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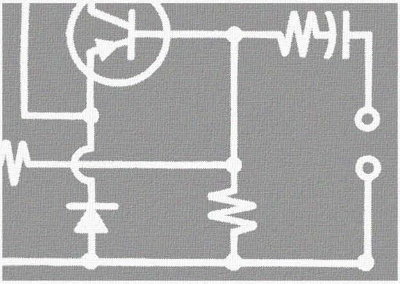
RCA TRANSISTOR SERVICING GUIDE



RCA ELECTRONIC COMPONENTS AND DEVICES, HARRISON, N. J.

RCA
TRANSISTOR SERVICING GUIDE

The introduction of transistors in home entertainment and industrial equipment has provided both an opportunity and a challenge to the service technician—the opportunity to expand his knowledge, and the challenge of a new technology. In practical terms, an understanding of transistor circuits results in faster, more accurate trouble-shooting, and therefore in greater customer satisfaction.

This guide, prepared by RCA, contains information on servicing transistor circuits used in radio and television equipment. Also included are sections on transistor theory, covering amplifier principles and considerations. The RCA Transistor Servicing Guide will be very useful to the service technician, enabling him to profit from this newer area of electronics.

Prepared by
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RADIO CORPORATION OF AMERICA
ELECTRONIC COMPONENTS AND DEVICES HARRISON, N. J.

CONTENTS

CHAPTER 1

TRANSISTOR AMPLIFIER PRINCIPLES

INTRODUCTION

All transistors can be considered as two semiconductor junctions "glued" back-to-back. These junctions are each made from two slightly different forms of a semiconductor material such as germanium or silicon. The different forms of a particular semiconductor material are produced by adding specific impurities to the molten semiconductor material in a process called *doping*. The two forms of semiconductor material which result, called *P* material and *N* material, are combined to form a *PN* junction.

In this chapter you will learn how two PN junctions are combined to form a transistor. One junction serves as a source, or injector of current carriers, and the other junction collects these injected current carriers. After studying the basic transistor principle, you will learn about the various circuit arrangements in which the transistor is used.

1. THE BASIC TRANSISTOR

PN Junction. Let's briefly discuss some important features of PN junction operation. Consider the forward-biased junction shown in Fig. 1. Due to the electric field set up by the forward-biasing battery:

holes move into the N region, and electrons move into the P region.

Now, the holes which are made to move into, or are *injected* into the N region, are minority carriers in that region. Likewise, the electrons that are made to move into, or are *injected* into the P region, are also minority carriers. Thus, a forward-biased junction is a *minority-carrier injector.* Also, the amount of minority carriers injected depends upon the amount of forward bias voltage.

Now, in a symmetrical PN junction, the injected minority carriers

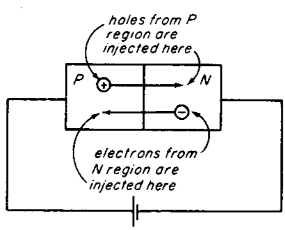


Fig. 1. A forward-biased junction is a minority-carrier injector.

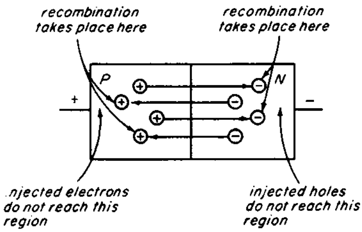


Fig. 2. In a symmetrical junction, recombination limits travel of injected carriers.

are not going to last very long, or get very far. There are many free electrons in an N region. Holes that are injected into an N region stand a very good chance of meeting an electron and becoming neutralized. This is shown in Fig. 2. Similarly, there are many holes in a P region. Electrons injected into a P region stand a good chance of meeting a hole and being neutralized. Thus, in a symmetrical PN junction an injected minority carrier cannot penetrate too far, or last too long, before it is neutralized.

The Unsymmetrical PN Junction. However, suppose that we make an unsymmetrical PN junction, as in Fig. 3. This junction is unsymmetrical in two ways. First, the N region is made very thin. Second, the P region is doped much more than the thin N region. When this junction is forward-biased, holes are injected into the N region, as before. But now, the injected holes have a very good chance to survive and reach the end of the N region. Because the N region is so thin, the injected holes can cross over the N region with little chance of meeting an electron. Also, because of the heavy P doping, there are many more injected holes than electrons. Thus, even if some holes do meet electrons, there are many holes left at the end of the N region, as is shown in Fig. 3. In a similar way, suppose we make a junction with a very thin P region, and a very heavily doped N region, as in Fig. 4. When this junction is forward-biased, electrons are injected into the P region as in the symmetrical junction. But now, there are many more injected electrons, and they have a good chance of reaching the end of the P region before meeting up with a hole.

The properties of forward-biased junctions in which one-half of the

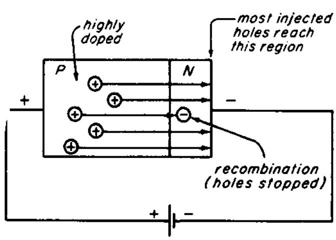


Fig. 3. Hole injection is improved by unequal doping and a thin N section.

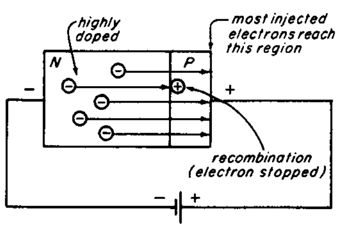


Fig. 4. Electron injection is improved by unequal doping and a thin P section.

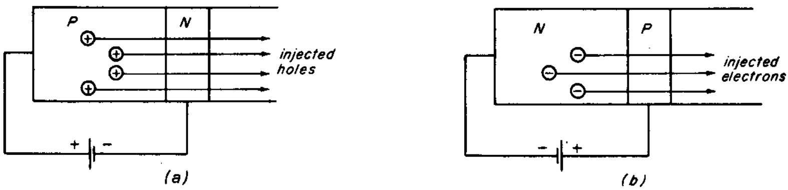


Fig. 5. (a) A forward-biased PN junction is a hole injector; (b) a forward-biased NP junction is an electron injector.

junction is very thin and the other half is heavily doped are summarized in Fig. 5:

A forward-biased PN junction is a hole injector, as in Fig. 5a. The number of injected holes depends on the amount of forward bias.

A forward-biased NP junction is an electron injector, as in Fig. 5b. The number of injected electrons depends on the amount of forward bias.

The Transistor Principle. Of course, the injected carriers do not spurt out of the N or P region into the air, as might be suggested by Fig. 5. A transistor is made by combining two junctions, as shown in Fig. 6. This combination may be either in the form of a PNP "sandwich," or an NPN "sandwich." In this two-junction transistor device, one junction is forward biased, and serves as an injector or source of current carriers, as shown in Fig. 7. This junction becomes the input junction when the transistor is used as an amplifier. The other junction is reverse biased, and becomes the output junction in an amplifier. When the injected carriers reach the output junction, the reverse-biased electric field across the output junction sweeps them into the output region and they are collected.

Let us see how this works specifically with the PNP transistor in

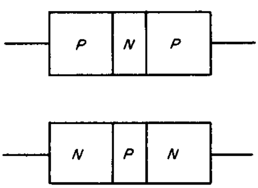


Fig. 6. The transistor is basically a PNP "sandwich" or an NPN "sandwich."

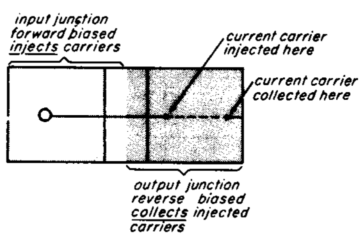


Fig. 7. A transistor consists of a forward-biased junction that injects carriers and a reverse-biased junction that collects the injected carriers.

Fig. 8. The forward-biased PN input junction injects holes from the heavily doped P region into the thin N region. Battery V_2 reverse biases the second, or NP junction. Note the tiny plus and minus signs in Fig. 8 along the lines *a* — *b*. These show the electric field set up across the NP junction by V_2. When a hole reaches the region between *a* — *b*, the direction of this electric field is such that the positively charged hole is forced across, into the right-hand P region. Once in this P region there is almost no chance of it meeting an electron, so the hole continues across to the negative battery terminal where it is "collected." That is, the arrival of the hole at the negative battery terminal allows an electron to enter the crystal and neutralize the hole. Thus, current flow into the input circuit produces a corresponding current flow in the output circuit; the amount of current flow in the output circuit depends on the amount of current flow in the input circuit.

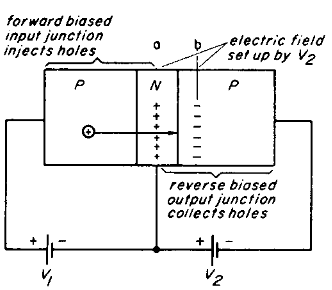


Fig. 8. In the PNP transistor, the forward-biased input junction injects holes; the reverse-biased output junction collects these holes.

Transistor Nomenclature and Symbols. Since the input P section of the PNP transistor supplies the current carriers, it is known as the *emitter*. The thin central section is called the *base*. The right hand P section, which collects the injected current carriers, is called the *collector*. These names, which are shown in Fig. 9*a*, apply to either PNP or NPN transistors. Note that the junction between emitter and base is called the emitter-base junction, and is always forward biased. On the other hand, the collector-base junction is always reverse biased.

Figure 9*b* shows the symbols for transistors. A straight line segment represents the base. The two inclined lines represent the emitter and collector. The inclined line with the arrowhead on it identifies the emitter; the other inclined line represents the collector. When the

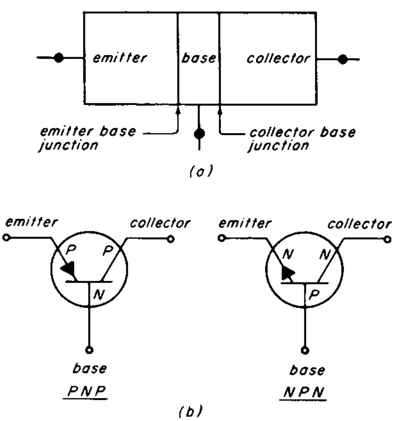


Fig. 9. Transistor nomenclature and symbols.

arrowhead points towards the base, the symbol represents a PNP transistor. If the arrowhead points away from the base, the symbol represents an NPN transistor. The arrowhead on the transistor symbol is in the direction of *conventional current* flow—electron flow is opposite the direction of this arrow.

NPN Transistor Action. Let's go over the basic transistor action again, this time using an NPN transistor as in Fig. 10. The emitter-base junction is the input junction. It is forward biased, and the heavily doped emitter injects electrons into the base region. Battery V_2 reverse biases the collector-base junction, as shown by the tiny polarity marks along the lines $a - b$.

The injected electrons travel through the thin base region with little chance of meeting a hole, and most of them reach the collector-base junction. Note the direction of the electric field across the collector-base junction. It is such as to sweep any negative charge across the junction. Once inside the collector region, the electrons continue across to the positive battery terminal, where they flow into the external battery circuit. Thus, electron flow into the emitter terminal produces a corresponding electron flow out of the collector terminal.

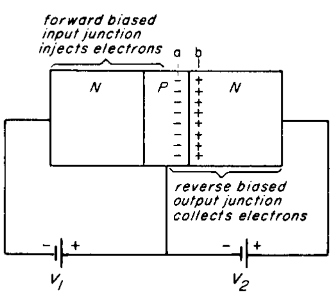


Fig. 10. In the NPN transistor, the forward-biased input junction injects electrons; the reverse-biased output junction collects these electrons.

In the remainder of this chapter we shall discuss the NPN transistor. In order to convert the discussion for a PNP transistor, you need only reverse the battery connections and think of hole flow rather than electron flow.

Emitter Base and Collector Currents. We now examine the currents flowing in the three terminals of the transistor, and refer to Fig. 11, in which a properly biased NPN transistor is shown. The resistor R_1 serves to limit the emitter current. We shall now use the letter designation V_{EE} for the emitter supply battery, and V_{CC} for the collector supply battery.

The application of forward bias to the emitter diode causes I_E, the

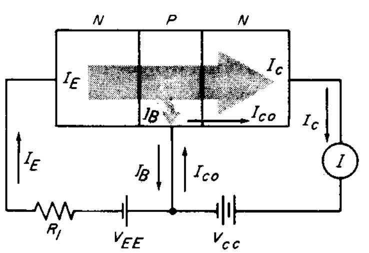


Fig. 11. Transistor currents.

d-c emitter current to flow. This current is made up primarily of majority carriers, electrons. When the electrons diffuse into the base region, the total electron flow splits and flows in two directions. Some 98% of the electrons emitted into the base reach the collector and go on to the positive terminal of the battery. The remainder of the injected electrons meet and combine with holes coming from the base terminal. Due to these electron-hole combinations, part of the total emitter current flows out of the base and back to the emitter battery. This small current which flows through the base terminal is the d-c base current, designated I_B. Hence, the emitter current is equal to the sum of the collector and base currents or:

$$I_E = I_C + I_B$$

The fundamental relationship should be remembered. It tells us that the base current and collector current must add up to the total emitter current.

Alpha. From Fig. 11, you see that the total collector current is the sum of the leakage current, designated I_{co}, and that percentage of the emitter current which reaches the collector, or:

$$I_C = I_{co} + \text{some fraction} \times I_E$$

In the previous section we assumed that this fraction of the emitter current which reached the collector was 98% or 0.98. However, this number is dependent upon the relative doping of the base and emitter, and the thickness of the base. These factors may vary from one transistor to the next, and the fraction of emitter current to reach the collector will likewise vary from transistor to transistor. Because this factor is important to the discussion of the transistor, it is given the special name *alpha* (α). Then, the collector current is expressed as:

$$I_C = I_{co} + \alpha\, I_E$$

The base current can also be expressed in terms of emitter current and alpha. Remember that base current I_B is the difference between the emitter and collector currents. Therefore since

$$I_C = I_{co} + \alpha\, I_E$$

it works out that:

$$I_B = I_E\,(1 - \alpha) - I_{co}$$

Note that the base current equals that portion of the emitter current which does not reach the collector *minus* I_{co}. This is clear from Fig. 11. Notice that I_B flows down the base lead while I_{co} flows up. In summation then, alpha represents that part of the emitter current which reaches the collector, and is always less than one, since some small part of the emitter current will always be diverted to the base. Typical values of alpha range from 0.98 to over 0.99.

2. HOW TRANSISTORS AMPLIFY

The transistor can be made to amplify by connecting a signal source between two of its terminals while a load is connected to two other terminals. Since there are only three terminals on a transistor, one terminal must be common to both input and output circuits. The transistor may be operated with any terminal common to input and output. Consequently, three amplifier circuits are possible. Let's first examine the configuration which is called the *common-base amplifier.*

The Common-Base Amplifier. In this circuit the signal source is connected between the emitter and base, while the load is connected between the base and collector as shown in Fig. 12. Because the base terminal is common to both the input and output circuits, this configuration is known as the *common-base circuit.* The batteries V_{EE} and V_{CC} establish the proper biasing while C_E and C_C bypass the power supplies so that the full a-c output signal is developed across the load R_L. The input signal causes the emitter current to vary and thereby causes the collector current to vary.

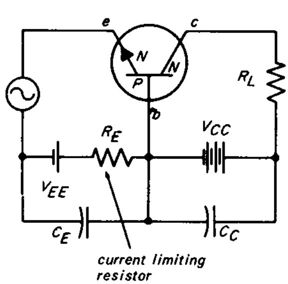


Fig. 12. Common-base circuit.

Current Gain. We now examine current gain of the common-base circuit. The signal source causes the emitter current to vary about the bias value, causing an a-c emitter current (I_e). The varying emitter current causes the collector current to vary about a fixed value, thereby developing an a-c collector current, I_c. We may disregard leakage current entirely, since it is a d-c current and will have no effect on the a-c output.

Current gain A_i is defined as the ratio of the a-c output current to the a-c input current. In this circuit the output current corresponds to the collector current (I_c), while the emitter current (I_e) represents the input current. Therefore,

$$A_i = \frac{I_c}{I_e}$$

Then the *current gain* of the *common-base* circuit is approximately *equal to alpha* and is therefore, *always less than one.* The common-base circuit does not have a current gain but exhibits a slight current loss. However, this circuit is capable of voltage and power gain.

Voltage Gain. Voltage gain (A_v) is defined as the ratio of output voltage to the input voltage. In the case of the common-base circuit, the output voltage is the collector voltage (V_c) and the input voltage is the emitter voltage (V_e).

Therefore, $$A_v = \frac{V_c}{V_e}$$

The *voltage amplification* of the common-base circuit is also equal to the product of alpha and the ratio of load resistance to the input resistance. Input resistance is the resistance that appears between the transistor's emitter and base terminals. This relationship is as follows:

$$A_v = \alpha \frac{R_L}{R_{in}}$$

Power Gain. Power gain G is the ratio of output power to input power. In the common-base circuit the output power is the collector power, while the input power is the emitter power.

Then: $$G = \frac{P_c}{P_e}$$

However, since power is equal to the product of current and voltage, power gain is also equal to the product of current gain and voltage gain:

$$G = \alpha^2 \frac{R_L}{R_{in}}$$

In other words, the power gain is equal to alpha *squared* times the resistance ratio in the common-base circuit.

Output-Signal Phase. Let us now examine the *phase* of the output voltage compared to the input voltage. Recall that in a vacuum-tube amplifier circuit there is a 180° phase shift between input and output voltage. Referring to the circuit in Fig. 13, when the signal voltage is going negative it increases the forward bias of the emitter

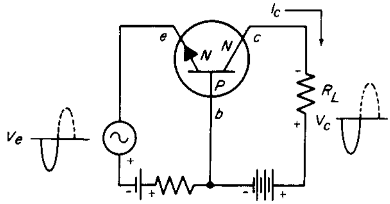


Fig. 13. Signal phase in CB circuit.

junction and causes more carriers to cross the junction and enter the collector. This causes the collector current to increase, which in turn, causes the voltage drop across the load resistor to increase. The increased voltage drop drives the top of R_L more negative and thereby causes the output voltage to go more negative. In the same way, a positive-going input signal causes the output signal to go more positive. In other words, in the common-base configuration, there is *no* phase shift between the input and output voltages.

Frequency Cutoff. The highest frequency signal that may be amplified by a vacuum-tube is limited by the small interval of time taken for electrons to traverse the cathode to anode region. This interval is known as *transit time*. Transit time also places a limit on the highest frequency that can be amplified by the transistor. As a result, the gain falls off more at higher frequencies.

This same decrease in gain can be simulated by an imaginary capacitor connected between base and emitter. At the low frequencies the capacitor has practically no effect on the gain of the amplifier. But as the frequency increases the capacitor begins to short out part of the input signal. Effectively we now have a smaller input signal and the size of the output signal decreases. As the frequency continues to increase, still more of the input signal is shunted through the capacitor and the magnitude of the output decreases further, until the transistor is effectively bypassed by the capacitance and becomes useless as an amplifier. The imaginary capacitance is termed the *diffusion* capacitance, since its effects, called *diffusion delay,* are due to transit time.

The effect of decreasing current gain with frequency is expressed by stating the frequency at which alpha falls to 0.707 of its low-frequency value. The frequency is referred to as the *alpha cutoff frequency* and is designated as $f\alpha$.

For instance, if we are told that a certain transistor has a low-frequency alpha of 0.98 and an alpha-cutoff frequency of 1 mc, its alpha at 1 mc is $0.707 \times 0.98 = 0.693$.

Summary. The important features of the common-base circuit are:

1. A current gain (A_i) equal to alpha, which is always less than one.

2. A voltage gain (A_v) equal to alpha times the ratio of R_L to R_{in}.

3. A power gain (G) equal to alpha squared times the ratio of R_L to R_{in}.

4. No phase shift between the input and output signals.

5. A cutoff frequency ($f\alpha$) at which alpha is equal to 0.707 of its value at low frequencies.

6. A relatively low input impedance, and a relatively high output impedance.

3. THE COMMON-EMITTER CIRCUIT

In the common-emitter circuit, the emitter terminal is common to both input and output circuits as shown in Fig. 14. The input signal is applied between the base and emitter, as in the common-base configuration, but the output signal is obtained between the emitter and collector terminals.

In Fig. 14 the emitter bias battery V_{BB} is installed as shown so that the P material base is made positive with respect to the N-type emitter. The resistor R_1 is a current-limiting resistor, which determines the bias current in the emitter diode. To obtain any desired bias current we may vary the value of R_1.

The collector diode is reverse biased by battery V_{CC}. At first

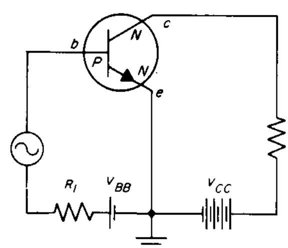


Fig. 14. Common-emitter amplifier.

glance the battery connections seem incorrect, but if you analyze the circuit, you can see that the N-type collector is positive with respect to the P-type base. Without a signal, the emitter-to-base current. (I_b) flows causing R_1 to drop the voltage V_{BB} and thereby make the potential at the base terminal approximately zero volts. A small voltage sufficient to overcome the barrier potential of the emitter junction is dropped across the emitter diode. At the same time the collector current I_c flows in the collector circuit. However, since the collector diode is reverse biased, it presents a higher impedance. This causes a larger voltage drop to be developed across the collector diode than the voltage drop across the forward-biased emitter diode. The result is that the collector is positive with respect to the base and the collector diode is properly reverse biased.

Notice that both collector and base are positive with respect to ground. This permits one of the batteries to be eliminated as shown in Fig. 15. The value of R_1 is now equal to V_{CC} divided by the base-bias current.

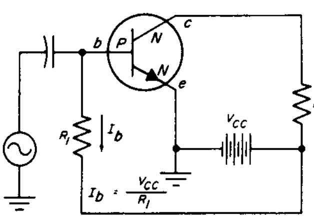


Fig. 15. Single-battery common-emitter amplifier.

Current Gain. Recall that current gain is the ratio of output current to input current. The ability of the common-emitter circuit to provide current gain lies in the fact that the input signal is applied to the base, where it adds to or subtracts from the very tiny base current. Since the base current is a fixed percentage of the emitter current, variations in the base current will cause proportional variations in the much larger emitter current. The varying emitter current will in turn cause variations in the collector current. In effect we are controlling the large collector current by variations in the small base current. Then in the common-emitter configuration, the base current is the input current and the collector current is the output current, so:

$$A_i = \frac{I_c}{I_b}$$

The current gain in the common-emitter circuit is also equal to alpha divided by one minus alpha. This quotient is known as the current-amplification factor and is given the name *beta* (β).

Then $$A_i = \beta = \frac{\alpha}{1-\alpha}$$

A typical value of beta may be calculated by assuming a typical value for alpha. For instance, assume alpha equals 0.98, then beta is 0.98/1 −0.98 = 0.98/0.02 = 49. Typical values of beta range from 20 to over 200. Remember that beta is the current gain of a common-emitter circuit.

Voltage Gain. Voltage gain is the product of current gain and the ratio of the load resistance to the input resistance. Recall that for the common-base amplifier:

$$A_v = \alpha \frac{R_L}{R_{\text{in}}}$$

In the common-emitter circuit the current gain is beta while the input resistance is the resistance at the input to the base and the load resistance is the resistance in the collector circuit. In order to calculate the voltage amplification we must first consider the input and load resistances. A typical value for common-emitter input resistance is 1000 ohms. A typical value of common-emitter output resistance is 20-k ohms.

The load resistance, placed in the collector circuit of the common-emitter amplifier seldom exceeds, and for practical reasons is often smaller than, the transistor's output resistance. A typical common-emitter amplifier load resistance is 10,000 ohms. Using the typical values of input and load resistances, voltage amplification can be calculated as follows:

$$A_v = \beta \frac{R_L}{R_{\text{in}}}$$

$$= 50 \times \frac{\text{10-k ohms}}{\text{1-k ohms}}$$

$$= 500$$

Signal Inversion. The phase of input and output signals can be determined as before, using Fig. 16.

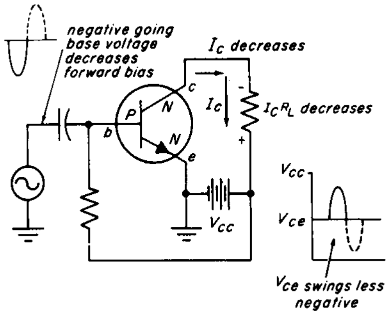


Fig. 16. Phase relationship in common-emitter amplifier.

A negative-going input signal now decreases the forward bias of the emitter junction and causes the collector current to decrease. The voltage drop across the load resistor decreases and collector voltage swings in the positive direction (closer to the supply voltage). Since in this case there is a negative-going input signal, but a positive-going output signal is produced, there is a 180-degree *phase inversion* in the common-emitter circuit.

Common-Emitter Amplifier Characteristics. The d-c current gain provided by this connection can be seen by selecting some value of base current and noting the collector current that results. Point *P* on the curve of Fig. 17 shows that at a collector voltage of +10 volts a base current of 40 micro amperes results in a collector current of 2.3 ma. However, to compute the a-c current gain (beta), a *change* in base current must be made, and the resulting *change* in collector current noted. This is done as shown in the figure by selecting a constant

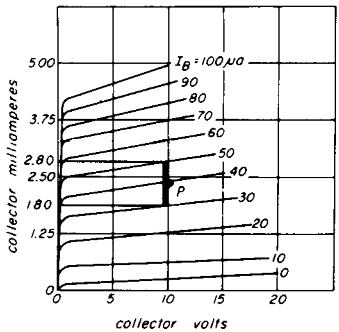


Fig. 17. Common-emitter characteristic curves.

value of collector voltage at +10 volts and using a 20-microampere change in base current from 30 to 50 microamperes. This results in a change in collector current from 1.80 to 2.80 ma or a change in I_c of 1.0 ma. Beta or current gain is the ratio of a change in I_c to a change in I_b, which in this case equals 50. Collector voltage is held constant in this computation by working on a vertical line representing +10 volts. This is done to obtain a value of beta that indicates the gain of the transistor itself and does not include other variables as would be added by the presence of a load resistor. If a load resistor is placed in the collector circuit, collector voltage varies and the current gain is therefore lower.

Note that beta varies at different points on the curve. Beta is higher at low values of collector current and decreases as collector current is increased. This is indicated by the crowding of the curves towards the top. Near the top of the graph a given change in base current yields a smaller change in collector current; beta is lower.

Leakage Current. The lowest curve in Fig. 17 is labeled I_b = zero, and represents the leakage current flowing from emitter to collector with the base terminal open. Note that it is much larger than the 10 microamperes of leakage current in the common-base circuit. This is to be expected, since the collector battery now places a slight forward bias on the emitter junction. The amount of leakage current that flows is approximately equal to beta times I_{co}. Leakage current in the common-emitter configuration is often designated I_{ceo}.

$$I_{ceo} = (\beta + 1)\, I_{co}$$

Frequency Cutoff. Recall that with the common-base circuit a certain frequency, designated as $f\alpha$, was defined as that frequency at which alpha dropped to 0.707 of its low-frequency value. This reduction in current gain was due to the action of the diffusion capacitance. In the common-emitter circuit the current gain of the transistor falls off more rapidly as frequency is increased.

4. THE COMMON-COLLECTOR AMPLIFIER

Another useful transistor circuit is the common-collector circuit. The common-emitter amplifier may be changed into a common-collector amplifier by simply moving the load resistor from the collector

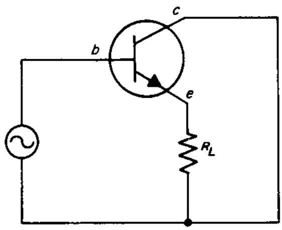


Fig. 18. Basic common-collector circuit.

circuit to the emitter circuit as shown in Fig. 18. Now the signal is applied between base and collector and the load is connected between emitter and collector. The collector is *common* to both input and output circuits. The complete circuit including a battery to supply bias is shown in Fig. 19. The battery is considered to have zero a-c impedance and thus the collector terminal is common to both circuits.

Current gain is determined by dividing output current by input current. In this circuit the input current is still the base current. The output current, flowing through the load resistor is now the emitter current (I_e). Current gain (A_i) is therefore:

$$A_i = \frac{i_e}{i_b}$$

Current gain is therefore slightly higher than in the common-emitter amplifier. However, the voltage gain (A_v) is less than one. Consider Fig. 20, where a battery is used to represent the bias voltage that normally appears between base and ground. As the signal swings negative, total emitter bias decreases and emitter current likewise decreases. This causes the voltage across the load to decrease. Since the input voltage is equal to the sum of the output voltage plus the voltage dropped between base and emitter, output voltage must be smaller than input voltage. It is smaller than the input by an amount amount equal to the voltage dropped between emitter and base. Thus, voltage gain is less than one, but very close to one, since the voltage drop across the forward-biased emitter diode is small.

Since base and emitter voltages are almost equal, very little voltage

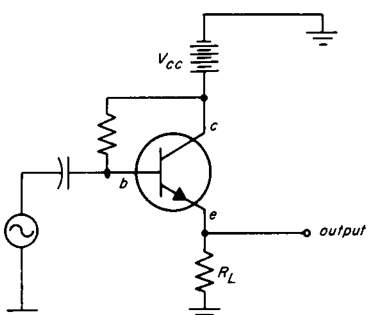


Fig. 19. Common-collector amplifier.

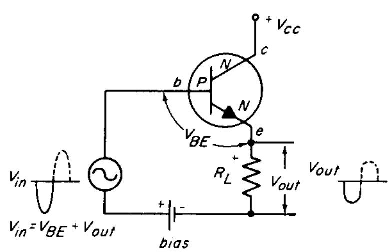


Fig. 20. Phase relationship in common-collector amplifier.

is dropped across the emitter junction and very little base current flows. Hence, a large change in base voltage produces only a small change in base current, and the input resistance to the common-collector amplifier circuit is very high.

Output resistance is low, since the output circuit is in the emitter leg and a small change in emitter voltage results in a large change in emitter current.

Base and emitter voltages are in phase in the common-collector circuit (there is no phase inversion of input and output signals). As shown in Fig. 20, a negative-going base decreases emitter forward bias and emitter current decreases. The voltage drop across R_L is as shown so that the output also swings negative.

The common-collector circuit is somewhat better in the matter of frequency cutoff than the common-emitter circuit.

The common-collector circuit amplifier is used most often as an impedance matching device to match a high-impedance source to a low-impedance load. An example is the matching of a high-impedance phono crystal to the low input-impedance of a common-emitter amplifier. In this respect the common-collector amplifier is similar to the cathode-follower electron-tube stage. Both have high input resistance, low output resistances, and voltage gains less than one. The common-collector circuit is also known as an *emitter follower*.

5. SUMMARY OF CHARACTERISTICS

The important characteristics of the three transistor amplifier circuits are summed up in Table 1. To give you an idea of practical

TABLE 1

SUMMARY OF TRANSISTOR CHARACTERISTICS

	Common Base	Common Emitter	Common Collector
Current Gain	Less than one, 0.98	High, 49	High, 50
Voltage Gain	High, 60 db	High, 54 db	Less than one
Power Gain	Medium, 32 db	High, 42 db	Medium, 16 db
Phase Inversion	None	180°	None
Input Resistance	Low, 175 Ω	Medium, 1200 Ω	High, 40 k Ω
Output Resistance	High, 500 k Ω	Medium, 75 k Ω	Low, 900 Ω
H-F Response	Highest	Lowest	Low

values, typical values are shown for input resistance, output resistance, etc. These values are for a typical low-power junction transistor and do not represent the entire gamut of transistor types.

The characteristics listed in the table govern the applications in which each circuit is found. The common-base amplifier, for example, does not lend itself well to the cascading of several stages. It is difficult to match the very high output impedance of the driving stage to the low input impedance of the following driven stage. Impedance matching is important because we are concerned with a transfer of power from one stage to another and not just voltage as is the case in electron tubes. The common-base circuit is useful as an impedance-matching amplifier where the driving source has a low output resistance and the amplified signal is delivered to a high-impedance load. An example is an amplifier which is fed from a low-impedance coaxial cable and drives the input terminals of a vacuum-tube amplifier. Very high-frequency amplifiers also utilize the common-base circuit to obtain the highest possible cutoff frequency.

The common-collector circuit is also employed mostly as an impedance-matching amplifier. Its high input impedance and low output impedance make it useful in matching high-impedance signal sources to following common-emitter stages.

The common-emitter circuit is the most widely used arrangement. Its power gain is the highest, and it lends itself best to cascading, because it has higher input impedance and lower output impedance than the common-base circuit.

6. COMPARISONS WITH ELECTRON TUBES

It is useful to study some similarities and differences between electron tubes and transistors as circuit elements.

Bias. It is important to note that tubes and transistors are exact opposites in the matter of bias. The transistor is forward-biased in the input circuit and reverse-biased in the output circuit. In the electron-tube circuit, the input signal is applied to the reverse-biased element. The grid is made negative with respect to the cathode. The plate, or output circuit, on the other hand is biased for easy current flow or is forward-biased. This also accounts for the opposite nature of input and output resistance for the two devices.

Cutoff Bias. In order to reduce the plate current of an electron tube to minimum, a negative bias voltage is applied to the grid. If bias is removed (reduced to zero) the tube conducts heavily. The transistor, on the other hand, is cut off when no bias is applied to the emitter junction and conducts heavily when a large forward bias is applied.

Voltage- and Current-Operated Devices. We are accustomed to thinking of electron tubes (in many applications) as voltage-operated devices. The reason is that input resistance is high, except at high frequencies, and signal and bias currents in the grid circuit are too low to be considered. In the simple bias circuit of Fig. 21*a*, the bias voltage applied to the grid is −6 volts because the grid-to-cathode resistance is so large that it is considered an open circuit. Thus, no voltage is lost across the 50-k ohm resistor, and the battery voltage is transferred to the grid.

The opposite is true of transistor circuits. Here input resistance (except in the common-collector circuit) is very low and both signal and bias currents are appreciable. In the bias circuit of Fig. 21*b*, the base-to-emitter resistance is low

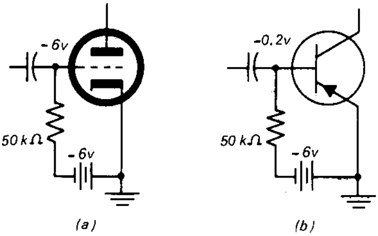


Fig. 21. Basic biasing of tube and transistor.

compared to the total resistance of the circuit, because it is a forward-biased germanium junction. Its resistance drops to only a few hundred ohms when bias voltage approaches the barrier potential. Typical bias voltages are in the neighborhood of 0.2 volts, and millivolt changes about this value result in large changes in input current. The voltage drop across the input terminals of the device is quite small compared to the voltage supply. In the matter of bias, therefore, it is

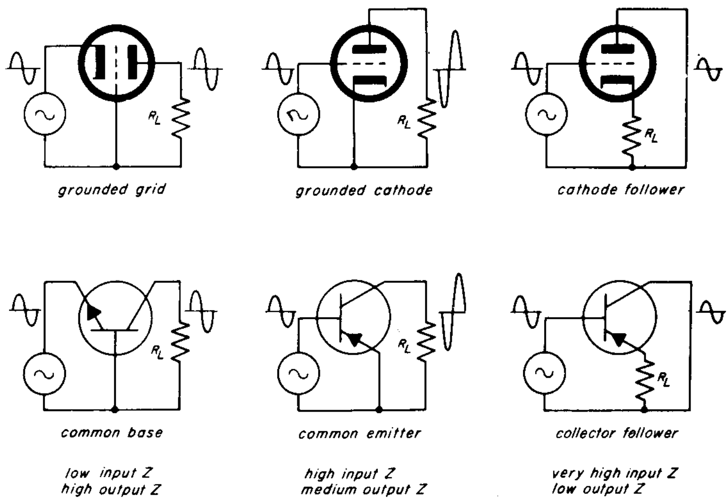


Fig. 22. Approximate comparison of tube and transistor circuits.

often more convenient to neglect input voltages and think in terms of input currents, and the effect of the external circuit upon them. For this reason collector characteristic curves are drawn showing the effect of the input *current* rather than input voltage. Also, transistors are compared in terms of *current gain* (alpha and beta), and the transistor is generally referred to as a *current-operated device.*

Electron-Tube Circuits Compared. Certain similarities exist between the three transistor arrangements and the three electron-tube amplifier circuits. These can be seen if we consider the emitter to have the same function as the cathode (a source of carriers), the base having the control function of the grid, and the collector as performing the same job as the plate. See Fig. 22. Note that all have the same relation with regard to phase inversion. Input and output impedances vary in the same direction but of course the actual values are far different.

CHAPTER 2

BASIC AMPLIFIER CONSIDERATIONS

INTRODUCTION

Now that you know how transistors work, you can begin to explore the ways in which transistors are used in electronic circuits. The basic function of the transistor is amplification. Hence, in this chapter you will study how transistor amplifier circuits are put together. You will see how the d-c operating conditions are chosen, and how the operating conditions are maintained despite temperature variations.

1. D-C OPERATING CONDITIONS

Transistors, like tubes, require certain supply and bias voltages in order to set the correct operating conditions for amplification. Bias is provided whenever an a-c signal is to be amplified. The input and output circuits of both tubes and transistors have rectifying properties; that is, current can flow in one direction only. D-c bias is provided, therefore, to act as a rest value for both input and output currents to ride upon. For instance, current in the plate circuit of a vacuum tube always flows in one direction but rises and falls from a previously selected rest or "no-signal" value. Similarly bias is provided in transistor circuits so that the "rest" currents lie somewhere between the limits of amplifying action. These limits are the states of *cutoff* and *saturation*.

Transistor Cutoff. Cutoff occurs in the common-emitter transistor circuit of Fig. 1, when the current flowing in the collector load drops to its lowest practical value. Actually, current does not fall to zero when the transistor is cut off as some *leakage* current flows in the reverse-biased collector junction. To reduce collector current to the leakage value (I_{cbo}), all current flow from the emitter to the base must stop. This happens, for all practical

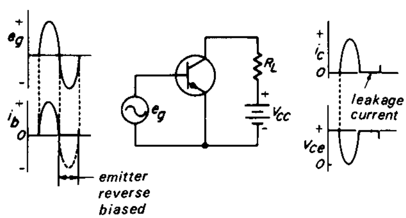


Fig. 1. The transistor is biased at cutoff when base-emitter bias is zero.

purposes, when the base-emitter junction has no forward bias. In other words when $V_{be} = 0$ and $I_b = 0$ the transistor is cut off. If a signal is applied to the base when the base-emitter bias is zero, the input signal will be rectified. After all, the base-emitter junction will act like an ordinary semiconductor diode. Consider the NPN transistor shown in Fig. 1. When the signal swings negative, the emitter junction will become reverse biased and no change will occur. On positive half cycles of signal the emitter will become forward biased and both base and collector current will increase in the usual way. Thus, the amplifier in Fig. 1 is biased at cutoff when it has no bias at all. The stage is acting as a class-B amplifier. Note that input current is rectified as well, as base current flows only during the positive half cycle.

Notice that when the transistor is cut off the emitter junction is reverse biased. Hence, in cutoff, both emitter and collector junctions are reverse biased.

In order to avoid the signal being rectified, a *minimum* amount of *forward* bias must be provided. This bias current when added to the a-c signal current prevents base current from falling to zero. In other words, to avoid cutoff, the emitter junction must remain forward biased for all values of signal voltage.

Saturation. The maximum forward bias that can be applied to the base is similar to that grid bias which causes plate-current saturation in a vacuum tube. However, the causes of saturation in a transistor are quite different from those in a vacuum tube. As the forward base-bias increases, increased collector current flows through the load resistance. The voltage drop across the load resistor increases, and the collector voltage decreases towards zero. In the case of an NPN transistor, as collector current rises, collector voltage decreases toward zero. But consider what's happening at the base as the collector voltage is approaching zero. The increasing base current is making the base-emitter voltage swing more positive. Thus, as the collector voltage approaches zero, the base voltage is more and more positive with respect to the emitter. At some point the base voltage actually becomes more positive than the collector. But at this point the collector junction is no longer reverse biased but has become forward biased. Thus, at the *saturation* point *both* junctions are forward biased. The transistor acts like a very small resistance and collector current is determined primarily by the collector load resistance and the V_{CC} supply.

Figure 2 shows the input- and output-voltage waveforms for an NPN transistor driven into saturation. Base-emitter voltage reaches a peak of +0.4 volt at the time that the voltage drop across R_L causes the collector voltage to drop to 0.2 volt. At that time the base is positive with respect to the collector

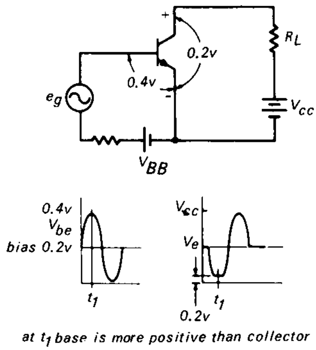


Fig. 2. Saturation occurs when both junctions become forward biased.

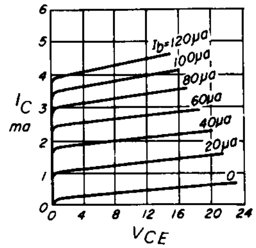


Fig. 3. Common-emitter output characteristic curves.

and the collector junction becomes forward biased. Note that the negative-going peak of the output waveform is clipped as the output voltage has fallen almost to zero and cannot go any lower.

Distortion- and Signal-Handling Ability. Transistor characteristics, like vacuum-tube characteristics, do not remain uniform between extreme limits of cutoff and saturation. For example, beta is high for low values of collector current, but decreases as collector current increases. This fact is illustrated by the common-emitter collector family of curves in Fig. 3. Note that the curves are spread out for low values of collector current and tend to bunch together for higher values of collector current. If a large signal is to be amplified, the peaks of the base signal will extend into regions where the current gain of the device will vary. This results in an unequal amplification whereby the negative half cycle of the input signal receives more amplification than the positive half cycle.

Another factor to be considered in determining the biasing conditions, is the size of the input signal that can be handled without severe distortion. The peak-to-peak signal must be limited to the difference in base current between cutoff and bottoming. That is, in order to avoid severe distortion, the transistor must be biased in such a way that the input will never cause cutoff or saturation. To handle the largest signal swing the bias must be chosen to fall halfway between these two conditions.

Classes of Operation. In the previous discussion we have specified that the transistor must never reach cutoff or saturation. These are the requirements which must be met in order to insure class-A operation. In class-A operation, signal current flows through the collector for the entire 360° of input cycle as shown in Fig. 4*a*. Transistor

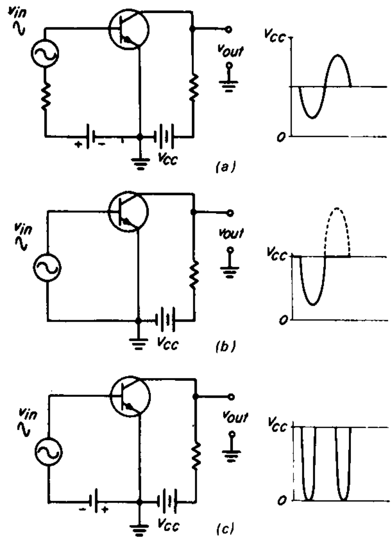


Fig. 4. Classes of operation: (a) class A; (b) class B; (c) class C.

amplifiers, like vacuum-tube amplifiers, may also be operated class B or class C.

Class-B operation is achieved by operating the emitter junction at cutoff bias. Cutoff is obtained by eliminating the forward bias from the base-to-emitter junction as shown in Fig. 4*b*. Collector and base currents flow only when the input signal swings positive, thereby producing forward bias. The negative excursion drives the emitter junction further into reverse bias, cutting off the injection of majority carriers into the base. Collector current flows for 180°, yielding class-B operation

Class-C operation is obtained by applying a d-c *reverse* bias voltage to the emitter diode, as shown in Fig. 4*c*. In this case, forward-bias current can flow only during that part of the positive portion of the input signal that exceeds the bias voltage. Collector current then flows for less than 180°.

2. BIAS STABILIZATION

The operating point of a transistor circuit refers to the no-signal or quiescent value of the collector current and collector voltage. This operating point is determined, once the transistor and load have been selected, by the no-signal bias of the circuit. Bias stabilization refers to the control of that bias so as to prevent, or at least minimize, shifting of the selected operating point.

Bias stabilization is needed in transistor circuits to prevent the operating point from moving away from its selected value due to the effects of temperature and variations in transistors themselves.

Preventing a shift in operating point is extremely important. Almost all the important features of a transistor circuit depend upon carefully chosen values of no-signal collector voltage and no-signal collector current. An increase in the no-signal current raises the power dissipation and may reduce the life expectancy of the unit. If the additional heat produced by the increased dissipation is not carried away quickly enough, a *thermal-runaway* condition may occur, which destroys the transistor. The change in operating point may reduce the signal-handling ability of the circuit, and thus bring about more distortion. In addition, the input resistance, output resistance, and current gain of the transistor change when the operating point is altered.

Temperature Effects. Both semiconductor leakage current and barrier potential are sensitive to temperature changes. An increase in temperature results in an increase in leakage current. It has been found that leakage current doubles for every 8° C rise in temperature in germanium junctions. In silicon units, leakage doubles for every 5° C rise in temperature. Although silicon units appear to be more sensitive to changes in temperature, the total leakage current in silicon transistors is about 100 times smaller than in similar germanium transistors. Consequently, temperature effects are less of a problem in silicon transistors.

One result of the increased leakage current is that the operating point moves up the load line. As the operating point is no longer in the center of the load line, the signal-handling ability of the amplifier decreases. Large signals will cause clipping by saturation limiting.

Thermal Runaway. Another temperature effect in transistors that must be guarded against is the possibility of thermal runaway. An external temperature increase raises the leakage current. This increased leakage adds to the collector current. More collector current raises the power dissipated in the transistor, and increases the amount of heat produced. This brings us back to the starting point, since increased temperature raises the leakage some more, and the cycle repeats itself, increasing the collector current more and more. This thermal runaway cycle can ruin a transistor in a very short time. It is guarded against by limiting the collector current, proper stabilization, and by use of a heat sink, which is designed to carry the heat away from the transistor.

Beta Variations. The transistors themselves cause changes in circuit-operating conditions due to variations in constants, notably beta variations, between transistors of the same type. Beta may vary between 50 and 100 for transistors of the same type.

Stability Factor. In order to compare various biasing methods, the concept of a *stability factor* (S), is used. The stability factor is defined as a change in collector current divided by the change in leakage current producing it.

$$S = \frac{\Delta I_c}{\Delta I_{co}}$$

Thus, the stability factor tells how much a change in collector-junction leakage current will be multiplied. A stability factor of 1, as exhibited by the common-base circuit, is considered to be the best possible rating. A higher stability factor represents a *less* stable circuit. The problem of stabilization is due to the fact that in the common-emitter amplifier, any change in leakage current is amplified by beta.

Now let us consider various ways in which circuit designers overcome the effects of amplified leakage current. While considering the following conditions, it might be helpful to bear in mind that the main objective is to establish some means of insuring a minimum change in collector current for a given change in leakage current.

The circuits of Fig. 5 approach highest stability when resistors R_1 and R_2 are made as small as possible and resistor R_e is made *as large as possible*. Small resistances result in very good voltage regulation at the tap, and base voltage remains constant. In practice R_1 and R_2 are moderately large to minimize battery drain through the voltage divider. A compromise is reached between stability factor and battery drain.

Figure 5 shows two forms of bias circuit. In Fig. 5*a* signal is coupled into the base circuit by means of a transformer. Note, that the signal is in series with the base-bias resistors. This is called a series fed arrangement and is used in transformer-coupled circuits. Capacitor C is added to bypass the a-c signal around the bias network.

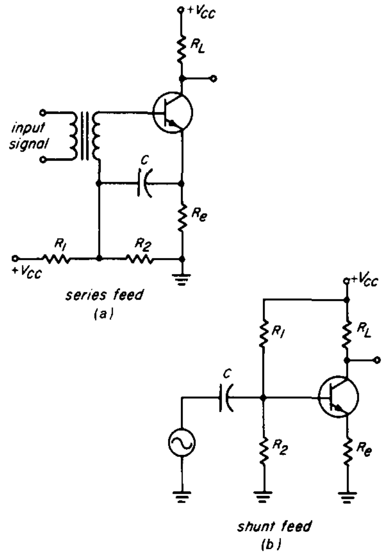


Fig. 5. Types of bias stabilization circuits.

Figure 5*b* shows a *shunt feed* system used with capacitive coupling. Here the signal is applied in shunt with the base-bias resistors. In this case R_1 and R_2 should be large, as both shunt the signal path. Note that the top of R_1 is at a-c ground (the battery voltage) and is therefore also across the signal source. The resistors represent a shunt path for signal that results in less signal current for the transistor. In this case a compromise must be made in choosing R_1 and R_2, between good bias stabilization and signal loss. For least signal loss R_1 and R_2 should be as large as possible; for best stabilization R_1 and R_2 should be as small as possible.

Feedback. The effectiveness of the bias stabilization of the circuits of Fig. 5 can be analyzed by considering the feedback voltage that is developed across R_e. Consider the circuit shown in Fig. 6. The base voltage divider holds the base voltage at 2 volts. About 0.2 volt is dropped between base and emitter, so that the voltage from emitter to ground is 1.8 volts. Thus, emitter current is 1.8 volts/1800 ohms = 1 ma. Now, suppose an increase in temperature causes collector current to increase. Emitter current will tend to rise as well. But an increase in emitter current raises the emitter voltage. Suppose it goes up to 1.9 volts. This leaves only 0.1 volt between base and emitter. In other words, an increase in emitter current results in a smaller base-emitter bias voltage. This in turn acts to reduce emitter current so that the original increase is cancelled to some extent.

The action described in the previous paragraph works best when the base voltage is held constant. Hence, R_1 and R_2 should be small to provide a low-impedance source of base voltage. In addition, R_e is made as large as possible because the feedback voltage, developed across R_e when emitter current changes, is directly proportional to the resistance of R_e. The maximum value for R_e is determined by how much of the total supply voltage can be allocated to the emitter resistance for bias-stabilization purposes.

Emitter Bypass. You saw that the voltage developed across R_e in Fig. 6 acts to oppose changes in collector current. This opposition acts to oppose *signal* changes in collector current as well. The circuit provides degenerative feedback that results in a decrease in stage gain. To restore signal gain, a bypass capacitor is placed across R_e. The capacitor prevents a-c voltage from being developed across R_e, but the *slow* changes in voltage resulting from temperature effects, still appear across R_e. The bypass capacitor must be quite large to be an effective bypass at low frequencies, as the total resistance from emitter to ground is quite small. Typical values for the emitter bypass capacitor are 50 to 1000 μf. Electrolytic capacitors must be used, but they are quite small physically as they have a very low voltage rating.

Collector Feedback. Another circuit used to stabilize the operating point by means of feedback is shown in Fig. 7. Here, the base resistor, R_b, is returned directly to the collector terminal instead of the supply voltage. Thus, the voltage applied to the base resistor is dependent upon the amount of collector current. If we consider the

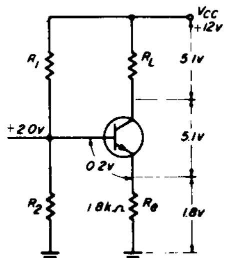


Fig. 6. Feedback voltage is developed across the emitter resistance.

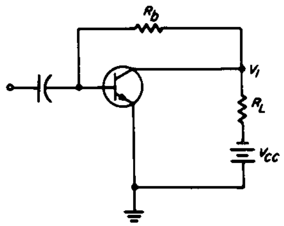


Fig. 7. Collector-feedback circuit.

base-emitter voltage to be negligible (normally a few tenths of a volt), the base current can be found by dividing the collector-to-emitter voltage by R_b.

To see how feedback via R_b acts to stabilize collector current, consider an increase in I_C (due, perhaps, to an increase in temperature). As I_C increases, collector voltage drops towards zero. But as collector voltage acts as the source of base current, base current falls as well. Thus, an increase in collector current is offset by a reduction in base-bias current. A decreasing collector current, on the other hand, will cause more base current to flow.

Realize that the feedback discussed in the previous paragraph is present at all frequencies as the base is resistor-coupled to the collector. Thus, degenerative feedback is active for signal as well as for the slow changes in collector current that might be caused by temperature effects.

Degeneration at signal frequencies can be eliminated by modifying the bias circuit as shown in Fig. 8. Resistors R_1 and R_2 are each one-half the value of the base resistor

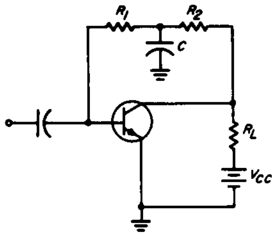


Fig. 8. Collector-feedback circuit with a bypass capacitor added to eliminate degeneration of signal frequencies.

R_b shown in Fig. 7. Thus, the total base resistance is unchanged. A bypass capacitor, *C*, is placed between the junction of R_1 and R_2 and ground. This bypass capacitor acts to short a-c signals to ground. Thus, a-c variations in collector voltage do not reach the base terminal. Resistor R_1 isolates the capacitor from the signal-input terminal so that input signal is not diverted from the transistor.

Collector feedback-stabilization circuits, such as the one shown in Fig. 8 are used in low-level RC coupled stages where the *d-c resistance* in the collector circuit is large. A large collector resistance is needed to develop the feedback signal. This system is seldom used in transformer-coupled circuits as the d-c resistance in the collector circuit is the small winding resistance.

3. METHODS OF COUPLING

Several ways are used to couple transistor circuits in cascade to produce amplifier chains. As with vacuum tubes, the most common coupling schemes are direct coupling, transformer coupling and RC coupling.

Direct Coupling. Direct coupling is used when very low-frequency signals are to be amplified. There are no reactive components in the coupling system so that frequency response remains flat down to zero cycles per second (d.c.).

Transistors lend themselves to direct coupling more easily than vacuum tubes because transistors will function with low values of collector voltage. An example of a direct-coupled transistor amplifier is shown in Fig. 9. Notice that the collector of Q_1 connects directly to the base of Q_2. You may think of R_{L1} and Q_1 acting like the voltage divider R_1 and R_2 of previous circuits to provide base voltage for Q_2. If R_{L1} is made very large, any *change* in collector current for Q_1 will appear as a change in base current for Q_2.

One of the disadvantages of the system shown in Fig. 9 is that it provides very little gain. Each stage has an unbypassed emitter resistance that introduces degenerative feedback and a considerable sacrifice in gain. The emitter resistors cannot be bypassed as any capacitor behaves as an open circuit at zero cycles (d.c.). Bypass capacitors would increase gain at higher frequencies but are no help at all at very low frequencies.

Another disadvantage of the direct-coupled amplifier is poor stability. Any change in the collector current of Q_1, due to thermal effects, shows up as a change in base current for Q_2. Thus, thermal changes are amplified. Small changes in the operating point for the first stage appear as very large changes in the operating point of the final stage.

A direct-coupled amplifier that has practical advantages is shown in Fig. 10. This amplifier consists of two (or more) direct-coupled emitter followers (common-collector amplifiers). Here the emitter current of Q_1 is the same as the base current for Q_2.

Over-all current gain is the product of the current gains of the individual stages. Thus current gain $= (\beta_1 + 1)\ (\beta_2 + 1)$. If beta is large compared to one, we can neglect the ones. Current gain, therefore, is approximately equal to the

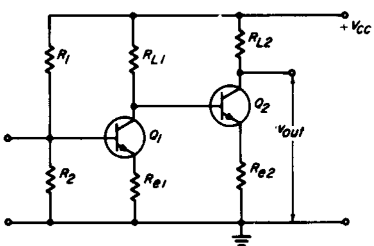


Fig. 9. Direct-coupled transistor amplifier.

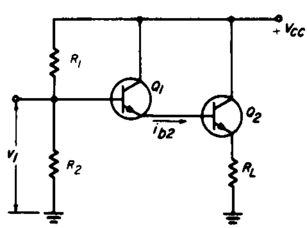


Fig. 10. Direct-coupled emitter followers.

product of the beta of the individual transistors. For this reason amplifiers of this type are sometimes called "super-beta" amplifiers.

Another interesting fact about the cascaded emitter followers is that they provide very large values of input resistance. The resistance looking into the base of Q_1 is approximately equal to the product of the individual betas times R_L. If both transistors have a beta of 50 and R_L is 500 ohms, the input resistance to the amplifier will be $(50)^2 \times 500$ ohms $= 2500 \times 500$ $= 1.25$ megohms.

You see that the circuit of Fig. 10 provides a very high current gain and a high input resistance. Its voltage gain, like that of a single emitter follower is less than one. Look at Fig. 10. You will see that the output voltage must be less than V_1 by the voltage drop across the base-emitter terminals of both transistors.

Differential Amplifier. All simple direct-coupled amplifiers are subject to drift as small changes in current of the early stages are amplified by subsequent stages. However, direct coupling is necessary in many applications. Hence, some means must be found to construct a direct-coupled amplifier that has little drift. The solution is the differential amplifier, one form of which appears in Fig. 11. Each stage of a differential amplifier requires two transistors. The transistors are connected so that only signal currents effect a change in output. Consider

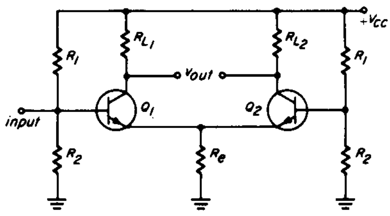


Fig. 11. Differential amplifier.

the effects of an increase in temperature. The leakage current of both transistors would increase. Collector voltage for both transistors would change by the same amount. However, the output voltage is the *difference* between the collector voltages for Q_1 and Q_2. Therefore, any factor that causes collector currents for both transistors to change by the *same amount* does not alter output voltage. To maintain this balanced condition the bias circuits and the transistors themselves should be matched. In addition, the transistors are sometimes mounted in a common heat radiator or mounting system so that both transistors are exposed to the same temperature conditions.

A signal applied to just one of the transistors acts to unbalance the system. Consider a positive signal applied to the base of Q_1. The transistor Q_1 will conduct more heavily and its collector voltage will drop towards zero. More current flows in the emitter circuit of Q_1 as well. As a result the emitter voltage tends to swing positive. But this acts to reduce the emitter-base bias on Q_2, and Q_2 conducts less. Thus, the positive signal causes the collector voltage of Q_1 to fall and the collector voltage of Q_2 to rise. The output voltage is the difference between these two voltages.

Note that the emitter resistor requires no bypass capacitor. Any signal current that acts to make Q_1 conduct more heavily also causes Q_2 to conduct less heavily. Thus, the total current through R_e tends to stay the same. If current through R_e remains substantially unchanged, no signal voltage will be developed across R_e, and hence there is no degenerative feedback developed in the emitter circuit. Figure 12 shows how differential amplifier stages are coupled together to form a direct-coupled amplifier chain. This amplifier has "floating" input terminals. Neither of the input terminals is referenced to ground. Potentiometer R_1 is a balance adjustment; it is set to balance the output voltage to zero under no-signal conditions.

Transistor Coupling. The most desirable feature of transformer coupling is its ability to achieve any degree of impedance match between a source and a load. Maximum power is transferred from a source to a load when the internal resistance of the source equals the load resistance. This is also true of coupling circuits. That is, maximum power is delivered from the transistor to its load when the load resistance equals the output resistance of the transistor. Similarly, maximum power is delivered from the signal source to the input of the transistor when signal-source impedance and input impedance "match." Maximum power gain occurs when the transistor receives maximum power from the signal source and transfers maximum power to the load. Only transformer coupling allows impedance to be changed to obtained "matched" conditions. The use of a transformer as an impedance-matching device is shown in Fig. 13. Here a transistor with an output impedance of 30,000 ohms is transformer-coupled to another transistor with an input impedance of 1000 ohms. The turns ratio of the transformer is chosen to effect an impedance match.

$$N = \sqrt{\frac{Z_p}{Z_s}}$$

$$= \sqrt{\frac{30{,}000}{1000}} = \sqrt{30} = 5.5$$

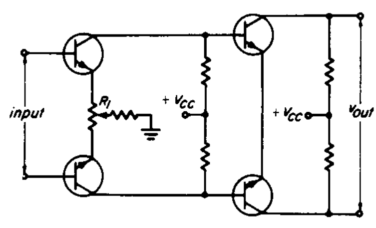


Fig. 12. Direct-coupled differential amplifiers.

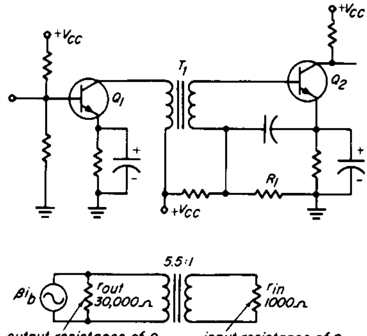


Fig. 13. Transformer coupling.

Thus, T_1 has a 5.5:1 step-down voltage ratio or a *current* step-up ratio of 5.5:1. Transformers used in transistor coupling are always voltage step-down transformers.

Transistor Q_1 now "sees" a load impedance of 30,000 ohms instead of the 1000-ohm load represented by the input resistance of the next stage. (The terms resistance and impedance appear interchanged in places. If no phase shift is involved, impedances can be considered resistive.)

As there is no direct coupling between primary and secondary, the d-c operating conditions of transformer-coupled stages are isolated from one another. This means that d-c changes resulting from temperature effects are not coupled from stage to stage.

There are several disadvantages of transformer coupling. Transformers are expensive. It is often cheaper to add an additional RC coupled transistor stage than to employ transformer coupling betwen a pair of stages. Transformers have generally poor frequency response. To have good response at low frequencies requires a high primary inductance and many turns of wire. High frequency response, on the other hand, is reduced due to the effects of wiring capacitance when large windings are employed. Transformers with good frequency response must be very carefully designed and are quite expensive.

RC Coupling. Resistor-capacitor coupling is probably the most common form of coupling in low-level audio stages (the preamplifier). A typical RC coupled stage is shown in Fig. 14. The coupling capacitor (C_c) passes the a-c signal from Q_1 to Q_2 but blocks the d-c voltages. Capacitor C_c is selected so that it offers negligible impedance at the lowest frequency to be amplified.

The base resistance of transistor Q_2 is quite low (around 1000 ohms) and its biasing resistors (R_1 and R_2) are effectively in shunt with the base resistance and R_c. Transistor Q_1 therefore has a load resistance of something less than 1000 ohms. This represents a serious mismatch, as the output resistance of the transistor itself is 50,000 ohms. Thus, this transistor circuit is never matched and power gain has been sacrificed. However, the low cost and good frequency response of the RC system makes up for the sacrifice in gain. Additional stages can be added to make up the required gain.

4. FREQUENCY RESPONSE

The factors that affect frequency response in transistor amplifiers are similar to those affecting vacuum-tube amplifiers. Some important differences turn up, however, due

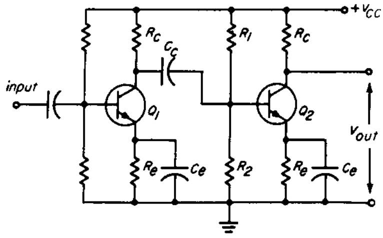


Fig. 14. An RC coupled stage.

to the low-input resistance of transistor amplifiers and the fact that beta changes with frequency.

Low-Frequency Response. The coupling capacitor in RC coupled amplifiers (Fig. 15) is in series with the signal path. At very low frequencies the coupling capacitor represents a large impedance that effectively breaks the signal path. At mid and high frequencies the reactance of C_c is negligible so that most of the signal current flows into the transistor's input terminals. The transistor's input resistance is usually much smaller than the output resistance of a previous stage (r_o) or the internal resistance of a signal source so that most signal current flows in the transistor's input terminals. At lower frequencies the impedance of the capacitor increases so that the division of current between r_o and the transistor changes. Less current flows in the transistor and more current is diverted through r_o. When the reactance of the transistor-capacitor branch approaches r_o, the division of current splits so that currents in the two branches become almost equal. At this point, current in the transistor's input terminals is about 70% of its mid-frequency value.

Note that the capacitor becomes important in diverting current away from the transistor terminals when its reactance approaches the resistance of the shunt path formed by r_o. To achieve a negligible decrease in transistor-input current, the reactance of C_c at the lowest operating frequency should be about one tenth the input impedance of the transistor. Typical capacitance values are large. Coupling capacitors are often electrolytic capacitors in the 1 to 5-μf range.

Emitter-Bypass Capacitor. The emitter resistor, used for bias stabilization purposes, introduces local degeneration into the common-emitter amplifier. To eliminate degeneration for signal frequencies, a capacitor is placed across the emitter resistance as shown in Fig. 16. This capacitor effectively puts the emitter at a-c ground and eliminates degenerative feedback. At low frequencies, however, the bypassing action of the capacitor diminishes as the reactance of C_e

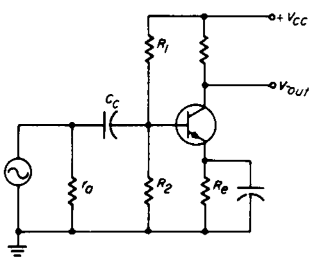


Fig. 15. Input circuit to a transistor stage.

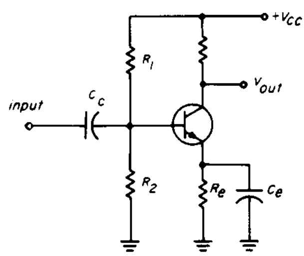


Fig. 16. Increasing reactance of C_e at low frequencies introduces degeneration.

rises. Thus, degeneration is introduced at low frequencies and gain is reduced.

To be an effective bypass capacitor, the impedance of C_e should be small compared to the total resistance between emitter and ground at the lowest frequency to be amplified. Unfortunately, the total resistance betwen emitter and ground is smaller than resistor R_e. An internal emitter resistance appears in shunt with the external-emitter resistor (R_e). The parallel combination of these two resistances represents the equivalent resistance that C_e must bypass. The internal-emitter resistance is approximately equal to the total resistance in the base circuit divided by the transistor's beta, hence it is usually quite low. Since the emitter bypass capacitor (C_e) must have a reactance about one-tenth of the total emitter resistance at the lowest frequency to be passed, the emitter bypass capacitor must be very large in order to be an effective bypass. Typical values for emitter bypass capacitors at audio frequencies range from 20 to 1000 microfarads. They must be electrolytic capacitors. However, they can be quite small physically as units with very small working voltages are chosen.

High-Frequency Response. Amplification decreases at high frequencies in part because of stray and junction capacitances that are in shunt with the signal path. These reactive shunt paths divert signal current around the amplifier load resistances. However, this effect is

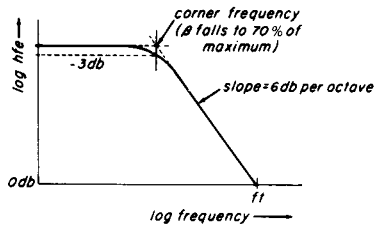


Fig. 17. A graph of beta versus frequency plotted on a log-roll scale.

not quite as pronounced as it is in vacuum-tube circuits because the load and input resistances are relatively small in transistor circuits.

The major cause for a decrease in gain at high frequencies is found in the transistor itself. Collector current cannot "follow" very rapid (high frequency) changes in base-emitter bias due to diffusion delay.

A useful way of showing the effect of the diffusion delay is to show how current gain or beta varies with frequency. A typical response curve showing beta versus frequency is shown in Fig. 17. The curve is horizontal at low frequencies. It begins to roll off and drops 3 db at the point where input current splits evenly between the emitter junction and the diffusion capacitance. At higher frequencies, current gain decreases rapidly. However, if frequency and gain are plotted using a log scale, as in Fig. 17, the rate at which gain decreases is a straight line that terminates at zero db or unity gain (the log of 1 is zero).

5. THE TUNED H-F AMPLIFIER

In this section, the considerations peculiar to tuned amplifiers are introduced.

In a superheterodyne receiver, when a tuned r-f amplifier rather than the mixer or converter is used as the first stage, it provides better noise performance and more selectivity. The selectivity is usually required to reduce interference from adjacent channels as well as from the image frequency. The type of coupling circuit employed between the antenna and the input terminal of the r-f stage depends on both the type of antenna and the bandwidth required.

R-F Amplifiers. At broadcast frequencies two main types of antennas are employed: the capacitive antenna, of the type typically used with automobile radios, and the inductive or loop antenna. The capacitive-antenna first stage is shown in Figure 18. For considerations relating to sensitivity and noise this type of antenna is usually tuned by varying the inductance (the so-called "perm" tuner). The variable-inductance antenna has a secondary winding to couple the antenna signal to the r-f stage. The sensitivity of the receiver is directly proportional to the unloaded Q of the resonant circuit.

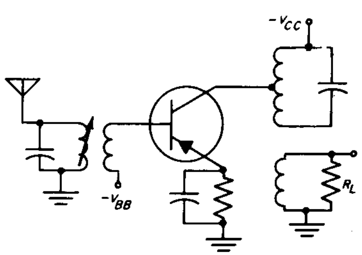


Fig. 18. Auto r-f stage.

The loop antenna is much less efficient than the capacitive antenna at broadcast frequencies, and is more frequently used with a mixer or converter first stage. Occasionally, however, a loop antenna r-f stage, of the type shown in Figure 19, is employed. The antenna winding, which is capacity tuned, employs a secondary winding to couple to the r-f stage input. At frequencies above the broadcast band, the loop antenna becomes increasingly efficient. At frequencies exceeding several megacycles, the efficiencies of the two types of antenna become comparable and both are frequently used.

At frequencies approaching the television band, the antenna typically becomes more complex, exhibiting both reactance and resistance. At these frequencies, the antenna may be either tuned, as illustrated, or somewhat decoupled from the tuned circuit. The purpose of decoupling is to preserve the selectivity of the antenna circuit as the resistance and reactance of the antenna varies with frequency.

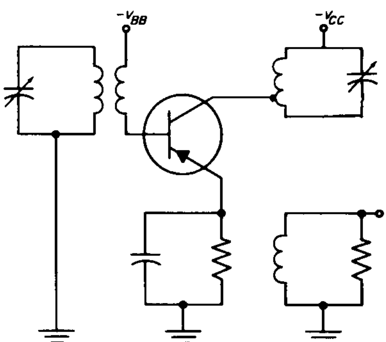


Fig. 19. Loop-antenna r-f stage.

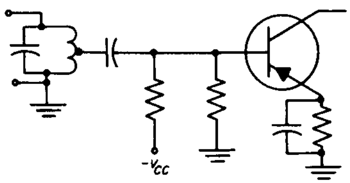


Fig. 20. TV r-f stage.

At television frequencies, the antenna is usually considered to be primarily resistive. Here the coupling network used, shown in Figure 20, is very similar to tube circuits. A single tuned circuit is employed, correctly tapped so as to efficiently couple the 300-ohm antenna (or perhaps 75 ohms) to the input resistance of the transistor. The circuit tuning is very broad to keep the insertion loss to a minimum and to prevent small amounts of reactance associated with the antenna from severely tilting the passband. Reactance associated with the input resistance of the transistor is tuned out, as at the lower frequencies.

I-F Amplifiers. Most of the important factors concerning tuned i-f amplifiers have already been discussed. There are several additional items, relating to cascaded stages and to overall receiver operation, which remain. A two-stage cascaded i-f amplifier is shown in Fig. 21. The amplifier employs three single-tuned transformers, the first transformer being fed from the mixer stage and the third transformer coupling to the detector stage. When several similar stages are cascaded as shown here, the increased overall amplification reduces the values of capacitance required to produce oscillation.

Overall selectivity of a cascaded i-f amplifier is determined by the product of the operating Q's of the individual stages.

A cascaded double-tuned i-f amplifier is shown in Fig. 22, and a combination double-tuned and single-tuned amplifier is shown in Fig.

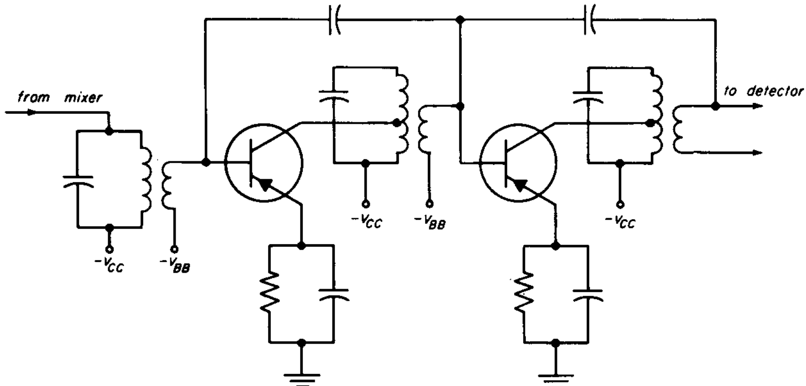


Fig. 21. I-f amplifier.

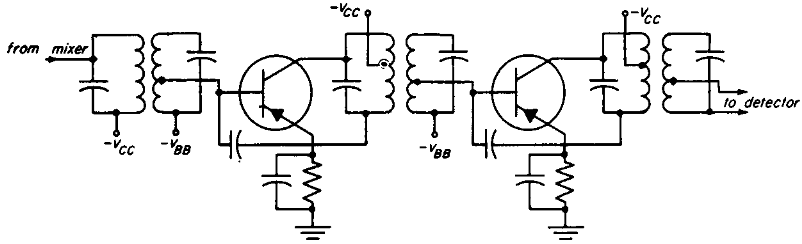


Fig. 22. Double-tuned i-f amplifier.

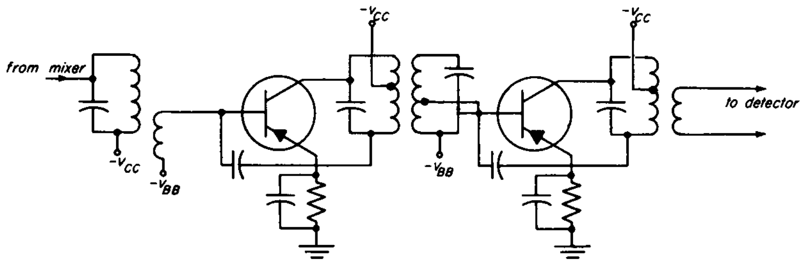


Fig. 23. Combination double-tuned and single-tuned i-f strip.

23. The double-tuned transformer provides a flatter response to the desired frequencies and gives more steeply sloping sides to the selectivity characteristic of the i-f amplifier.

6. GAIN CONTROL

The variation in amplifier gain that may be obtained by varying the emitter current for a transistor-tuned amplifier is shown in Fig. 24. The gain shown is the circuit power gain, that is, the transistor gain less insertion loss of the input and output transformers. This is the type of control that may be achieved in a receiver and permits a single-stage control range in excess of 50 db. Two factors bring about this variation in power gain. The reduction in emitter current results in an increase in both input and output

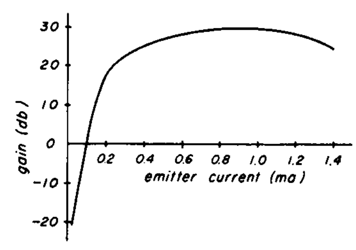


Fig. 24. Gain vs. emitter current.

resistance of the transistor. This increase in transistor resistances results in transformer mismatch and increased insertion loss in the transformers. Of greater importance, however, is the reduction in transistor g_m.

Base-Voltage Control. There are two general methods of varying the emitter current to obtain gain control. One method consists of varying the base-to-emitter voltage and thereby varying the emitter current; the other in supplying a control current to reduce the emitter current directly. These methods are illustrated in Fig. 25 and Fig. 26. In the circuit of Fig. 25, below gain-control threshold, the voltage at the base is sufficient to provide normal emitter current in the transistor, say about 1.0 milliampere. The gain of the stage is then reduced by causing the voltage at the base to become more positive, for the PNP transistor shown. When the base voltage is equal to zero, the emitter current is also zero and the stage is cut off. The amount of attenuation provided by the stage (usually in the order of 20 db) then depends on capacitive feedthrough. One feature of this type gain control, which is sometimes a significant disadvantage, is the required voltage swing at the base. As shown, some emitter resistance is usually used for d-c bias stability. This provides reasonable interchangeability of transistors. The larger the value of this resistance for a given initial bias current, the greater the required voltage change at the base. For example, if the base voltage to ground is one volt, the gain-control voltage must be at least one volt to cut the transistor off.

Emitter-Current Control. In the arrangement of Fig. 26, the point (gain-control threshold) where gain-control action starts, the stage is constant-emitter-current biased in the conventional fashion. For the zero-signal condition, substantially no d-c emitter current flows in the detector circuit. As signal is increased, detector d-c emitter current is developed, and flows through the resistor *R*, in the emitter circuit of the first i-f stage. Since the

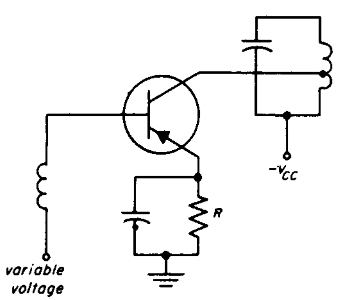


Fig. 25. Base-voltage gain control.

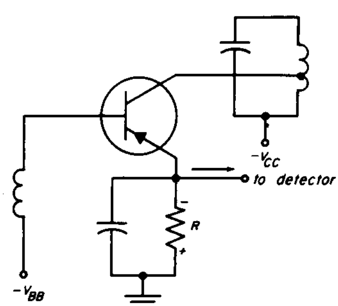


Fig. 26. Emitter-current gain control.

d-c bias arrangement of the first i-f stage holds the current in R essentially constant, the detector may be considered to "rob" the i-f stage of emitter current, i.e., the d-c emitter current shifts from the i-f stage to the detector. Thus, the transistor current decreases in proportion to the applied signal. When the control current multiplied by R is equal to the magnitude of the base voltage, the stage is cut off. At cutoff the magnitude of the control current is about equal to the original bias current of the controlled stage.

In all transistor receivers gain control is applied to the first i-f amplifier. In those receivers employing an r-f first stage, gain control is applied to a minimum of the r-f stage and the first i-f amplifier. Occasionally, additional stages, such as the mixer, are also gain controlled. As has been indicated, the control range obtainable with one transistor amplifier is quite large. The reason for control of more than one stage is to obtain flatness of the control characteristic and to avoid overload conditions. Invariably, the gain-control signal is obtained from a stage following the controlled stages. Generally, the higher the gain preceding the control signal source and the larger the number of stages controlled, the flatter the control characteristic. Overload of a transistor stage may occur in two ways. First, if the peak signal swing at the output becomes larger than the supply voltage. (This is frequently termed bottoming after a similar overload condition in tube amplifiers.) The second type of overload is due to excessive signal swing at the input of the transistor. This is the type of overload usually encountered in gain controlled stages. The overload is due to clipping caused by the nonlinearity of the transistor input characteristic. At low values of bias current, the transistor exhibits significant nonlinearity for values of input signal exceeding about 20 millivolts. This merely means that the signal at the input to a gain-controlled stage must not exceed this value if linear operation is to be maintained.

The predominant effect of gain control on band shape is caused by the increase in transistor input and output resistance which accompanies the reduced emitter current. This leads to an increase in circuit Q for the transformers which couple the controlled stages. The magnitude of the increase depends in part on the frequency of operation and in part on the degree of stability for which these stages are designed. In a practical receiver the increased Q may be offset by additional loading of the detector coupling transformers. (The detector input resistance decreases with increasing input signal.) The two effects may be made to cancel each other to such an extent that negligible change in overall receiver bandwidth occurs. Detuning effects in transistor amplifiers are generally not significant because capacitive loading of the interstage networks is usually small compared to the resistive loading.

7. NEUTRALIZATION

The principle of neutralization is that feedback can be balanced out by connecting another feedback path, having equal amplitude but opposite phase, back to the input, as shown in Fig. 27. In this diagram, C_f represents the internal feedback, and C_n the neutralizing capacitance. The sum of these two feedback signals is then zero and there is no net feedback. The amplifier acts as though both feedback paths were disconnected. A reverse-phase voltage is obtained from a secondary winding on the output inductance. If the magnitude of the voltage appearing across the secondary winding is equal to the voltage across the tuned circuit, the feedback may be balanced out by making C_n equal to C_f. This process of balancing out reactive feedback is termed neutralization. Whatever the turns ratio from primary to secondary is, the neutralizing capacitor, C_n may be adjusted to provide neutralization. For instance, if a 3 to 1 stepdown winding is employed the size of C_n must be 3 times the the size of C_f.

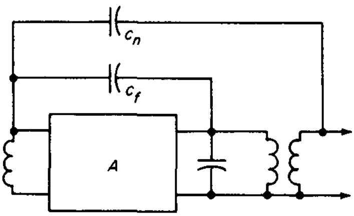


Fig. 27. Principle of neutralization.

A more sophisticated form of neutralization is called *unilateralization.* In this process a resistive element is included in series or shunt with the neutralizing capacitor to exactly match the phase shift (away from 180°) in the positive-feedback path caused by resistive elements inside the transistor.

Practical Circuits. Two forms of unilateralized transistor amplifier stages are shown in Fig. 28. Both circuits employ a tuned transformer for coupling to a load. The secondary winding is poled so as to

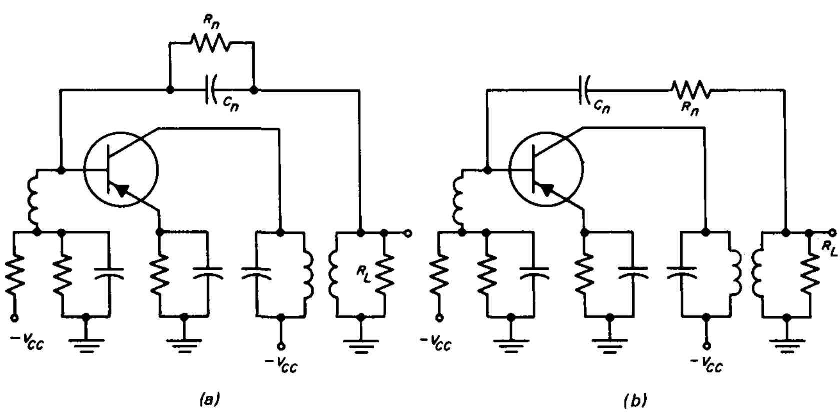


Fig. 28. Unilateral i-f stages.

provide the desired phase reversal. In the circuit shown in *a*, a shunt resistor is employed to provide the required phase shift. The size of the resistor depends on the operating frequency, being in the order of 100-k ohms for typical broadcast i-f transistors. In circuit *b* a series resistor is employed. his resistor is usually quite small, again depending on the operating frequency and the size of C_n. The size of C_n is almost the same for both circuits, for frequencies where the reactive feedback is large compared to the resistive feedback. C_n is closely equal to C_c multiplied by the output transformer turns ratio. The series feedback arrangement *b* closely approximates the internal feedback circuit and provides an accurate unilateralization signal over a relatively wide frequency range. The parallel circuit *a* approximates the impedance and phase shift of the series arrangement for a narrow frequency range and provides satisfactory unilateralization for most narrow-band tuned amplifiers.

Some forms of neutralized amplifier stages are illustrated in Figs. 29, 30, and 31. The circuit in Fig. 29 is similar to the unilateralized stage, but resistive feedback is not employed. The circuit in Fig. 30 uses a tapped inductance to obtain the phase reversal for neutralization. The circuit in Fig. 31 employs a pi section with capacitive voltage division. The voltages across the two capacitors are out-of-phase, the requirement being that the capacitor from which the feedback voltage is obtained have low reactance

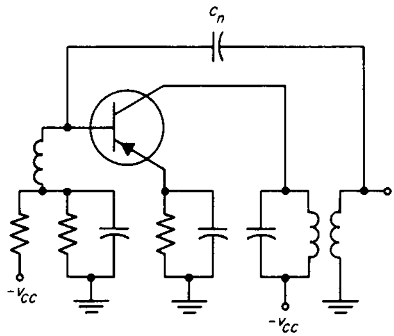


Fig. 29. Neutralized i-f stage.

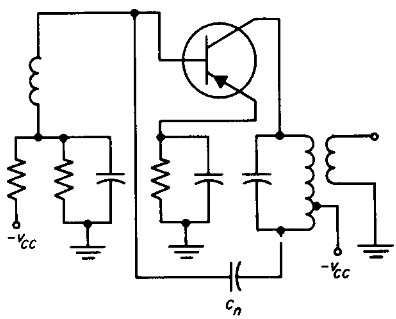


Fig. 30. Tapped-inductance type neutralization.

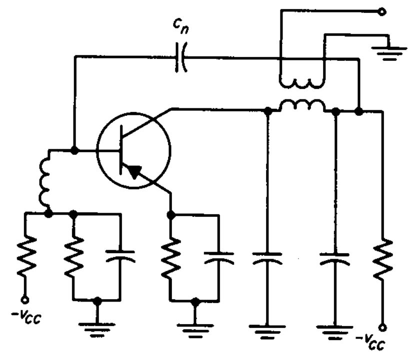


Fig. 31. Feedback voltage obtained from pi-network.

compared to the shunt resistor. This capacitor and shunt resistor may also be used to provide decoupling from the supply. The latter two circuits are most frequently used in circuits employing double-tuned transformers.

Stabilization by Gain Reduction. The other method of stabilization is to reduce the gain. When this method is used the positive feedback is still present, but the gain is reduced to a point where the feedback is not effective.

A transistor amplifier stabilized by gain reduction has the disadvantage of lower gain than the neutralized transistor. However, the neutralized transistor amplifier has the disadvantage of requiring close tolerance components to maintain accurate cancellation of the positive feedback. Furthermore, the neutralization afforded by a fixed circuit is limited to a narrow frequency range.

In general, these high-frequency amplifier circuits are therefore designed to allow for some imperfection in neutralization. Not only is neutralization an approximation to the desired unilateral condition but variations among transistors lead to further imperfection. Variations of the order of ± 20 percent in collector capacitance are typical of many transistor types. For a transistor with a nominal C_c of 12 micromicrofarads, this leads to a misneutralization range of ± 2.4 $\mu\mu$f. This amount of feedback can be serious at certain frequencies if the source and load impedances are high. Further variation in collector capacitance is caused by supply-voltage variations which may add a like amount of increase in collector capacitance. Still another error may arise when a fixed capacitor is used for neutralization. Tolerances of ± 10 percent are usually used for this component so that another micromicrofarad may be added to the neutralization error.

If any transistor of a given type is to be used with one fixed neutralization circuit, there is therefore a definite possibility that some transistors will not be perfectly neutralized. In these cases the amount of feedback capacity that is not cancelled out is called the residual capacity, C_r. In order to design a circuit that has fixed neutralization, and yet stabilize transistors with different collector-to-base capacitances, the residual capacity must be taken into account. Some gain reduction must be used to make C_r ineffective. The amount of gain reduction needed is dependent on C_r. The worst condition for stability is when C_r is a maximum. If enough gain reduction is used to stabilize the circuit under these circumstances, stable operation of all transistors of this type is assured.

CHAPTER 3

TRANSISTOR RADIO CIRCUITS

INTRODUCTION

In this chapter you will see how the transistor principles and circuit techniques discussed earlier are applied to complete transistor broadcast receivers. The important aspects of receiver operations are covered, and those problems unique to transistor circuits are emphasized. A number of different circuit arrangements for performing each circuit function required are shown, and the operating features of each circuit are explained.

1. RECEIVER FIRST STAGE

The first stage of a receiver can process the signal from the antenna in either of two ways: the signal may be amplified by an r-f amplifier; or the signal may be converted to an intermediate frequency and then amplified.

If the signal is converted first and then amplified, the signal-to-noise ratio at the output of the frequency converter is poorer than if the signal from the antenna is first amplified by an r-f stage and then converted. Furthermore, when a frequency converter is used in the first stage there is generally only one r-f tuned circuit, that of the antenna circuit. When an r-f stage precedes the frequency converter there are two r-f tuned circuits, the antenna tuned circuit and the interstage tuned circuit between the r-f stage and the frequency converter. The additional r-f selectivity that the r-f stage provides improves the image and i-f rejection of the receiver.

R-F Stage Requirements. There are four requirements an r-f stage must fulfill. These are: sufficient stage gain; low internal noise; sufficient selectivity; and linear characteristics throughout the entire range of the input circuit levels.

The transistor r-f stage is a tuned amplifier and is therefore bounded by certain stability conditions. The amplifier must be stabilized by neutralization or gain reduction. For operation at a fixed frequency, neutralization may be used. If the amplifier is to be tunable over a band of frequencies it is difficult to maintain accurate neutralization for all frequencies within the band. In this case stabilization is achieved by gain reduction. The amount of reduction required depends on the frequency. In order to insure stability over the entire frequency band the amplifier must be stabilized at the highest frequency in the band.

Noise. In order to provide the receiver with an optimum signal-to-noise ratio, the r-f stage must have low internal noise and efficient coupling to the antenna. Low internal noise is important because the signal-to-noise ratio at low signal levels depends upon how much noise is generated with the r-f stage. If the internal noise is low the signal-to-noise ratio is high. The internal noise of the r-f stage depends on the type of transistor that is used.

Following similar reasoning, if the signal is high, the signal-to-noise ratio is also high. It is important, therefore, to maintain efficient coupling to the antenna so that maximum signal transfer takes place. Optimum coupling between the antenna and the r-f stage occurs when the transistor is matched to the antenna circuit. This means that the connection to the antenna is so arranged that the dynamic resistance of the antenna circuit at this point equals the input resistance of the transistor. When this connection is made, maximum signal transfer is possible. Both capacitive and inductive coupling may be used, as shown in Fig. 1, to transfer the signal and obtain impedance matching. The desired impedance match is obtained by adjusting the relative values of C_1 and C_2. When transformer coupling is used, as in Fig. 1*b*, impedance matching is obtained

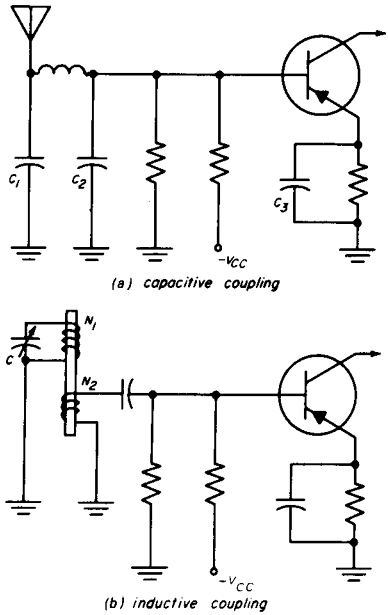


Fig. 1. Input circuits.

by adjusting the turns ratio N_1 to N_2.

When this procedure is followed the input of the transistor is matched and there is no reduction in gain at this point. However, to stabilize the transistor its gain must be reduced by a predetermined amount (the amount depends on whether or not the stage is neutralized). All of the gain reduction must be accomplished in the output circuit. (The input circuit must remain matched for best signal-to-noise ratio.) By properly tapping the primary of the interstage transformer, as shown in Fig. 2, the required amount of gain is obtained.

Linearity and Automatic Gain Control. To complete the discussion of the r-f stage, linearity of

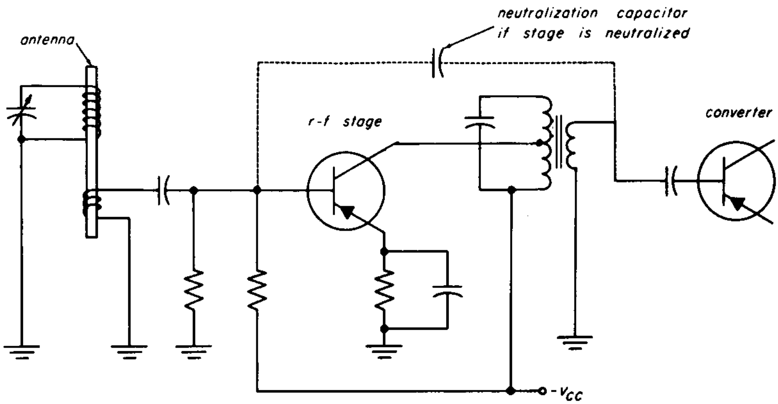


Fig. 2. Gain reduction by tapped tank circuit.

operation must be included. Linearity of operation is necessary for accurate reproduction of the modulation envelope. When considering linearity the effects of automatic gain control are important. Gain control may be applied to a transistor by means of *forward gain* control or *reverse gain* control. Forward gain control is the process of reducing the gain of a transistor by decreasing its collector voltage. This method of gain control is very seldom used for AM receivers because it requires high power from the automatic gain control circuit. Also, the detuning effect on the coupling transformers is objectionable.

Reverse gain control is the process of reducing the gain of a transistor by decreasing its collector current. This method is almost universally used in transistor AM receivers. However, reverse gain control is not without fault, for if the agc circuit was not properly designed, serious non-linearity occurs at certain signal levels. In order to reduce the collector current, the base to emitter voltage, V_{BE}, is decreased. At large signals where considerable agc is needed, V_{BE} approaches zero. At still stronger signals, V_{BE} goes through zero and the base becomes reverse biased.

Reverse gain control (reduction of the collector current) can be applied either by changing the base voltage as shown in Fig. 3, or by

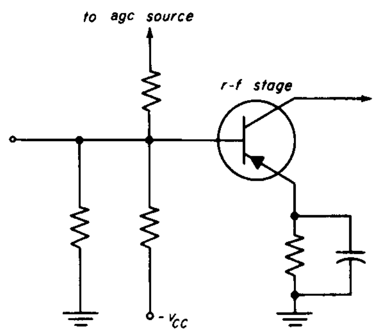


Fig. 3. Agc applied to the base.

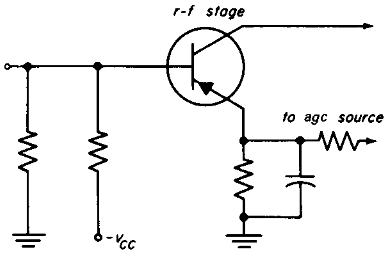


Fig. 4. Agc applied to the emitter.

changing the emitter voltage "emitter current control" as shown in Fig. 4.

The agc voltage swing required for either base-voltage control or emitter-current control is about the same. If the base voltage is controlled, the base current must be changed. Since the base current is very small, only a small amount of power is needed. If the emitter voltage is controlled, the collector current must be changed. Since the collector current is a relatively high current a large amount of power is needed. In most cases when a diode second detector is used, the base voltage is controlled. If a transistor second detector is used the emitter voltage is controlled.

Capacitance-Tuned R-F Stage. A capacitance-tuned r-f stage is one which uses a variable capacitor to accomplish the tuning. This type of circuit is used when the antenna is a fixed indicator. Figure 5 shows such a typical circuit using a ferrite-rod antenna. In the broadcast frequency range, 540 kc to 1600 kc, the usable stable gain is about 22 db.

Inductively Tuned R-F Stage. Figure 6 illustrates a typical r-f stage which is tuned with a variable inductor. If the same transistor, frequency, and bias conditions are used as in the capacity-tuned example, the amount of gain reduction needed for stability and the usable stable gain are the same.

2. OSCILLATORS

Most of the oscillator configurations that are used with vacuum tubes may also be employed with

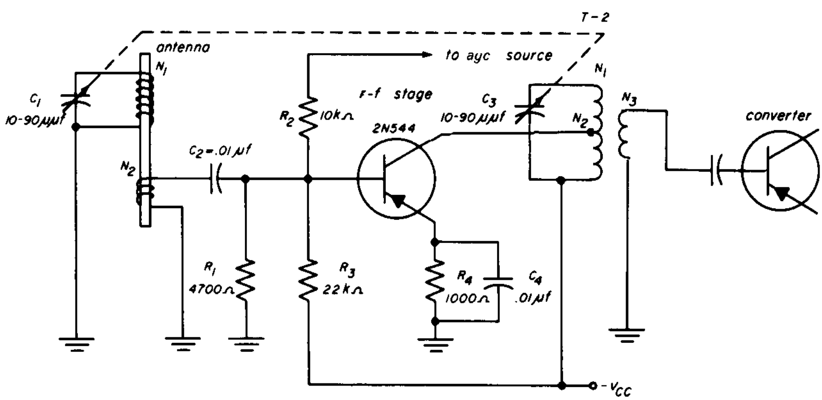


Fig. 5. Capacitively-tuned r-f stage.

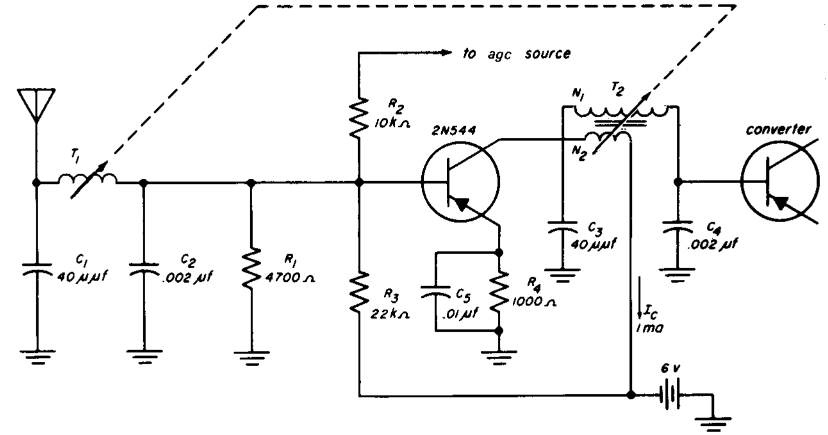


Fig. 6. Inductively-tuned r-f stage.

transistors. The oscillating conditions for high-frequency transistor circuits, however, frequently involve power transfer.

In tube amplifiers, fixed oscillator bias is rarely used. Grid current is drawn at the positive peaks of the signal swing which charge up an RC network in the grid circuit. The voltage across this RC network serves as the signal-developed bias for the oscillator. Thus, there is no bias when oscillations are just starting, and the bias builds up as the oscillations build up. This mode of operation is desirable for tube oscillators, because the gain of the stage is maximum at zero bias, making the oscillator starting easy. On the other hand, a transistor with zero forward bias has minimum gain. Thus, transistor oscillators work best when a d-c forward bias is supplied. Some signal-developed bias, produced by charging an RC filter, is usually also used in addition to the d-c forward bias, to stabilize the amplitude of the oscillations.

Figure 7 shows a tuned-collector oscillator employing tickler feedback to the base. The biasing arrangement is similar to that employed in a common-emitter amplifier. Constant emitter-current biasing is desirable in the oscillator, just as for the amplifier, since it stabilizes the transistor operating point

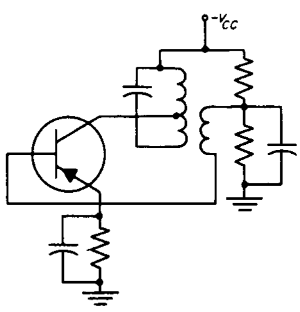


Fig. 7. Tuned-collector oscillator.

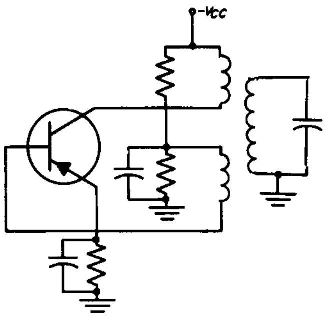


Fig. 8. Oscillator with separate tank circuit.

for changes in temperature and supply voltage, and allows for interchangeability of oscillator transistors. Oscillator operation is similar to that in an amplifier in which the output signal is developed across a tuned circuit having a reflected load from a secondary winding. In this case, however, the reflected load is the input impedance of the oscillator transistor. A portion of the output signal is thereby fed to the base of the transistor which in turn amplifies the signal to a level sufficient to sustain oscillations.

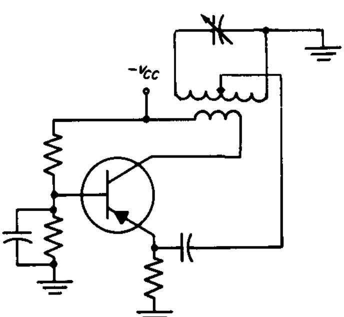


Fig. 9. Oscillator with feedback to the emitter.

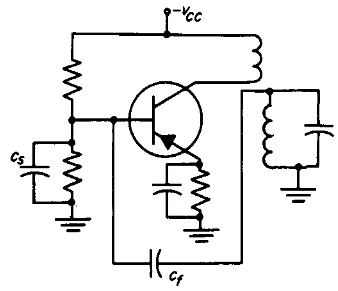


Fig. 10. Capacitive-feedback circuit.

The oscillator shown in Fig. 8 is nearly equivalent to the previous circuit. Here the collector is inductively coupled to the tank circuit via a step-down winding, feedback again being fed to the base. A circuit employing feedback to the emitter is shown in Fig. 9. In this circuit the emitter bypass of the previous circuit is used as the feedback coupling capacitor. This circuit is shown with a variable tuning capacitor since it is a type which is frequently employed in broadcast receivers. Another circuit employing capacitive feedback is shown in Fig. 10. The magnitude of feedback is conveniently controlled in this circuit by the ratio of C_f to C_S.

A very large number of other oscillator circuit configurations are possible. The foregoing illustrations are merely representative. In all of the circuits the phasing of the windings should be such as to cause positive feedback. That is, the feedback signal is in phase with an applied signal with the feedback disconnected.

3. FREQUENCY-CONVERSION CIRCUITS

A frequency-conversion circuit may occupy either of two positions in a receiver. It may be the stage immediately following the antenna, or the stage immediately following the r-f amplifier. In either case, the frequency-conversion circuit may be either a mixer with separate oscillator, or a self-oscillating mixer (often referred to as an autodyne converter).

Two conditions are necessary to achieve frequency conversion: a local oscillator signal must be present, and a non-linear resistance must be present. The non-linear resistance is present because of the input characteristics of a transistor. The local oscillator power can be obtained from either of two sources: a separate transistor can be used as the oscillator, or the transistor which is used for the mixer can serve a dual purpose and act as its own oscillator. When the first arrangement is used, the result is a mixer-oscillator combination. The second arrangement results in a converter.

The advantages of a mixer-oscillator combination over a self-oscillating mixer is the independence of the local oscillator from the mixer circuit. This enables agc to be applied to the mixer stage without any effect on the local oscillator. In the case of the self-oscillating mixer this is not possible. If agc is applied to this circuit, the oscillator is affected in a manner which causes it to change frequency or even to stop oscillating. The advantage that the self-oscillating mixer offers is that it uses only one transistor to perform the function of frequency conversion. The self-oscillating mixer (autodyne converter) is very commonly used in transistor radio receivers.

Although the oscillator-mixer combination and the self-oscillating mixer have different local oscillator sources, their basic mixing action is similar and the same analysis can be applied to both.

Conversion Gain. The conversion gain of a transistor frequency converter or mixer depends upon:

1. The r-f input circuit.

2. The i-f output circuit.

3. The oscillator-injection voltage (oscillator voltage between the base and emitter).

The gain and stability conditions differ somewhat from those of a straight transistor amplifier. In an amplifier, feedback from the output circuit can appear across the input circuit and cause instability. In the case of the frequency converter the output is tuned to the i-f frequency while the input is tuned to the r-f frequency. Any feedback from the output circuit is at the i-f frequency. If this feedback at the i-f frequency appears across the input which is tuned to the r-f frequency, it is greatly attenuated. Thus, the frequency converter can operate at high gain without the stability problems of an amplifier. It is general practice to match the input of the frequency converter to the r-f source (this can be an antenna or an r-f stage) and the output to the i-f load. The proper output connections are covered in the i-f amplifier section. At optimum oscillator injection the conversion gain of a frequency-converter stage is the maximum available gain of the transistor of the i.f. less approximately 10 db for conversion loss.

Oscillator Injection. Optimum conversion gain is obtained when the combination of d-c bias and oscillator injection voltage places the modulated r-f frequency signal at the steepest (highest g_m) part of the input characteristic. The optimum injection voltage for germanium transistors falls within the range of 0.08-volt rms to 0.13-volt rms. A curve showing the effects of oscillator injection on gain is given in Fig. 11. If the injection voltage is lower than this range, the gain decreases rapidly. Above this value the gain decreases slowly, usually because of involved frequency effects relating to the input and output impedance. It is, therefore, desirable to operate the mixer with somewhat more than optimum injection so that small changes in injection do not lead to large changes in gain. The output load may be made as large as necessary for efficient transfer of power. Stability is not a problem in mixers since the signal developed at the output is at a different frequency than the input.

Mixers. Mixers are usually operated with values of emitter resistance comparable to that employed with amplifiers. This acts in the usual fashion to stabilize the transistor operating point at the average value of current as determined by the oscillator injection. Occasionally some small forward bias is also used so that the bias current does not depend entirely on the oscillator injection. Figure 12 shows

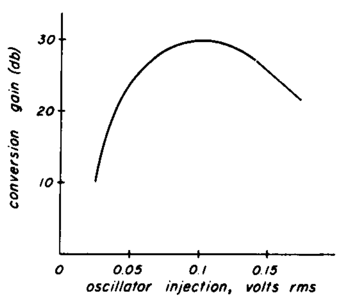


Fig. 11. Conversion gain vs. oscillator injection voltage.

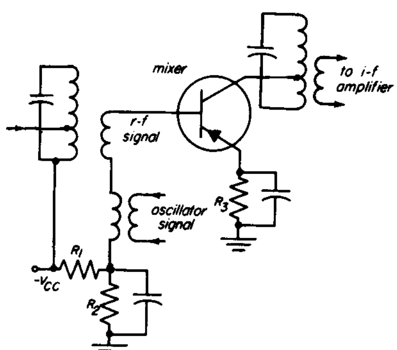


Fig. 12. Typical mixer using base injection.

a practical transistor mixer circuit where forward-bias voltage is applied to the base via resistors R_1 and R_2. The combination of this bias voltage, the oscillator voltage (fed via a transformer winding in series with the signal winding) to the base of the transistor, and the emitter resistor, R_3, determines the average mixer current. This current is typically in the 0.5 to 2 milliampere range.

For good mixer operation, the oscillator secondary winding should present a low impedance to both the r-f and i-f signal frequencies. The impedance seen by the r-f secondary winding is then determined primarily by the average mixer current and secondarily by the peak mixer current. Typically, this value is about twice the impedance of the same transistor operated as an amplifier at the r-f frequency. The r-f stage output impedance is then "matched" to the mixer input impedance as previously discussed. The mixer output impedance is also about twice the conventional i-f amplifier-output impedance, since both impedances are increased by the cutoff portion of the oscillator signal.

The mixer shown has both r-f and oscillator signals applied to the base. Either one or both of these signals may be applied to the emitter. A circuit employing oscillator feed to the emitter is shown in Fig. 13. Again the oscillator winding should have low impedance to both the r-f and i-f signals. Generally,

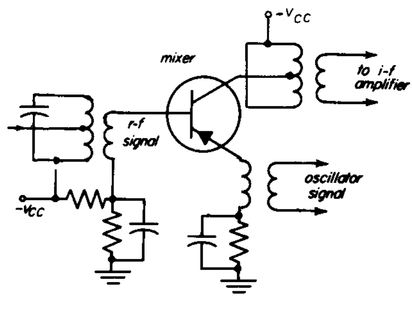


Fig. 13. Typical mixer with emitter injection.

the r-f signal is fed to the base since this connection has higher gain than the emitter input connection. The terminal for oscillator feed makes no difference in available gain, provided the impedance conditions are met.

A complete schematic diagram of a mixer employing capacitively-coupled oscillator injection to its emitter is shown in Fig. 14. The injection-voltage takeoff winding (N_3) has comparatively low impedance at the signal frequency, so capacitor C_2 also acts as an emitter bypass for the mixer at the r-f and i-f frequencies. With an average transistor, such as a 2N412, a conversion gain of 30 db can be expected at broadcast-band frequencies.

Converters. A converter combines the mixer and oscillator functions in a single transistor stage. A converter circuit with the oscillator portion employing tickler feedback to the base is shown in

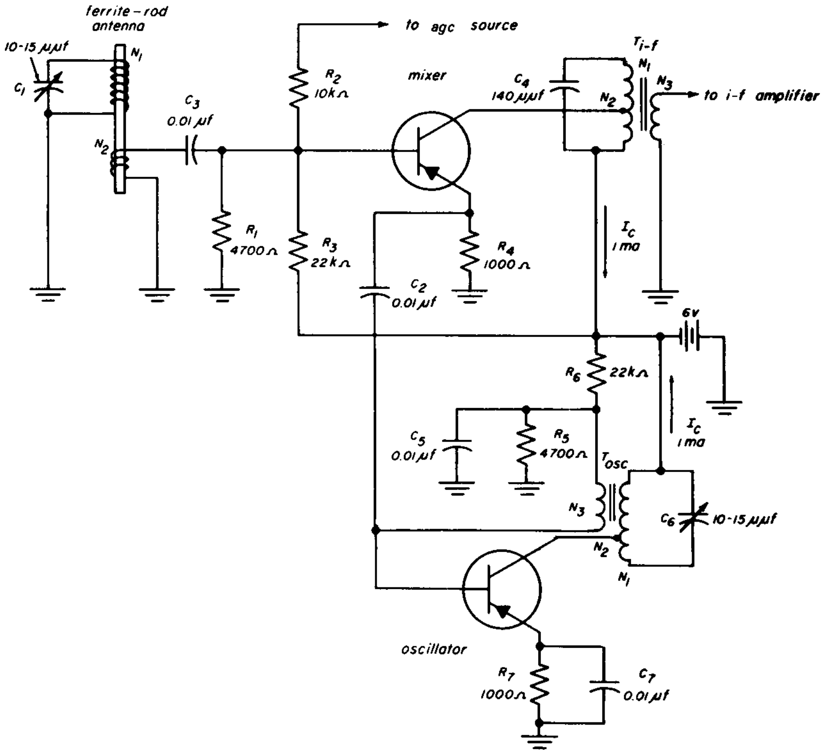


Fig. 14. Typical mixer with capacitively-coupled oscillator-voltage injection to its emitter.

Fig. 15. Here the oscillator and r-f feed to the base are in series, while the oscillator and i-f tuned circuits are connected in series in the collector path. The collector of the transistor is connected to the i-f transformer, $T_{i\text{-}f}$, which in turn is connected to the oscillator transformer, T_{osc}. The impedance of the i-f circuit at oscillator frequencies is negligible, and similarly the impedance of the oscillator circuit is negligible at the i-f frequency. Oscillation is sustained by connecting the base to the oscillator transformer, T_{osc}, through a secondary winding, N_3. Both N_2 and N_3 of the oscillator transformer T_{osc} are adjusted to provide the proper injection voltage with minimum variation under various operating conditions. The r-f signal is applied to the base of the transistor from the secondary of the r-f transformer. Notice that the r-f and oscillator secondary windings are in series with each other. The r-f circuit presents a negligible impedance to the oscillator frequency, and the

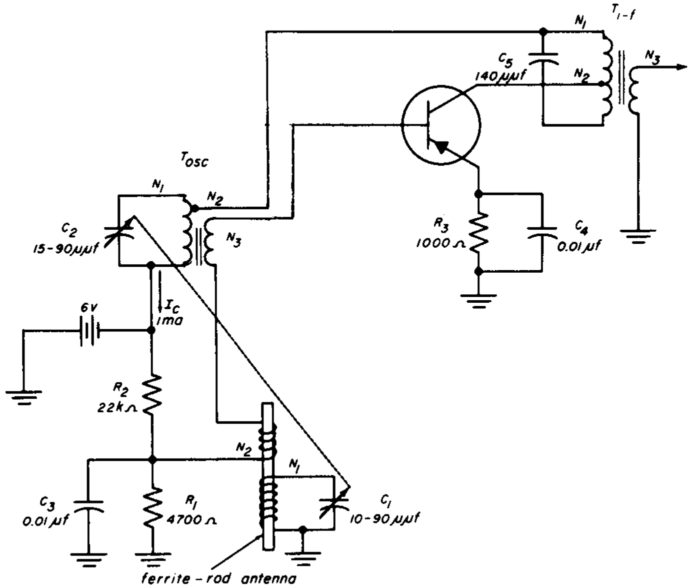


Fig. 15. Typical converter with base injection.

oscillator circuit presents a negligible impedance to the r-f frequency.

The requirements for this type circuit are much the same as the individual requirements of the oscillator and the mixer. A very important additional requirement that is imposed on the circuit concerns the amplitude of the oscillator voltage excursion at the collector. (In the mixer, the oscillator signal is not developed at the collector but in the converter an oscillator signal at the collector is necessary to sustain oscillations.) For the converter, the oscillator voltage swing must not be large enough to cause the collector to run out of voltage (frequently termed bottoming). When bottoming occurs the conversion gain of the converter becomes very low. This is easily overcome by reducing the impedance of the oscillator tuned circuit at the collector. The impedance is reduced as usual by tapping down on the tuned circuit, which effectively reduces the L/C ratio at the collector.

The mixing action of the converter is identical to that taking place in the mixer. The oscillator voltage swing at the input (base or emitter) swings the transistor from a cutoff condition to a region of

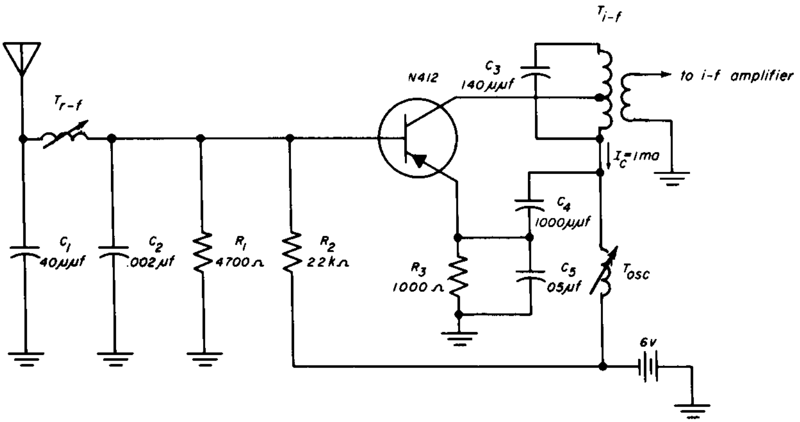


Fig. 16. Emitter injection of the oscillator-feedback voltage in a converter.

high transconductance. Of course in the converter forward bias is always used so that the oscillator functions properly. The range of oscillator signal at the input for the converter circuit is then from 0.05 to 0.2 volt, being somewhat less than that required for a non-forward-biased mixer.

An example of emitter injection of the oscillator-feedback voltage is shown in Fig. 16, where tuning is accomplished by means of variable inductors. In this circuit, the oscillator feedback is achieved by using a capacitance divider (C_4 and C_5), and injecting the oscillator signal into the emitter. Proper oscillator injection is obtained by the correct choice of capacitor, C_5. At oscillator frequencies the base is bypassed through the capacitor, C_2. At the r-f the emitter is bypassed through the capacitor C_5. The gain of this circuit is also similar to the oscillator-mixer combination.

An alternative circuit, where oscillator feedback is applied to the emitter from a tap on the oscillator coil is shown in Fig. 17. This circuit is similar to that employed in

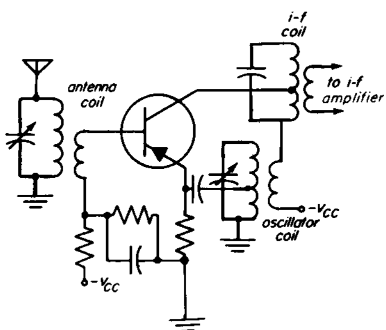


Fig. 17. Converter with emitter injection from a tapped coil.

many RCA broadcast receivers. The loop antenna is coupled to the converter base via an antenna secondary winding. The oscillator tuned circuit is coupled to the collector from a secondary winding and is capacitively coupled to the emitter. The i-f signal is coupled from the collector of the converter to the i-f amplifier with the usual tuned circuit.

4. INTERMEDIATE-FREQUENCY AMPLIFIERS

The major function of an i-f amplifier is to provide selective gain. In fulfilling this function the amplifier must maintain linear operation throughout the entire range of signal levels. The number of stages constituting the i-f amplifier depends on the particular receiver application. In most cases a two-stage amplifier provides sufficient gain. However, there are applications where the gain of a one-stage amplifier suffices. In most cases enough gain is available in a two-stage amplifier so that it may be stabilized by gain reduction, and neutralization is not required.

Selectivity. There are two aspects to consider when determining the selectivity of an i-f amplifier. They are:

1. The bandwidth required to pass the modulation frequencies, and

2. The amount of rejection required to suppress adjacent channel stations. Most radio receivers are designed to receive speech and music information. The frequency range of this information extends to about 10 kc. If the receiver is designed for high-quality reproduction, the bandwidth must be wide enough to pass modulation frequencies up to 10 kc. Therefore, the i-f selectivity cannot attenuate signals in a frequency band ±10 kc from the intermediate frequency. If this degree of quality is not necessary (as in the case of most portable receivers) the bandwidth can be made much narrower. The bandwidth found in most home or portable receivers is about ±3 kc. This is sufficient for most applications.

Conflicting somewhat with the bandwidth requirements is the necessary selectivity for adjacent-channel rejection. If single-tuned circuits are used, there is a definite relationship between the flat portion of the response and the skirt selectivity. The flat portion determines the bandwidth, while the skirt selectivity determines the adjacent channel attenuation. Figure 18 shows the selectivity that can be expected from

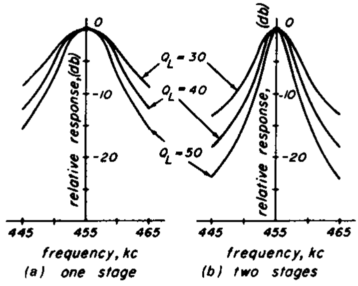


Fig. 18. Response of single-tuned circuits.

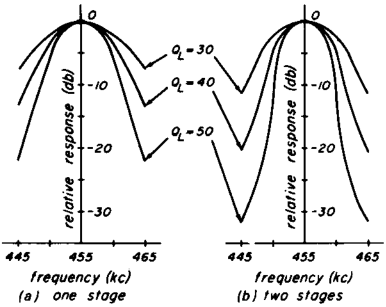


Fig. 19. Response of double-tuned circuits.

a one- or two-stage i-f amplifier using single-tuned circuits at 455 kc for different values of loaded Q. If double-tuned circuits are used, the relationship between the flat portion of the response and the skirt selectivity depends upon the coupling factor. With this type of circuit, the skirt selectivity can be improved while still maintaining adequate bandwidth. The coupling factor that provides optimum response characteristics is called critical coupling. In practice the coupling factor is always slightly less than critical. This insures that for normal production variances, critical coupling is not exceeded. Figure 19 shows the selectivity that can be expected from a one- or two-stage i-f amplifier using double-tuned circuits at 455 kc. A typical i-f amplifier, using 2 stages and single-tuned circuits is shown in Fig. 20.

Linearity. It is essential for the i-f amplifier to maintain linear operation throughout the range of signal levels that it will encounter. Any non-linearity distorts the modulation information and renders the receiver useless. There are two ways in which non-linearity may occur in the i-f amplifier. First, the signal level may exceed the power handling capabilities of the amplifier. Secondly, if agc is applied to the amplifier, the same type of overload distortion may occur as was described in the r-f stage section.

5. DETECTORS

From the antenna to the detector, the signal is in the form of an

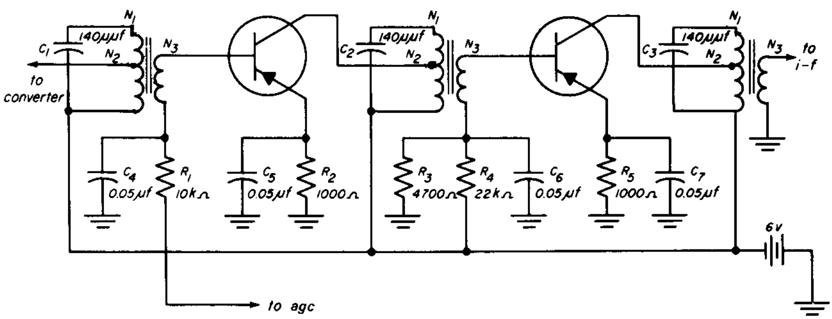


Fig. 20. Typical 2-stage i-f amplifier.

amplitude-modulated carrier. It is the detector's function to separate the modulation from the r-f carrier. The information contained in the modulation is fed into an audio amplifier and amplified to a power level sufficient to operate an electromechanical device such as a loudspeaker. The second function of a detector is to provide a source of agc.

Diode Detector. The operation of a diode detector, shown in Fig. 21, is the same as any rectifier. The i-f signal is rectified and the capacitor C_1 charges up to a voltage V_1 that is proportional to the amplitude of the i-f signal and equal to its peak value. This d-c voltage can be used for agc. With an amplitude modulated signal applied to the detector, the voltage V_1 varies in proportion to the modulation. In this case the modulation is the audio signal, and hence, audio voltage also appears across the load resistor R_1. In most cases the input resistance of the audio amplifier is approximately 1000 ohms, and the value of R_1 about the same. For efficient rectification capacitor C_1 must be large enough to present a low impedance to the i-f signal. The upper limit for the value of C_1 is limited by the amount of shunting it presents to the audio frequencies. A typical value for C_1 is 0.05 μf. The d-c component of the voltage across R_1 may be used for agc. For this function the audio components of the signal are filtered out by the low-pass filter, R_2 and C_2. The agc voltage is applied to the base of the r-f stage, the i-f stage, or both. Depending on the input signal level, the power loss of this detector varies from 25 to 35 db. The diode detector is very widely used and provides economy of parts with linear operation over a wide dynamic range.

Transistor Detector. Where additional agc power is required, or where more gain following detection is needed a transistor detector is used. Figure 22 shows a typical transistor detector. Rectification takes place in the input circuit and the current gain of the transistor

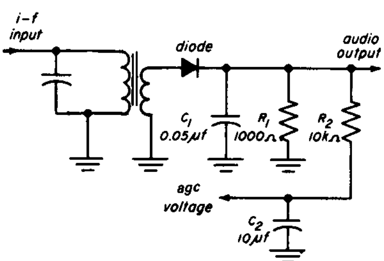


Fig. 21. Diode detector.

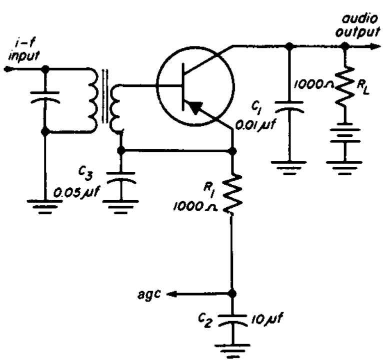


Fig. 22. Transistor detector.

amplifies the d-c and audio components of the rectified signal. The rectification process is the same as for the diode detector. The associated limitations on low-level operation also hold true. The amplification process of the transistor detector is the same as in any amplifier, and therefore the amplified signal is limited by the collector current and the available supply voltage.

The net power gain of a transistor detector is a combination of the power loss of the rectification process and the power gain of the amplification process. The combination of the two results in approximately 5 db of gain. In the circuit of Fig. 22, the emitter of the detector transistor is connected through the low-pass filter R_1 and C_2 to the r-f stage, the i-f stage, or both. With this connection the collector current of the detector affects the collector current of these stages, and provides automatic gain control. The audio output is taken across R_L. The function of C_1 is to bypass the i-f to prevent it from affecting the following audio stages.

Typical Detector Circuits. There are many types of detector circuits which can be used. These circuits are all similar in that their operation is basically that of the diode detector, or the transistor detector. The differences arise because the particular receiver system in which the detector circuit is used has its own specific requirements.

6. AUDIO AMPLIFIERS

The function of the audio amplifier is simply to amplify the audio signals. The important considerations of this amplifier are its power-output capabilities, power gain, frequency response, linearity, and efficiency.

Power-Output Stage. There are several types of power-output stages that are commonly found in transistor receivers. They are:

1. Transformer-coupled class-B amplifiers.

2. Class-B amplifier using a tapped speaker.

3. "Stacked" output stages using class-B amplifiers.

4. Class-A amplifier.

The efficiency of class-B circuits is relatively good for high output levels and also for average listening levels. Class-B operation requires the use of two transistors, and is used for most portable applications.

Transformer-Coupled Class-B Amplifier. A transistor operating in a class-B mode draws current for only one half the total cycle. The average current depends on the signal level; when the signal is small the current is small. This results in the high efficiency which is the major advantage of class-B operation. Figure 23 shows the general principles involved, and the complete

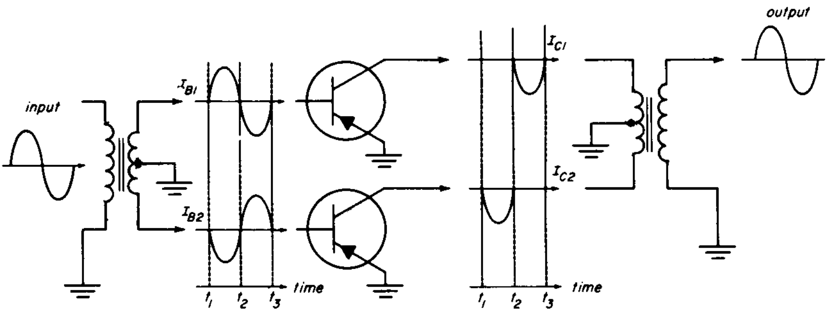


Fig. 23. Operation of a class-B amplifier.

circuit is given in Fig. 24. An input transformer is used to obtain two out of phase currents, I_{B1} and I_{B2}. Since both transistors are turned off for one polarity and on for another polarity, the two transistors are alternately turned on and off. The input currents I_{B1} and I_{B2} are then amplified by the current gain of the transistor and the output currents are combined in the output transformer to give the final output waveform. In order to use a particular type of transistor in a class-B circuit, the rated power dissipation of the transistor must be at least 27% of the maximum power output.

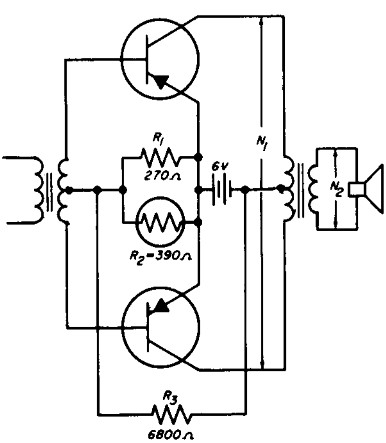


Fig. 24. Typical push-pull class-B amplifier.

Ideally a class-B amplifier has no d-c bias. However, in practice the effect of the non-linear portion of the transfer characteristics produces distortion similar to that shown in Fig. 25. In order to avoid this crossover distortion, the transistors are biased so that they operate in the linear region of their characteristics. The bias is applied from the voltage divider consisting of R_1, R_2, and R_3. R_2 is a temperature-sensitive resistor which maintains the proper bias point over the desired temperature range.

A further reduction of crossover distortion is possible by decreasing

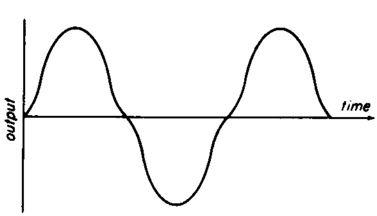


Fig. 25. Crossover distortion in a class-B amplifier.

the gain of the amplifier at high frequencies. By doing this, the higher harmonics that cause crossover distortion are eliminated. The reduction can be accomplished in either of two ways:

1. By shunting the output transformer with a capacitor of about 0.1 μf, or

2. By connecting a capacitor (about 0.02 μf) between the collector and base of each transistor.

Class-B Amplifier Using a Tapped Speaker. Another configuration which may be used with class-B operation involves the use of a tapped speaker. The tap is arranged so that the impedance of the speaker is divided into two equal parts. Fig. 26 shows this circuit. Essentially the operation of this amplifier is the same as in the transformer-coupled circuit. The difference is that the transformer is eliminated, and the proper phasing of the output current is achieved by using the tapped speaker.

Stacked Output Stage. One of the ways of eliminating the output transformer in a class-B circuit is to use a "stacked" arrangement as in Fig. 27. One side of the speaker is connected to a point common to the positive terminal of one battery, and the negative terminal of another battery. The other side of the speaker is connected to a point common to the emitter of one transistor and the collector of the other. The secondary of the input transformer is divided into two windings. One winding is connected between the base and emitter of the first transistor and the other winding is connected betwen the base

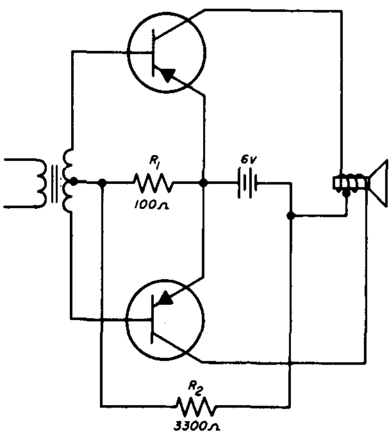


Fig. 26. Tapped-speaker output stage.

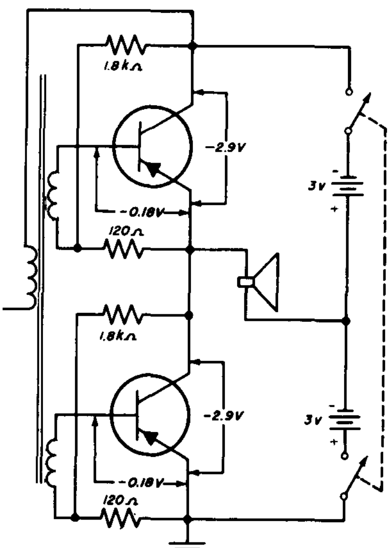
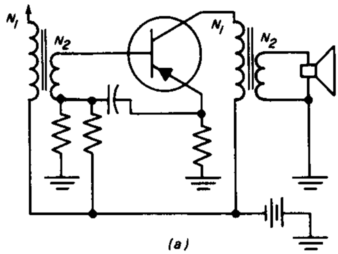
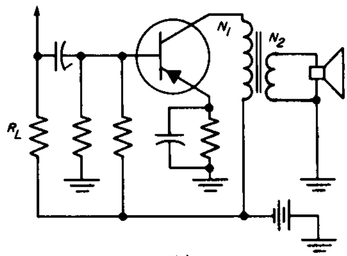


Fig. 27. Stacked output stage.



(a)



(b)

Fig. 28. Class-A output stages.

and emitter of the second. The two windings are phased so that the two transistors conduct alternately on every other half cycle. When one transistor conducts, the output current flows through the speaker in one direction. When the other transistor conducts the output flows through the speaker in the other direction. Thus, the proper output wave form is reproduced and class-B operation is achieved.

Class-A Amplifier. Figure 28 shows typical class-A power-output stages. Class-A operation requires that current flows through the transistor during the entire cycle. When a class-A output stage is used, the driver can be transformer coupled as in Fig. 28*a* or capacity coupled as in Fig. 28*b*. If transformer coupling is used, the design of the driver stage is the same as with a class-B output stage. If capacity coupling is used, the driver stage has less gain. In Fig. 28*b* the value of R_L is approximately equal to the input impedance of the output stage. The collector current of the driver is again adjusted to provide optimum gain and output capabilities.

Typical Audio Amplifier. Figure 29 shows a typical audio frequency-amplifier section of a portable transistorized receiver. The output of

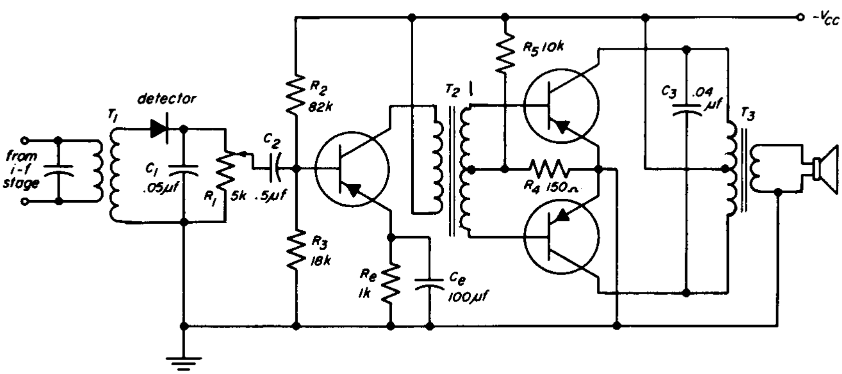


Fig. 29. Typical audio section.

the detector is fed to the input stage which is operated as a class-A amplifier. Because the output from the detector contains both r-f and a-f signals, a low-pass filter consisting of the parallel combination of R_1 and C_1 is used. Otherwise, amplified i-f signals may be fed back to the i-f circuit and cause oscillation. The desired a-f signal is developed across R_1, and serves as the volume control. The output of the class-A audio stage is transformer coupled to the input of the class-B push-pull output stage. The class-A stage is generally referred to as the driver stage and is designed to have sufficient signal handling capabilities to deliver the power input needed by the output stage. The output stage of most portable receivers is usually operated as a class-B amplifier because of the low standby power consumed at zero signals. This mode of operation also provides good operating circuit efficiency, and is discussed in detail in the section on large signal amplifiers.

Variable Gain or Volume Adjustment. In most audio systems, there is a provision for manual adjustment of the gain of the amplifier. When a preamplifier is used, this adjustment is usually incorporated in the preamplifier circuit. The gain variation is obtained by using a variable resistance which can be arranged to attenuate the signal level at a particular point in the amplifier. The gain control, or volume control, adjusts the signal level without causing any change in the overall frequency response of the amplifier.

The volume control can be arranged either as a *voltage* or *current divider*. When operated as a voltage divider, the volume control usually operates into a high resistance load with the signal voltage being supplied by a relatively low resistance voltage source. This is the case in vacuum-tube circuits. As a current divider, the volume control works into a low load resistance (the transistor input resistance) and is usually driven from a high resistance current source (output resistance of the preceding transistor). The volume control employed in most transistor applications is placed between a high resistance source and a low load resistance, as shown in Fig. 30, by potentiometer R. For this reason, it acts as a current divider.

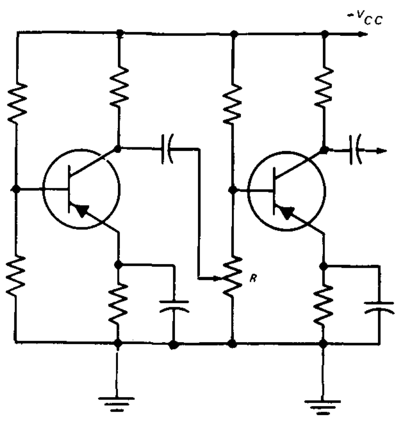


Fig. 30. Current-divider volume control.

7. OTHER RECEIVER CONSIDERATIONS

Overload Diode. When the first stage of a receiver is a self-oscillating mixer, automatic gain control is only applied to the i-f amplifier. However, as previously discussed, the amount of gain control that is available from the i-f amplifier is in the order of 20 to 30 db. In most cases this is insufficient, and if gain control is not provided between the antenna and i-f amplifier, the receiver will overload at strong signals. An effective way to provide this extra gain control is by use of an overload diode in the manner shown in Fig. 31. The diode CR_1 is connected across the first i-f transformer. The collector current of the first i-f amplifier flows through R_2. At low signal levels, the voltage drop in R_2 is high enough so that the diode is reverse biased. When the diode is reverse biased it has a high impedance, and therefore does not affect the signal. As the input signal is increased the agc reduces the current in the first i-f amplifier. This reduction in collector current causes a decrease in V_2. As V_2 is reduced the diode is biased toward the forward direction. In this region the diode impedance is relatively low and the i-f transformer is loaded more heavily. The shunting effect of the diode reduces the power delivered to the i-f amplifier, and the receiver now operates at higher signals without overloading.

Reflex Amplifiers. The term reflex amplifier refers to an amplifier which simultaneously amplifies signals at two frequencies. When used in a radio receiver, the most common arrangement is that shown in Fig. 32. Here the second i-f stage also acts as an audio amplifier. This circuit eliminates the need

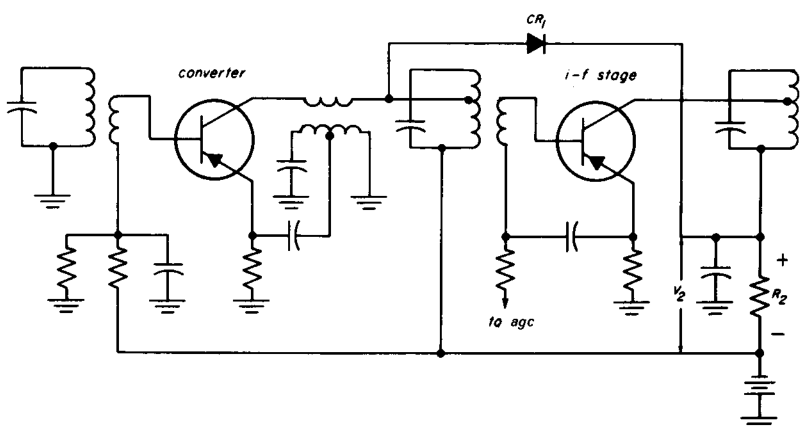
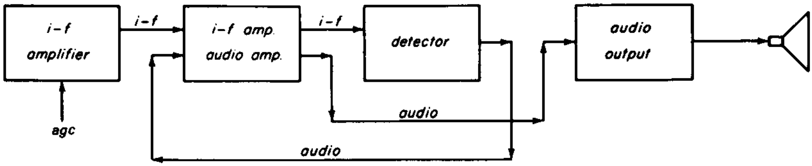
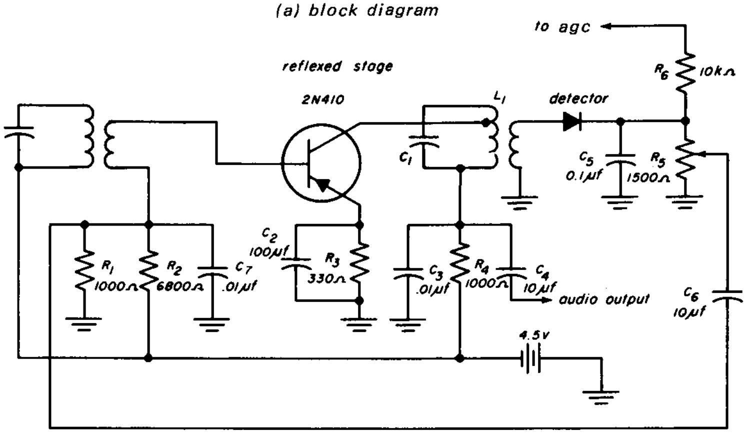


Fig. 31. Overload-diode circuit.



(a) block diagram



(b) circuit

Fig. 32. Reflexed stage.

for a separate audio driver transistor. Performance is relatively poor, and this circuit is no longer used now that the cost of transistors is quite low.

A typical circuit arrangement is shown in *b* of the figure. The tank circuit L_1C_1 is the amplifier load at i-f frequencies, and C_3 shorts out R_4 at these frequencies. For audio frequencies, the tank L_1C_1 is shorted, and R_4 serves as the load resistor for the audio signal.

Power Supply and Filtering. In most receivers the internal resistance of the power supply is large enough to require some isolation between the receiver and the power supply. If this isolation is not present, motorboating or other forms of instability may result. The most common way to accomplish this isolation is to insert RC filters between the receiver and the power supply.

From these filters power is supplied to the various stages in a receiver. The circuit used to supply this power is between a ground connection and a "hot" line or power lead. Normally the ground connection is the chassis or some other prominent part of the receiver such as the tuning capacitor, or the speaker brackets. The other side of the supply circuit is connected to

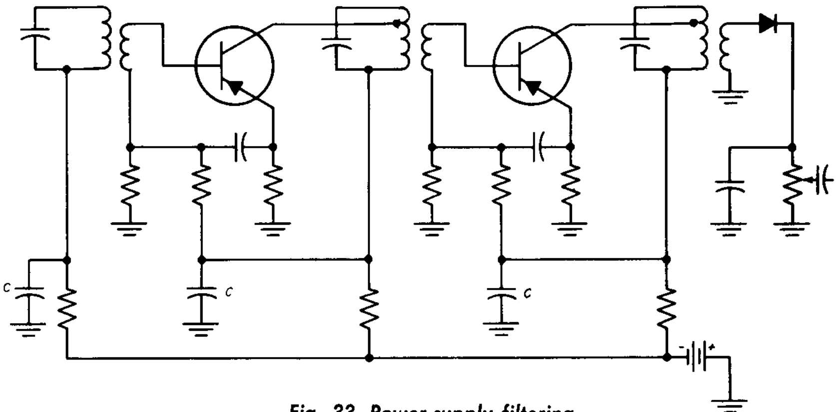


Fig. 33. Power-supply filtering.

the different stages by a single wire, or if printed circuits are used, by a single copper ribbon. The various stages are connected to this lead at different points along its length. This supply lead also acts as an a-c return path to ground. In most cases it is of such a length that the inductance of the wire is an appreciable impedance at r.f. and i.f. Since the high frequency current of the r-f, converter, and i-f stages returns to ground through this common impedance, there is a possibility of positive feedback and therefore overall instability. To avoid this condition isolation in the form of filters must be inserted in the supply lead so that each stage can have an independent a-c return path to ground.

When PNP transistors are used there are two ways in which the necessary filtering may be provided. The positive terminal of the power supply is connected to ground and RC filters are put in the supply lead, as shown in Fig. 33. Capacitor *C* shunts the output signal around the battery.

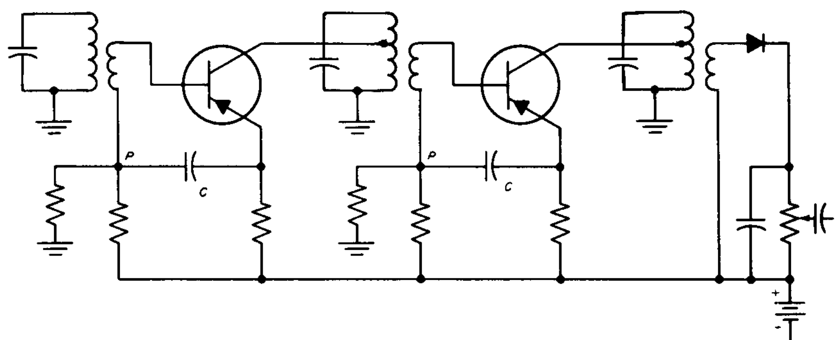


Fig. 34. Alternate bypassing method.

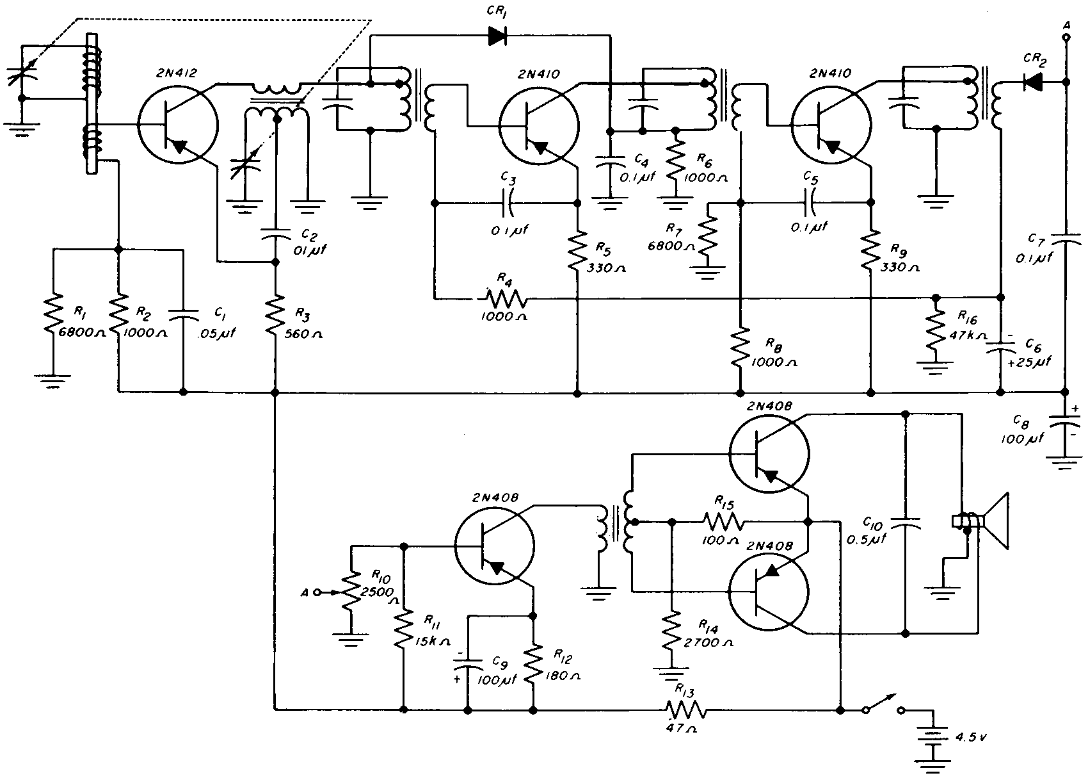


Fig. 35. Typical receiver.

Alternately, the negative side of the power supply is connected to ground and the emitter resistors and bypass capacitors utilized as filters, as in Fig. 34. Capacitor *C* puts an a-c short between point *P* and the emitter, so no signal on the supply line can enter the input circuits. The second of these two methods is more desirable because filtering is obtained without extra components.

8. COMPLETE RECEIVERS

Figure 35 shows the complete circuit of a portable receiver. This receiver is designed to have sufficient sound for outdoor use, long battery life, high gain for good sensitivity, and a minimum number of parts for reliability and low cost.

To provide both high sound output and long battery life, a class-B output stage is used. A tapped speaker is used in the output circuit and thus eliminates the need for an output transformer and a temperature sensitive bias network. The maximum power output of this particular amplifier is 150 mw.

The audio driver is transformer coupled to the output stage. Base bias is derived from the network consisting of R_{10} and R_{11}. R_{10} also serves as a volume control. The overall gain of the audio amplifier including the driver and output stages is 60 db.

A diode is used as both a detector and agc source. The diode is connected so that the position of the volume control determines the d-c operating point of the first i-f stage as well as the amount of audio power delivered to the audio amplifier. At low signal levels the bias of the first i-f stage depends on the voltage at point *A*. At strong signals the bias depends upon the voltage at point *A* and the rectified voltage across the diode. Thus the gain of the i.f. depends upon the signal level and the volume control position. The advantages of this volume control arrangement are:

1. Low minimum volume can be attained without a critical minimum-resistance specification for the volume control.

2. The elimination of an audio coupling capacitor between the detector and audio circuits.

The i-f amplifier consists of two stages. Signal is applied to the input of the transistors by connecting the secondaries of the transformers directly to the bases and through capacitors C_3 and C_5 to the emitters. So far as the signal is concerned, this input connection puts the emitter resistors in the ouput circuit. In most cases the emitter resistors are small when compared with the output load, and can be left unbypassed without any noticeable loss in gain. For any given collector current, the value of the emitter resistors (R_5 and R_9) is a compromise between temperature stability and

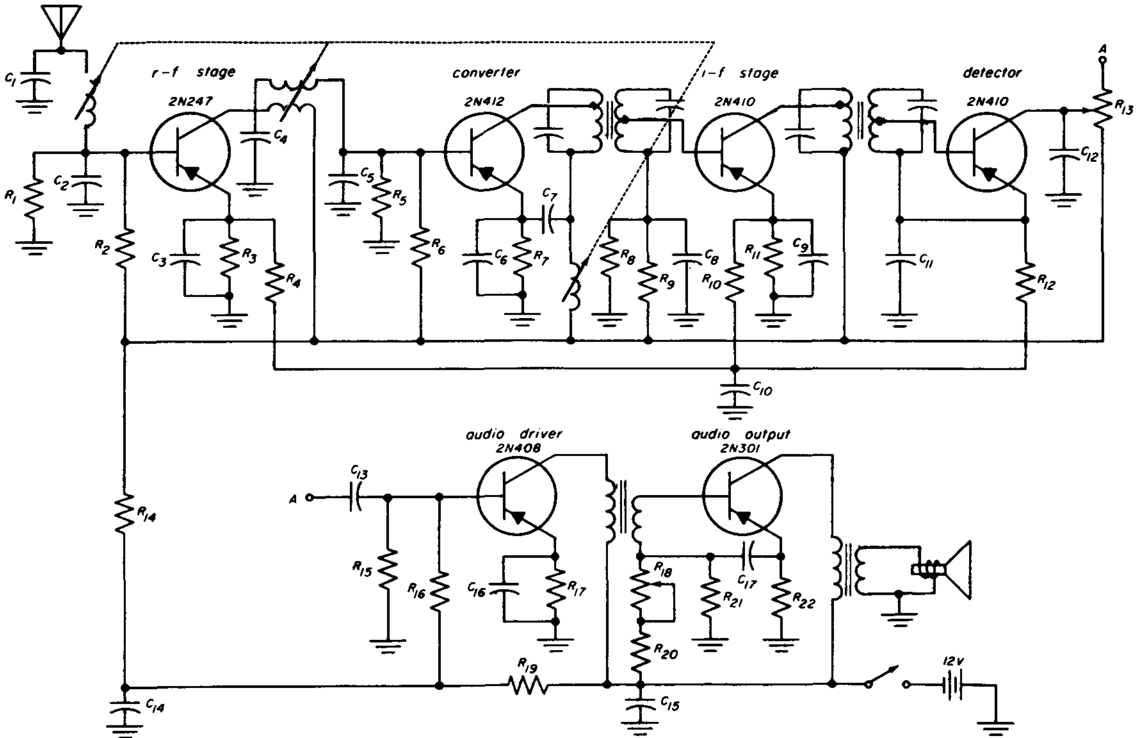


Fig. 36. Typical receiver with r-f stage.

R_1 = 4700 Ω
R_2 = 33-k Ω
R_3 = 4700 Ω
R_4 = 470 Ω
R_5 = 4700 Ω
R_6 = 33-k Ω
R_7 = 4700 Ω
R_8 = 4700 Ω
R_9 = 33-k Ω
R_{10} = 1500 Ω
R_{11} = 1200 Ω
R_{12} = 220 Ω
R_{13} = 1000 Ω
R_{14} = 47 Ω
R_{15} = 5600 Ω
R_{16} = 33-k Ω
R_{17} = 680 Ω
R_{18} = 400 Ω
R_{19} = 47 Ω
R_{20} = 100 Ω
R_{21} = 27 Ω
R_{22} = 1 Ω
C_1 = 40 μμf
C_2 = 0.002 μf
C_3 = 0.01 μf
C_4 = 40 μμf
C_5 = 0.002 μf
C_6 = 0.005 μf
C_7 = 1000 μμf
C_8 = 0.1 μf
C_9 = 0.1 μf
C_{10} = 100 μf
C_{11} = 100 μf
C_{12} = 0.05 μf
C_{13} = 25 μf
C_{14} = 100 μf
C_{15} = 300 μf
C_{16} = 100 μf
C_{17} = 300 μf

interchangeability on the one hand, and loss of collector voltage and gain on the other hand. The total gain of the i-f amplifier is 50 db.

The frequency converter is a self-oscillating mixer which uses emitter injection. The conversion gain of this circuit is 30 db. To supplement the agc at strong signals, an overload diode circuit is used between the converter and the first i-f stage. The overload diode circuit provides about 50% of the gain control of the receiver.

High Performance Receiver. Figure 36 shows a receiver circuit that provides the type of performance needed in mobile equipment. Six transistors are used and the tuner is inductively tuned.

The audio output stage is operated class A and provides four watts of audio output power. R_{18} is a factory adjustment which is used to adjust the collector current to its proper value. This adjustment is necessary because of the non-uniformity of reasonably-priced power transistors. The audio driver is transformer coupled to the output stage and the combination of the driver and output stages provide between 60 and 70 db of power gain.

A transistor detector is used providing a 5 db gain. The output circuit of the detector is connected to the volume control in such a way that a reduction in volume control setting reduces the resistance in the collector circuit. This action prevents voltage limiting at strong signals. The combination of R_{12} and C_{10} is a filter for the agc current.

The i-f amplifier is a single stage amplifier using double-tuned transformers. The gain of this stage is 30 db. The frequency converter is an inductively-tuned self-oscillating mixer using emitter injection. The r-f stage is also inductively tuned and the combined gain of the frequency converter and the r-f stage is 55 db.

Automatic gain control is applied to both the r-f stage and i-f stage. However the amount of agc applied to the i-f stage is very small and is not effective until strong signals are reached. Approximately 80% of the gain control comes from the r-f stage. At strong signals the r-f stage is reverse biased, and therefore, the receiver does not overload for any reasonable signal.

CHAPTER 4

TRANSISTOR TELEVISION CIRCUITS

INTRODUCTION

The major advantage of a transistorized TV receiver is a very large reduction in total power consumption. Small size and weight are also advantageous but the reduction in power drain makes possible a true portable, capable of operating from is own self-contained power pack. A typical tube-equipped portable receiver consumes about 165 watts, of which about 60 watts is used to heat the filaments. This tube portable operated from a standard automobile battery yields only a few hours of viewing time. A transistorized receiver eliminates all heater power except that required for the kinescope, and possibly the high-voltage rectifier. In addition, the d-c power consumed by certain stages is drastically reduced. For example, a 6CB6 i-f amplifier dissipates about 2 watts of d-c power, while a transistorized i-f stage may consume only 5 milliwatts. Complete receivers may consume only 25 to 40 watts. One small-screen unit has been made that dissipates less than five watts.

The block diagram of a transistorized receiver is the same as that of a conventional tube-equipped receiver. Also, the functions of each block are the same. Thus we need only consider the special features, limitations, and requirements of transistorized circuits in each of the television sections. The r-f, i-f sound, and video sections will be covered first. Following a brief discussion of the switching characteristics of transistors, the synchronization and scanning circuits will be studied.

1. THE RADIO-FREQUENCY SECTIONS

The Tuner. The functions of the tuner are the same in both tube and transistor versions. The primary job is to convert the selected channel frequencies to the intermediate frequencies employed in the receiver. Some gain and selectivity are also provided by the tuner but these are of secondary importance. Some of the more important tuner considerations are: (1) high signal-to-noise ratio at the output of the mixer; (2) low local-oscillator radiation; (3) good image rejection; (4) variable gain (for agc purposes); and (5) the prevention of signal overload when receiving strong signals.

High signal-to-noise ratio at the mixer output requires a maximum signal amplitude at the input to the mixer stage. Transistorized tuners are very similar to tube-type tuners in that both contain three stages performing the functions of r-f amplifier, mixer, and local oscillator. The r-f stage also serves to improve image rejection and reduce oscillator radiation. Its gain may be varied for agc purposes.

The tuned circuits employed in transistorized tuners are similar to those found in vacuum-tube tuners. Channel changing is accomplished by varying circuit inductance. Both turret and switch type (incremental inductance) tuners may be used.

R-F Amplifiers. Either common-base or common-emitter amplifiers may be chosen for the r-f stage. The common-emitter circuit is generally preferred because its gain characteristics are affected less by variations in input and load impedances.

A basic common-emitter r-f amplifier circuit is shown in Fig. 1. The antenna input circuit is conventional in employing an elevator transformer or a balun to match the 300-ohm balanced transmission line to the single-ended amplifier input. A parallel-resonant trap is placed in series with the output of the elevator transformer to attenuate signals at the receiver's intermediate frequency.

The input tuned circuit, consisting of L_2C_2, the series circuit of C_3 and the input impedance of the transistor is loaded with two low-resistance loads. Both the transformed antenna resistance and the low input resistance of the amplifier load this tank. By comparison, the input resistance of a vacuum-tube amplifier only lightly loads the input-tuned circuit. To secure the desirable bandwidth of 6 mc with this degree of loading requires a tuned circuit of abnormally high unloaded Q. In practice standard coils and capacitors are employed

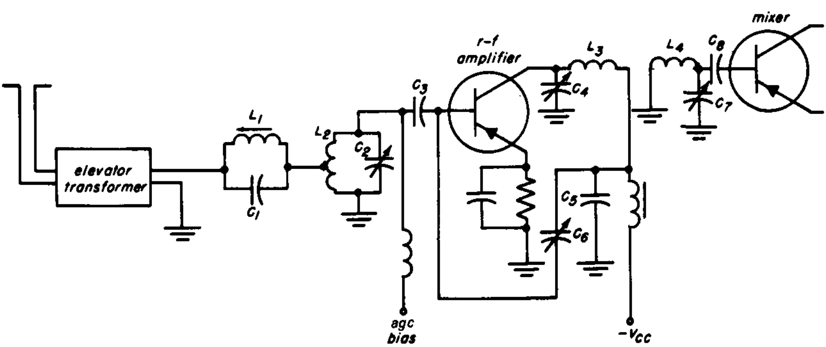


Fig. 1. A simplified schematic diagram of a TV r-f amplifier.

resulting in a heavily damped, wide-bandwidth circuit. The selectivity required to reject image and other unwanted signals is then achieved in the coupling circuit between the r-f and mixer stages. Note that more elaborate filters could be employed in the input circuit of the r-f amplifier. However, these would complicate the switching circuit and attenuate the signal. In the interest of a high signal-to-noise ratio, the largest signal possible is applied to the base of the r-f stage.

The coil L_2, is tapped to match the antenna input to the total tank impedance. Likewise, the transistor input is matched to the tank by means of the capacitive divider formed by C_3 and the transistor input (base-emitter) capacitance. In some cases a slight mismatch is arranged between the tank and the transistor to achieve a better signal-to-noise ratio. The base-circuit tank is tuned by a capacitive trimmer, C_2. This capacitor is made as large as possible to minimize the effects of variations in the input capacitance of the transistor upon circuit tuning. In this way the capacitance added by the transistor is a small fraction of the total tank capacitance. Transistor input and output capacitance vary due to changes in bias voltages and currents. Capacitance variations also occur due to differences between replacement transistors of the same type.

The coupling circuit between the r-f amplifier and the mixer is designed to provide the steep-sided selectivity curve required to reject signals outside the 6-mc bandpass. This coupling circuit is very similar to those found in conventional vacuum-tube tuners. It employs two tuned circuits, one in the collector circuit of the r-f amplifier and the other in the base circuit of the mixer. These tanks are coupled magnetically as shown in the figure. As an alternate system, the signal may be developed across a mutual impedance. The response curve shape is determined by the degree of coupling between the tuned circuits, the Q, and the resonant frequency of both tanks. The primary tuned circuit, consisting of L_3, C_4 and C_5 is loaded by the relatively high output impedance of r-f transistor. In that secondary circuit the tank is matched to the low input impedance of the mixer by means of the capacitive divider formed by C_8 and the input impedance of the mixer. The mixer could also be driven from a low-impedance tap on the coil but tapped coils require more complicated channel-switching arrangements.

Preventing Instability. Internal feedback occurs in the transistor from the collector circuit to the base via the collector-junction impedance. This impedance consists of the high collector resistance shunted by the junction capacitance. In drift-field transistors the resistive component is very high and may be ignored. This leaves a feedback path via the collector junction capacitance, C_c, as in Fig. 2. Note that the feedback signal appearing

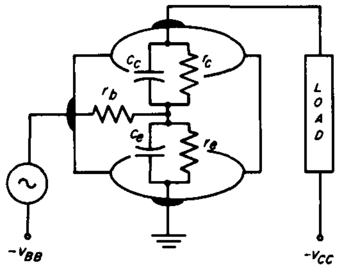


Fig. 2. This simplified drawing shows the relation of the junction capacitances to the load and source impedances.

at the base is that produced by a voltage divider consisting of C_c and the base-to-emitter impedance. If the collector and the base-to-emitter impedance is resistive, the feedback is degenerative due to the phase inversion between base and collector voltages. In tuned amplifiers, however, collector and base impedances become reactive at off-resonant frequencies. The feedback signal then undergoes a triple phase shift as it is developed across the collector impedance, fed through the collector capacitance, and is developed across the base-to-emitter impedance. When this phase shift adds up to 180 degrees, the feedback signal becomes regenerative. Oscillation may then occur if the transistor's power gain just equals or exceeds the attenuation that the feedback signal undergoes in the feedback path. Oscillation in an amplifier is a form of instability and it is prevented by removing one of the conditions under which oscillation may occur.

One method of stabilizing the amplifier is to reduce its gain. In that case the feedback signal is never large enough to produce oscillation even though the phase of the feedback signal is regenerative. The gain reduction need not be severe. It is only necessary to lower the forward gain to a value that is just below the attenuation provided by the internal feedback system. The total gain around the loop then has a net negative value and oscillation cannot take place.

The second method of stabilizing the amplifier is to cancel the effect of the internal feedback voltage with an externally applied feedback signal of opposite phase. This system, called neutralization, is shown in the r-f stage shown in Fig. 1. A feedback signal 180 degrees out-of-phase with the collector voltage is obtained from a capacitive voltage divider across the coil L_3. C_4 and C_5 make up this divider. This provides a signal at the bottom of the coil that is inverted in polarity compared with the collector signal. The same type of inverted signal may be obtained by grounding a tap on the coil, but tapped coils present a switching difficulty. In addition, by means of the capacitive tap the amplitude of the feedback signal remains a substantially constant fraction of the collector signal amplitude. Thus neutralization is maintained despite coil changes when switching channels. The neutralizing signal is fed back to the base of the amplifier via the trimmer capacitor C_6. Complete neutralization is obtained when C_6 is adjusted so that the neutralizing voltage just equals the internal feedback voltage. When this condition is met the collector capacitance is effectively removed and the input and output circuits are isolated except for the normal, forward-amplifying action.

Neutralization is chosen over gain reduction in this case to allow the r-f amplifier to operate at maximum gain. This provides a higher amplitude signal at the base of the mixer. Usually a combination of both methods is employed to allow for differences in collector capacitance and imperfect neutralization. Neutralization also affects the gain control characteristics of the amplifier. When very strong signals are received, the agc system may be set up to reverse bias the emitter junction of the r-f amplifier. The amplifier is cut off and acts as an attenuator. In this condition, the signal is passed through via the collector capacitance. If the amplifier is perfectly neutralized, the feedthrough signal is cancelled in the collector load and extremely high attenuation results. The degree of attenuation may be adjusted to some extent by varying the degree of misneutralization.

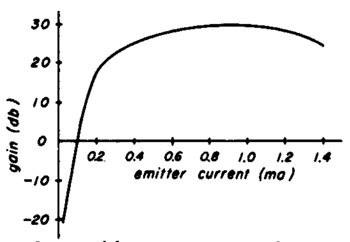


Fig. 3. Amplifier gain versus d-c emitter current for an agc-controlled i-f stage.

Agc Action. The gain of the r-f amplifier is varied by altering its d-c bias conditions. Two methods may be employed: reverse agc and forward agc. In reverse agc the gain of the amplifier is reduced by reducing the d-c emitter current from some initial value. Gain control is based on the drop in beta of the transistor at low values of emitter current. Also input and output resistances change with changes in the d-c operating point, and the resultant mismatch to source and load impedance further reduces the power gain of the amplifier. Figure 3 shows a plot of power gain versus emitter current for a representative high-frequency amplifier. Control of emitter current may be obtained by means of a control signal applied either at the base or the emitter of the amplifier, as in Fig. 4. In

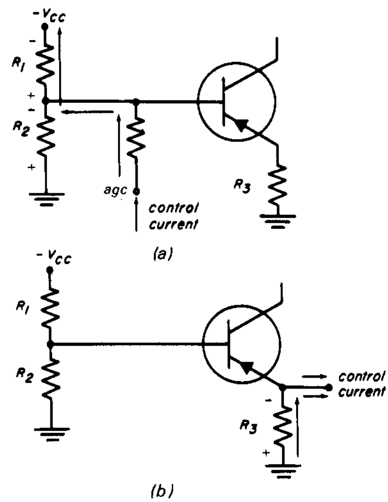


Fig. 4. Automatic gain control is achieved by altering the d-c collector current. (a) Base voltage control; (b) emitter current control.

a of the figure, emitter current is controlled by varying the base voltage. The bias network (R_1, R_2) provides forward bias for the transistor in the absence of the control signal (weak signal conditions). The application of a positive control voltage causes current to flow through the bias network as shown, reducing the base voltage towards zero. As the control voltage is raised still more, the base voltage swings positive with respect to the emitter, cutting the stage off. The amount of signal transferred through the stage then depends upon feedthrough as discussed earlier.

Reverse agc, by emitter current control, is obtained as shown in *b* of the figure. No-signal bias is obtained, as before, so that collector current starts out at about one milliampere. The application of a control current as shown tends to make the emitter voltage more negative, and since the base voltage is held constant by the voltage divider, emitter bias is reduced and emitter current drops. As the control current is raised, more of the total current flowing through R_3 becomes control current and the emitter current fraction diminishes. The voltage across R_3 remains fairly constant and just equals the voltage developed across R_2 when the emitter current drops to zero. At this point the emitter-base bias voltage is zero. A further increase in control current causes the emitter to swing negative with respect to the base and signal attenuation again depends upon feedthrough. Forward agc is discussed in the next paragraphs.

Overload. Strong signal overload is an important consideration in the television r-f amplifier. If reverse agc is employed an increasing signal amplitude causes the base-to-emitter voltage to fall towards zero. Under these conditions the signal voltage, which may approach a volt in very strong signal locations, is rectified by the base-emitter diode and signal overload occurs. One solution to the problem is to arrange the agc bias so that the signal level is still small as the base bias voltage passes through the very nonlinear region near the origin of the emitter-base diode characteristic. Note, as shown in Fig. 5, that the curve is linear in both regions of forward and reverse bias as shown. Signal distortion is most likely in the region centered about zero. Distortion is minimized by ensuring that signal level is low as this region is traversed.

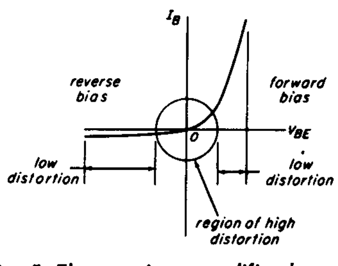


Fig. 5. The transistor amplifier becomes a nonlinear amplifier at low values of V_{BE} and I_B. To reduce distortion the signal level should be low when the transistor is operated at low values of V_{BE}.

Forward agc offers another solution to the overload problem. In this system, gain control is achieved by controlling collector voltage. At low collector voltages (below one volt) the beta and the output impedance of the transistor decrease with decreasing collector voltage. Collector voltage is controlled by inserting the bypassed resistor R_c in the collector circuit, as in Fig. 6. The control current is then applied so as to *increase* forward bias with increasing signal strength. The rise in collector current increases the drop across the collector resistor, reducing V_{CE} towards zero. Note that agc control causes the base-emitter voltage to increase, placing the operating bias further away from the nonlinear region. The improvement in overload characteristics is limited, however, because the stage gain operates nonlinearly as collector voltage approaches zero. A disadvantage of forward agc is the change in collector-junction capacitance with collector voltage.

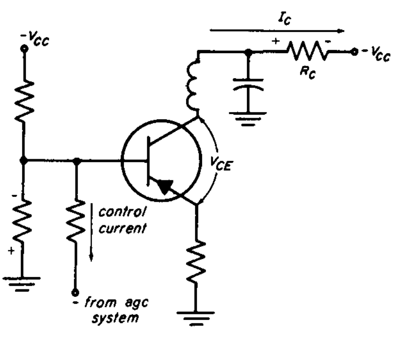


Fig. 6. In the forward agc system collector current is made to increase with an increase in signal level, but collector voltage drops due to the bypassed resistor R_C.

This results in detuning of the collector load under varying signal conditions. The detuning effect outweighs the slight improvement in overload characteristics so that the use of forward agc system is limited. In the i-f amplifier, however, the detuning effect may be used to advantage to secure a desired change in the receiver's overall frequency response at different signal levels.

Mixer Circuits. Heterodyning takes place in any mixer circuit by passing the signal and local-oscillator voltages through a nonlinear impedance. In transistor mixers, the nonlinear impedance is the rectifying base-to-emitter junction. The modulation envelope resulting from the addition of a CW and a modulated signal is applied to a base-to-emitter diode having a slight forward bias. Due to the nonlinearity of the diode in this region, the base current emerges with an a-c component of signal at the intermediate or difference frequency. The base current is amplified by the transistor so that the mixer also provides useful gain at the intermediate frequency.

As far as mixing action is concerned, it makes no difference if either the base or the emitter is grounded. Also the signal voltage and oscillator voltage may both be applied at the same terminal or one may be applied to the base and the other may be applied at the emitter. The common-emitter circuit is usually employed, however, as it provides the greatest power gain at the intermediate frequency. Note that the mixer does not have to amplify the oscillator and r-f signals but only amplifies the 45-mc difference-frequency signal.

A basic mixer circuit is shown in Fig. 7. The r-f signal is developed in L_1 and tapped capacitively to match the input impedance of the mixer. The oscillator signal may be injected either at the base or at the emitter as shown. Best mixing action is achieved if the oscillator signal is large enough to make the emitter-base diode reverse biased for a small part of each cycle. (This requires the oscillator peak voltage to exceed the d-c forward bias developed by resistors R_1, R_2, and R_3. Typical injection voltages lie in the range of 0.1 to 0.3-volt rms. The amount of forward bias provided by the bias network is a compromise between gain and mixer efficiency. Gain increases at higher emitter currents, but the signal moves up into the more linear region of the emitter diode's characteristic, and detection efficiency decreases.

Neutralization is unnecessary in the mixer stage, since the output-signal frequency is different than the input-signal frequency. Thus, little of the collector feedback signal can be developed in the base circuit. However, any i-f signal fed back is degenerative, tending to reduce mixer gain. Hence, the mixer gain may be improved somewhat by further reducing the base impedance at the intermediate frequency. The series resonant trap L_2C_2, tuned to 45 mc, accomplishes this job.

The Local Oscillator. The common-base amplifier circuit is well suited for the local oscillator. Since the collector voltage is in phase with the emitter voltage, the problem of obtaining feedback of the proper phase is simplified.

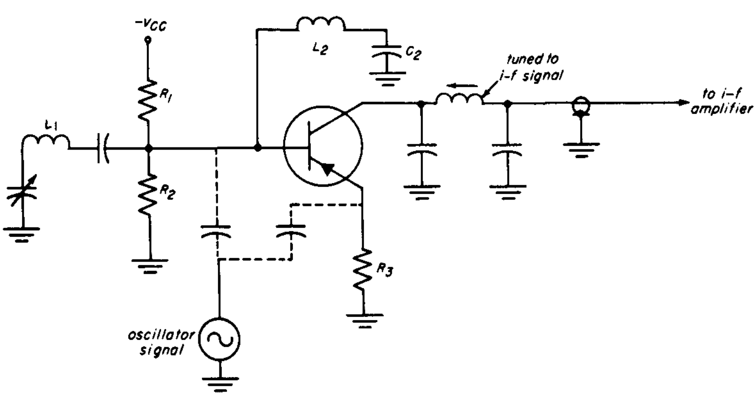


Fig. 7. In the mixer circuit the r-f and oscillator signals are added across the emitter junction.

The basic oscillator circuit is shown in Fig. 8. It is a common-base amplifier with an external feedback path provided between collector and emitter by the capacitor C_f. A large value emitter resistor allows the required fraction of the collector voltage to be developed between emitter and ground. The base is returned to r-f ground by means of a feed-through type of bypass capacitor which has very low lead inductance. As can be seen, the circuit resembles the Colpitts oscillator circuit in that feedback is obtained from a capacitive voltage divider consisting of C_f and C_e. Forward bias is provided by the network consisting of R_e, R_1, and R_2. Oscillator power is adjusted by controlling the emitter current and hence the beta of the transistor. Greatest output is obtained at collector currents of about 1 ma. An oscillator of this type, using a VHF drift-field transistor, can provide much more r-f power than is required by the mixer stage.

The feedpoint from which r-f energy is obtained depends upon the way in which signals are applied to the mixer. If both r-f and local oscillator voltages are applied to the same mixer terminal, then the feedpoint is obtained from a high-impedance point in the oscillator circuit. This prevents shunting of the r-f signal by the source impedance presented by the oscillator. A high-impedance feed is obtained by means of a small coupling capacitor, C_1, from the collector terminal. If oscillator and r-f signals are applied at different terminals, a low source impedance is required. A low impedance source of oscillator voltage is obtained as shown in the figure. C_2 is in series with the tuning capacitance C_t, and is

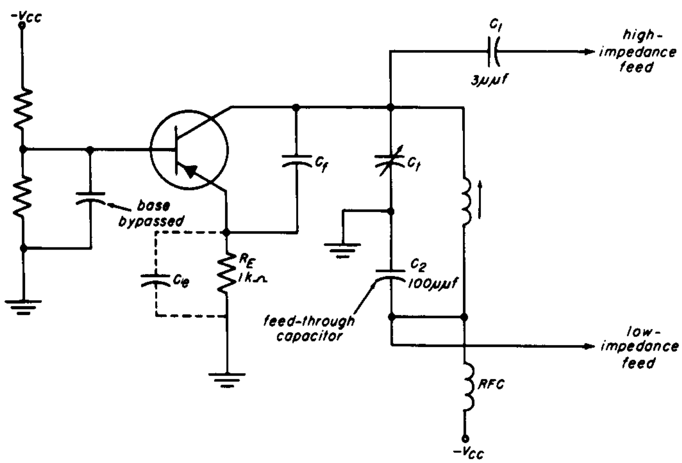


Fig. 8. This local oscillator circuit uses a common-base amplifier circuit. Feedback is from collector to emitter.

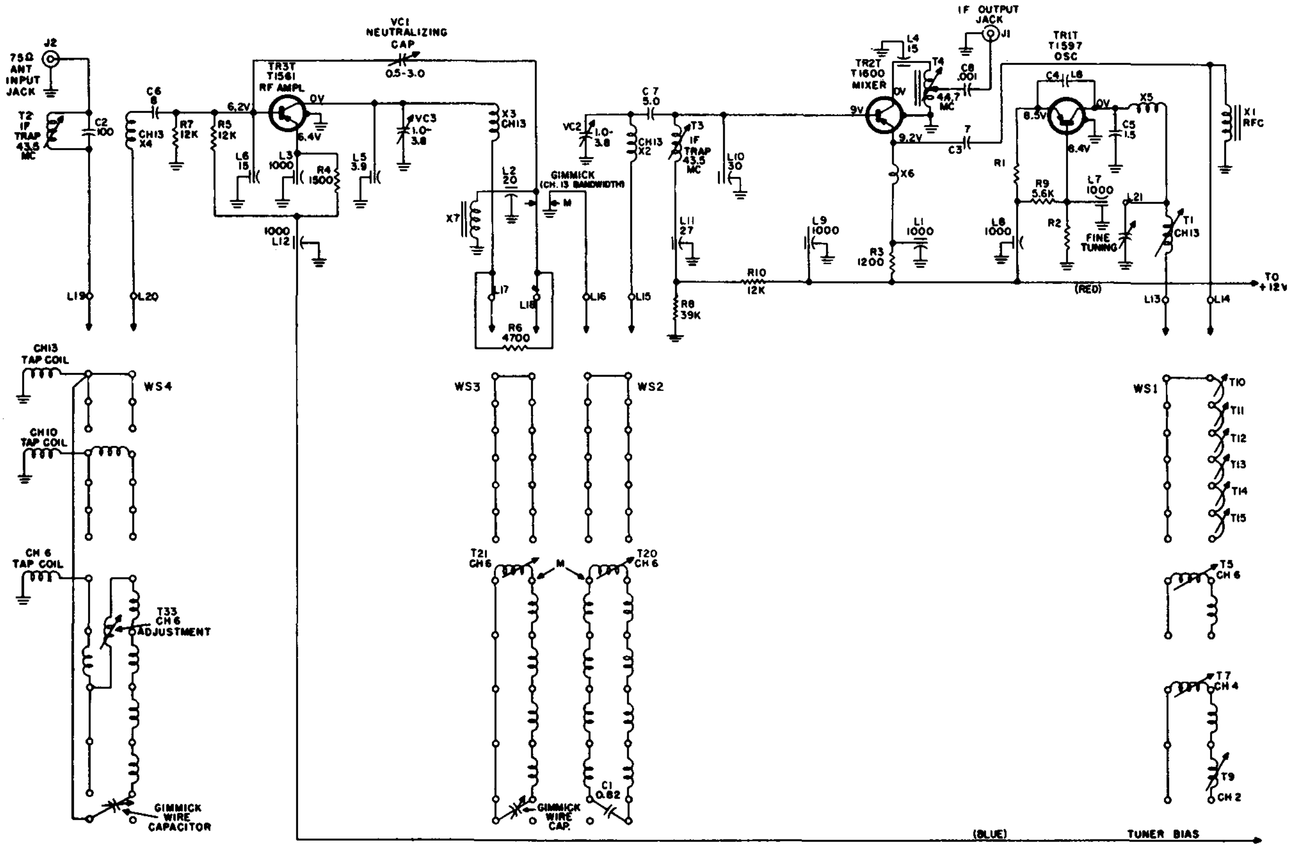


Fig. 9. A representative transistorized TV tuner.

large compared to C_t. Thus the oscillator voltage developed at this point is low. When the voltage developed across C_2 is applied to the emitter (or base) of the mixer, the large value of C_2 acts to keep the emitter or base impedance low at the intermediate frequency. Thus the gain of the mixer at the intermediate frequency is kept at a maximum.

Fine tuning is accomplished by the same means employed in conventional tuners. A generally larger tuning range is provided to allow for increased frequency drift due to changes of transistor parameters with temperature.

Complete Transistorized Tuner. A practical TV tuner is shown in Fig. 9. This tuner is designed for use in a portable receiver. It is fed from a whip type of antenna which provides a single-ended input signal. Circuit inductance is changed by means of an incremental inductance or switch-type system.

The r-f amplifier is similar to the simplified circuit shown earlier. The input signal is applied through a parallel-resonant trap T_2C_2 tuned to the center i-f bandpass. A match between the input tank circuit and the antenna is obtained by means of the tapped inductance selected by switch contacts L_{19} and L_{20}. A common tap-coil serves this function for Channels 13, 12, and 11. Another coil is switched in for Channels 10, 9, 8, and 7 and a third matches the antenna impedance for the low channels. The small "gimmick" tunes the unused low-channel coils to a frequency that prevents energy absorption by these coils when the set is tuned to the higher channels. A capacitive voltage divider consisting of C_6 and feedthrough capacitor L_6 matches the loaded tank impedance to the low input impedance of the r-f transistor.

Bias for the r-f amplifier is conventional except that the collector is returned to chassis ground and a *positive* supply voltage is applied at the emitter. Forward bias for the base-emitter junction is provided by the resistive divider R_7 and R_5 which holds the base negative with respect to the emitter. A large (1500 ohms) emitter resistor ensures adequate bias stabilization. The emitter resistor is bypassed by a large feedthrough capacitor to prevent degeneration at r-f frequencies.

In the collector circuit of the r-f amplifier, the total inductance, consisting of X_3 and the inductance added by the channel-selector switch, tunes with feed-through capacitor L_5, tuning capacitor V_{C3}, and the output capacitance of the transistor. The low end of inductance X_3 is held above r-f ground by the inductance X_7. The voltage developed across the feedthrough capacitor L_2, is fed back via V_{C1} for neutralizing purposes. The switch decks holding the inductances tuning the r-f amplifier output and the mixer input circuits are in close proximity and careful placement of the coils provides the desired degree of coupling.

A divider consisting of capacitors C_7 and L_{10} matches the base circuit tank to the base impedance of the mixer. A series resonant trap, consisting of T_3 and L_{11}, is also placed between the base and ground to reduce the base impedance at the intermediate frequency to a very low value. This prevents feedback signals at 43.5 mc from being developed in the base circuit. Since the collector circuit is resistive at this frequency, such feedback is normally degenerative. The trap, therefore, resuls in an increase in mixer gain. Another feature of the base circuit is that L_{10} and its lead inductance are chosen to resonate at some frequency above 215 mc. This reduces the base impedance to local-oscillator frequencies when tuned to the higher channels, and acts to reduce oscillator radiation.

Oscillator injection voltage is developed across the coil X_6 in the emitter circuit of the mixer. X_6 is actually a small part of the total oscillator tuning inductance and connects the cold side of the tuning inductance to r-f ground via C_3 and L_1. Note that the collector circuit of the oscillator is returned to d-c ground through the choke X_1 which isolates the tuning inductance from ground at this point. X_6 offers a low impedance to r-f and i-f signals and so has little effect upon mixer gain. Small variations in injection voltage can be obtained by varying the inductance of X_6 which is a small "hairpin" coil.

Inductance T_4, in the collector circuit of the mixer, is tuned to 44.7 mc with L_4, a 15-$\mu\mu$f feed-through capacitor, and the output capacitance of the mixer transistor. The coil is tapped down to a low impedance point in order to match the coaxial line that supplies the mixer output to the i-f amplifier.

The oscillator is similar to the simplified circuit shown earlier. It employs the common-base amplifier circuit with feedback adjusted by choosing proper values for C_4 and L_8.

Agc is not applied to the r-f amplifier in this tuner. Instead the gain of the r-f stage is controlled manually in three steps corresponding to weak, medium, and strong signal conditions. The manual selector switch (not part of the tuner) chooses the appropriate collector voltage to suit signal conditions. Collector voltage should be low for strong signals, slightly higher for medium strength signals and high for weak signals. For the first two conditions, the collector voltage is under one volt so that gain reduction is obtained. For weak signals, the collector voltage is made high so that full gain is obtained. The collector voltages chosen for the three signal conditions are selected to minimize nonlinearity.

2. INTERMEDIATE-FREQUENCY AMPLIFIERS

I-F Amplifier Requirements. The greatest portion of the total receiver gain is obtained in the intermediate-frequency amplifier. The approximate gain required may be calculated as follows. Assume that the minimum usable signal is 50 microvolts developed across 300 ohms. This gives an input power of 8.3×10^{-12} watts. At the picture tube, about 70 volts developed across a 10-k ohm resistor is 490×10^{-3} watts. Total power gain required from the antenna terminals to the input of the kinescope is then:

$$\text{power gain (db)} = 10 \log \frac{490 \times 10^{-3}}{8.3 \times 10^{-12}}$$

$$= 10 \log 5.9 \times 10^{10}$$

$$= 108.3 \text{ db}$$

The r-f amplifier and the mixer contribute about 20-db gain, and the video amplifier provides 30-db power gain, leaving a total power gain of approximately 58 db to be obtained in the i-f amplifier. A four-stage i-f amplifier then requires about 14 db of gain per stage. This relation holds for four wide-band amplifiers if each is tuned to the same frequency. However, to obtain the response curve needed to handle the vestigial sideband properly, stagger tuning is required. Stagger tuning introduces some loss into the system so that a greater stage gain, about 17 db per stage, is required.

In addition to the maximum gain required for weak signals, the i-f amplifier is required to maintain a nearly constant output for input voltages approaching one volt in amplitude. This means that the agc system should be capable of reducing the i-f gain from a maximum of about 60 db to a low of practically zero db.

Basic I-F Amplifier. A drift-field type of transistor in the common-emitter arrangement is used in the circuit shown in Fig. 10. The stage is stabilized against changes in the d-c operating point by the bypassed emitter resistor R_3 and the base voltage divider R_1R_2. In the absence of an agc current, the bias system is designed to develop a collector current of about one milliampere with a collector-to-emitter voltage of about five volts. The transformer is then chosen so that the impedance at its input terminals matches the output impedance of the transistor at these bias conditions. In this way, maximum power gain is obtained. The application of reverse agc then tends to both reduce the current gain of the

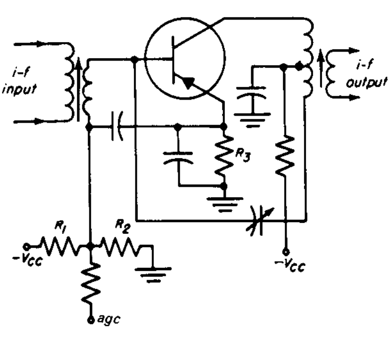


Fig. 10. The basic common-emitter i-f amplifier.

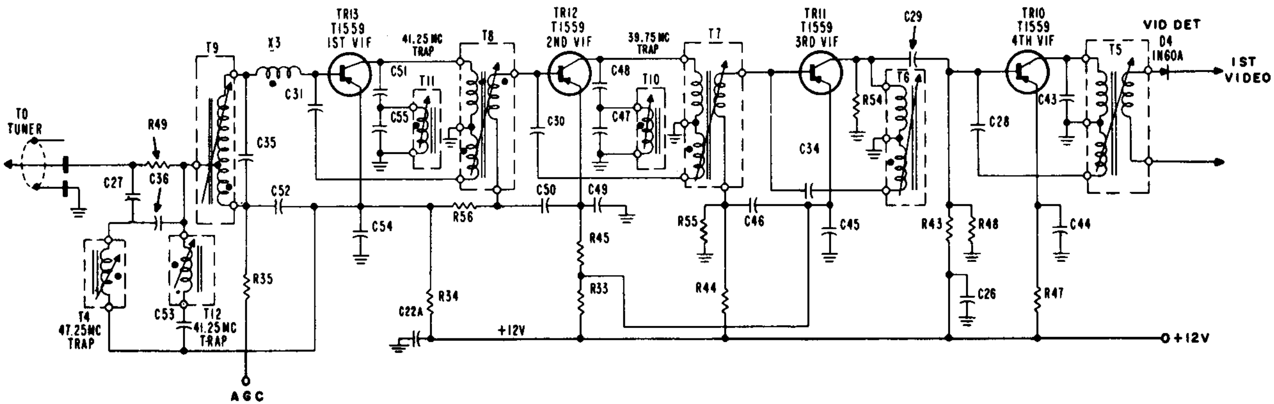


Fig. 11. A representative i-f amplifier.

transistor and upset the match between the transistor and its load. The latter occurs because the output impedance of the transistor increases with a decrease in emitter current.

In addition to lowering power gain, the effect of an increase in output impedance is to reduce the loading on the tuned circuit, resulting in higher circuit Q and narrower bandwidth. In systems using reverse agc, bandwidth tends to decrease as signal strength increases. When forward agc is applied, the reverse condition occurs. In this case an increase in signal causes emitter current to increase. Current gain drops due to a lowering of collector voltage and again a change in output impedance results in a mismatch between transistor and load. Output impedance decreases, however, and bandwidth tends to broaden with increasing signal.

The i-f amplifier may be stabilized against oscillation by gain reduction, as discussed earlier, or by one of the several neutralization circuits. Figure 10 shows an example of neutralization by means of a feedback signal obtained from the transformer primary. The tap of the transformer primary is grounded for a-c signals, and the voltage obtained at the bottom of the coil is therefore inverted with respect to the collector voltage.

A representative i-f amplifier is shown in Fig. 11. It employs four common-emitter amplifiers driving a conventional diode detector. Agc is applied to the first three i-f stages. The final stage bias is not varied, as is common practice, to provide minimum distortion of the amplified i-f signal.

Before discussing the agc and bias circuitry let us consider the i-f signal path by itself. The signal from the tuner is coupled via a low-impedance coaxial cable. T_9 is an impedance matching transformer inserted to terminate the cable in its characteristic impedance and match the cable to the higher input impedance of the first i-f amplifier.

Two series-resonant traps, tuned to the selected- and adjacent-channel sound i-f frequencies shunt the cable. The traps become parallel resonant at some frequency inside the video bandpass and so are effectively removed from the circuit at these frequencies.

Transformer T_9 is tuned by C_{35} and the signal developed across this tank is coupled via the coil X_3 to the base of the transistor. Coil X_3 is part of the impedance matching system. It prevents excessive loading of the tuned circuit by the transistor. The signal is applied between base and emitter of the transistor. Bypass capacitor C_{52} provides a return path for r-f signals but prevents shorting of the d-c base-to-emitter bias voltage. An emitter bypass capacitor, C_{54}, returns the emitter to a-c ground.

In the collector circuit the amplified signal is developed across the primary of the matching transformer, T_8. The primary of T_8 resonates with the output capacitance of the transistor at the desired frequency. The low impedance secondary matches the input impedance of the following stage. A capacitively-coupled trap, of the same type employed in vacuum-tube i-f circuits, is placed across the collector circuits of both the first and second i-f amplifiers. The first (T_{11}) attenuates the selected-channel sound signal to the proper level for intercarrier detection. The second trap (T_{10}) attenuates the adjacent-channel picture carrier signal.

A tap on the primary of all the i-f transformers is grounded directly to the chassis. This provides a feedback voltage, inverted with respect to the collector voltage at the lower end of the primary winding. The feedback voltage is coupled back to the base (via C_{31} in the first stage) for neutralizing purposes.

The i-f circuits of the remaining amplifiers are similar to the first with the exception of the coupling circuit between the third and fourth amplifiers. A capacitive voltage divider, formed by C_{29} and the input impedance of the fourth i-f stage is used for coupling in this case. The input impedance of the final i-f stage is somewhat higher than that of the other stages. The reason is that its load, the detector, provides a relatively low collector load impedance. It is characteristic of the common-emitter circuit that the input impedance becomes larger as the load impedance is made smaller.

Agc Circuit. At first glance it appears as though agc is applied to the first stage only. Closer examination shows that the agc is direct coupled through the amplifiers and applied to the second and third amplifiers as well. Forward agc is utilized in the first two amplifiers while reverse agc is applied to the third amplifier. We shall consider agc action in the first stage and then trace the application of agc to the following stages.

In the absence of an agc signal (no antenna signal) a positive reference voltage of about 2.1 volts appears at the agc terminal. Since the emitter is returned through R_{34} to a $+12$ volt supply, the base is negative with respect to the emitter and the emitter junction is forward biased. The circuit is shown in simplified form in Fig. 12. Conduction of the transistor causes the voltage at the emitter to level off at about 2.4 volts so that a forward bias voltage of 0.3 volts exists from base to emitter. Under these conditions the total drop across the emitter

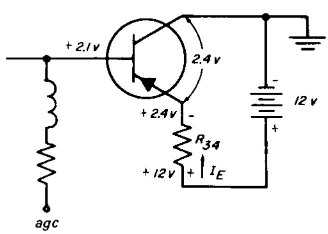


Fig. 12. Simplified bias circuit for the first i-f amplifier in Fig. 11.

resistor is 9.6 volts, and the collector-to-emitter voltage is 2.4 volts.

As the signal strength increases from zero, the voltage supplied to the agc terminal swings less positive. This increases the forward bias at the emitter junction, I_E rises, and the voltage drop across the emitter resistor also rises. The resultant drop in collector-to-emitter voltage provides the forward agc action. As collector-to-emitter voltage drops below one volt, gain diminishes rapidly. Varying the collector voltage between one and two volts causes little change in gain, so that the system has a sort of "built-in" agc delay.

Note that the first-stage emitter voltage, with respect to ground, drops from a high of plus 2.4 volts toward zero as agc is applied. This emitter voltage provides agc bias for the following stage. It is coupled to the base of the second stage through isolating resistor R_{56}. In the second stage, agc control is just like that of the first. An increase in signal causes the base voltage to drop down from a high of 2.4 volts. In a similar manner, the emitter voltage of the second stage is applied as control for the third stage. In this case, however, the second stage acts to control the emitter voltage and not the base voltage. The base voltage is held at a fixed value by the voltage divider R_{44} and R_{55}. The emitter voltage is tapped from voltage divider (R_{33} and R_{45}) in the emitter circuit of the second stage. Under no-signal conditions (no agc) the voltage tapped off at the junction of R_{33} and R_{45} makes the emitter slightly positive with respect to the fixed base voltage. The no-signal bias is calculated to provide maximum gain of the third stage. As signal increases the emitter voltage of Stage *2*, and likewise that of Stage *3*, becomes less positive. The emitter of Stage *3* then becomes less positive with respect to its base and forward bias decreases. Thus *reverse* agc is employed at the third stage. In employing both forward and reverse agc in the same i-f amplifier, the opposing effects upon bandwidth tend to cancel and the over-all bandwidth is stabilized over a wide range of signal levels.

Video Detector. Diode detectors are capable of operating with a minimum of distortion over a wide range of input voltages. Basically, the detector circuit is unchanged from those employed in tube receivers. An example of a diode detector used to drive a transistorized video amplifier is shown in Fig. 13. The detector circuit uses

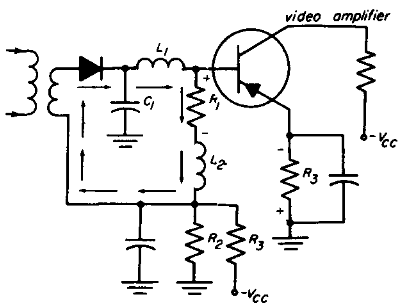


Fig. 13. This detector circuit is coupled directly to the video amplifier.

a series diode. A low-pass filter, L_1 and C_1, removes the i-f signal component. The load consists of R_1 in series with the shunt-peaking coil L_2. Both of these components are shunted by the relatively low input impedance of the transistor amplifier. In this circuit the detector is direct-coupled to the video amplifier. The entire circuit is above ground, and the voltage developed across the load is applied in series with the bias voltage developed across R_2. Direct coupling preserves the d-c component of the detected signal. This makes possible the extraction of an agc signal at the output of the first video amplifier.

Note that the diode polarity is reversed compared to the conventional video detector circuit. This provides, at the base of the video amplifier, a signal in which sync extends in the positive direction. Negative-going sync is normally applied to vacuum-tube amplifiers to achieve some measure of impulse noise protection. In that case noise pulses which exceed sync amplitude exceed cutoff and do not appear in the plate signal. The same action may take place in the transistor amplifier as shown in Fig. 14. In this case a PNP transistor is employed and the bias supply holds the base negative with respect to the emitter. A positive signal excursion, which exceeds the voltage developed across R_2, causes the emitter-base junction to become reverse biased, cutting off the amplifier. Limiting action may also be obtained by driving the transistor

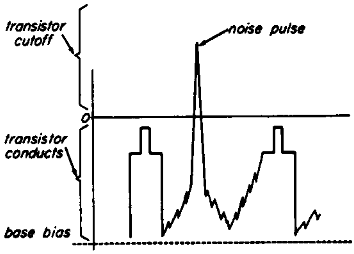


Fig. 14. Noise pulses at the output of the video detector in Fig. 13 drive the video amplifier into cutoff.

further into conduction. In this case limiting occurs when collector current rises to a value very close to the maximum that can be supplied from a given supply voltage and load resistor. A transistor driven to this condition is said to be in *saturation*. In this condition, the collector voltage is very close to zero and collector current is practically equal to the supply voltage divided by the load resistance. Since collector current cannot increase beyond this limit, base input signals exceeding the forward bias necessary to produce saturation do not appear in the collector signal. (We shall examine saturation in greater detail later.) If saturation clipping is employed, the diode shown in Fig. 14 should be reversed so that a negative-going sync signal is developed.

An important factor to be considered is the load that the video amplifier places upon the detector. In vacuum-tube circuits the load presented by the input to the video amplifier is negligible and the detector works into a load impedance of 3000 to 5000 ohms. A transistor video amplifier, however, may have an input impedance of only 500 ohms. This means that the total a-c load impedance on the detector is less than 500 ohms. The effects of a low load impedance are: a reduction in output voltage; lower efficiency; and higher distortion. In regard to efficiency and output voltage, a load resistance of 500 ohms is about equal to the forward conducting resistance of the diode. Thus the detector diode and its load split the total input voltage between them. Note that output voltage is a major consideration. Since the kinescope is a voltage operated device, voltage gain and not power is the purpose of the video amplifier. A low detector output voltage therefore requires increased voltage gain in the video amplifier.

Distortion occurs unless the detector is operated at a relatively high power level. Consider a detector developing ½ volt across 5000 ohms. The power delivered by the detector circuit is 0.05 milliwatts. The same power delivered to a 500-ohm load develops only 0.16 volt. A diode operated at this voltage level produces more distortion than one operated at higher voltage levels due to the curvature of the characteristic in this region, as shown in Fig. 15. To prevent these effects, the first stage of the video amplifier is designed to present a high input impedance as will be shown.

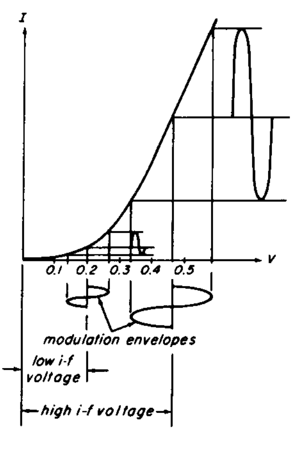


Fig. 15. Distortion occurs at low-signal levels in the diode detector.

Agc voltage for the i-f amplifier shown in Fig. 11 may be developed by rearranging the detector and first video amplifier in the manner shown in Fig. 16. Under no-signal conditions the voltage developed across the detector load R_2 is zero. A positive voltage appears at the emitter resistor which is a few tenths of a volt above that developed across R_1. When signals are being received, a d-c component of voltage appears across R_2 with the polarity shown. This voltage increases the forward bias applied to

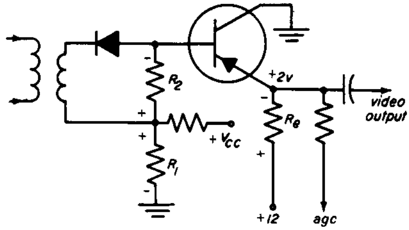


Fig. 16. This common-collector video amplifier provides a high-impedance load for the video detector and delivers a d-c agc output voltage.

the video amplifier and its emitter current increases. An increase in voltage drop across R_e causes the emitter voltage, and hence the agc voltage, to drop towards zero.

3. VIDEO AMPLIFIERS

Video Amplifier Requirements. The primary purpose of the video amplifier is to provide the driving voltage for the kinescope. A video output voltage of 50 to 100 volts peak-to-peak is required. Thus the final stage of the video system must be a voltage amplifier designed to work into the high impedance presented by the control circuits of the kinescope. In addition to voltage gain, the video amplifier must provide a uniform output voltage from 60 cycles to about 3.5 or 4 mc. Auxiliary functions of the video amplifier include furnishing feed points for 4.5-mc sound signal, synchronization signals, and in some cases agc signals.

The problem of obtaining wide-band amplification is similar to that in vacuum-tube amplifiers. Gain must be sacrificed to secure wide-band amplification. Maximum gain is obtained when the signal source resistance matches the input resistance of the amplifier and the load matches its output resistance. At high frequencies, gain drops off due to shunt capacitance. An important contributor to the drop in high-frequency gain is the capacitance that effectively shunts the input signal around the emitter junction. This capacitance is shown as $C_{b'e}$ in Fig. 17. Neglecting other factors, gain drops to 0.707 of its low-frequency value when the reactance of $C_{b'e}$ drops to a value equal to the sum of R_g and $r_{b'b}$ (the intrinsic base resistance).

Another factor effecting the apparent size of the capacitance shunting the emitter junction is internal feedback through the collector junction capacitance. This capacitance provides an alternate path for input signals. The effect of this capacitance is the same as though another shunt capacitance is added across $C_{b'e}$ whose value is the collector junction capacitance

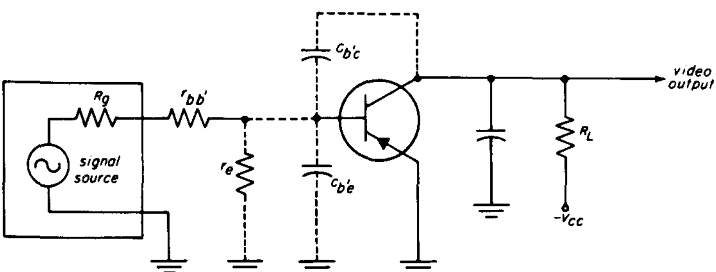


Fig. 17. Location of important junction capacitance in the simplified video amplifier.

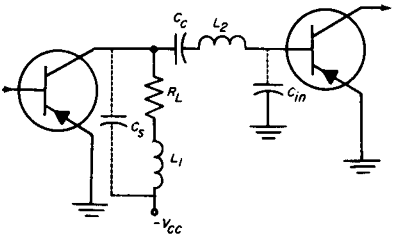


Fig. 18. Shunt and series peaking coils in the transistorized video amplifier.

times the voltage gain of the device. The over-all effect is very similar to the Miller effect affecting the input capacitance of triodes. The capacitance added by this feedback effect may be reduced by lowering the voltage gain of the amplifier. Voltage gain is lowered by reducing the value of the load resistance. In addition to reducing the effective input capacitance, a reduction in load resistance also minimizes the effect of the total capacitance shunting the load resistance. The capacitance across the load R_L is the sum of the output capacitance of transistor, wiring capacitances, and the input capacitance to the following stage.

A further increase in bandwidth may be secured by "tuning out" the reactances presented by the transistor by means of shunt and series peaking coils. Peaking is accomplished in much the same way as it is in vacuum-tube circuits and is shown in Fig. 18. L_1, the shunt peaking coil, tunes with the total capacitance across R_L forming a broad parallel peak at resonance. L_2, the series peaking coil, tunes with the input capacitance to the following stage.

Practical Video Amplifier. A practical video amplifier appears in Fig. 19. It employs two stages of amplification providing a bandwidth of 3.5 mc with a voltage gain of 40. The output stage employs a

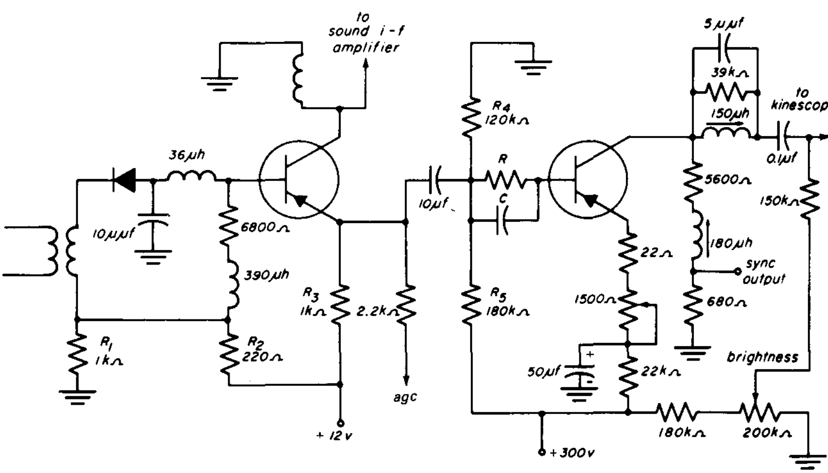


Fig. 19. A representative video amplifier.

single transistor in the common-emitter circuit. The BV_{CEO} for the output transistor used in this circuit is in excess of 100 volts.

The first stage is a common-collector or emitter-follower amplifier. This type of amplifier presents a very high input impedance and a low output impedance. These characteristics are desirable as they allow the detector to work into a high-impedance load and provide a low-impedance driving source for the output stage. The former allows the detector to operate at a higher voltage level which minimizes the nonlinearity occurring at low signal voltages. A low-impedance driving source improves the high-frequency response of the output stage.

Direct coupling is employed between the detector and the first video stage. The d-c component of the detected signal, appearing at the output of the detector, thus also appears across the load resistor of the emitter follower. This voltage is tapped off through a 2.2-k ohm isolating resistor and applied through a low-pass filter (not shown) to the controlled i-f stages. Direct coupling is achieved by inserting the detector load resistance between the base of the first stage and the base-bias voltage supply. Base bias is obtained from the voltage divider R_1 and R_2.

A positive agc voltage is produced which decreases with signal level. With no signal applied, a positive voltage appears at the emitter of the first stage which is a few tenths of a volt more positive than the base-bias voltage. As signal level rises the base swings in the negative direction increasing forward bias for the stage. The rise in emitter current increases the voltage drop across R_3 causing the emitter voltage to go less positive.

A single inductance in the collector circuit provides an impedance across which the 4.5-mc sound signal is developed. This feed system isolates the input impedance of the 4.5-mc sound circuits from the video circuitry.

An RC network is shown in series with the base lead of the output stage. It acts as a high-pass filter and corrects the natural roll-off in high-frequency response. A 10-μf coupling capacitor performs the task of d-c blocking and minimizes the loss of low-frequency signal components. Forward bias for the output stage is provided by R_4 and R_5.

Collector voltage for the output stage is obtained from a 300-volt supply. This high supply voltage permits the insertion of a 22-k ohm resistor in the emitter circuit of the stage. This large resistor determines to a large extent the d-c collector current of the stage and it acts to stabilize collector current against temperature variations. The 22-k ohm resistor is heavily bypassed and so does not affect amplifier performance in the normal video bandpass. An unbypassed section of emitter resistance is employed as the contrast control. An unbypassed emitter resistance reduces the gain of the stage by degeneration. Gain is raised to a maximum by shorting out the 1500-ohm emitter resistance. Since this resistance is a small fraction of the total d-c emitter resistance, the change in bias incurred by rotating the contrast control is small.

The coupling circuit to the kinescope cathode is conventional. It employs shunt-and-series-peaking coils to obtain the required high-frequency response. The collector load is tapped at a low-impedance point to provide a video-feed point to the sync separator circuit.

In some receiver designs, where cathode drive is necessary, the video amplifier of Fig. 20 may be employed. Here the video output stage is a common-base amplifier and no polarity reversal occurs with this configuration. Most of the above comments relative to the common-emitter amplifier also apply for this amplifier. The difference resides mainly in the magnitude of input resistance which is several times smaller for the common-base amplifier. This requires the output resistance of the common-collector amplifier to be lower but simplifies the coupling between stages. As shown, direct coupling may be employed without undue biasing difficulties.

Voltage Limitations. A primary consideration in video amplifiers designed for television receivers is the maximum voltage that may be applied to the transistor before breakdown occurs. The output stage may be required to supply a peak-to-peak voltage swing of 100 volts. Fortunately the transistor can be operated so that its collector swings from almost zero to the supply voltage with little distortion in the

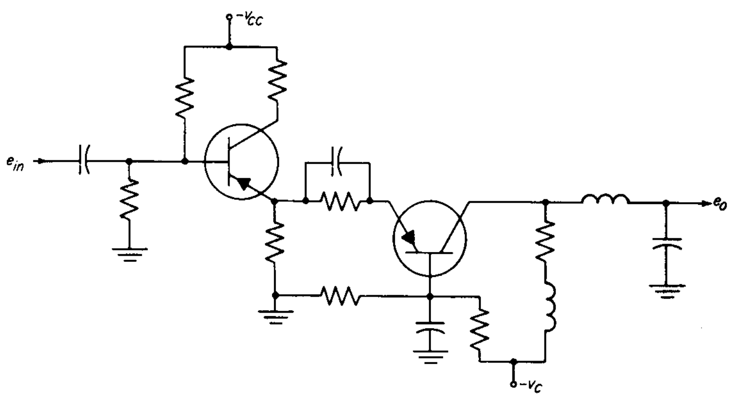


Fig. 20. No polarity reversal producing video amplifier.

output wave. See Fig. 21. Note that the knee of the characteristic curves occurs at extremely low values of collector voltage. Thus, the available peak-to-peak voltage swing practically equals the supply voltage.

When the video-output transistors used have a breakdown voltage rating either close to or lower than the required output swing, special techniques are employed.

Two means of operating the output transistors at lower supply voltages are shown in Fig. 22. In *a* of the figure, the grid and cathode of the kinescope are driven in push-pull from two transistor output stages. The input signal drives the grid negative while the cathode is driven positive. The total grid to cathode voltage is equal to the sum of the outputs of the individual amplifiers. A 100-volt driving signal

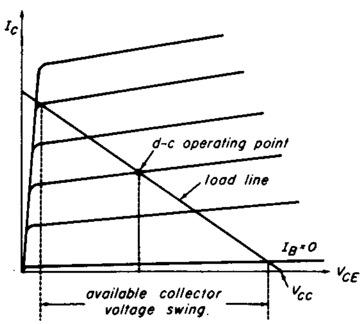


Fig. 21. Since collector voltage can come very close to zero without producing excessive distortion, the peak-to-peak output voltage swing in a transistor video amplifier is nearly equal to the supply voltage.

may then be obtained with only 50 volts from each of the output stages.

In *b* of Fig. 22, two transistor amplifiers are "stacked" in series across the supply voltage. The total supply voltage is shared between transistors. Input signals are applied to the base of Q_1, a common-emitter amplifier. Q_1 is direct coupled to the emitter of Q_2 which functions as a common-base amplifier. A voltage divider (R_1 and R_2) from the collector of Q_2 to ground applies base bias to the common-base amplifier Q_2. The base voltage so provided is chosen to make the base slightly negative with respect to the collector of Q_1, thus providing forward bias for Q_2. The voltage divider also provides negative feedback from the collector of Q_2 and makes it possible for both transistors to share the output voltage swing between them. To understand this action, consider R_1 and R_2 to be returned to a fixed supply voltage. In that case the base of Q_2 remains at a fixed voltage and a voltage swing of only a few tenths of a volt at the collector of Q_1 is sufficient to overdrive Q_2. The circuit behaves like a conventional cascaded amplifier with practically all of the output voltage provided by Q_2.

Now consider the action with R_1 and R_2 returned, as shown, to the collector of Q_2. If the input swings negative, the conduction of Q_1 increases and its collector voltage swings less negative. As the emitter

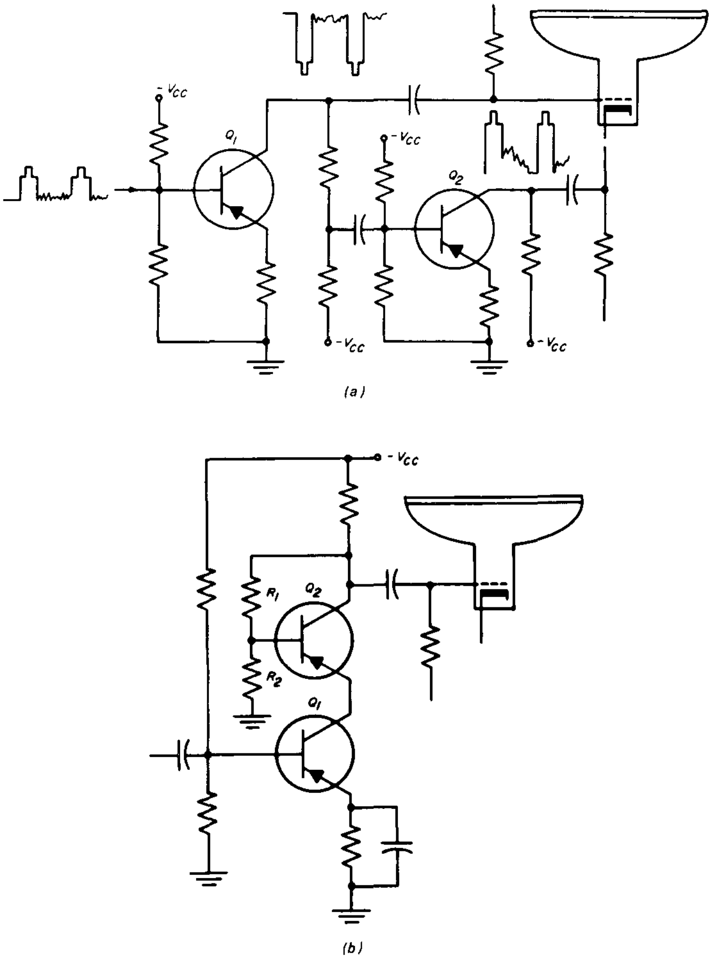


Fig. 22. Two systems that can be used to secure the high peak-to-peak video voltages needed to drive the kinescope.

of Q_2 goes less negative, the forward bias applied to Q_2 increases and it also conducts more heavily. Increased I_C causes the collector voltage, and the output voltage to swing less negative. But the base of Q_2 also swings less negative offsetting the effect of the emitter voltage change. This feedback works to reduce the gain of Q_2 severely, but it accomplishes the end of allowing a greater voltage swing at the collector of Q_1. Thus the total output-voltage swing is the sum of the changes at both collectors and the share of output voltage swing assumed by each transistor is more evenly divided.

4. THE SOUND SECTION

As in vacuum-tube receivers, the sound section consists of a 4.5-mc amplifier, an FM detector, and an audio-amplifier section. Basically the circuits of the i-f and audio sections are similar to the i-f and audio sections of transistor-radio receivers. Therefore, these circuits are not covered in detail, except to point out minor basic differences in design considerations.

The audio section consists of a class-A driver and an output stage. Output stages may use either a single power transistor in class A or a pair of transistors operated class B. The latter is chosen in practically all portable radios because of its low average current demand and hence its low battery drain. In the television receiver the advantage of a class-B audio output is less important as the audio output stage draws only a fraction of the total battery current. In receivers designed to operate from the power line, class-B operation is less desirable. The reason is that the current drawn by the output stage varies with sound level and a rectifier supply with extremely good regulation is required. Thus, in receivers with rectifier supplies, class-A operation is more advantageous. In battery operated portables both class-A and class-B push-pull output stages are found.

One requirement of the television sound i-f amplifier not found in AM radio i-f amplifiers is limiting. Some form of limiting may be required depending upon the amplitude-modulation rejection capabilities of the detector. Transistor limiters can be made with characteristics similar to their vacuum-tube counterparts. For PNP transistors, the positive going half-cycle may be clipped by biasing the stage so that the base-emitter junction becomes reverse biased during positive signal peaks. The negative going half-cycles are limited by placing a large resistor in the collector circuit. The resistor is chosen so that collector voltage falls almost to zero on negative signal peaks. Thus, one half of the input cycle is limited by base-to-emitter cutoff; the other half-cycle is limited by collector saturation. Limiters of this type cannot handle a large range of input signal amplitudes because the d-c operating conditions of the stage change due to signal rectification at the base-to-emitter junction.

FM discriminators and ratio detectors employing semiconductor diodes are found in many vacuum-tube equipped receivers. Operation is the same as in vacuum diode circuits of the same type, except that the semiconductor circuits may operate satisfactorily at a slightly lower signal voltage level.

5. TRANSISTOR SWITCHING CHARACTERISTICS

In most of the circuits discussed so far, the transistor has been operated as a *linear amplifier,* designed to produce as little distortion as possible. In the sync section that follows, the transistor is purposely overdriven in order to achieve sync clipping action. Also, the deflection oscillators and sweep circuits require the transistor to be operated as an electronic switch, just as was shown for vacuum tubes in earlier lessons. Before studying the circuits themselves, we must first examine the transistor when it is overdriven, and find the transistor equivalents of cutoff and plate current saturation or the point when grid current is drawn. With regard to electronic switches we shall find that the transistor is superior in many ways to vacuum tubes.

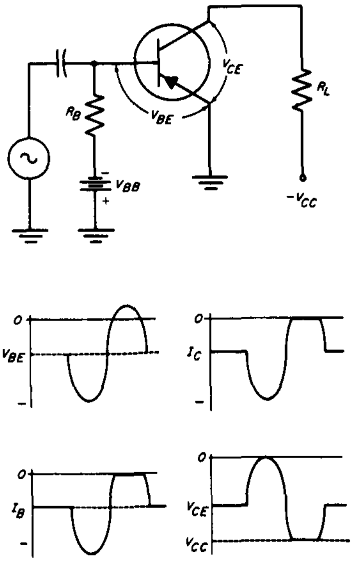


Fig. 23. The transistor amplifier is driven into cutoff when the base-emitter junction becomes reverse biased.

Transistor Cutoff. Suppose that a vacuum tube is overdriven so that the grid signal swings negative enough to exceed cutoff and positive enough to allow grid current to be drawn. This causes both the negative and positive peaks of the output signal to appear clipped. Let's find the equivalent of these conditions for the PNP transistor stage in Fig. 23. With no signal applied, the emitter junction is forward biased by V_{BB} and R_B yielding a no-signal base-bias current of I_B. The *positive excursion* of input signal opposes the bias battery and causes base current to fall. If the positive peak of the input signal should exceed V_{BE}, then the base becomes positive with respect to the emitter, the emitter junction becomes *reverse biased,* and collector current falls to a very low value (leakage current). *This is the cutoff condition.* Since there is a negligible voltage drop across R_L during cutoff, collector voltage is approximately equal to V_{CC}, the collector supply voltage. This causes the negative peak of the output signal to be clipped off as shown. It is interesting to note that the cutoff stage is achieved by removing bias while in vacuum tubes bias must be increased.

Transistor Saturation. Now let's get back to the amplifier circuit and see what factors cause clipping on the alternate half cycles of input signal. Now we shall consider the input voltage to swing negative. In this case the forward bias applied to the emitter junction increases, and the collector current increases. Some factor must limit the maximum amount of current flow. In the vacuum tube the grid may go positive and draw current, limiting further increase in grid voltage, or the tube may reach saturation because all the electrons emitted by the cathode are being used, and no more are available. In the transistor circuit of Fig. 24*a*, saturation occurs when both junctions become *forward biased*. This condition occurs as follows: As the base is made more and more negative, collector current increases and the voltage drop across the load resistance causes the collector voltage to drop. As collector voltage approaches very close to zero, a point is reached where the base is

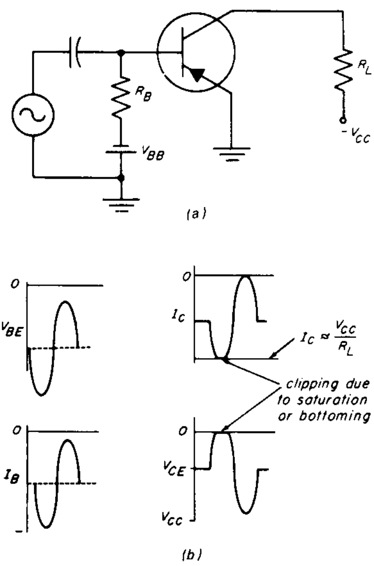


Figure 24. Clipping due to saturation occurs when the voltage drop across the load impedance is nearly equal to the supply voltage.

actually more negative than the collector. Both junctions become forward biased and the transistor acts like a very low resistance. Collector current is approximately equal to the supply voltage divided by the total load resistance. The positive-going peaks of the input sine wave are clipped off at almost zero volts as in *b* of the figure. In this condition the transistor is referred to as being in *saturation*, or in the *bottomed state*.

The Transistor Switch. An ideal switch is characterized by zero resistance in the ON state, infinite resistance in the OFF state, and the ability to change state, from ON to OFF, in zero time. Vacuum-tube switches far surpass mechanical switches in speed of operation, and approach ideal conditions in the OFF state. However, vacuum tubes offer a high series resistance and hence develop a high voltage drop in the ON or conducting state. The transistor, on the other hand, has an emitter-to-collector resistance of only a few ohms to a few tenths of an ohm when driven into saturation, and so has excellent characteristics in the ON state. This advantage is slightly offset by the fact that leakage currents lower the OFF resistance below that of a cutoff vacuum tube. In most applications, however, small leakage currents can be tolerated. Thus, the static switching characteristics of transistors are considered to be superior to those of vacuum tubes.

A basic switching circuit is shown on the graph of Fig. 25. A

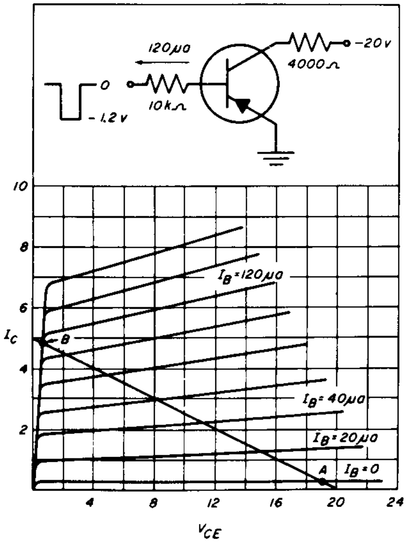


Fig. 25. The amplifier is cut off when base current is zero. A base current of 120 μa produces the saturated state.

current of approximately 5 ma is to be switched into a 4-k ohm load as shown. The ON and OFF conditions may be shown graphically by plotting the load line on the collector static curves. The load line is constructed in the same manner as is employed for vacuum tubes. First, if I_C is taken as zero (transistor collector circuit has infinite resistance), V_{CE} equals the supply voltage and a point is located at $V_{CE} = 20$ volts, $I_C = 0$. Another point is found where V_{CE} equals zero (the transistor is considered a short). Here current is limited only by the load resistance and $I_C = 20$ v/4-k ohms $= 5$ ma. The load line connects these two points as shown.

Before the negative switching pulse is applied, base current is zero and the transistor is cut off. Collector current still flows, however, due to leakage, and the OFF operating point is located at point *A* in the figure. (Note: this leakage current in the absence of ON pulses may be reduced considerably if a reverse bias is applied to the base.)

The transistor is switched to the ON state by applying sufficient base current to move the operating point below the knee of the curve as shown at *B*. Here the 120-microampere base current curve intersects the load line at a point where the characteristic is nearly vertical. The steepness of the curve at this point represents the condition of saturation. Its low resistance is shown by the fact that a very small change in collector voltage results in a large change in collector current in this region.

In the case shown, 120 microamperes is about the minimum amount that will result in saturation. The minimum value may be calculated without the curves if the beta of the transistor is known. For example, if we assume that V_{CE} drops to zero at saturation, then the load current will be 5 ma. If the large-signal beta of the transistor is 45 then the minimum required I_B can be calculated as:

$$I_B = \frac{I_C}{\beta} = \frac{5 \text{ ma}}{45} = 111 \ \mu\text{a}$$

In practice I_B is chosen to be higher than this minimum in order to ensure positive switching despite changes in parameters due to temperature or transistor replacement. The value of I_B also affects switching speed as we shall see.

Switching Speed To apply the transistor in television sync and scanning circuits, it is necessary to know how fast the switch may be turned on or off and what factors affect the speed of operation. The response of the transistor to an abrupt switching pulse is shown in Fig. 26. Before the negative on pulse is applied to the base, a positive bias voltage maintains a reverse bias across the emitter junction. In this condition, the emitter junction capacitance is charged to the bias voltage V_{BB}.

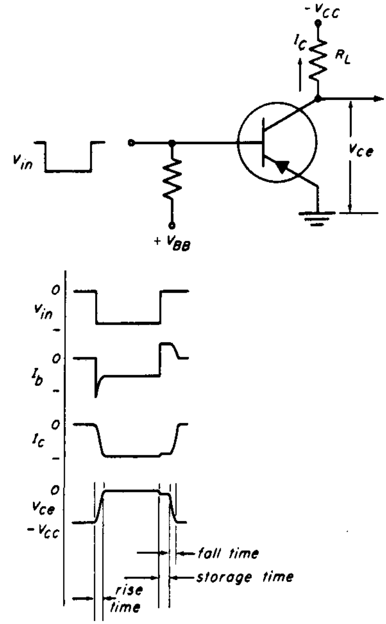


Fig. 26. Switching delays associated with the on-off transition.

When the negative-going *on* pulse arrives, the charged emitter capacitance must be discharged. This accounts for the transient that appears at the leading edge of the base-current waveform. At the first instant the total emitter current flows out through the base, and collector current remains at zero. Collector current then begins to rise and I_C levels off at its maximum value.

Turn-Off Time. When the on pulse is removed the transistor should stop conducting or turn off. Examination of the collector current waveform of Fig. 26 shows, however, that collector current continues to flow for some time at a slightly diminished value, and then finally falls to zero at about the same rate as the rise time. This delay in turning off is called the *storage delay*.

Nonsaturated Operation. In some switching applications speed of operation is of prime importance and the storage delay is avoided entirely by preventing the condition of saturation. This is accomplished by various *clamping circuits* that either prevent I_B or I_C from rising to the saturation value, or prevent the collector voltage from falling to the saturation value. An example is the circuit shown in Fig. 27. As long as the collector voltage is greater than two volts

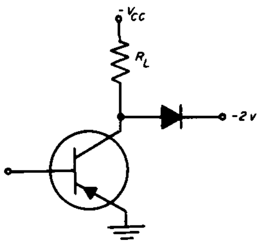


Fig. 27. This diode bias clamp prevents collector voltage from reaching the saturation voltage.

negative, the diode remains non-conducting and does not affect circuit operation. When the transistor is turned on, collector voltage tends to fall to a few tenths of a volt. However, when V_{CE} falls below two volts, the diode conducts, effectively shorting the collector to the two-volt supply. Thus collector voltage cannot fall below two volts (neglecting the voltage drop across the diode).

6. SYNCHRONIZATION CIRCUITS

Transistors offer a degree of flexibility in pulse handling circuits not found in vacuum-tube circuitry. Both collector-current cutoff and collector saturation may be employed in squaring and clipping circuits. Also, the use of both PNP and NPN types in the same circuit permits pulses of either polarity to be handled without the need for phase-inverting circuits.

Three jobs must be performed in the sync circuits: (1) the sync pulses must be separated from the video signal; (2) the separated sync pulses must be squared off and limited to a fixed amplitude; and (3) vertical pulses must be separated from horizontal pulses. In addition, the sync pulse should remain free of video information and noise for a wide range of input signal amplitudes.

The tasks are accomplished in much the same way as they are in vacuum-tube circuits. One- and two-stage systems perform the three jobs in the order stated above. Major differences arise due to transistor circuit input impedances, saturation delays, and the means of providing automatic bias.

Driving considerations. Several points must be considered in coupling the input signal from the transistor video amplifier to the sync separator. The input impedance of the transistor sync clipper is several hundred ohms during periods of active amplification and rises up to over a megohm when cut off. If driven into saturation, the input resistance falls to a few ohms. If this type of load is coupled into a high-impedance point in the video amplifier, the video circuit becomes heavily loaded during sync intervals and almost open circuited during the remainder of the video-signal interval. To minimize video-amplifier loading, a buffer stage may be inserted between the video amplifier and the sync clipper. A common-collector amplifier is a good choice for this purpose as it possesses a very high input impedance and a low output impedance.

An alternative is to obtain driving signals from a low-impedance point in the video amplifier. One low-impedance source is the detector circuit. However, the additional shunt capacitance provided by the emitter capacitance of the sync clipper, even when the clipper is cutoff, results in lowering of the detector's high-frequency response unless the detector's load resistance is also lowered. The latter reduces the efficiency of the detector circuit.

An alternative to a low-impedance source is to couple to a high-impedance point in the video amplifier with a large isolating resistor, as in Fig. 28. The sync separator is isolated from the video amplifier by the resistor R_s, but this severely reduces the rise time of the separator stage as explained earlier. Another drawback is that the full video voltage appears at the base of the clipper stage during the intervals that the clipper is cut off. High-amplitude video signals may be bypassed around the clipper by the collector-base junction capacitance, thereby resulting in incomplete sync separator action.

Practical solutions to the coupling problem are shown in Fig. 29

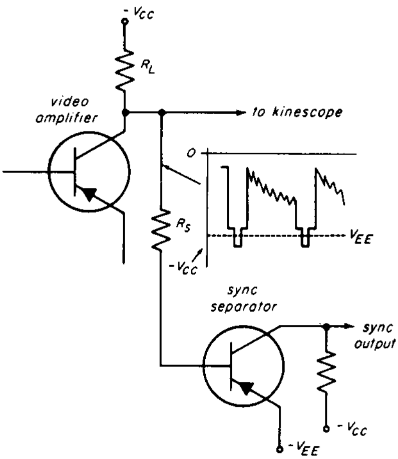


Fig. 28. This coupling from the video amplifier to the sync separator employs a large series resistance to minimize the load on the video amplifier.

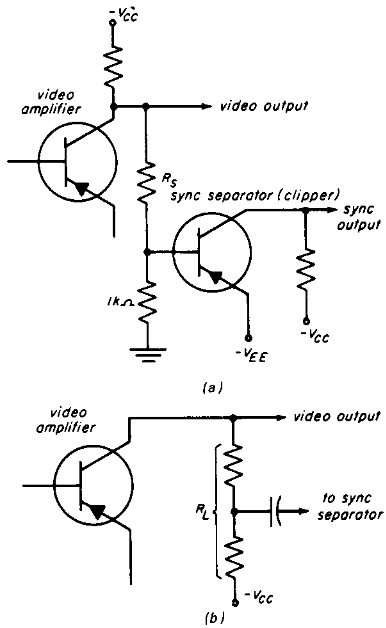


Fig. 29. Practical methods of coupling signals from the video amplifier to the sync separator place a light load on the video-amplifier, but represent a low-impedance signal source to the sync separator.

in *a* of the figure, the isolating resistor is tapped down at a low impedance point to provide a low-resistance driving source for the clipper stage. This system also allows direct coupling to the clipper. The total resistance of the divider must be quite high, however, to prevent the collector voltage of the video amplifier from being lowered. Thus the signal currents flowing in the divider are low and considerable signal loss is incurred. An alternate system is to tap down on the collector load impedance as shown in *b* of the figure. In this case, practically all of the a-c signal current flows into the clipper. Since capacitive coupling is employed, the clipper must employ some means of re-inserting the d-c component of the video signal.

Biasing the Sync Clipper. A simple sync clipper employing fixed bias is shown in Fig. 30. The applied signal is direct-coupled from the video amplifier and extends in the negative direction as shown. The emitter junction of the transistor remains reverse biased until the

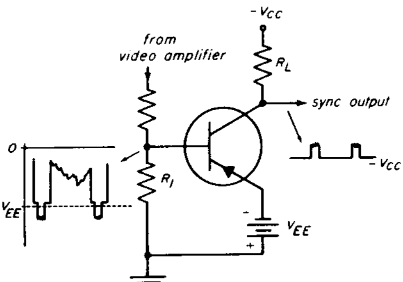


Fig. 30. The basic transistor sync separator. Negative-going sync pulses drive the stage into conduction.

base is made more negative than V_{EE}. The signal causes the base to become negative with respect to the emitter during the sync pulse interval as shown. Collector current flows at this time and collector voltage drops from the V_{CC} value producing a positive-going output pulse.

Fixed bias systems of this type are not practicable in television receivers because peak-signal amplitude is not constant but varies over a wide range due to signal conditions. Also, if a-c coupling is employed, either between stages of the video amplifier or between the video amplifier and the sync circuits, the d-c component of the signal is lost. Sync pulse height then varies due to changes in average brightness. The solution is to replace the fixed bias supply (V_{EE} of Fig. 30) with a voltage supply that adjusts itself automatically to changes in sync pulse height.

Signal bias may be added to the simple sync clipper by replacing the voltage supply, V_{EE}, with an RC network as shown in Fig. 31. When a negative-going signal is applied to this circuit, the emitter junction becomes forward biased. The transistor conducts, and C_1 charges towards the peak of the applied voltage (minus the small voltage drop across the conducting emitter junction). The capacitor charges rapidly through the conducting resistance of the transistor, which, looking in at the emitter terminal, is very low. At the end of

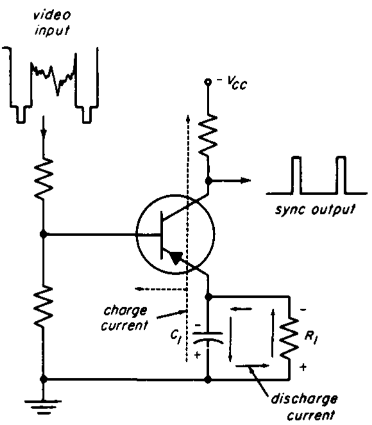


Fig. 31. Automatic bias for the sync separator is provided by the RC network in the emitter circuit.

the sync pulse the base voltage becomes less negative, and C_1 begins to discharge. However, emitter current cannot flow in the opposite direction, and C_1 must discharge through the high resistance of R_1. The time constant of C_1 and R_1 is chosen so that C_1 loses only a fraction of its charge between sync pulses. The emitter voltage then remains at a higher negative potential than the base and the transistor remains cut off until the next sync pulse arrives. When the next sync pulse arrives, a forward bias appears from base to emitter equal to the difference between the peak voltage of the video signal and the charge remaining on C_1. This capacitor then recharges to the peak negative signal voltage, and a pulse of collector current produces the positive-going sync pulse in the collector circuit.

The reverse bias voltage measured at the emitter stays at a fixed percentage of the peak signal. Therefore, the bias voltage automatically adjusts for changes in peak negative voltage, whether the change is due to variations in signal amplitude or loss of the d-c component.

The RC bias network may also be placed in the base circuit as shown in Fig. 32. In this case, C_1 charges through the input resistance of the transistor as seen at the base input terminal. The discharge of C_1 makes the base positive with respect to the emitter holding the transistor cutoff between sync pulses. Circuit operation is the same as the emitter-biased circuit shown earlier, except that the value of C_1 may have to be made smaller to maintain a short time constant in the charging circuit. One possible disadvantage is that collector-junction-leakage current can discharge C_1, and changes in leakage

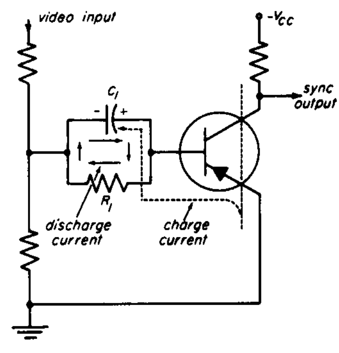


Fig. 32. Bias for the sync separator may also be provided by an RC network in the base circuit.

current with temperature may seriously alter the clipping level. The same problem exists with emitter-junction leakage current (I_{EO}) when the bias network is placed in the emitter leg. I_{EO} is usually smaller than I_{CO}, however.

Although the clipping level remains between the tip of sync and the blanking level, providing separator action, the amplitude of the output pulse is affected by input signal amplitude. To illustrate, suppose that the negative peak signal is one volt and that V_{C1} drops to about 80% of this peak voltage by the time the next pulse arrives. The forward bias applied to the emitter junction during the next sync pulse is 1 volt minus 0.8 volt or 0.2 volt. If the input signal is doubled, then the charge remaining on C_1 between sync pulses is 1.6 volts and the forward bias applied to the emitter junction is 0.4 volt during the conducting interval. Pulse amplitude therefore, is a function of input signal amplitude. Subsequent sync-limiting stages are provided to amplify and limit sync-output pulses to a uniform height.

The Amplifier and Limiter. An additional limiter stage is required following the sync separator, to amplify and square off the sync pulses, limit any noise pulses that may accompany the signal, and provide constant-amplitude output pulses. To insure constant-amplitude output pulses, the stage should be driven between cutoff and saturation. Output pulse height then depends upon circuit parameters alone and is not affected by input signal amplitude. A minimum amplitude driving pulse must be applied to ensure that the stage is driven past these two limits.

Two possibilities present themselves: the limiter stage may be biased to saturate the transistor and the input pulse applied so as to drive the transistor into cutoff. Or, the stage may be biased at cutoff with the input pulse driving the stage into saturation. (With the transistor in cutoff or saturation between sync pulses, noise occurring between pulses does not pass through the stage.) The first choice has a disadvantage because the sync pulse acts to switch the transistor off. Thus the leading edge of the output pulse is delayed by the storage delay. If the second alternative is chosen, the transistor is switched on by the sync pulse and any storage delays follow at the *trailing edge* of the pulse. Since timing accuracy of a triggered oscillator depends upon the leading edge of the sync pulse, the second choice is the logical one.

A negative-going pulse is required to turn on a PNP transistor. Since the output of the sync separator is a positive pulse, some means of inverting the pulse is required. An alternative is to employ an NPN transistor as the limiter. As shown in Fig. 33*a*, this stage may be biased at zero base current or with a slight reverse bias. The positive-going input pulse is of

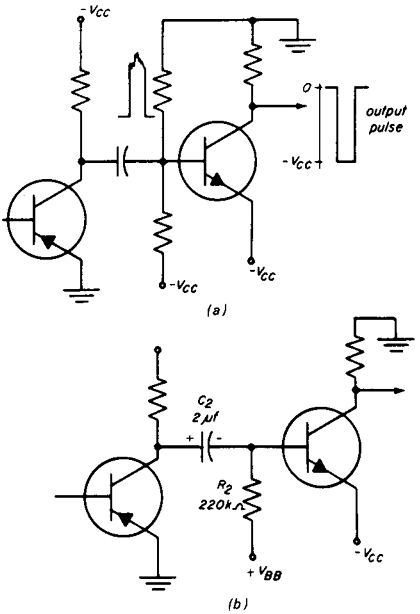


Fig. 33. The sync output stage is driven from cutoff to saturation to produce an output pulse of uniform amplitude.

sufficient height to saturate the transistor and the output waveform swings from the supply voltage during cutoff, to almost zero during saturation. The peak output voltage is almost equal to the collector supply voltage.

This simple clipper has two major drawbacks, however. First, weak signals may not provide an input pulse of sufficient height to saturate the transistor. Decreased limiting action and poor noise immunity would accompany weak signals. Secondly, very strong signals would drive the transistor further into saturation. This increases the storage delay and the turn-off time. The result is slower turn-off and output pulses of longer duration as signal amplitude increases. This effect is sometimes called the "back porch" effect.

The practical solution to the above problems is shown in *b* of Fig. 33. Here a small forward bias is applied to the base through a large resistor. The bias voltage is adjusted to ensure that the weakest usable sync pulses saturate the transistor. C_2 and R_2 act to develop a reverse bias that varies with pulse height. C_2 charges during the sync pulse interval with the polarity shown, and discharges slowly through R_2 between sync pulses. This variable reverse bias holds the stage cut off between sync pulses when normal amplitude signals are received. On strong signals the bias increases and prevents excessive minority-carrier storage in the base region.

Single-Stage Sync Circuits. It is possible to perform both functions of sync separation and sync limiting in a single stage. The sync separator of Fig. 31 may be modified by adjusting the load resistance and the collector supply voltage so that the negative peak of the sync pulse drives the transistor into saturation. The output pulse is then separated from the video signal, amplitude limited, and has a constant amplitude that is practically equal to the collector supply voltage.

Bias considerations are somewhat different in the single-stage separator-clipper. Driving the dual-function single stage into saturation to obtain clipping removes the automatic-biasing feature. When the stage is not permitted to saturate, the current charging C_1 (Fig. 31) is proportional to peak signal amplitude and is higher than the base current by the current gain of the transistor. Bias therefore constantly adjusts for changes in input signal level. When the stage is driven to saturation, however, the charging current, which is largely collector current, is determined mainly by collector voltage and the resistance in the collector circuit. Therefore no automatic bias setting action occurs. This condition is remedied in part by increasing the base current that flows when the transistor is saturated. Greater driving power and lower driving source impedance helps to provide the necessary increase in base current.

Noise Immunity. As in vacuum-tube sync separators, the time constant of the signal-bias circuit plays an important role in determining the noise immunity of the system. A short-time constant-bias network recovers quickly after large peak noise pulses have temporarily blocked or "backed off" the separator. Short time-constant filters work best with horizontal-deflection systems as the afc circuit tends to ignore the loss of a few sync pulses. On the other hand, vertical sync is best served by a long time-constant bias circuit in the separator as the effects of single isolated sync pulses do not cause the deep serrations in the output signal that may trigger the vertical oscillator. One solution uses two separator circuits, one for the horizontal and the other for the vertical systems. Each can then employ the appropriate time constant for its bias network. A more economical solution uses a single separator whose bias is established by two RC networks. The time constants of each network, one short and the other long, are chosen to provide optimum noise immunity for both deflection systems. A single-stage sync separator, employing two RC bias networks, is shown in Fig. 34. This technique does not work out as well in transistor circuits as it does in vacuum tube separators. The reason is that the short time-constant network acts as a high-pass filter and materially reduces the gain of the stage

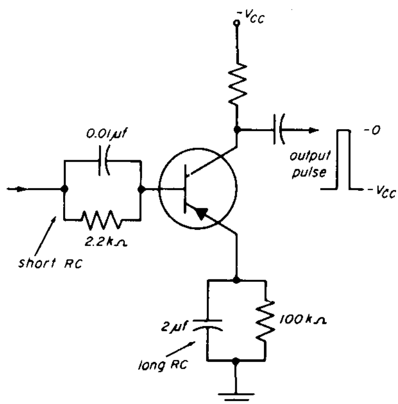


Fig. 34. This single-stage sync circuit performs the functions of the sync separator and sync output stage.

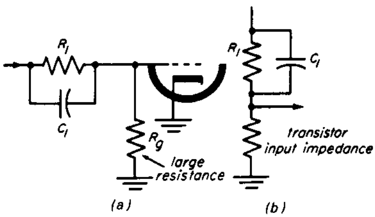


Fig. 35. The short time-constant network forms a voltage divider that attenuates low frequencies severely when connected to the low input impedance of the transistor clipper.

at *low* frequencies. As shown in Fig. 35*a*, this effect does not occur in vacuum-tube circuits because the short time-constant network is in series with the high-input impedance of the tube. When the same RC circuit is placed in series with the low-input impedance of the transistor stage, a voltage divider is formed in which the signal attenuation decreases with frequency, as in *b* of the figure. To minimize the loss at low frequencies, the value of R_1, in the short time-constant filter, is made as small as possible. Compensating filter networks may also be employed to boost low-frequency response before the signal is applied to the separator.

Automatic Time-Constant Control. The circuit shown in Fig. 36 makes use of the economy of the double time-constant bias network but eliminates most of the disadvantages of the compromise necessary when two separators are not used. It does so by switching the long time-constant circuit in and out as needed. During the intervals that the sync pulse is applied, the transistor conducts and a negative voltage appears at the emitter. The voltage places a forward bias on the diode, D_1. The diode conducts, placing the long time-constant circuit R_1C_1 in the emitter leg. This long time-constant network stabilizes the bias and maintains an accurate separation level. At the

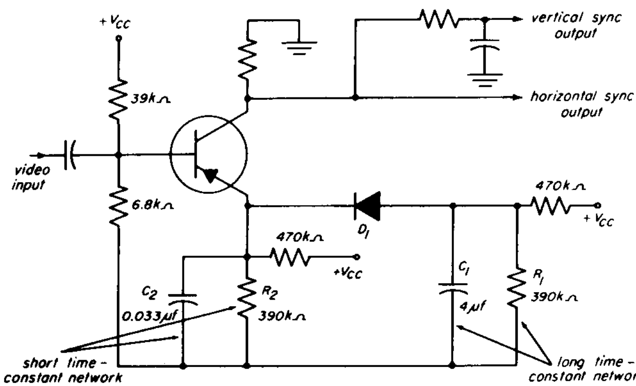


Fig. 36. The diode in this sync separator effectively disconnects the long time-constant network after the separator has been subjected to a large noise-pulse. This action prevents periods of extended blocking.

termination of each sync pulse (or noise pulse) the emitter voltage drops and C_1 begins to discharge. The voltage across C_1 remains more negative than the emitter voltage and the diode becomes reverse biased. Thus the long time-constant network is effectively disconnected at the end of the charging pulse and cannot block or back-off the separator. The short time-constant network, R_2C_2, which had charged in parallel with R_1C_1, discharges rapidly and the separator recovers quickly. The network consisting of C_1 and R_1 effectively absorbs the effects of large noise pulses without causing back-off in the separator.

Noise-Switched Sync Separator. The sync separator shown in Fig. 37 is protected by a noise switch. In the absence of noise pulses, Q_2 conducts due to the negative bias applied at the base. When conducting, Q_2 forms a low-resistance return for the emitter of the sync separator, Q_1, and the separator functions normally. Q_2 is switched off by noise pulses, disabling the sync separator by opening its emitter circuit. Since the separator cannot conduct when Q_2 is switched off, noise pulses cannot develop an excessive back-off bias.

Noise pulses to actuate the switch are obtained from the video detector. Sync and noise pulses are positive-going at the output of the video detector and so are of the correct polarity to cut off Q_2. The diode D_2 serves as a *noise separator*

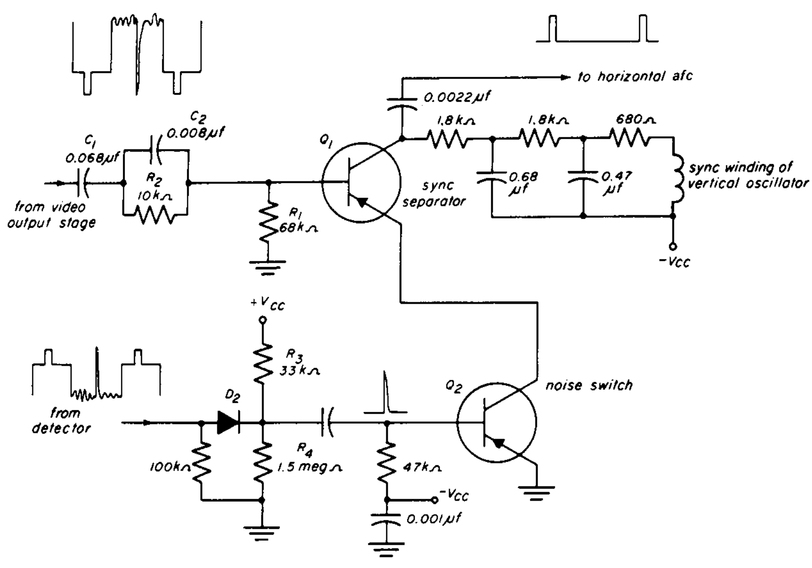


Fig. 37. A sync separator protected by a noise switch. Noise pulses, separated from the video signal by D_2, cut off Q_1 and disable the sync separator.

and conducts only when the peak-noise pulse amplitude exceeds the voltage developed at the voltage divider R_3R_4.

7. VERTICAL-DEFLECTION SYSTEMS

Transistors are well suited to magnetic-deflection systems. Functioning as current amplifiers with low-output impedances as compared to vacuum tubes, they can provide the high peak current required for deflection without the need of matching transformers.

The Output Stage. We shall consider the characteristics of the output stage first so that we may develop the type of driving signal required from the deflection oscillator.

The final product of the vertical-deflection system is a sawtooth of current that rises at a uniform rate in the deflection coils. The peak value of this current depends upon the sensitivity of the yoke (number of turns), the accelerating potential applied to the kinescope, and the deflection angle. For a typical 17-inch kinescope, employing a sensitive (48 mh) commercially available yoke, the required current swing is about 400 ma peak-to-peak. This current swing can be handled easily by transistors in the power category. Since the transistor can supply the yoke current directly, the current step-up transformer neded in vacuum tube circuits may be deleted.

A simplified vertical-output circuit is shown in Fig. 38. The common-emitter circuit is chosen because it provides the greatest power gain and so reduces the a-c power demanded of the driving circuits. Since the load on the transistor is largely resistive at the vertical-scanning frequency, we may treat the stage as a conventional class-A current amplifier. A linear sawtooth of collector current produces

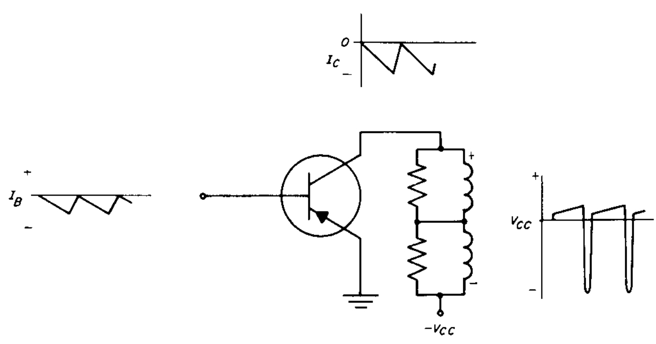


Fig. 38. The PNP sweep output stage requires a negative-going sawtooth so that the retrace pulse at the collector does not cause the collector junction to become forward biased.

the desired deflection current. The voltage drop across the yoke during the trace is as shown. During the short retrace interval collector current drops back towards zero, and the rapid collapse of the deflection field induces a high pulse of voltage to appear across the coils. The polarity of coil voltage changes during retrace causing this pulse to appear in the coil voltage waveform as a negative spike. Note that the voltage produced across the coil has a trapezoidal shape due to this spike.

Yoke Damping. The input waveform need not be peaked (trapezoid shaped). Peaking is required in some vacuum-tube circuits so that the grid voltage goes quite negative during retrace. This prevents the positive pulse at the plate from bringing the tube out of cutoff and causing excessive damping. In the transistor circuit, however, the high reverse pulse developed in the collector circuit has little influence upon base-to-emitter bias and so cannot bring the transistor back into conduction. Some damping is provided by leakage current, but most of the required damping is provided by damping resistors.

Other factors that should be noted are the polarity of the driving signal and the maximum voltage the transistor can withstand across the emitter and collector terminals before breakdown occurs. As regards the polarity of the driving signal, it is seen that thc negative-going sawtooth applied to the base of the transistor in Fig. 38 results in a negative spike of collector voltage during retrace. If the polarity of the driving signal is reversed to provide a sawtooth that rises in a positive direction, then a high positive pulse of voltage is developed across the yoke during retrace. A high positive pulse forward biases the collector junction, placing almost a short circuit across the yoke. Excessive damping results, causing a very slow retrace. Therefore, PNP output stages require negative-going sawtooth driving signals. NPN output stages require positive-going signals for the same reason.

The transistor also behaves like a short circuit if the voltage applied between collector and emitter is sufficient to cause breakdown. An avalanche effect takes place in the reverse-biased collector junction, similar to the firing of a gas tube, and collector current rises to a value determined only by collector voltage and the external resistance in the collector circuit. If the high retrace pulse should exceed the breakdown voltage, the yoke becomes very heavily damped and the completion of retrace is delayed. To keep the retrace pulse below the breakdown rating, the inductance of the deflection yoke must be limited. The pulse height is proportional to the inductance of the coils and to the rate of change of current during retrace. In addition to excessive damping caused by operating the transistor past its breakdown rating, permanent damage to the transistor is likely.

Yoke Coupling. In the simple deflection circuit of Fig. 38 the deflection-coil current varies from zero to its maximum value during the trace period but does not reverse direction. As a result, all deflection is entirely above or below the center of the screen. In addition, excessive power is dissipated in the yoke windings due to the d-c component of collector current. Recentering may be accomplished by means of strong permanent magnets but the problem of power dissipation in the yoke remains. One solution is to use an isolation transformer with a one-to-one turns ratio to couple to the yoke as shown in Fig. 39*a*. This eliminates the d-c component and provides a current sawtooth varying equally above and below zero. A transformer, however, produces distortion due to the onset of magnetic saturation in the core.

A solution is shown in Fig. 39*b*. Here a very high-inductance, low-resistance choke is placed across the deflection coils. To the a-c sawtooth, the choke appears as an open circuit and the a-c component of the signal flows in the deflection coil. The d-c component sees two parallel paths and most of the d-c current flows through the low resistance choke. The d-c component of deflection-coil current cannot be eliminated entirely in this fashion using a choke of practical size. D-c centering magnets can make the necessary small shift in centering, however, and power dissipated in the yoke is limited to a-c power. D-c deflection coil current is eliminated entirely by adding a coupling capacitor, as shown in *c* of Fig. 39. However, a very large capacitor of 1000 μf or more is required.

Providing the Sawtooth. A sawtooth of voltage is generated in vacuum-tube circuits by permitting a capacitor to charge to a fraction of the applied voltage. The voltage developed across the capacitor is the signal driving the output stage. This system represents no problems

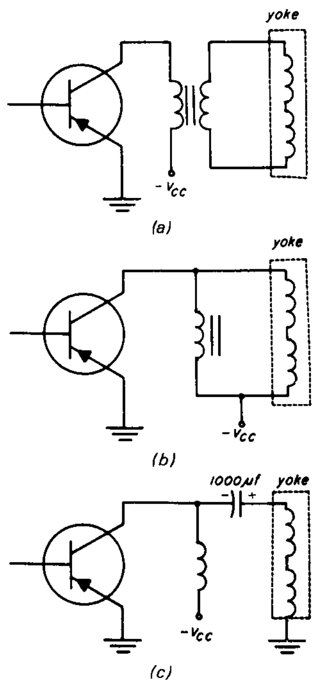


Fig. 39. Methods of coupling the vertical-output stage to the yoke.

when the load on the capacitor is the very high input impedance of the vacuum tube. But in the circuits to be considered the load is a few hundred ohms at best and draws a considerable amount of current. This load on the sweep capacitor lowers the sweep voltage and destroys sweep linearity.

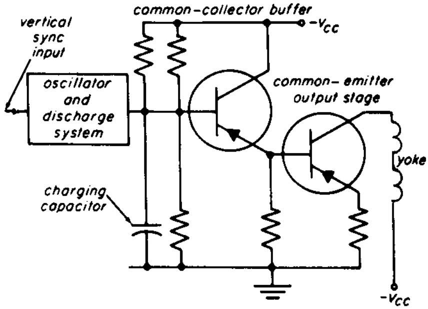


Fig. 40. The common-collector buffer amplifier has a high input impedance and places a small current load on the sawtooth capacitor.

Buffer Amplifiers. A solution to the problem of loading the sweep capacitor is an additional amplifier to isolate the input impedance of the deflection amplifier from the sweep-generating network. An amplifier used for this purpose is called a buffer. The common-collector amplifier, sometimes called an emitter follower, is ideally suited for this purpose. It has a high-input impedance, a low-output impedance and provides some power gain. Thus it does not load the capacitor across which the voltage sawtooth is developed, and provides the low-impedance voltage source required by the output stage. The basic circuit arrangement of the emitter-follower buffer is shown in Fig. 40. The stage can be directly coupled to the output amplifier, dispensing with the need for a very large coupling capacitor.

Common-emitter amplifiers are also employed as buffers. However, they present a low-input impedance and must be modified to prevent loading of the sweep capacitor. A series resistance, R_s in Fig. 41, serves to raise the amplifier's input impedance. The high gain of the common-emitter circuit more

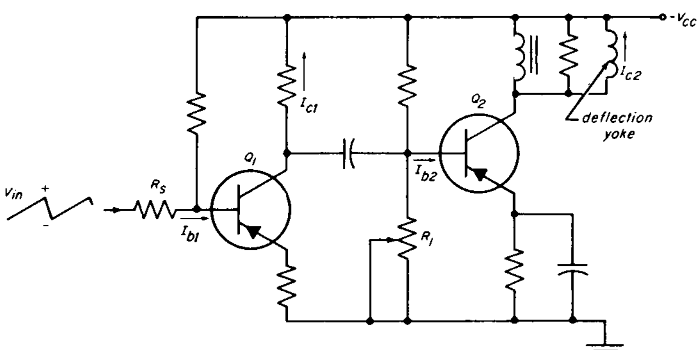


Fig. 41. The common-emitter buffer amplifier uses a large series resistance to minimize loading of the sawtooth capacitor.

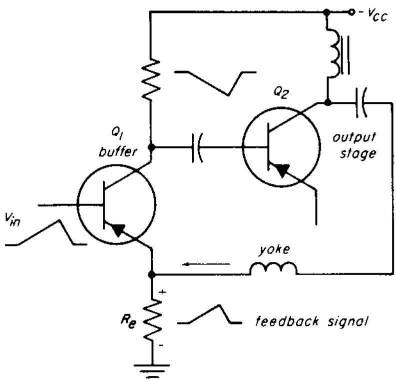


Fig. 42. A feedback signal, proportioned to yoke current, is obtained by passing the yoke current through the low emitter resistance of the buffer stage. The feedback is negative and minimizes distortion.

than makes up for the loss in signal incurred in this resistor. Feedback systems such as an unbypassed emitter resistor also serve to raise the input impedance of the buffer amplifier.

Linearity may also be improved by means of a negative-feedback loop around both stages. In one system a feedback signal having the waveshape of the yoke current is applied to the emitter of the buffer stage as in Fig. 42. The yoke current feedback signal is obtained by allowing the yoke current to flow through a low value emitter resistor in the buffer stage. The polarity of the feedback signal is as shown. A positive-going sawtooth input signal causes a negative-going sawtooth to appear at the collector of the buffer stage. This signal causes yoke current to increase as the sweep progresses. Yoke current flowing through R_e causes the emitter voltage to increase in a positive direction. Note that the input signal is acting to reduce forward bias on the buffer stage while the feedback signal acts to increase it. The total signal acting upon Q_1 is therefore the input signal minus the yoke-current signal and feedback is degenerative.

Deflection Oscillators. The buffer amplifier allows a conventional RC charging circuit to be used in generating the sweep waveform. The next stage is an electronic switch (an oscillator and discharge circuit) to allow the sweep capacitor to charge and discharge at the proper rates. Transistorized versions of both multivibrator and blocking oscillators are used. The blocking oscillator is usually chosen because it is far more stable with regards to frequency. Multivibrators are frequency-sensitive to small changes in supply voltage.

One form of a transistor blocking oscillator is shown in Fig. 43. This circuit operates very much like its vacuum-tube counterpart. One unique feature is that a single RC network, R_s and C_s, performs both the functions of the frequency-determining network and the sweep-generating network.

Circuit operation is as follows: At the instant the supply voltage is applied, a negative voltage appears at the base due to the voltage divider R_1 and R_2. C_s is uncharged

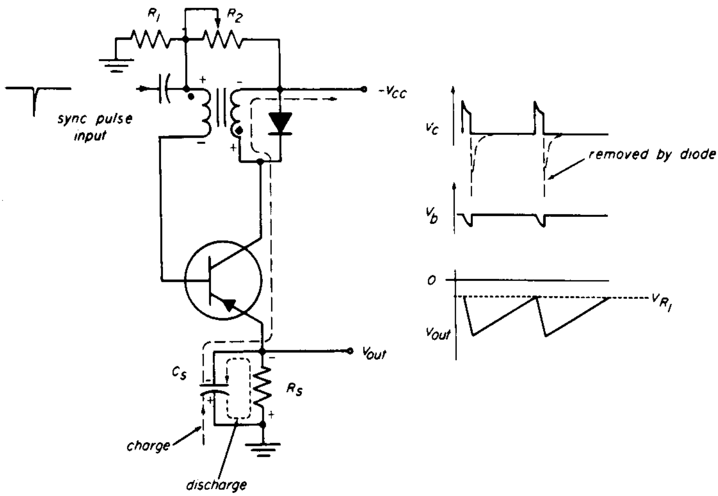


Fig. 43. A transistor blocking oscillator. The RC network in the emitter leg determines repetition rate and provides the output sawtooth.

so that the emitter is at ground potential and the entire voltage developed across R_1 appears as forward bias between base and emitter. The transistor conducts and C_s begins to charge to the base voltage. As soon as collector current increases in the primary of the transformer, a voltage is induced into the secondary with the polarity shown. The base is driven to a higher negative voltage, transistor conduction increases and C_s charges very rapidly. The regenerative cycle in which C_s charges to a high negative voltage continues until transistor saturation causes collector current to level off at the saturation value. As there is no voltage induced into the secondary at the instant collector current stops rising, the base voltage drops towards the voltage developed across R_1. The emitter is then at a higher negative potential than the base due to the charge on C_s, and the transistor cuts off. The field about the transformer collapses rapidly inducing a high-voltage pulse in both windings which is opposite in polarity to that shown in the figure. The collector transient dies out quickly, but the transistor remains cut off until the voltage across C_s drops to the voltage developed across R_1. When this occurs the base becomes forward-biased, the transistor conducts, and the regenerative charging cycle begins again. The relatively slow discharge of C_s through R_s serves both to develop the sawtooth and establish the trace interval.

Synchronization is obtained by applying a negative-going sync pulse to the base as shown. Just as in vacuum-tube blocking oscillators, the natural repetition rate of the oscillator must be slightly lower than the sync frequency in order to achieve synchronization. The sync pulse acts to bring the transistor into conduction before the discharge of C_s brings about conduction naturally.

The free-running frequency of the oscillator is controlled mainly by the R_sC_s network in the emitter leg. Note that during the discharge cycle, the transistor is cut off and the C_sR_s network is *isolated* from the rest of the circuitry. This helps to provide good frequency stability. Frequency may be changed over a small range by altering the d-c bias applied at the base. The manual frequency or hold control R_2 is placed in the bias network for this purpose. An increase in forward bias applied to the base allows the transistor to come into conduction sooner and so raises the frequency.

The voltage transient appearing at the collector and base terminals at the beginning of the cutoff or trace cycle can be troublesome, particularly if it exceeds the breakdown potential of the transistor. The transient is removed by placing a diode across the primary winding as shown. The diode conducts when the transformer polarities reverse from those that are shown in the figure. This permits the coil to discharge slowly and harmlessly through the diode.

The sweep circuit shown in Fig. 43 requires a phase inverter to drive the common-collector buffer/output stage shown earlier. A negative-going sawtooth may be obtained directly from the circuit shown in Figure 44. This circuit is somewhat like the vacuum-tube blocking oscillator in that the sweep capacitor charges during the trace period and discharges during retrace. Circuit operation is as follows: At the time that V_{CC} is applied, the charge on the sweep capacitor C_s is zero, and the base voltage, which is obtained from the voltage divider across C_s, is also zero. The transistor is cut off by the negative voltage that is developed across R_3 and applied to the emitter. The trace begins as C_s charges through R_s towards V_{CC}. As the voltage across C_s rises, a

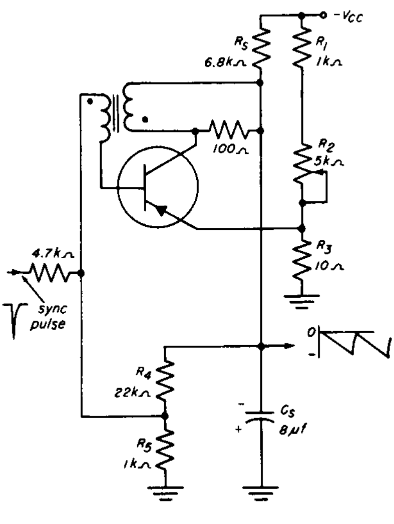


Fig. 44. The blocking oscillator provides the negative-going sawtooth required by the PNP output stage.

smaller voltage obtained from the divider R_4R_5 causes the base voltage to rise. The sweep is terminated when the negative voltage developed across R_5 just exceeds the emitter voltage. Transistor conduction begins at this instant and transformer action causes the base voltage to swing highly negative, driving the transistor towards saturation. C_s discharges rapidly through the transistor and the 10-ohm emitter resistor. At the end of this rapid regenerative cycle C_s has discharged almost to zero, the transistor is cut off again, and the next trace interval begins.

Frequency control is again achieved by adjusting the amount of d-c base-to-emitter voltage. In this case, an increase in negative voltage applied at the emitter increases the time during which the transistor is held at cutoff, and the sweep frequency decreases. A 100-ohm damping resistor across the transformer primary reduces the retrace pulse to safe limits.

Linearity Controls. Good linearity is obtained in the buffer and output stages by careful matching of transfer characteristics and the use of negative feedback. However, some control of waveshape must be provided to adjust for changes in transistor characteristics and nonlinearity in the sawtooth provided by the deflection oscillator. Two types of linearity control are employed. One is somewhat similar to the linearity control in vacuum-tube output circuits in that its serves to vary the d-c operating point of one of the amplifiers (usually the output stage). By shifting the d-c operating point the sector of the transfer characteristic included in the base-current swing is altered. Resistor R_1 in Fig. 41 is an example of this type of control. It also serves to allow for differences in transistor characteristics between replacement units.

Another form of linearity control uses a waveshaping network in the signal path or in a feedback network. This type of control must be used when the common-collector buffer circuit is employed. The reason is that the nonlinearity in the input sawtooth cannot be corrected by utilizing the nonlinearity of the buffer and output circuits. An example of a waveshaping network in a feedback loop is shown in Fig. 45. The yoke voltage waveform is applied to a differentiating network consisting of C_f and R_f. A differentiated sawtooth is developed across R_f and applied through R_g to the input of the buffer stage. The retrace transient is attenuated in the feedback signal by the sweep capacitor C_s so that only the low-frequency components of the feedback signal, occurring during the trace interval, are effective. Note that the differentiated sawtooth, when subtracted from the nonlinear input signal, acts to produce a linear sawtooth at the base of Q_1. Two linearity controls are available. Resistor R_f determines the degree of differentiation by adjusting the time constant of the circuit; resistor R_g

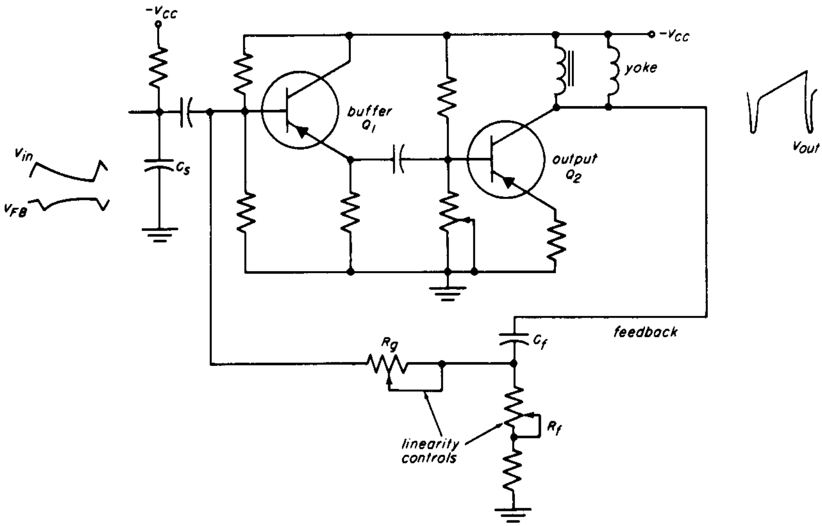


Fig. 45. A vertical-output stage equipped with feedback linearity correction.

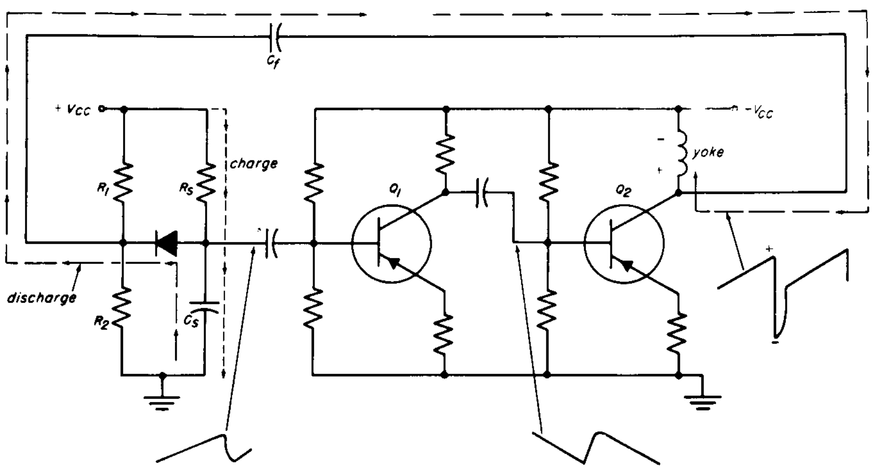


Fig. 46. A complete vertical-deflection system in which a large sweep capacitor is discharged by a diode.

adjusts the amount of feedback signal.

Two-Transistor Multivibrator Circuit. The trend in vacuum-tube vertical-deflection circuits has long been away from separate oscillator and output-stage circuits to a combined transformerless circuit in which positive feedback is obtained from the output stage. The same principles may be applied to the transistorized deflection system. An example is the circuit shown simplified in Fig. 46. This circuit employs the common-emitter arrangement for both buffer and output stages. The electronic switch used to charge and discharge the sweep capacitor utilizes a diode and feedback from the output stage. At the instant power is applied, the voltage across the sweep capacitor, C_s, is zero. The diode is connected between C_s and a positive voltage tapped off at the junction of the voltage divider consisting of R_1 and R_2. As the diode is reverse biased in this condition, its high impedance isolates the sweep capacitor from R_1 and R_2, and from the feedback system including C_f. Thus C_s charges through R_s to the applied positive voltage, and as long as the diode remains reverse biased, the sweep circuit is not connected to the feedback network. This charging cycle constitutes the linear trace period. The positive-going sawtooth is applied to the buffer where it is inverted and appears as a negative-going sawtooth at the base of the output stage. Collector current in the output stage rises linearly, causing a voltage drop to appear across the yoke with the polarity shown.

The trace interval is terminated when the charge on C_s just exceeds the voltage developed across R_2. This brings the diode into conduction, preventing any further increase in voltage across C_s and connecting the sweep capacitor to the feedback network. Since the sawtooth has stopped rising the voltage drop across the yoke drops rapidly towards zero. A regenerative discharge cycle begins in which C_s discharges through the diode, C_f, and the yoke. The collapse of the field about the yoke causes the expected high negative pulse to appear at the collector of V_2 which causes C_s to discharge very rapidly. The discharge is oscillatory involving the resonant circuit formed by the yoke, C_f and C_s. When C_s discharges to zero, the yoke voltage begins to swing positive and the diode cuts off, disconnecting the feedback loop, and allowing C_s to charge as before.

A sawtooth of charging voltage in this case may be obtained without the need for a buffer stage having a high input impedance. A large capacitor may be employed, with a correspondingly small charging resistor, so that the charging current is much larger than the base current of the driver stage. The high dischage current that results from this system presents no problem in this circuit because it is handled by the diode circuit, and does

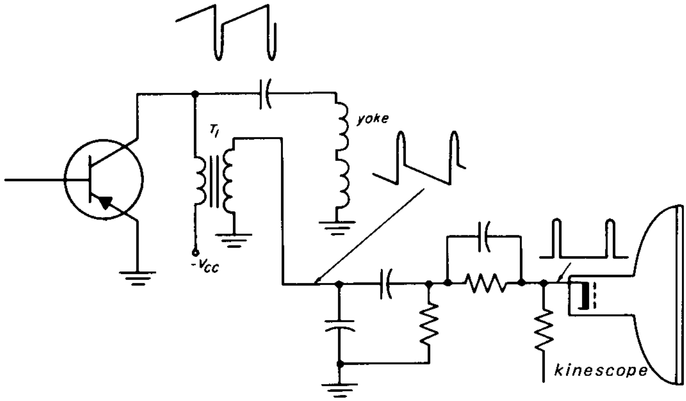


Fig. 47. The transformer serves as a coupling choke. The secondary winding provides retrace blanking pulses.

not require a transistor with a high-current rating.

Retrace Blanking. The vertical-output circuit is also called upon to provide a retrace pulse to the kinescope control circuits for blanking purposes. A blanking pulse of the desired amplitude and polarity may be obtained from a secondary winding wound upon the coupling choke as shown in Fig. 47. This transformer is sometimes called a "vertical output transformer" but the name is not correct as it does not couple the sweep signal to the yoke by transformer action. The RC network in the signal path corrects the sawtooth component of the signal, so that only positive pulses are applied to the cathode of the kinescope.

8. HORIZONTAL-DEFLECTION SYSTEMS

The horizontal-output circuit permits maximum advantage to be taken of the unique switching characteristics of the transistor. Transistor horizontal-output stages can be very efficient, and involve relatively little power dissipation in the transistor itself.

An Ideal Switch. We have seen how the transistor, driven between cutoff and saturation, comes closer to the characteristics of an ideal switch than the vacuum tube. This is particularly true in the conducting state. Therefore, let us examine the possibilities of a deflection system equipped with an ideal switch and then replace it with a transistor, noting the changes and sacrifices that must be made.

At the horizontal scanning rate, the inductive reactance of the deflection yoke is very high compared with its resistance. Thus the yoke appears reactive. If the series resistance added by the switch is also very small, the yoke may be considered as if it were a pure inductance. The basic deflection circuit appears in Fig. 48. Now, in order to produce a sawtooth of current in a pure inductance, a constant voltage is required across the coil. This may be understood if we consider a rising, linear sawtooth of current in the coil. The rate of field expansion in the coil is constant, in this case, and hence the cemf induced into the coil windings is also constant. Thus a linear sawtooth of current is obtained by connecting the coil to a constant source of voltage, by means of an ideal switch.

In the circuit shown in Fig. 48 coil current is started flowing downwards in the coil by closing the switch. If the circuit resistance is zero, coil current continues to rise in a linear fashion, indefinitely. In the practical circuit, the switch is held in the closed position until one half of the trace is completed and then it is opened. When the switch

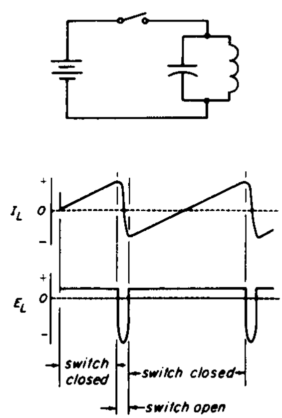


Fig. 48. In this lossless system a linear sawtooth of coil current is produced using a simple switch and a voltage source.

opens, the coil current continues flowing in the same direction but flows into the capacitor. Current collapses following a sine curve whose period is determined by the LC product. The switch remains open so that one half cycle of oscillation may be completed. During the interval coil current falls to zero as the capacitor reaches maximum charge. Coil current then builds up in the opposite direction and is at its peak value (flowing upwards) when the switch is closed again. The coil then discharges slowly and linearly through the switch and the battery. In this lossless system no power is consumed.

Notice that the switch is required to pass current in *both* directions (a bilateral switch). There is no vacuum tube counterpart for this type of switch as current can only flow in one direction through a tube. However, a transistor held in the saturated state, is a bilateral switch, and will pass current in both directions. Consider a transistor held in saturation with current flowing from emitter to collector. If the collector-to-emitter voltage, V_{CE}, is reversed, the roles of emitter and collector also reverse, but the high current-saturation state remains. Thus, a single transistor, operated between saturation and cutoff can replace the ideal switch shown in Figure 48.

Drive Requirements. The driving requirements for the transistor switch are quite simple. All that is required is a signal to hold the transistor in saturation during the

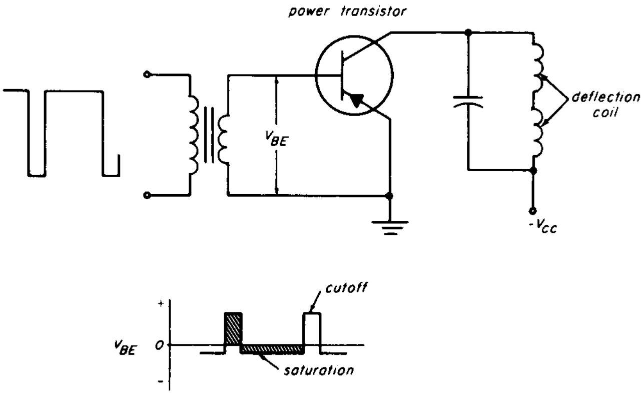


Fig. 49. A transistorized version of the simple switching circuit of Figure 48.

trace period and drive the transistor into cutoff during retrace. This can be accomplished by a d-c bias system that normally holds the transistor in saturation and a pulse signal to cut the transistor off during the retrace. A simpler system utilizes a pulse signal that is transformer coupled to the base-emitter circuit as shown in Fig. 49. The action of the transformer is such that the pulse waveform resolves itself about the zero axis, at the secondary of the transformer, so that the area of the signal above zero equals that below it. Thus, the signal is highly positive during retrace and is negative enough during the remainder of the cycle to maintain the saturated condition. The collector current waveform rises linearly prior to the arrival of the cutoff pulse, as shown in Fig. 50. When the cutoff pulse is applied, collector current falls to zero at a rate determined by the cutoff delay of the transistor. Yoke current continues to flow as the transistor cuts off and charges the circuit capacitance. The voltage across the coil reaches its peak as coil current is passing through zero, yielding the voltage waveform as shown. At the end of

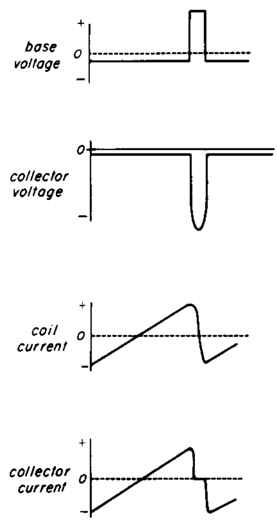


Fig. 50. Waveforms in the simple horizontal-deflection circuit of Figure 49.

the cutoff pulse, transistor conduction resumes, but current now flows in the opposite direction until the coil current discharges to zero. Note that the coil current swing below the line has a lower peak amplitude than that above the line. This accounts for the energy losses in the system during the retrace phase. A centering shift also occurs due to this unbalance in positive and negative coil currents but it is easily corrected by conventional centering systems.

Voltage and Current Requirements. The peak-to-peak current to be handled by the transistor switch is determined by the peak power requirements of the deflection system and the maximum reverse voltage the transistor can withstand during the retrace interval. Peak power requirements are the same as those found in vacuum-tube circuits and have been well established. A typical 90-degree deflection system with a standard yoke and accelerating potential requires a peak power of about 1200 volt-amperes. The reverse voltage is determined by the *rate of change of current* in the yoke at the center of retrace and the yoke inductance. Yoke inductance can be calculated for a given retrace time to present a retrace pulse that does not exceed the breakdown potential of the transistor. If the transistor can withstand 100 volts and the yoke inductance is chosen to yield this peak value, then the transistor must handle a peak-to-peak current of 1200 VA/100 volts or 12 amperes. Coil inductances are higher, and current peaks lowered for transistors having higher breakdown potentials. Silicon transistors, having high-breakdown potentials, are used extensively for the horizontal-output stage.

The reverse voltage peak may be reduced by adding a tuned circuit resonant at the third harmonic of the yoke circuit, as in Fig. 51. This circuit is shock excited by the collector signal, and develops a third-harmonic signal. The addition of the third-harmonic signal when the

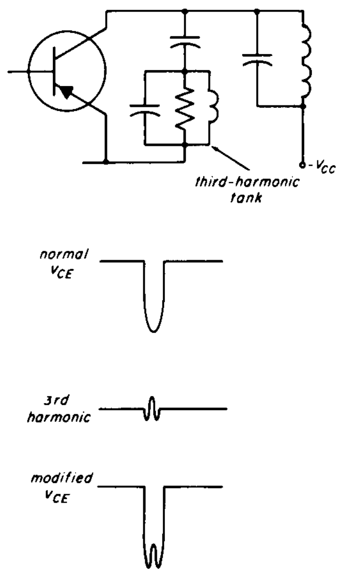


Fig. 51. The retrace pulse at the collector of the horizontal-output stage is reduced in amplitude by adding a circuit tuned to the third harmonic of the natural resonant frequency of the deflection coils.

added circuit rings results in the dip at the voltage peak as shown. In actual practice, the added circuit also forms part of the high-voltage system.

The Driver Stage. The negative portion of the input signal provides sufficient base current to maintain saturation throughout the trace interval. The minimum amount of base current is determined by the forward peak of current (at the end of the trace) and the current gain of the transistor. A 7-ampere forward peak and a transistor with a beta of 70 therefore requires a base current of 7 amperes/70 or 100 ma.

A driver stage is required to supply the power needed to cut off the output stage rapidly and to act as a buffer for the deflection oscillator. The driver circuit, shown in Fig. 52, supplies a pulse type of waveform just like that obtained from the output stage. The signal from the oscillator holds the driver transistor in conduction during the scan interval and cuts the driver off during retrace. A flyback pulse appears at the coupling-transformer primary as the field collapses and charges the total shunt capacitance. Here again the transformer is permitted to ring for one-half cycle producing the pulse needed to cut off the output stage. The transformer has a step-down ratio to match the low-impedance input to the output stage and provide the correct amplitude of cutoff pulse. Forward bias for the driver stage is obtained in the same fashion as that shown for the output stage. In addition the d-c operating point may be altered somewhat by adjustment of R_1. R_1 and C_1 form a signal bias network that develops a small amount of reverse bias. This bias voltage works against the forward bias resulting from the a-c signal averaging itself

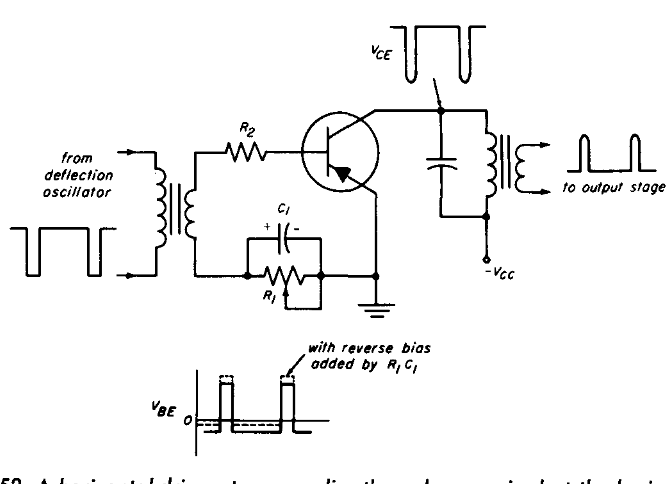


Fig. 52. A horizontal-driver stage supplies the pulses required at the horizontal-output stage.

about zero, as shown earlier. Thus R_I can be adjusted for optimum bias and the storage delay associated with too much forward bias during scan can be reduced. The series resistor R_2 is sometimes used to limit the base current during saturation.

Deflection Oscillators. The deflection oscillator can be a blocking oscillator similar to those shown in the section dealing with vertical deflection. Triggered sync is unsatisfactory, however, and automatic control of oscillator repetition rate is required. This may be accomplished by controlling the d-c base bias of the oscillator. A phase-sensitive detector comparing incoming sync pulses with the scan signal provides the d-c correction signal.

A practical automatic frequency-controlled blocking oscillator is shown in Fig. 53. The oscillator is a blocking oscillator employing a transformer to secure positive feedback. It conducts during the short retrace interval, charging C_6 and remains cut off during the trace period while C_6 discharges through R_7. The time constant of the R_7C_6 network in the emitter leg controls the trace period and hence oscillator frequency. Control over a

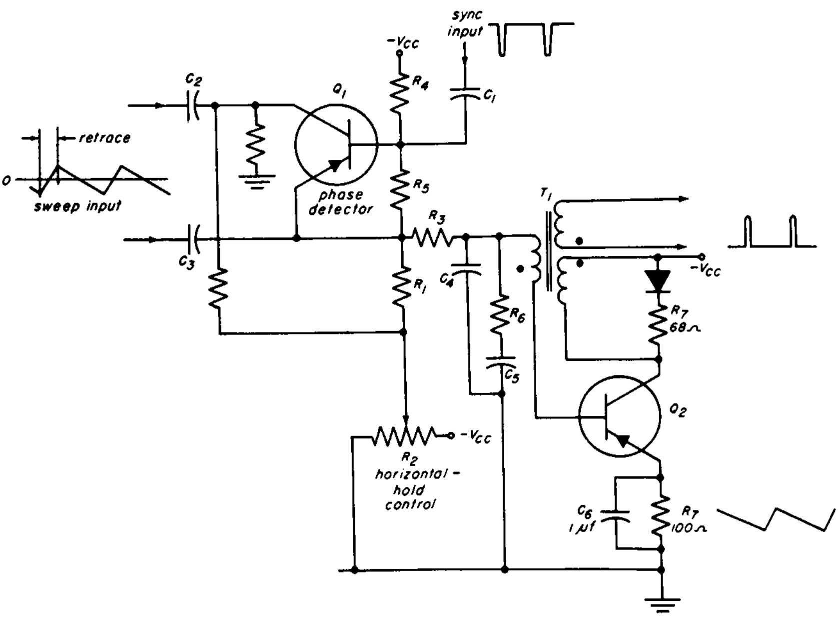


Fig. 53. A transistorized horizontal afc system. A transistor phase discriminator controls the blocking oscillator.

narrow range of frequencies is obtained by varying the base-bias voltage. The retrace interval begins when C_6 discharges to the voltage applied to the base so that varying the base voltage controls the duration of the trace interval. This bias is manually controlled by potentiometer R_2 which is the hold control. In addition to the voltage picked off at the arm of R_2, an additonal control voltage, obtained from the afc phase detector Q_1, is developed across R_1. This voltage adds to or subtracts from the fixed bias voltage to automatically correct the oscillator's repetition rate.

The control voltage developed across R_1 can be either positive or negative. The phase detector, Q_1, is a PNP transistor having nearly identical emitter and collector regions, so that it provides the same characteristics when the roles of collector and emitter are reversed. Negative-going sync pulses are applied at the base and act to bring the transistor into conduction. C_1 is charged by base current at this time and the discharge of C_1 holds the transistor cut off between sync pulses. Now the collector-to-emitter voltage is a sawtooth obtained from the output stage and RC coupled between the emitter and the collector terminals. As shown, the collector-to-emitter is negative during the first half of retrace and the transistor acts normally. During the latter half of retrace the collector is positive and acts as the emitter. If the oscillator is phased properly in the free-running state the center of the sync pulse just straddles the point where the sweep signal passes through zero during the retrace interval. Under this condition, during the time that the sync pulse holds the transistor in conduction, current flows first from emitter to collector and then reverses and flows from collector to emitter. The peak amplitude of current in either direction is equal, and the net correction voltage developed across R_1 is zero. Suppose now the oscillator period increases. The sync pulse then occurs early, compared to retrace. The sync pulse then arrives when the collector is negative and transistor conduction increases in the normal direction. The result is a voltage across R_1 that is negative at the top. This makes the base of the oscillator more negative and tends to bring the oscillator into conduction sooner. Thus the period of the oscillator is reduced, giving the required frequency correction. Note that the correction voltage is applied through a low-pass filter consisting of C_4, R_6, and C_5. These components serve the same purpose as in vacuum-tube circuits. They reduce the loop gain of the system at high frequencies to prevent oscillation, while permitting the system to operate at high gain at low frequencies. The latter helps to increase the phase accuracy of the system.

Practical Output Systems. The horizontal-output stage must provide facilities for producing high voltage as well as horizontal-deflection currents. In addition, an all-transistorized receiver operating from a 12-volt battery, requires several intermediate voltages above the supply voltage. These are the supply voltages for the video amplifier, and first-anode and focus-anode voltages for the kinescope. All of these are obtained from the horizontal-deflection circuit. Two systems of driving the yoke and the high voltage circuits are shown in Fig. 54. In *a* of the figure the deflection coil is placed directly in the collector circuit. Coil inductance must be limited to a value that results in a flyback pulse that does not exceed the collector breakdown of the transistor. A diode D_1 is shown in shunt with the transistor. This diode conducts part of the total current that flows when the coil is discharging during the first part of the scan. It is used when the transistor characteristics are not symmetrical in both directions.

The primary of the high-voltage transformer is in parallel with the yoke and bleeds off part of the total collector current. A diode D_2 may be added, as shown, to isolate the transformer from the yoke circuit during the retrace interval. Alternatively, the diode may be omitted, and the reflected inductance

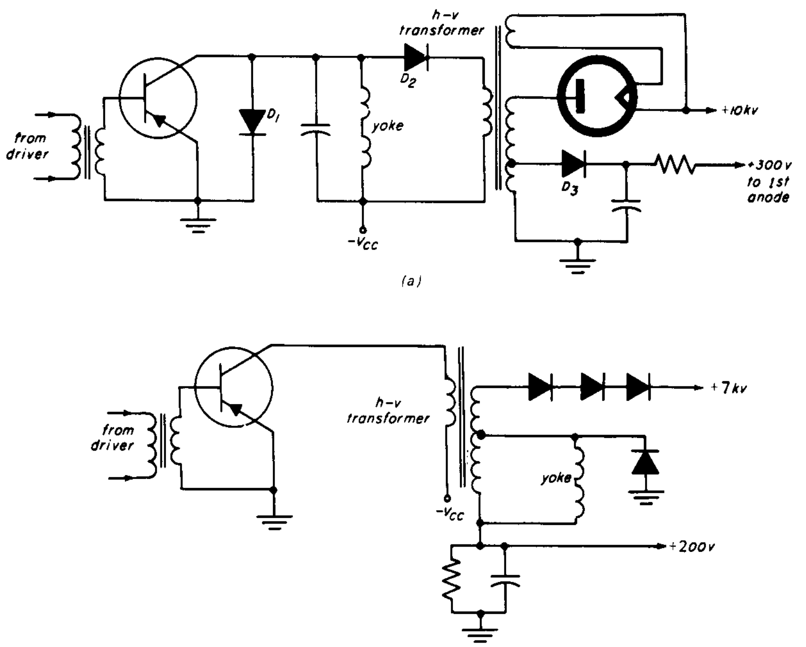
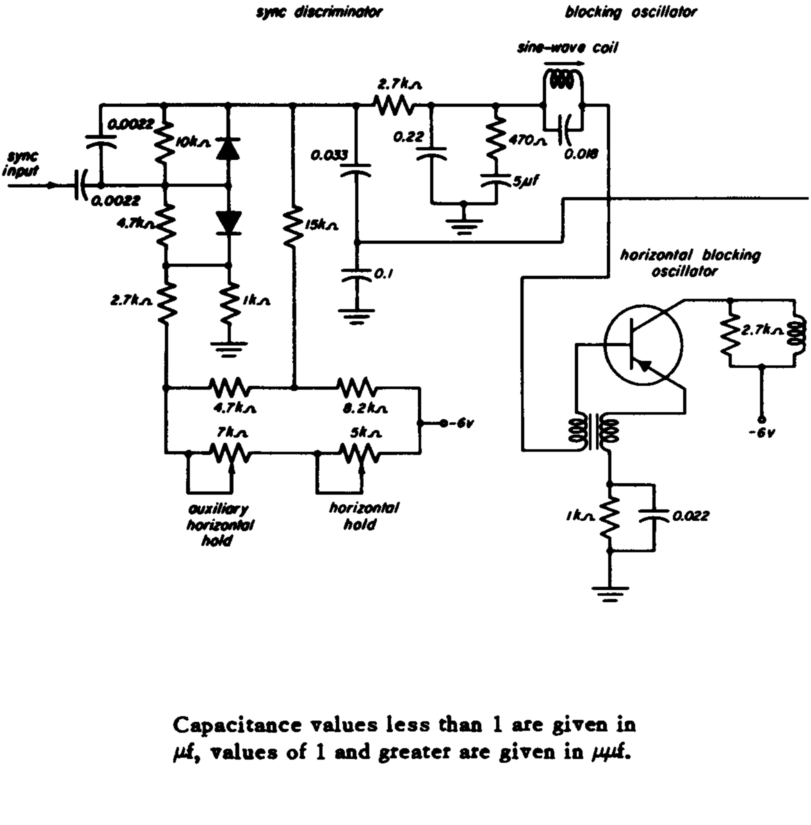


Fig. 54. Types of transistorized horizontal-output systems.



Capacitance values less than 1 are given in μf, values of 1 and greater are given in μμf.

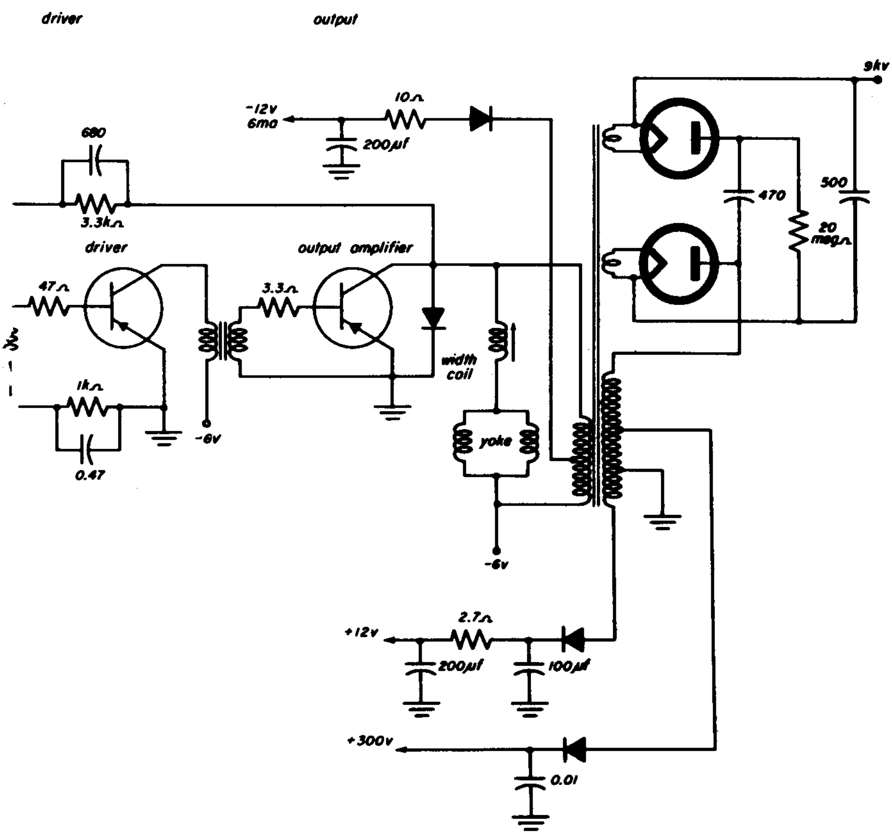


Fig. 55. A representative horizontal-deflection system.

and shunt capacitance of the high-voltage transformer used to provide the third-harmonic signal needed to reduce the collector voltage pulse.

The high-voltage rectifier circuit shown in Fig. 54*a* is similar to that found in vacuum tube circuits. It employs a miniature vacuum diode whose heater voltage is obtained from a single-turn secondary wound upon the core of the transformer. Additional intermediate voltages are obtained as shown by means of a silicon diode rectifier D_3 tapped down to the appropriate voltage level on the secondary.

An alternative output circuit is shown in Fig. 54*b*. In this circuit a voltage step-up transformer is used to couple the deflection yoke to the transistor. Circuit operation is the same, but the transformer permits standard value deflection yokes to be used while maintaining a low peak flyback pulse at the collector. This circuit, like most small-screen transistor portables (where the anode voltage required is comparatively low) uses a solid-state h-v rectifier.

A Representative Horizontal System. Figure 55 shows the horizontal-deflection system of a Philco receiver. It employs a blocking-oscillator circuit that is similar to the one described in this section. Feedback is obtained from a transformer in the emitter leg. A separate transformer is placed in the collector circuit to couple the pulse to the driver stage. The RC network in the emitter circuit of the oscillator establishes the repetition rate of the oscillator. A variable base bias serves to alter the repetition rate over a narrow range. This bias is provided by two potentiometers and an automatic-correction voltage provided by the diode afc discriminator. The sync discriminator is similar to the type found in vacuum-tube circuits. It is driven with a positive-going sync pulse and a sawtooth obtained from the horizontal-ouput circuit. Note also that a ringing coil is added in series with the base circuit for sine-wave stabilization. Stabilization is accomplished in exactly the same way as it is in vacuum-tube deflection oscillators.

The coupling transformer applies a positive-going pulse to the base of the driver transistor. Base current drawn in between cutoff pulses is limited by both the 47-ohm resistor, and the reverse bias developed across the 1-k ohm resistor and the 0.47-μf capacitor. A negative-going pulse is developed at the collector of the driver stage due to flyback action when the stage is cut off. This pulse is inverted by the next coupling transformer and applied to the output stage as a positive-going pulse. The output stage is similar to that shown in Fig. 54*a*. Both the yoke circuit and the high-voltage transformer primary are in parallel and share the total collector current of the output stage. A parallel-connected yoke is employed to minimize ringing problems. A variable inductor, in series with the yoke, adjusts the input impedance of the yoke circuit and so regulates the total yoke current and the width of the picture.

The high-voltage transformer drives a conventional vacuum-diode voltage doubler to provide high voltage for the kinescope. In addition, the secondary is tapped down to provide +300 volts for the first-anode potential. Three isolated diode rectifiers provide + and −12-volt supply voltages to be used by the i-f, sync separator, tuner, and video circuits. Note that the horizontal-output stage is able to provide these supply voltages for the receiver, even though the primary supply voltage is only 6 volts.

9. COMPLETE TV RECEIVER

The remainder of this chapter analyzes the circuits of an actual TV receiver, the RCA Institutes HSK T1. This receiver is very similar to the RCA Cherub portable TV, and uses a KCS-153X chassis.

First, we shall discuss the overall operation of this set by means of a block diagram. Following this, we shall analyze each of the individual stages. A complete schematic of this receiver appears at the end of this book.

Overall Operation. The HSK T1 uses relatively simple television circuitry, even though transistors are used for all but the picture-tube and high-voltage rectifier functions. The sound video, and sync circuits are very much like the familiar vacuum-tube circuits. The deflection circuits, though more of a departure from their vacuum-tube counterparts, are based on the basic electronic techniques and circuits with which you are familiar.

A block diagram of the HSK T1 is shown in Fig. 56. Signals from the TV antenna enter one of the tuners, are converted to the intermediate frequency, and are then amplified by the picture i-f amplifiers. The video detector demodulates the i. f. and provides output to the video-amplifier stages and the agc (automatic gain control) circuit. The video-amplifier stages boost the video signal to sufficient amplitude to operate the picture tube, and also to feed the sync separator and sound takeoff. The sound signal is then amplified, demodulated, and amplified again by the sound i. f., sound detector, and audio-amplifier stages. The sync-separator stage processes the sync pulses for application to the sweep circuits. Vertical-sync pulses are fed directly to the vertical-oscillator stages, where they adjust the operating frequency of this stage. However, the horizontal-sync pulses are fed to an afc (automatic frequency control) circuit, which provides superior control of the operating frequency of the horizontal-oscillator stage.

The vertical oscillator supplies the sweep voltage for the vertical-deflection system. This voltage is

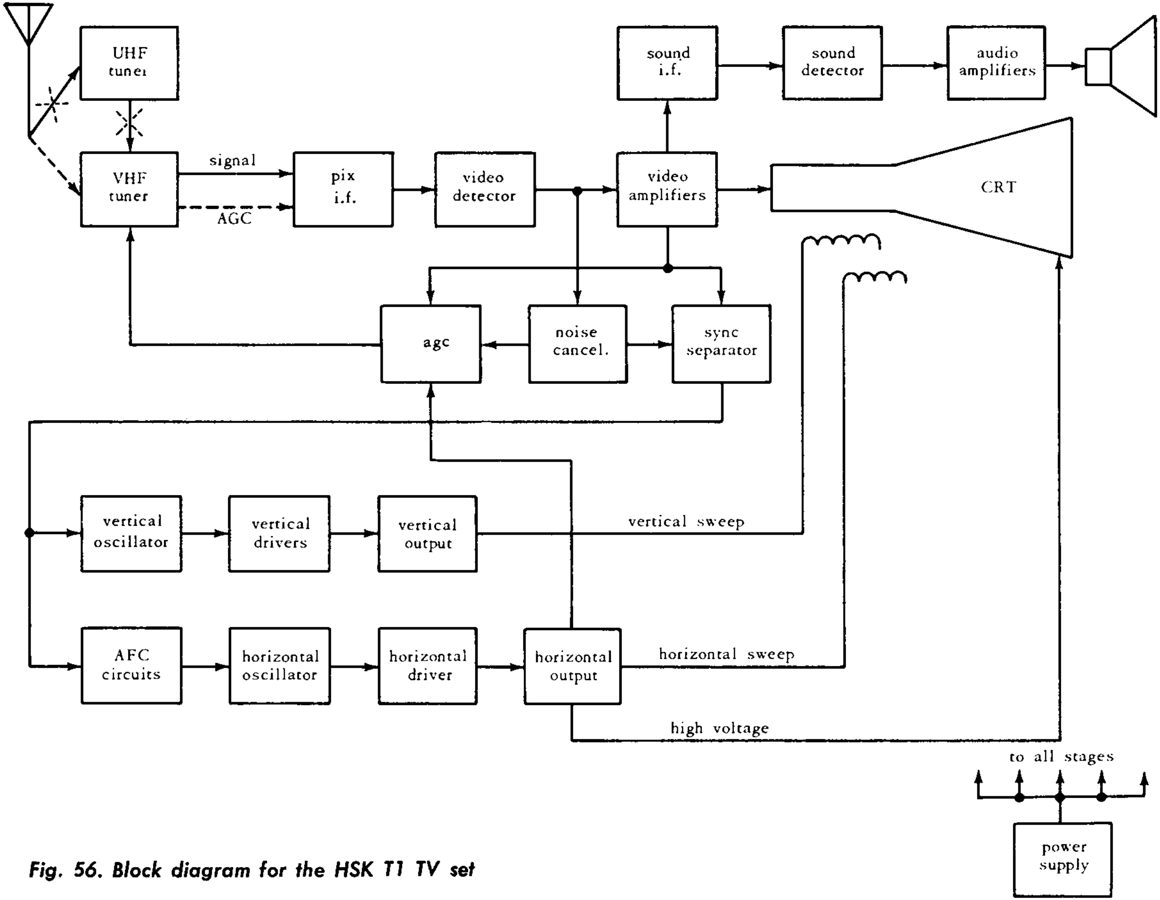


Fig. 56. Block diagram for the HSK T1 TV set

applied to the vertical-output stage by the vertical drivers, which are essentially buffer/amplifier stages. The vertical-output stage supplies the coils in the deflection yoke with heavy deflection current of the proper waveform. The signal flow in the horizontal deflection is essentially the same, with important additions. The horizontal-output stage does more than drive the horizontal-deflection coils in the yoke; it also provides high-voltage for the picture tube and a keying pulse for the agc circuit.

The power supply provides filtered d.c. at the proper voltages for operating the other stages in the set.

Power Supply. The solid-state power supply used in the HSK T1 provides two d-c output voltages and a-c voltage for the picture-tube heater. A simplified schematic diagram of this circuit appears in Fig. 57.

The main d-c supply provides +30 volts at up to 1.8 amperes for powering most of circuitry in the HSK T1. The other d-c supply provides +140 volts at approximately 50 ma for operating the audio-output and video-output stages, and for biasing one of the vertical drivers.

The +30-volt supply consists of a full-wave bridge rectifier employing silicon diodes, and an electronic ripple-filter. The +140-volt supply consists of a half-wave silicon-diode rectifier fed by a separate winding

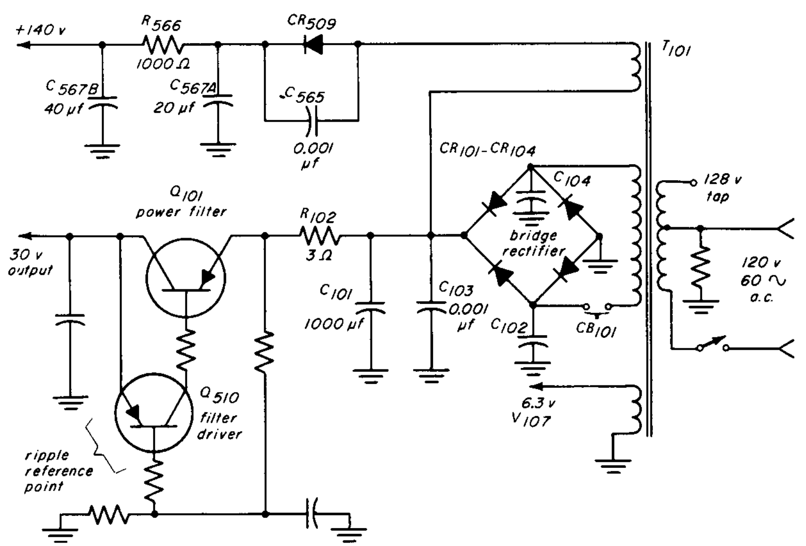


Fig. 57. Simplified schematic diagram of the HSK T1 power supply.

of the power transformer. This supply is "stacked" on the +30-volt supply and filtered by a capacitor-input RC filter. This type of arrangement permits use of a circuit breaker that is common to both supplies.

The electronic-ripple filter in the +30-volt supply is used to produce extremely smooth (low-ripple) output without using large and expensive filter chokes and capacitors. Low-ripple supply voltage is necessary for stages susceptible to hum, such as the low-level amplifier stages. Any ripple in the output appears between the emitter and base of filter-driver transistor Q_{510}, which amplifies, inverts, and applies the ripple voltage to power transistor Q_{101}. The amplified and inverted ripple voltage cancels out the original ripple voltage in this transistor, resulting in an extremely low amount of ripple (about 50 millivolts) in the output of transistor Q_{101}. Furthermore, the circuit is as effective in reducing ripple at 15 kc (the horizontal-sweep frequency) and 30 cps (lowest video frequency) as it is at the 120-cps power-supply ripple frequency. This gives the +30-volt supply a very low a-c impedance, thereby preventing unwanted coupling between stages common to this supply.

Circuit breaker CB_{101} is in a transformer winding common to both the +30-volt and +140-volt supplies. This way, a short across the output of either power supply will cause the circuit breaker to open. The circuit breaker is designed to operate at 2.7 amperes, well above the normal total circuit drain.

Picture I-F Section. The picture i-f section consists of three stages of amplification and a detector, as shown by the block diagram in Fig. 58. A partial schematic of this section is given in Fig. 59. The collector circuit of each i-f stage is tuned and impedance matched to the input of the succeeding stage. This provides sufficient amplification so that approximately 1.5 volts of video signal is developed at the detector under average conditions.

Standard intermediate frequencies in the 41- to 48-megacycle range are used in the HSK T1. The picture carrier is centered at 45.75 mc, and the sound carrier is centered at 41.25 mc.

I-f output signal from the VHF tuner is applied to the base of the first pix i-f amplifier (Q_{201}) through special coupling networks. T_{204}, coil L_{201}, capacitor C_{201}, and resistor R_{201} form a tuned circuit which matches the VHF tuner's output circuit to that of Q_{201}. The T-notch filter consisting of C_{206}, C_{207}, R_{202}, and T_{205} is a sound trap tuned to 47.25 mc. This filter, and the 41.25-mc absorption trap in the collector circuit of Q_{201} eliminate adjacent sound and sound carrier, respectively, from the pix i-f strip.

Blocking capacitor C_{203} couples signal from the T-notch filter to the

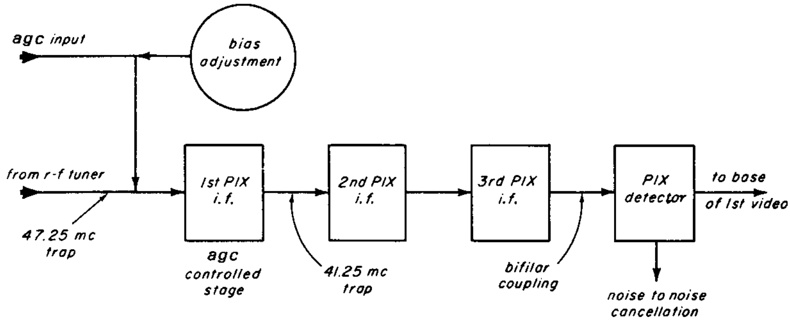


Fig. 58. Block diagram of the picture i-f stages.

base of transistor Q_{201}. The biasing arrangement for the base of Q_{201} is dependent on the setting of the BIAS DELAY ADJ control (R_{112}) and the agc voltage derived from the r-f stage. The r-f stage is used as a d-c amplifier providing output which *reverse* biases the first i-f stage (Q_{201}). A maximum of about 35 db of attenuation is obtainable by reducing the gain of the first i-f stage in this manner.

At the collector of the first i-f stage a tuned circuit is used to couple to the base of the 2nd i-f transistor. This circuit is parallel resonant and matches the base of the next stage by a resistive "tap." This type of transformer is referred to as an *Rx* transformer; a 10:1 voltage step down is obtained. The 2nd and 3rd pix i-f stages are fixed biased by means of conventional voltage dividers. The coupling circuit between the 2nd and 3rd i-f stages is similar to that between the 1st and 2nd i-f stages. The output of the 3rd i-f is tuned by a transformer with separate primary and secondary windings giving a 2:1 step down to match impedances between the collector circuit of Q_{203} and the video detector.

The video detector uses a semiconductor diode (CR_{201}) as a half-wave rectifier directly connected to the base of the 1st video amplifier. Because of this d-c coupling, the entire video-detector circuit is above d-c ground by the amount of bias voltage required by the base of transistor Q_{204}.

Video-Amplifier Stages. The video-amplifier stages consist of emitter follower Q_{204} and video-output stage Q_{205}. The large amount of d-c coupling used between these and adjoining stages results in excellent low-frequency performance; high-frequency response extends to about 3.2 mc.

The 1st video stage uses transistor Q_{204} as an emitter follower to drive the video output, agc, and

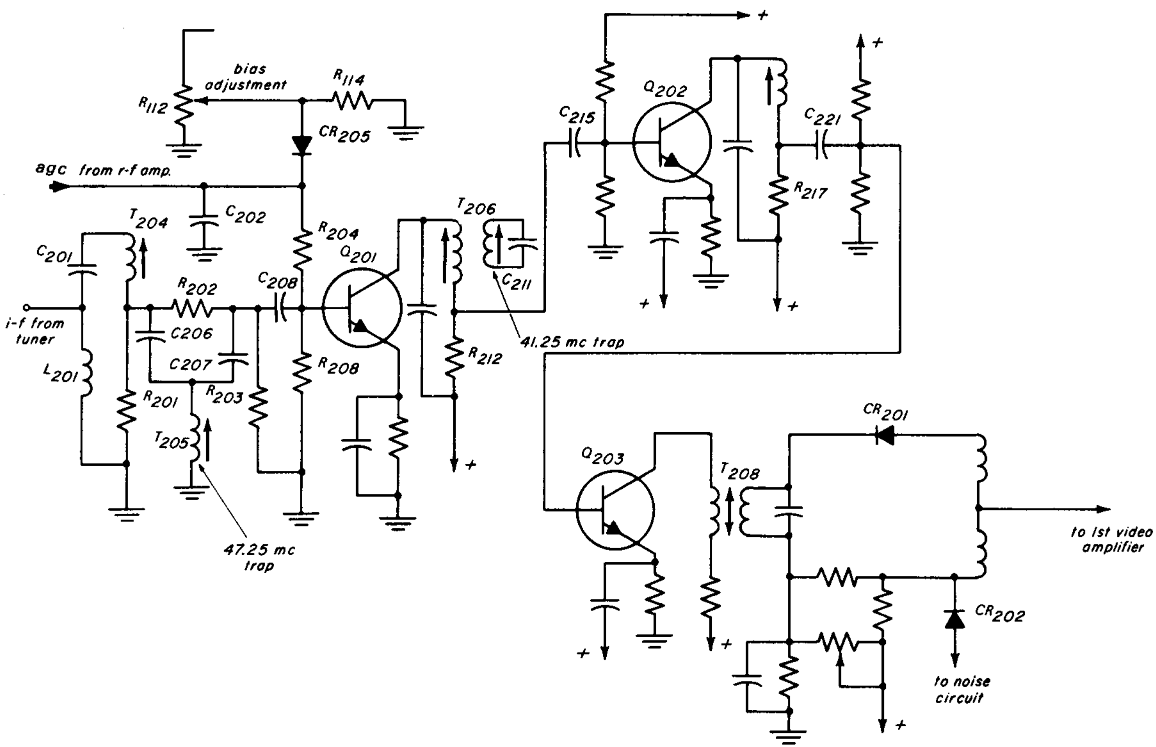


Fig. 59. Partial schematic diagram, picture i-f section.

sound-takeoff circuits, and as a low-gain amplifier for providing video to the sync separator. Output from the video detector is directly connected to the base of transistor Q_{204}, as indicated by the waveform in Fig. 60. This voltage is inverted and slightly amplified, appearing across collector resistor R_{234}. This voltage is used by the sync circuits. Video voltage also appears at the emitter of Q_{204}; this voltage is not inverted and serves several purposes. It drives the agc circuits, the sound-takeoff circuit, and video-output stage Q_{205}, the base of which is directly coupled to the emitter of Q_{204}.

Transistor Q_{204} amplifies and inverts the video signal; a positive-going video signal appears across the load impedance in its collector circuit. This load impedance consists of the CONTRAST control (R_{242}) and resistor R_{245}. Most of the video signal taken from the tap on the CONTRAST control is coupled to the cathode of the picture tube through series-peaking coil L_{206} and d-c blocking capacitor C_{237}. Also, approximately 10% of the d-c component is coupled to the picture tube through resistor R_{243} in conjunction with resistor R_{246}. This gives good rendition of scene values and good action when the picture is fine-tuned.

The 4.5-mc trap in the emitter circuit of Q_{205} causes degeneration at the intercarrier sound frequency, so sound interference in the picture is eliminated. Capacitor C_{233} begins to have a relatively low impedance at medium video frequencies,

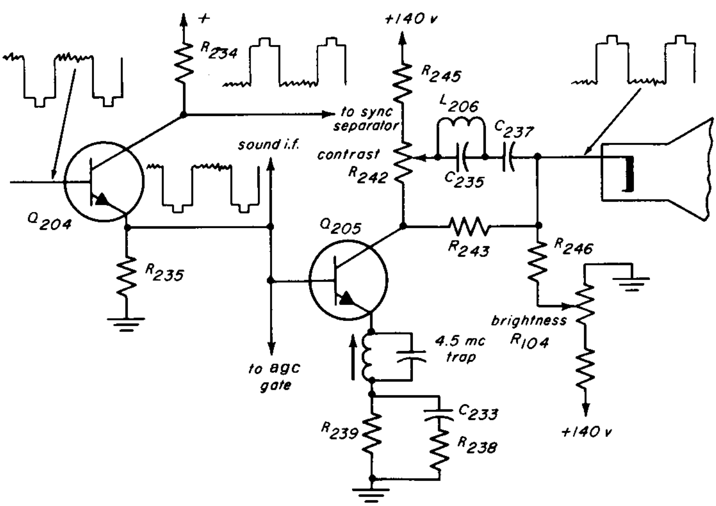
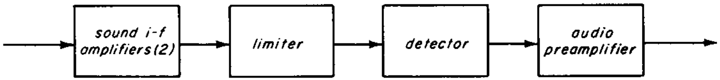
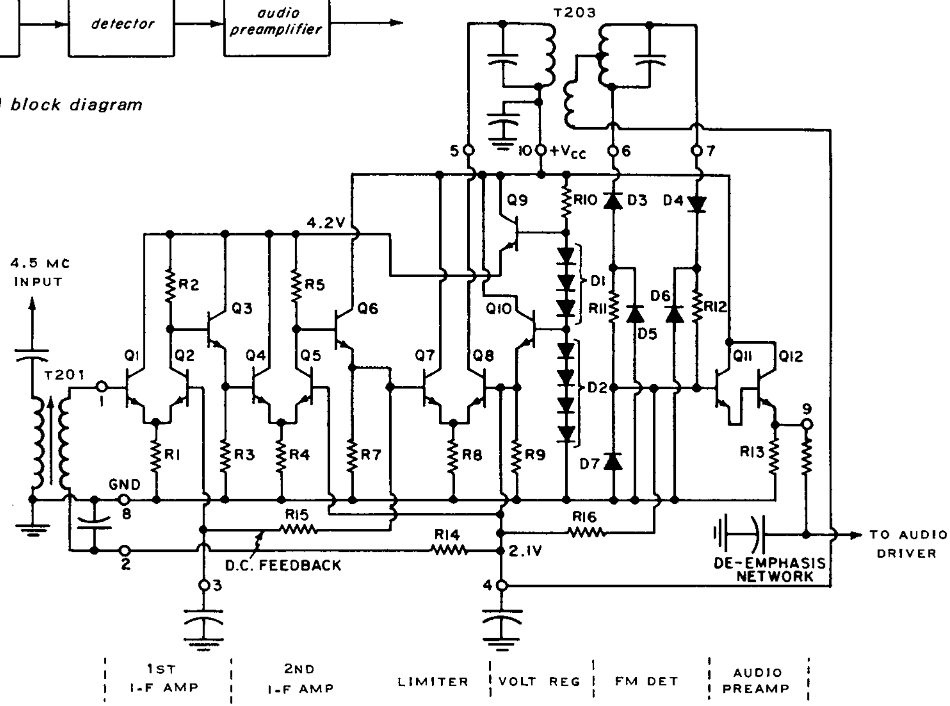


Fig. 60. Partial schematic diagram, video-amplifier section.



(a) block diagram



(b) schematic diagram

Fig. 61. The sound i-f section.

reducing the degeneration caused by emitter resistor R_{239} and thereby increasing the gain of Q_{205} at these frequencies. The output stage operates with a high collector voltage (150 volts) to allow sufficient voltage swing to properly drive the picture tube. The 80 volts peak-to-peak of positive-going video signal needed here is produced by an input to the 1st video stage of approximately 1.2-volts peak-to-peak of negative-going video. A sheet-metal radiator is used on Q_{205} to aid in dissipating heat.

Sound I-F and Detector Stages. The entire sound i-f amplifier, detector, and audio pre-amplifier section is built around a single integrated-circuit chip! A block diagram of this section of the TV set is given in Fig. 61*a*, while Fig. 61*b* is a schematic diagram showing the external- and internal-circuit elements.

The 4.5-mc signal from the emitter of the 1st video amplifier is coupled to the 1st sound i-f amplifier by means of transformer T_{201}. The 1st sound i-f amplifier, consisting of transistors Q_1, Q_2, and Q_3, is directly coupled to the 2nd sound i-f amplifier, which consists of transistors Q_4, Q_5, and Q_6. The greatly amplified signal is now coupled to the limiter stage consisting of transistors Q_7 and Q_8. After limiting, the sound signal is demodulated by the ratio-detector circuit consisting of transformer T_{203}, detector diodes D_3 and D_4, and an unusual type of load resistance consisting of resistors R_{11}, R_{12}, and reverse-biased diodes D_5, D_6, D_7. The capacitance of the reversed-biased diodes filters the r.f. from the detector output. The output of the ratio detector is directly coupled to the audio-pre-amplifier stage, which consists of transistors Q_{11} and Q_{12}. Their output signal is applied to the audio-driver stage through a de-emphasis network consisting of resistor R_{231} and capacitor C_{223}.

The portion of the circuit including diodes D_1 and D_2, and transistors Q_9 and Q_{10} is an internal-voltage regulator supplying the critical portions of the circuit.

Sync Circuits. The sync circuit provides noise-free sync pulses of high amplitude to synchronize the vertical oscillator and horizontal oscillator. The sync circuit will separate sync pulses from composite video if the sync level of the incoming signal is at least 18% (most signals have at least 20-25% sync). A simplified schematic diagram of the sync and associated circuits is shown in Fig. 62.

Composite video with positive-going sync pulses is coupled from the collector circuit of the 1st video amplifier to the input circuit of the sync separator by means of blocking capacitor C_{501}. The RC network consisting of resistor R_{506} and capacitor C_{503} prevents the blocking capacitor from charging too heavily on the broad vertical-sync pulses in the composite video. Resistors R_{510}

and R_{507} reverse bias the base of the sync separator (Q_{502}), thereby increasing the dynamic input range of the sync-separator circuit. Base current flows on sync tips when a signal is present, causing saturation of Q_{502}. Capacitor C_{501} discharges through resistors R_{500} and R_{510} during the sweep time; the negative voltage thus developed cuts off Q_{502} between sync pulses. In this manner, collector circuit of sync separator transistor Q_{502} swings between cutoff and saturation. Since the collector current of Q_{502} also flows through resistor R_{511}, the current variations develop a pulse voltage at the collector whose repetition rate corresponds to the interval between sync pulses, as shown in Fig. 62.

Noise in the sync output can cause erratic sweep because it can trigger a sweep oscillator into premature conduction. To prevent this, a noise-cancellation circuit consisting of diode CR_{202} and transistor Q_{503} is used. Noise-cancellation transistor Q_{502} is normally conducting; it serves as a ground return for the emitter of the sync-separator transistor (Q_{502}), However, when noise appears in the video-detector output, diode CR_{202} couples the noise pulse to the base of transistor Q_{503} through capacitor Q_{505}. A negative-going noise pulse at the base of Q_{503} cuts off this transistor. Since this operation opens the emitter circuit of Q_{502}, it also prevents the sync separator from operating during a noise pulse.

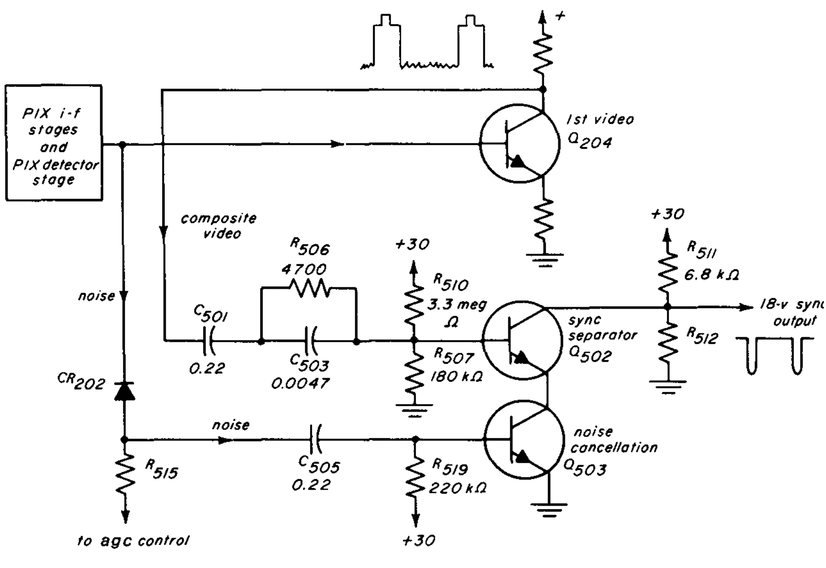


Fig. 62. Simplified schematic, sync separator and noise-cancellation circuits.

Agc Circuit. The agc system used consists of a closed loop made up of the agc gate, the r-f amplifier, i-f amplifiers, video detector, and 1st video amplifier. A block diagram of this system is given in Fig. 63*a*, and a simplified schematic diagram in Fig. 63*b*.

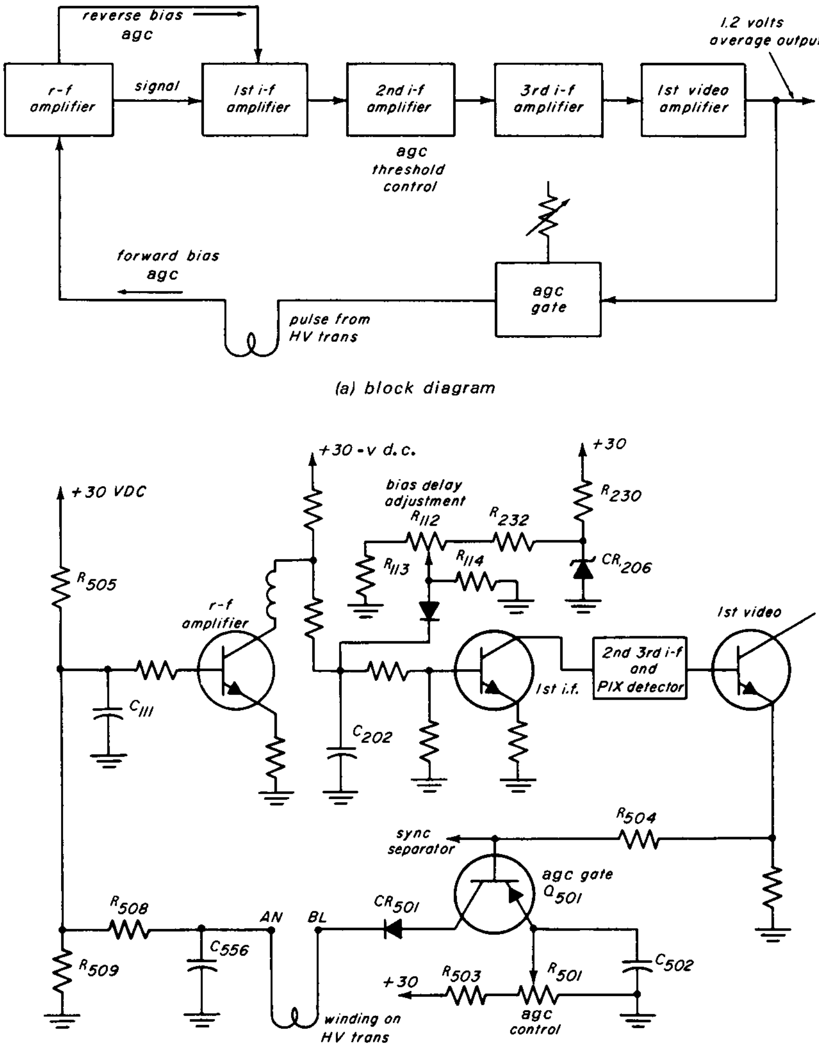


Fig. 63. Block diagram and simplified-schematic diagram of the agc system.

The system maintains a relatively constant 1.2-volt output at the emitter of the 1st video amplifier over a wide range of signal inputs. It is a *gated* or *keyed* agc system in which an agc voltage is developed at horizontal-sync time and sustained for the duration of the horizontal-scan time. Sync tips *only* are utilized to produce the control voltage; the system is noise immune, and not affected by scene variations.

Operation of the agc circuits is as follows: As the input signal to the r-f amplifier increases, the output of the 1st video amplifier tends to increase. The increased video level is applied as an input signal to the *agc gate*. The agc gate is rendered operative at horizontal-sync time by a 30-volt negative pulse from the h-v transformer which is applied to the collector of the agc gate transistor (Q_{501}). At that time, the agc gate transistor amplifies the sync signal which is simultaneously occurring at the base. A positive agc voltage is then developed, and is retained during scan time by the long time-constant of the agc bus. In order to prevent the collector-to-base junction of agc transistor Q_{501} from becoming forward biased by this voltage, diode CR_{501} is inserted between the agc gate collector and the agc bus. Such a condition would short out the agc voltage. The positive agc voltage so formed is then applied as *forward* bias to the r-f amplifier transistor, reducing the gain of the r-f stage.

The r-f amplifier has a dual function. In addition to amplifying the r-f signal, it also functions as a d-c amplifier for the agc circuit. It amplifies and inverts the agc signal, delivering it to the base of the first i-f amplifier as *reverse* bias. This type of bias makes use of the cutoff characteristic of the i-f transistor to reduce the gain of the 1st i-f amplifier.

> NOTE: *Either* reverse bias *or* forward bias will cause a reduction of gain in a transistor amplifier. In one case the cutoff characteristic of the transistor is utilized, and in the other case the saturation characteristic is utilized.

In this manner, both the r-f amplifier and the 1st i-f amplifier act to reduce the gain of the system. In order to maintain optimum gain and signal-to-noise ratio, and to prevent possible cross-modulation from occurring in the mixer, the r-f amplifier is permitted to give nearly full gain on weak-to-medium input signals, most of the control being in the i-f amplifier. As signal gets increasingly stronger, the r-f amplifier assumes more control. About 35 db of attenuation is possible by means of agc action on the 1st i.f., and an additional 35 db attenuation is introduced on strong signals in the r-f amplifier.

Control over the point at which the agc becomes effective on the r-f amplifier is provided by the BIAS DELAY ADJ control, R_{112}. This control sets the minimum gain of the 1st i-f amplifier. The agc voltage applied to the r-f stage is filtered by means of a network consisting of capacitors C_{556} and C_{111}, and resistors R_{508} and R_{509}. Additional filtering for the agc voltage applied to the i-f amplifier is provided by resistor R_{13} and capacitor C_{202}. The time-constant of the system is slow enough to prevent "hang-up" of sync or agc due to horizontal scan being out of sync; but still fast enough to permit agc action on fast fluctuations.

Audio Section. The audio section of the HSK T1 consists of a two stage, d-c coupled amplifier driving a 3-inch speaker; a simplified schematic diagram of the circuit is shown in Fig. 64. It incorporates both d-c feedback for stability, and blowout protection for the power-output transistor.

Audio from the sound detector is applied to R_{105}, the volume control. The audio-output voltage of this control is coupled from its wiper contact to the base of audio-driver transistor Q_{208} by means of d-c blocking capacitor C_{255}. Transistor Q_{208} amplifies the audio voltage and applies it to the base of audio-output transistor Q_{104} by direct coupling. The high-power audio appearing in the collector circuit of transistor Q_{104} is applied to the speaker through output transformer T_{104}. Transformer T_{104} matches the low (3.2 ohm) impedance of the speaker to the much higher output impedance of the audio-output transistor. Capacitor C_{109} provides high-frequency de-emphasis for a more pleasing tone and helps to protect the collector junction of Q_{104} from high-voltage transients. A protective jumper in the speaker connector removes the supply voltage from the collector circuit of Q_{104} when the speaker is disconnected. The very high audio voltage which develops at the collector of transistor Q_{104} when the load is disconnected and an input signal is applied might

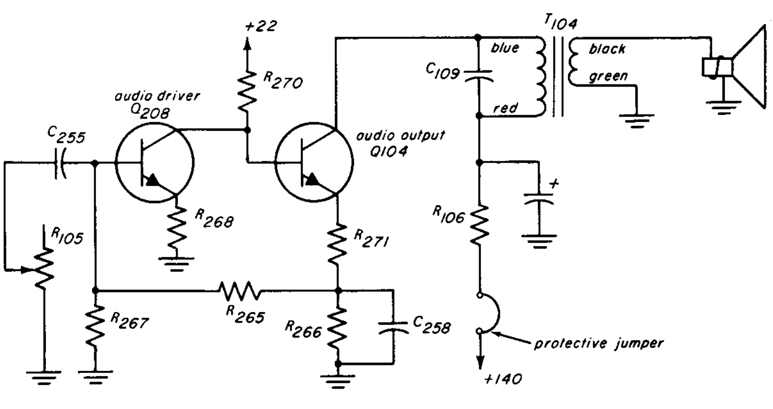


Fig. 64. Simplified schematic diagram of the audio section.

cause a voltage breakdown or "blowout" in the transistor.

Because transistor Q_{104} is direct-coupled to transistor Q_{208}, a d-c negative feedback loop is used. This loop, consisting of resistors R_{271}, R_{265}, R_{266}, and R_{267}, provides both proper biasing and stabilization of the transistors' operating points.

Vertical-Deflection Circuit. The vertical-deflection circuits of the HSK T1 consist of a vertical oscillator, pre-driver and driver stages, and a vertical-output stage, as shown in Fig. 65.

The vertical oscillator, synchronized by the incoming vertical sync pulses, produces a sawtooth sweep signal. The pre-driver and driver stages are buffer-amplifier stages which isolate the vertical oscillator from the low-input impedance of the vertical-output stage. The vertical-output stage boosts the power of the sweep signal to a level suitable for driving the coils in the deflection yoke, and also alters the waveform of the sweep signal so that the output voltage will produce a linear sweep. Feedback from the vertical-output stage to the vertical oscillator is used to sustain oscillation.

The fine performance of the vertical circuits of the HSK T1 results in a linear and stable vertical scan. Circuit refinements insure good noise immunity, freedom from line voltage variations, independence from varying transistor characteristics, and full control of the vertical-sweep waveform with the usual height, linearity and hold controls.

A more detailed analysis of the operation of the vertical-deflection circuit is possible by referring to Fig. 66, a simplified schematic diagram of the entire circuit. The oscillator and output transistors work as a feedback system and are grounded emitter circuits. Oscillation is sustained by the positive feedback from the yoke to the base circuit of transistor Q_{504}, the vertical oscillator. This feedback is coupled through resistor R_{534} (8200Ω) and capacitor C_{513} (0.22 μfd). The pre-driver and driver stages are d-c amplifiers which isolate the transistor from the sweep capacitor (C_{516}), and are "emitter followers" or common-collector circuits.

The vertical oscillator can be considered as a switch, which is *closed* (ON) at retrace time, and *open* (OFF) during vertical scan

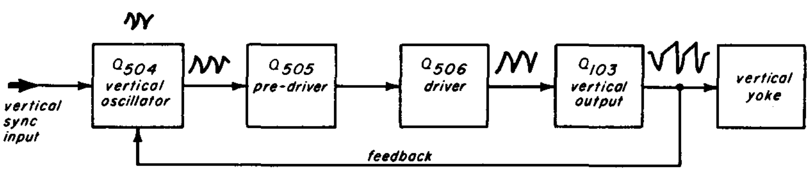


Fig. 65. Block diagram of the vertical-deflection circuits.

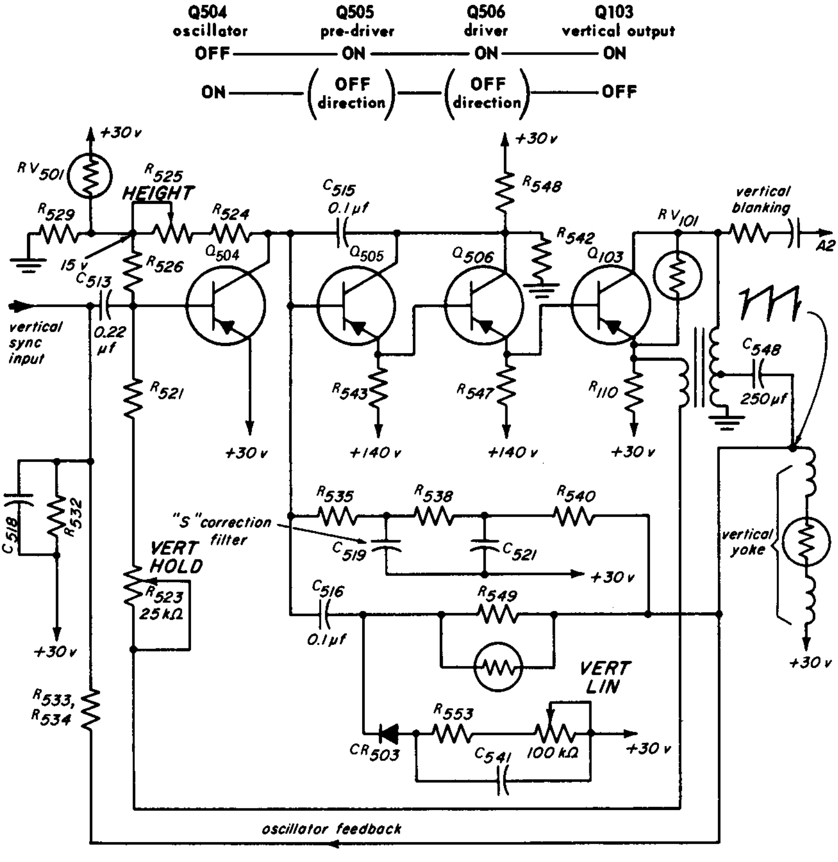


Fig. 66. Simplified schematic diagram of the vertical-deflection circuits.

time. This switching action is utilized at the collector of the vertical oscillator to quickly short sweep capacitor C_{516} to +30 v at retrace time, and permit a linear decrease in voltage on this capacitor during scan time or while the oscillator transistor is nonconducting (OPEN). Spurious oscillation at high frequencies is prevented by capacitor C_{515} which supplies feedback at high frequencies.

Scan time (that time during which transistor Q_{504} is "open") is controlled by the large time constant of capacitor C_{513} (0.22 μfd) and resistor R_{251} (6800Ω), and VERT HOLD control R_{523} (25 kΩ). Conduction of the oscillator occurs when the voltage on the base of the vertical oscillator Q_{504} becomes lower than the emitter. This happens suddenly, and lasts for a very short duration; feedback from the emitter circuit of output transistor Q_{103} insures self starting of the oscillator, and modifies the waveshape at the oscillator base to make the point of conduction independent of transistor characteristics.

The sweep capacitor (C_{516}) is discharged rapidly during retrace time due to the 130-volt flyback pulse. During scan time the capacitor voltage starts a gradual exponential build up, however, this build up is linearized by the feedback from the output transistor, which tends to keep the charging current constant. This produces a very linear voltage rise. This type of circuit is known as a "Miller-feedback amplifier." The extreme linearity so obtained must in fact be somewhat "modified" to produce a linear travel from top to bottom of the picture tube. This is known as "S" correction and is accomplished by adding a filtered portion of the output waveform to the input of transistor Q_{505}. When the output of the filter is combined with the existing linear waveform, the rate of sweep is slightly slowed at the top and bottom of the screen, giving a truly linear picture in the vertical plane.

The pre-driver and driver stages perform the function of isolating the vertical-output stage from the charging circuit of capacitor C_{516}. It is important that the output transistor is fully OFF when the oscillator is conducting. (This corresponds to flyback time.) To insure this condition, the emitters of transistors Q_{505} and Q_{506}, the drivers, are returned to the +140-volt source. Since transistors Q_{505} and Q_{506} are essentially emitter followers, a high-emitter voltage appears on these transistors when the oscillator is ON and this voltage is tied to the base of Q_{103}, thereby insuring that Q_{103} is OFF at that time.

The sweep signal from the vertical drivers is amplified by power transistor Q_{103}, which serves as the vertical-ouput amplifier. The operating point of this transistor is chosen to produce an output voltage like that shown in Figs. 65 and 66. This voltage waveform results in a linear sawtooth circuit flow through an inductive load, specifically the vertical-deflection coils in the yoke. Vertical-output transformer T_{103} serves as an impedance matching auto-transformer to feed the deflection yoke through a d-c blocking capacitor, C_{548}.

An additional winding on T_{103} is used to pick up some a-c component of the deflection waveform to improve the waveform at the base of the vertical oscillator. As capacitor C_{513} discharges through resistor R_{521} and the VERT HOLD control (R_{523}), an exponential decay occurs. This means that a very gradual change in voltage is occurring when the base of oscillator Q_{504} reaches the conduction point. Under these conditions, slight variations in transistor characteristics or any voltage variations could cause initiation of a new cycle at random times. The waveform from transformer T_{103} has the effect of "steepening" the waveform so that conduction occurs at a fixed time and is thereby independent of transistor characteristics.

The d-c component at the emitter of transistor Q_{103} is also coupled to the base circuit of the oscillator through a winding of T_{103}. This assists in self-starting of the oscillator.

Variations in the size of the flyback pulse can cause a "hunting" condition where alternate cycles of output also vary. Diode CR_{503} and capacitor C_{541} are employed to prevent such variations. The VERT LIN control is also inserted at this point of the circuit. It affects the starting point of the output transistor Q_{103} and controls linearity at the top of the picture.

Other circuit features of the HSK T1 vertical circuit include a *VDR* (varistor) across the output transistor which protects it from the large flyback pulse. A *VDR* is also used to stabilize the bias voltage of the oscillator transistor.

Horizontal-Sweep Circuit. The horizontal-deflection circuit of the HSK T1 performs the usual functions of providing linear-deflection current in the horizontal coils of the yoke and providing high voltage for the second anode (A_2) of the picture tube. In addition, the circuit also provides voltage for the accelerating anode of the picture tube. The usual circuit requirements such as horizontal blanking, automatic-frequency control, and protection of circuit components are featured.

As shown in Fig. 67, a block diagram, each circuit performs a specific function with the final result being horizontal deflection.

Horizontal-sync pulses from the sync separator enter the phase splitter. The phase splitter produces two output voltages; each output consisting of horizontal-sync pulses equal in amplitude but opposite in polarity to the other output. These dual outputs are used as references

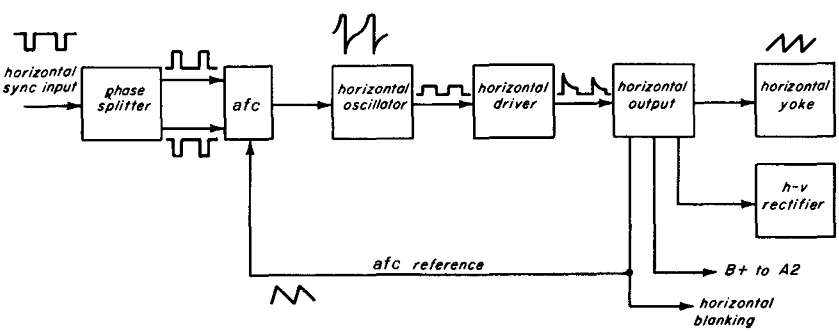


Fig. 67. Block diagram of the HSK T1 horizontal-deflection circuit.

by the afc circuit, where the phase and frequency of the horizontal-output voltage is compared to them. The afc circuit puts out an error signal proportional to the phase or frequency difference. This signal brings the horizontal-oscillator frequency into agreement with the sync frequency. The horizontal oscillator generates the original sweep voltage, which is then amplified and shaped by the horizontal driver. Output from the driver operates the horizontal-output stage. This stage provides scanning current for the horizontal coils in the yoke, low-voltage B+ for the picture tube, and operates a flyback power supply which produces high-voltage d.c. for the picture tube.

Incoming horizontal sync is applied through capacitor C_{551} to the phase splitter, an NPN transistor with equal load resistors in the emitter (R_{561}) and collector (R_{560}), as shown in Fig. 68. This produces sync pulses of equal and opposite polarities for use in the afc detector.

The afc circuit uses two diodes, CR_{504A} & $_B$, to compare the sync pulses with a reference waveform from the horizontal output. When both agree in frequency and phase, no correction voltage results; when they differ, a positive or negative voltage results and is applied to the horizontal oscillator to bring it into agreement with incoming sync.

The horizontal oscillator (Q_{508}) generates the 15,750 cycle waveform for use in developing a sweep waveform in the horizontal-output stage. This stage, shown in Fig. 69, uses an NPN transistor as a blocking oscillator with refinements to obtain high stability of operation. These refinements consist of extra d-c supply voltage filtering (C_{579} and R_{585}), and thermistor (RT_{501}) bias stabilization. The HORIZ STABILITY coil (L_{101}) adjusts the waveform of the blocking oscillator, while the HORIZ HOLD control (R_{580}) adjusts the oscillator frequency. Within certain limits the frequency is automatically corrected

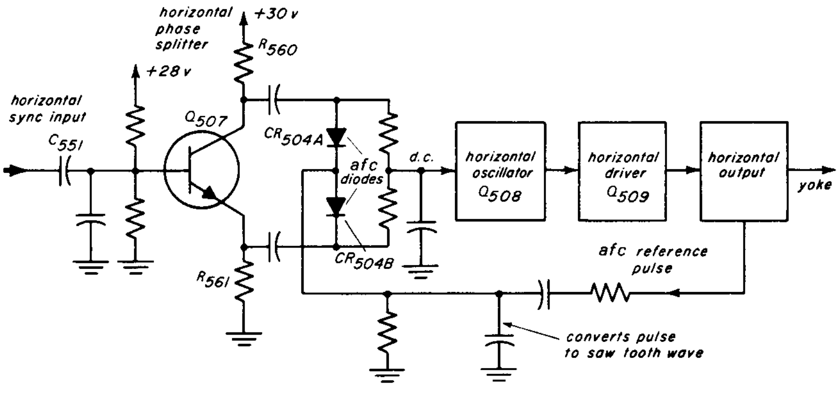


Fig. 68. Simplified schematic diagram of the afc circuit.

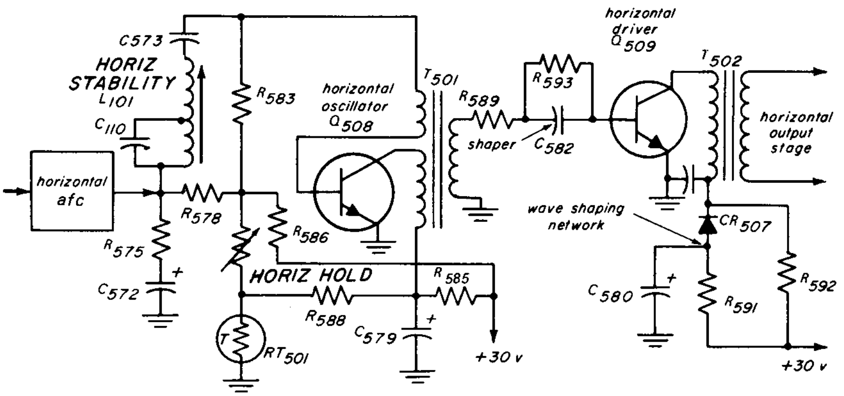


Fig. 69. Simplified schematic diagram of the horizontal-oscillator and driver stages.

by the afc correction voltage. The output of the blocking oscillator is a square wave which is applied to the horizontal driver for further processing.

An additional stage, not usually found in tube-type receivers is the "horizontal driver." This stage takes the square wave produced in the oscillator circuit, amplifies it and

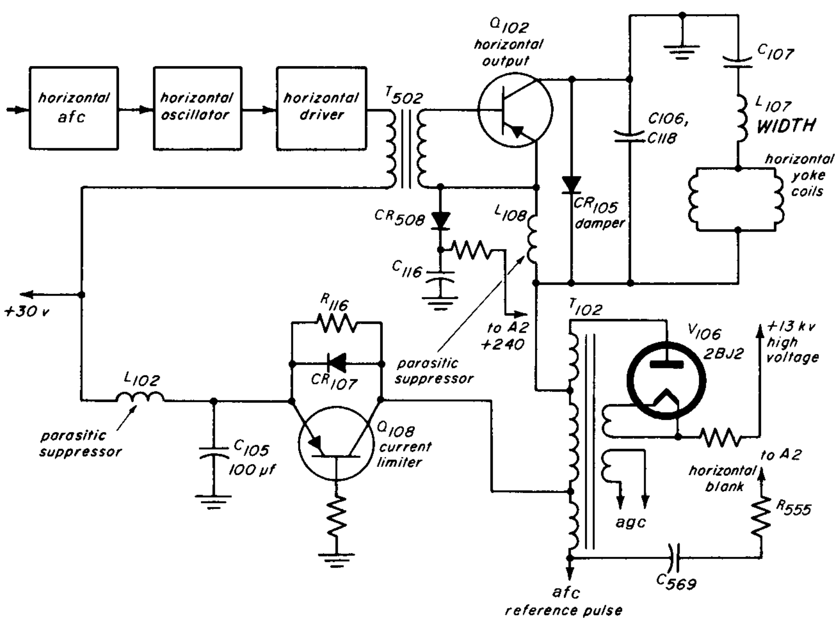


Fig. 70. Simplified schematic diagram of the horizontal-output circuits.

shapes it to a waveform suitable for driving the horizontal-output stage. This waveform has a very steep rise at the start, then becomes a block of some 18 microseconds duration.

Horizontal-Output Stage and H-V Power Supply. The horizontal-output circuit (Fig. 70) uses a PNP power transistor (Q_{102}) in a grounded-collector circuit. The h-v transformer (T_{102}), yoke, and damper (CR_{105}) form the emitter load of this transistor. The output stage comes into conduction at approximately the center of the sweep, and produces a linear rise in current for the remainder of the sweep, at which time it is suddenly turned off by the input waveform furnished by the driver. The familiar "flyback" pulse then occurs. This produces retrace and brings damper CR_{105} into conduction. The damper then conducts for the first half of the sweep, decaying to zero current at the approximate center of sweep, at which time the cycle repeats itself. The relative time relationships of the sequence of events just described are shown in Fig. 71.

Special provisions are made to protect the output transistor from excessive current by means of current-limiter transistor Q_{108}. Diode CR_{107}, in turn, protects the current-limiter transistor from breakdown due to spike voltages. Width coil L_{107}, in series with the horizontal-yoke coils, is to insure adequate size of the horizontal scan. This coil influences the efficiency of the yoke windings.

High voltage is produced as a result of the flyback pulse occurring during retrace time. The flyback pulse is stepped up by autotransformer T_{102} and then rectified by a 2BJ2 vacuum-tube rectifier (V_{106}). The +13 kv thus obtained is utilized as the ultor voltage for the picture tube. The flyback pulse at the emitter of the output transistor is also rectified and filtered, and used as a voltage source for the accelerating anode (A_2) of the picture tube. Diode CR_{508} rectifies the flyback pulse and provides about +240 volts d.c. for the accelerating anode.

The horizontal-output transformer also furnishes the reference pulse for afc comparison, and the keying pulse for keyed agc operation.

UHF Tuner. The UHF tuner, whose schematic diagram is shown in Fig. 72, employs the familiar diode-mixer and transistor-oscillator arrangement.

UHF signals from the antenna are inductively coupled by coil L_1 to the first input tuned circuit. This circuit, consisting of coil L_2 and capacitor C_2 selects the desired signal frequency. Another tuned circuit, consisting of coil L_3 and capacitor C_4, is mutually coupled to the first tuned circuit. L_3-C_4 are also tuned to the signal frequency, and therefore aid in improving the image rejection of the tuner. The signal energy from L_3-C_4 is mixed with local oscillator (Q_1) energy coupled

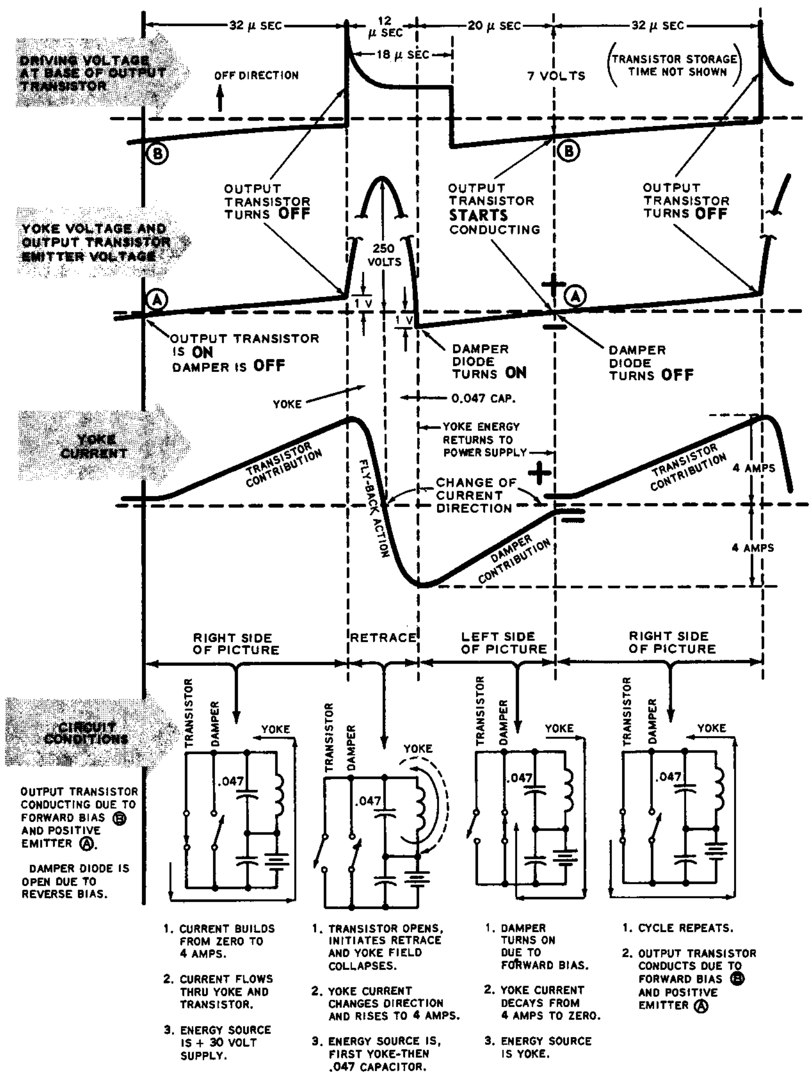
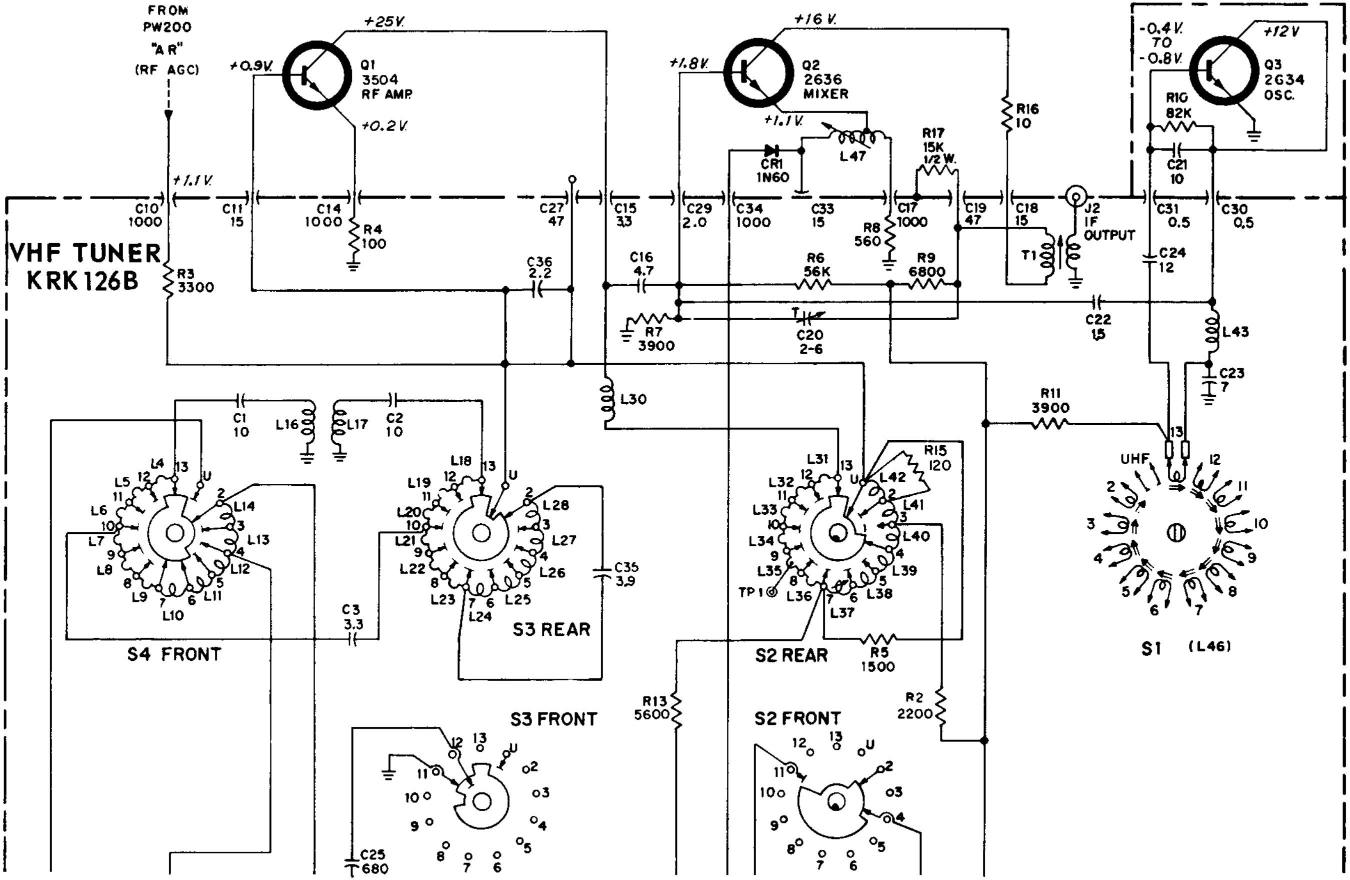
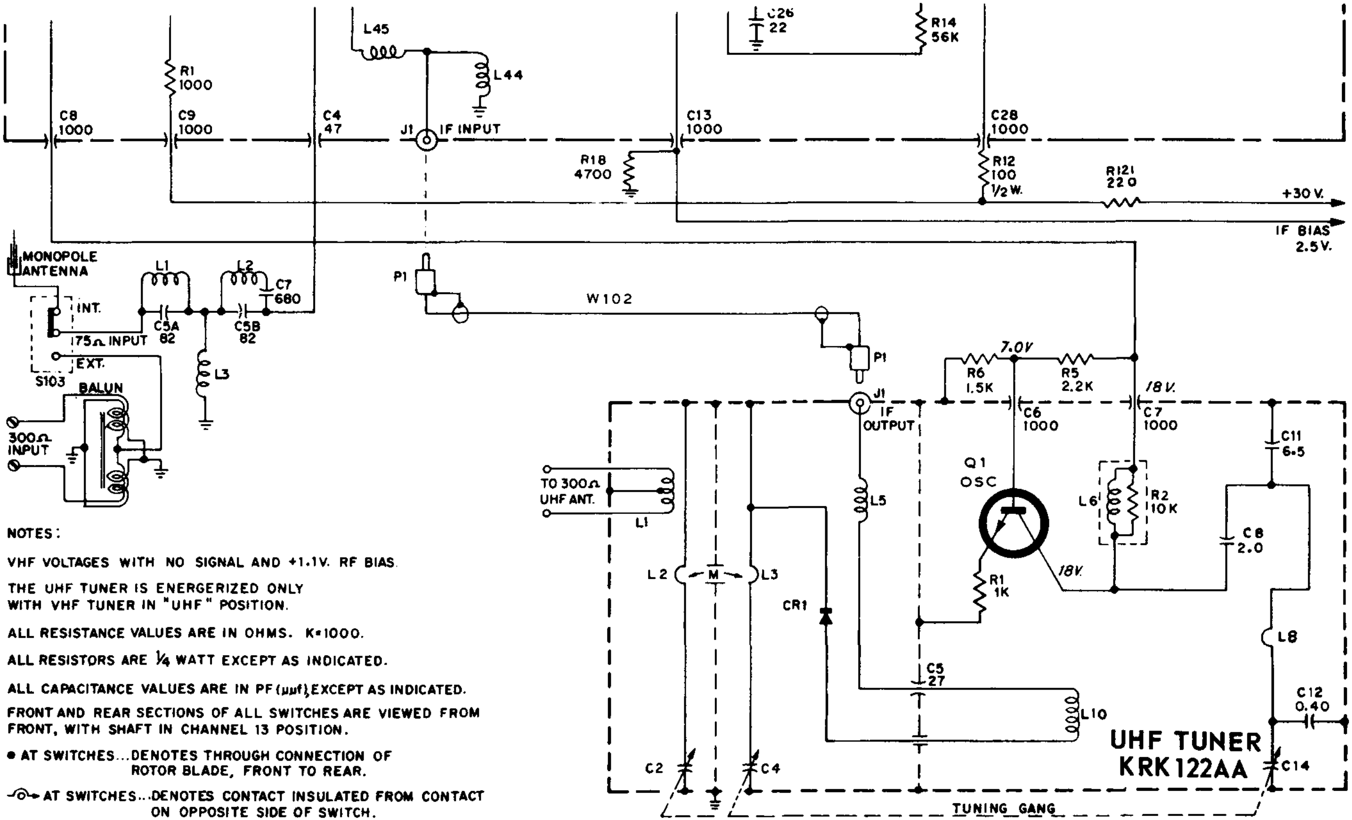


Fig. 71. Relative time relationship of the horizontal-ouput operational cycle.


FROM PW200 "AR" (RF AGC)
VHF TUNER KRK126B
Q1 3504 RF AMP
Q2 2636 MIXER
Q3 2G34 OSC.
+25V.
+0.9V.
+0.2V.
+16 V.
+1.8V.
+1.1V.
+12V
-0.4V. TO -0.8V.
+1.1V.
C10 1000
C11 15
C14 1000
R4 100
C27 47
C15 33
C29 2.0
C34 1000
C33 15
C17 1000
C19 47
C18 15
J2 IF OUTPUT
C31 0.5
C30 0.5
R3 3300
C36 2.2
C16 4.7
R6 56K
R9 6800
R8 560
R7 3900
C20 2-6
C22 1.5
C24 12
L43
C23 7
R16 10
R17 15K 1/2 W.
CR1 1N60
L47
R10 82K
C21 10
T1
L30
R11 3900
C1 10
C2 10
L16
L17
S4 FRONT
S3 REAR
S2 REAR
S1 (L46)
UHF
C3 3.3
C35 3.9
R15 120
R5 1500
R13 5600
R2 2200
TP1
S3 FRONT
S2 FRONT
C25 680



NOTES:

VHF VOLTAGES WITH NO SIGNAL AND +1.1V. RF BIAS.

THE UHF TUNER IS ENERGERIZED ONLY WITH VHF TUNER IN "UHF" POSITION.

ALL RESISTANCE VALUES ARE IN OHMS. K=1000.

ALL RESISTORS ARE 1/4 WATT EXCEPT AS INDICATED.

ALL CAPACITANCE VALUES ARE IN PF (μμf) EXCEPT AS INDICATED.

FRONT AND REAR SECTIONS OF ALL SWITCHES ARE VIEWED FROM FRONT, WITH SHAFT IN CHANNEL 13 POSITION.

● AT SWITCHES...DENOTES THROUGH CONNECTION OF ROTOR BLADE, FRONT TO REAR.

-o- AT SWITCHES...DENOTES CONTACT INSULATED FROM CONTACT ON OPPOSITE SIDE OF SWITCH.

Fig. 72. Tuner circuit schematic diagram.

into the mixer circuit by coil L_{10}. The non-linear characteristics of diode CR_1 generate energy having the same modulation characteristics as the signal frequency, but at the difference frequency between the signal and local oscillator frequencies. Output at the difference or *intermediate* frequency is coupled to the VHF tuner through coil L_5 and cable W_{102}.

VHF Tuner. The VHF tuner, whose schematic diagram is also shown in Fig. 72, consists of an r-f amplifier stage, mixer stage, and local oscillator.

Signal for the r-f stage is obtained either from an external antenna via a balun, or directly from the set's own monopole antenna. Components L_1, L_2, L_3, C_5, and C_7 comprise the various filters and traps required at the front end of the VHF tuner. This network couples VHF TV frequencies to the input tuned circuits of r-f amplifier Q_1. The r-f stage has two tuned circuits at its input. One circuit consists mainly of L_4 to L_{14}, L_{16}, C_1 and stray capacitance; the other tuned circuit consists of L_{17} to L_{28}, C_2, and stray capacitance. These circuits are inductively coupled through L_{16} and L_{17}, and select the desired signal frequency for amplification. The selected signal is applied to base of Q_1, amplified, and applied to another tuned circuit. This circuit, consisting of L_{30} to L_{42}, C_{15}, C_{16}, C_{29}, and stray capacitance, is also tuned to the desired signal frequency. It therefore provides a high load impedance for the r-f amplifier only at the desired signal frequency.

Signal voltage from the third tuned circuit is coupled to the base of mixer Q_2 by capacitor C_{16}. Injection voltage from the local oscillator (Q_3) is coupled to the base of mixer Q_2 by capacitor C_{22}. The mixer transistor is biased for non-linear operation, so the signal and local-oscillator voltage mix and produce variations in the mixer's collector current at their difference frequency. I-f transformer T_1 is resonant at the resulting intermediate frequency, so the collector current variations develop an output voltage across the primary winding of T_1. By means of connector J_2 and the recording winding of T_1, the i-f output voltage is applied to the picture i-f circuits.

Capacitor C_{19}, which normally acts as the a-c return for the collector circuit, is of small value so that some signal voltage will be developed across it for use in neutralization of the mixer stage. The out-of-phase voltage appearing across the capacitor is coupled to the base of the mixer through trimmer capacitor C_{20}.

Picture-Tube Circuit. The picture-tube circuit is essentially the same as that used in all vacuum-tube and transistor TV sets. Both the video signal and variable d-c bias voltage are applied to the cathode of the 12BNP4A picture tube, while the control grid bias is fixed

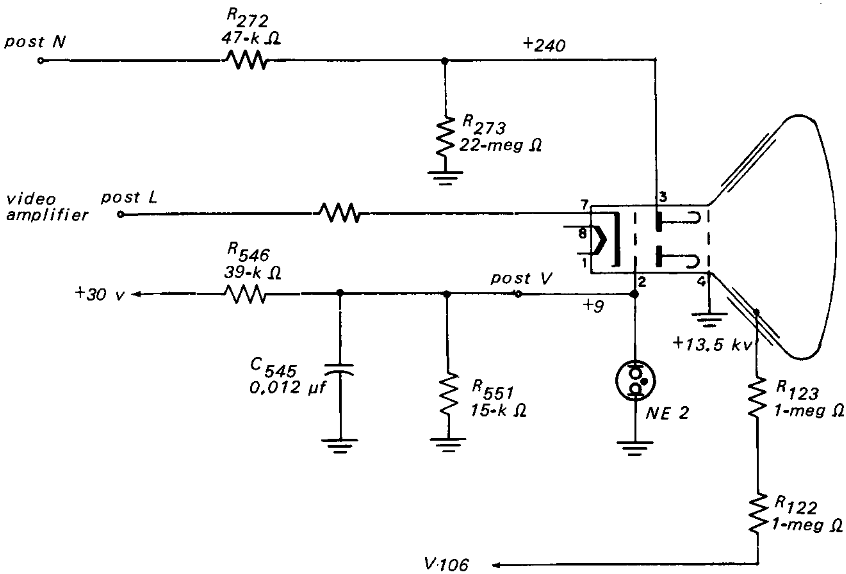


Fig. 73. Picture-tube circuit.

at +9-volts d.c. and grounded to a.c. (Fig. 73). A neon lamp on the control-grid bus protects the picture tube from damage due to arcing caused by the vertical-blanking pulse applied to the control grid.

The focus electrode is at zero-volts potential, and the accelerating anode is supplied with about +240-volts d.c from the junction of resistors R_{272} and R_{273}. The ultor (2nd-anode) potential is about +13 kv when the cathode bias is adjusted for minimum brightness. The voltage is supplied with considerable ripple from the flyback power supply, but the internal capacitance of the picture tube and resistors R_{122} and R_{123} (in the anode-cable assembly) filter it to produce smooth d.c.

CHAPTER 5

SERVICING TRANSISTOR CIRCUITS

INTRODUCTION

The servicing techniques applied to transistorized equipment are the same, in principle, as those employed in vacuum-tube circuitry. Differences exist only in details, but these details are very important. If the details are overlooked much time can be wasted and needless damage to equipment may result. It is the purpose of this chapter to point out these details and present a systematic approach to transistor-circuit servicing so that the technician can attack the problem in an organized manner and recognize symptoms without being led astray by false indications.

1. TEST-EQUIPMENT REQUIREMENTS

In general, the standard test equipment employed for radio and television servicing serves equally well for transistor circuits. However, the conventional volt-ohm-milliammeter and some vacuum-tube voltmeters have limitations when certain readings are made. This does not mean that special equipment must be obtained. If you are aware of the limitations you will not be confused by false readings and can develop alternate ways of making the measurement. Other equipment, such as signal generators and signal tracers need only some attention to coupling details. These specific requirements are discussed below.

Voltmeters. The range of voltages measured in a typical germanium transistor stage in a receiver circuit are shown in Fig. 1. Note the range of base-bias *voltages*. In previous lessons, base bias has usually been mentioned in terms of current, but base-current measurement requires the circuit to be broken in order to insert a current meter. Therefore, a direct measurement of bias *voltage* is more convenient and can be used to establish bias conditions. For this measurement to be useful, however, the voltmeter must be fairly accurate. To illustrate,

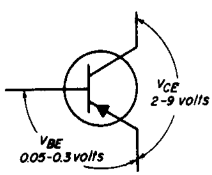


Fig. 1. Typical transistor operating voltages.

consult Fig. 2, which shows collector current versus base-to-emitter voltage for a 2N109 transistor. Note that a change in bias of only 0.1 volt, from 0.25 to 0.35 volt, results in a change in collector current of about 26 ma. This represents a considerable shift in operating point. In order to detect an error in base-bias voltage, the meter should read accurately to a few tens of millivolts. A VTVM with an accuracy of ±3 percent on the 1.5-volt range provides useful readings. (However, it is well to remember that the actual voltage may be plus or minus 45 millivolts on this range.) An RCA WV77E or WV98C is suitable for this purpose. If a less accurate meter is employed, or troubles like distortion are being investigated, it is best to resort to indirect measurements of bias current, as will be shown later.

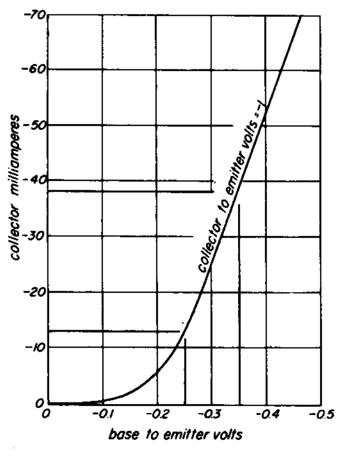


Fig. 2. Transconductance characteristic of type 2N109.

Voltmeter loading is a particularly important consideration when making measurements in small-signal RC-coupled amplifiers. A low-sensitivity meter (1000Ω/volt) has a total resistance of only 1500 ohms on its 1.5-volt range; this will definitely upset the transistor's biasing and give false readings. A 20,000-Ω/volt meter is the *minimum* usable sensitivity in these cases, and a VTVM is preferred.

Ohmmeters. Ohmmeters must be used with great care in transistor circuits. It must be remembered that the ohmmeter has an internal voltage source. Also some instruments are capable of delivering high currents.

In-circuit measurements are often misleading because of the shunt paths provided by transistor junctions that become forward biased by the ohmmeter's supply. An example is shown in Fig. 3. With the ohmmeter connected in this fashion, the internal battery places a forward bias on the emitter junction of the

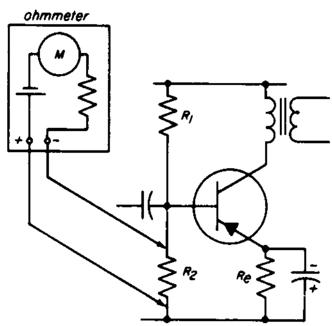


Fig. 3. False resistance reading caused by forward-biased emitter junction.

transistor. This effectively places R_E in shunt with R_2.

In addition to shunt paths provided by transistor junctions, the physically small electrolytic capacitors employed as coupling and bypass capacitors also provide low resistance shunt paths if the ohmmeter's internal supply should polarize them incorrectly. These components can be permanently damaged by even a low voltage of incorrect polarity. Damage can also be caused if the polarity is correct, but the working voltage of the capacitor is exceeded by the ohmmeter's supply. Some capacitors are rated at only 3 volts, while the internal supply of many ohmmeters is 7.5 volts and may run as high as 30 volts. (Most VTVM's employ a 1.5-volt supply, and are therefore always safe as far as voltage breakdown is concerned.)

As a general rule, ohmmeter measurements in the transistor circuit should be made by disconnecting one lead of the component to be checked. This removes any possibility of a shunt path. However, the reading in Fig. 3 could be made if the ohmmeter leads are reversed, in which case the emitter junction becomes reverse biased. This requires a knowledge of the polarity of the ohmmeter voltage at the meter's test leads. In most volt-ohm-milliammeters the same jacks on the meter are used for both voltage and ohmmeter readings. The jack marked plus for voltmeter use is usually (but not always) connected to the positive side of the internal battery when the instrument is employed as an ohmmeter. If doubt exists as to the polarity of the meter leads, connect the ohmmeter to another voltmeter as shown in Fig. 4. An upscale voltage reading identifies the positive lead of the ohmmeter as the one connected to the positive terminal of the voltmeter.

Ohmmeter connections that reverse-bias transistor junctions sometimes run the risk of exceeding the breakdown potential of the junction. This is particularly true in the case of the emitter junction, which breaks down at lower reverse voltages. If the BV_{EBO} of the transistor is not known and if a battery voltage of several volts or more is used in the ohmmeter, it is best to disconnect the transistor.

Ohmmeter readings that, intentionally or accidentally, forward bias a transistor junction also run the risk of causing excessive current to flow through the transistor. The forward-biased junction is practically a short, so that the total current flowing is determined mainly by the ohmmeter's voltage supply

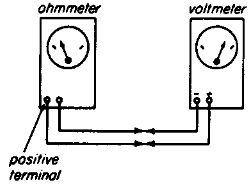


Fig. 4. Determining the polarity of the voltage source in an ohmmeter.

and its internal resistance. Many ohmmeters, including VTVM's, supply a short circuit current of *100 ma* when used on the Rx1 scale. This current can damage many transistors. To check for the maximum current an ohmmeter can supply, connect the ohmmeter to the terminals of a milliammeter (at least 200 ma full scale), as in Fig. 5. To prevent any danger of damage, only use those resistance ranges where the short circuit current is below 1 ma. For most service-type instruments, use of the Rx100 and Rx10K ranges is safe. Do not read forward-bias currents on the Rx1 scale.

Summing up, ohmmeter readings require some judgment before they are made. You need to know three things about the ohmmeter before making measurements: the polarity of the voltage at the leads; the voltage of the internal battery; and the short circuit current. Also, unless shunt paths can be definitely eliminated by proper polarization, one lead of the component to be checked must be disconnected.

Signal Generators. Both audio and r-f signal regenerators, such as the RCA WA-44C and WR-50B, require blocking capacitor in series with the output cable. In vacuum-tube circuits a blocking capacitor is employed to keep supply voltages in the circuit from appearing across the output attenuator of the generator. When signals are injected at the grid, the blocking capacitor may be omitted. In the transistor circuit the blocking capacitor is needed *at all times* to prevent bias changes in the circuit. For example, a signal injected between the base and ground of the r-f amplifier shown in Fig. 6 places the low (50-ohm) output resistance of the generator between the base and the negative supply. The increase in base-bias current causes excessive conduction of the transistor and may result in a burnout.

In selecting the values of blocking capacitors, you should bear in mind the low-input impedance of

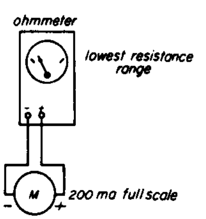


Fig. 5. Determining the short-circuit current of an ohmmeter.

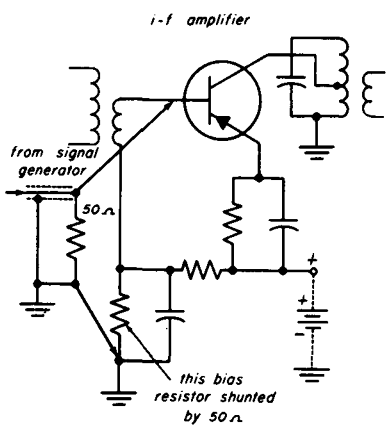


Fig. 6. Effect of signal generator output resistance on transistor bias.

transistor amplifiers. Capacitor values are much larger than those used in vacuum-tube circuits. For general service work a 0.005-μf capacitor and a 0.1-μf capacitor may be used for r-f and audio generators, respectively. A capacitor working-voltage rating of 25 to 50 volts is all that is required in most cases. If the generator is equipped with a calibrated output attenuator and exact input-voltage measurements are to be made, larger values of blocking capacitors should be used. In this case, the r-f generator requires a blocking capacitor of 0.1 μf and the audio generator requires a 100-μf capacitor in series with the output cable.

Oscilloscopes. Waveform observation is helpful in tracking down causes of distortion. Distortion is a common complaint in transistor radios. Since audio voltages are very low, even in the driver and output stages of a transistor amplifier, a scope with high vertical sensitivity, such as the RCA WO-91B, is required. A deflection sensitivity of 3.5 millivolts per inch or better is needed.

Tools and Accessories. Transistorized circuitry is always miniaturized and is often fabricated on a printed-circuit board. Any repairs to this type of circuitry requires a well tinned, pencil type soldering iron. Care must be taken in soldering, both to prevent damage to the printed board and to make sure that the transistor itself is not overheated. A good quality 60/40 (60% tin, 40% lead) solder helps to make joints quickly with a minimum of heat. When transistors must be removed or installed, some means must be provided to conduct heat away from the transistor junction. Fig. 7 shows how the long nose pliers may be used as a heat sink when unsoldering transistor leads.

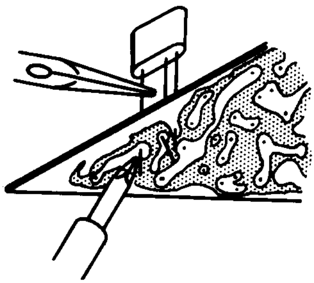


Fig. 7. Use of long-nose pliers as a heat sink.

Leakage between the heating element and the soldering-iron tip can cause the tip of the iron to be above ground potential. This voltage may cause transistor damage if the chassis has a return to ground. Remove all electrical connections to the chassis while soldering, or check the soldering iron for leakage as shown in Fig. 8. A voltage reading

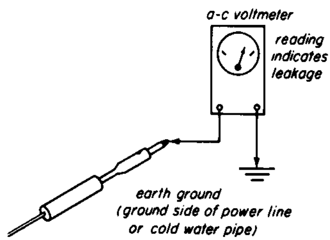


Fig. 8. Checking a soldering iron for electrical leakage.

indicates leakage. The check should be made twice with the a-c plug reversed for each check. Do not use a VTVM.

The same considerations about line leakage also apply to the use of your oscilloscopes and signal generators. Even though the equipment may have a power transformer, the "line-filter" capacitors which are usually connected between the transformer primary and chassis ground permit an a-c flow between chassis ground and earth ground. Therefore, if you connect the ground lead of such test equipment to a sensitive point in transistorized equipment having an earth ground, damage will result.

Meter test probes with sharp points facilitate checks on the printed-circuit board. They minimize the danger of accidentally bridging adjacent conductors. Also, the needle points easily pierce resin, varnish or surface corrosion on the conductors. False readings are often the result of not making good connections on the printed-circuit board.

Because *transistorized* equipment is usually also *compact* equipment, you will not have as much room between components to "poke around in." This makes miniature tools and connectors a necessity. *Really* fine needle-nosed pliers, a good tweezer, goose-neck magnifying glass with stand, and needle-point diagonal cutters are all extremely useful. Use miniature alligator clips *exclusively* for making connections to compact circuitry; the standard-sized alligator clips usually short against adjacent leads and components.

2. TESTING TRANSISTORS

A transistor that is operated within its ratings, with respect to voltage, power dissipation, and temperature, is expected to have an almost unlimited life. Failures in transistorized circuits are more often the result of damage or malfunctioning of some other component. This is particularly true when miniatured transformers and electrolytic capacitors are employed. Despite the reliability of the transistor itself, failures occur due to shorts or opens in the bias circuitry, temporary overloads, physical damage or even mishaps while servicing. Thus, occasions doubtless occur when an evaluation of the transistor becomes necessary.

A great number of transistor testers and analyzers are available. Some only check leakage and current gain while others are capable of measuring all of the transistor parameters. From a servicing viewpoint, a few simple tests are enough to reveal the vast majority of troubles. These tests, to be described below, reveal shorts, opens, excessive leakage, and provide a rough check of current gain. Fortunately, little equipment is required. Some of the tests require only an ohmmeter. The more elaborate checks can be made with just a few additional components.

Testing the Junctions. The transistor contains two PN junctions or diodes. Most of the characteristics of the transistor are tied in with the behavior of the junctions, while the rest of the crystal simply serves as connective material. Damage to the transistor therefore almost always shows up as a malfunctioning of one of the rectifying junctions. The fault may be an open or shorted junction, or excessive reverse current (leakage).

A rough but useful check of the condition of the junctions may be made with an ohmmeter. First, the forward resistance of each junction is measured, as shown in Fig. 9. In this figure, the connections for a PNP transistor are shown. The negative terminal of the ohmmeter is connected to the base. The forward resistance of both junctions is checked by touching the emitter and then the collector terminal in turn with the positive lead. A high reading indicates an open junction. A normal unit should show a reading below 500 ohms. Observe the precautions given earlier for using the ohmmeter. The forward resistance of the junctions of an NPN unit is checked with the same set up shown in Fig. 9 but with the leads to the ohmmeter reversed.

To check for shorts or excessive leakage, reverse the ohmmeter connections, and switch to a higher resistance scale, as shown in Fig. 10. Now the ohmmeter places a reverse bias on each junction in turn, and leakage current is registered on the meter. A low resistance reading indicates a shorted or leaky junction. Low and medium power germanium transistors should show a resistance reading of at least 500-k ohms. Typical readings taken with an ohmmeter on the Rx10K scale are 700-k ohms to 1.5 megohms. Silicon transistors give much higher resistance readings. Power transistors have larger junctions and therefore greater leakage currents. Reverse-bias resistance readings should be 50-k ohms or greater for power transistors.

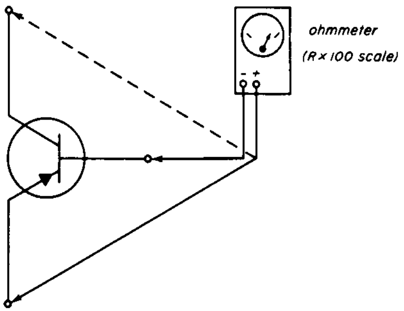


Fig. 9. Method of checking the forward resistance of both junctions.

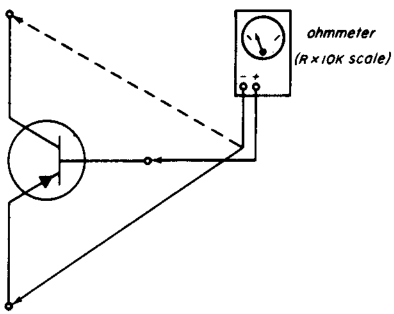


Fig. 10. Method of checking the reverse resistance of both junctions.

Reverse-resistance checks on NPN transistors are made by reversing the ohmmeter leads, so they are opposite to that shown in Fig. 10. Note that the actual numerical reading in ohms is meaningless as the ohmmeter can only measure linear resistances. The specific ohms-reading changes from meter to meter and is not the same for different settings of the range switch. The minimum and maximum values given here apply in the majority of cases. To increase the accuracy of the ohmmeter check, the readings should be compared with those made on a known good transistor of the same type.

A better check of the condition of the junctions can be made by measuring the leakage current directly. This can be done with the simple setup shown in Fig. 11. Maximum leakage currents (I_{CBO} and I_{EBO}) are stated in the manufacturer's specifications. For most low- and medium-power transistors, emitter and collector leakage currents should not exceed 10 to 15 microamperes at room temperature. Power transistors may exhibit leakage currents of 100 microamperes or more. Leakage in silicon junctions is usually measured in fractions of a microampere. A transistor which has been damaged due to momentary overloads or high temperatures usually displays excessive leakage current.

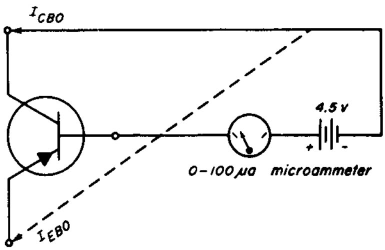


Fig. 11. Measuring the leakage current of both junctions.

If a microammeter is not available a leakage check can be made with a 0-1 ma or 0-10 ma meter using the test setup shown in Fig. 12. I_{CEO} is measured in this arrangement and the leakage current is amplified by the current gain of the transistor ($I_{CEO} = I_{CBO} \times h_{FE}$). The reading for low- and medium-power transistors should be below 1 ma. This type of leakage-current check is employed in the less expensive transistor testers, as it does not require a costly microammeter. However, this test may sometimes be misleading as a *normal* reading may result from a combination of high leakage and low current gain. A direct measurement of both I_{CBO} and I_{EBO} provides a better evaluation of junction conditions.

Any check of leakage should make allowance for ambient temperature. The values stated here apply at normal room temperature,

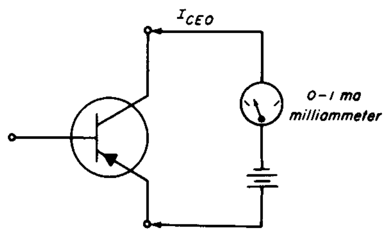


Fig. 12. Leakage measurement using a less sensitive meter.

approximately 20°C or 68°F. Estimate leakage currents at other temperatures by doubling the expected leakage current for 10°C rise in temperature.

Current Gain. Transistor action may be checked with an ohmmeter using the set up shown in Fig. 13. The meter registers I_{CEO} before the 500-k ohm resistor is touched to the base. Connecting the resistor allows a small base bias to be applied, and the meter shows an increase in current (decrease in resistance reading). To make more quantitative measurements either alpha or beta should be measured.

Transistor alpha may be checked using the set up shown in Fig. 14. The basis of the check is a measurement of the d-c alpha as shown in *a* of the figure. It may be assumed that if the collector current reading is at least 0.95 of that shown by the meter in the emitter circuit the transistor has near normal gain. It may be assumed that the a-c gain

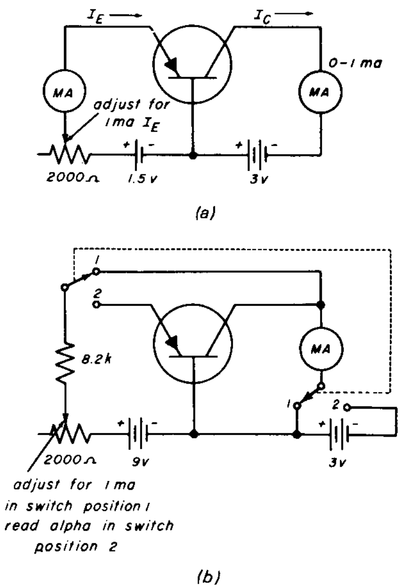


Fig. 14. Setup to measure alpha.

characteristics are normal if the d-c alpha appears normal. The set up in *b* of the figure requires only one meter, and accuracy is somewhat improved since the same meter measures both I_E and I_C. With the switch in position 1, the rheostat is adjusted to give a full scale reading of 1 ma on the meter. The switch is then turned to position 2, which applies one milliampere to the emitter circuit and the collector current is read on the meter. A reading below 0.8 ma indicates a defective unit.

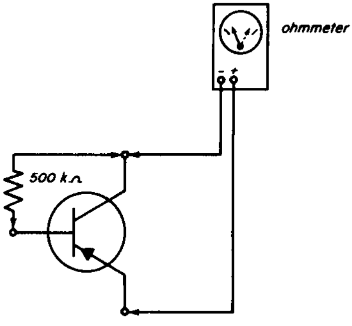


Fig. 13. Rough gain measurement using an ohmmeter.

Beta Measurements. Beta checks allow a more accurate evaluation of transistor gain, as changes in alpha too small to be detected in the above measurements show up as large changes in beta. A simple beta checker is shown in Fig. 15. The 450-k ohm base bias resistor supplies a calibrated base current of 10 microamperes. D-c beta is the ratio of the collector-current reading divided by the base current. If the meter reads 1 ma, then

$$\beta = \frac{I_C}{I_B} = \frac{1000\mu a}{10\mu a} = 100$$

The collector current measured in this way includes leakage current and yields a measurement of d-c beta which may be different from the a-c current gain of the device. To measure a-c beta a collector-current reading is first made with the base open. Then, the 10 microamperes of base current is applied and the *change* in I_C noted. A-c beta may then be calculated as the change in $I_C/10$ microamperes.

Direct-Reading Beta Checker. A simple arrangement giving accurate direct measurements of a-c beta is shown in Fig. 16. This setup allows current gain to be measured at an operating d-c collector current of 1 milliampere, and a collector voltage of 4.5 volts. This approximates the operating point used

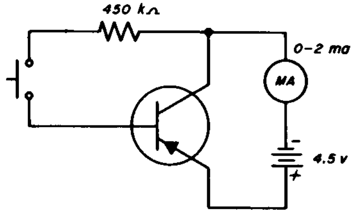


Fig. 15. Setup to measure d-c beta.

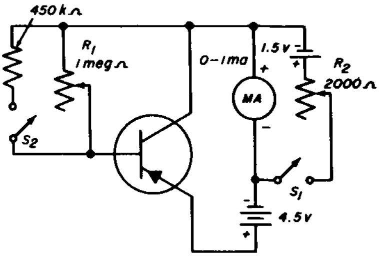


Fig. 16. Setup to measure a-c beta at an I_C of 1 ma.

to specify the parameters of many low- and medium-power transistors. To set up the meter, all switches are open and R_1 is adjusted to yield a full scale reading, establishing I_C at 1 ma. S_1 is then closed and R_2 is adjusted to zero the meter. Note that the 1.5-volt battery supplies a current in the opposite direction through the meter. Closing S_2 supplies an additional base current of 10 microamperes. The meter then registers the change in I_C above the original 1 ma. Beta is read out directly by multiplying the meter scale by 100. A full-scale reading yields:

$$\beta = \frac{\text{change in } I_C - 1 \text{ ma}}{\text{change in } I_B - 10\ \mu a} = 100$$

The circuit and component values shown in Fig. 16 apply for PNP low- and medium-power transistors. For NPN units the supply batteries and the meters should have their leads reversed.

Power Transistors. The same tests described above also apply to power transistors (one- to 40-watt collector dissipation). The difference is only in the magnitude of current measured in the tests. Leakage currents are higher, typically 100μa for I_{CBO} and I_{BEO} and 10 milliamperes for I_{CEO}. Check the manufacturer's data for maximum values. Ohmmeter checks using the Rx1 scale are quite safe on power transistors.

Current-gain measurements on power transistors are meaningless if made at low collector currents. Therefore the tester shown earlier needs to be modified to measure gain at collector currents of 0.1 or 1 ampere, depending upon the transistor type. Lantern batteries, heavy-duty No. 16 dry cells or storage batteries can supply the above currents for short periods of time.

Commercial In-Circuit Transistor Tester. The RCA WT-501A Transistor Tester (Fig. 17) is an accurate instrument whose portability and in-circuit measuring ability make it ideal as a service instrument. The ease with which in-circuit tests can be conducted is demonstrated during the following procedures:

1. Turn off the equipment under test.

2. Turn the PNP-OFF-NPN switch to the position corresponding to the type of transistor you are about to test.

3. Turn the inner CAL knob fully counterclockwise until it clicks into the ICEO position.

4. Turn the range-function switch to the CAL position of a suitable current range. Select the current range as follows:

Type of Transistor	*Range*
low power	10mA or 1mA
medium power	100mA
high power	1A

If the 1-ampere range is selected, make this test as quickly as possible to minimize the drain on the battery.

5. Connect the test leads to the corresponding electrodes of the transistor under test.

6. Adjust the IN CIRCUIT ZERO ADJ knobs until the meter pointer indicates 0.

NOTE: The inability to zero or calibrate the WT-501A in this procedure indicates a defective transistor or other circuit component.

7. Turn the inner CAL knob clockwise until the meter pointer is over the BETA CAL X1 mark on the mirrored scale.

8. Turn the range-function switch to the BETA position of the selected current range. Read the circuit current gain (in-circuit beta) directly on the BETA scale.

For a more accurate beta measurement, and to measure leakage currents, use the WT-501A as an

Fig. 17. RCA Model WT-501A Transistor Tester.

out-of-circuit tester. To measure beta, proceed as follows:

1. Connect the test leads to the corresponding electrodes of the transistor to be tested, or plug the transistor into the appropriate socket.

2. Turn the PNP-OFF-NPN switch to the position corresponding to the type of transistor under test.

3. Turn the inner IN CIRCUIT ZERO ADJ knob fully counterclockwise until it clicks into its OUT OF CIRCUIT position.

4. Turn the range-function switch to the CAL position of the current range to be used. Select the current range as you did in the previous procedure.

5. Adjust both CAL knobs for a full-scale reading on the mirrored scale (BETA CAL X10 mark).

6. Turn the range-function switch to the BETA position of the selected current range. Multiply the reading on the BETA scale by 10 to determine the current-gain (beta) of the transistor under test.

While set up to measure beta, you can also measure the collector-to-base leakage (I_{CBO}) by turning the range-function switch to the ICBO 100μA position. Read the leakage current directly on the I_{cbo} scale of the meter. For most silicon transistor I_{CBO} will be below 1μ.

To measure the collector-to-emitter leakage, (I_{CEO}) proceed as follows:

1. Turn the inner CAL knob fully counterclockwise until it clicks into its ICEO position.

2. Turn the range-function switch to the CAL position of the 1 mA range. Read the collector-to-emitter leakage current (I_{ceo}) on the Icbo scale of the meter, where full scale now represents 1 ma. If the meter pins at full scale, switch to the CAL position of a higher current range until you get an on-scale reading. Keep in mind that the value of the full scale reading during *this* test depends on the current range you are using.

Identifying Transistor Leads. Many case and base wiring systems are in use at present and the manufacturer's data should be consulted for proper lead identification. Some of the more popular base systems are shown here.

The *linotetrar* base is very common. It is shown in Fig. 18*a*. This

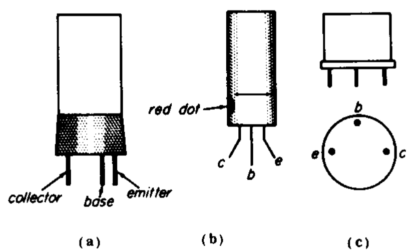


Fig. 18. Base connections for home-entertainment transistors.

base has space for four leads in a straight line, with one space omitted. The lead standing off by itself is the collector, the base is between the collector and emitter terminals. Fig. 18*b* shows three equally spaced leads arranged in a line. The collector terminal is identified in this case by a red dot or a line at one end of the case. The base is again the center lead.

The triangular-shaped arrangement shown in Fig. 18*c* is also popular. The leads may be identified by holding the transistor so that the leads form an arrowhead pointing upwards as shown. The collector terminal is then on the right with the base and emitter lead identified by counting counterclockwise from the collector. (Some transistors using this type of base have the positions of the collector and emitter leads reversed.)

Many power transistors use the case style shown in Fig. 19. The collector terminal is returned to the case itself and connections are made

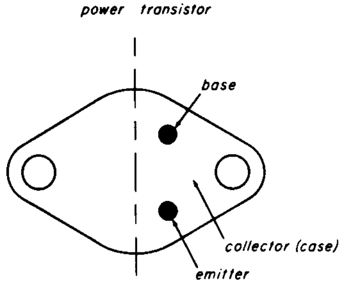


Fig. 19. Base diagram for a popular power-transistor case.

by mounting bolts, or sometimes, a central threaded stud. The base and emitter leads are identified by holding the case so the two wire leads are to the right of center as shown. The base is then the top lead.

Identifying Leads on Unmarked Transistors. Occasionally identifying marks may be obliterated on the transistor case. The leads may then be identified with the few ohmmeter checks shown in Fig. 20. In step 1, ohmmeter checks are made between each pair of leads in *both* the forward and reverse directions. Low readings (below 500 ohms) will be found when the ohmmeter places a forward bias across emitter and collector junctions. The highest forward reading is obtained when the meter is placed across the emitter and collector leads. This establishes the base lead as the one that is *not* involved in the high forward-resistance reading. It is not yet known which of the two remaining leads is the collector or whether the transistor is an NPN or PNP type. Step 2 identifies the transistor type. An ohmmeter check is made between the base and one other lead. If a low-resistance reading is obtained when the negative side of the ohmmeter is connected to the base, the transistor is a PNP type. A low resistance reading when the base is positive indicates an NPN unit.

Step 3 identifies the collector and emitter leads. Assume the type has been identified as a PNP. Two readings are then made between emitter

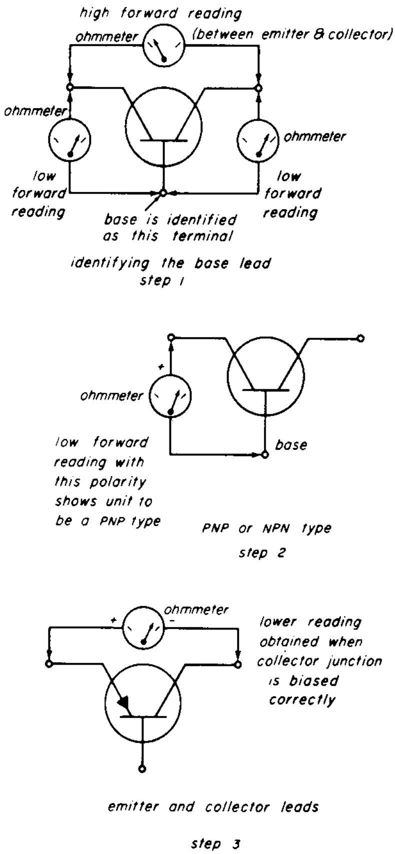


Fig. 20. Identifying the terminals of an unmarked transistor.

and collector, reversing the ohmmeter leads for the second check. One reading shows a lower resistance reading than the other. When the lower resistance reading is obtained the terminal connected to the negative lead of the ohmmeter is the collector. For NPN units the collector is connected to the positive ohmmeter lead when the lower resistance reading is obtained. Step 3 depends upon two factors. First the heavy emitter doping yields a slightly higher alpha when forward bias is applied to the emitter junction. Second, the collector junction is larger in area and therefore has a higher leakage current. Since total leakage current is approximately βI_{CO} more current flows (lower resistance reading) when the applied V_{CE} is in the proper direction.

3. TROUBLESHOOTING INOPERATIVE OR WEAK RECEIVERS

In this section we shall begin the discussion on how to trouble-shoot the popular portable transistor radios. We shall start with a common fault to which straightforward troubleshooting methods may be easily applied: a completely dead receiver.

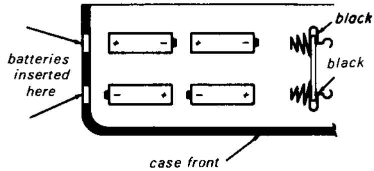
Preliminary Checks. Many causes of complete failure may be localized with a few simple tests. The tests shown below can be made in a few minutes and do not require any measurements to be made other than those available at the battery terminals. These tests are:

1. visual inspection
2. battery voltage
3. total current drain
4. speaker circuit-click test
5. local-oscillator test
6. agc test

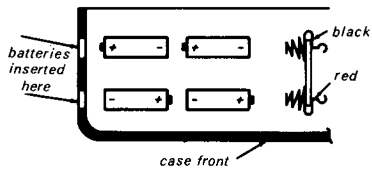
Visual Inspection. Many failures follow accidental dropping of the radio. A visual inspection should be made to locate physical damage. Check the fine wires leading from the loopstick antenna for breaks at tuning capacitor or printed-circuit terminals. Inspect the loopstick for cracks and see if the tuning capacitor turns freely. Leads to the volume control, speaker, earphone jack and battery should also be inspected for breaks. The printed circuit board should be carefully inspected for cracks. Flex the board at several points by applying slight pressure with the eraser end of a pencil. Momentary return of sound indicates a crack in a printed circuit conductor or a poor connection to a component lead. Careful inspection in the sensitive area may locate the fault.

In sets using the small penlite type of battery, make sure the batteries have been inserted correctly. An instruction sheet is usually glued in place near the battery case, or the ends of the case may have polarity marks. Inverting the battery polarity may result in burnout of transistors and electrolytic filter and decoupling capacitors. Note that mercury cells have polarities opposite to those of their penlite counterparts, as shown in Fig. 21. If these are used they must be inserted in the opposite direction to that of the ordinary penlite cell.

Battery Voltage. The more frequent cause of failure is a weak battery. The receiver should operate if the battery voltage is about ⅔ of its normal rating. At lower voltages the receiver may be weak and



(a) 4 mercury cells



(b) 4 penlite cells

Fig. 21. Polarity difference between penlite and mercury cells.

distorted or may motorboat. Complete loss of sound may also result due to failure of the local oscillator. Check the battery voltage under load, by checking terminal voltage while the battery is connected to the receiver with the switch turned on. As an alternative, the battery may be checked by placing an external load resistor across its terminals. Choose a resistor to draw about 20 ma at the rated terminal voltage. If the battery shows a terminal voltage less than ⅔ of its rating under load, replace it with a fresh unit.

Total Current Drain. Many circuit failures alter the d-c operating conditions of one or more stages. A failure which greatly alters the operating point of a single stage may be detected by measuring the total current drain of the receiver. This is done by simply placing a milliammeter in series with the battery leads. Total current drain is given by some manufacturers in the service notes. The figure is usually given for no-signal conditions. A high reading indicates a shorted or leaky filter or decoupling capacitor, or a transistor stage which is conducting too heavily. A low reading is usually the result of an open transistor or a break in the d-c circuit of a stage.

If the normal no-signal current drain is not known it may be estimated roughly by allowing 0.5 ma to 1 ma for the converter stage; 1.5 ma for each i-f and audio-driver stage and 4 ma for a class-B push-pull stage. If a personal portable uses a single class-A output stage, allow about 4 ma for the output stage. The detector, whether a diode or a transistor, draws negligible current under no-signal conditions. The currents given for each stage include the current drawn by the base-bias circuits. The current drain of a six transistor portable using a push-pull output stage and two i-f stages is estimated as follows:

converter	0.5 ma
i-f amplifiers (2)	3 ma
audio driver	1.5 ma
output stage	4 ma
Total	9.0 ma

If the measured current drain is zero, a break is indicated in the supply circuit. The small on-off switches employed in personal portables are a common cause of this fault.

A current reading which differs from the estimated no-signal current drain ±50 percent indicates either a radical change in the operating conditions of a stage, or a shorted decoupling capacitor. Signal-tracing tests may then be dispensed with and the trouble localized by d-c tests alone. Methods for measuring the current drain of individual stages are shown later. If current drain is about normal, the following preliminary checks may localize the fault. If not, more exhaustive checks must be applied.

Speaker-Circuit Click Test. This test localizes faults in the collector circuit of the output stage. Hold the speaker close to the ear and turn the power switch on and off. A definite click should be heard. The mechanical click of the control may mask the click in the speaker. If so, leave the switch on and make and break one of the battery connections. If no click is heard, an open exists either in the speaker, or the primary or secondary of the output transformer, or the connections to the earphone jack. As shown in Fig. 22, a bent or broken normally-closed contact on the earphone

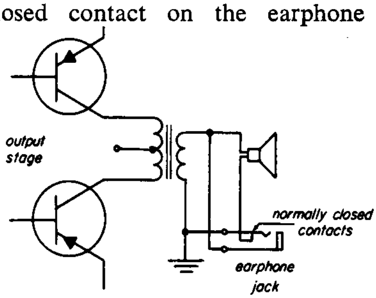


Fig. 22. Speaker-headphone circuit found in most transistor portables.

jack, or a broken terminal connection on the jack is a common cause of trouble.

Local-Oscillator Test. A quick check of local-oscillator operation may be made without opening the case of the receiver. Tune another receiver to a station between 1000 kc and 1500 kc. Place the transistor receiver under test close by, and rotate the tuning dial slowly through the range of 545 kc to 1045 kc. If the oscillator is operating, it radiates signals in the 1000-kc to 1500-kc range. Oscillation is established when the radiated signal beats with the signal to which the auxiliary receiver is tuned. A loud whistle is heard in the speaker of the auxiliary receiver. If no zero beat can be heard, the trouble has been localized to the local oscillator.

Agc Test. The collector current of controlled i-f stages varies from one milliampere to a few tenths of a milliampere when signal conditions vary from very weak to very strong. This change in current can be detected by monitoring the total drain of the receiver. If a dip in total current drain is noted while slowly rotating the tuning dial, the converter, i-f amplifier, detector, and agc systems must be functioning. A trouble is therefore indicated in the audio section. This test may not be conclusive unless there is at least one strong local station. The receiver must be tuned carefully as the drop in total current is small and may be missed if the set is tuned too rapidly.

Note that an *increase* in total current drain when tuning in a station indicates audio drive to the class-B output stage. If a large increase in total drain is noted, and no sound can be heard the trouble is probably in the speaker circuit.

Localizing the Defective Stage. The preliminary checks should have eliminated the local oscillator and the collector circuit of the output stage as possible causes of trouble. The agc test might also have indicated whether the trouble precedes or follows the detector stage. To further localize the trouble to a single state the reeciver chassis has to be removed from the case and signal-tracing or signal-injection methods applied.

Signal-injection methods are the same, in principle, to those employed in vacuum-tube circuits. The ground side of an audio generator is connected to either terminal of the battery. The test signal is then injected through a 0.1-μf capacitor to various points in the audio circuit. Starting at the collector terminal of the output stage, adjust the generator for an output signal that is just audible. The free end of the coupling capacitor is then moved back towards the volume control touching in sequence the points numbered in Fig. 23. Note a signal *voltage* is being injected. All transformers in the transistor radio have a voltage step-down ratio. Thus, when moving the injected signal from point *2* to *3*, a definite drop in volume should be

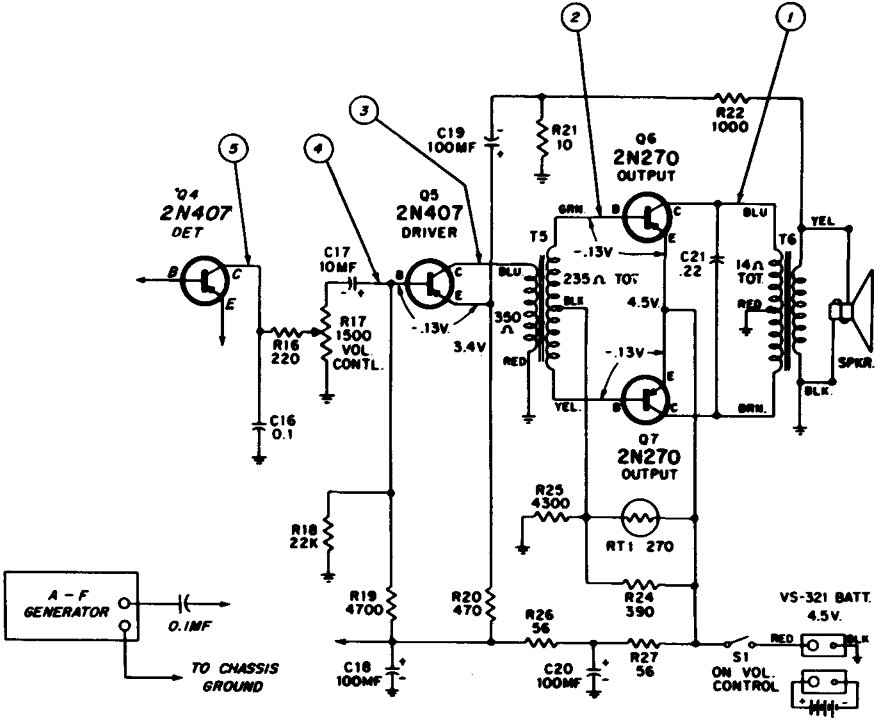


Fig. 23. Signal injection points in the audio circuits.

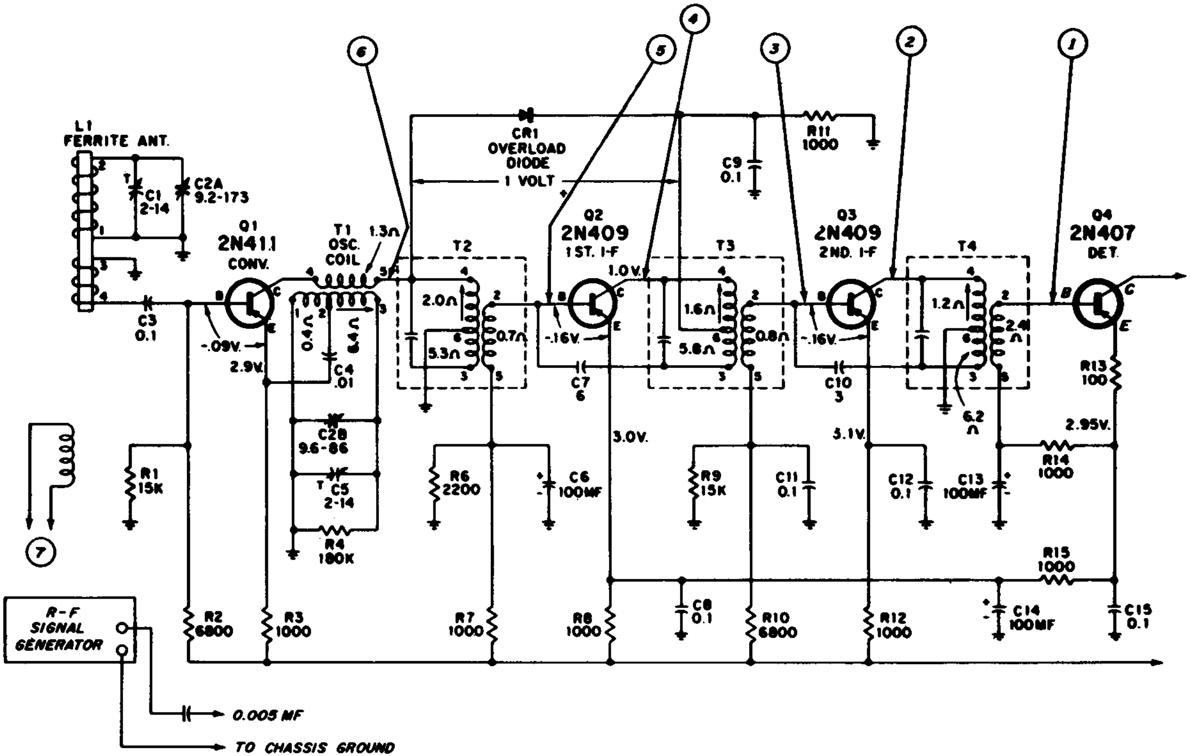


Fig. 24. Signal injection points in the r-f and i-f circuits.

expected. When moving the probe from the collector to base (from point *3* to *4*) a large increase in volume should be heard. Always reduce the attenuator setting of the signal generator to maintain a low output volume, or the effects of agc or signal overload may hide the true change in volume. Failure to produce an output at any point localizes the trouble between the point where the signal was last heard and the point where no output can be produced. D-c voltage checks should then be made to localize the fault to a component.

If an audio output can be produced with a low-level signal applied at the volume control (about 3 millivolts), the audio section is eliminated, and trouble is indicated ahead of the detector. (A method for determining a 3 millivolt signal is given later.) Signal injection should continue as shown in Fig. 24 using a modulated r-f signal generator tuned to 455 kc. A 0.005-μf or larger capacitor should be used for coupling the generator to these stages. The same procedure applies: a drop in volume should be noted when the probe is moved from the secondary of a transformer to the primary; and a definite rise in volume should be noted when moving the injection probe from collector to base of a stage.

Trouble is indicated in the r-f tuning section if a low-level (10 microvolt) signal applied to the base of the converter at 455 kc produces an output. To check the r-f tuning section, the r-f signal should be injected in such a way that the output resistance of the signal generator does not load the resonant tank. A suitable coupling system can be made using a loopstick antenna or an air core coil. The latter may consist of 15 to 20 turns of hookup wire wound upon a two-inch coil form. The coil is connected across the output of the signal generator, and is positioned about 6 inches to 12 inches away from the loopstick antenna in the receiver under test. The amount of coupling may be adjusted by varying the spacing. This coupling arrangement is also very useful for aligning the r-f section of the receiver.

Some technicians prefer the signal-tracing method. This system employs a high-gain amplifier preceded by a detector, and is used to trace the signal from the antenna to the speaker (reverse of signal injection). In this case an r-f signal may be injected by means of the coupling coil. The signal-tracer probe is placed across the loopstick antenna in the receiver and the volume control on the signal tracer adjusted to obtain an output in the headphones. The probe is then moved from point to point back towards the speaker in the reverse order from that shown in Figs. 23 and 24. Here again, remember that the coupling transformers give a voltage step down, so that volume normally drops when going from primary to secondary.

Click Tests. Most technicians are accustomed to injecting signals into vacuum tube audio circuits by the "finger" method. Placing the finger on the arm of the volume control in a vacuum-tube set, for example, produces a loud hum in the speaker. This method does not work at all in transistor circuits because of the very low impedance of the input circuit as compared to that of the vacuum tube's grid circuit.

Another favorite is the click test in which a transient signal is injected by momentarily shorting the grid of a tube to ground (chassis). This test may also be applied in transistor circuits but certain precautions must be observed or damage may result. To illustrate, consider the audio circuits shown in Fig. 25. In *a* of the figure, shorting the base to the chassis (or the ground conductor on the printed-circuit board) removes the forward bias applied to the transistor. A momentary removal of bias cannot damage the transistor and the disturbance produces an audible click in the speaker if this stage and the following stages are functioning. If the same test is applied in the circuit shown in *b* of the figure, burnout of the transistor is the likely result. In this case the short returns the base directly to the negative side of the battery, placing a high forward bias on the emitter junction. Many receivers are wired as shown in *b* of the figure with the negative side of the battery returned to ground.

To insure safe click tests the set up shown in Fig. 26 is recommended. A 10-k ohm to 20-k ohm resistor limits the current flowing in the test probe even if it is accidentally placed in the wrong spot. By returning the probe to the positive side of the battery the base bias of a transistor is *reduced* when the base is touched. If NPN transistors are used, return the probe to the negative side of the battery.

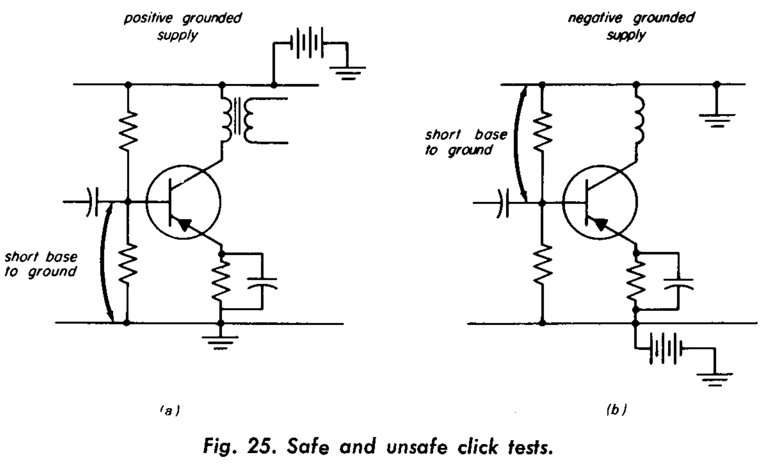


Fig. 25. Safe and unsafe click tests.

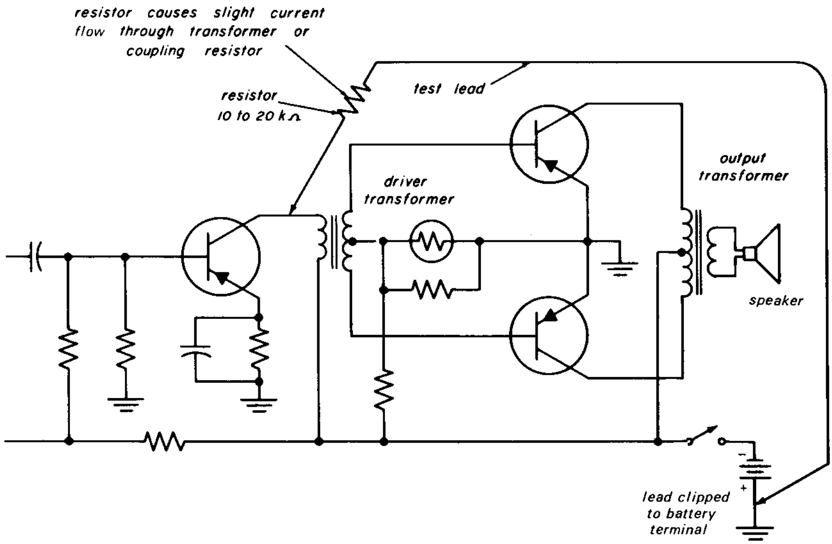


Fig. 26. Making click tests without danger of transistor damage.

The click test is used just as in signal injection by starting at the output stage and working towards the antenna. A break in the signal path is indicated when touching the base of a transistor does not produce a click in the speaker.

Weak Receivers. A receiver that does not produce sufficient volume on some or all stations requires a little more judgment to localize the fault than the clean-cut problem involving a complete break in the signal path. In general, signal-tracing and signal-injection methods are applied in the same fashion, but this time you are looking for a stage whose gain is below normal and not completely dead.

Listening tests can often localize the faulty stage to either the r-f, i-f and detector sections or the audio section. Loss of gain in the audio section produces low volume on all channels. A weak i-f section results in low sensitivity, but strong local signals may produce near normal output, while weaker signals may not be heard at all.

Preparation and Use of a Calibrated Signal Source. Guess work in localizing a weak stage is minimized if measurements of stage gain can be made. This can be done by adding quantitative measurements to the signal-injection method of troubleshooting. To illustrate, an increase in volume is normally noted when the signal injection probe is moved from the collector to the base of a transistor stage. The increase indicates that the stage is working. To change this check to a measurement of stage gain we *could* measure the increase in output voltage. However, this requires the output of the generator to be held constant, and chances are there would soon be overload and agc problems as the probe is moved closer to the antenna. The more acceptable method is to maintain a low and constant output at the receiver terminals and measure the change in input voltage required to produce the same output. A low output is required so that little agc is developed and the receiver operates at or near maximum gain.

To illustrate the procedure, suppose you wish to know the voltage gain of the second i-f amplifier in Fig. 27. The signal is first injected at the base of the detector and the output is adjusted for a just audible output. The voltage at the speaker terminals and the settings of the generator's output attenuator are noted. The probe is then moved to the base of the second i-f stage and the generator's attenuator reset to produce the same reading at the output meter. The gain of the stage is then measured by noting the factor by which the output of the generator is reduced. This factor is the ratio of the two attenuator settings.

Note that voltage ratios and hence voltage gain is measured in the above tests. Power gains are usually considered for transistor circuits. Voltage gains however are easiest to measure, and provide data accurate enough for troubleshooting purposes.

Voltage-gain measurements are easiest to make if the output voltage of the signal generator is known. A few service generators provide a metered output but most do not. If a wide band a-c voltmeter or a d-c VTVM equipped with an r-f probe is available, generator output-voltage measurements

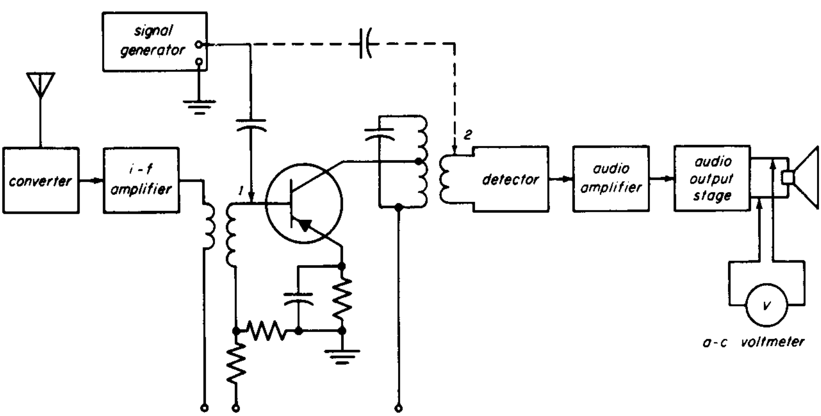


Fig. 27. Setup for measuring the gain of an i-f stage.

in the millivolt and microvolt range may be made by using a suitable attenuator. Attenuators for use with audio generators (600-ohm output) and r-f generators (50-ohm output) are shown in Fig. 28. (The values given are the closest standard values that give the voltage ratio required and add up to the desired total impedance.) The meter reads the input to the attenuators. The meter reading is multiplied by the selected attenuator setting to obtain the output voltage. In the r-f attenuator, moving each switch down attenuates the signal voltage by 1/10th (20-db steps). Each switch turned downwards adds another 20 db of attenuation. Two switches attenuate the signal by 1/100, three give a 1/1000 attenuation, and all four switches attenuate the signal by 1/10,000. Thus a 0.1-volt input to the attenuator yields an output of 0.1/10,000 or 10 microvolts when all the pads are in the circuit.

With the calibrated signal source, a signal may be injected at the base of each stage, starting at the output stage and proceeding towards the converter. The gain of each stage is noted by the ratio of output to input voltage required to produce a previously selected low voltage at the speaker terminals. Fig. 29 shows the input voltage required at the base terminal of each stage to provide the indicated output for a typical receiver. Readings necessarily vary from one type of receiver to another, and the technician

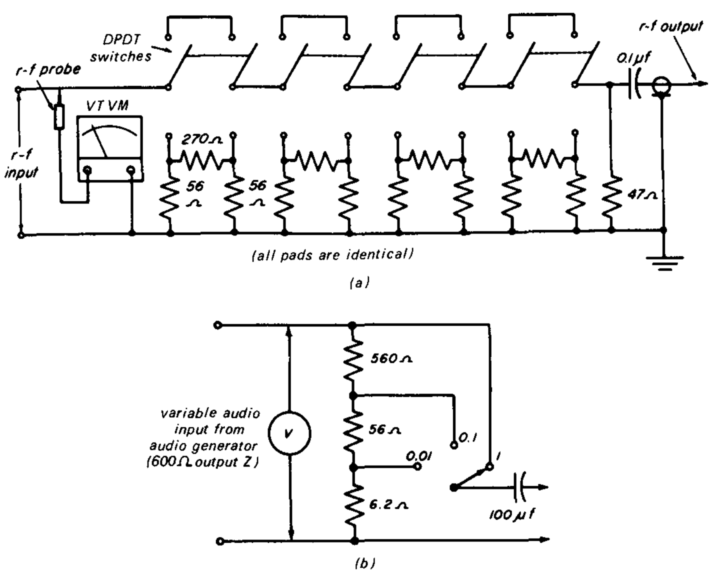


Fig. 28. Calibrated attenuators: (a) r-f type; (b) a-f type.

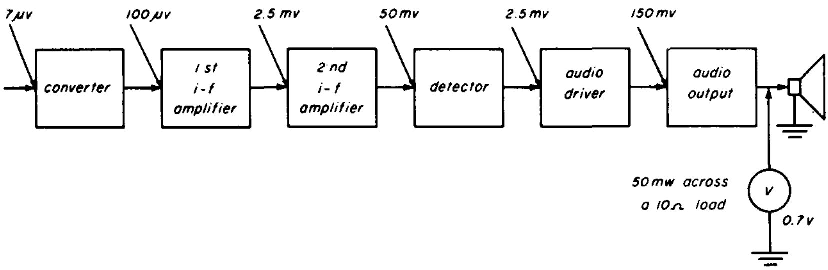


Fig. 29. Input voltage requirements for a typical 6-transistor receiver.

should compile a set of tables such as that shown for every receiver type that he encounters. When no information is available, the approximate voltage gain for various stages may be estimated as: 30 to 60 for audio drivers; 15 to 25 for the second i-f amplifier; 25 to 50 for the first i-f amplifier; and 10 to 30 for the converters. Diode-detector stages usually show a drop in voltage gain of about 10 to 25, while transistor detectors provide some gain.

TV troubles can be localized by the process of elimination. There are only six major "signal paths" or functions in a TV set. These are the r-f, video, and sound-signal paths, vertical and horizontal sweep, and power supply. Localization to one (or more) of these major sections is, in fact, easier to do on a TV set than a radio, primarily because of the "built-in test instrument" in every TV set, the picture tube! The presence of a raster on the screen eliminates the power supply; clear audio eliminates the power supply, r-f, and audio circuits; etc. In other words, the same localization techniques which you used on vacuum tube sets is just as useful in servicing transistorized sets. The difficulties begin *after* you locate the defective section.

4. TROUBLESHOOTING THE DEFECTIVE STAGE

The foregoing tests should localize the cause of a weak or inoperative receiver to a single stage, or perhaps to a group of stages. The next checks should localize the defective component. In vacuum-tube circuitry, the tube in the defective stage is always checked first, either by a tube checker or by substitution. However, in transistor circuitry, the transistor is usually checked as a last resort. Transistors are generally more reliable than other miniature components. Also, in most late-model receivers, the transistors are soldered in place, and removal is inconvenient and exposes the transistor to unnecessary heat damage.

D-C Measurements. As in vacuum tube circuits, most component failures alter the d-c operating conditions of the stage. Once abnormal bias conditions are determined, voltage, current, and resistance checks are used to localize the bad component.

Two voltage measurements, V_{CE} and V_{BE} reveal the general bias conditions of the stage. An abnormal collector voltage reading points to either a faulty component in the collector circuit or a change in the amount of conduction (I_C) of the transistor. If we assume that the transistor itself is satisfactory, then any abnormality in conduction is due to a fault in the biasing system. The measurement of V_{BE} is made to determine the emitter junction bias condition.

In the circuit of Fig. 30, the normal voltage from collector to emitter is seven volts. V_{BE} is 0.1 volt. An error in V_{BE} is ± 0.05 volt

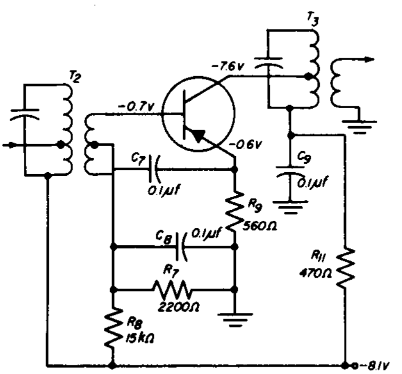


Fig. 30. Typical receiver terminal voltages (to ground) for a grounded-positive supply.

points to a substantial change in operating conditions. To illustrate, suppose previous checks indicate low gain in this stage. A check of V_{CE} shows a higher than normal reading indicating a low voltage drop across R_{11} and low collector current. If a check of V_{BE}, for example, reads 0.04 volts, this points to a reduction in forward bias of the emitter junction. Possible causes are: an increase in the value of R_8 or R_9; a low value R_7; or a leaky C_7 or C_8. Any of these component changes affects a reduction in V_{BE}.

Current Measurements. Measurements of V_{BE} may be difficult to interpret, as very small changes in V_{BE} produce large changes in collector current. Therefore, to definitely establish the bias conditions, a measurement of I_C or I_E is more satisfactory. Collector current can be measured indirectly without breaking into the circuit by measuring the voltage drop across a resistor in the collector or emitter circuit and applying Ohm's law. You may make the simplifying assumption that I_E equals I_C. For example, in Fig. 30, a 0.6-volt drop appears across the 560-ohm emitter resistor, R_9. I_E (and I_C) is therefore 0.6 volt/560 ohms or 1.1 ma. A word of caution here. An open emitter resistor will render the stage inoperative, yet the emitter voltage will be very close to the correct value! This is because the voltmeter will form enough of a d-c return to forward bias the *BE* junction; and the voltage drop across a forward-biased junction is very small.

The collector current may also be determined from the drop across the 470-ohm decoupling resistor, R_{11}. The voltage drop is 8.1 v − 7.6 v or 0.5 v. I_C equals 0.5 v/470 = 1.1 ma. A change in collector current by more than ±50 percent of the indicated value suggests a fault in the bias system. If the stage is controlled by agc, make sure that the current is measured under the conditions indicated in the schematic. On most schematics voltages are given for no-signal conditions.

In PNP circuits in which the positive side of the battery is grounded, the emitter voltage, measured with respect to ground, is a good indication of stage conduction. This voltage is employed in troubleshooting in the same way as the cathode voltage is used to investigate tube conduction in vacuum tube circuits. A higher than normal voltage reading indicates higher than normal collector current, and a low emitter voltage indicates either reduced forward bias applied between emitter and base of a weak transistor. The value of the emitter resistance should be checked when employing this method of determining I_C. Note that changes in the emitter resistance have little effect upon the emitter voltage drop. V_E is always a few tenths of a volt below the voltage applied at the base. But changes in R_E affect I_C materially. An increase in R_E acts to reduce V_{BE} and lowers the collector current.

Grounded-Negative Supplies. In many late model receivers the negative side of the battery is grounded. In PNP circuits using this system, the collector circuit is returned to ground and the supply voltage is applied to the emitter, as shown in Fig. 31. (If vacuum tube circuits were made in this way the plates would be returned to the chassis and the cathodes to the negative supply.) This reversal of polarities leads to some confusion while troubleshooting. Remember that direct measurements of V_{CE} and V_{BE} remain the same regardless of which side of the battery is grounded.

If we wish to make use of the voltages printed in the schematic, which are measured with respect to ground, your thinking must be oriented slightly. For example, as in Fig. 31, an increase in I_C causes the emitter voltage to *decrease* (become less positive) when measured with respect to ground. To make use of the voltage drop across the emitter resistance as an indication

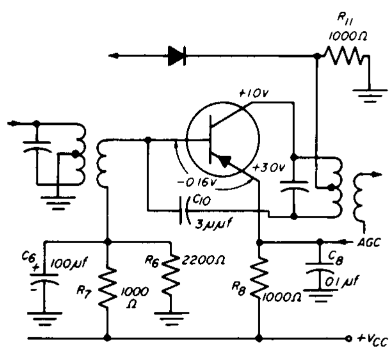


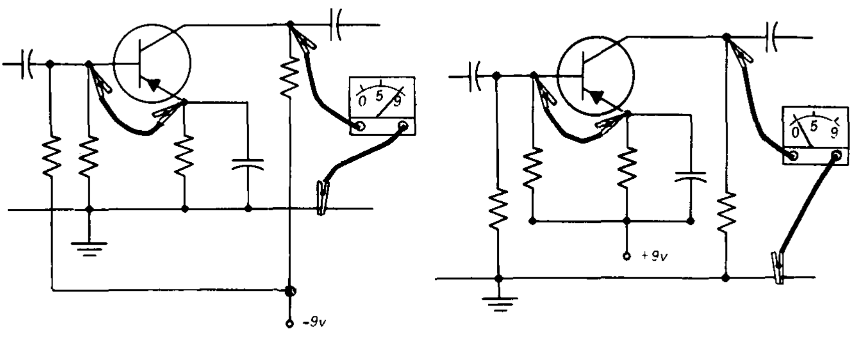
Fig. 31. Typical receiver terminal voltages (to ground) for a grounded-negative supply.

of I_C, the voltmeter should be placed *across the resistor* and not from emitter to ground. In the circuit shown, the voltage drop across the collector decoupling resistor R_{11} is a better indication of I_C. Here an *increase* in I_C results in an *increase* in V_C with *respect to ground.* The voltage drop across R_{11} is one volt giving a normal I_C of one milliaampere for the circuit.

To gain a little familiarity with circuits of the type shown in Fig. 31, consider the effects of a short in C_6. The short returns the base to ground, and since the emitter is returned to the positive supply the forward bias applied to the emitter junction increases. Collector current rises, V_E (with respect to ground) drops and V_C (with respect to ground) rises.

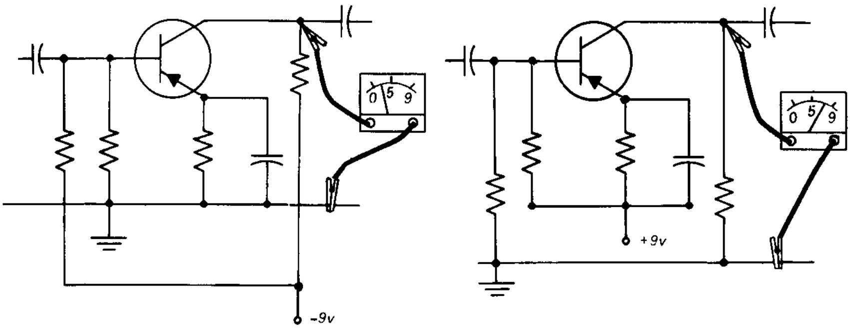
A rough (and quick) indication of whether a transistor is working can be obtained by simply increasing or decreasing the bias on its base-emitter junction. The easiest way is by shorting the base and emitter leads by means of a jumper. This removes bias from the transistor, cutting off the collector current. When the collector current is cut off, the collector voltage of most circuits will increase to the full supply voltage, as in Fig. 32*a*. However, if the collector side of the battery is grounded, and the emitter is returned to the hot side of the battery as in Fig. 32*b*, the collector voltage will be nearly zero when the transistor is cut off.

Conversely, when the base-emitter junction bias is increased by connecting an additional resistor from collector to base, the collector current will increase. This increases the voltage drop across the collector circuit resistance and results in a lower collector voltage for circuits like that shown in Fig. 33*a*, and higher collector voltages for circuits like that shown in Fig. 33*b*. While somewhat crude, these techniques allow a quick evaluation of whether or not a transistor is in working



(a) usual battery connection *(b) reversed battery connection*

Fig. 32. Effects of cutoff on collector voltage.



(a) usual battery connection

(b) reversed battery connection

Fig. 33. Effect of increased bias on collector voltage.

condition. On the basis of the results of these tests, you can go on to another stage or make further tests on this stage.

Direct Measurements of Collector Current. In some cases it may be desirable to measure I_C directly by inserting a milliammeter into the circuit. On a printed circuit board, a conductor may be opened temporarily to insert the meter by cutting the printed conductor with a sharp knife or razor blade. The cut may be bridged with solder after the measurement is completed. When inserting the meter always place the meter in the collector circuit and never in the emitter circuit. If placed in the emitter circuit the internal resistance of the meter upsets the bias network.

Determining Collector Dissipation. All transistors except those in the output stage are operated at a fraction of their maximum collector dissipating ratings. However, abnormal bias conditions may result in excessive power dissipation. Faults in the bias systems of output stages in particular may cause the transistor to operate at dangerous temperatures. This fault is likely in circuits which have required repeated transistor replacement. Collector dissipation is found by multiplying V_{CE} by I_C. I_C may be determined as shown above. In class-B output stages, V_{CE} and I_C should be measured at maximum volume.

Troubles Not Affecting D-C Voltages. In some cases, a stage may be found to have low gain and a check of d-c collector voltage and current reveals normal operating conditions. The tests to be applied should then be aimed at finding component failures that do not affect bias conditions.

In audio circuits, such as that shown in Fig. 34, an open coupling capacitor, C_1, or emitter bypass capacitor, C_2, produces a loss in gain

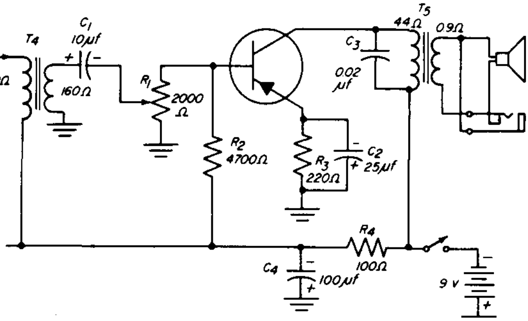


Fig. 34. Typical class-A output stage.

without disturbing d-c bias. The subminiature electrolytic capacitors employed in these circuits often lose a good part of their capacity, which results in similar symptoms. The capacitors may be checked by bridging them with a known, good test capacitor whose capacity is the same as, or larger than that indicated on the schematic.

The capacitor across the primary of the output transformer can develop a short. Since the short is across the very low d-c resistance of the primary no change in collector voltage or current can be detected, but the a-c signal is shorted out.

A frequent cause of trouble is shorted turns in miniature audio transformers. The change in d-c resistance is not great enough to change the collector current, but ohmmeter checks can detect the condition in most cases. The d-c resistance of transformer windings is usually given on the schematic.

Loss of Gain in the I-F Section. If a weak stage has been traced to the i-f section and a check of d-c bias reveals correct operating conditions, the following procedure should be applied. First peak align the i-f section. This may be done quickly and easily by applying a 455-kc signal to the receiver antenna system by means of the radiating coil mentioned earlier. Signal generator output and coil coupling is adjusted to produce a low volume output with the receiver's volume control at maximum. All i-f transformers are then peaked for maximum output. As the output increases, reduce the coupling of the radiating coil to keep the output low. In many cases realignment restores the receiver's sensitivity. If not, circuit faults may be revealed during the alignment procedure. A stage that tunes broadly is probably defective. A quarter-turn of the transformer core each side of the peak position should produce a noticeable change in volume. If detuning by several turns produces only a slight change in volume, excessive loading of the tuned circuit is indicated. (Note that the last transformer should tune more broadly than the others due to the normal loading of the detector circuit.)

Circuit Q may be lowered by shorted turns in the transformer itself or by an open decoupling capacitor. If the decoupling capacitor opens, the decoupling resistor is placed in the tuned circuit as shown in Fig. 35, and acts to load the circuit. To check the decoupling capacitor, bridge it with a known good capacitor of the same value and repeak the transformer. Other bypass capacitors in the case and emitter circuits should be bridged if the stages peak normally but gain can not be brought up to the normal value.

If the above component checks do not restore i-f gain, the i-f transistors should be checked by substitution. In some cases the high-frequency cutoff of the transistors becomes lowered due to some fault in the transistor. The technician has no convenient check for frequency cutoff and must resort to substitution. Always replace with an exact replacement transistor, and repeak the i-f amplifier after the new transistor has been installed.

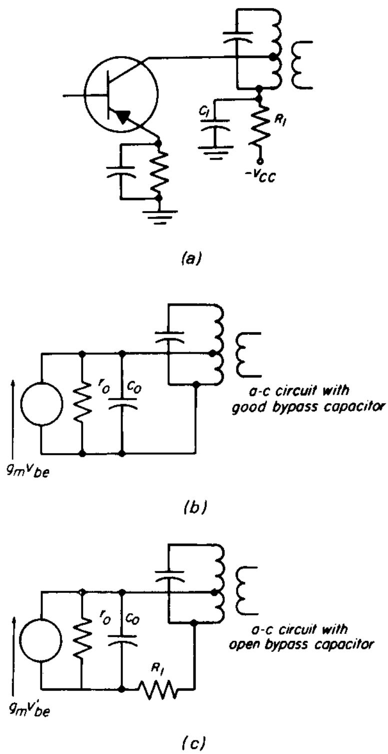


Fig. 35. Effect of an open decoupling capacitor on the collector tank circuit.

Overload Diodes. Many receivers employ an overload diode circuit such as that shown in Fig. 36, to reduce the gain of the i-f section when very strong signals are received. Normally the diode is reverse biased until agc action causes the collector voltage of V_2 to become more negative than that of V_1. Conduction of the diode then loads the primary of T_1 and gain drops considerably. In troubleshooting this

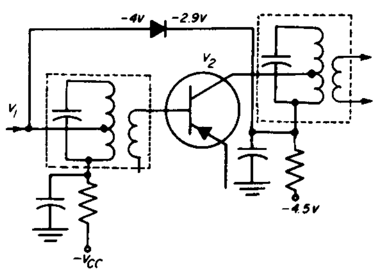


Fig. 36. Checking the reverse-bias voltage on the overload diode.

circuit, make sure the diode is reverse biased under no signal conditions. The anode should be negative with respect to the cathode by a volt or more under no-signal conditions. Typical no-signal voltages on each side of the diode are shown in the figure. A shorted diode causes a severe reduction in gain under any signal condition. An open diode results in distortion when very strong signals are received.

Checking the Local Oscillator. The simple radiation test described earlier only checks to see if the oscillator is operating at the low end of the band. In some cases the oscillator may quit at higher frequencies due to a drop in the cutoff frequency of the converter transistor. (This fault also occurs if an i-f transistor is placed in the converter circuit by mistake.) In many cases this condition develops when the battery voltage drops slightly below its rated value. To check the operation of the oscillator a test is required that can be applied at any frequency and that gives at least a relative indication of oscillator output. A sensitive wide band a-c voltmeter or an oscilloscope can be used to measure the voltage developed at the collector of the oscillator or at the injection point (base or emitter) of the converter. Emitter injection is shown in Fig. 37. A typical voltage measurement at either the base or emitter injection point is 0.2 to 0.7 volt peak-to-peak.

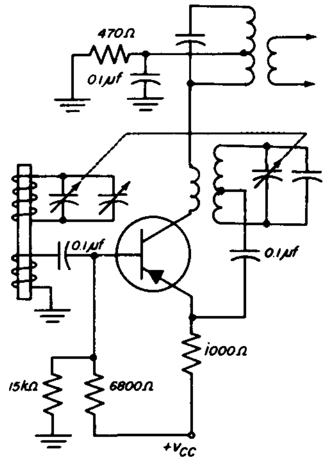


Fig. 37. Representative converter circuit.

In vacuum tube circuits, oscillators are tested by measuring signal or grid leak bias, an easy d-c measurement. Grid leak bias is proportional to signal strength and drops to zero if the oscillator quits. Signal bias is also used in transistor oscillators but its measurement must be interpreted differently. In the transistor oscillator a forward bias is applied to the base emitter junction and the signal bias, developed by an RC network in the base or emitter leg acts to *reduce* this bias as oscillator strength increases. Note the difference: the bias voltage is *always present* and you must look for a *change*. In any circuit V_{BE} may be measured directly with a VTVM. In oscillator circuits V_{BE} is normally low. To check if the oscillator is stopped, simply place a wet finger on the stator plates of the oscillator section of the tuning capacitor. If the oscillator had been working the forward V_{BE} voltage shows an *increase*. If the oscillator was not working, no change in bias voltage is noted. This check can also be made by measuring the change in I_C. In the circuit of Fig. 37, I_C may be checked by measuring the voltage drop across the collector decoupling resistor. If the voltage drop across this resistor, R_1, rises when the oscillator is loaded the oscillator was working.

A very convenient way of checking the oscillator is by means of a simple r-f wavemeter, shown in Fig. 38. The wavemeter detects oscillator signals coupled into its coil by transformer action. It is made from a ferrite rod antenna equipped with two tuning capacitances that tune the wavemeter at two check points in the oscillator band. (A variable capacitor may also be used to tune the wavemeter continuously through the band.) The instrument shown checks the oscillator at 1095 kc and 1695 kc. These are the oscillator frequencies when the set is tuned to 640 kc and 1240 kc respectively (the Conelrad frequencies). The wavemeter is calibrated by holding

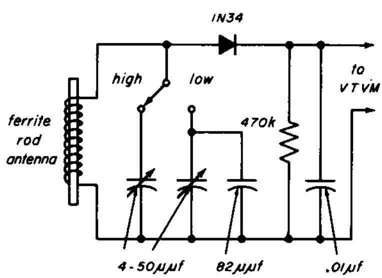


Fig. 38. A wavemeter to check the operation of the local oscillator.

the loopstick near the unshielded oscillator coil of a good receiver. The set is tuned to 640 kc and 1240 kc and the appropriate trimmers peaked for maximum reading on the voltmeter. To use the wavemeter hold the loopstick about an inch away from the oscillator coil and tune the set to the check frequencies. If the oscillator is working, a voltage reading of 0.5 volts to five volts is registered on the voltmeter.

Oscillator failure may be due to a faulty or incorrectly wired oscillator coil, a faulty i-f transformer coupling the converter to the first i-f amplifier, or an open bypass capacitor. Shorting capacitor plates or poor ground connections at the tuning capacitor are also common causes of failure. A faulty transistor is indicated if the oscillator quits at the high end of the band.

5. DISTORTION AND REGENERATION

Distorted, mushy or raspy sound and squeals, tweets, and birdies are among the most common faults in portable transistor radios.

Localizing the Cause of Distortion. Distortion usually occurs in the audio section and is due to excessive signal or a shift in the d-c operating point in an audio stage. In some cases distorting may originate in the detector stage or may be the result of regeneration in the i-f section. Listening tests can help to localize the trouble to a major section. First note the distortion carefully at low and high volume settings. If no change in distortion occurs, if sound is equally bad at all volume levels, distortion is probably occurring ahead of the volume control. The detector is the likely suspect in this case. If the distortion gets worse as the volume control is advanced, trouble in the audio section is indicated. The likely cause is a stage being driven into cutoff or saturation due to a shift in operating bias. If distortion seems worse at low volume settings, crossover distortion in the push-pull output stage is indicated.

If distortion seems the same at all volume levels check the sound for different stations. An increase of distortion on weak signals points to a lack of forward bias at the detector. Normally a slight forward bias is applied to a diode detector to eliminate the non-linearity that exists at very low values of i-f voltage. If this bias is absent or if the diode has been installed backwards distortion occurs which appears worse at low signal levels. If the diode has been installed backwards distortion also occurs on strong signals due to failure of the agc system. A forward bias of about 0.1 volt should be measured across the diode with no signal applied.

If tuning seems very critical and breaks into a hiss or whistle as the tuning dial is rotated, regeneration is occurring in the r-f or i-f section. Procedures for treating this condition will be given shortly.

Distortion in the Audio Section. If distortion is suspected in the audio section, d-c measurements of V_{CE} and V_{BE} should be made in the output and driver stages to detect any abnormalities in the bias conditions. Most cases of distortion are the result of a shift in bias conditions that cause a stage to be driven to cutoff or saturation. Particular attention should be paid to V_{BE} to detect changes in bias. A low forward bias of 0.09 to 0.12 volt is applied to the transistors in class-B output stages to eliminate crossover distortion. If this bias is low or absent crossover distortion occurs. This form of distortion is worse at low volume settings. Any abnormality in bias voltage should be followed up by resistance checks of the components that establish bias for the stage.

An oscilloscope offers the best means for detecting and localizing distortion in the audio section. To make a check of the audio section. an audio signal at 400 or 1000 cycles is injected at the input to the volume control. If the set employs a diode detector, the diode should be disconnected from the circuit by unsoldering one of its leads, or by making a knife cut in its printed conductor. This is necessary to prevent rectification of the output of the audio generator. Connect the oscilloscope across the speaker terminals and advance the output of the generator until a waveform is observed. Now if the audio section is functioning normally it should be capable of producing its rated output without any sign of clipping or distortion on the output waveform. Advance the output of the audio generator until the rated output is produced. To determine this it is necessary to know the undistorted power output rating of the receiver (150 to 300 milliwatts for most portables employing push-pull output stages) and the impedance of the speaker voice coil. The voltage developed across the speaker coil is then calculated by:

$$V_{rms} = \sqrt{P_{ac} \times R}$$

For a typical receiver equipped with a 3.2-ohm speaker and rated at 250 milliwatts, the output voltage should be 0.9 volt. This may be read with an a-c voltmeter across the speaker terminals, or the CRO may be calibrated to measure 2.6 volts peak-to-peak.

If the waveform shows signs of distortion before the rated output is reached, check the waveforms at the collector terminals of the audio stages preceding the output stage. Distortion in these stages usually appears as clipping, as in Fig. 39. Once distortion has been localized to a stage, the bias conditions of that stage should be checked. If the bias circuitry appears normal the transistor should be removed and checked for leakage and gain, or a new transistor substituted.

If the waveform appears normal at the collector terminal of the driver stage, distortion is occurring in the output stage. If the output stage is operated in class-B push-pull, two forms of distortion may take place. One is crossover distortion resulting from operation close to the knee of the V_{BE}—I_C characteristic at low values of V_{BE}. It results from insufficient base bias on the output transistors. The output waveform appears as shown in Fig. 40. Distortion may also occur due to unbalance between the two output transistors. Balance is checked by measuring the waveform at the collector of both output transistors. If the collector signal voltages are unequal, either the transistors are not matched or the driver or output transformers are defective. An ohmmeter check of both transformer windings should be made. If both halves of the center-tapped winding show equal resistances, from the outer terminals to the centertap, the

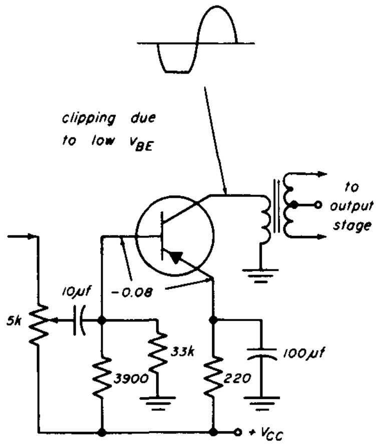


Fig. 39. Distortion due to insufficient base-emitter bias.

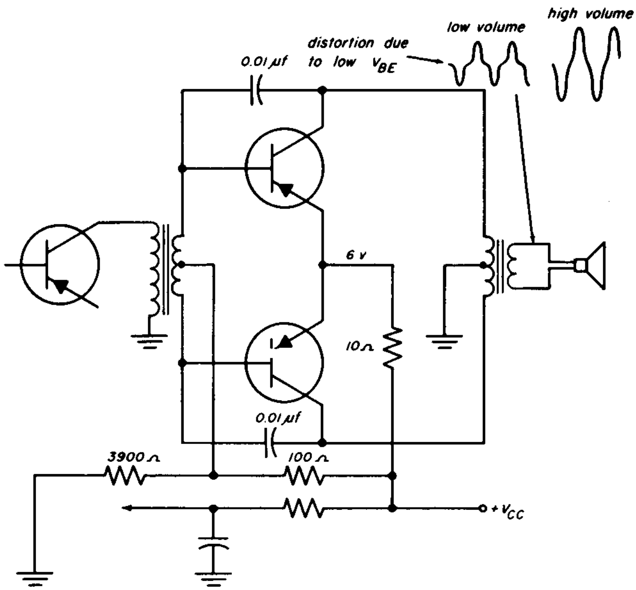


Fig. 40. Crossover distortion in class-B push-pull stages.

transformers are probably satisfactory, and one or both of the output transistors should be replaced.

Other possible causes of distortion are a stuck or rubbing speaker voice coil, or magnetic saturation occurring in the audio transformers. The speaker should be checked in the usual manner. Saturation sometimes occurs in the transformers due to the very small dimensions of these components. Check to make sure the primary d-c currents are not excessive. In rare cases the transformers may have to be replaced.

Regeneration. Squeal, hiss, motorboating, and very critical tuning are all symptoms of regeneration in the receiver. Regeneration may be the result of feedback in a single stage, or a feedback loop in which several stages may be included. The trouble may be localized to either the r-f, i-f, or audio section by rotating the volume control. If the squeal persists without changing in frequency but only in volume, it is reasonably safe to assume that the regeneration loop is confined to the r-f amplifier-i-f amplifier-detector-agc section of the receiver. If the pitch of the sound can be altered or if regeneration stops or starts at different settings of the volume control, the audio section is included in the feedback loop.

The most common cause of regeneration is a high resistance ground connection. In transistor circuits, any resistance in the common ground or supply voltage returns constitutes a common impedance across which feedback signals may be developed. Connections must have a much lower impedance than can be tolerated in vacuum tube circuits. The reason is that a few ohms is low compared to the input and output impedances of vacuum tubes, but is appreciable when compared to transistor impedances.

If feedback includes the audio section, a check should be made of all ground connections, battery connections and the battery itself. Weak batteries have high internal resistance and are a common cause of regeneration. Electrolytic capacitors used to decouple the supply voltage should also be checked by bridging them with a substitute capacitor. Ground connections to the volume control, battery leads, and transformer lugs should be resoldered, and battery terminals cleaned if corrosion is present.

If regeneration is localized to the r-f amplifier-agc section, an inspection and resoldering of all ground connections to tuning capacitor lugs and transformer shield cans should be made. Check bypass capacitors, by substitution (bridging), in the converter and i-f stages, and also check the bypass capacitors used to decouple the agc line.

Oscillation may occur in a single i-f stage due to excessive gain or insufficient neutralization. Oscillation may be detected in the i-f stages by removing the input signal and measuring the r-f voltage at the collector terminals of each stage with a wideband oscilloscope or a VTVM equipped with an r-f probe. Presence of an r-f signal with no antenna signal indicates oscillation. To completely remove any input signals solder a 330-ohm resistor temporarily across the high-impedance terminals (those connected across the tuning capacitor) of the local oscillator coil. To localize the oscillating stage each stage in turn is momentarily loaded down with another 330-ohm resistor, as in Fig. 41. First, place the other 330-ohm resistor across the high impedance terminals of T_1. If oscillation can still be detected, the converter stage is not included in the feedback loop and the trouble is localized to the two i-f stages. Now place the 330-ohm resistor across the high impedance windings of T_2 and T_3 in turn. If oscillation can still be detected when T_2 is loaded, the final stage is oscillating. If oscillation is detected when T_3 is loaded, the first i-f stage is oscillating. If oscillation stops when either T_2 or T_3 are loaded, both stages are included in the feedback path and the feedback signal is developed across an impedance common to both stages. Recheck bypass capacitors and ground connections. In some cases various ground points on the printed circuit board may have to be connected together with additional wire jumpers to insure a low impedance ground.

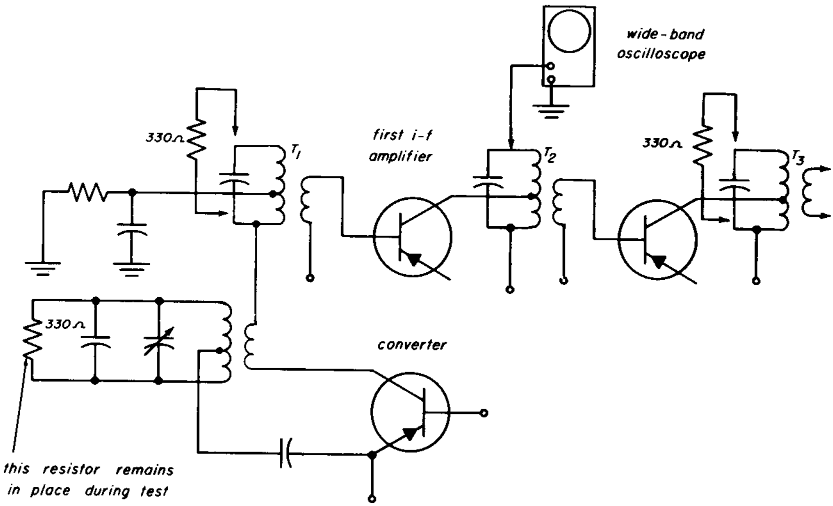


Fig. 41. Localizing the cause of oscillation in the i-f amplifier.

If a single stage is oscillating, the fault may be excessive gain. A "hot" transistor or a converter transistor installed in an i-f stage is a frequent cause of trouble. Check to make sure the transistors are of the type specified in the schematic. In some cases interchanging the two transistors in the i-f section may cure the trouble. As a last resort, a slight detuning of the i-f amplifier may prevent oscillation without producing a noticeable drop in receiver gain.

If the i-f amplifier makes use of a neutralizing system to suppress oscillations, an increase in the size of the neutralizing capacitor may be required. A small increase in capacity may be secured by installing a "gimmick" across the neutralizing capacitor, as in Fig. 42. About 2 inches of insulated hookup wire are soldered to each lead of the neutralizing capacitor. The free ends are twisted together as shown. Twist together only enough of the leads to cause oscillations to stop, and cut off the excess.

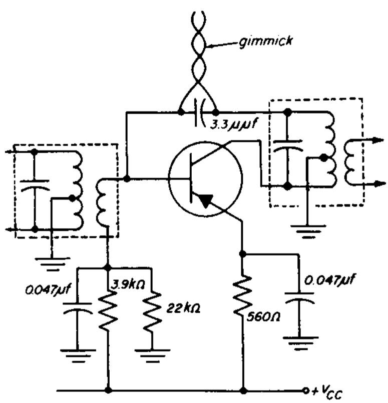


Fig. 42. Increasing the value of the neutralizing capacitor.

Noise. Another form of distortion is excessive noise. This is evidenced by hiss or rushing sounds in the speaker. In vacuum tube radios this trouble is practically nonexistent and can only result from excessive noise being generated in the r-f or converter circuit. In the transistor receiver, any stage, including the audio amplifier, can produce large amounts of noise. A noisy audio amplifier is evidenced by hiss in the speaker which is not affected by the volume control setting. The cause is usually a faulty transistor, and the trouble develops in transistors having excessive leakage currents. If loud noise is the symptom, look for a stage which has abnormally high values of collector current.

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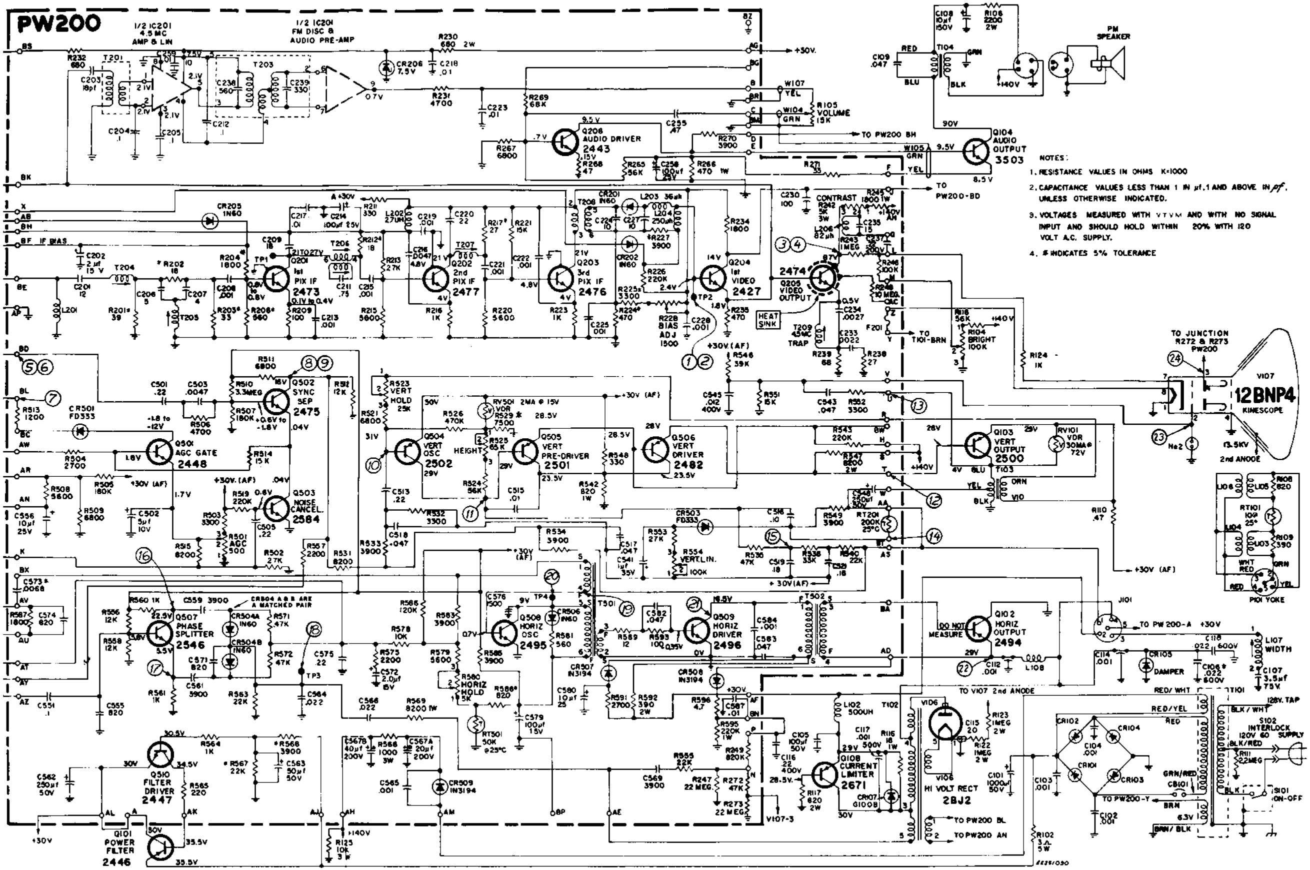
RCA SK-Series replacement semiconductors have electrical characteristics comparable with, or superior to, those of the devices they replace. In most cases the recommended SK-Series replacement transistor or rectifier can be installed without changes in mechanical mounting arrangements, circuit wiring, or operating conditions.

The 23 transistors in the RCA SK-Series can replace more than 6,500 types of transistors used in entertainment electronic equipment, including U.S.A. industry-standard EIA types, foreign types, and types identified only by device-manufacturers' or equipment-manufacturers' parts numbers. The silicon rectifiers in this series can replace more than 2300 selenium and silicon rectifiers having comparable ratings and used in entertainment type equipment.



SCHEMATIC DIAGRAM OF THE HSK T1 MAIN CHASSIS