DEPARTMENT OF WEAPONS TRAINING
LOWRY AIR FORCE BASE
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TRANSISTORS

IN SERVICE TRAINING COURSE

CHAPTER 6

TRANSISTOR CIRCUITS

17 October 1961

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CHAPTER 6

TRANSISTOR CIRCUITS

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Abstract: The basic differences between common emitter, common base, and common collector amplifiers are explained and justified. Biasing circuits are given, and the effects of temperature on transistor circuits are shown. Temperature compensation and bias stabilization techniques are covered. Small-signal, direct-coupled, and power amplifiers are described; practical circuits and coupling methods are given. Furthermore, the special problems associated with high-frequency operation are considered, and typical tuned and untuned (video) amplifiers are discussed. Both sine-wave and non-sinusoidal oscillators are treated. The latter category includes blocking oscillators, sawtooth generators, and multivibrators. Finally, the operation of the transistor as a switching device, rather than as an amplifier, is analyzed.

INTRODUCTION

The previous chapters have been devoted to the physics of semi-conductor devices. This chapter will discuss their use in practical circuits. The transistor will receive the greatest attention because of its obvious importance. Nonetheless, other devices will be mentioned in conjunction with transistor circuits.

Frequently in the analysis of transistor circuits, transistors are compared with vacuum tubes. Although the transistor, like the vacuum tube, is a three terminal amplifying device, it does have many peculiarities. Among these are:

(1) The transistor is not strictly a voltage amplifier. The input impedance is often quite low so an appreciable current must be supplied by the driving source.

(2) The voltages and impedances encountered with transistors are considerably lower.

(3) The transistor has a far greater temperature sensitivity than the vacuum tube. Hence, a considerable portion of its circuitry may be devoted to temperature compensation.

(4) The transistor has more internal feedback which means that the input characteristics are affected by the load impedance and the output characteristics are affected by generator impedance.

These and other differences limit the value of vacuum tube-transistor analogies. Therefore, they will be avoided in this chapter.

BASIC AMPLIFIER CIRCUITS

Before going into transistor circuitry, the basic amplifier circuits will be discussed. These are the common base, common emitter, and common collector configurations shown in Figure 6.1. In each of these circuits biasing potentials are provided to reverse bias the collector-base junction (about 20 volts) and forward bias the emitter-base junction (about 0.5 volt). In all cases amplification takes place because the input signal varies the forward bias on the emitter-base junction which causes variations in the collector current.

Common Emitter. Perhaps the most straightforward of these circuits is the common emitter amplifier. The input signal is applied between the emitter and base; it varies the forward bias on the emitter junction. The resulting variations in collector current cause an output voltage to be developed across the load
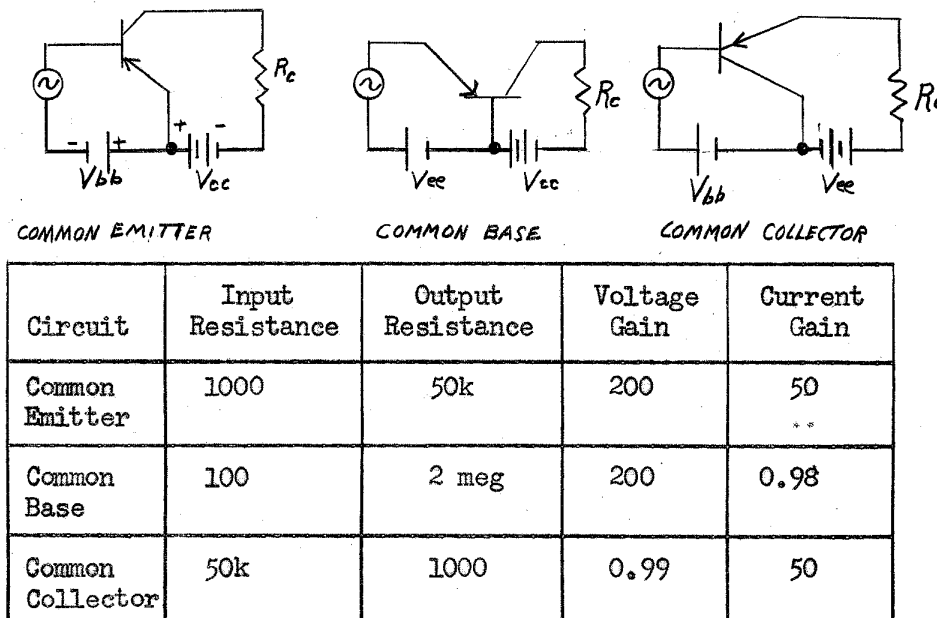
COMMON EMITTER   COMMON BASE   COMMON COLLECTOR

Circuit	Input Resistance	Output Resistance	Voltage Gain	Current Gain
Common Emitter	1000	50k	200	50
Common Base	100	2 meg	200	0.98
Common Collector	50k	1000	0.99	50

Figure 6.1. Basic Amplifier Circuits and Approximate Characteristics.

resistance. Only the base current flows in the input circuit, and it is small compared to the collector current so a current gain will be realized. The voltage gain of the common emitter amplifier can be high because small changes of input voltage across the forward biased emitter junction can produce large changes in collector current. The changing collector current can then produce a comparatively large voltage swing across the load if its impedance is sufficiently high. Current gains of 50 and voltage gains of 200 are not uncommon in practice.

As can be seen from Figure 6.2, the collector current of a common emitter stage is primarily determined by the base current and not too greatly affected by collector voltage. The collector current is affected somewhat because increasing collector voltage widens the collector depletion region and decreases base width. This increases the current gain and, consequently, the collector current. This effect is not too great so the output resistance ($R_o = \frac{\Delta V_c}{\Delta I_c}$) of a common emitter amplifier is moderately high. On the other hand, the base current changes rather rapidly with base voltage. Hence, the input resistance ($R_{IN} = \frac{\Delta V_b}{\Delta I_b}$) is moderately low.

Common Base. In the common base configuration the input signal is applied to the emitter and the output is taken from the collector. The base is the common terminal. An a.c. signal applied to the input will vary the emitter-base voltage and, therefore, the collector current. The changes in collector current will be slightly less than the corresponding changes in emitter current - differing, of course, by the base current. Hence, the current gain will be slightly less than one. If the changing collector current passes through a high impedance load, relatively large variations in collector voltage will result. It follows, then, that the voltage gain of the common base amplifier can be large. Current gains of 0.98 and voltage gains greater than 200 are not unusual with the configuration.
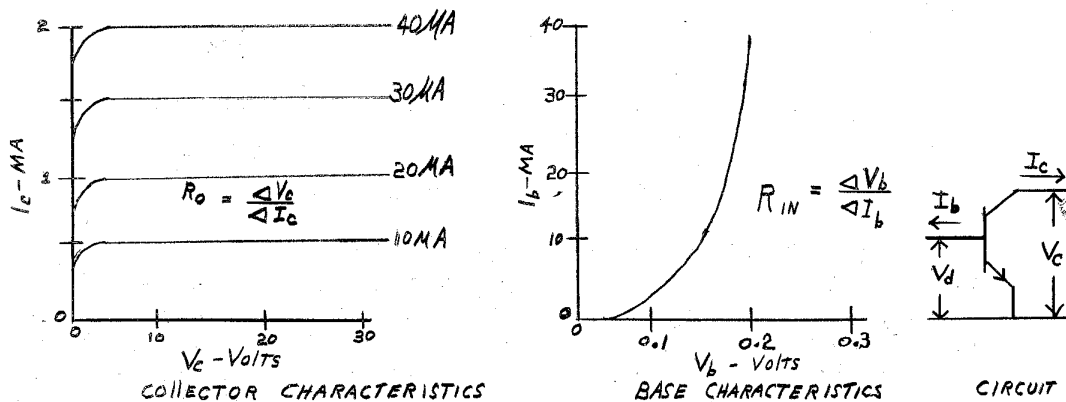

Figure 6.2. Common Emitter

The input resistance ($R_{IN} = \frac{\Delta V_c}{\Delta I_c}$) of the common base amplifier is considerably less than that of a common emitter amplifier as can be seen from Figure 6.3. The change in emitter current for a given change of emitter-base voltage is greater than the change in base current. The emitter current is approximately equal to the base current times the common emitter current gain so the input resistance is about equal to the common emitter input resistance divided by this current gain.

The effect of base width on common base current gain is minimal. Although significant changes in common emitter current gain are produced by varying collector voltage, they are reflected as comparatively small changes in common base current gain. Since the collector current is affected less by collector voltage in the common base configuration, the output resistance ($R_0 = \frac{\Delta V_c}{\Delta I_c}$) is much higher. It turns out that the common base output resistance

is about equal to the common emitter output resistance times the
common emitter current gain. In other words, since the effect
of collector voltage on the common base characteristics is so
small, the output resistance is nearly equal to the resistance
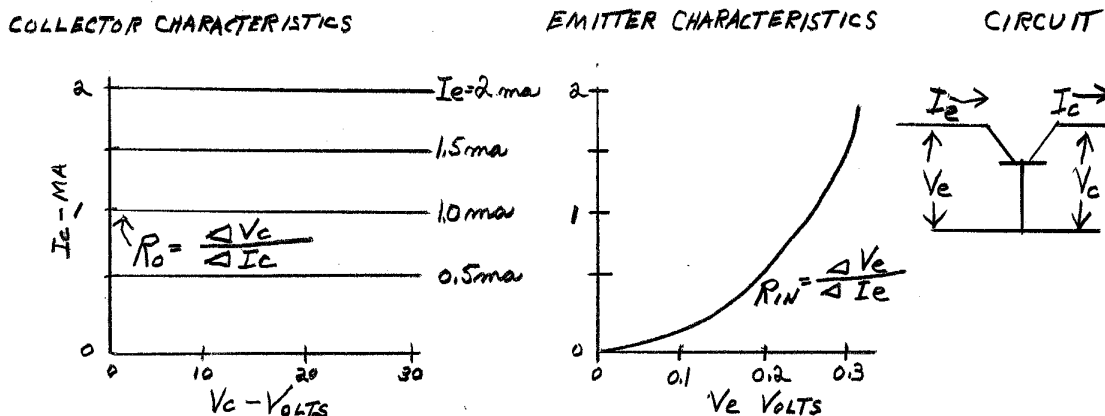of the reverse biased collector junction alone.

COLLECTOR CHARACTERISTICS   EMITTER CHARACTERISTICS   CIRCUIT

Figure 6.3. Common Base

Common Collector. With the common collector configuration, the
input signal is applied to the base, and the load is connected to
the emitter. As before, the amplifying action takes place when the
input signal varies the forward bias on the emitter-base junction.
A changing base current will vary the emitter current and produce
an a.c. voltage across the load. The emitter current will be consid-
erably greater than the base current so the circuit will have a rather
high current gain - slightly greater than that of a common emitter
amplifier. The voltage gain, however, will be somewhat less than
unity.

Considering the circuit shown in Figure 6.4, if the base current
is increased by an increased input voltage, the emitter current will
also rise. This will raise the voltage drop across the load resistance.
In addition, the larger base current will also increase the emitter-
base voltage. Now from Figure 6.4 it can be seen that the increase
in input voltage required to produce this effect will be equal to
the voltage change across the load plus the increase in emitter-base
voltage. Hence, the change in output voltage will be less than the
change in input voltage, differing by the change in emitter-base
voltage. Since the emitter-base voltage will be small in comparison
to the output voltage, the voltage gain will be only slightly less
than one. Current gains of 50 and voltage gains of 0.99 are practical
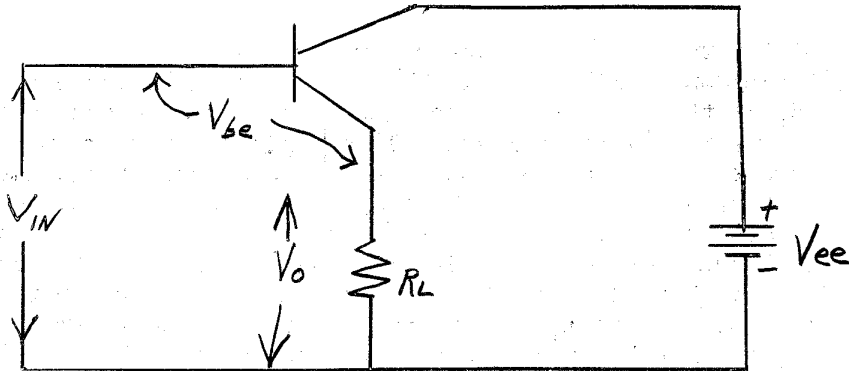with this circuit.

Figure 6.4. Common Collector Amplifier

The input resistance ($R_{IN} = \frac{\Delta V_b}{\Delta I_b}$) of the common collector amplifier is approximately equal to the product of the current gain and the load resistance. The output voltage is nearly equal to the input voltage, so for a given a.c. input voltage, the load current and, consequently, the base current will depend on the load resistance. As the load resistance is decreased, the base current must increase. For very low load resistances the input resistance will be equal to that of a common emitter amplifier, while for high load resistances the input impedance may be several megohms. The latter case is illustrated in Figure 6.5.
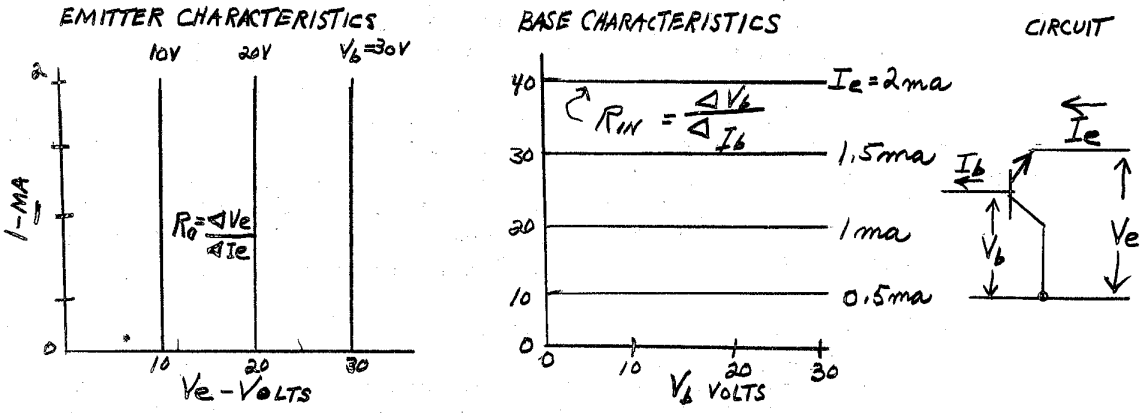
Figure 6.5. Common Collector

Similarly, the output resistance ($R_o = \frac{\Delta V_e}{\Delta I_e}$) is about equal to the source resistance divided by the current gain. If the base is fed by a constant voltage source (low impedance) a change in emitter voltage will be felt entirely across the emitter base junction. Hence, small

6-5

changes in emitter voltage will produce large changes in emitter current so the output impedance will be low. This is shown in Figure 6.5. On the other hand, if the base is fed from a constant current source (high impedance) changes in emitter voltage will not produce any great changes in emitter current. In this case, the output resistance is equal to that of a common emitter amplifier; and the output characteristic curves will be similar to those of a common emitter stage. For intermediate values of source resistance, a change in emitter voltage will be felt almost undiminished on the base through the forward biased emitter junction. This change in base voltage will vary the base current in inverse proportion to the generator resistance. The resulting variations in emitter current will, in turn, be proportional to the current gain. Further analysis will show that the output resistance is roughly equal to the source resistance divided by the current gain.

This interaction between the input and output circuits is not confined to the common collector configuration. It is present to a lesser degree in both the common emitter and common base circuits. This phenomenon is the result of internal feedback within the transistor. It will be discussed further in a later section.

Conclusions. The common emitter amplifier exhibits both a voltage and a current gain; it has the highest power gain of all three configurations. As a result, the common emitter amplifier is used where high gain per stage is required. In addition, the moderately low input impedance and the moderately high output impedance of the common emitter amplifier do not create too much of an impedance mismatch when several stages are cascaded without impedance matching devices. These characteristics favor the use of the common emitter in most general purpose applications.

The power gain of the common base amplifier is between that of the common emitter and common collector amplifiers. The current gain of this type approaches unity, but the voltage gain can be quite high. The low input impedance and high output impedance of a common base stage necessitates impedance matching if the stages are to be cascaded but are sometimes useful in special cases.

Near the maximum operating frequencies of a transistor, the gain of the common base amplifier is about equal to that of the common emitter amplifier. However, the common base stage has the additional advantage that neutralization is not required. Consequently, the common base configuration is most frequently used in high frequency amplifier circuits.

The common collector amplifier has the lowest power gain of the three types. The current gain is high, but the voltage gain is less than one. The common collector amplifier is essentially a common emitter stage with 100 percent negative voltage feedback. Therefore,

the signal distortion will be small even at high levels if it is driven by a low impedance source.

The common collector configuration finds most frequent application as a high impedance input stage, a driver where low output impedance is required, and as a power amplifier where large voltage swings are needed with a minimum of distortion.

BIASING METHODS

The biasing of transistor circuits is not as simple as that of vacuum tube circuits. In most cases, some form of bias stabilization or temperature compensation is required. The problem with transistors is that the collector current tends to increase rapidly with temperature. There are two principle causes of this temperature instability: the increase of the reverse saturation current of the collector junction with temperature and the decrease in emitter junction impedance with increasing temperature. The current gain frequently increases with temperature and adds to the problem.

Reverse saturation current is the current that flows across the reverse biased collector junction. It is produced by thermally generated minority carriers that are swept across the junction so it will increase with temperature. Figure 6.6 shows the increase in reverse saturation current with temperature for typical germanium and silicon transistors.

The saturation current is only troublesome at high temperatures. For example, if the germanium transistor in Figure 6.6 is biased to a collector current of 1 ma, it can be seen that the saturation current will contribute little to the total collector current until a temperature of about 50°C is reached. However, at higher temperatures the saturation current becomes an appreciable portion of the collector current. When this happens, the bias must be decreased with increasing temperature in order to maintain the collector current constant. This compensation is effective until the saturation current becomes equal to or greater than the quiescent collector current. At these higher temperatures, the input circuit loses control of the collector current since the saturation current is independent of the emitter junction bias. Therefore the transistor ceases to function as an amplifier. This places an upper limit of about 90°C on germanium transistors. The lower saturation current of silicon transistors permits operation up to about 175°C.

With the common emitter configuration, the effect of the reverse saturation current is even more serious. If the base is open circuited, the minority carriers swept across the collector junction will cause a charge unbalance in the base and will forward bias the emitter junction. Hence, additional carriers will be injected into the base and will contribute to the collector current. The collector current will then be
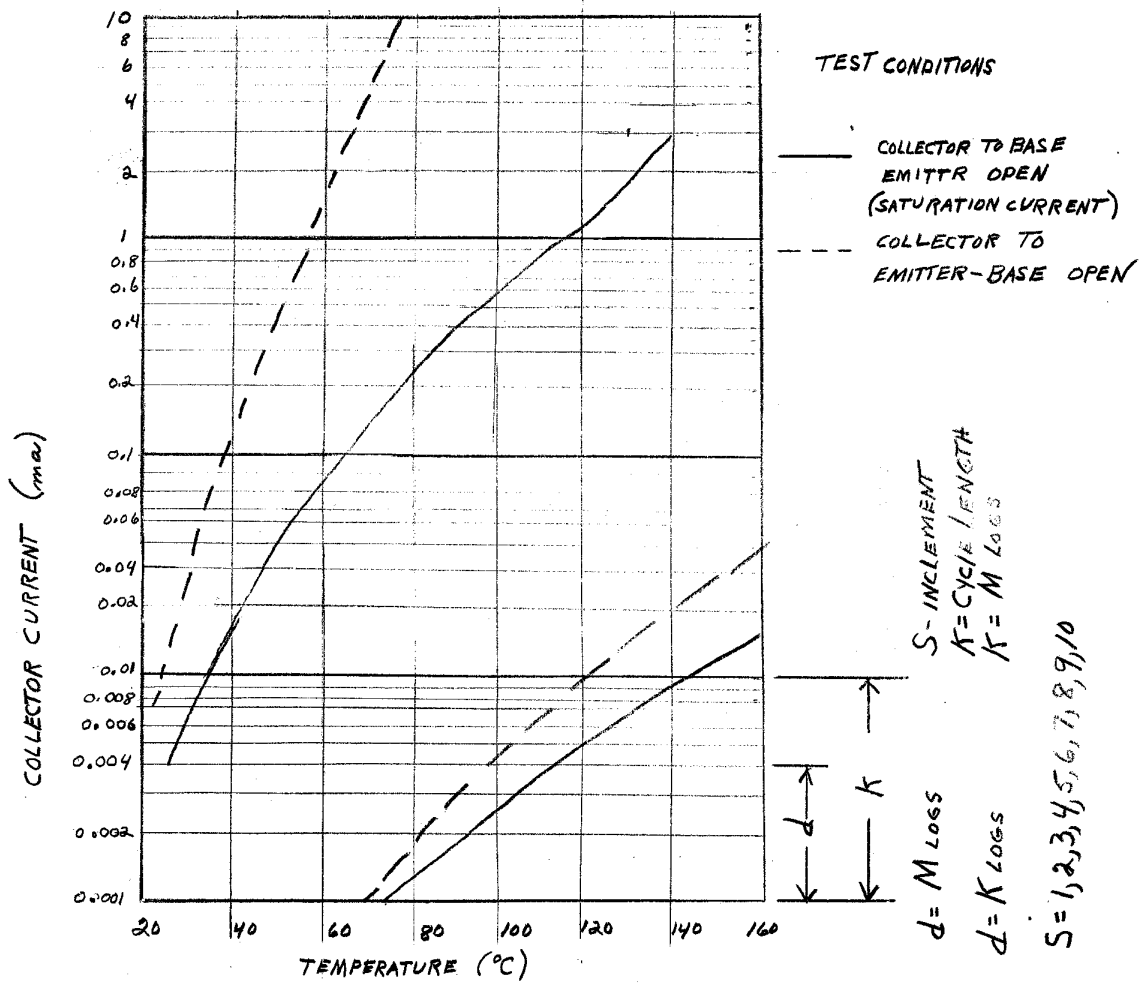
Figure 6.6. Plot of the Collector Cutoff Current for Silicon and Germanium Transistors as a Function of Junction Temperature.

equal to the saturation current times the current gain. Since the current gain is quite small at very low collector currents, the collector cutoff current will be slightly greater than the saturation current at low temperatures. As temperature increases, the saturation current will increase sharply, and the current gain will increase with the collector current. Therefore, the collector cutoff current will increase more rapidly than the saturation current alone. This is shown in Figure 6.6.

If the base is shorted to the emitter, the collector cutoff current will be only slightly larger than the reverse saturation current. Most of the excess charge in the base will then be drained off through the base terminal so the forward bias on the emitter junction will be reduced. This will significantly reduce the collector current.

Constant Voltage Bias. Figure 6.7 illustrates a possible biasing circuit. A constant d.c. bias voltage is applied to the emitter-base junction. This method has the advantage that there is a low d.c. resistance between the emitter and the base. Therefore, the excess charge produced in the base by the reverse saturation current will be drained off and multiplication of the saturation current by the emitter junction will be minimized.

Nonetheless, this method is entirely unsatisfactory. The collector current will increase steadily with temperature over the entire operating range of the transistor.

The forward bias lowers the emitter junction barrier so that current carriers can diffuse across the junction. As temperature is increased, thermal motion is more rapid so a greater number of carriers can diffuse across the junction if the barrier height remains the same. The result is that the emitter current and, consequently, the collector current will increase steadily with temperature for a given emitter-base bias. This phenomenon manifests itself over the entire temperature range of the device. Hence, to maintain a constant collector current, the forward bias voltage must be decreased with increasing temperature. This is shown in Figure 6.7. It can be seen from the figure that the bias must be decreased linearly with increasing temperature until the reverse saturation current becomes appreciable. Then the bias must be decreased more rapidly to compensate for the increased saturation current in addition to the changing emitter junction resistance.

Temperature sensitive elements are available that will drop the biasing voltage as temperature increases. These will be discussed later.
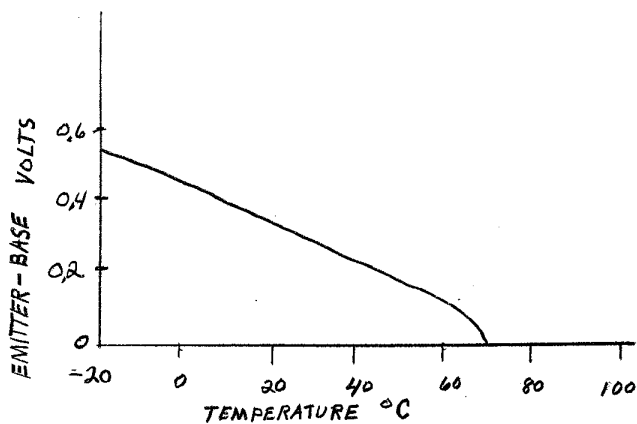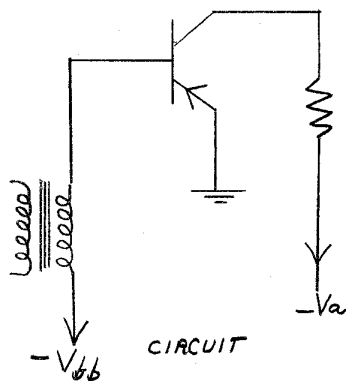
Figure 6.7. Emitter-Base Bias Voltage Required for Constant
Collector Current with Increasing Temperature.

Constant Current Bias. A transistor can also be biased by
supplying a constant current to the base. This could be accomplished
by forward biasing the base through a large resistance from a high
voltage (for example, the collector supply voltage). Such a circuit
is shown in Figure 6.8. The emitter-base resistance of the transistor
will be very much smaller than the biasing resistor so changes in
the emitter-base resistance will have little effect on base current.
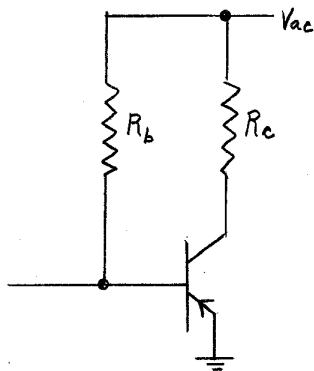
Figure 6.8. Constant
Current Bias.

Although it is much better than
constant voltage bias, constant current
bias is still unsatisfactory for two
reasons. First, the d.c. current gain
of a transistor usually increases with
temperature. Consequently, if the base
current is constant, the collector current
will increase with temperature. Second,
the collector current will increase
uncontrolled as the collector saturation
current rises. Since the base current is
constant, the saturation current must cross
the emitter junction. As a result, the total
increase in collector current produced by
the saturation current of the collector
junction will be equal to the saturation
current times the current gain. This

causes instability even at low temperatures.

Collector Feedback Bias. It should be obvious that some form
of bias stabilization is required to keep the collector current
constant over any appreciable temperature range. One method of
bias stabilization is collector feedback bias. As shown in Figure
6.9a, the base current is supplied through a resistor connected
to the collector. Therefore, if the collector current increases,
the collector voltage will drop, reducing the bias current propor-
tionally. This feedback action tends to maintain the collector
current constant.

Another circuit employing collector feedback bias is shown
in Figure 6.9b. This circuit has the advantage that the external
emitter-base resistance is reduced. The saturation current of the
collector junction will pass through the external resistance rather
than cross the emitter junction. Thus, multiplication of the
saturation current by the emitter junction is minimized. This
resistance does, however, lower the input resistance of the amplifier
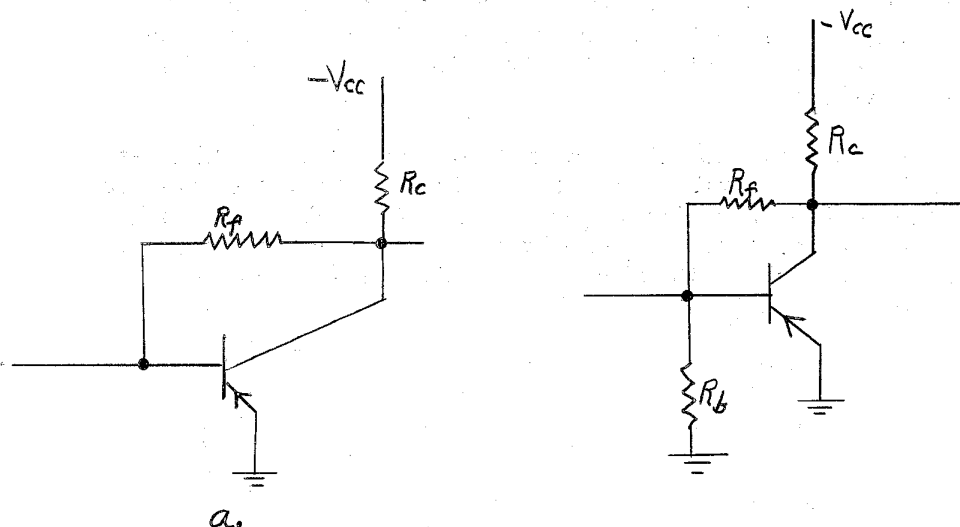which might be detrimental in some cases.

Figure 6.9. Examples of Collector Feedback Biasing.

A characteristic common to both these circuits is that negative
feedback is introduced by the collector-base resistance. This will
lower the stage gain, but it will reduce distortion and stabilize
the gain. The feedback can be eliminated, if desired, by tapping
the feedback resistance at a suitable point and connecting a large
capacitor between the tap and ground. This will prevent feedback
of the a.c. signal but will not affect the d.c. feedback.

Collector feedback bias is only effective when there is a large d.c. resistance in the collector circuit. If the collector load were a transformer primary winding with a low d.c. resistance, increases in the collector current would not change the collector voltage so the stabilizing action could not take place.

Emitter Feedback Bias. Another method of bias stabilization is illustrated in Figure 6.10. This is emitter feedback bias. As shown in Figure 6.10a, the base is forward biased from a constant voltage source; a resistor is included in the emitter lead. The voltage developed across the emitter resistor opposes the forward bias applied to the base. Thus, if the emitter current increases, the voltage drop on the emitter resistor will increase thereby reducing the bias on the emitter-base junction. This action stabilizes the emitter current against changes. The collector current will also be stabilized since it is very nearly equal to the emitter current.

One of the advantages of emitter feedback bias is that the base is fed from a constant voltage source so very little of the collector saturation current will cross the emitter junction. In addition, the circuit can be used when the collector load has a low d.c. resistance since stabilization is provided by the emitter resistance.

It is desirable to make the emitter resistance as high as possible to realize stable operation. However, the voltage drop across the emitter resistance does reduce the effective supply voltage and causes a loss of supply voltage. For this reason, there are practical limitations on the size of the emitter resistance.
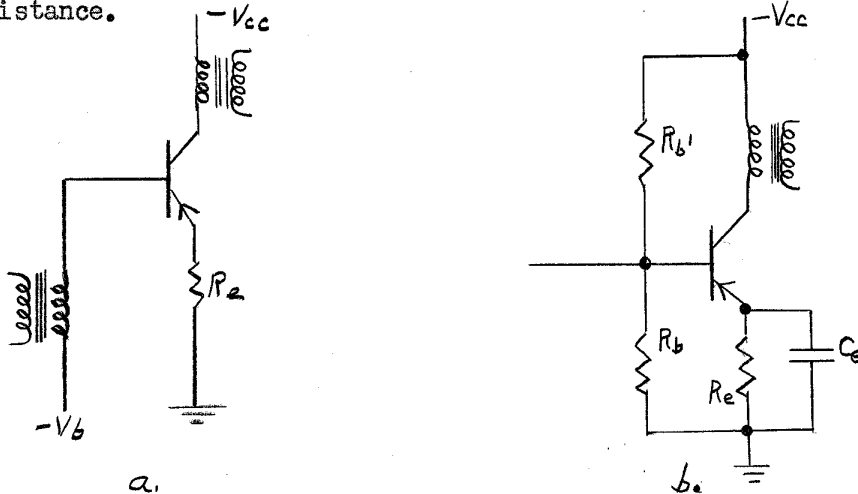

Figure 6.10. Examples of Emitter Feedback Bias.

The base bias voltage can also be supplied from the collector supply through a voltage divider. This is shown in Figure 6.10b. The resistance of this divider should be as low as possible -- without excessive shunting of the input signal -- to approximate a constant voltage source. If the base is fed from a high resistance divider (nearly constant current) the desirable effects of the emitter resistor will be nullified because the base current will remain nearly constant regardless of the emitter voltage drop.

Emitter feedback biasing will produce degeneration of the a.c. signal, unless the emitter resistance is by-passed as shown in Figure 6.10b. The by-pass capacitor prevents a.c. voltages from appearing across the emitter resistor.

Conclusions. The biasing circuits discussed thus far are compared in Figure 6.11. The curves are for germanium transistors. In each case, the bias was adjusted to give 1 ma collector current at room temperature.

Constant voltage bias is unsatisfactory since the collector current is very unstable even over a restricted temperature range.

With constant current bias the collector current is reasonably constant at low temperatures, but rises rapidly at junction temperatures above 40°C. Nonetheless, even if a circuit is to be operated at a nearly constant temperature, constant current bias will not give completely satisfactory results. The collector current is directly proportional to the common emitter current gain, and this can vary greatly with age and from transistor to transistor. Thus, constant current bias is suitable only for experimental work.

Collector feedback bias provides adequate stabilization for most commercial applications, but its use is restricted to circuits where the collector load has a relatively high d.c. resistance. It cannot be used in transformer coupled stages. For this reason, collector feedback bias is usually limited to resistance coupled, audio-frequency amplifiers. With collector feedback bias, the collector current can be made nearly independent of the transistor characteristics. This is done by using a low resistance divider to feed the base from the collector.

Emitter feedback bias is about the best general-purpose biasing method. The stability of this method is evidenced in Figure 6.11. As with collector feedback bias, the collector current can be made nearly independent of transistor parameters by using a sufficiently high emitter resistance. This is highly desirable for interchangeability of transistors. Emitter feedback bias is not too frequently used in power amplifiers since the emitter resistance can cause an excessive loss of power. Other biasing techniques are generally used in power stages. These will be discussed in the next section.
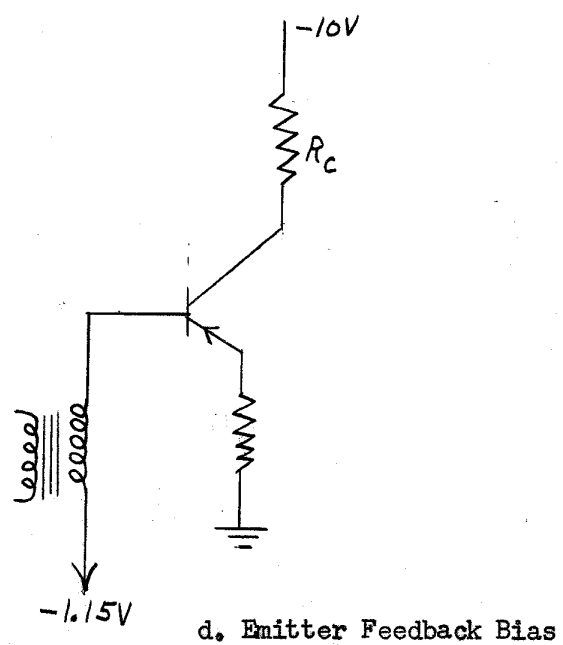
a. CONSTANT VOLTAGE BIAS

b. CONSTANT CURRENT BIAS
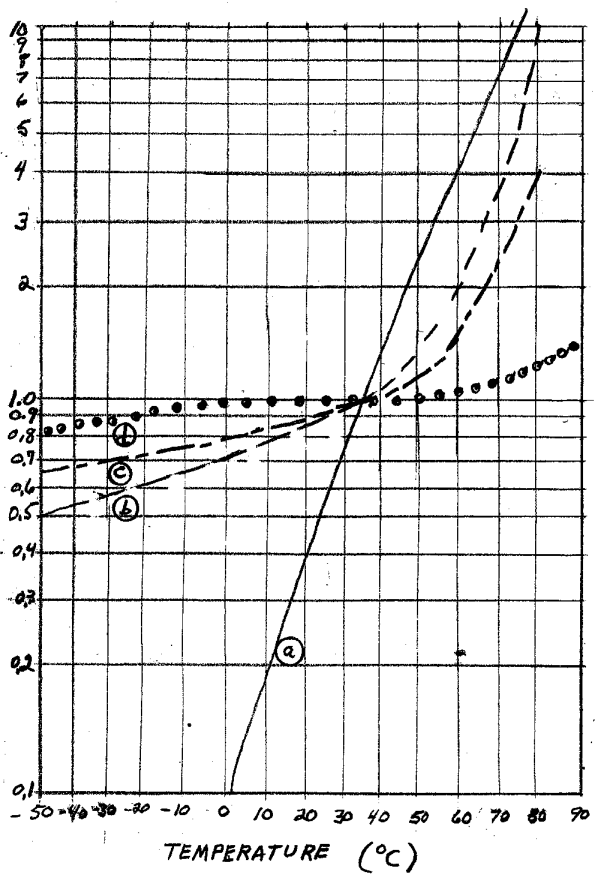
c. COLLECTOR FEEDBACK BIAS
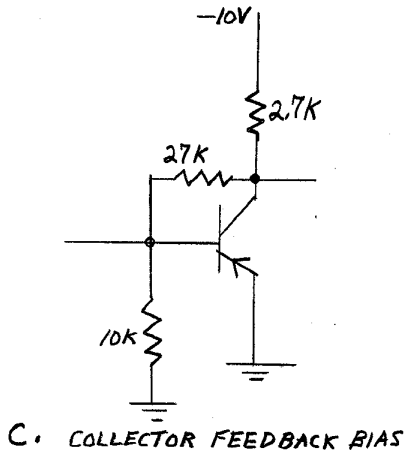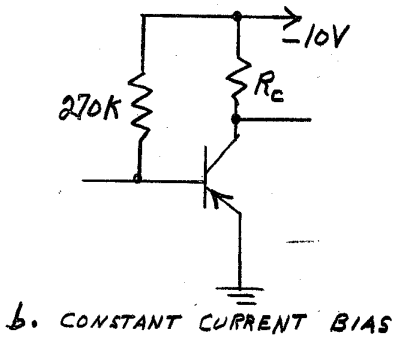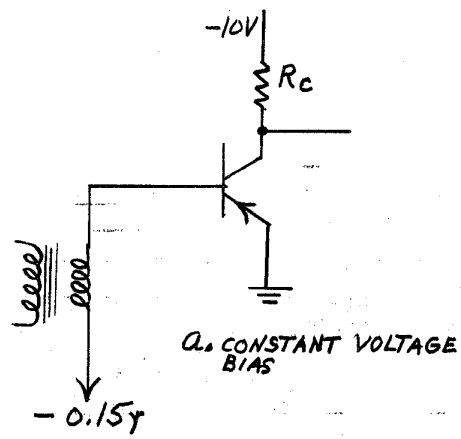
d. Emitter Feedback Bias

Figure 6.11. Variation of Collector Current with Temperature for Typical Germanium Transistor Using Different Biasing Circuits.

The biasing of common base and common collector stages is similar to that of common emitter stages. In Figure 6.12a, the biasing used is identical to the emitter feedback bias described previously. The base is fed from a voltage divider connected to the collector supply, and a resistor is included in the emitter lead. If the collector current should rise, the voltage drop across the emitter resistor will increase and lower the bias current. The only difference with this circuit is that the input signal is fed to the emitter, and the base is by-passed to ground.

Another common base circuit is given in Figure 6.12b. Here, the base is fed from a voltage divider connected to the collector. Hence, when the collector current increases, the collector voltage will drop and will lower the forward bias on the base. This is identical to the collector feedback bias circuit for a common emitter stage.

Common collector biasing circuits are given in Figure 6.13. The arrangement in Figure 6.13a is the same as the collector feedback circuit given in Figure 6.9a. The circuit in Figure 6.13b employs a combination of collector and emitter feedback. When the emitter current increases, the increased voltage drop across the 4.7K resistor lowers the voltage supplied to the voltage divider supplying the base as with collector feedback; and the increased voltage drop across the 1K resistor opposes the forward bias as with emitter feedback.

Thermistor Compensation. The thermistor is a resistor with a large, negative temperature coefficient. That is, its resistance will decrease quite noticeably as temperature increases. If a thermistor is used as part of a voltage divider supplying bias to a transistor, as shown in Figure 6.14a, the bias will be reduced as temperature increases. This compensates for the tendency of the collector current to increase with temperature. It can be seen from the plot in Figure 6.11 that this compensation is not exact. The temperature coefficient of the thermistor is not matched to that of the transistor. Hence, over part of the temperature range the compensation is too great while over other parts it is not enough. If the thermistor were not shunted with a resistor (Figure 6.14a), the compensation at low temperatures would be far too great. At low temperatures when the resistance of the thermistor is high, R_2 will determine the voltage division ratio of the bias network and there will be practically no compensation. As temperature increases, the resistance of the thermistor becomes comparable to that of R_2 and the bias voltage decreases with increasing temperature. At medium temperatures this compensation is too great until the collector saturation current becomes appreciable. Finally, the rapidly increasing saturation current causes a net increase of collector current in spite of the thermistor.
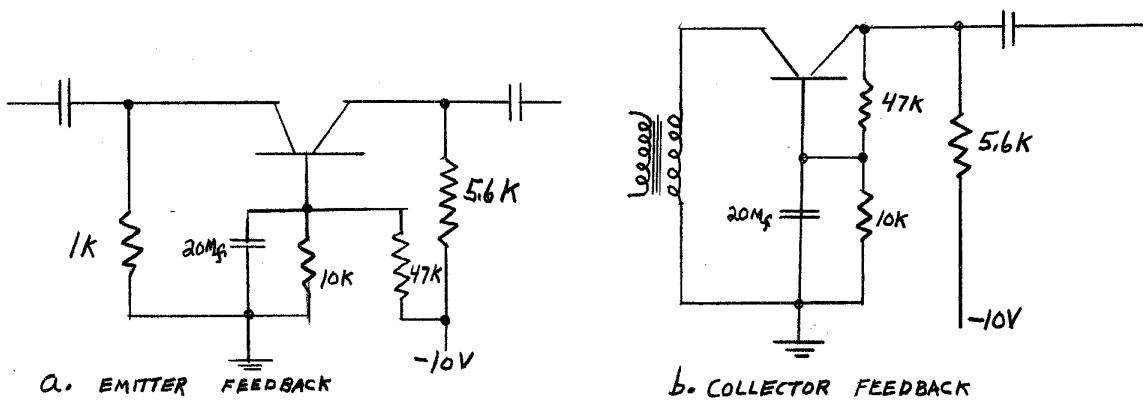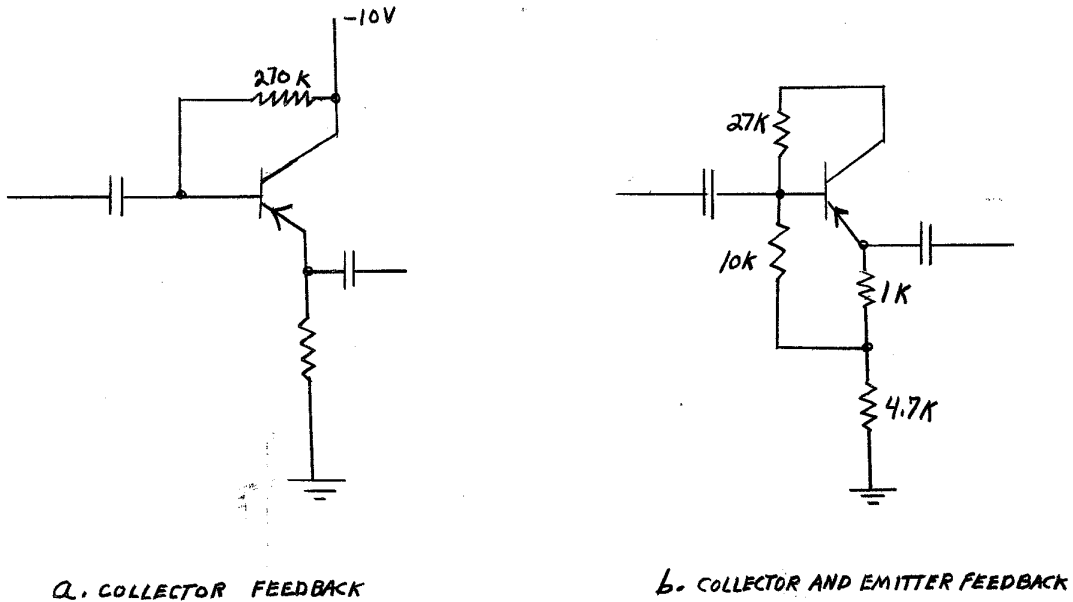
Figure 6.12. Common Base Biasing Circuits

a. COLLECTOR FEEDBACK

b. COLLECTOR AND EMITTER FEEDBACK

Figure 6.13. Common Collector Biasing Circuits.

b. DIODE COMPENSATION

a. THERMISTOR COMPENSATION

COLLECTOR CURRENT (ma)
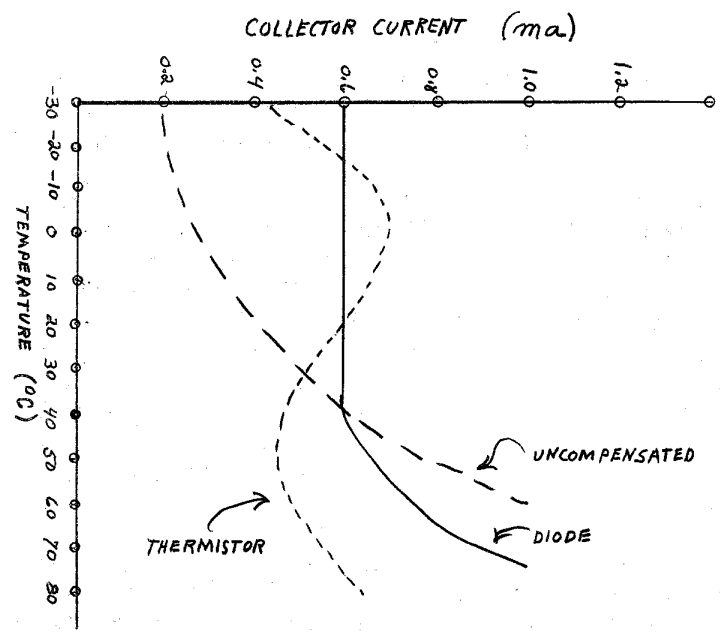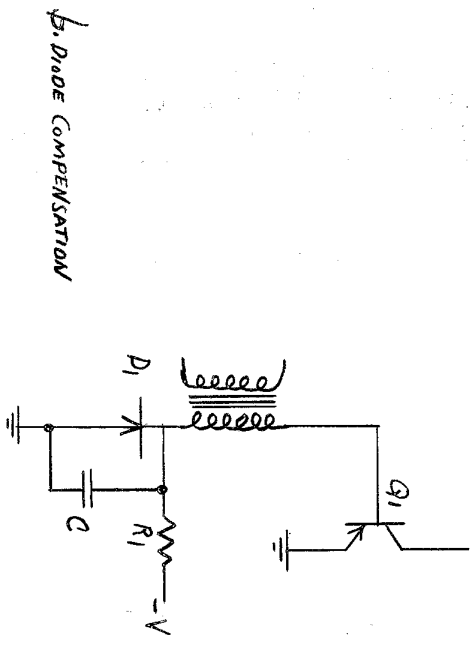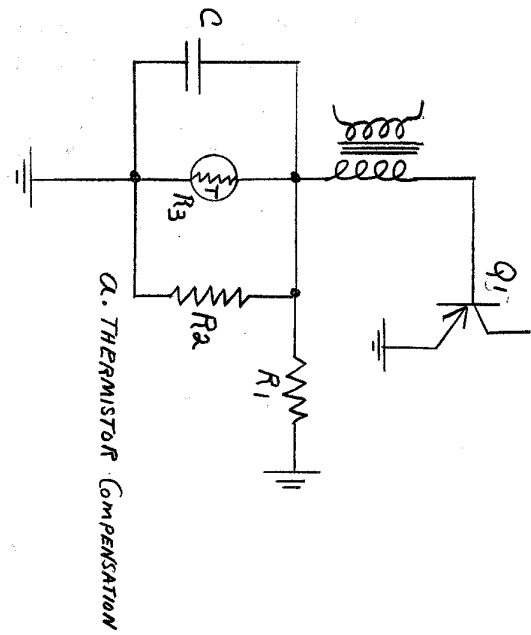
TEMPERATURE (°C)

UNCOMPENSATED

DIODE

THERMISTOR

Figure 6.14. Negative Temperature Coefficient Thermistor and Semiconductor Diode Compensation.

Recently, positive temperature coefficient, silicon thermistors have been made with thermal characteristics closely resembling those of a silicon transistor. When these elements are used in conjunction with feedback circuits, almost exact compensation is realized. An example is shown in Figure 6.15.
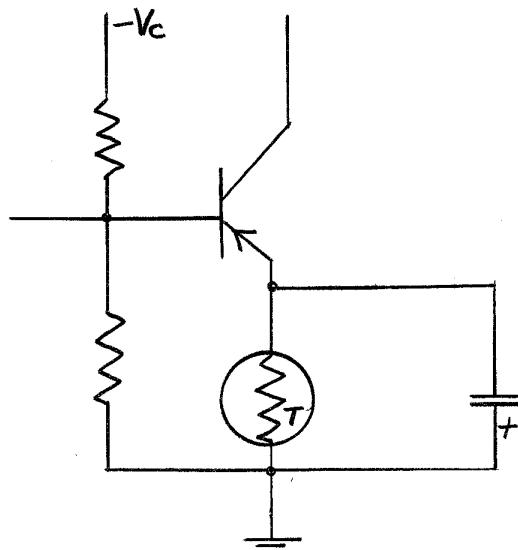

Figure 6.15. Example of Temperature Correction Using a Positive Temperature Coefficient Thermistor in Conjunction with Feedback.

Diode Compensation. An ordinary junction diode can be used to compensate transistor bias. In Figure 6.14b, the transistor bias voltage is developed across the forward-biased diode (D_1). The forward resistance of the diode varies with temperature in the same manner as does the impedance of the emitter-base junction of the transistor. If the current through the diode is large compared to the base current of the transistor, the diode alone will determine the bias voltage. This system will provide almost perfect compensation as long as the saturation current is small. This is evidenced by the graph. If the transistor is to be operated at higher temperatures, some auxiliary means of compensation will be necessary.

When temperature compensation is used, it is important to locate the temperature sensitive element close to the compensated transistor. This is necessary because the transistor may become noticeably hotter than its surroundings, particularly if power is being dissipated in the unit.

SMALL SIGNAL AUDIO AMPLIFIERS

The transistor is basically a non-linear device: the input and output impedances, as well as the current gain, vary with collector current. However, when the signal level is low, these changes will be small so some definite magnitude can be assigned to the impedances and the gain. Then, the transistor can be considered to be a linear amplifier. In this section it will be assumed that this condition is satisfied. Furthermore, only low-frequency operation will be considered so the deleterious effects of junction capacitance and transit time can be neglected.

Single-stage amplifiers -- common emitter, common base, and common collector -- were discussed previously. This section will be primarily concerned with methods of coupling, or cascading, these basic amplifiers to realize higher gains than might be possible with a single stage.

Although specifying the voltage and current gain of an amplifier is frequently useful, the determination of these values becomes somewhat complex with cascaded amplifiers. These two terms can be combined into one -- the power gain which is the product of the voltage and current gains. The power gain gives an absolute measure of amplifier performance. The input to an amplifier, be it from a high-impedance crystal microphone or a low-impedance pickup, is at a certain power level. This signal must be amplified to supply power to a load whether it is a set of headphones or a large motor.

From the preceding discussion it follows that it is desirable to produce maximum power gain per stage, the voltage or current gain alone being of minor significance. To do this, it is necessary to match impedances through the amplifier. Any mismatch will result in a lower gain per stage.

Transformer Coupling. From an electrical viewpoint, the simplest way to couple two stages is with a transformer. The output of one stage is fed to the input of the following, as shown in Figure 6.16. The transformer provides isolation of d.c. voltages and impedance matching.

Transformer coupling has the advantage that very little d.c. power is wasted: the d.c. resistance of the transformer primary is small compared to the load impedance. As compared with resistive coupling, equal results can be obtained with half the supply voltage. Nonetheless, a transformer does have certain disadvantages: for example, comparatively high cost, limited bandwidth, and large phase shifts at high frequencies. Size is not a particular disadvantage because miniature transformers can be fabricated to operate at the low power levels of transistors.
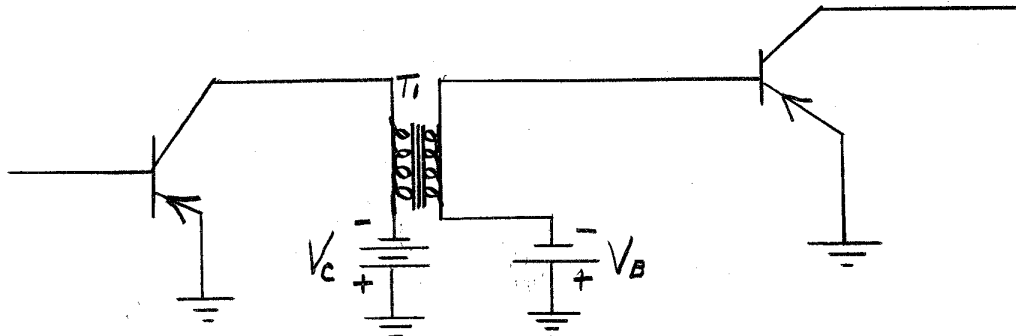
Figure 6.16. Example of Transformer Coupling.

A two-stage, transformer-coupled audio amplifier is shown in Figure 6.17. Input and output transformers, as well as an inter-stage transformer, are used. The input transformer must match the source to the 1000 impedance (1000-Ω) input of the common emitter amplifier. The interstage transformer must match the low impedance input of the second stage to the medium output impedance (15,000-Ω) of the first stage. The third transformer matches the load to the second transistor output.

It should be noted that maximum power gain for each stage can be realized because each transistor can be made to work into an optimum load impedance by appropriate adjustment of the transformer turns ratio. For this reason, the common emitter configuration is most frequently used because it is capable of a higher gain per stage since it has both current and voltage gain.

In Figure 6.17 two methods of bias feed are shown. Both stages use emitter feedback resistors, and the base bias is supplied from a voltage divider. In the first stage the bias is applied directly to the transistor; the transformer is connected to the base via a coupling capacitor which prevents shunting the d.c. bias. In the second stage the transformer is connected directly to the base, and the bias voltage is fed through the transformer primary from the voltage divider. A bypass capacitor is required across the voltage divider to prevent loss of the input signal across the divider resistance. Generally speaking, the latter method is more satisfactory because the voltage divider resistance does not shunt the input signal; however, the two circuits are equivalent as far as biasing is concerned.

If the emitter resistance is not bypassed, the output current flowing through the emitter would produce a voltage across this
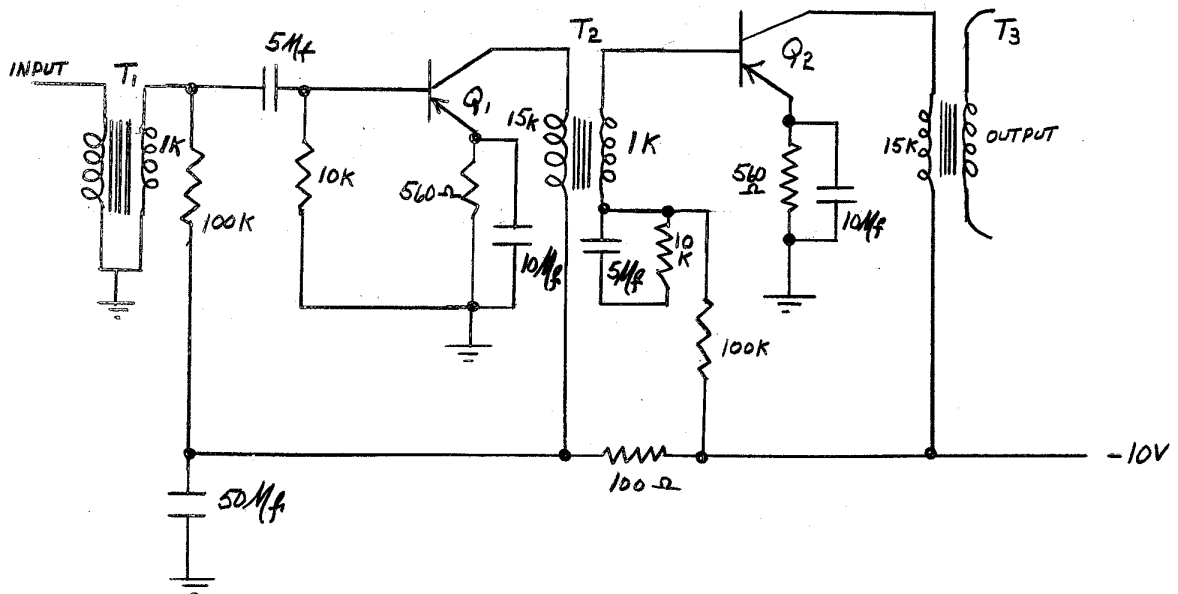
Figure 6.17. Two-Stage, Transformer Coupled Amplifier
              Illustrating Methods of Bias Feed.

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resistance which would oppose the input signal being applied between the base and emitter. Also, the signal developed across the resistor would not appear in the output, giving an additional loss.

Because of the low d.c. resistance of the transformer primary it is necessary to employ effective bias stabilization with transformer coupled amplifiers. The increased collector current caused by increased temperature will not materially reduce d.c. collector voltage, so the power dissipation of the transistor will increase as the square of the collector current. The increased power dissipation will increase heating and collector current, causing ultimate destruction of the transistor if precautions are not taken.

A decoupling network is shown in the collector supply between the two stages. Although this is not necessary for a two-stage amplifier it is usually required when there are three or more stages. This network attenuates any low-frequency being fed back through the collector supply. Lack of adequate decoupling usually results in instability and low-frequency oscillations (motor boating).

R-C COUPLED AMPLIFIERS

Another method of cascading several stages is resistance-capacitance coupling. This scheme is illustrated in Figure 6.18. Direct current voltage is fed to the collector of the first stage (Q_1) through a resistor (R_c). Variations in the collector current of Q_1 develop a signal voltage across the collector resistor. This signal is coupled to the base of the following stage (Q_2) through a capacitor (C_c), and bias is supplied to this second stage through the resistor R_b.

It should be recognized that the purpose of the coupling network is to deliver the signal power, generated in the collector circuit of Q_1, to the base of Q_2, therefore it is necessary that the shunt resistance of R_c and R_b be high compared to the input impedance of Q_2 to prevent loss of power in these elements. Also, the capacitance reactance of C_c should be small compared to the input impedance of Q_2 at the lowest frequency to be amplified.
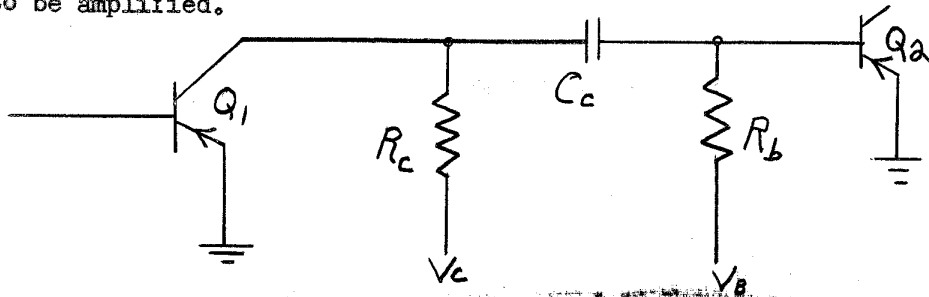

Figure 6.18. Example of R-C Coupling.

There is a limit on the maximum value of collector resistance (R_c). Because the d.c. power must be supplied to Q_1 through this element, too high a resistance will limit collector current and, therefore, the maximum current swing of the transistor.

Figure 6.19 illustrates some of the limitations of the r-c coupled amplifier. The loadlines of both an r-c coupled amplifier and a transformer coupled amplifier, both operating with a collector supply voltage of 5v and quiescent current of 0.3 ma, are plotted. Referring to the loadlines of the r-c amplifier, the d.c. loadline is determined by the collector resistor. This loadline determines the static conditions of the transistor as all possible combinations of static collector voltage and current lie on this line. The impedance presented to the collector is considerably lower for an a.c. signal. The a.c. load impedance consists of the parallel combination of the collector resistor and the base resistor and input impedance of the following stage. This results in a separate, lower-impedance, a.c. load line. This load line must pass through the operating point as shown in the figure. It can be seen that the voltage swing is restricted, but this in itself is not too serious because the voltage swing requirements for the input of the following transistor are small.

The bias stabilization requirements of an r.c. coupled amplifier are not as stringent as with transformer coupling because of the comparatively large d.c. resistance in the collector circuit. Thermally generated increases in collector current will also decrease collector voltage; hence, the collector dissipation will not increase too rapidly with collector current. In fact, if the collector resistance is sufficiently high, the collector dissipation will decrease with increasing d.c. collector current. All this aside, thermally induced changes of collector current will shift the operating point which is undesirable so some degree of stabilization is still required.

An example of a two-stage, r-c coupled amplifier is given in Figure 6.20. Collector feedback bias stabilization is employed on the first stage. The voltage divider supplying bias to the base of the first transistor is fed from the transistor collector, so an increased collector current will lower the forward bias by decreasing the collector voltage. The second stage uses emitter feedback bias stabilization. A voltage divider from the negative collector supply forward biases the second transistor, and a resistor in the emitter circuit provides bias stabilization by decreasing base to emitter voltage when the collector current increases. The emitter resistor is bypassed with a 10 μf capacitor to prevent degeneration at signal frequencies.

The input signal is coupled into the amplifier through a 5 μf capacitor, amplified by the first stage, and coupled into the second stage through another 5 μf capacitor. This capacitor couples the

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COLLECTOR CURRENT

1.25 Ma

1.0 Ma

0.75 Ma

0.5 Ma

0.25 Ma

R-C COUPLED AMPLIFIER

OPERATING POINT

AC LOAD LINE (1 kΩ)

DC LOAD LINE (10 kΩ)

OPERATING POINT

TRANSFORMER COUPLED AMPLIFIER (15K LOAD)

$I_b = 10$ Ma

20 Ma

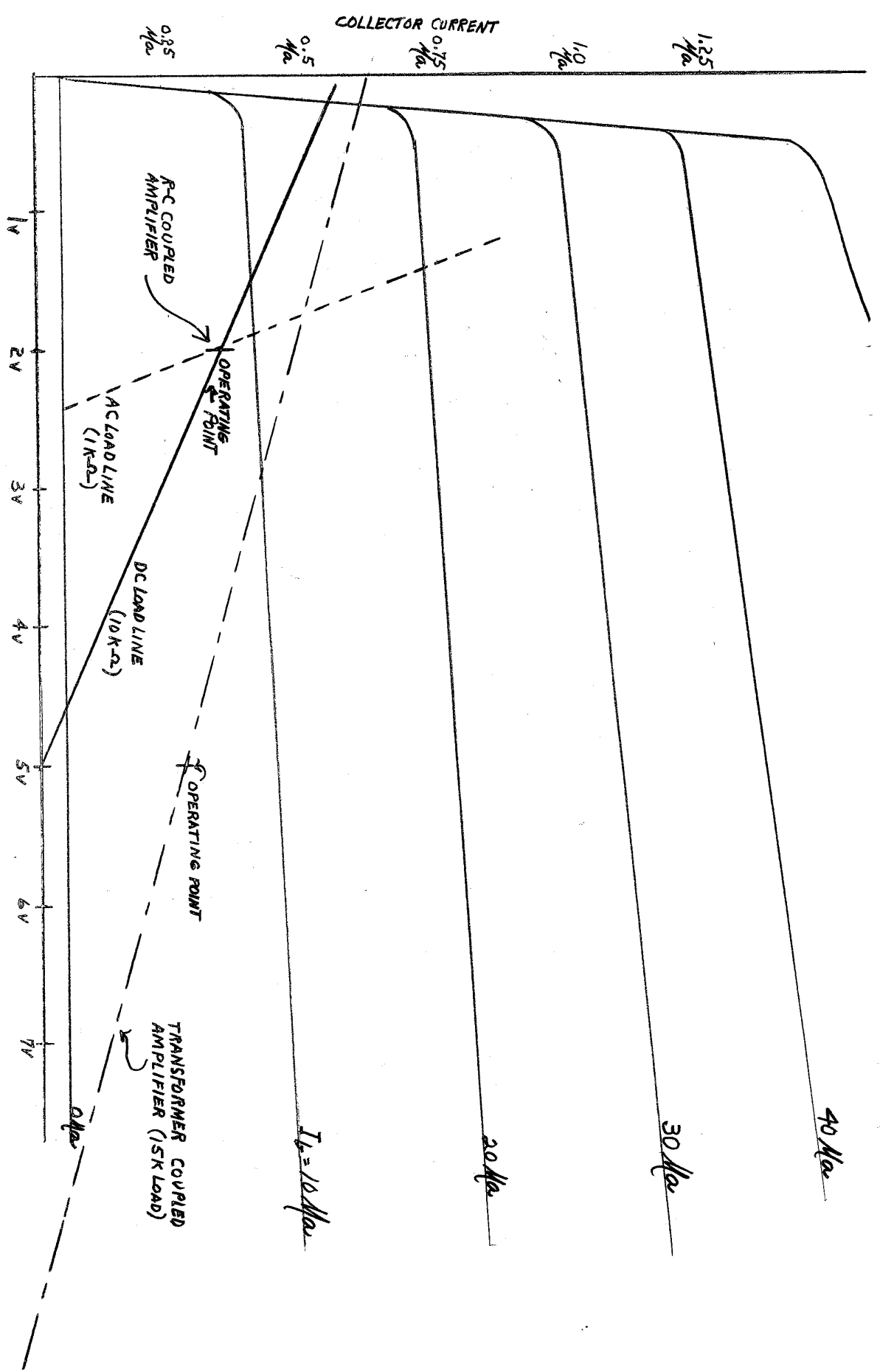30 Ma

40 Ma

0 Ma

1v 2v 3v 4v 5v 6v 7v

Figure 6.19. Comparison of Transformer and R-C Coupled Amplifiers.
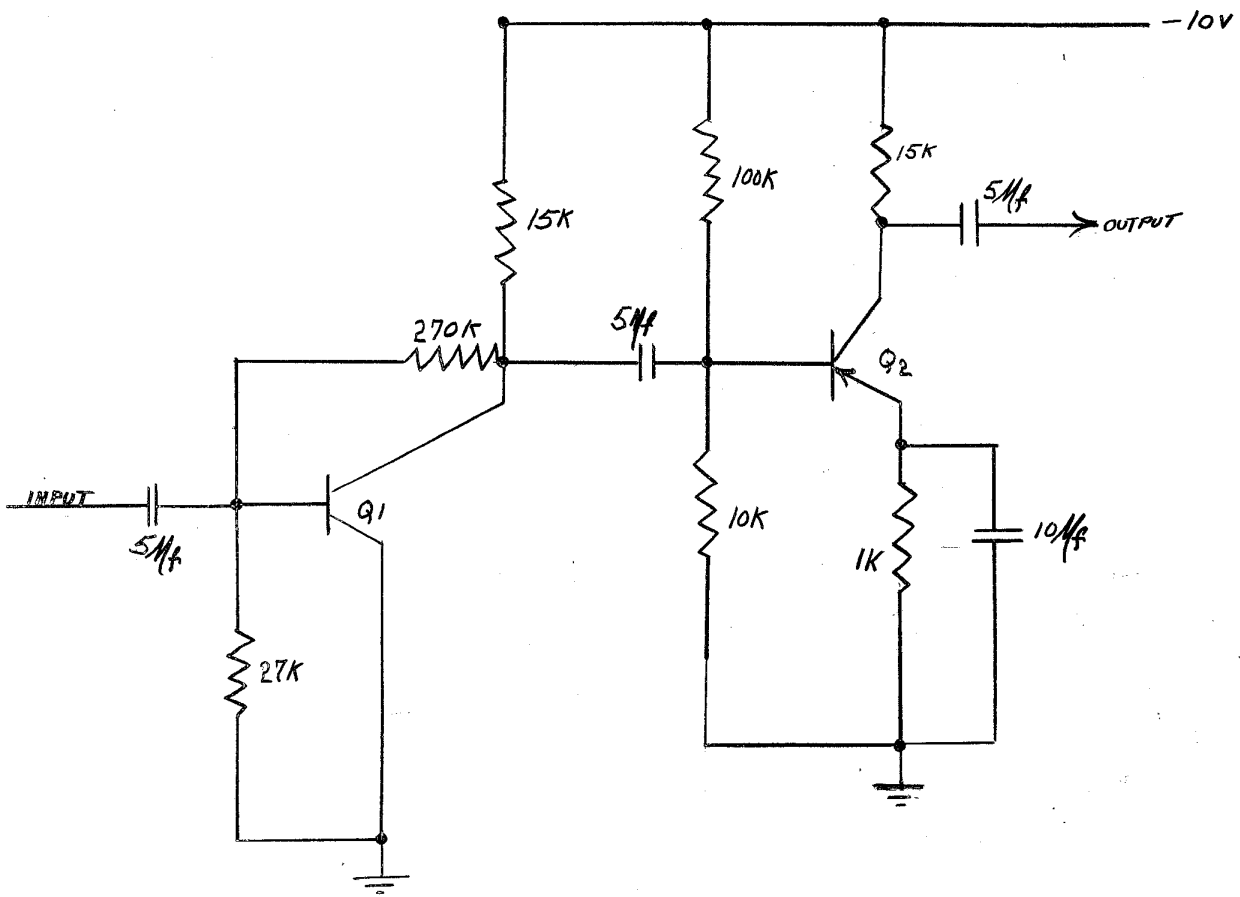
Figure 6.20. Two-Stage, Resistance-Capacitance Coupled
            Amplifier.

15000 Ω impedance of the collector circuit to the 1000 Ω input impedance of the following transistor. Because of this mismatch the maximum gain of Q_1 is not realized. The second transistor amplifies the signal further, and it is finally coupled out from the collector circuit of this stage through still another coupling capacitor. The gain of the second stage will depend upon the impedance of the load to be driven by the amplifier. If the load impedance is very much different from the collector impedance of Q_2 the power gain of the stage will be reduced. Also, as shown in Figure 6.19, the voltage swing of the amplifier will be restricted by low impedance loads.

Resistance-capacitance coupling is generally used where wide-band frequency response is needed or where the amplifier is to be employed within a negative feedback loop (because of the lower phase shift per stage of r-c coupled amplifiers as compared to transformer coupled amplifiers). Also, in some cases, r-c coupling is less expensive than transformer coupling. However, the realizable gain per stage is lower with r-c coupling and it is only practical between stages with input and output impedances in the same order of magnitude because large impedance mismatches produce excessive loss of gain. Moreover, r-c coupling cannot be used on high-level stages because the nonlinear input characteristics of the transistor cause charging of the coupling capacitor and a shift in operating point -- but this will be discussed further along with high-level amplifiers.

Direct Coupled Amplifiers. The characteristics of the transistor and also the availability of PNP and NPN types makes this device well suited for direct coupled amplifiers. Several configurations for direct-coupling transistor stages are shown in Figure 6.21. Figure 6.21a shows a common collector stage coupled directly to a common emitter stage. The emitter current of Q_1 passes through the base of Q_2 with the input resistance of Q_2 acting as the load for Q_1 and the voltage V_1 acting as the collector supply. In Figure 6.21b a two-stage common collector amplifier is shown. Again the output current of the first stage is the input current of the second and the voltage source (V_1) has been connected to act as a collector supply for both stages. Figures 6.21 (c through e) illustrate the use of both NPN and PNP transistors. Figure 6.21e is of particular interest because it clearly shows that there is no intrinsic limit on the number of stages that can be cascaded. It can be seen from the figure that the collector circuit of the intermediate stages is in series with the input circuit of the following stage and a voltage source. The transistors and voltage sources must be connected so there is a path for current flow around this loop in such a direction as to forward bias the base-emitter junctions and reverse bias the collector junctions.

One thing that should be immediately apparent in these circuits is that temperature stability can become quite a problem. An increase in temperature will cause increases in collector current in all the stages, but the collector current increase of the first stage will be amplified by successive stages which will result in intolerable increases in the final stages, if precautions are not taken. Therefore, the circuits, as shown in Figure 6.21 are not practical.
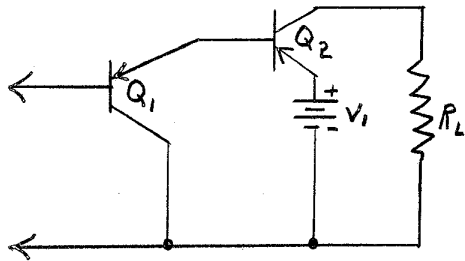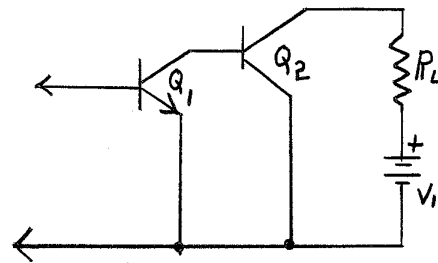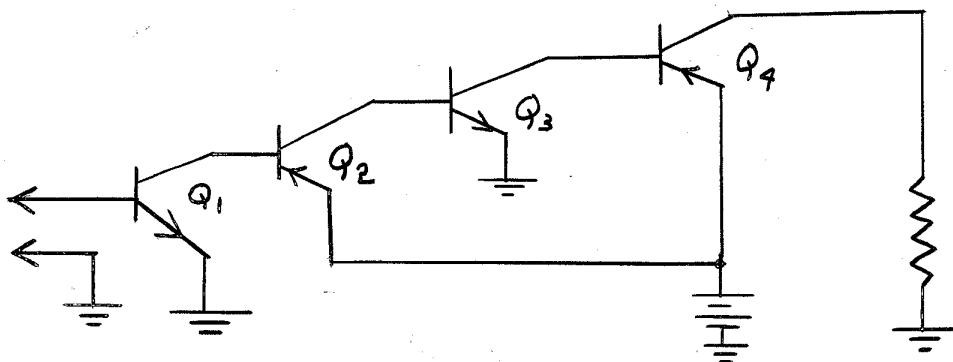
a. COMMON COLLECTOR, COMMON EMITTER
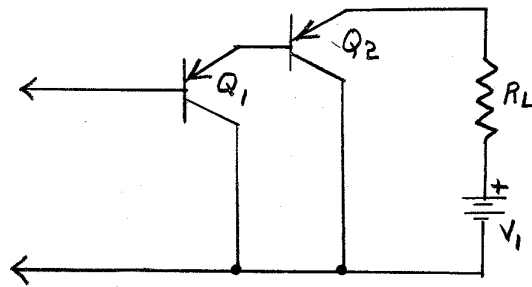
b. TWO STAGE COMMON COLLECTOR
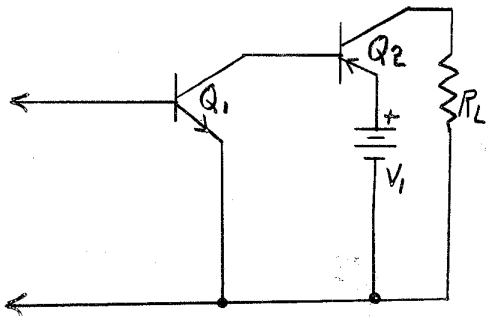
c. TWO STAGE COMMON EMITTER

d. COMMON EMITTER, COMMON COLLECTOR

e. CASCADED COMMON EMITTER AMPLIFIER

Figure 6.21. Elementary, Direct-Coupled Amplifiers.

Figure 6.22 illustrates a practical two-stage common emitter, direct-coupled amplifier. Diode compensation is employed on the first stage. The voltage across D_1 will vary inversely with temperature, reducing the bias on Q_1 at higher temperatures. The input signal is developed across R_3 which delivers the bias voltage to Q_1. The resistor R_2 produces the bias on the compensating diode from the collector supply. A resistor (R_1) is placed on the input of Q_2 (base to emitter) to prevent forward biasing of Q_2 by its own reverse saturation current when Q_1 is cut off. The diode (D_2) is a zener diode. The voltage regulating qualities of this device are used to prevent signal variations of the collector-supply voltage. This diode takes the place of a bypass capacitor which would not be effective at very low frequencies. However, a capacitor could also be employed for additional bypassing at high frequencies.

Although temperature compensation alone is usually adequate for two-stage amplifiers, some form of feedback must be employed if more stages are used or if a wide range of temperatures is to be encountered. Figure 6.23 indicates how this might be accomplished. The circuit of Figure 6.23 uses diode compensation, three local feedback loops, and overall feedback. The purpose of the feedback is not only to improve d.c. stability but also to decrease nonlinear distortion.

The first stage employs diode compensation as did the previous circuit; however, the emitter resistor (R_3) also provides bias stabilization. In addition, the output voltage (across R_L) is fed back to the input stage via. R_7. As the conduction of any of the transistors increases, regardless of the cause, the voltage across R_L will increase. A portion of this increase will be coupled back across R_3 through R_7, decreasing bias on the first stage. This degeneration is also present for the signal so the overall gain is reduced and distortion decreased. The second transistor (Q_2) is connected in the common-collector configuration. The load of this stage (R_5, R_6, and the input impedance of Q_3) serve as an emitter resistor and provides local feedback on the stage. The third stage (Q_3) also incorporates an emitter resistor for local feedback. The input resistor (R_5) is connected between the base and emitter across R_6 so that the voltage variations across R_6 might be felt in the input circuit of Q_3. In addition to providing bias stabilization, the local feedback loops also improve the transistor input characteristics, thereby reducing distortion in the previous stages resulting from driving a non-linear impedance.

In this circuit the zener diode (D_2) serves to keep signal variations out of the supply voltage.
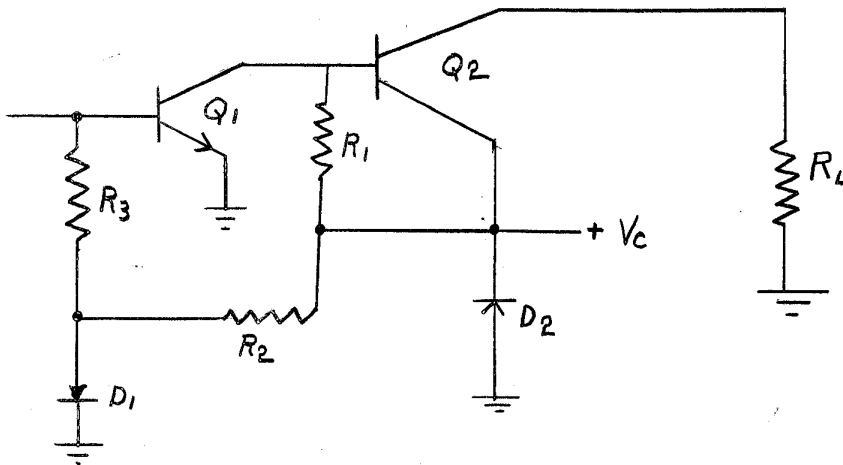
6-28

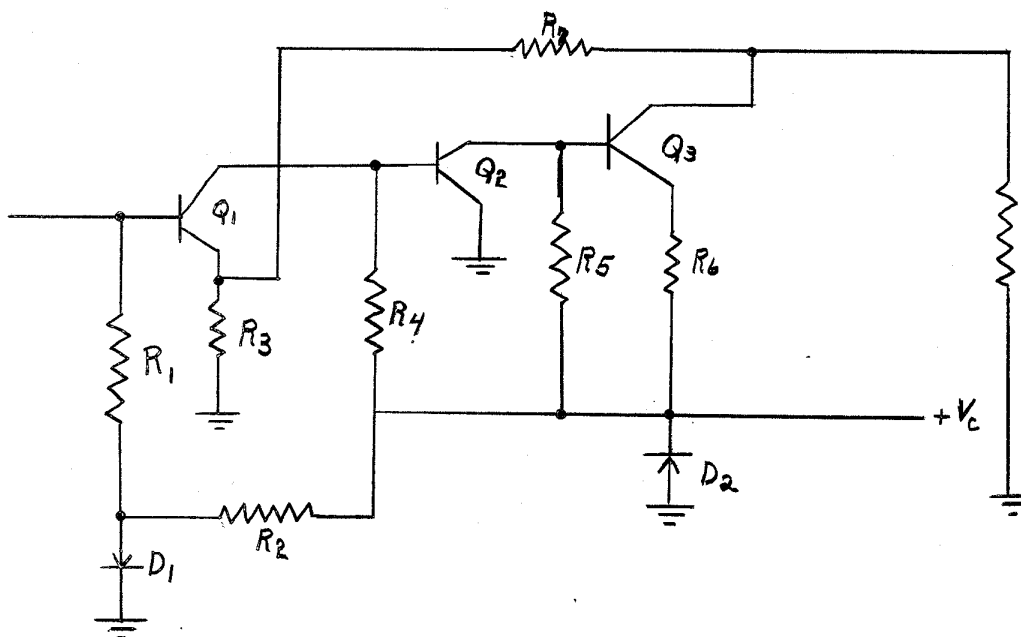Figure 6.22. Temperature Compensated D-C Amplifier.

Figure 6.23. Temperature Compensated D-C Amplifier with
Feedback to Improve Linearity and Stability.

The use of direct coupled amplifiers along with NPN and PNP transistors can be extended further to make many useful and interesting circuits. One of the most outstanding of these is the complementary-symmetry amplifier shown in Figure 6.24. This amplifier is most frequently used as a power amplifier, but its operation will be introduced here.

The input signal is fed to both Q_1 and Q_2, but these transistors have opposite polarity: one is an NPN and the other PNP. On the positive half cycles of the input signal, Q_1 will conduct while Q_2 is cut off. As the collector current of Q_1 passes through the base of Q_3, this transistor will also conduct, reproducing the positive half cycle of the input signal across R_L. On the negative half of the input signal Q_2 becomes forward biased and conducts, and Q_1 cuts off. This causes Q_4 to conduct and impresses some portion of Y_2 across the load, reproducing the negative half cycle. When the input signal is zero, no current passes through R_L so the output is also zero.

The circuit as shown in Figure 6.24 is somewhat simplified. Some form of bias stabilization is required. Also it is generally desirable to employ negative feedback to equalize the gain of the upper and lower channels. However, the techniques discussed previously can be applied here so these points will not be discussed further.

In establishing some relative merit for direct coupling, it can be said that it has many advantages over both transformer and r-c coupling. It is frequently used in low frequency signal amplifiers and also in wide band amplifiers using large amounts of negative feedback. The absence of any reactive components (capacitors and inductors) gives the d.c. amplifier a favorable position as regards low-frequency phase shift. The elimination of coupling components may permit reductions in equipment size -- particularly in the case of high-power circuits. However, some of the limitations of r-c coupling are also present with direct coupling. In particular, the direct coupling of stages does not permit impedance matching so maximum gain cannot be realized. Nonetheless, higher gains than with r-c coupling are possible because none of the signal is lost in a collector resistor. Moreover, the absence of the collector resistance eliminates the inefficient use of d.c. power as in r-c amplifiers.

POWER AMPLIFIERS

Thus far low-level amplifiers have been considered. The signal levels involved have been assumed to be low enough so that operation of the transistor could be considered linear. If this condition is satisfied, some definite value can be assigned to the transistor parameters, such as input impedance, output impedance, and current
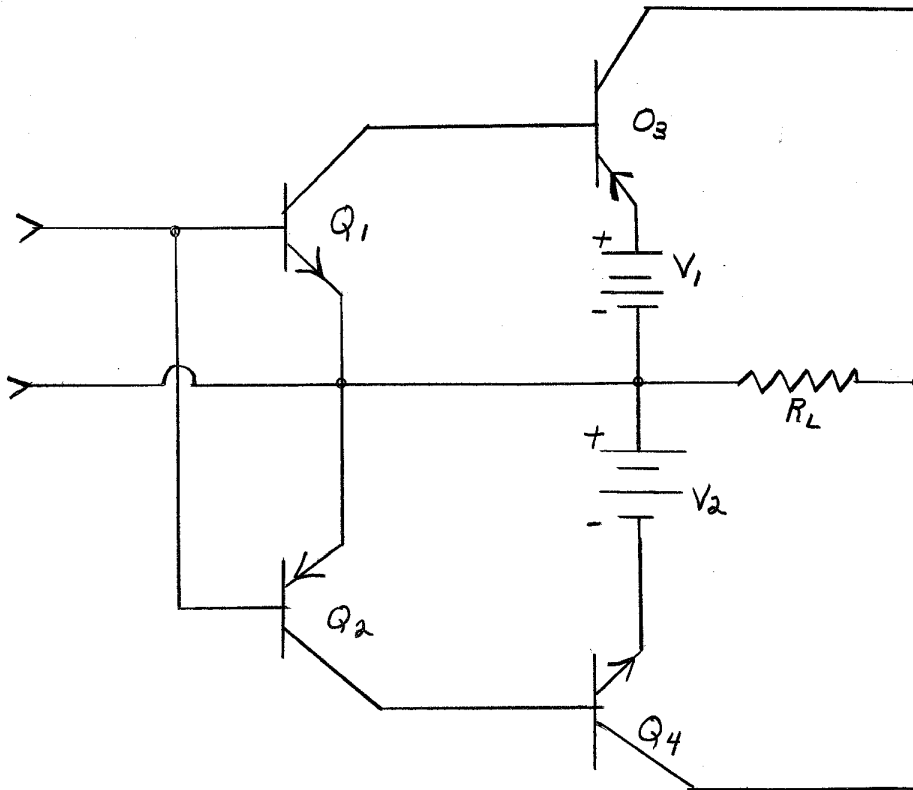
Figure 6.24. Single-Ended, Push-Pull Class B Amplifier.

gain. Moreover, the nonlinear distortion produced in the amplifier could be neglected. When the signal amplitude is increased, as in a power amplifier, this is no longer true: the input impedance will vary with the instantaneous input voltage, and the current gain will vary with instantaneous input voltage, and the current gain will vary with instantaneous collector current. Because of these nonlinearities, the amplifier will produce distortion and care must be taken in the design of circuitry to keep the distortion at a minimum while producing a large output signal.

The Output Circuit. Before going into the drive requirements of power transistors and the reduction of nonlinear distortion, the generation of power in the output circuit will be related to the maximum ratings of the transistor.

As was mentioned, the output impedance of a transistor does not determine the optimum load impedance for a power amplifier. Instead, it is the maximum ratings of the device that must be considered. In order to produce a large power output, it is necessary to have a large voltage and current swing, so the collector reverse breakdown voltage and the maximum collector current will be limiting factors. Moreover, in the operation of the amplifier, power will be dissipated into heat, so the maximum permissible power dissipation must be considered.

Amplifier Circuits. The discussion of low-frequency power amplifiers will be limited to two types: the single-ended, class A amplifier and the push-pull, class B amplifier. Other types (push-pull class A and class AB) will require only simple modification of the given information.

A single-ended, class A amplifier is shown in Figure 6.25a. The transistor is biased to some quiescent collector current, and the input signal varies the collector current about this point. For maximum output power the collector current varies between zero and twice the quiescent value while the collector voltage varies between zero and twice the supply voltage. The collector characteristics of a transistor are shown in Figure 6.25b. The loadlines (R_1, R_2, and R_3) represent three different values of collector load for the transistor in Figure 6.25a. R_1 is the optimum load resistance because it allows the maximum voltage and current swings. The loadline, R_2, represents the result of too low a load impedance: the full current swing can be realized, but the voltage swing is restricted. The loadline, R_3, illustrates the effect of too high a load impedance: the full voltage swing can be realized, but the current swing is restricted.

It can be seen from Figure 6.25b, that with an optimum load impedance, the available power output at a given supply voltage (V_c) will be determined by the quiescent collector current: the larger
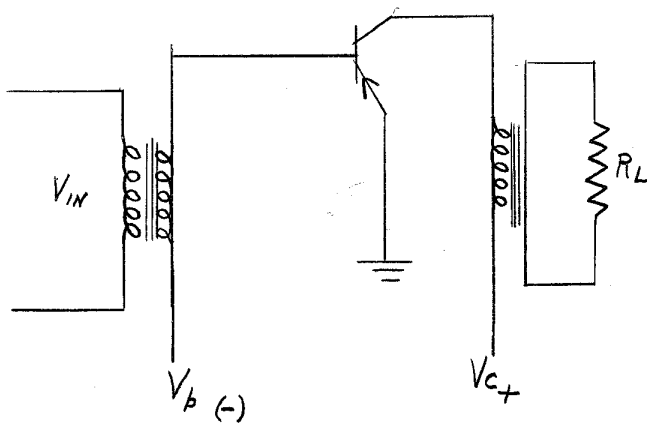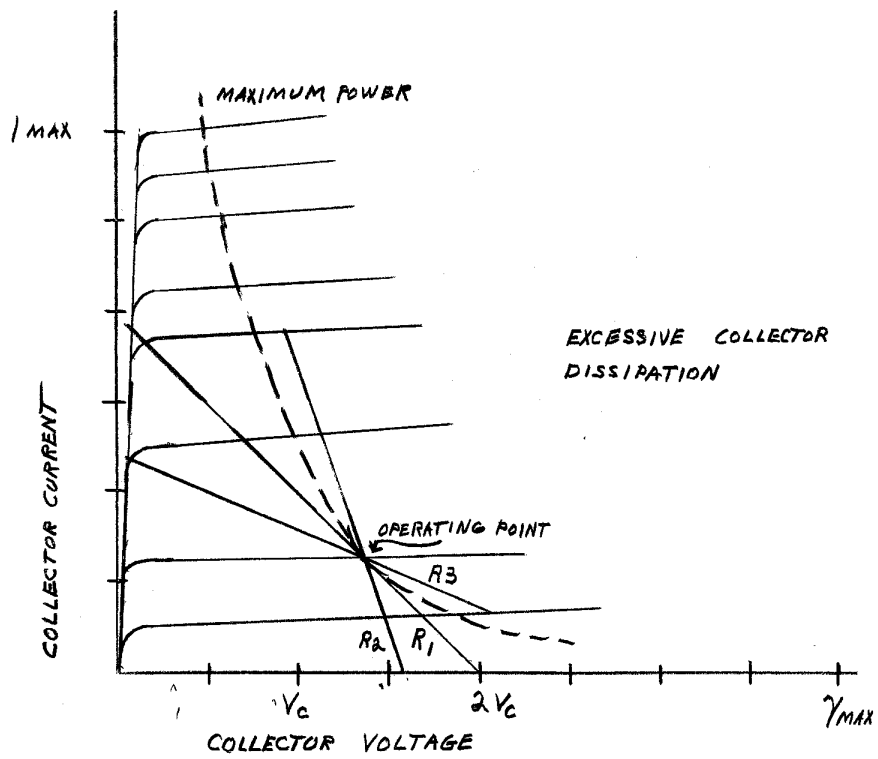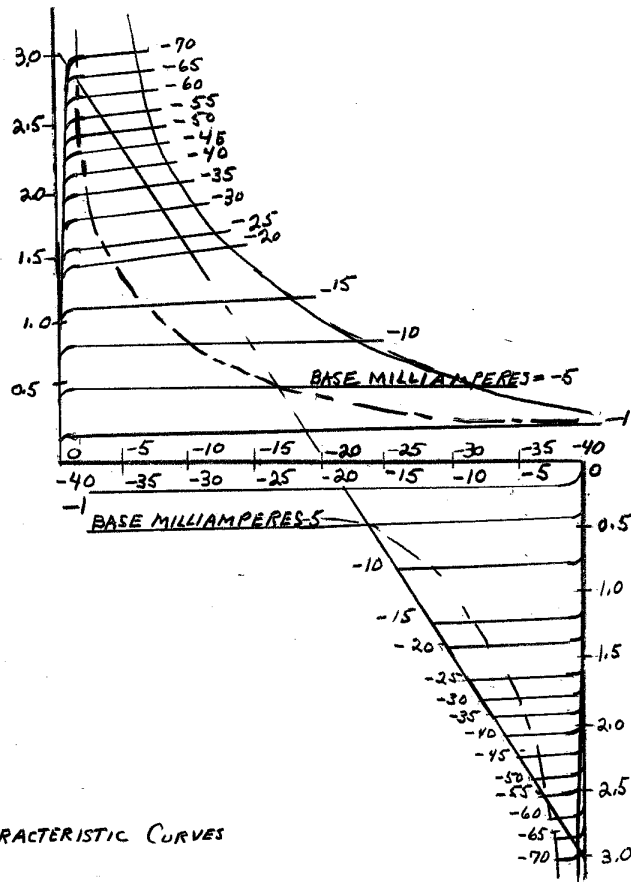
a. AMPLIFIER CIRCUIT

Figure 6.25. Single-Ended, Class A, Power Amplifier.

6-33

the quiescent collector current, the larger the current swing. However, the magnitude of the quiescent current is limited by the maximum permissible power dissipation of the transistor. The maximum theoretical power output of a class A amplifier is equal to one-half the zero-signal power dissipation, which indicates an overall collector circuit efficiency of 50%. This theoretical figure can only be approached in practice. Efficiencies in the order of 40-45% can be realized with transistor circuits.

One disadvantage of the class A power amplifier is that large amounts of d.c. power are consumed, even under no-signal conditions. This is particularly undesirable with amplifiers operating from batteries. The push-pull class B amplifier, illustrated in Figure 6.26b, overcomes this limitation. Power is drawn from the d.c. circuit only when there is an output signal. With reference to the circuit, it is evident that neither transistor is conducting under zero-signal conditions, as no forward bias is provided. However, when a signal is applied, the transistors will conduct on alternate half cycles. Hence, each transistor amplifies half the input signal and remains nonconducting during the other half.

The loadline for the push-pull amplifier can be represented using the composite characteristic curves shown in Figure 6.26a. To draw these curves it is necessary to align the two individual curves so that the collector voltages correspond at the supply voltage, as shown in the figure. The loadline must then pass through zero collector current at this point of correspondence. With the push-pull class B amplifier, the output power increases with decreasing load resistance: the maximum voltage swing will be very nearly equal to twice the supply voltage regardless of the load resistance, so a smaller load will give a larger current swing and a higher power output. The minimum load resistance, and therefore the maximum power output, is limited by the peak current or maximum power ratings of the transistor - whichever would be exceeded first.

The theoretical, maximum efficiency of a class B amplifier is 78.5%. Ideally, the efficiency is independent of load resistance because, as the power output is increased or decreased by changing the load resistance, the input power will change proportionally. This condition is closely approximated in transistor amplifiers, and efficiencies greater than 65% are not difficult to realize over a wide range of load impedances. The maximum, average-power dissipation of the output transistors does not occur at maximum power output with class B amplifiers. It occurs at somewhat less than full output where the output efficiency falls to approximately 50%. This must be taken into consideration with the permissible dissipation of the transistor.

a. COMPOSIT CHARACTERISTIC CURVES
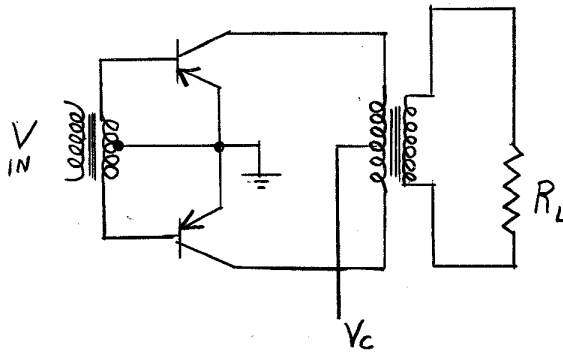
b. AMPLIFIER CIRCUIT

Figure 6.26. Push-Pull, Class B Power Amplifier.

In comparing the output circuits of the class A and class B amplifiers, it can be said that the letter is more conservative from the standpoint of power drain, both because of its higher efficiency and because of its low standby current. The advantages of the class A amplifier arise from the economy of using only one transistor and also from the less stringent control of transistors required to reduce distortion to a given level. It should be obvious that the characteristics of the output transistors of a class B amplifier must be closely matched to produce an acceptably low distortion level. This task is difficult, and often some form of negative feedback is required.

It should be noted that the previous discussion covered only common-emitter amplifiers. This is not meant to imply that other types are not used; however, as far as the output circuit is concerned, common-emitter and common-collector amplifiers are identical. The difference lies in the input circuit. The output characteristics of the common-base circuit are slightly different, but the same considerations apply. Again, with the common-base circuit, the main difference lies in the drive circuits.

Drive Requirements. It has already been stated that no particular valve can be assigned to such parameters as the input impedance and current because of variations with signal level. Nevertheless, under certain conditions these parameters can be assigned a nominal value if operation is sufficiently linear.

In coupling the driver to the output stage the objective is not to transfer maximum power; but instead, the driver should be matched to the output stage in such a way as to produce minimum distortion. This point is clearly illustrated in Figure 6.27 which gives the transfer characteristics of a common-base amplifier. Figure 6.27a shows that the voltage transfer characteristics of a common-base amplifier are extremely nonlinear: that is, if a sine wave voltage were applied to the input of a common-base amplifier, the collector current would not be a sine wave -- an appreciable amount of distortion would be generated. However, if a current sine wave were applied to the input, it can be seen from Figure 6.27b that the collector current would be a sine wave; and very little distortion would be generated. Hence, it is desirable to drive the common-base amplifier from a high-impedance source.

This behaviour can be explained with the help of the simplified equivalent circuit of Figure 6.28. The reason for the nonlinearity in the voltage transfer characteristics is the variation in the input impedance with instantaneous signal voltage; therefore the input current will not vary linearly with input voltage. The current transfer characteristics, on the other hand, are linear because the output current is equal to the product of the input current and the alpha current gain (α) which is always very nearly equal
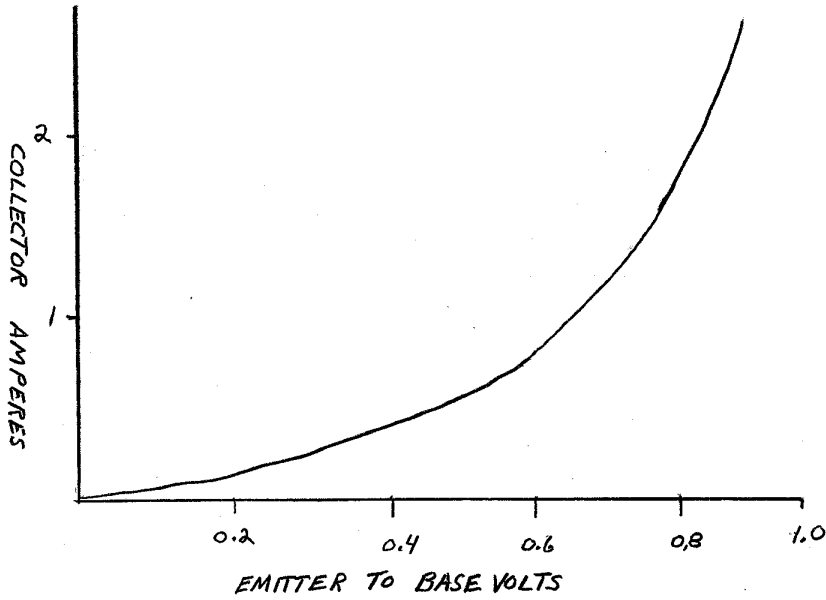
a. VOLTAGE TRANSFER CHARACTERISTICS
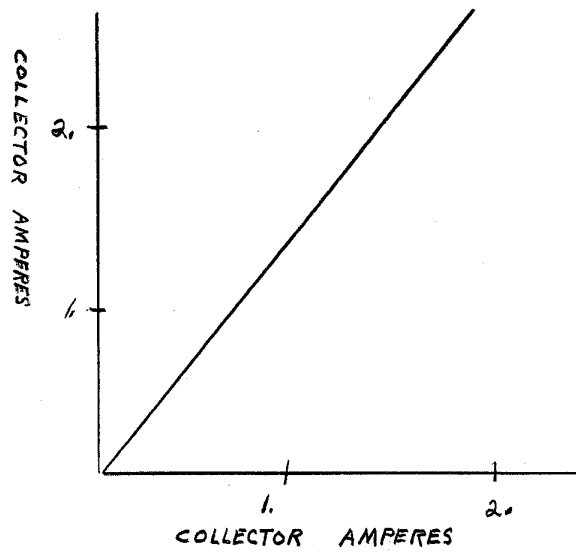
b. CURRENT TRANSFER CHARACTERISTICS

Figure 6.27. Transfer Characteristics for a Common-
Base Amplifier.

6-37

to one. In Figure 6.28, the driver is represented by a constant voltage source with its internal impedance in series (R_g). If this internal impednace is high compared to the low input impedance of the transistor, the current flowing into the emitter will depend on the source voltage and the internal impedance: the fluctuations of input impedance in the transistor will produce very small changes in the total impedance seen by the voltage source, so the input current will have the same waveshape as the source voltage. If, however, the driver impedance is low compared to the input impedance, fluctuations of the latter quantity will produce appreciable changes in the total impedance seen by the voltage source; and the input current will depend not only on the source voltage and internal impedance, but also on the input impedance which is not constant. In this case the input current does not have the same waveshape as the source voltage, and distortion is produced.

This argument shows that it is undesirable to match the driver impedance to the input impedance of the output stage. Instead, it is desirable to have the driver impedance very much higher than the input impedance. This makes for inefficient transfer of power between the driver and output stage. Moreover, the low (unity) current gain of the common-base amplifier puts excessive current demands on the driver. For these reasons, the common-base configuration is not used frequently as a linear power amplifier.

In the case of a common-emitter amplifier, both the voltage and current transfer characteristics are nonlinear to some extent - although the nonlinearity is not as great as with the common-base configuration. The base input characteristics in Figure 6.29a, however, do resemble those of the common base connection: the input impedance decreases with increasing input voltage. A given change in base voltage will be more effective in producing a change in base current at higher input voltages. Nevertheless, this effect is not so pronounced in the voltage transfer characteristics because of the fall-off of common-emitter (α_E) current gain at high current levels. This fall-off explains the nonlinearity in the current-transfer plot (Figure 6.29c).

Because of the nonlinearity in both the voltage and current transfer characteristics, it cannot be stated whether a high or low impedance driver will produce the lowest distortion. Nevertheless, a comparison of the two plots will show that the voltage characteristic has an increase in slope at high collector currents while the current characteristic has a decrease in slope at high collector currents. It would seem, therefore, that some intermediate value of driver impedance would produce the lowest distortion. This is indeed the case, and because of this it is possible to have a reasonable impedance match between the driver and the output stage consistent with low distortion. The common-emitter amplifier will therefore,
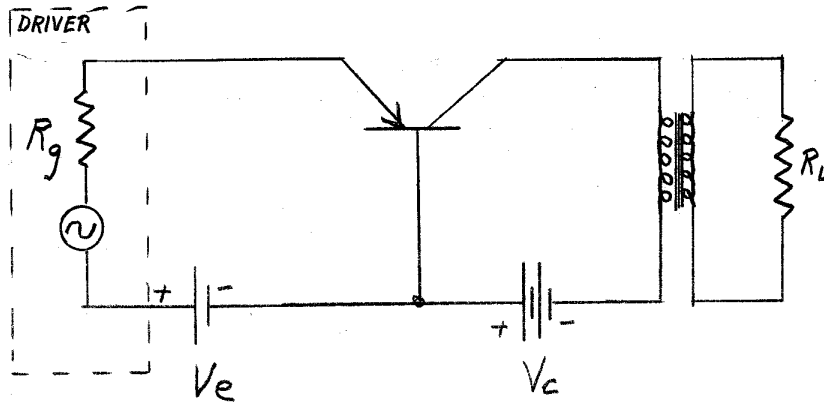
6-38

Figure 6.28. Simplified Equivalent Circuit for Common-
Base Amplifier and Driver.

$a.$ BASE IMPUT CHARACTERISTICS $b.$ VOLTAGE TRANSFER CHARACTERISTICS $C.$ CURRENT TRANSFER CHARACTER
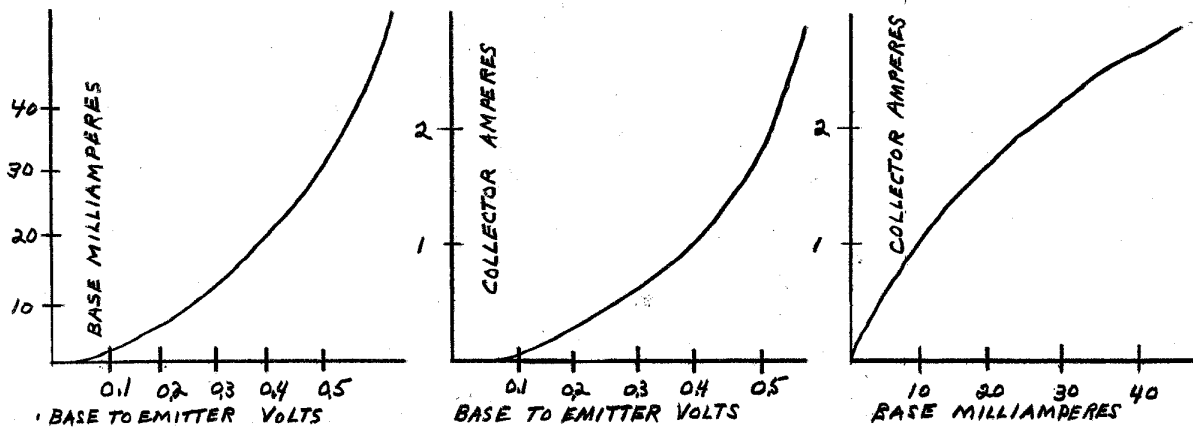

Figure 6.29. Transfer Characteristics for a Common
Emitter Amplifier

make efficient use of driver power; moreover, both the driver and the output stage will have a much greater power gain as compared to the common base amplifier. For these reasons the common emitter amplifier is commonly used in power amplifiers.

The common-collector power amplifier can be considered to be a common-emitter stage with 100% negative voltage feedback, applied in such a way as to increase the apparent input impedance and linearize the voltage transfer characteristic. It is clear, therefore, that the common-collector amplifier should be driven from a low-impedance source. One disadvantage of this configuration is the large voltage-swing requirement that must be met by the driver: the negative feedback reduces the voltage gain to less than unity. Driven from a constant voltage source, the common-emitter amplifier has a very low distortion; but if it is driven from a constant current source, its operation will be identical to that of a common-emitter amplifier.

The above results can be summarized as follows: the common-base power amplifier should be fed from a constant-current (high-impedance) source, and the common-collector amplifier should be fed from a constant-voltage (low-impedance) source. Moreover, the former will tend to maintain a constant current through a varying load impedance in its output circuit so it can be considered to have a high impedance output, while the latter will try to maintain a constant output voltage across a varying load so it can be considered a low impedance source. Both these configurations have a low power gain when driven properly and put excessive current or voltage requirements on the driver. Therefore, they are not too frequently used in practice.

The common-emitter power amplifier has a high power gain. In addition, for minimum distortion, the driver is very nearly matched to the output stage so there is a reasonable efficiency in using the driver power. This configuration is the most practical for use as a power amplifier so the major portion of the circuits given will be confined to the common-emitter circuit.

Driver Coupling. The driver can be r-c coupled to a class A output stage as shown in Figure 6.30a. However, because some significant amount of power must be generated in the driver, this method is frequently too inefficient. Furthermore, the nonlinearity in the transistor input resistance can cause charging of the coupling capacitance at high signal levels, which will shift the operating point of the transistor. Hence, r-c coupling is used only in low-power amplifiers, such as driver stages.

An example of an r-c coupled, class B amplifier is given in Figure 6.30b. Diodes must be included in the input circuit to prevent

FROM DRIVER

R_L

$-V_c$

a. CLASS A AMPLIFIER

R_L

V_c
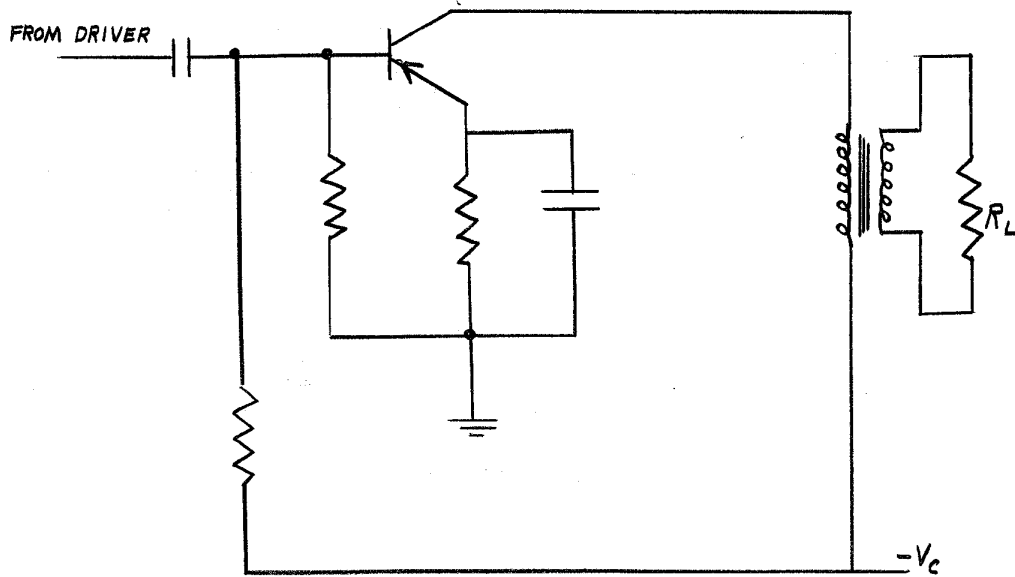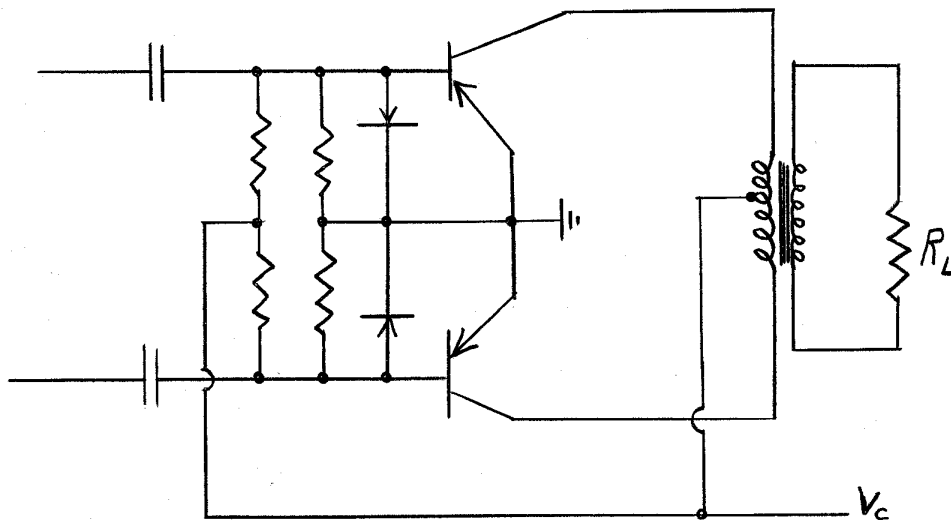
b. CLASS B AMPLIFIER

Figure 6.30. Resistance-Capacitance Coupling.

charging of the coupling capacitors by the half-cycle conduction of the output transistors. A small, forward bias is applied to the output transistors, as with any class B stage, to minimize "crossover distortion" when the conduction shifts from one transistor to the other. As with class A amplifiers, the use of r-c coupling is quite limited.

The usual method of coupling the driver to the output stage is transformer coupling. A transformer allows a good efficiency in the driver stage as well as giving any desired impedance ratio. A driver transformer also eliminates the necessity of a phase-inverter in push-pull circuits.

A practical class A amplifier is shown in Figure 6.31. The driver stage is transformer coupled to the output transistor. The output transistor is operating with a split load: that is, half the load is in the emitter and half in the collector circuit. This combines the advantages of common collector with common emitter by applying 50% negative feedback. Amplifiers of this type can be built to deliver powers greater than 5 watts using germanium transistors.

A transformer coupled, class B amplifier is shown in Figure 6.32. A class A amplifier is used as a driver. The output stage has a small forward bias to minimize crossover distortion. This bias is temperature compensated with a germanium diode. Negative feedback is applied to the emitter of the driver from the secondary of the output transformer to minimize the distortion caused by inequalities in the output transistors. Class B amplifiers with outputs as high as 200w can be built using available transistors.

Direct coupling can also be used to couple the driver to the output stage, but this method presents the same impedance-mismatch problems as are experienced with r-c coupling. Also, there is the problem of d.c. instability arising from temperature effects. Direct coupled power amplifiers are used, however, where it is desirable to eliminate the size and weight as well as the frequency and phase shift limitations of the transformer. An example of a direct-coupled power amplifier, the class B complementary-symmetry amplifier, has already been given (Figure 6.24). This circuit uses complementary (matched PNP, NPN) power transistors which are difficult to obtain commercially so this circuit has only limited uses. Another circuit, the quasi complementary amplifier, (Figure 6.33) has a direct-coupled, push-pull output stage and uses only PNP power transistors. In this circuit, the transistors (Q_1, Q_2, Q_3, and Q_4) are not conducting under no-signal conditions so there is no voltage across the load. A negative input signal on the base of Q_3 drives this transistor (Q_3) and, consequently, Q_1 into conduction; and the negative half of the signal appears across the load. A positive signal, applied at the
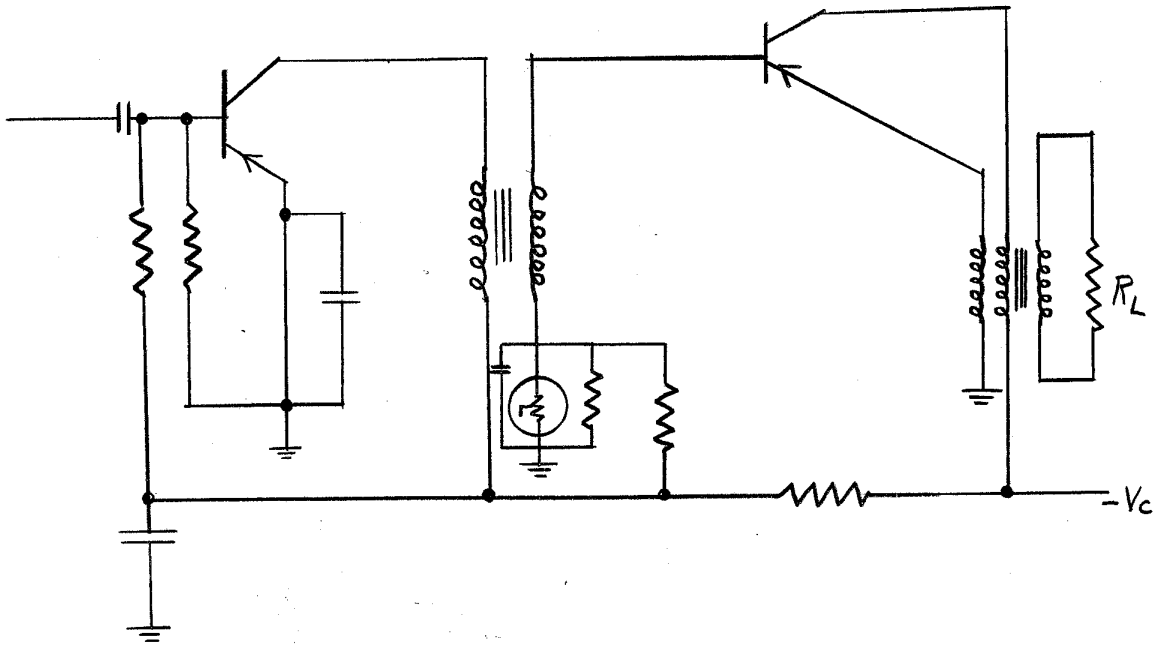
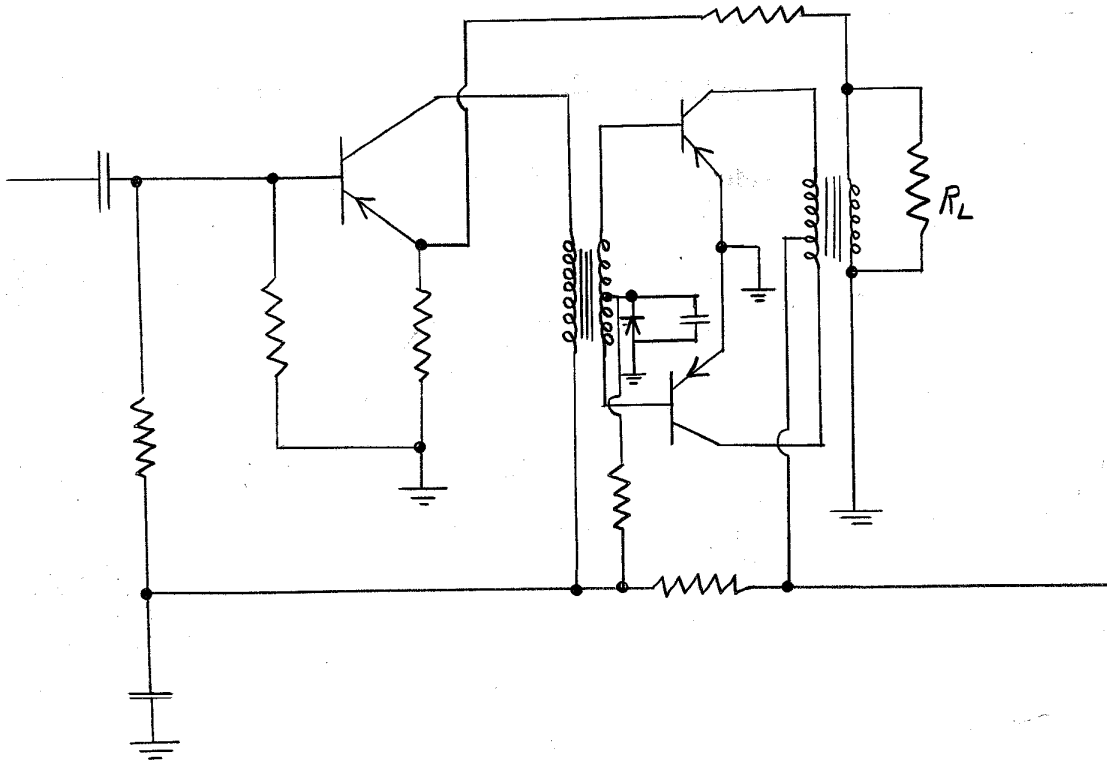Figure 6.31. Class A Power Amplifier.

Figure 6.32. Class B Power Amplifier.

same point, drives the NPN transistor, Q_4, into conduction. When Q_4 conducts, Q_2 will also conduct and the positive half of the signal will appear across R_L. The upper (Q_3 and Q_1) and lower (Q_4 and Q_2) channels are not symmetrical, but the application of 100% negative feedback eliminates signal unbalance. The upper channel uses two PNP transistors in the common-collector connection. The input signal is applied to the base of Q_3, and its emitter is connected to the input (base) of Q_1. The emitter of Q_1 goes to the load. The collectors of both Q_3 and Q_1 are brought to signal ground at the negative terminal of V_1. Because of this common collector connection, the voltage gain of the upper channel is very close to one. The lower channel uses a NPN and a PNP transistor cascaded in common emitter connection. The input signal is connected to the base of Q_4, and its collector is direct coupled to the base of Q_2. The collector of Q_2 is connected to the load. The voltage gain of cascaded common-emitter stages is normally very much greater than one, but 100% negative feedback is applied to this channel by connecting the emitter of Q_4 to the output rather than to signal ground. The feedback reduces the voltage gain of this channel to slightly less than one; therefore, the voltage gain of both the upper and lower channels is nearly equal; and they are electrically balanced. Signal is fed to the amplifier by the class A driver, Q_5. If it is assumed that Q_5 is operating at its maximum swing, it can be seen that the bases of Q_3 and Q_4 will alternately be connected to the negative battery (through R_1) when Q_5 is nonconducting and to the positive battery when Q_5 is saturated.

Direct-coupled power amplifiers are particularly applicable in high-fidelity systems. The elimination of the output transformer can reduce cost considerably while equivalent circuit performance is maintained.

Conclusions. There is much to be said for transistor power amplifiers, their high (near theoretical) efficiency being only one advantage. Furthermore, transistor amplifiers can operate directly from standard d.c. systems (6, 12 and 28V), thus eliminating the need for power converters. Also, their small size permits reduction in equipment size. However, they are not without limitations: one is that the beta cutoff frequency of many alloy-junction power transistors is near the high end of the audio frequency band; a second is the restricted temperature range of germanium transistors which imposes severe limits on collector dissipation above about 70°C. Both of these limits have been appreciably extended with silicon transistors employing a diffused-base structure, but at the present time the cost of these units is quite high.

HIGH-FREQUENCY AMPLIFIERS

There are two important limitations to the high-frequency performance of a transistor: the first is the finite transit time for current
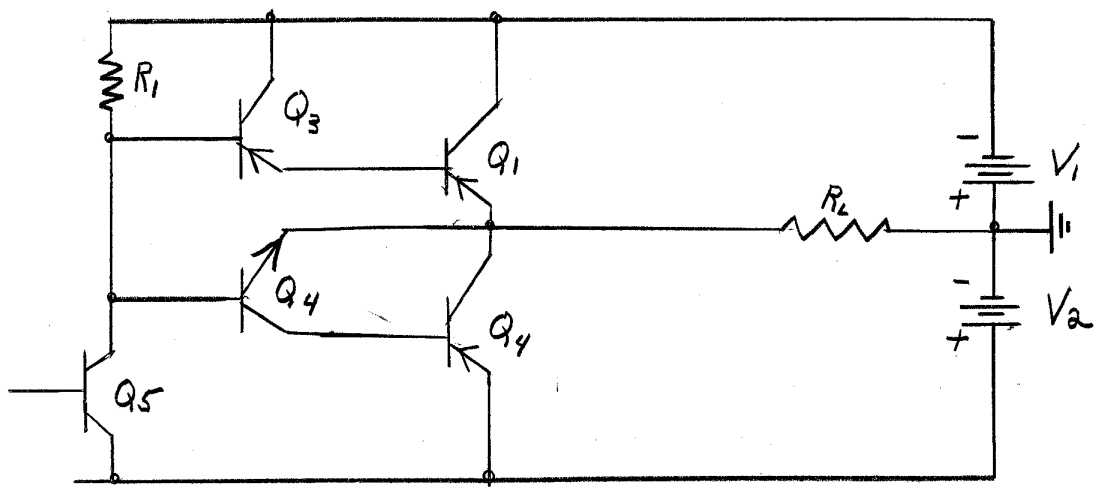
Figure 6.33. Simplified Schematic of the Quasi-Complementary Class B Amplifier.
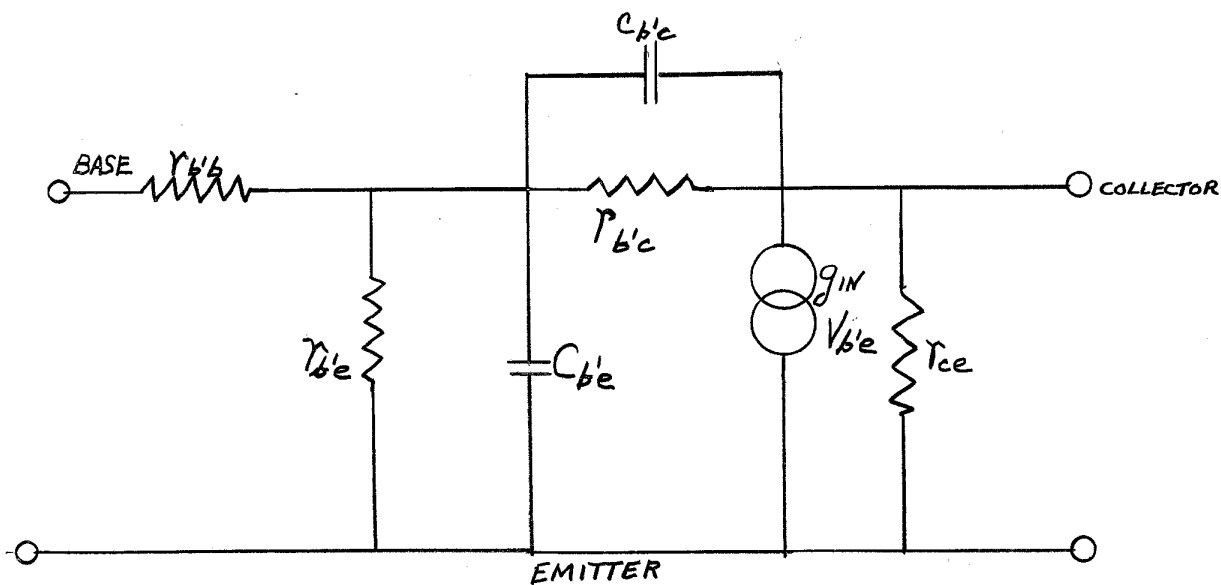

Figure 6.34. Transistor Equivalent Circuit.

carriers crossing the base, and the second is the collector-junction capacitance. The transit time effect is reflected in the alpha cut-off frequency ($f_{\alpha_{co}}$) rating of the device. This is an absolute limitation; that is, there is no way to design around, or compensate, for this frequency restriction. On the other hand, the collector capacitance can be tuned out or compensated for to some extent.

An equivalent circuit can be used to approximate the characteristics of a transistor. These circuits do not describe transistor performance but only approximate it over the useful frequency range for purposes of circuit analysis. Figure 6.34 gives one possible equivalent circuit convenient for analysis of common emitter circuits. The capacitance shunting the input circuit ($C_b{}^1{}_e$) represents the alpha cutoff frequency limitations, while the collector junction capacitance is shown as $C_b{}^1{}_c$. It can be seen that $C_b{}^1{}_c$ provides a feedback path between the output and input circuits in addition to shunting the output circuit ($C_b{}^1{}_e$ is very much greater than $C_b{}^1{}_c$). This feedback path frequently necessitates the use of neutralization in tuned amplifiers.

Video Amplifiers. High-frequency amplifiers will be divided into two groups: video amplifiers which are intended to cover a wide band of frequencies from about 10cps to about 10mcps, and tuned amplifiers which employ resonant circuits and cover a relatively narrow frequency range.

The appearance of transistor video circuits is quite similar to the low-frequency amplifiers already discussed. Common base, common emitter, and common collector stages are possible. Because of the wide band of frequencies to be passed, it is necessary to use either direct or r-c coupling as it is generally impossible to build transformers with the required bandwidth. In order to cascade single-stage amplifiers and realize any gain, then, there must be a reasonable impedance match between the input and output impedances of successive stages since there is no means available for matching. For this reason, video amplifiers almost always use the common-emitter configuration except, possibly, in the input or output stages where there is some special impedance requirement. Hence, the discussion will be confined to this circuit.

In common-emitter amplifiers the beta cutoff frequency is of interest; it is related to the alpha cutoff frequency by
$$f_{\beta_{co}} = f_{\alpha_{co}} (1-\alpha_o)$$
where α_o is the low-frequency alpha current gain. From this equation it can be seen that the beta cutoff frequency is considerably lower than the alpha cutoff frequency. How much lower will depend on the low-frequency alpha current gain; therefore, transistors with exceptionally high current gain are not generally desirable.

Figure 6.35 shows the characteristics of a common-emitter video amplifier that has its upper limit determined by the beta cutoff frequency. Lowering the load resistance below the optimum value reduces gain but does not raise the upper frequency limit. However, the insertion of a compensating inductor will increase the maximum frequency, because the coil raises the effective load impedance at higher frequencies.

As a rule, alloy-junction transistors and some grown-junction transistors suffer from this type of frequency limitation. These transistors do not have an accelerating (drift)field across the base region so the transit time is comparatively large.

Transistors that have their maximum frequency imposed by collector-junction capacitance can be compensated to some extent as shown in Figure 6.36. In this case lowering the load resistance will increase the bandwidth in addition to reducing gain because the shunt impedance of the output capacitance will not be comparable to the load resistance until a higher frequency is reached. On this basis, gain can be traded evenly for bandwidth. Moreover, compensation will extend the maximum frequency in either case because, in this case, the compensating inductor will form a low-Q parallel-resonant circuit with the output capacitance and effectively remove it from the circuit over a limited frequency range.

Transistors having a graded base giving a drift field, such as diffused junction and some grown junction transistors, have their high-frequency performance limited by collector-output capacitance. However, in most cases, the maximum frequency of these devices is considerably greater than those without the drift field.

Another circuit worth mentioning is shown in Figure 6.37. This is a video amplifier using a tetrode transistor. The same consider-ations apply here as did in the previous discussion. The only unusual feature of this circuit is that a cross-base bias must be supplied to the transistor to electrically reduce the junction area. Bias of the correct polarity is applied as shown in the figure.

As was pointed out, video amplifiers are very much like direct or r-c coupled, low-frequency amplifiers except that high-frequency transistors must be used. There are, nevertheless, other points that should be considered. First, the cutoff frequency is raised by increasing the collector voltage. The increased voltage widens the collector depletion region and, therefore, reduces output capacitance. Furthermore, the depletion region extends into the base so base width and, consequently, transit time is reduced. Secondly, both the input and output impedances are considerably lower at high frequencies. This is an important point where stages are to be cascaded.
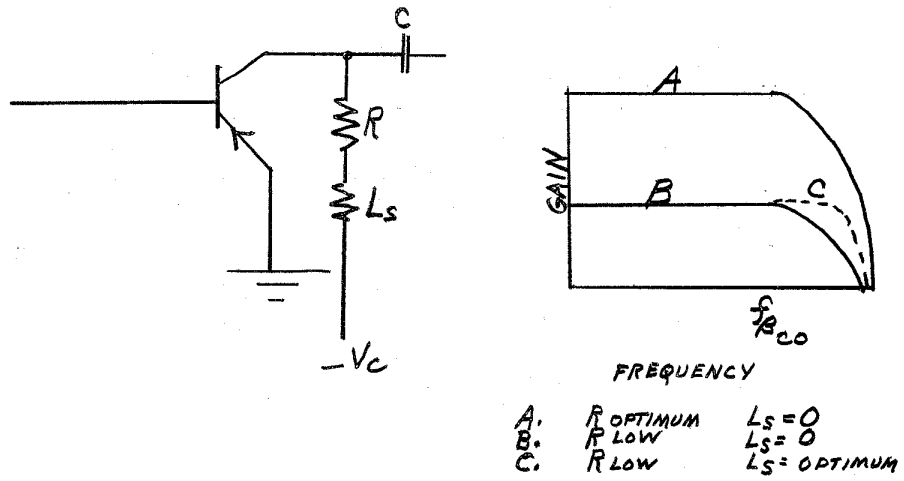
Figure 6.35. Output Circuit of Video Amplifier Limited
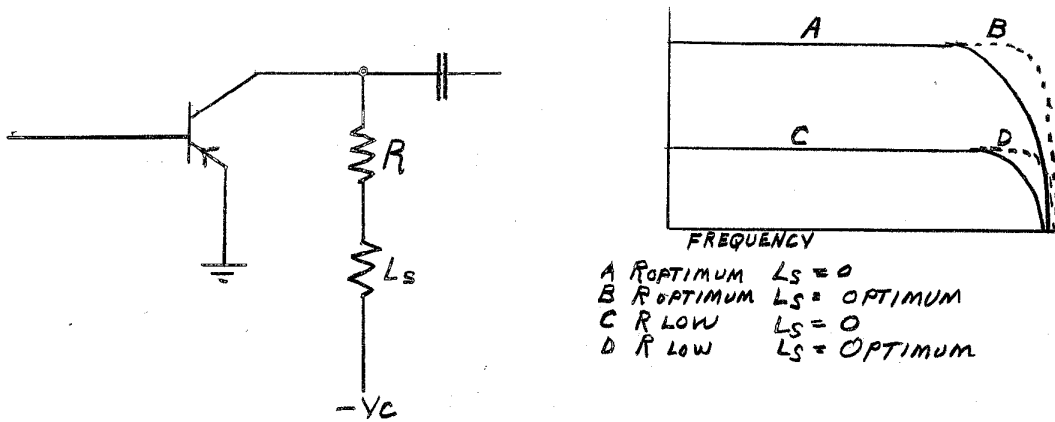By Beta-Cutoff Frequency.

Figure 6.36. Output Circuit of Video Amplifier Limited
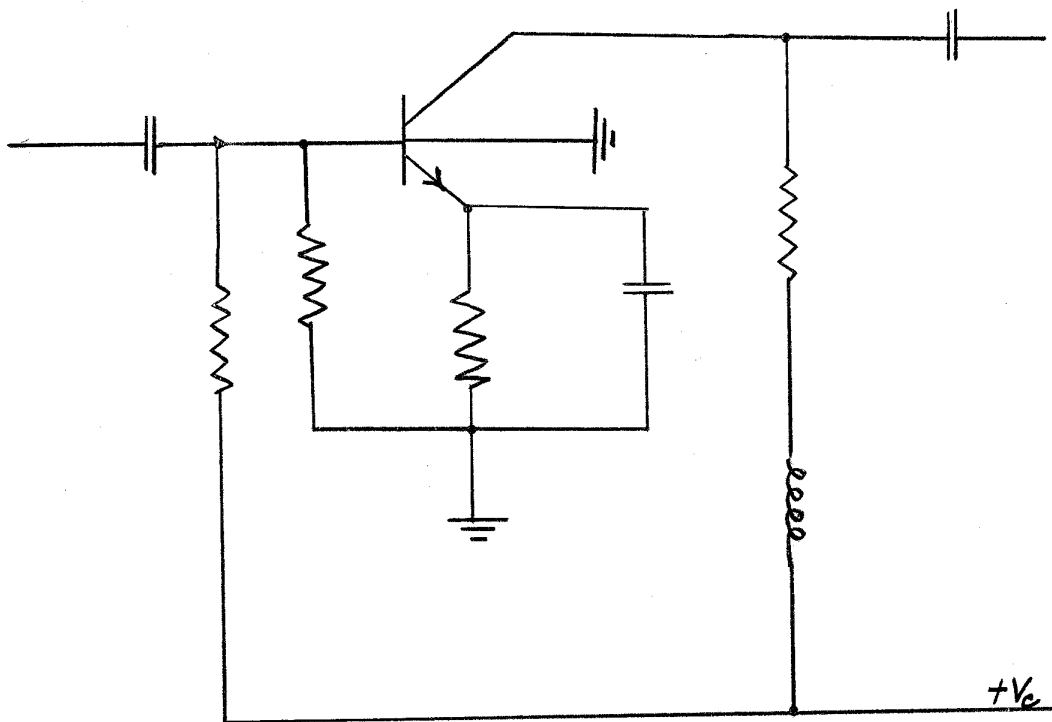by Collector-Junction Capacitance.

Figure 6.37. Video Amplifier Using a Tetrode Transistor.

Tuned Amplifiers. In a tuned amplifier, the major frequency limitation is the alpha-cutoff frequency since the collector junction capacitance can be made part of the resonant circuit in the collector. This capacitance, nevertheless, does provide a feedback path and in some cases neutralization must be provided for stable operation.

Impedance transformation is no problem with tuned circuits as is shown in Figure 6.38, so in tuned amplifiers impedances are usually matched to realize maximum gain. This, of course, is imperative with the common-base connection because the entire gain arises from the difference between input and output impedances. The matched-impedance power gain of the common emitter configuration is higher than both common collector and common emitter so this connection is generally preferred. However, at frequencies near the alpha-cutoff of the transistor, the difference between the common emitter and common base connections becomes small; and the latter is frequently used in the interest of circuit stability.

With present day transistors, the collector-junction capacitance is small enough so that a common emitter amplifier can be operated without neutralization. But in this case the stage cannot be designed for maximum gain and still be stable. Neutralization permits operation with maximum gain; however, circuit adjustment will be more critical and transistor interchangeability will be poorer. Nevertheless, one neutralized stage is roughly equivalent to two unneutralized stages.
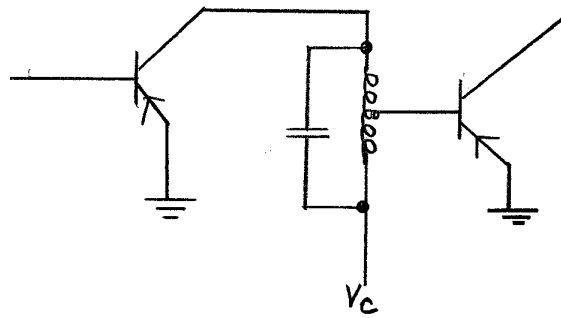
Figure 6.39 illustrates the difference between neutralized and unneutralized r-f amplifiers. In the neutralized stage, double-tuned transformers are used. Moreover, the low-impedance input (base) of the transistor as well as the collector are tapped down on the tuned circuits to prevent loading which would reduce effective Q. The primary winding in the collector circuit is tapped; and a signal that is 180° out of phase with the collector voltage is fed back to the base through the neutralizing capacitor (C_n). This signal cancels that which is fed back to the base through the collector-junction capacitance. Because of variations in parameters from unit to unit of the same type, this circuit may become unstable with a particular transistor requiring readjustment of the neutralizing capacitance.

In the unneutralized stage (Figure 6.39b), single-tuned transformers are used; and the transformers are not tapped for optimum performance. This reduces stage gain, but at the same time it permits operation without neutralization. This, however, requires careful design to achieve a certain degree of stability with a minimum loss of gain.
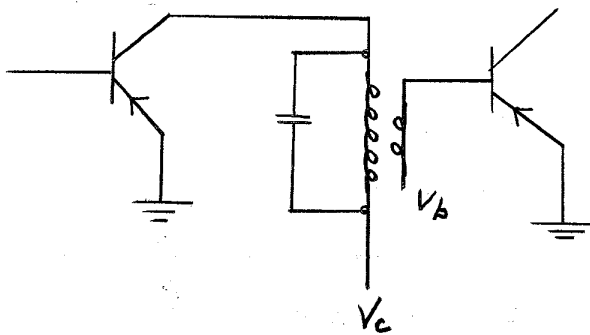
As was mentioned before, the common-base connection is used at frequencies near the alpha cutoff frequency of the transistor
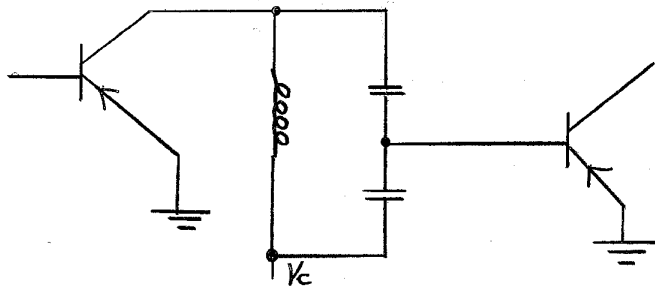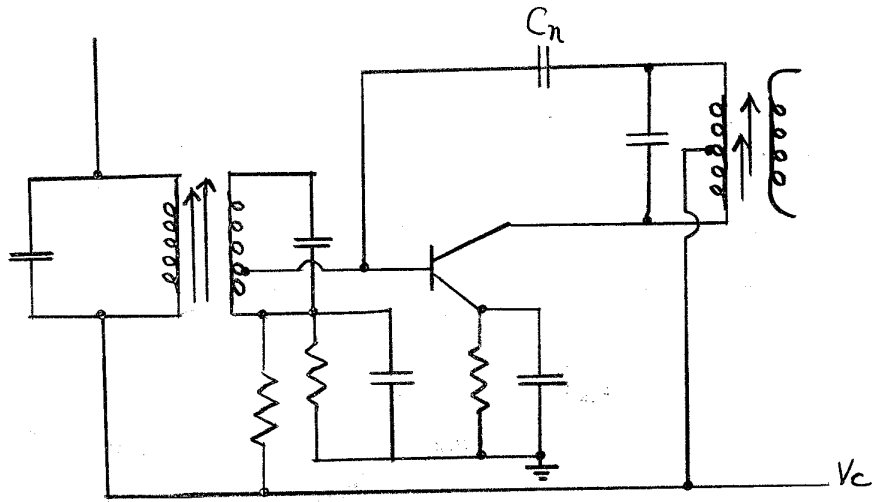
a. TAPPED CAPACITOR
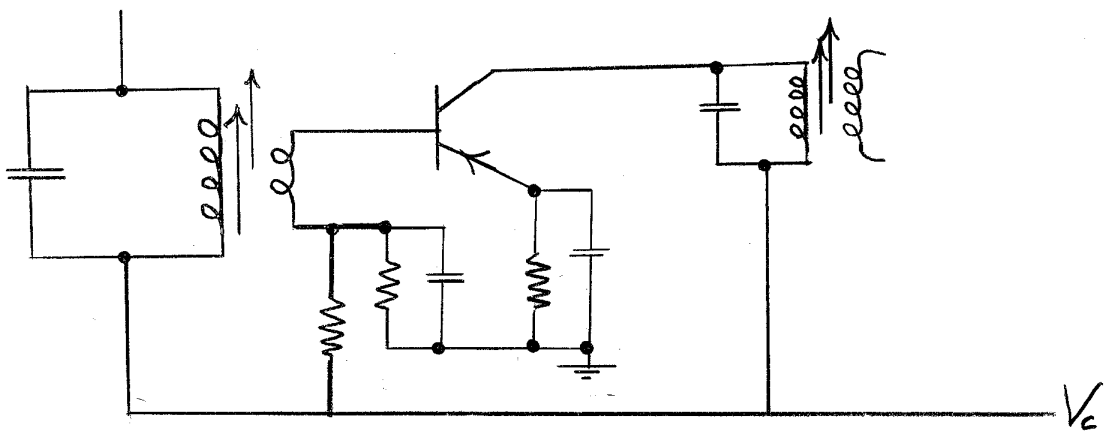
b. TAPPED INDUCTOR

c. SINGLE-TUNED TRANSFORMER

Figure 6.38. Interstage Impedance Matching for Tuned
Amplifier.

a. NEUTRALIZED AMPLIFIER

b. UNNEUTRALIZED AMPLIFIER

Figure 6.39. Tuned R. F. Amplifiers.

where the difference between common-emitter and common-base power gain becomes small. A common-base V.H.F. amplifier is shown in Figure 6.40. The circuit is straightforward, and it infrequently requires neutralization. Careful design should be used to conserve all possible gain because the transistor is operating close to its maximum frequency, and the available gain is low.

In most r-f receiving equipment, it is desirable to incorporate some form of automatic gain control (a.g.c.). The easiest way of accomplishing this is to vary the transistor bias, but there are many effects that must be considered. If the collector current is reduced below some optimum value, the power gain will be reduced. This can be accomplished by changing the d.c. bias. However, a change in bias will also effect the input and output impedances of the device. If the circuit is adjusted so that all the impedances are matched for maximum gain, this impedance shift will also contribute to the gain reduction which will add to the a.g.c. control of the stage. Care must be taken so that the change in impedance does not cause instability.

Conclusions. The previous section showed that high-frequency operation is more dependent on transistor characteristics than it is on circuitry. The circuit used for a particular application does not change greatly as the frequency is advanced. However, at higher frequencies, more care must be taken in insuring d.c. stability of the transistor. Steps taken to increase the operating frequency of a transistor usually decrease the maximum power dissipation so, even in low-power circuits, transistors are operating near this limit.

OSCILLATORS

An oscillator, as referred to here, is an electronic device that generates an a.c. signal of some kind. Oscillators will be divided into two general categories: sinusoidal oscillators which generate a single-frequency sine wave and nonsinusoidal oscillators, such as blocking oscillators, multivibrators, and sawtooth generators, which generate some particular nonsinusoidal waveform.

Sine-Wave Oscillators. An oscillator contains an amplifier and a positive feedback path of some sort. A sinusoidal oscillator must also have some frequency-sensitive element - such as a resonant tank, a quartz crystal, or a phase shift network which determines the frequency of oscillation. In order for the oscillator to function, the amplifier must be capable of supplying the power lost in the load and various circuit elements as well as its own input power.

Several oscillators using a resonant tank as a frequency-determining element are shown in Figure 6.41. The first two use an additional winding on the resonant tank to supply the feedback which
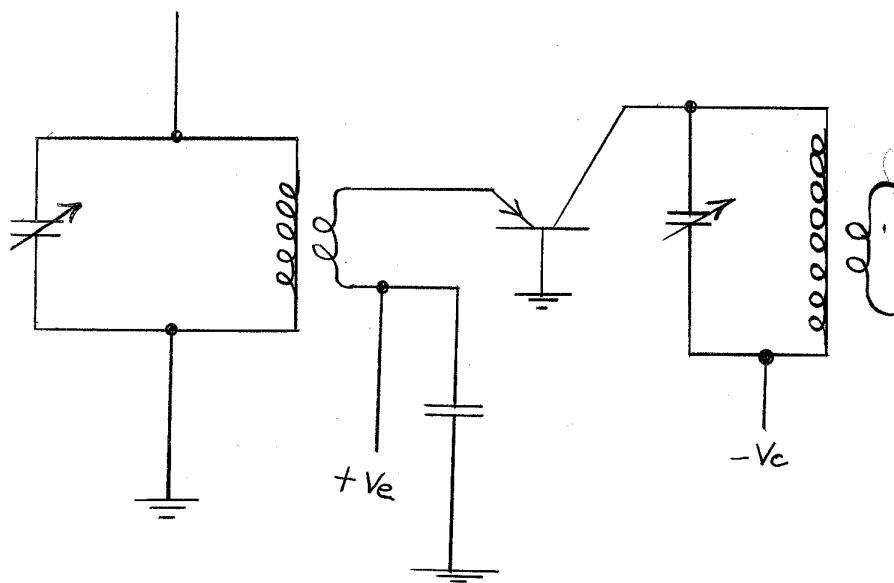
Figure 6.40. Common-Base V.H.F. Amplifier.

maintains oscillation. The remaining circuits use a tapped (either capacitive or inductive) tank to obtain the correct feedback between the input and output.

Figure 6.41a shows a common-emitter amplifier with a resonant tank in the output circuit. Oscillation is maintained by coupling part of this signal back to the transistor input. Figure 6.41b shows a similar circuit except that the transistor is connected in the common-base configuration. In Figure 6.41c that portion of the tank between the emitter and ground acts as the emitter load for the transistor while that portion between the emitter and base supplies the transistor drive. Figure 6.41d is identical to Figure 6.41c except that a capacitive tap is used on the tank. The circuits in Figure 6.41e and f are identical except that a capacitive tap is used on the former and an inductive tap is used on the latter. That portion of the output voltage across the lower portion of the tank appears between the emitter and the base to supply transistor drive.

The a.c. performance of all these circuits is the same. The only reason that one circuit would be preferred over another would be to simplify the d.c. circuitry or to put one particular transistor terminal at a.c. ground (e.g. whichever terminal is connected to the transistor case).

The circuits shown in Figure 6.41 are not self starting. When power is applied to the circuits no current will flow because the transistors are cut off. A small emitter to base bias must be applied to start operation. Figure 6.42 illustrates a method of supplying a starting bias to the circuit of Figure 6.41b. The voltage dropped across R_2 appears between the emitter and base of the transistor to provide forward bias. If the bleeder current through R_1 is made greater than the normal base current, this bias will remain after oscillation has started; and it can be adjusted for class A operation of the oscillator. If the bleeder current is of the same magnitude as the base current which is in the opposite direction, there will be very little voltage dropped across R_2; and the operation will be essentially class B. If the bleeder current is made still smaller and if R_2 is sufficiently large, the base current will charge the capacitor and the emitter-base junction will become reverse biased, permitting current flow only at the peak of the input cycle. Even higher values of R_1 and R_2 will cause blocking or "squeegeeing" which will result in intermittent operation or pulsing of the oscillator.

At frequencies above the alpha cutoff frequency a transistor still has some gain. This gain is low so the transistor is not too useful as an amplifier, but it can still function as an oscillator. Generally speaking, a transistor can be made to oscillate considerably above its cutoff frequency.
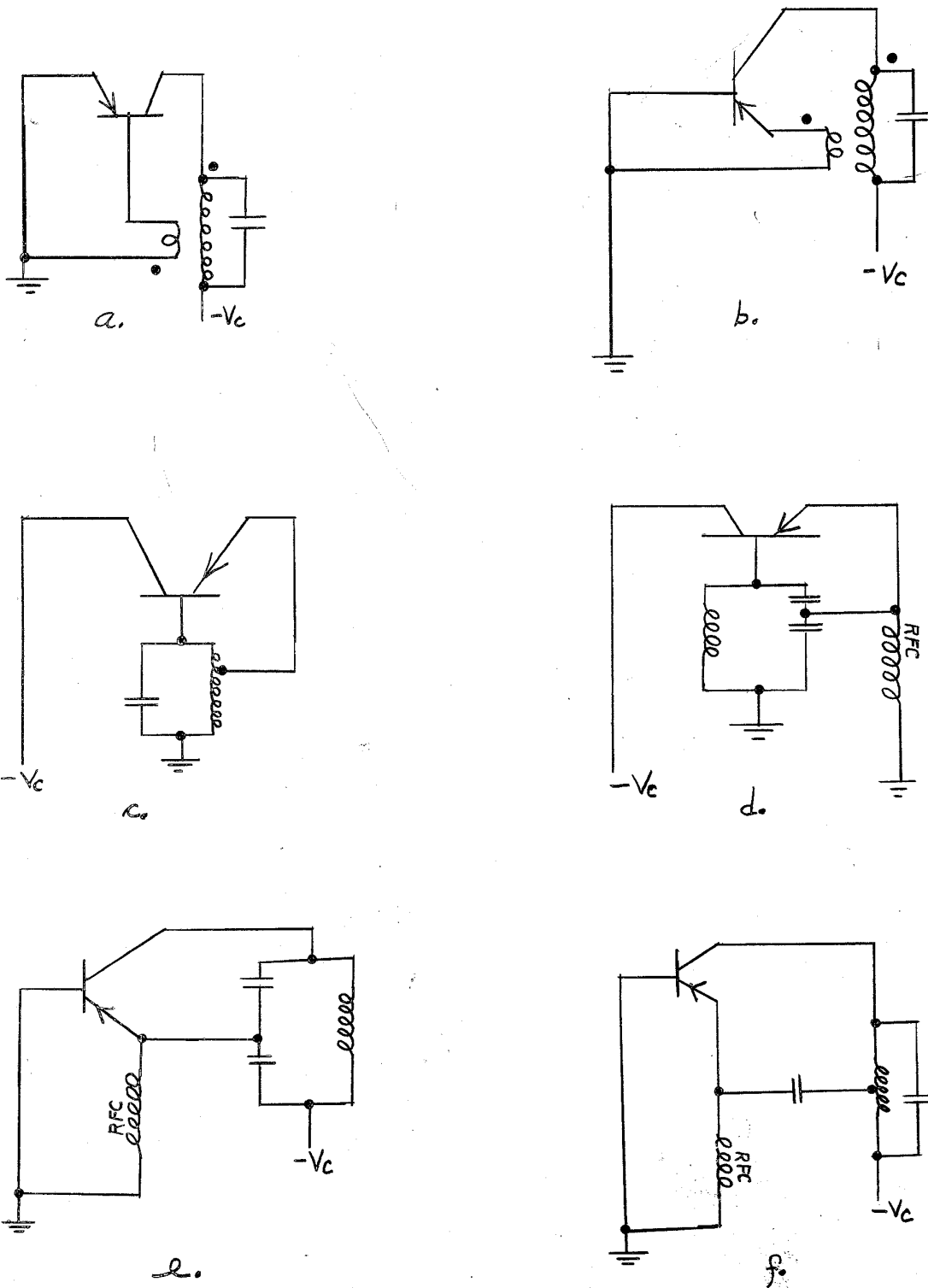
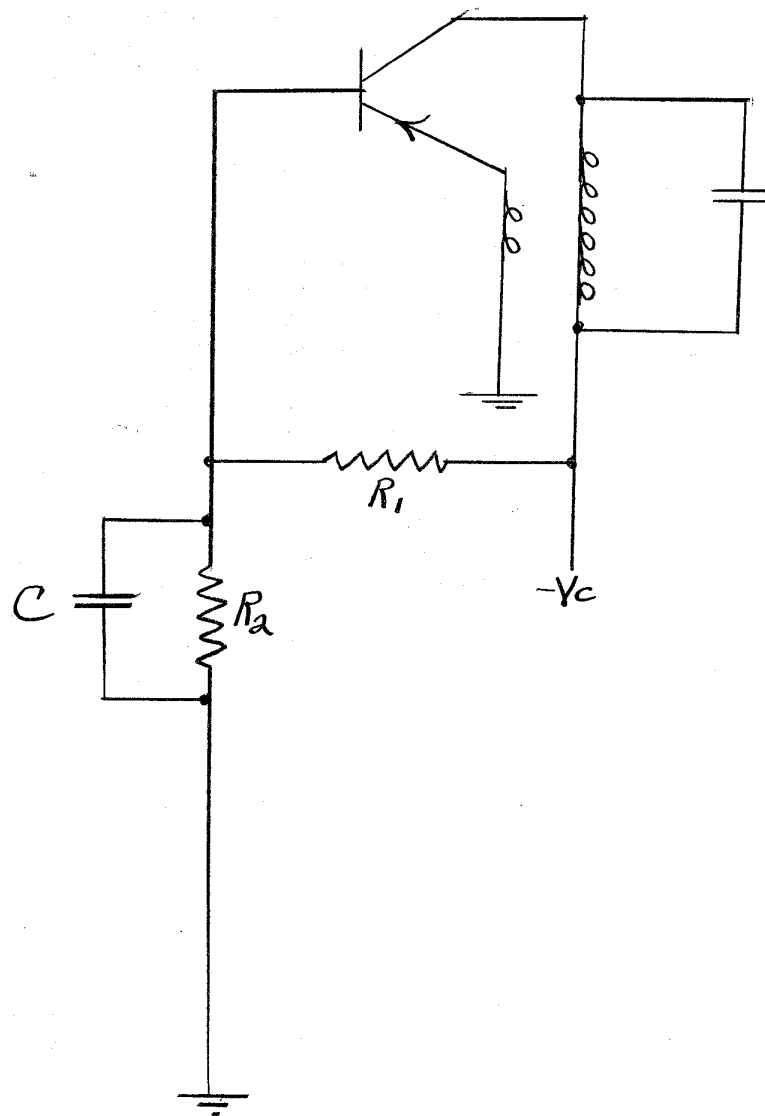Figure 6.41. Several Transistor Oscillator Circuits.

Figure 6.42. Oscillator Biasing Circuit.

Even at moderately high frequencies the phase shift within a transistor due to transit time effects becomes appreciable. This phase shift alters the frequency of oscillation slightly. Because the transit time will vary with temperature, collector voltage, etc., frequency instability may be encountered with transistors operating at higher frequencies.

There are no special requirements for high-frequency operation. However, because available power gain is usually low, particular care should be taken to match impedances and minimize circuit losses. Near the frequency limit of the transistor the emitter input may appear inductive. It is advisable to tune this out. One method is shown in Figure 6.43.

Sine-wave oscillators can be built using only resistance-capacitance networks. An example is the phase-shift oscillators in Figure 6.44. In the circuit of Figure 6.44a, the output of Q_1 is fed through a r-c network which shifts the phase $180°$. The $180°$ phase shift will occur at only one frequency so the oscillations will be sinusoidal as long as the feedback voltage does not become so great as to overdrive Q_1. The resistance of R_1 should be low compared to the input impedance of the transistor so that changes in the latter quantity do not affect the oscillation frequency. The emitter resistor, R_4, is left unbypassed to reduce distortion in Q_1 and to increase the apparent input impedance of the transistor.

Because of the low impedance of the r-c network needed to obtain frequency stability, the output circuit may become so heavily loaded that oscillation is not possible. Hence, a common collector stage is sometimes added to drive the low impedance network as shown in Figure 6.44b. This stage has no phase shift so it does not affect the operating frequency. Q_2 does reduce the collector load on Q_1 and so it increases the gain of this stage.

The phase-shift oscillator is only useful at low frequencies where the gain of the transistors is high. The feedback circuit is lossy and does not provide impedance matching so it will not operate unless there is enough gain available.

Nonsinusoidal Oscillators. A nonsinusoidal oscillator, as the name implies, is a generator of nonsinusoidal a.c. voltages. This class includes pulse generators, square-wave generators, sawtooth generators, etc. The period, or basic frequency, of these circuits is usually determined by the time constants of R-C and/or R-l circuits rather than by resonant elements.

Blocking Oscillators. The circuit of a transistor blocking oscillator along with pertinent waveforms is shown in Figure 6.45. A pulse transformer is used to obtain feedback from the transistor
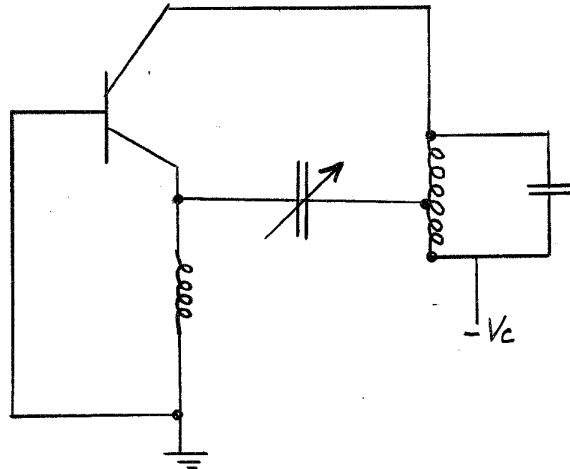
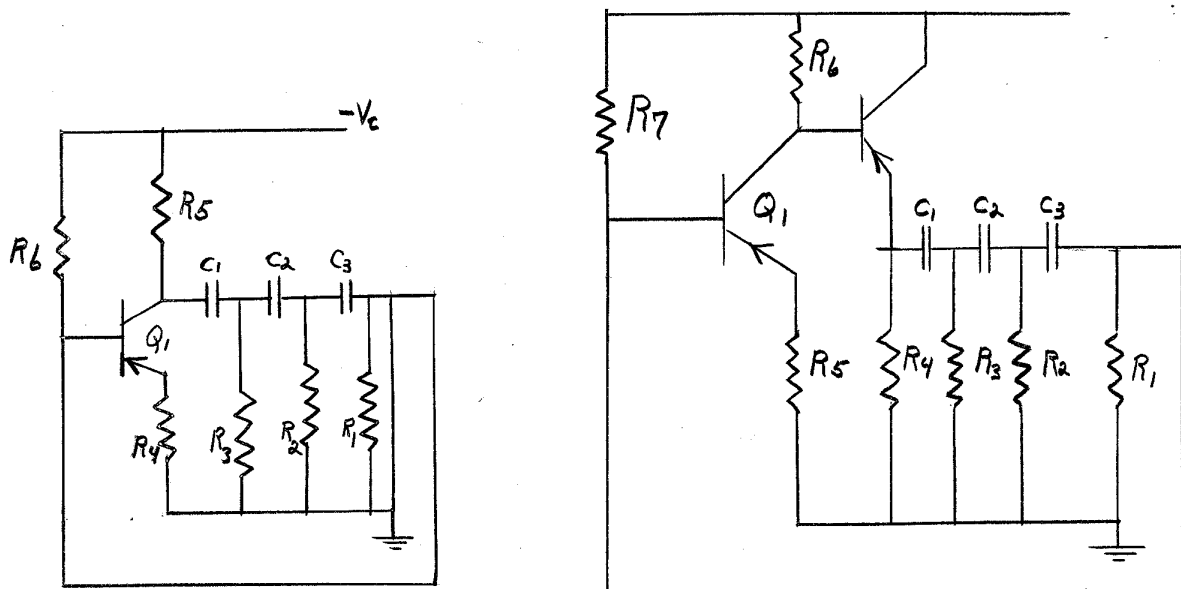Figure 6.43. High-Frequency Oscillator.

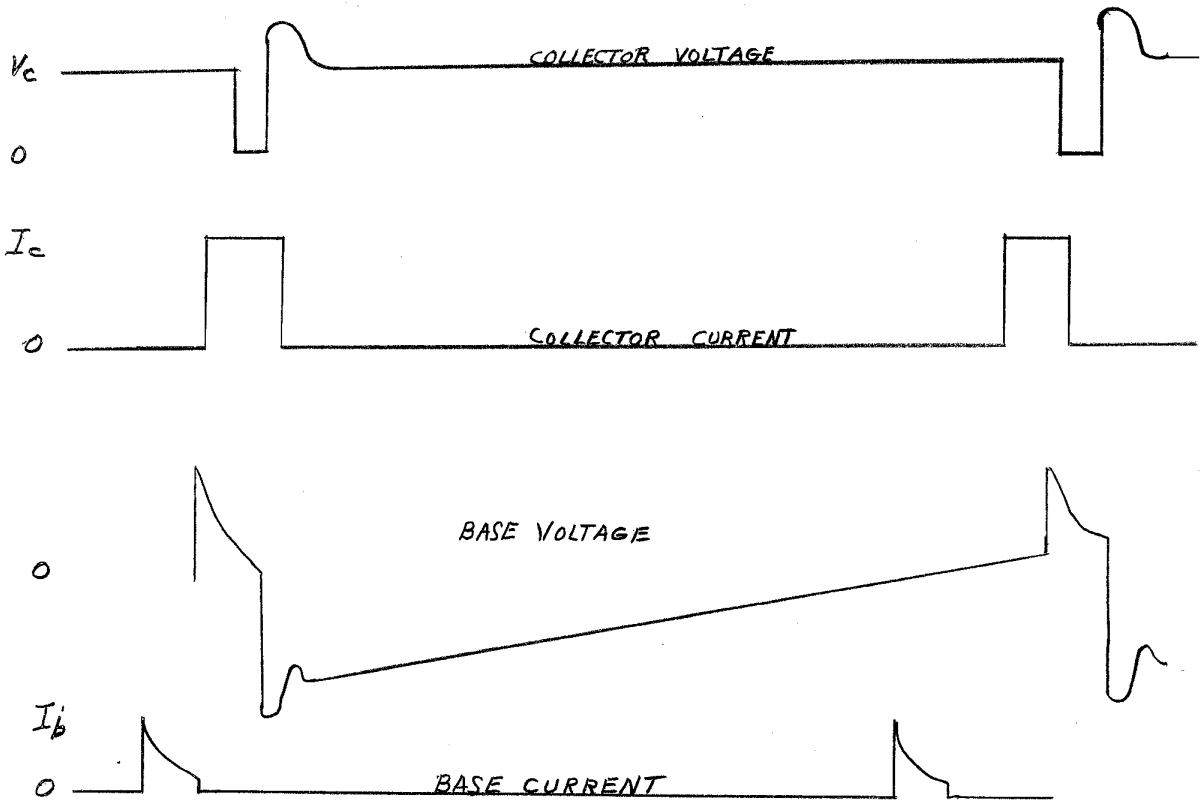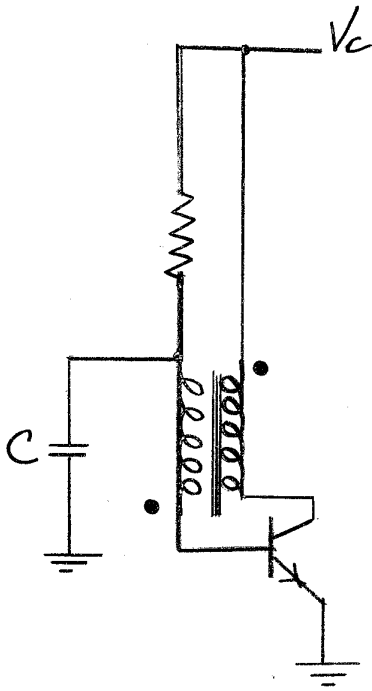Figure 6.44. Phase-Shift Oscillator Circuits.

Figure 6.45. Circuit and Waveforms of a Transistor Blocking Oscillator.

output (collector) circuit to the input (base) circuit. A r.c. combination is used to determine the basic operating frequency (repetition rate). Operation is as follows:

When power is applied, the forward bias voltage on the transistor through R will cause conduction. The increase of collector current through the transformer primary will induce a voltage in the secondary in such a direction as to increase the transistor drive. This process is cumulative so the transistor rapidly switches into saturation.

Assuming that the collector current is constant for the duration of the output pulse - although it may either increase or decrease depending on the mode of operation - the phenomenon responsible for maintaining an induced secondary voltage during the "on" portion of the blocking-oscillator operation will be investigated.

After the transistor has been switched to the "on" condition, the large base-drive current will begin to charge the capacitor. When this happens, the base current will decrease because more and more of the secondary voltage is being dropped across the capacitor leaving less available for base drive. This action does not affect the collector current because the transistor is driven far into saturation and the input circuit no longer has control of collector current. The flux setup in the transformer core due to secondary current opposes the flux established by the primary current so the decreasing secondary current causes an increase in net core flux, which maintains the secondary voltage. Hence, the transistor will conduct until the base drive voltage has dropped to the point where the input circuit regains control of the collector current. At this point the decreasing base voltage decreases the collector current which decreases the base voltage further through the action of the transformer. Again the process is cumulative and the transistor quickly switches off.

After conduction has ceased, the charge stored in the capacitor holds the transistor in the nonconducting state until this charge is drained off through the resistor, R. When this happens, the transistor will begin conducting, and the cycle will repeat.

During the "on" portion of the cycle, energy is stored in the magnetic field of the transformer. When the transistor is switched off, this energy is dissipated by the "inductive kickback" voltage as shown in the figure. If there is no damping resistance across one of the windings, this inductive voltage swing will become excessive.

One particularly effective way of reducing this voltage to a negligible value is the use of a diode across one winding. The diode is nonconducting on the useful portion of the output pulse, but it conducts on the backswing rapidly dissipating the energy of the

magnetic field in its forward resistance.

Sawtooth Generators. The blocking oscillator can be conveniently used in a sawtooth generator circuit as shown in Figure 6.46. The large emitter current flowing during the conducting portion of the blocking oscillator cycle is used to charge a capacitor. After operation of the blocking oscillator ceases, the capacitor discharges slowly through a large resistor producing a nearly-linear sawtooth. When the capacitor discharges to the point where the emitter becomes positive with respect to the base, the transistor again conducts; and the cycle repeats.

A step-up ratio in the blocking oscillator transformer of about 3 to 1 provides adequate current drive while producing enough secondary voltage to allow the capacitor to charge nearly to the supply voltage. The capacitor discharges until the emitter voltage drops to about 0.1 volt less than the fixed, base supply voltage. When this happens, the transistor will become forward biased; and the circuit will recycle. The sawtooth can be made nearly linear by adjusting the fixed base voltage so that only a small portion of the discharge is used, or by returning the discharging resistor to a positive voltage rather than to ground.

Another scheme used for linearizing the sawtooth is shown in Figure 6.46b. In this circuit the constant-current characteristics of the common-base configuration is used to provide a constant discharge current for the capacitor.

The transistor, Q_2, is forward biased by the application of a positive voltage to the emitter. The current flowing will be practically independent of transistor parameters because the resistor, R_1, is made large in comparison to the emitter-base impedance of Q_2. The resulting collector current will be equal to the emitter current times the d.c. alpha current gain of the transistor which is always very close to unity. Since the collector current of Q_2 is used to discharge the timing capacitor, this discharge will be extremely linear. The period of the output wave can be adjusted with R_2 which varies the emitter current of Q_2; and, therefore, the discharging current of the capacitor.

It is also possible to build a sawtooth generator using negative-resistance devices such as the avalanche transistor or the PNPN diode or transistor; but these schemes are generally less satisfactory because the devices mentioned are not, as a rule, capable of supplying the large capacitor discharge currents. At any rate, this will be discussed further in the section on special devices.

Free-Running Multivibrators. The output of a free running multivibrator is normally an asymmetrical rectangular wave. Such a circuit is shown in Figure 6.47. During operation, the transistors, Q_1 and Q_2, are alternately driven into conduction by a regenerative switching
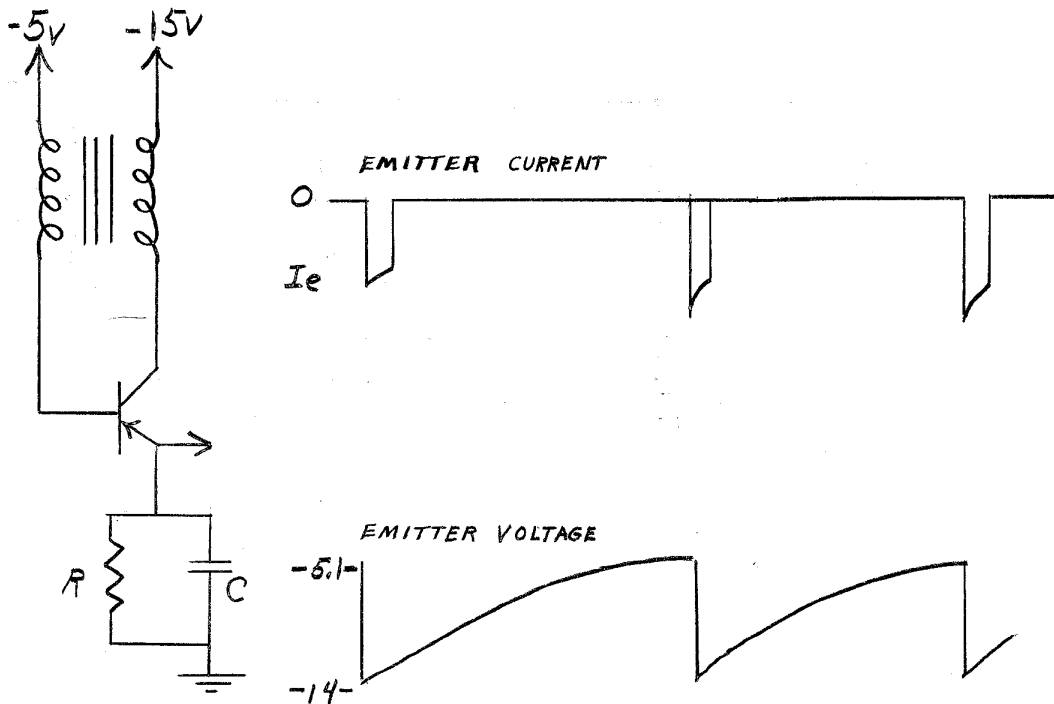
a.   Basic Sawtooth Generator.
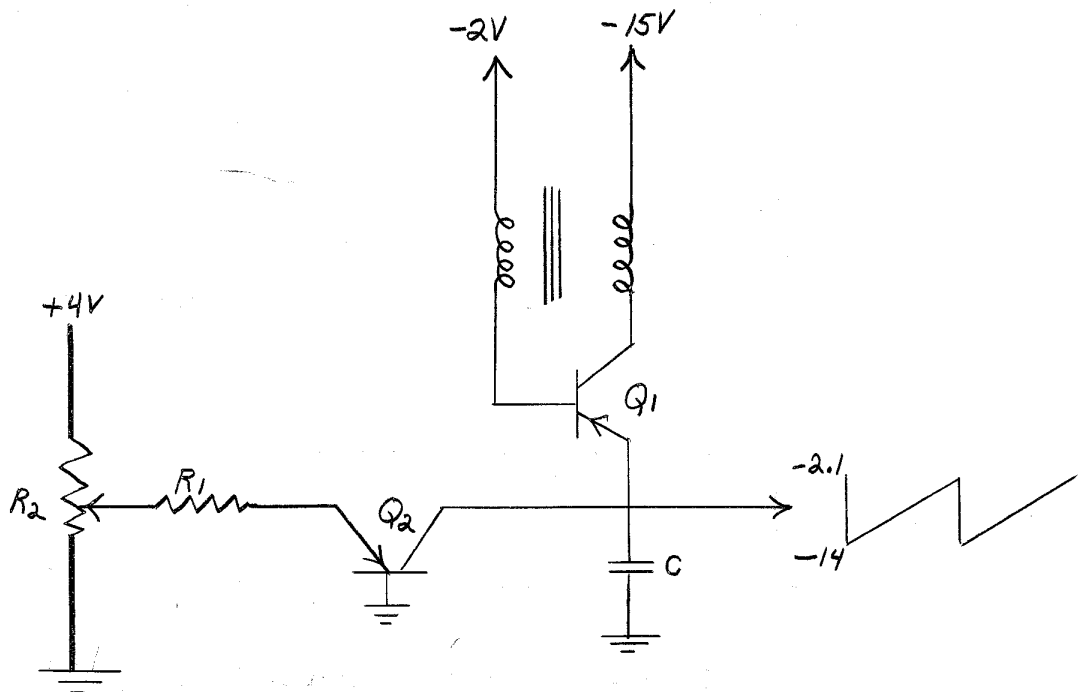
b.   Linear-Sawtooth Generator.

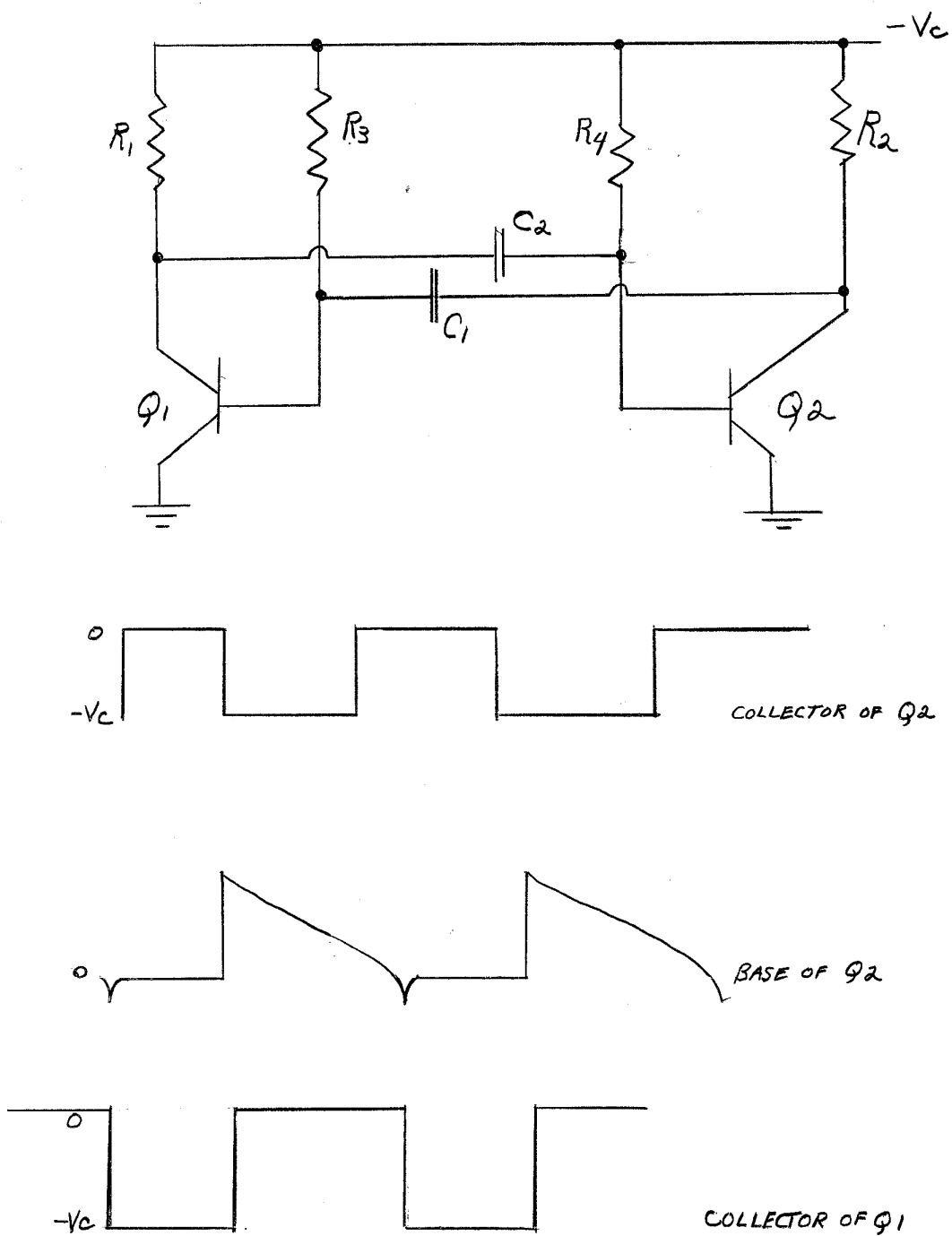Figure 6.46.   Sawtooth Generator Circuits.

Figure 6.47. Asymmetrical Multivibrator Including Waveshapes.

action. The period and asymmetry of the output wave form is controlled by adjusting the time constants of the R-C circuits in the feedback network.

Assuming that the transistor Q_1 is conducting the action is regenerative so Q_1 rapidly switches into cutoff and Q_2 switches into conduction. This condition will prevail until C_1 discharges to the negative supply voltage through R_3, and Q_1 again begins conduction.

When power is first applied to the circuit, one of the transistors will conduct harder because of a circuit dissimilarity or a random noise voltage and this transistor will switch into conduction. Therefore, it can be seen that the circuit will be self starting.

The conducting time of each transistor can be adjusted independently by controlling the time that the other transistor is cut off. This factor is solely dependent on the R-C time constant in the transistor base circuit. Therefore, even though the components have the same nominal value, the output of this oscillator is not, generally, a symmetrical square wave.

The Monostable Multivibrator. The monostable and the bistable multivibrators cannot be considered oscillators, in any strict sense, because they are not free running. Nevertheless, their similarity to the free-running variety justifies a discussion of these circuits here.

The monostable, or one-shot, multivibrator will be discussed first. This circuit is designed to produce a rectangular output pulse of known duration when excited by a trigger pulse. (The shape of the input pulse as long as its amplitude, is sufficient). Because of the many schemes available for direct-coupling transistors, there are many possible monostable multivibrator circuits. Several are given in Figure 6.48.

In the first circuit a two-stage, direct-coupled amplifier is used. In the stable condition, both transistors are conducting, and the collector voltage on Q_2 is low. A positive input pulse will reduce conduction of Q_1 which, in turn, will reduce conduction of Q_2. The collector voltage of Q_2 will then rise, and this positive-going change will be coupled through the feedback capacitor, C_f, to the base of Q_1 reducing its conduction further. Thus, both transistors are regeneratively switched off. This condition will prevail until C_f charges through R_1, and the base voltage of Q_1 becomes negative which will cause Q_1 and, therefore, Q_2 to conduct. The collector voltage on Q_2 will drop, and this negative-going drop will be coupled back to the base of Q_1 switching both transistors back into conduction until the arrival of another positive input pulse.
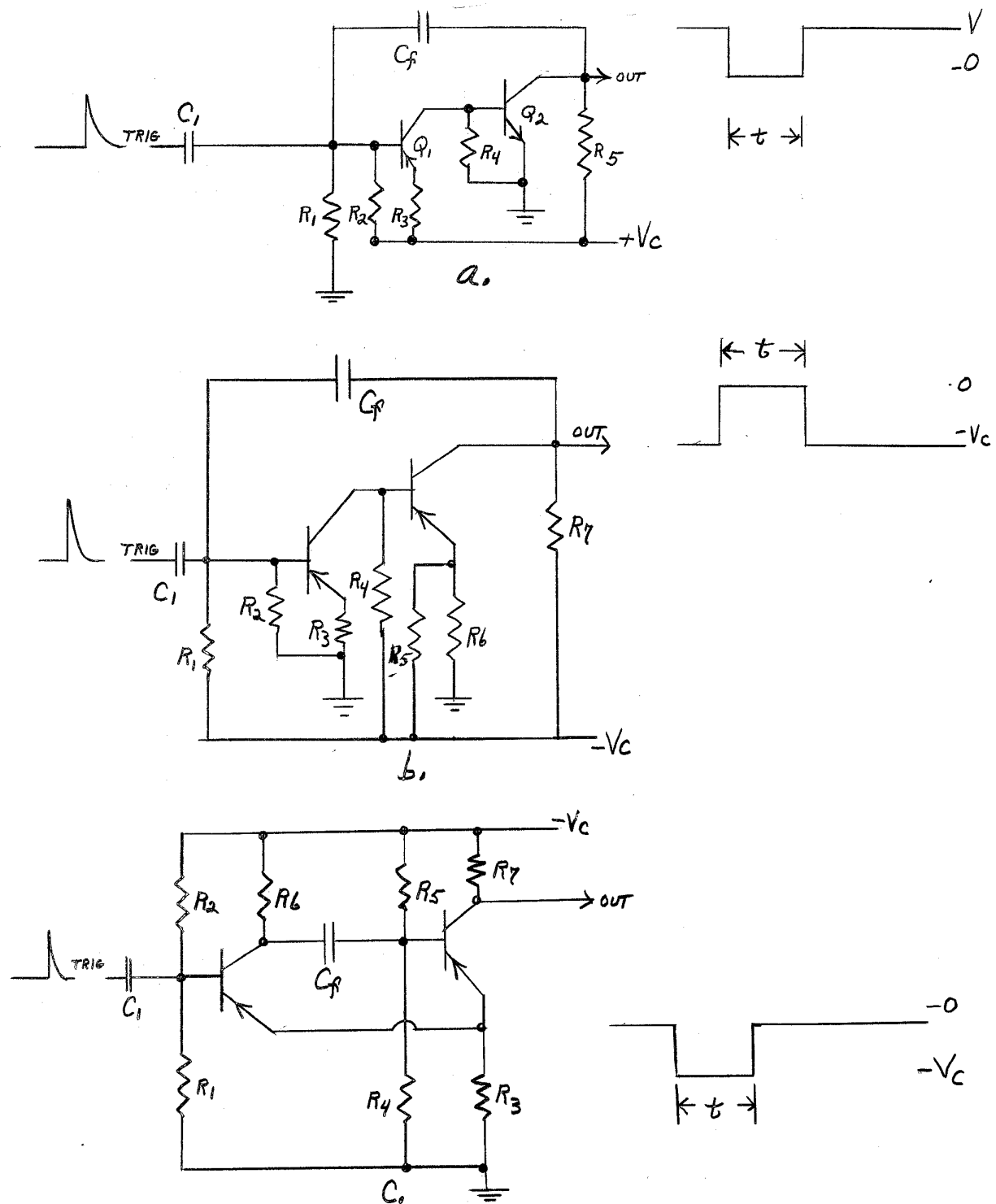
Figure 6.48. One-Shot Multivibrator Circuits.

The "on" state will only be stable if one of the transistors is in saturation. Then a definite trigger amplitude is required for operation, and triggering from random noise voltages is prevented.

The circuit in Figure 6.48b is somewhat different. In the stable state, Q_1 is conducting so that its collector voltage is less negative than the emitter voltage of Q_2. Therefore, Q_2 will be nonconducting. When a positive trigger pulse arrives, Q_1 conducts less so its collector voltage and, consequently, the base voltage of Q_2 becomes more negative. Q_2 will then begin to conduct, and its collector voltage will drop. This positive going change will be felt back on the base of Q_1 through C_f, re-enforcing the input signal and cutting off Q_1. Q_2 then goes into full conduction. The circuit will remain in this unstable state until C_f discharges through R_1 and R_2. The multivibrator will then switch back to the stable state, and wait for another positive input pulse.

The circuit shown in Figure 6.48c is another variation. The bias resistors are adjusted so that Q_2 is conducting and Q_1 is nonconducting. When a trigger is applied, Q_1 will conduct; and a signal will be coupled through C_f cutting off Q_2. The feedback loop is completed by the resistor, R_3, which is common to the emitters of both transistors. When Q_2 cuts off, the voltage across R_3 will drop allowing Q_1 to switch into the conducting state. When C_f charges, the circuit will recycle.

The Bistable Multivibrator. The bistable multivibrator, or "flip flop", will remain in either of two stable states until excited by a trigger pulse. There is no internal timing in this circuit, and output is entirely dependent on the input pulse rate. The schematic of a bistable multivibrator is given in Figure 6.49. It will be shown that the two stable states are Q_1 conducting with Q_2 cut off and Q_1 cut off. Moreover, it will be seen that the multivibrator can be switched between these states with a trigger pulse properly applied to the device.

If Q_1 is conducting, the emitters of both transistors will be at some negative voltage because of the drop in resistor, R_7. Furthermore, the collector of Q_1 will be at nearly the same potential as its emitter. The base of Q_2 will then be at some potential less negative than its emitter because it is being fed through a voltage divider from the collector of Q_1. Therefore, Q_2 will be nonconducting and the collector potential of Q_2 will be nearly equal to the supply voltage. It is this high collector voltage on Q_2 that supplies the conducting bias to Q_1 through the voltage divider. The circuit will remain in this state indefinitely because the charging or discharging of capacitors does not determine the bias levels of the transistors.

This situation can be disturbed by impressing a trigger pulse on the circuit in such a way as to cut off Q_1 or drive Q_2 into
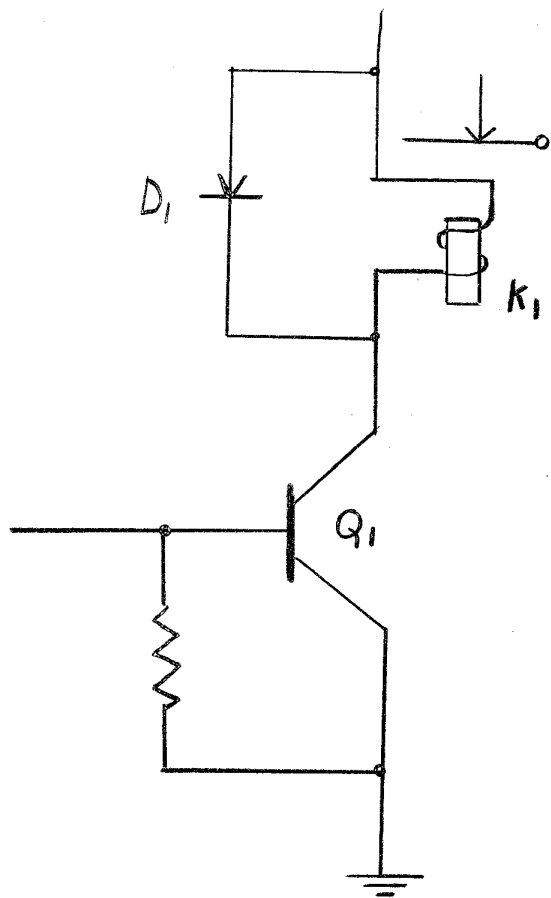
Figure 6.48. Illustrating Use of Protective Diode with
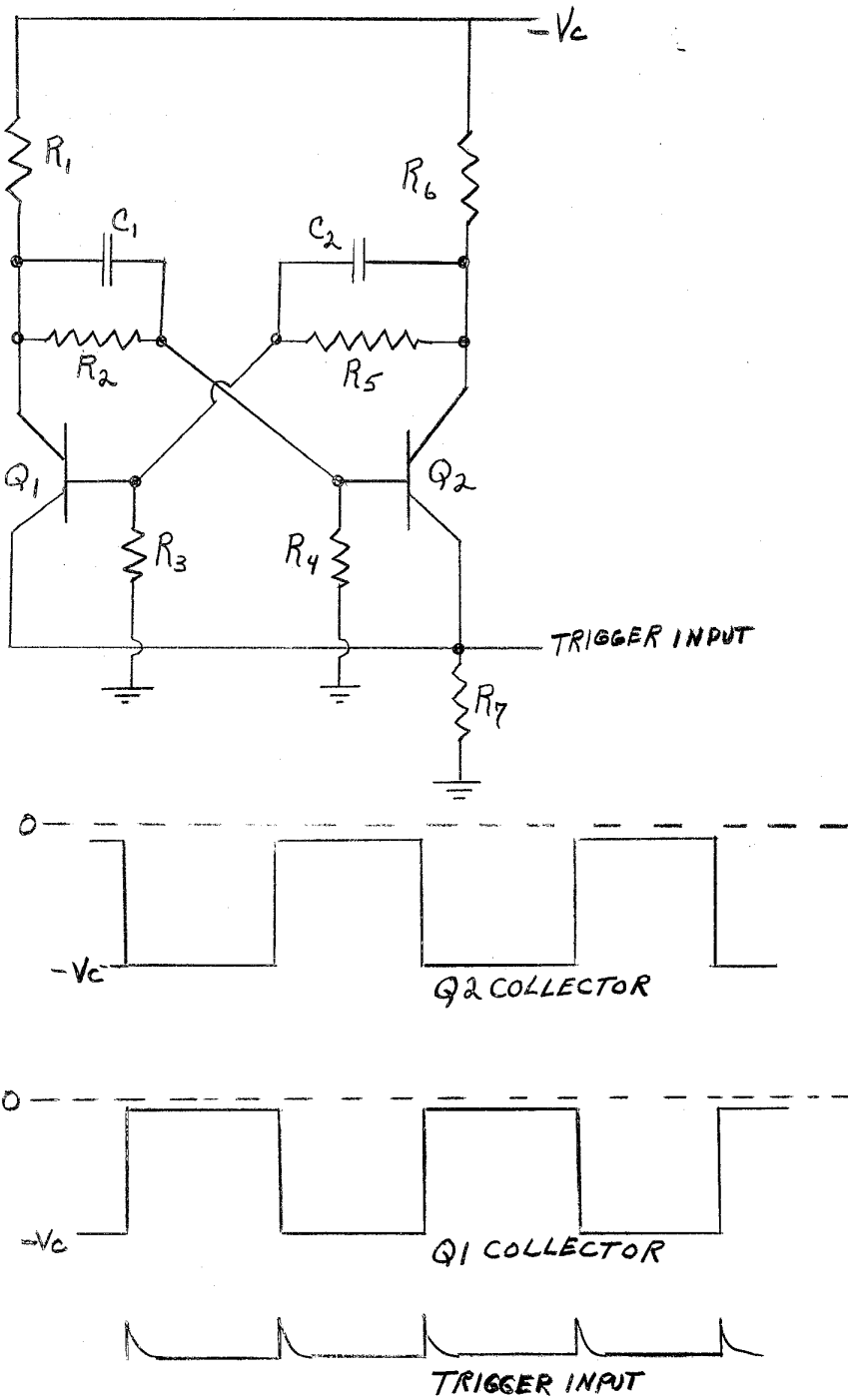Inductive Loads.

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Figure 6.49. Circuit of a Bistable Multivibrator
Including Waveshapes.

conduction. One such trigger arrangement is shown in Figure 6.49. Positive trigger pulses are fed in across the common emitter resistor. These positive pulses will not affect Q_1 because it is already conducting. However, the emitter of Q_2 will become sufficiently positive so that it conducts. When this happens, the collector voltage of Q_2 will drop, removing the forward bias on Q_1; and the multivibrator will switch states.

It can be shown in a similar manner that the next positive pulse will switch the "flip flop" back to its original state.

A negative trigger can also be used because it will cut off the conducting transistor causing its collector voltage to rise and drive the other transistor into conduction.

The switching time of the bistable multivibrator is an important consideration as the repetition frequency of the trigger pulses increases. In order to reduce switching time, "speed up" capacitors (C_1 and C_2) are shunted across the base voltage dividers. These capacitors are small so they couple the rapid, switching signal without attenuation while not disturbing the quiescent, d.c. level.

Although the circuit in Figure 6.49 is typical, many variations are possible. For example, the trigger signal may be applied to the transistor base through triggering diodes, or to the collectors in a similar manner.

Moreover, for improved high-frequency performance, special circuits may be employed to prevent saturation of the transistors. This latter detail will be covered in the following section.

Transistor Switches. The transistor makes an efficient switch. When in the conducting state, it presents a very small resistance and can pass large currents with a voltage drop of a few tenths of a volt. In the nonconducting state, it presents a high resistance, and only a small leakage current flows.

When the transistor is saturated, the low resistance of the collector circuit is referred to as the collector saturation resistance. Its value is given by the slope of the collector saturation curves. For the transistor described in Figure 6.46a, this value is in the order of 0.1 ohm. Using an expanded scale for the saturation region shows that the collector-current curves do not merge in the saturation region (Figure 6.46b). This expanded scale more accurately describes the saturation characteristics and is frequently useful in switching circuits where large currents must be passed.
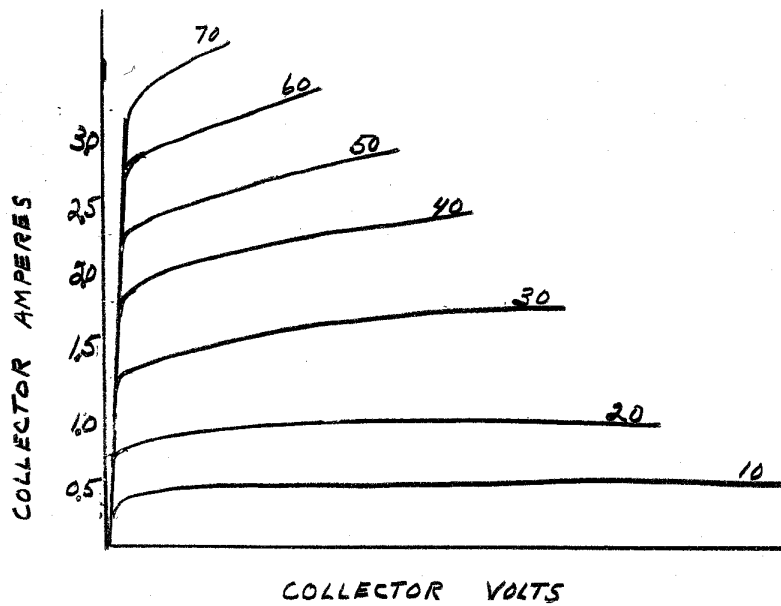
a.
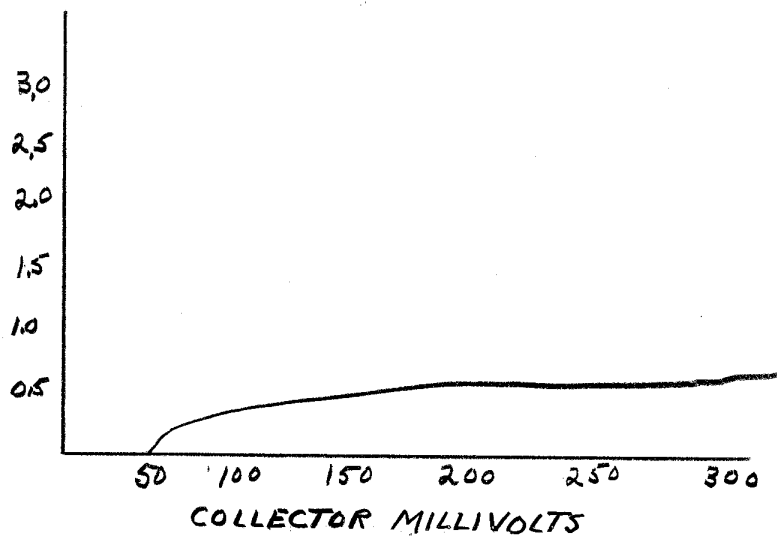
b.

Figure 6.46. Transistor Characteristic Curves Including an
Expanded Scale of the Collector Saturation Region.

The "off" current of a transistor can be divided into two components. The first is the result of thermally-generated carriers within the semiconductor. This current is the reverse saturation current of the collector junction. This current will not depend on collector voltage, except at very low voltages when the reverse junction potential is not high enough to sweep across all the carriers or at high reverse potentials where carrier multiplication will occur. The second current component is surface leakage current. This current arises from resistive leakage across the junction due to the presence of contaminating agents on the semiconductor surface. This component, being resistive in nature, does depend on collector voltage. These two components produce the characteristic shown in Figure 6.47.

The variation of collector junction saturation current with temperature can easily be calculated; however, the surface leakage component cannot. Hence, no fixed relationship between cutoff current and temperature can be established; and manufacturers data must be consulted. Nonetheless, in the absence of more accurate data, an approximation can be used: the collector cutoff current will double for every $10^\circ C$ rise in temperature.

The power dissipation of a transistor in the "on" or "off" state is small. In one case the collector current can be quite large but the collector voltage is small. In the other case the collector voltage is large but the current small. In switching between those states the dissipation will increase; and, even though the time involved is small, this can result in damage to the transistor if it does not have the necessary peak power capacity. This is particularly true when inductive loads are switched off or capacitive loads are switched on. With inductive loads, such as a relay, a diode can be used to absorb the energy in the collapsing magnetic field, thus protecting the transistor from voltage surges and sustained current during switching. This is illustrated in Figure 6.48.

High-Speed Switching. Because of the high frequency limitations of transistors, it is only reasonable to expect that some restriction will be imposed on switching speed. Referring now to Figure 6.49, when an input signal is applied to a transistor, base current flows immediately; but the corresponding increase of collector current is delayed because of the carrier transit time across the base region. After the collector current does begin to rise, some finite time will be required for it to reach its final value both because of the irregular diffusion rate of current carriers crossing the base and because of the collector junction capacitance.

The delay time (t_d) is shown to be the time between the application of the input signal and the point where the collector voltage reaches 10% of its final value. The rise time is the time taken for the collector voltage to go from 10% to 90% of its final value.
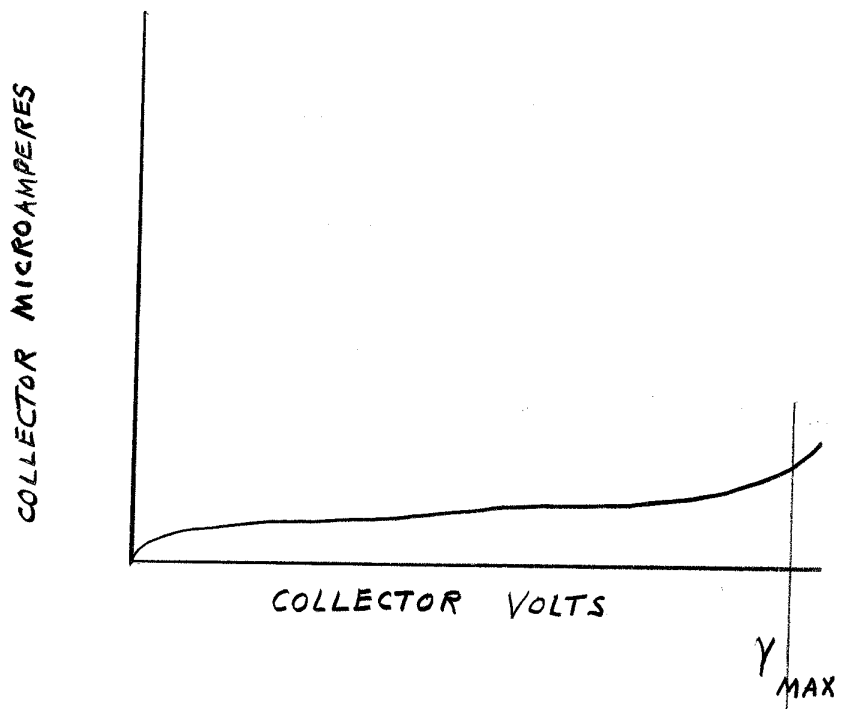
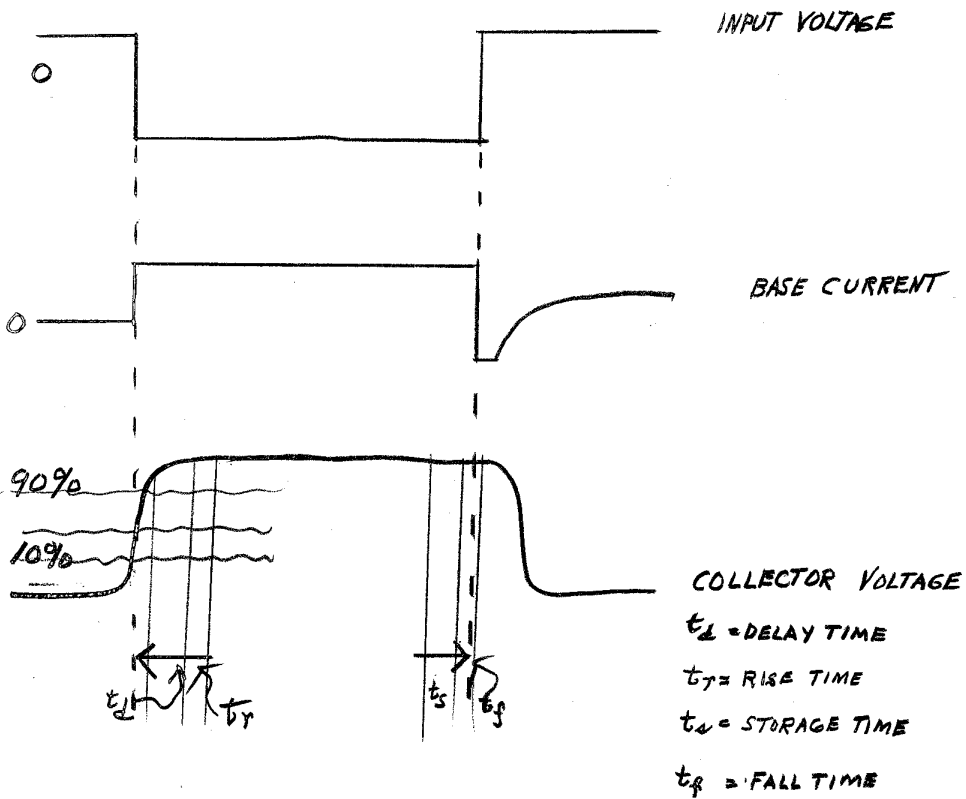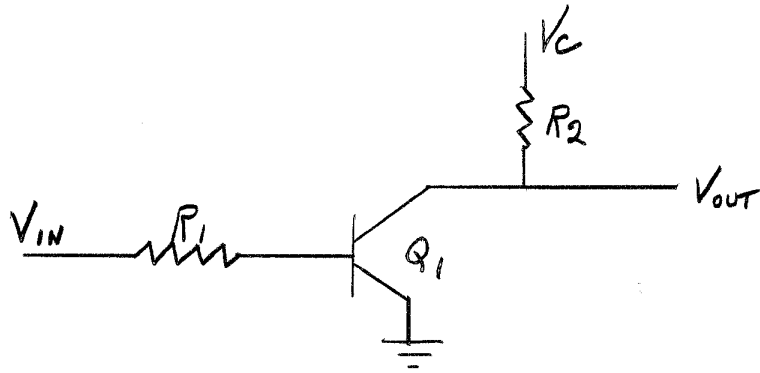Figure 6.47. Variation of Collector-Cutoff Current with Collector Voltage.

Figure 6.49. Description of Transistor Switching Time.

In this case it is assumed that the transistor is driven into saturation; that is, the input signal is of such a magnitude that the collector voltage drops below the base voltage; and both the emitter and collector junctions become forward biased. When this happens, both the emitter and collector inject minority carriers into the base region. The base will then become charged with excess current carriers as shown in Figure 6.50.
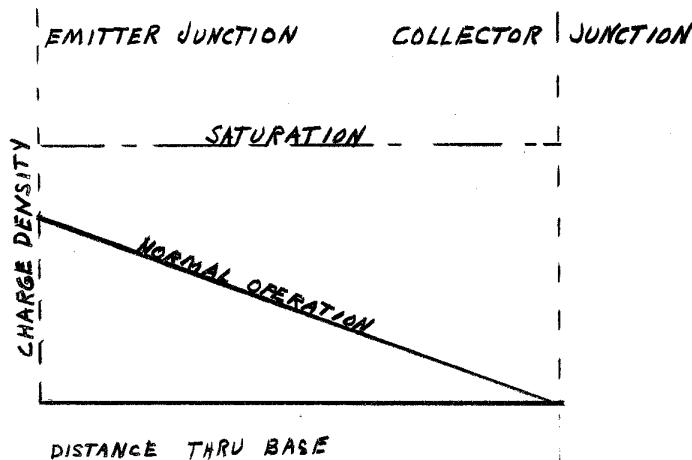

Figure 6.50. Injected Carrier Density in Base Region for Normal and Saturated Operation.

When the turnoff signal arrives, the excess charge must be drained from the base before the collector current can fall which produces a delay in the output waveform. This delay is defined as the storage time (t_s). Because the base is not abruptly cleared of charge and also because of the collector junction capacitance, an additional increment of time (t_f) is required for the transistor to switch off. The fall time is defined in Figure 6.49.

In high-speed switching circuits, the switching time can become appreciable compared to the pulse duration. It is therefore desirable to keep these time lags at a minimum. One method suggests itself almost immediately: if the collector is kept from going into saturation, it is obvious that the storage time can be reduced. The most direct method of accomplishing this is to use a diode clamp to prevent the collector voltage from dropping below a particular value. Such a circuit is illustrated in Figure 6.51a. This circuit suffers from the disadvantage that the collector current in the "on" state can vary over a wide range. When the collector voltage is held at some

value by a clamp circuit, the collector current will depend on the base drive current and the alpha current gain. The maximum possible variation of base current drive will depend on circuit design, but the current gain will vary considerably with aging and temperature. A collector current variation of ten to one would not be unusual under these conditions.

A circuit that does not suffer from this limitation is shown in Figure 6.51b. A feedback circuit is used. When the collector voltage drops below some design value, the diode will conduct and prevent any further increase of base current by limiting the voltage at the junction of R_1 and R_2. Hence, when the transistor is in the "on" condition, the collector will be at a higher voltage than the base by the amount of the voltage drop across R_2; and it is impossible for it to go into saturation.

In all the circuits mentioned thus far, the switching time can be reduced by shunting a capacitor across the series, base resistor. This capacitor will provide a large base drive to charge and discharge the base region during switching, but it will not affect the steady-state current values.

Another technique is available for reducing the rise and fall times but at the expense of output amplitude. If the collector is clamped to prevent the collector voltage from reaching the supply voltage when the transistor cuts off, the switching time will be reduced. This is illustrated in Figure 6.52. With the clamp circuit, the collector current will already have started switching by the time the collector voltage begins to change so the slow initial rise does not appear in the voltage waveform. A similar situation is present when the transistor cuts off. The final portion of the charge curve for the collector junction capacitance is eliminated by the clamp circuit thus reducing the fall time.

Figure 6.52 illustrates a typical high speed transistor switch, where D_1 prevents collector saturation, D_2 provides the cutoff clamp of 5 volts below the collector supply voltage, and C_1 gives a switching transient to quickly charge or discharge the base region. The improvement in pulse shape (particularly the reduction of storage time) is evident from the accompanying waveforms.
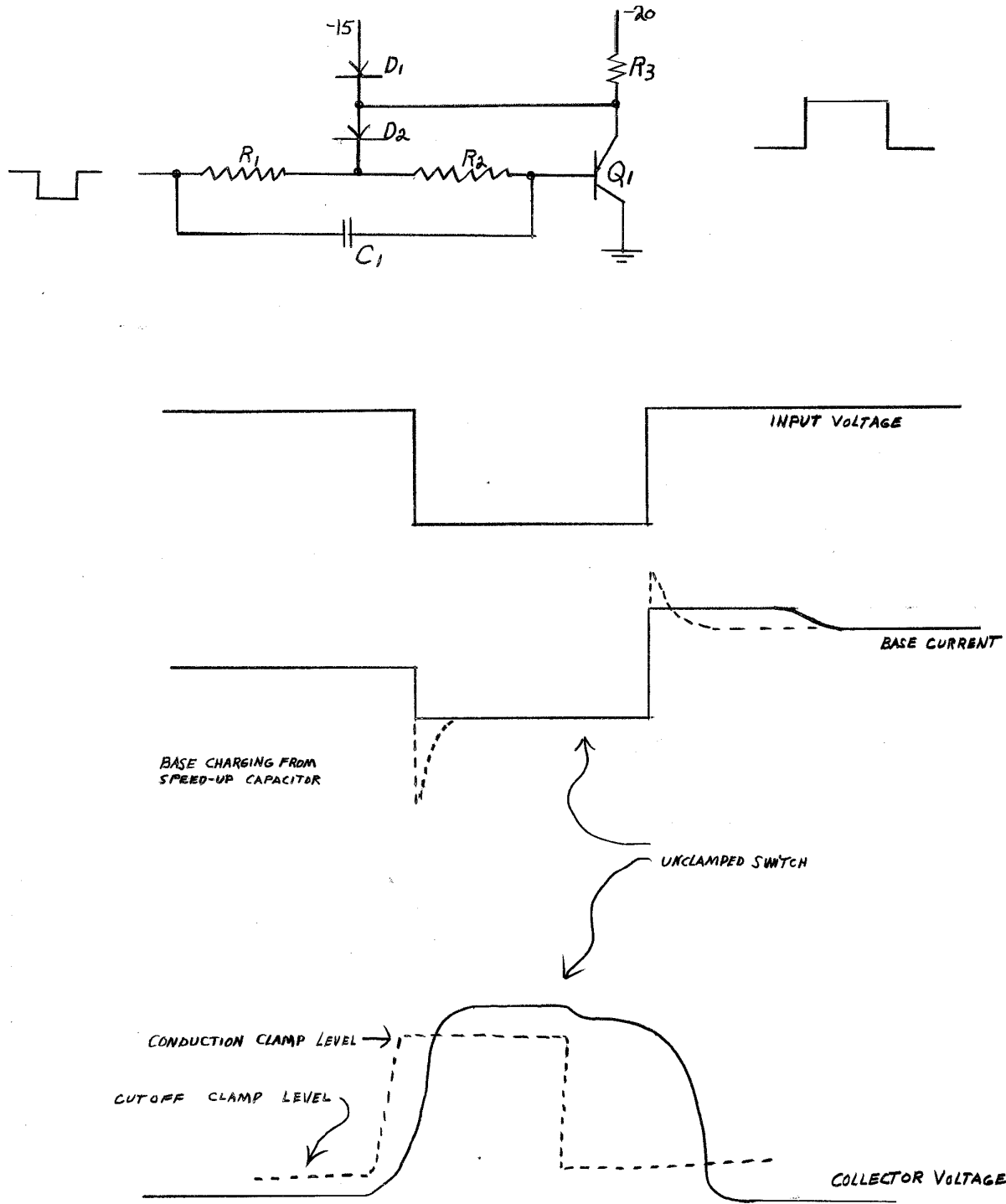
Figure 6.52. High Speed Switching Circuit.