera signal. It appears at the output of the picture detector in a color receiver, and is commonly fed through the video and Y amplifiers to the cathodes of the color picture tube, as shown in Fig. 8-1. In some receivers, the Y amplifier may be fed to the center tap on the secondary of the chroma-demodulator transformer, as depicted in Fig. 8-2. The result is the same in either case.

The chroma signal also appears at the output of the picture detector; however, it is trapped out and does not pass through the Y amplifier. Instead, the chroma signal is fed to the chroma bandpass amplifier, and then to the chroma demodulators. Although terminology differs, we often call the video-amplifier section past the delay line the Y amplifier, and the section prior to the delay line the video amplifier. Thus, the video sec-



(A) Screen pattern.



(B) Vector relationships.

Fig. 8-4. NTSC generator output.

tion carries both Y and chroma signals, while the Y section carries the Y signal only. Chroma phases are measured with respect to the burst phase. Color-TV transmission is a quadrature system in which two phases, I and Q, of the subcarrier are modulated as shown in Fig. 8-5.

Although I and Q demodulation is possible, it is comparatively complicated, and practically all modern color receivers demodulate the chroma signal on axes other than the I and Q phases. This is possible because the I and Q signals have components on any axis that might be selected. Regardless of the demodulation axes that are used, every chroma demodulation system must produce output signals that have the R − Y, B − Y and G − Y phases, as depicted in Fig. 8-6A. These three signals are required for operation of the color picture tube. The R − Y signal is fed to the red grid, the B − Y signal to the blue grid, and the G − Y signal to the green grid.

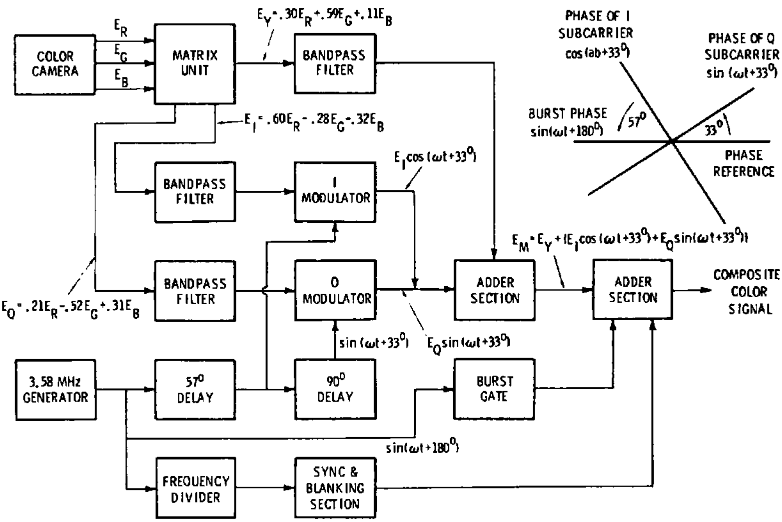


Fig. 8-5. Partial block diagram of color-TV transmitter.

CHROMA SIGNAL PROCESSING

In a basic chroma demodulation system, R − Y and B − Y demodulators are used, and a chroma matrix is utilized to form the G − Y signal. The chroma matrix mixes the R − Y and B − Y signals in suitable polarities and amplitudes. Note in Fig. 8-2 that the chroma demodulators also operate as matrices; however, these are not chroma matrices, but RGB matrices.

This simply means that the demodulators outputs are red, blue, and green signals, as depicted in Fig. 8-4. We also find X and Z demodulation axes in common use, as shown in Fig. 8-6B. After the X and Z demodulators, R, G, and B amplifiers are used; these amplifiers operate as matrices, and their outputs are R − Y, B − Y, and G − Y signals. Thus, the RGB terminology might be misleading in this case, and the beginner should carefully note the circuit action.

Some color receivers use an R − Y demodulator, a B − Y demodulator, and a G − Y demodulator. Fig. 8-2 is a typical example. It follows from Fig. 8-5 that an analogous color receiver

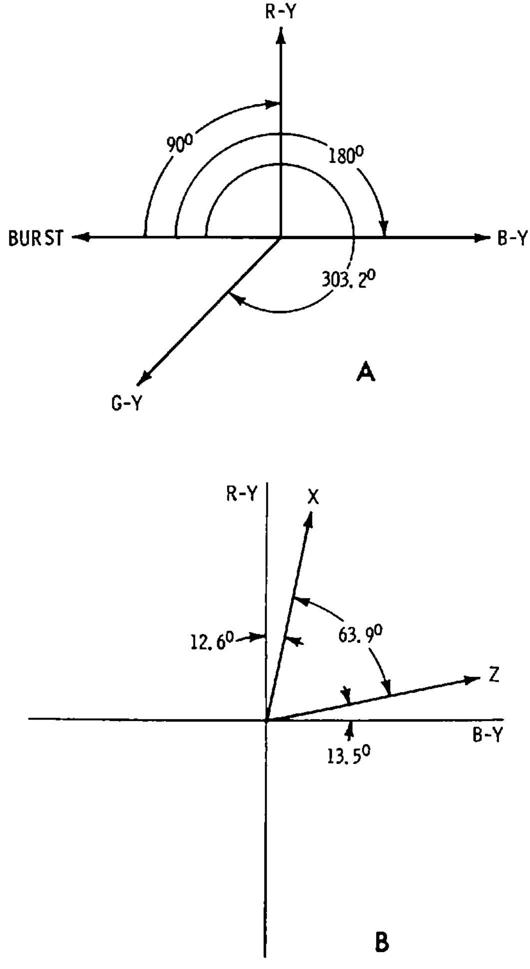


Fig. 8-6. Standard demodulation axes.

will have a chroma demodulator with an input consisting of an I signal and a Q signal. There is a phase separation of 90 degrees between these two signals. The receiver will also have another chroma demodulator with the same I and Q signal input. The 3.58-MHz color subcarrier is also fed from the subcarrier oscillator into each of the chroma demodulators. However, the I demodulator operates with an injected subcarrier that has the I phase, and the Q demodulator operates with an injected subcarrier that has the Q phase. The I and Q signals are separated in the process of demodulation.

Waveform checks will show that injected 3.58-MHz subcarrier sine waves have comparatively large amplitudes; thus, the peak of the injected waveform drives the demodulator transistor (or semiconductor diodes) into conduction for a small interval of time. Thus, the demodulator transistor or diode is cut off most of the time, and conducts briefly at the peak of the subcarrier waveform. This is just another way of saying that the I and Q signals are sampled by the chroma demodulators. Sampling serves to separate one of the chroma signals from its companion quadrature signal. With reference to Fig. 8-7, a summary of the principal chroma demodulator arrangements is presented. All of these systems employ the same circuit actions in various ways, and the result is the same for all.

The basic principle of chroma demodulation is seen in Fig. 8-8. There is a 90-degree phase difference between the I and Q signals. Therefore, when the I signal reaches its peak value, the Q signal is going through zero. Since the I demodulator is driven into conduction on the peak of the I signal, as shown at A by the heavy bar, only the I signal produces an output voltage during the conduction interval. Conversely, since the Q demodulator is driven into conduction on the peak of the Q signal, as shown at A by the light bar, only the Q signal produces an output during the conduction interval. Thereby, the two chroma signals separate the I and Q signals by a sampling process. Note that the I signal might be positive during the sampling interval (B), or it might be negative during the sampling interval (D). Similarly, the Q signal might be positive (C), or it might be negative (E).

This is called quadrature, or two-phase demodulation. It follows from Fig. 8-6A that if R − Y and B − Y demodulators are used, they operate as quadrature demodulators. However, X and Z demodulators are modified quadrature demodulators; as seen in Fig. 8-6B, the X and Z phases are separated by only 63.9 degrees. Complete separation of the X and Z phases is accomplished by sampling the X signal as the Z signal goes through zero, and sampling the Z signal as the X signal goes through zero. Normal operation of a chroma demodulator depends on adjusting the phase of the injected 3.58-MHz oscillator signal correctly, so that the desired signal is separated cleanly. This is done on the basis of waveform checks, as will be explained.

Let us consider the "phasing" of an R – Y and B – Y demodulator system. We apply simultaneous R – Y and B – Y test signals from a color-bar generator to the receiver; a scope is connected in turn at the demodulator outputs to check the waveform of the demodulated signal. Fig. 8-9 shows results of these tests. The demodulators should separate the R – Y and B – Y signals almost completely. If the demodulators are phased correctly, a single-bar (square-wave) signal is observed at each output. Incorrect adjustment results in the display of two bars

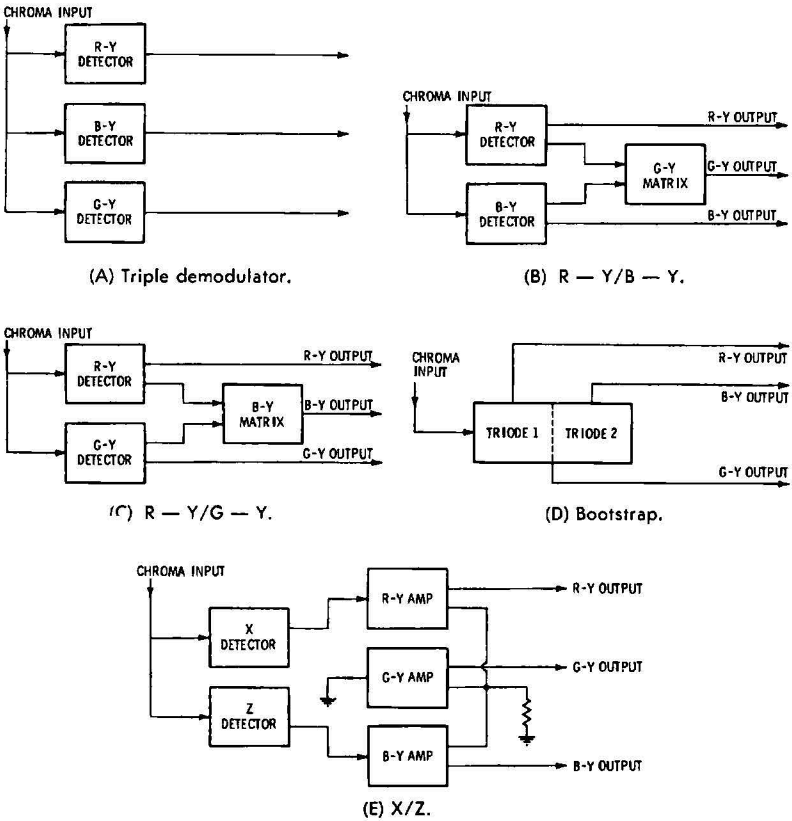


Fig. 8-7. Common demodulator systems.

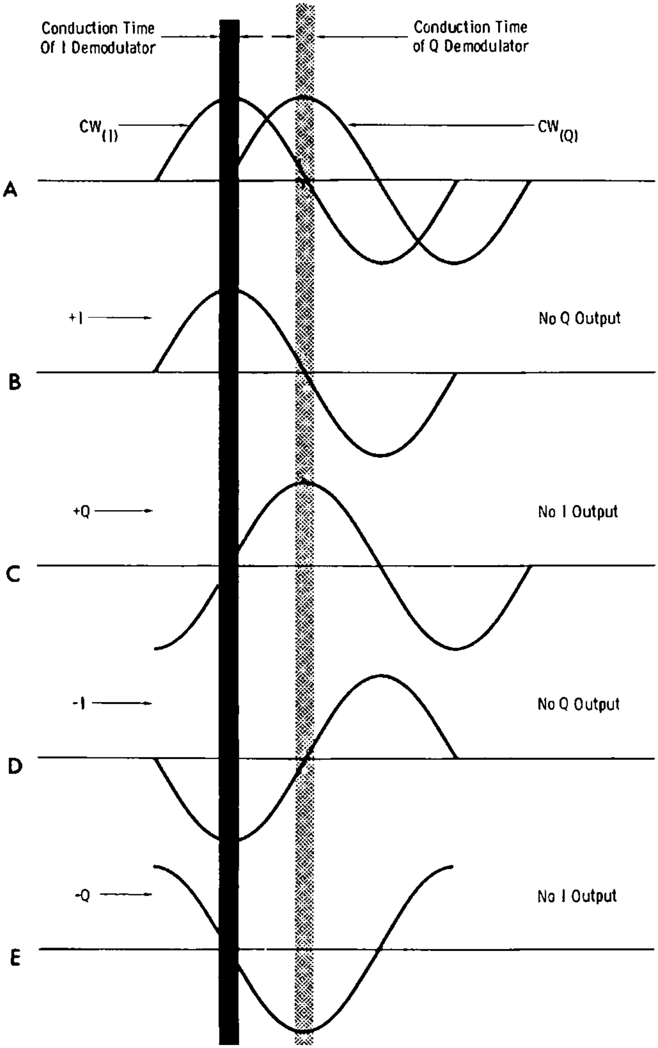


Fig. 8-8. Sampling I and Q signals.

(square waves), which may have various relative amplitudes, at the output of each demodulator. If we are checking an R − Y demodulator, the desired output waveform appears as shown in Fig. 8-9B. The waveforms in A and C represent incorrect phase adjustments.

A chroma matrix (Fig. 8-7) is checked in the same basic way as a chroma demodulator. That is, we apply a quadrature signal to the matrix, and check to see if a null pattern is displayed on the scope screen. Chroma demodulators and chroma matrices can also be checked by means of a keyed-rainbow signal. A keyed-rainbow signal provides 10 chroma bars, and an eleventh bar that serves as a color burst. Each bar has a phase advance of 30 degrees. A keyed-rainbow waveform, and its corresponding bar pattern is shown in Fig. 8-10. An R − Y demodulator normally nulls on the sixth bar; a B − Y demodulator on the third and ninth bars; a G − Y demodulator or matrix normally nulls on the first and tenth bars. An I demodulator would normally null on the fifth bar, and a Q demodulator on the second and eighth bars.

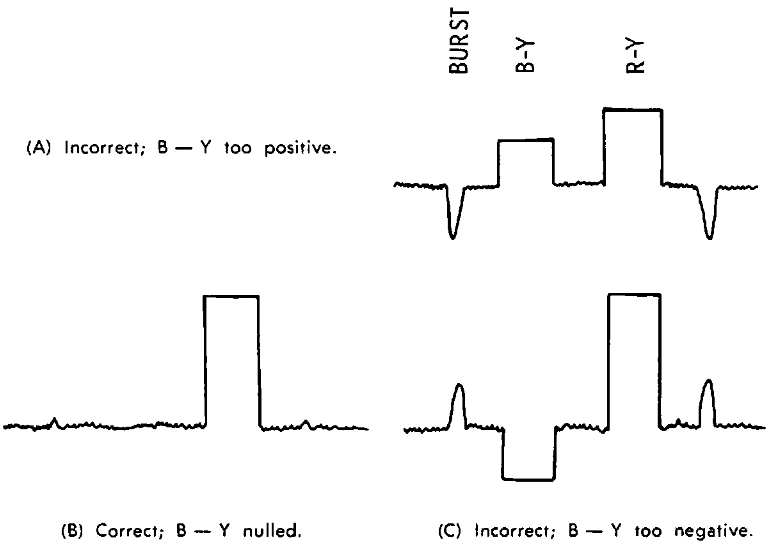
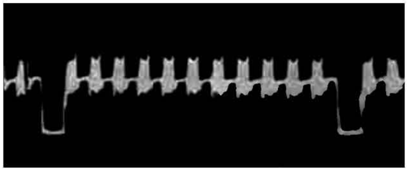
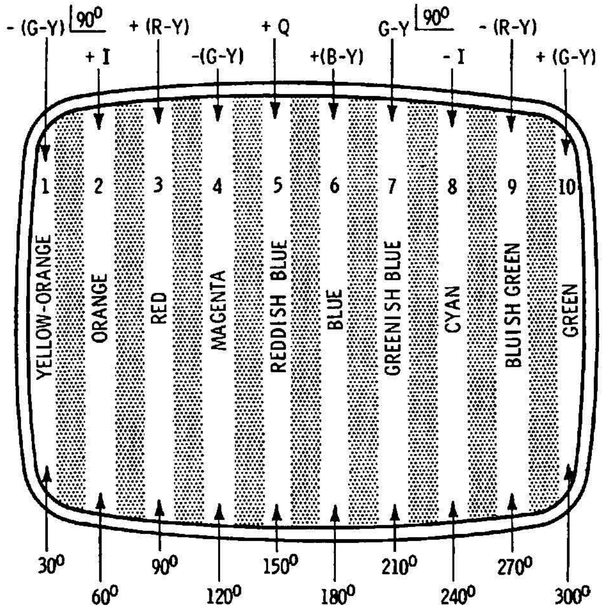


Fig. 8-9. Nulling B — Y output.

In other words, if we apply a keyed-rainbow signal to a color receiver, the waveforms at the outputs of the various chroma channels normally display the nulls depicted in Fig. 8-11. Note that the chroma demodulators in Fig. 8-2 also operate as RGB matrices. Therefore, a definitive test is made by applying an NTSC color-bar signal to the receiver (Fig. 8-12). If there is no component defect present, the red bar is displayed on a scope at the output of the red demodulator with full amplitude. The blue, green, and cyan bars are nulled. The yellow, magenta, and white bars also appear in the "square wave" with the red bar, as seen in Fig. 8-13. Next, if the scope is connected at the output of the green demodulator, we normally observe two "square waves," as shown in Fig. 8-13. When the scope is connected at the output of the blue demodulator, we normally see the broad "square wave" depicted in Fig. 8-13.



(A) Keyed rainbow signal.



(B) Chroma phases or bar display.

Fig. 8-10. Rainbow signal generator.

Some color bar generators provide one signal at a time, as illustrated in Fig. 8-14. It makes no difference whether we test with one signal at a time, or with a simultaneous multibar signal. In the case of a multibar signal, we merely count the bar intervals to locate a given chroma phase, or a given color signal. A keyed-rainbow generator supplies a multibar signal, as shown in Fig. 8-10A. Similarly, a simultaneous NTSC generator supplies a multibar signal, as depicted in Fig. 8-12. An NTSC generator commonly supplies color-difference signals in pairs, as shown in Fig. 8-15.

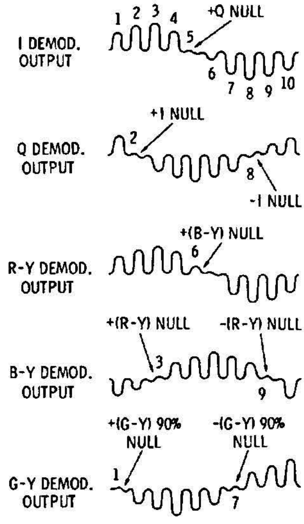


Fig. 8-11. Nulls of keyed rainbow signal.

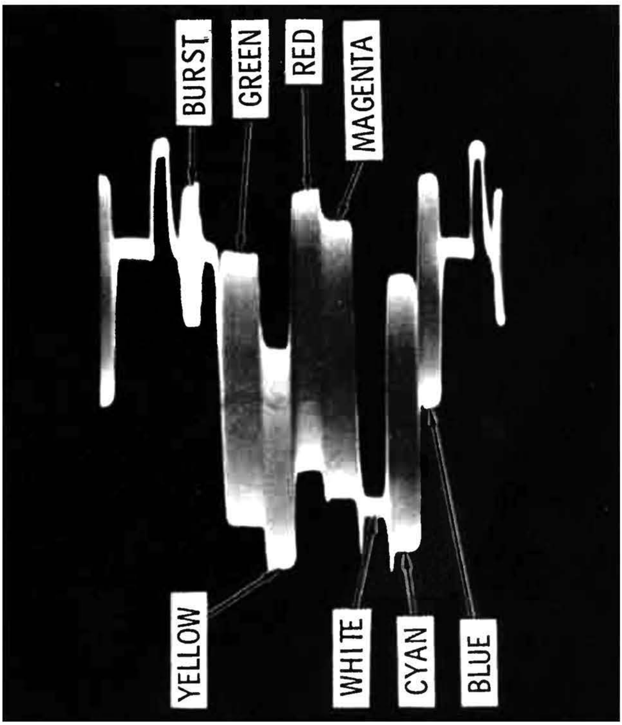


Fig. 8-12. The color signal from color-bar generator.

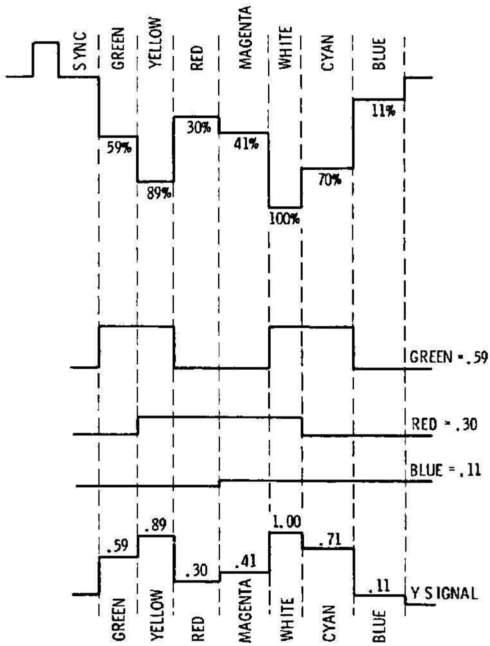
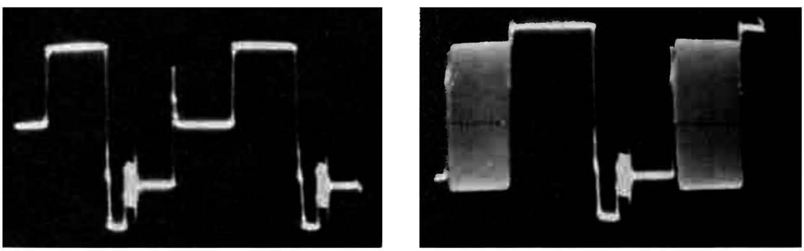


Fig. 8-13. Y component of NTSC signal.

The basic distinction between a keyed-rainbow generator and an NTSC generator is seen in Figs. 8-10 and 8-12. In a keyed-rainbow generator, the chroma signals are all lined up at black level. That is, the keyed-rainbow generator provides color-difference signals only. On the other hand, in an NTSC generator, the chroma signals are placed on their respective monochrome (Y) levels to provide true and fully saturated primary and complementary colors. When an NTSC generator is set to supply color-difference signals (Fig. 8-15), the



(A) Y signal only. (B) Complete signal (Y and chroma).

Fig. 8-14. Single-bar NTSC output.

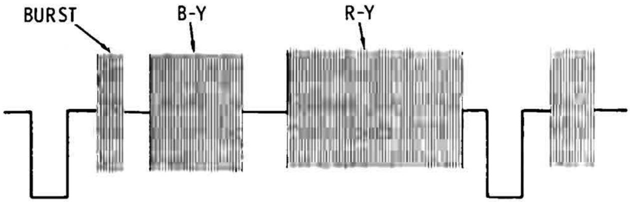
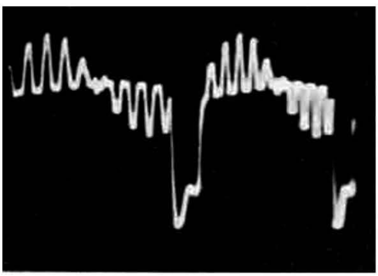


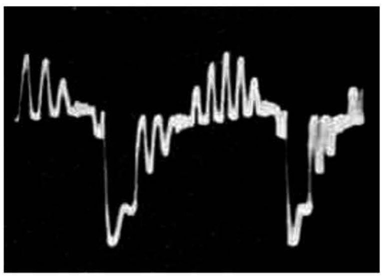
Fig. 8-15. R-Y/B-Y waveform supplied by NTSC generator.

waveform is basically the same as provided by a keyed-rainbow generator. Most NTSC generators provide R − Y, B − Y, I, and Q signals. Some NTSC generators also provide quadrature G − Y signals.

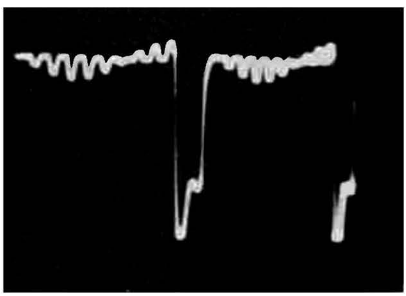
A keyed-rainbow generator does not supply pure and saturated color signals. However, this is of no concern in practical servicing procedures, because we are concerned only with reproduced waveforms such as those depicted in Fig. 8-11. These waveforms have been somewhat idealized for the sake of clarity. In actual practice, we analyze waveforms such as illustrated in Fig. 8-16. If you compare these waveforms with the diagrammatic representation in Fig. 8-11, you will observe that the same nulls are indicated. We analyze the foregoing



(A) R — Y signal.



(B) B — Y signal.



(C) G — Y signal.

Fig. 8-16. Normal keyed-rainbow signals from R — Y, B — Y, and G — Y circuits.

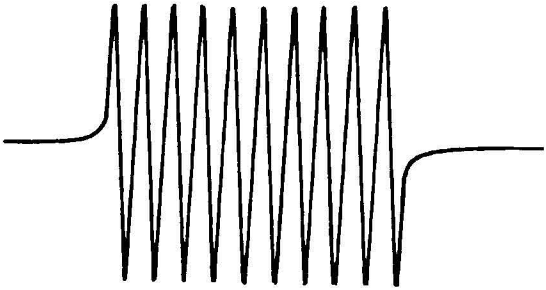
waveforms in terms of their nulls, rather than their peaks, because a null is sharply defined in a scope pattern. On the other hand, the determination of maximum peak amplitude is usually less definite. As the receiver's chroma-phasing control is varied, the peak and null points shift in the waveforms of Fig. 8-16. At any control setting, we will find that the null indications are always quite definite.

The long pulse that extends downward in Fig. 8-16 is the retrace blanking pulse. In some receivers, the blanking pulse is comparatively wide, and tends to merge more or less with the first or the last chroma pulse. Sometimes, this can make it a bit difficult to "count bars." That is, we might not be certain where to start counting bars in the pattern. In such case, we can temporarily disable the blanking pulse at the input of the chroma section. Then, all 10 chroma pulses will be clearly displayed on the scope screen. Subcarrier phases in the chroma-demodulator circuits are determined by R, C, and L components, as seen at the right-hand edge in Fig. 8-2. If incorrect nulls are observed, a defective component is indicated. Capacitors are the most common troublemakers.

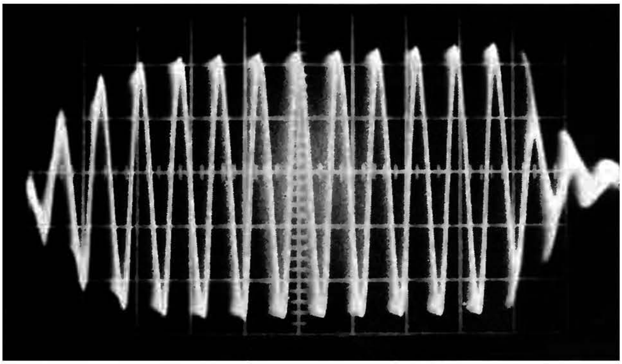
CHROMA SYNC WAVEFORMS

Chroma sync action starts with the color burst. With reference to Fig. 8-2, the complete chroma signal is applied to the gated color-sync amplifier. This stage is also called a burst amplifier. Note that Q7 is gated by a pulse developed from the flyback pulse and the horizontal-sync pulse. In turn, this gating pulse is timed to coincide with the color burst (see Fig. 8-15). Therefore, Q7 in Fig. 8-2 conducts during the burst interval, and passes the color burst to the 3.58-MHz quartz crystal. The crystal is shock-excited by the burst, and rings continuously in phase with the burst signal. Since the crystal has a very high Q value, its ringing waveform decays very little from one burst to the next. Thereby, the color subcarrier is reconstituted.

However, since the output waveform from the crystal is not perfectly uniform, the crystal output is used to synchronize the locked oscillator Q8. The output from this color oscillator has a completely uniform amplitude. Next, to obtain a wide range for the chroma-phasing (hue) control, the output from the color oscillator is fed to phase splitter Q11. In turn, Q12 amplifies the color-subcarrier signal and feeds it to the R, G, and B demodulators. The color burst from a generator appears as seen in Fig. 8-14. If a triggered-sweep scope is used, the



(A) Ideal waveform.



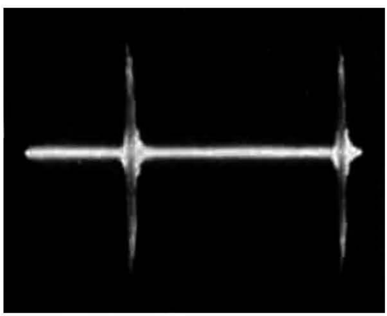
(B) Practical waveform.

Fig. 8-17. Color burst waveforms.

burst can be expanded as illustrated in Fig. 8-17. This is not necessary in ordinary service procedures, and is done chiefly when adjusting a color-bar generator for optimum output waveforms.

The output waveform from the burst amplifier normally appears as illustrated in Fig. 8-18. We are concerned chiefly with

Fig. 8-18. Normal burst-amplifier output waveform.



the peak-to-peak voltage of the burst waveform when a normal signal input is applied to the color receiver. This value is specified in the receiver service data. Subnormal amplitude can be caused by a defect in the burst-gating circuit, such as a leaky capacitor; misalignment of tuned circuits in the signal channel can also be responsible. Misalignment is generally associated with poor picture quality, whereas defects in the burst-gating circuit impair color sync only.

FREQUENCY-RESPONSE CURVES

Details of chroma alignment procedure cannot be covered in this book; however, interested readers may refer to alignment handbooks for procedural details. Receiver service data also provide the essential instructions. Frequency-response curves for chroma bandpass amplifiers vary considerably from one receiver to another. Therefore, alignment should not be attempted without guidance from the receiver service data. A typical bandpass response curve is illustrated in Fig. 8-19. We are concerned with the frequencies at the ends of the top, the tilt of the top (if specified as uphill, downhill, or flat), and the shape of the skirts. When a tilt is specified, it is employed to compensate for rising or falling frequency response in a previous section of the signal channel.

The purpose of the bandpass specified for the curve is to accept most of the chroma sidebands, and to reject the low-frequency components of the Y signal. Of course, the high-frequency components of the Y signal are passed with the chroma sidebands. However, these Y signal components cancel out because the chroma signal is frequency-interleaved with the Y signal. Details of this signal action are explained in books on the theory of color television. The Y amplifier has a frequency response that is approximately the inverse of the bandpass-amplifier response. Fig. 8-20 shows a typical frequency-response curve for a Y amplifier. Note that the lower-frequency components of the Y signal are passed at high amplitude, while frequencies in the vicinity of the color subcarrier are attenuated. A subcarrier trap is provided at 3.58 MHz.

Y-amplifier response curves vary considerably from one receiver to another. In some cases, the subcarrier trap merely places a fairly deep notch in the curve, and does not reduce the response to zero at 3.58 MHz. Greater response is provided between the subcarrier trap and the sound trap than is seen in Fig. 8-20, in some cases. The amount of subcarrier trapping and high-frequency attenuation that is provided depends considerably on the amount of linearity developed by particular color picture tubes. If a tube is comparatively linear, more response at high video frequencies is permissible. In any case, beat interference between the chroma sidebands and the Y signal is held below a visible level. Beat interference has the form of color-picture "crawl." It also has the aspect of 920-kHz bars; that is, the difference between 3.58 MHz and 4.5 MHz (chroma and sound signals) is 920 kHz. Therefore, the sound signal must be extensively trapped.

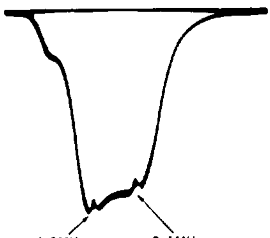


Fig. 8-19. Frequency-response curve for bandpass amplifier.

Fig. 8-20. Frequency-response curve for Y amplifier.

The waviness seen along the top of the Y-amplifier response curve in Fig. 8-20 is caused by residual resonances in the delay line. On a square-wave test, residual resonances tend to cause ringing. Delay lines vary considerably, and high-quality delay lines have very small residual resonances. In such cases, the top of the Y-amplifier response curve is practically as smooth as the top of the video-amplifier response curve. Since the majority of service shops do not have video-frequency sweep generators available, the Y-amplifier response curve is not usually checked. However, most sweep generators can sweep a bandpass amplifier over the 2.5 to 4.5 MHz range, and alignment of bandpass amplifiers is commonly checked.

OTHER COLOR RECEIVER WAVEFORMS

Transistor color-TV receivers also have dynamic convergence waveforms that are absent in monochrome receivers. These convergence waveforms are merely noted here and are specified for particular receivers in the pertinent service data. These waveforms are analyzed in case the dynamic convergence controls lack sufficient range, or are otherwise ineffective in obtaining satisfactory convergence of the color picture tube. Convergence procedure is somewhat involved, and requires both study and experience. Interested readers may consult specialized color-TV servicing books for detailed explanations.

REVIEW QUESTIONS

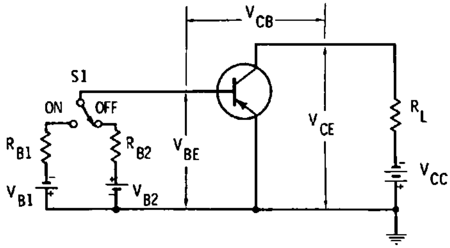
1. Name the principal sections in a transistor color-TV receiver that are not present in a monochrome receiver.
2. What is the function of the chroma bandpass amplifier?
3. Describe the function of a chroma demodulator.
4. How does a color-killer operate?
5. Explain the purpose of an automatic chroma-control circuit.
6. Discuss the composition of an NTSC color-bar signal.
7. Why is a color burst required in the complete color signal?
8. State the electrical characteristics that correspond to brightness, hue, and saturation.
9. Define the Y signal.
10. Give an example of quadrature chroma signals.
11. Name the three color-difference signals that are applied to the color picture tube.
12. What is the function of a chroma matrix?
13. Describe the sampling process in chroma demodulation.
14. Define a null in a chroma waveform.
15. Explain the characteristics of a keyed-rainbow signal.
16. What is the distinction between an NTSC signal and a keyed-rainbow signal?
17. Why is a burst amplifier gated?
18. How does misalignment affect the output waveform from the burst amplifier?
19. State the approximate passband of a chroma bandpass amplifier.
20. Describe the function of a chroma bandpass amplifier.
21. What are the chief characteristics of a Y-amplifier response curve?
22. How is the color subcarrier reconstituted in a color receiver?
23. Explain how a quartz crystal is activated by a succession of color bursts.
24. Discuss the phasing of the injected subcarrier in the chroma demodulators.
25. What type of scope is required to expand a color burst?

9

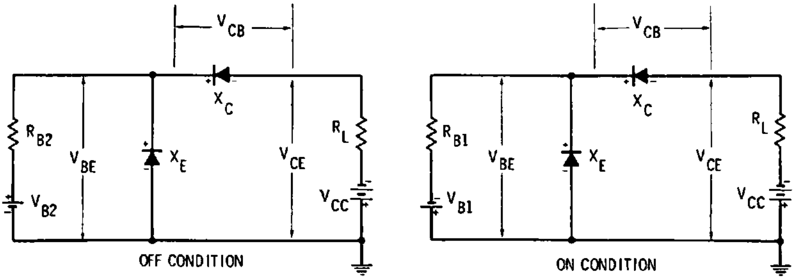
Transistor Electronic Computers and Waveforms

In the first analysis, the basic principles of scope application and waveform evaluation developed in preceding chapters apply to transistor electronic computers. However, we will find that more sophisticated concepts are involved, because computers are designed to operate at maximum practical speeds. We are concerned with the fastest possible rise and fall times, the maximum practical repetition rates, and the minimum practical pulse widths. We work with circuits that have the greatest practical bandwidth, and semiconductor devices that have the fastest response time attainable. We are concerned with problems of deterioration in pulse waveshape, and processes for pulse regeneration. Moreover, reliability of circuit action must approach 100 percent, even if speed of operation is sacrificed. Let us consider some of the problems of pulse-circuit action encountered in transistor computers.

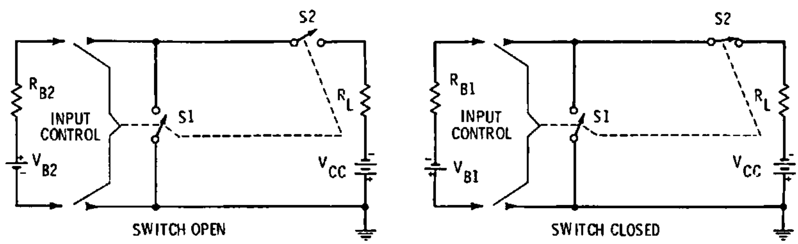
Fig. 9-1 illustrates a simple transistor switching circuit. When S1 is thrown from *off* to *on*, and then back to *off*, an input current pulse waveform I_B is generated, as shown in Fig. 9-2A. In this large-signal operation, the input waveform I_B drives the transistor from cutoff to saturation, and back to cutoff. The output current pulse I_C in Fig. 9-2B is formed, differing from I_B because a transistor cannot respond instantly to a change in signal level. This waveform indicates the transient response of the switching circuit. It is evident that the transient response of a transistor basically determines the maximum repetition rate (switching speed) at which the circuit may be operated. That is, the transistor is generally the bottleneck in limitations of switching speed. The output pulse



(A) Transistor switching circuit.



(B) Diode equivalent of circuit at (A).



(C) Switch equivalent of circuit at (A).

Fig. 9-1. Analysis of transistor switching circuit.

characteristics are governed basically by the ac characteristics of the transistor.

RISE-TIME CONSIDERATIONS

We know that rise time is measured between the 10-percent and 90-percent points on the leading edge of a waveform. Factors affecting rise time are the nonlinear characteristics of the transistor, the energy storage in the semiconductor substance, and, in some cases, the characteristics of the external circuit.

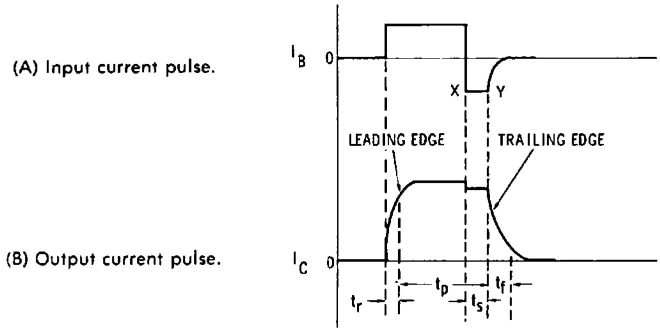


Fig. 9-2. Current pulse characteristics.

Charge carriers (holes or electrons) moving from emitter to collector in the transistor do not have completely free flow paths. The charge carriers collide occasionally with semiconductor atoms, and become dispersed and diffused. Therefore, all charge carriers that start out do not reach the collector output terminal at exactly the same time. We will find that overdriving a transistor results in decreased rise time. Therefore, overdrive is minimized in practical switching circuits. Rise time is denoted by t_r in Fig. 9-2B.

Following the rise interval, the pulse waveform attains a maximum value for the so-called pulse time, t_p, as depicted in Fig. 9-2B. The pulse time is also called the duration time of the waveform, and denotes the length of time that the pulse remains at or near its maximum value. Pulse-time duration is measured from the point on the leading edge where the pulse has attained 90 percent of its maximum value, to the point on the trailing edge at which the amplitude has fallen to 90 percent of its maximum value.

STORAGE TIME

The storage time is denoted by t_s in Fig. 9-2B. This is also called the minority carrier storage time. When the input current I_B is cut off, the output current I_C does not fall to zero instantaneously, but remains at almost maximum value for a certain length of time before decaying to zero. This storage time is also referred to as saturation delay time. It results from injected minority charge carriers that are present in the base region of the transistor at the moment when the input current is cut off. We recall that electrons are minority carriers in p material, and that holes are minority carriers in n material. These charge carriers have a certain mobility, and require a certain length of time to be collected. Transistors with specialized internal construction are used to minimize switching time.

Holes have a mobility of 1700 cm/sec per volt/cm; electrons have a mobility of 3000 cm/sec per volt/cm. Therefore, it is advantageous to use npn transistors in switching applications, because electrons are then minority carriers in the base region, and electrons have a comparatively high mobility. The length of the storage time depends on the degree of saturation into which the transistor is driven, and the time that the charge carriers spend in saturation. The base-current I_B reversal that occurs between points X and Y in Fig. 9-2A at the end of the input pulse, is the result of the stored carriers contributed by the current gain of the transistor, multiplied by the initial input current I_B.

When this current value decays to a value equal to the maximum current value at saturation, the collector-base junction becomes reverse-biased, and both I_B and I_C thereafter decay exponentially to zero. We recognize that for high-speed switching, storage time is undesirable. Minority carrier storage can be avoided by switching a transistor from its *off* state to a point in the active region, thereby avoiding collector saturation. We will find that a technique called collector clamping can be used to prevent operation of a transistor in its saturation region.

FALL TIME

In the fall time, or decay time, of a pulse we know that the amplitude decreases from 90 percent to 10 percent of its maximum value. The fall time of a pulse is determined by essentially the same factors that determine its rise time. We will find that fall time can be reduced slightly by application of a reverse current at the end of the input pulse. Fig. 9-3 shows a superimposed display of rise and fall times by a triggered-sweep scope. The turnoff time of a switching transistor is the sum of its storage time and fall time. Decreased turnoff time results from decreasing either the storage time or the fall time. If the turnoff time is decreased, the maximum permissible switch rate (repetition rate) is thereby increased.

In transistor switching applications, the output current should be zero when the transistor is cut off. However, zero output current cannot be realized in practice. In the CB configuration of Fig. 9-4A, with the emitter current equal to zero

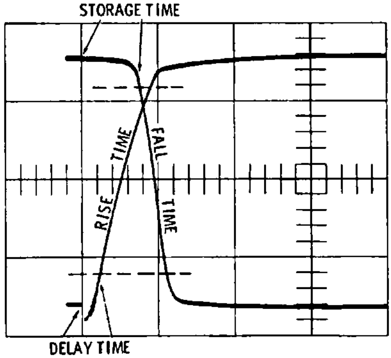
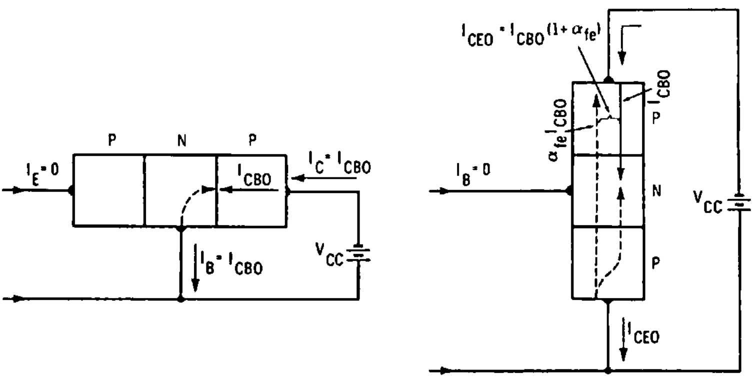


Fig. 9-3. Superimposed display of rise and fall times by triggered-sweep scope.

($I_E = 0$), a small reverse-bias collector current flows. This current is produced by minority carriers in the collector and base regions. Reverse-bias collector current I_{CBO} results from thermal generation of electron-hole pairs, and this current increases with temperature. In the CB configuration, the value of I_{CBO}, also called leakage current, is normally quite small and is measured in microamperes. Consequently, it can be tolerated in most applications.

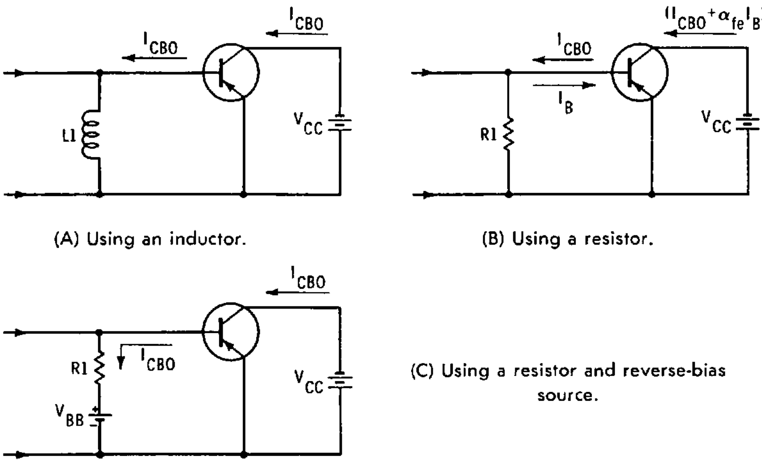
In the CE configuration of Fig. 9-4B, leakage current I_{CEO} from emitter to collector is much larger than in the CB configuration. I_{CEO} is measured from collector to emitter with the base terminal open ($I_B = 0$). Note that I_{CEO} stems from I_{CBO}. That is, reverse-bias minority carriers (electrons, solid-line arrow) in the collector region enter the base region and combine with holes (dashed-line arrow) from the emitter region. Before combining, however, the electrons cause a heavy hole current to flow from emitter to collector. In other words, the



(A) I_{CBO}. (B) I_{CEO}.

Fig. 9-4. Leakage current.

electrons from the collector region act as a base-bias current to cause an amplified collector-current flow. The magnitude of this current is approximately the product of beta and I_{CBO}. Since beta ranges from 25 to 50 for typical switching transistors, leakage current is of major concern.



(A) Using an inductor.

(B) Using a resistor.

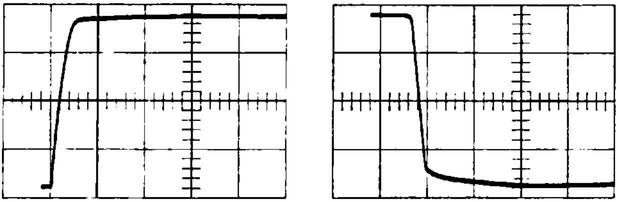
(C) Using a resistor and reverse-bias source.

Fig. 9-5. Circuits for reduction of leakage current I_{CEO}.

REDUCTION OF I_{CEO}

The value of I_{CEO} can be reduced by means of several circuit arrangements. However, we cannot reduce I_{CEO} below the value of I_{CBO}. As shown in Fig. 9-5A, an inductor, L1, may be used instead of a base-return resistor. In turn, there is negligible dc voltage drop between base and emitter, and when $I_B = 0$, $I_{CEO} = I_{CBO}$. It is not always possible to use an inductor in a switching circuit; accordingly, a low-valued resistor, R1, may be used instead, as depicted in Fig. 9-5B. Flow of I_{CBO} through R1 produces a small forward bias, resulting in a small flow of base-emitter current I_B. In turn, I_{CEO} is approximately equal to beta times I_B. To minimize I_B, the value of R1 is made as small as is practical. When a high base-input resistance is required, I_{CEO} can be minimized by means of a reverse-bias source V_{BB}, as shown in Fig. 9-5C. This reduces I_B to zero, so that only the leakage current I_{CBO} flows.

In a properly designed switching circuit, and with a transistor in normal operating condition, the rise and fall times are comparatively rapid. A collector load resistance of 500 ohms is typical (not shown in Fig. 9-5). The drive waveform may have an amplitude of 0.5 volt pk-pk, with a source resistance of 50 ohms in a typical switching circuit. A collector supply voltage of 3.5 volts may be utilized. The output pulse waveform will have a typical amplitude of 7 mA pk-pk and a peak-to-peak voltage of 3.5 volts. This drive condition switches the transistor from cutoff to collector saturation. Corresponding rise and fall times are illustrated in Fig. 9-6; the time base of the scope is set for 0.5μs per horizontal division.



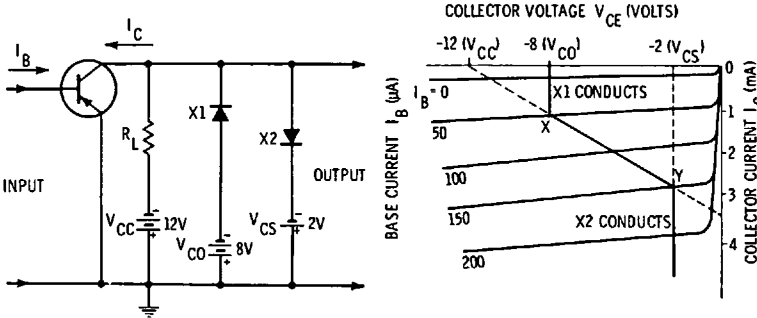
(A) Rise time. (B) Fall time.

Fig. 9-6. Rise time and fall time in switching transistor.

CUTOFF AND SATURATION CLAMPING

When a transistor is driven to saturation and cutoff, the output waveform may be distorted. Variations in collector potential occur when either the temperature-dependent cutoff current drifts, or the load impedance varies. The resulting change in collector potential can cause unreliable operation of following stages. We know that when a transistor is driven into saturation, minority-carrier storage delay occurs. This widens the pulse waveform and reduces the permissible repetition rate of the switching circuit. Junction diodes are commonly employed to avoid transistor operation in saturation, or at cutoff. Other methods are also used to avoid objectionable saturation.

In Fig. 9-7, a pnp transistor is used in the CE configuration for a simple switching circuit. Drive current I_B is sufficiently great to drive the transistor from cutoff to saturation without the clamping diodes X1 and X2, and bias batteries V_{CO} and V_{CS}. Fig. 9-7B shows the transistor output characteristics with load line R_L. V_{CC} is 12 volts in this example. X1 and bias battery V_{CO} are used for cutoff clamping. X2 and V_{CS} are used for saturation clamping. Clamping of the upper and lower levels in the output waveform facilitates substitution of one transistor for another. Since operation with clamping diodes usually provides operation over the linear portion of the output characteristics, low cutoff current and low saturation voltage are eliminated. The average power dissipation is increased, and the load line must be selected so that transistor operation falls within the permissible power-dissipation region, under the maximum power-dissipation curve noted previously.

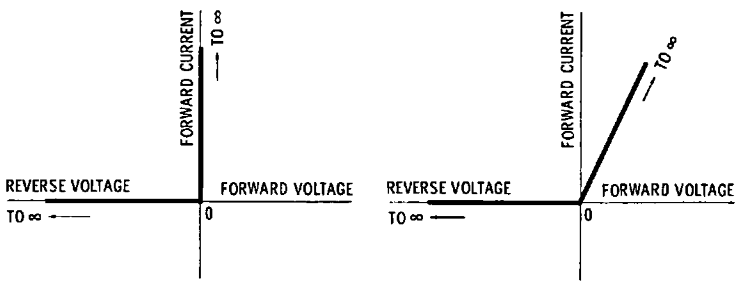


(A) Schematic.

(B) Characteristic curves.

Fig. 9-7. Cutoff and saturation clamping.

Let us consider cutoff clamping action. In Fig. 9-7B, we may assume that I_B has started to fall from 150μA at point Y. The collector voltage increases negatively from 2 volts at this point. Battery V_{CO} applies 8 volts to X1. As the collector potential increases from 2 volts toward 8 volts, X1 remains reverse-biased and nonconducting. When the collector potential reaches point X ($I_B = 50\mu$A) on the load line (8 volts), X1 becomes



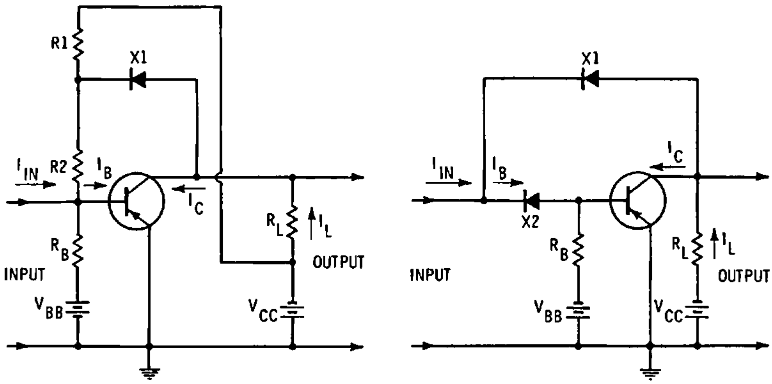
(A) Ideal characteristic.

(B) Approximate practical characteristic.

Fig. 9-8. Clamp characteristics.

forward-biased and starts to conduct. Fig. 9-8 depicts ideal and approximate practical clamp characteristics. Further decrease in I_B toward zero has no effect on the collector potential, which remains fixed at 8 volts, even though I_C decreases (V_{CO} holds V_{CE} at 8 volts).

Current flow through R_L in Fig. 9-7 is maintained at X (about 1 mA); it consists of I_C and the current through forward-biased X1 and V_{CO}. The voltage drop across R_L is equal to the difference (approximately 4 volts) between V_{CC} and V_{CO}. Any change in I_C that does not cause it to exceed the value at point X is compensated by current from V_{CC}. Thus, equilibrium is established and the output potential is fixed at 8 volts. Note that X2 is reverse-biased and is effectively an open circuit in the foregoing clamping process. In high-speed switching circuits, the transient response of diodes becomes a matter for consideration.



(A) Single diode. (B) Double diode.

Fig. 9-9. Collector-to-base saturation clamping.

Next, let us follow the circuit action in saturation clamping. Disregarding X1 and V_{CO} in Fig. 9-7, we may assume that I_B is increased to 150μA at point Y. The collector voltage, indicated by the vertical projection to the collector-voltage axis, falls to 2 volts. The voltage drop, produced by increasing I_C through R_L, causes the collector voltage to fall. Bias voltage from V_{CS} is 2 volts. A slight further increase in I_C, due to increase in I_B, further decreases the negative potential on the collector, and forward-biases X2, which starts to conduct. When X2 conducts, its resistance is very small, and V_{CS} is effectively applied between collector and ground. This voltage cannot change and since it is in parallel with V_{CC} and R_L, the voltage across this branch cannot change. Current through R_L is fixed to provide the necessary voltage drop so that the collector-to-ground potential is equal to V_{CS}. In this example, the load current is approximately 3 mA.

Although the collector current may increase further, because of an increase in base current, the collector potential remains fixed at 2 volts. Additional collector current is drawn from V_{CS} through X2. The vertical extension of the load line from Y represents the zero-resistance load line for forward-biased X2. The normal load line, on the other hand, extending from Y to the collector-current axis (dashed line) would enter the saturation region, which falls beyond the knee of the curve of $I_B = 200\mu A$. The collector-base diode would be forward-biased in this region. When the drive current is reduced to zero, I_C falls rapidly from its maximum value to point Y, and then follows the normal load line toward point X. However, any minority-carrier storage introduced by the clamping diode (Fig. 9-7) would affect the width of the output pulse. To the left of point Y on the load line, the collector potential is more negative than 2 volts. Diode X2 is reverse-biased, and is therefore in a nonconducting state.

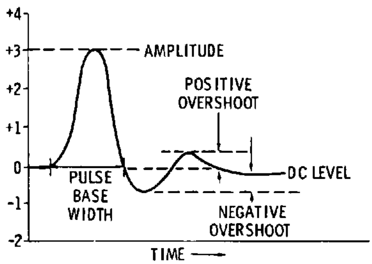


Fig. 9-10. Deteriorated pulse-drive waveform.

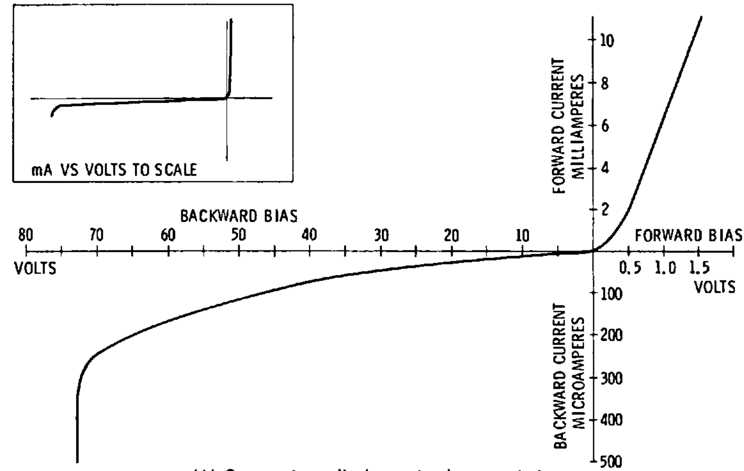
Thus, the foregoing switch circuit provides combination clamping. The effect of cutoff and saturation clamping is to keep the collector potential in the range of −2 to −8 volts, thereby preventing waveform distortion due to cutoff and saturation conditions in the transistor. More efficient switching action can be obtained by means of collector-base saturation clamping with a single diode or a duo-diode. Single-diode clamping is depicted in Fig. 9-9A. In the cutoff condition of the transistor, X1 and the emitter-base junction are reverse-biased by R1, R2, and R_B in series with V_{CC} and V_{BB}. When V_{CC} and V_{BB} are equal, R1 + R2 is made slightly greater than R_B. This provides the required initial reverse bias for the emitter-base junction. R1 is much larger than R2, the ratio being determined by the desired clamping voltage when the transistor is conducting.

It is assumed that the drive pulse to the switching circuit has a more nearly ideal waveshape than can be reproduced in the collector circuit. Pulses become progressively deteriorated, as shown in Fig. 9-10. Therefore, computer systems employ waveform regenerators when necessary, to reconstitute approximately ideal drive waveforms. Details of pulse regeneration are explained subsequently. Let us assume that I_{IN} in Fig. 9-11A forward-biases the emitter-base junction, and drives the transistor into saturation. The potential at the junction of R1 and R2 is negative, and is effectively fixed near the very low saturation voltage of the transistor. The collector voltage falls from V_{CC} to this fixed potential. X1 is nonconducting and $I_B = I_{IN}$. $I_L = I_C$, and is the amplified drive potential I_{IN}. When the collector potential falls just below the fixed potential, X1 conducts.

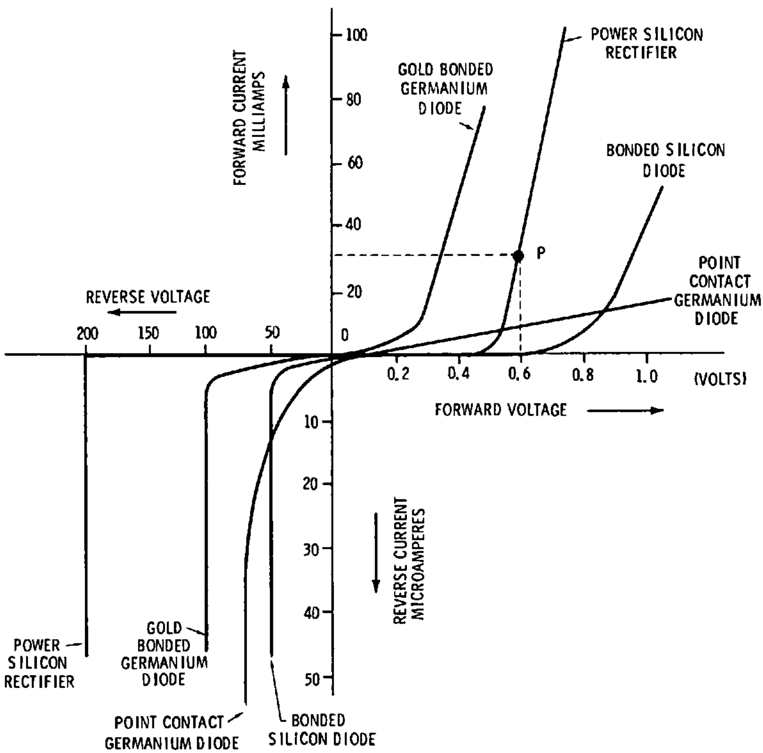
As X1 conducts and I_{IN} increases further, I_B remains essentially constant, and the excess current is shunted through R2 and X1. I_C increases by the small value of excess current input, rather than by the amplifier input current prior to clamping action. Since relatively small currents are passed through X1, the minority-carrier storage delay contributed by the clamping diode is smaller than that contributed by the clamping diode in Fig. 9-7. Battery power dissipated in R1 and R2 is avoided by using double-diode clamping, as shown in Fig. 9-9B. X1 functions as a clamping diode, in the same manner as previously explained, while X2 substitutes for R2 in Fig. 9-9A.

The emitter-base junction is reverse-biased by V_{BB} in Fig. 9-9B through R_B. Collector reverse bias is provided by V_{CC} through L. X2 remains forward-biased throughout the circuit action. A germanium diode is generally used for X1 because it has a low forward-voltage drop. A silicon diode, with a slightly higher forward-voltage drop, is used for X2. (See Fig. 9-11). In a typical application, X1 in Fig. 9-9B provides a forward-voltage drop of 1 volt, and X2 provides a forward-voltage drop of 1.2 volts. This essentially maintains a reverse bias of 0.2 volt between collector and base after clamping action ensues.

Circuit action in the duodiode arrangement is similar to that of the single-diode clamping circuit. All of the input current I_{IN} is applied to the base, while X1 remains reverse-biased. This condition continues until I_L drops the collector potential just below the potential at the junction of diodes X1 and X2. The base potential is more positive than the potential at this point by the value of voltage drop across X2. X1 becomes forward-biased and shunts excessive input current of the collector. The smaller drop across X1 maintains the reverse collector-base bias at the difference between the drops across X1 and X2. During the clamped condition, I_L is equal approximately to beta times I_B, and I_C is equal to the sum of I_L and the shunted portion of I_{IN}. The latter is small and contributes little to output pulse widening. Therefore, the repetition rate of the switching circuit is increased.



(A) Germanium diode static characteristics.



(B) Comparative diode static characteristics.

Fig. 9-11. Static characteristics of solid-state diodes.

PULSE REGENERATION

A basic pulse-regenerator circuit is shown in Fig. 9-12. This is a one-shot multivibrator configuration; the output pulse shape is independent of the input pulse shape. Also, the width of the output pulse is independent of the width of the input pulse. Therefore, if a deteriorated pulse waveform is used to trigger the pulse regenerator, a standard pulse is obtained at the output. The common-emitter resistor contributes to fast rise and fall of the output waveform. The diode is called a steering diode, and is a one-way device that permits only a negative-going leading edge to trigger the circuit. By choosing a suitable time constant, the output pulse can be made to have as narrow a width as desired, consistent with the transient response of the transistor.

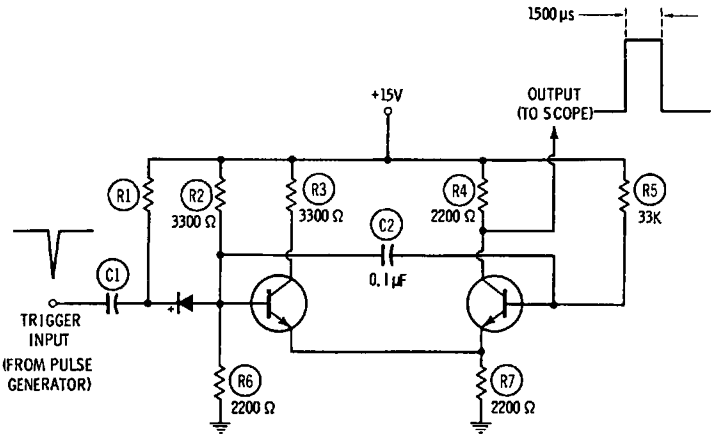
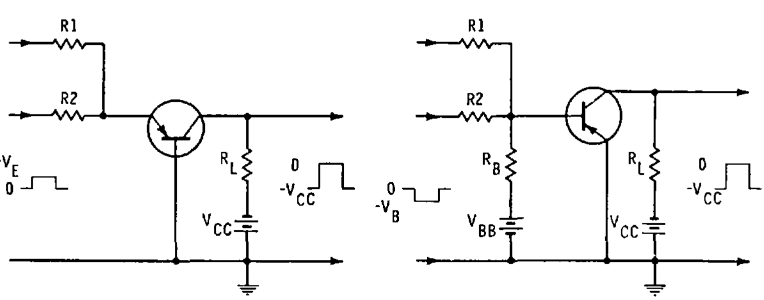


Fig. 9-12. Basic pulse-regenerator circuit.

MACHINE-LOGIC CIRCUITS

Machine logic is an extensive topic, and can only be briefly discussed here. Logic circuits are used in combination with counter circuits to perform operations in arithmetic. Basic transistor OR circuits are depicted in Fig. 9-13. An OR gate provides an output when a positive pulse is applied at either of the input resistors. That is, an output pulse is obtained by applying an input to R1 or to R2. Pulses can be applied to both resistors, and an output pulse is again obtained, although the increased input current tends to produce a wider output pulse due to storage effects.

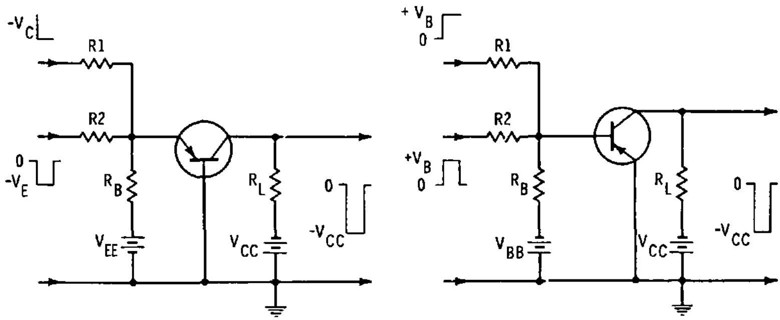


(A) OR gate. (B) NOR gate.

Fig. 9-13. Single transistor OR gates with multiple inputs.

The configuration in Fig. 9-13B is similar to that in A, except that it is driven by a negative input pulse, and is called a NOR gate. Note that positive input pulses produce no output. No output is obtained, whether a positive pulse is applied to R1, R2, or to both R1 and R2. However, an output pulse is obtained if a negative pulse is applied to either R1, R2, or to both R1 and R2. Fig. 9-14 depicts AND gates. The AND gate is similar to the OR gate, but the transistor is biased to collector saturation. An input pulse at R1 is insufficient to bring the transistor out of saturation. However, if an input pulse is also applied to R2, the sum of input 1 and input 2 then drives the transistor from saturation and into cutoff. Thus, inputs 1 and 2 together produce an output pulse. The NOT AND gate provides a negative output pulse when positive pulses are applied simultaneously to the inputs. However, no output is obtained with negative applied pulses.

A flip-flop circuit is shown in Fig. 9-15. This is the basic element in a counter section, and is used for information storage in electronic computers. It is basically a bistable multivibrator, and requires two input trigger pulses in succession to



(A) AND gate. (B) NOT AND gate.

Fig. 9-14. Single transistor AND gates with multiple inputs.

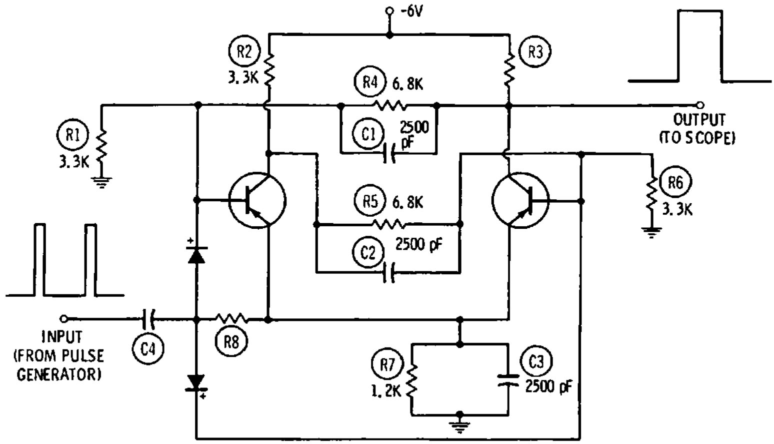


Fig. 9-15. Flip-flop configuration with trigger circuit.

produce an output pulse. Therefore, the repetition rate of the output pulse is one-half the repetition rate of the input pulses. It is evident that if we connect two flip-flops in series, one output pulse will be obtained for four input pulses. Three flip-flops connected in series provide one output pulse for six input pulses, and so on. If neon bulbs are connected in the collector circuits of a flip-flop chain, we obtain readout of stored binary numbers. Interested readers may refer to computer arithmetic books for discussion of binary numbers and their relation to ordinary decimal numbers.

The diodes in Fig. 9-15 are steering diodes that direct input pulses automatically to the transistor that should be triggered for proper flip-flop action. The transistors conduct alternately from one input pulse to the next, and the diodes are alternately reverse-biased by the circuit voltages, so that each pulse flows into the correct transistor. C1 and C2 are speedup capacitors, used to obtain maximum speed of response. They overcome the integrating effects of R4 and R5. C3 is a conventional emitter bypass capacitor. R7 serves as a temperature-stabilizing resistor. Direct coupling is used in computer flip-flops to minimize the number of components that are required.

REVIEW QUESTIONS

1. Explain the distinction between a switching circuit and an amplifier circuit.
2. What is the cause of input pulse distortion in a switching circuit?

3. What are the causes of output pulse distortion in a switching circuit?
4. Why is overdrive minimized in pulse switching circuits?
5. Discuss the meaning of storage time in a switching transistor.
6. Do electrons or holes have greater mobility?
7. How is leakage current I_{CEO} minimized in a switching circuit?
8. State typical voltage and current amplitudes in a switching circuit.
9. What rise and fall times might we find in a switching circuit?
10. Define cutoff and saturation clamping.
11. Discuss the effect of storage time in a clamping diode.
12. Why are two diodes often used in a saturation clamping circuit?
13. Describe deterioration of a pulse waveform in a computer system.
14. Briefly explain the differences in forward characteristics of semiconductor diodes.
15. How is a one-shot multivibrator used as a pulse regenerator?
16. Explain the configuration of an OR gate.
17. How is a NOR gate distinguished from an OR gate?
18. Describe the configuration of an AND gate.
19. How does a NOT AND gate differ from an AND gate?
20. What is the function of machine logic?
21. Discuss the operation of a flip-flop counter section.
22. Why are flip-flops connected in cascade?
23. How is information stored in a counter chain?
24. Explain how readout can be obtained from a counter chain.
25. Is readout obtained in the binary system or in the decimal system?

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