SECTION 3
GENERAL INFORMATION ON SEMICONDUCTOR CIRCUITS

3.1 DEFINITIONS OF LETTER SYMBOLS USED.

The letter symbols used in the diagrams and discussions on semiconductor circuits throughout this technical manual are those proposed as standard for use in industry by the Institute of Radio Engineers, or are special symbols not included in the standard. Since some of these symbols change from time to time, and new symbols are developed to cover new devices as the art changes, an alphabetical listing of the symbols used herein is presented below. It is recommended that this listing be used to obtain the proper definitions of the symbols employed in this manual, rather than to assume an erroneous meaning.

3.1.1 **Construction of Symbols.** Semiconductor symbols are made up of a basic letter with subscripts, either alphabetical or numberical, or both, in accordance with the following rules:

a. A capital (upper case) letter designates external circuit parameters and components, large-signal device parameters, and maximum (peak), average (dc), or root-mean-square values of current, voltage, and power (I, V, P, etc.)

b. Instantaneous values of current, voltage, and power, which vary with time, and small-signal values are represented by the lower case (*small*) letter of the proper symbol (i, v, p, i_e, v_{eb}, etc.)

c. D-C values, instantaneous total values, and large-signal values, are indicated by upper case subscripts (i_C, I_C, v_{EB}, V_{EB}, P_C, p_C, etc.)

d. Alternating component values are indicated by using lower case subscripts; for example, i_c, I_c, v_{eb}, V_{eb}, p_c, P_c.

e. When it is necessary to distinguish between maximum, average, or root-mean-square values, maximum or average values may be represented by addition of a subscript m or av; for example, i_{cm}, I_{cm}, I_{CM}, I_{cav}, i_{CAV}.

f. For electrical quantities, the first subscript designates the electrode at which the measurement is made.

g. For device parameters, the first subscript designates the element of the four-pole matrix; for example, I or i for input, O or o for output, F or f for forward transfer, and R or r for reverse transfer.

h. The second subscript normally designates the reference electrode.

i. Supply voltages are indicated by repeating the associated device electrode subscript, in which case, the reference terminal is then designated by the third subscript; for example, V_{EE}, V_{CC}, V_{EEB}, V_{CCB}.

j. In devices having more than one terminal of the same type (say two bases), the terminal subscripts are modified by adding a number following the subscript and placed on the same line; for example V_{B1-B2}.

k. In multiple-unit devices the terminal subscripts are modified by a number *preceding* the electrode subscript; for example, V_{1B-2B}.

3.1.2 **Alphabetical List of Semiconductor Letter Symbols.** Since the rules above are somewhat complex, an alphabetical list of all letter symbols used herein is presented below for easy reference.

Symbol	**Definition**
α	Current amplification factor (common-base current gain - Alpha)
α_{FB}, α_{FC}, α_{FE}	Short-circuit forward current transfer ratio, static value
α_{fb}, α_{fc}, α_{fe}	Small-signal short-circuit forward current transfer ratio
A_G	Available gain
A_I	Current gain
A_p	Power Gain
A_V	Voltage gain
B or b	Base electrode
β	Common-emitter current gain - Beta
BV or V_{BR}	Breakdown voltage (formerly P_{IE} or P_{IV})
BZ	Large-signal breakdown impedance
bz	*Small-signal* breakdown impedance
C or c	Collector electrode or capacitor
CB, CC, CE	Common base, collector, and emitter, respectively
C_c	Collector junction capacitance
C_{cc}	Coupling capacitor
C_e	Emitter junction capacitance
CG	Current gain
γ	Collector current gain —— Gamma
CG_o	Over-all current gain
C (dep)	Depletion layer capacitance
C (diff)	*Diffusion layer capacitance*
C_{ib}, C_{ic}, C_{ie}	Input capacitance for common base, collector, and emitter, respectively
C_{ij}	Input capacitance
C_{ibs}, C_{ies}, C_{ics}	Input terminal capacitance with output terminals shortcircuited to ac, for common base, emitter, and collector
C_L	Load capacitance
C_{ob}, C_{oc}, C_{oe}	Output terminal capacitance for common base, collector, and emitter, respectively
C_{obo}, C_{oco}, C_{oeo}	Output terminal capacitance, a-c input terminals open-circuited, for common base, collector, and emitter, respectively
D	Distortion
E or e	Emitter electrode
E_D	Drain-terminal supply voltage, unipolar transistor
E_G	Gate-terminal supply voltage, unipolar transistor
E_B, E_C, E_E	Same as V_{BB}, V_{CC}, V_{EE}
e_b, e_c, e_e	Same as v_b, v_c, v_e
E_{bb}	Battery supply voltage
E_i or E_1	Input voltage, 4-terminal network
E_o or E_2	Output voltage, 4-terminal network
EF	Emitter follower, computer logic circuit
f or 21	Forward transfer

Symbol	Definition
f_{ab}, f_{ac}, f_{ae}	Alpha cutoff frequency for common base, collector, and emitter, respectively
f_c	Cutoff frequency
f_{Gp}	Power conductance cutoff frequency
f_{max}	Maximum frequency of oscillation
F_n	Noise figure
f_o	Theoretical cutoff frequency, or zero (basic) frequency
f_{osc}	Maximum frequency of oscillation
f_{pg}	Power gain cutoff frequency
f_r	Resonant frequency
f_T	Transition frequency
f_1	Frequency of unity current transfer ratio
G_b, G_c, G_e	Power gain for common base, collector, and emitter, respectively
g_{MJ}	Static transconductance
g_m	Intrinsic transconductance
g_{mj}	Small-signal transconductance
G_{MJ}	Large-signal transconductance
h	Hybrid parameter
h_{FB}, h_{FC}, h_{FE}	Short-circuit forward-current transfer ratio, static value for common base, collector, and emitter, respectively
$h_{fb}, h_{fc}, h_{fe}, h21$	Short-circuit forward-current transfer ratio, small-signal value, for common base, collector, and emitter respectively
h_{IB}, h_{IC}, h_{IE}	Short-circuit input resistance, static value for common base, collector, and emitter, respectively
$h_{ib}, h_{ic}, h_{ie}, h11$	Short-circuit input impedance, small-signal value
h_{OB}, h_{OC}, h_{OE}	Open-circuit output conductance, static value
$h_{ob}, h_{oc}, h_{oe}, h22$	Open-circuit output admittance, small-signal value
h_{RB}, h_{RC}, h_{RE}	Open-circuit reverse-voltage transfer ratio, static value
$h_{rb}, h_{rc}, h_{re}, h12$	Open-circuit reverse-voltage transfer ratio, small-signal value
I	Direct current (dc)
i	Alternating current (ac)
I_B, I_C, I_E	D-C base, collector, and emitter current, respectively
I_b, I_c, I_e	RMS value of a-c signal current for base, collector, and emitter, respectively
i_b, i_c, i_e	Instantaneous value of a-c base, collector, and emitter current, respectively
I_{BM}, I_{CM}, I_{EM}	Maximum value of total base, collector, and emitter current, respectively
I_{bm}, I_{cm}, I_{em}	Maximum a-c component value of base, collector, and emitter current, respectively
$I_b(av), I_c(av), I_e(av)$	Average (d-c) value of alternating component of base, collector, and emitter current, respectively
$I_B(AV), I_C(AV), I_E(AV)$	Average (d-c) value of total base, collector, or emitter current, respectively with signal applied
$I_{(BR)}$	Total current at breakdown voltage; use additional subscripts to identify electrodes measured and conditions
I_{BCO} or I_{BO}	D-C base current, base reverse-biased with respect to collector, emitter to collector open
I_{BCS} or I_{BS}	D-C base current, base reverse-biased with respect to collector, emitter shorted to collector
I_{BEO}	D-C base current, base reverse-biased with respect to emitter, collector to emitter open
I_{BES}	D-C base current, base reverse-biased with respect to emitter, collector shorted to emitter
I_{CBO} or I_{CO}	D-C collector current, collector reverse-biased with respect to base, emitter to base open
I_{CBS} or I_{CS}	D-C collector current, collector reverse-biased with respect to base, emitter shorted to base
I_{CEO}	D-C collector current, collector reverse-biased with respect to emitter, base to emitter open
I_{CES}	D-C collector current, collector reverse-biased with respect to emitter, base shorted to emitter
I_D	D-C drain current, unipolar transistor
I_{EBO} or I_{EO}	D-C emitter current, emitter reverse-biased with respect to base, collector to base open
I_{EBS} or I_{ES}	D-C emitter current, emitter reverse-biased with respect to base, collector shorted to base
I_{ECO}	D-C emitter current, emitter reverse-biased with respect to collector, collector to base open
I_{ECS}	D-C emitter current, emitter reverse-biased with respect to collector, base to collector open
I_F	Diode d-c forward current
i_F	Instantaneous forward diode current
I_G	D-C gate current, unipolar transistor
I_{GT}	Gate trigger current in a PNPN type switch
I_H	Holding current in a PNPN type switch
I_O	D-C output current
I_I	D-C input current
I_n	Electron current
I_{nc}, I_{ne}	Electron current through collector and emitter junction, respectively
I_p	Hole current
I_1	Input current, 4-terminal network
I_2	Output current, 4-terminal network

Symbol	Definition
i_{b1}	Base input turn-on current (switching)
i_{b2}	Base input turn-off current (switching)
i_i	A-C input current, instantaneous
i_R	Instantaneous diode reverse current
I_R	Diode d-c reverse current
I_s	Saturation current
J or j	Electrode, general
MAG	Maximum available gain
MIN, min	Minimum value
NPN	Transistor consisting of one P-type and two N-type semiconductor junctions
N-type	Semiconductor with donor impurity
o or 22	Output (used as a subscript)
P	Total average power dissipation of all electrodes of a semiconductor device
P_b, P_c, P_e	Average power dissipation of base, collector, and emitter, respectively
PG	Power gain
PG_o	Over-all power gain
P_i or P_1	Input power
P_o or P_2	Output power
P_{om}	Maximum output power
P_t	*Point contact*
P_{bm}, P_{cm}, P_{em}	Peak power dissipation of base, collector, and emitter, respectively
PN	Combination of P-type and N-type semiconductors
PNP	Transistor consisting of one N-type and two P-type semiconductor junctions
P-type	Semiconductor with acceptor impurity
R	Resistance, resistor
R_B, R_C, R_E	External series resistance for base, collector, and *emitter, respectively*
R_b, R_c, R_e	Internal resistance of base, collector, and emitter, respectively
r_b, r_c, r_e	A-C resistance of base, collector, and emitter, respectively, for low-frequency T-equivalent circuit
R_f	Feedback resistance
R_i	Input resistance
r_i	4-terminal network input resistance
R_{im}	Matched input resistance
r_m	A-C transfer resistance for T-equivalent circuit
R_L	Load resistance
R_o	Output resistance
r_o	4-terminal network output resistance
R_{om}	Matched output resistance
r'_b	Equivalent base high-frequency resistanc
r_{11}	Small signal, open-circuit input resistance
r_{12}	Small-signal, open-circuit, reverse transfer resistance
r_{21}	Small-signal, open-circuit, forward transfer resistance
r_{22}	Small-signal, open-circuit output resistance
S_I	Current stability factor
S_i	Silicon high-temperature transistor
S_V	Voltage stability factor
T	Absolute temperature, or transformer
T_A	Ambient temperature
t_c	Time constant
T_C	Case temperature
TCBV	*Temperature coefficient* of breakdown voltage
T_J	Junction temperature
t_d	Ohmic delay time (switching)
t_f	Pulse fall time, 90% to 10% of pulse (switching)
T_{max}	Absolute maximum temperature
T_{stg}	Storage temperature
t_{fr}	Diode forward recovery time
t_p	Time of pulse
t_r	Pulse rise time, 10% to 90% of pulse (switching)
t_{rr}	Diode reverse recovery time
t_s	Storage time from turnoff pulse to 90% decay time
t_w	Pulse average time
μ_{FB}, μ_{FC}, μ_{FE}	Open-circuit forward voltage transfer ratio, static value for common base, collector, and emitter, respectively
μ_{fb}, μ_{fc}, μ_{fe}	Small-signal open-circuit forward voltage transfer ratio for common base, collector and emitter, respectively
μ_{RB}, μ_{RC}, μ_{RE}	Open-circuit reverse voltage transfer ratio, static value for common base, collector, and emitter, respectively
μ_{rb}, μ_{rc}, μ_{re}	Small-signal open-circuit reverse voltage transfer ratio for common base, collector, and emitter, respectively
V	*D-C voltage*
v	A-C voltage
V_B, V_C, V_E	D-C voltage for base, collector, and emitter, respectively
v_b, v_c, v_e	A-C voltage for base, collector, and emitter, respectively
V_{BB}	D-C base supply voltage
V_{BBE}	Base-to-emitter d-c supply voltage
V_{BC}	Base-to-collector d-c voltage
v_{bc}	Base-to-collector a-c voltage
V_{BE}	Base-to-emitter d-c voltage
v_{be}	*Base-to-emitter a-c voltage*
V_{BEF}	D-C base floating potential, collector reverse-biased with respect to emitter, base to emitter open
V_{BF} or V_{BCF}	D-C base floating potential, emitter reverse-biased with respect to collector, base open
V_{BR}	Breakdown voltage
V_{CB}	Collector-to-base d-c voltage
v_{cb}	Collector-to-base a-c voltage

Symbol	Definition
V_{CBF} or V_{CF}	D-C collector floating potential, emitter reverse-biased with respect to base, collector to base open
V_{CC}	Collector d-c supply voltage
V_{CCB}	Collector-to-base d-c supply voltage
V_{CCE}	Collector-to-emitter d-c supply voltage
V_{CE}	Collector-to-emitter d-c voltage
v_{ce}	Collector-to-emitter a-c voltage
V_{CEF}	Collector d-c floating potential, base reverse-biased with respect to emitter, collector to emitter open
V_{CEO}	D-C collector-to-emitter voltage with collector junction reverse-biased, zero base current (specify I_c)
V_{CER}	Similar to V_{CEO}, except with a resistor (of value R) between base and emitter
V_{CES}	Similar to V_{ceo}, except with base shorted to emitter
V_{CE} (SAT)	Collector region saturation voltage at specified I_C and I_B
V_D	D-C drain voltage, unipolar transistor
V_{EB}	Emitter-to-base d-c voltage
v_{eb}	Emitter-to-base a-c voltage
V_{EBF} or V_{EF}	Emitter d-c floating potential collector reverse-biased with respect to base, emitter to base open
V_{EC}	Emitter-to-collector d-c voltage
v_{ec}	Emitter-to-collector a-c voltage
V_{ECF}	Emitter d-c floating potential, base reverse-biased with respect to collector, emitter to collector open
V_{EE}	D-C emitter supply voltage
V_F	Diode d-c forward voltage
v_F	Diode instantaneous (a-c) forward voltage
V_G	D-C gate voltage, unipolar transistor
VG	Voltage gain
VG_o	Over-all voltage gain
V_I	D-C input voltage, 4-terminal network
V_i	A-C input voltage, 4-terminal network
v_i	A-C input voltage
v_n	Noise voltage
V_o	A-C output voltage, 4-terminal network
v_o	A-C output voltage
V_O	Pinch-off voltage, unipolar transistor
V_{oc}	Open-circuit voltage
V_R	Diode d-c reverse voltage
v_R	Diode instantaneous (a-c) reverse voltage (peak inverse)
V_{RT}	Reach-through (punch-through) voltage
V_s	Source voltage, general
V_{SAT}	Saturation voltage, general
V_T	Thermal noise voltage
W	Transistor base region thickness
X	Reactance, general
X_c	Capacitive reactance
X_L	Inductive reactance
y	Short-circuit admittance parameter
y_{fb}, y_{fc}, y_{fe}	Small-signal forward-transfer admittance, a-c output shorted, for common base, collector, and emitter, respectively
y_{ib}, y_{ic}, y_{ie}	Small-signal input admittance, a-c output shorted, for common base, collector, and emitter, respectively
y_{ob}, y_{oc}, y_{oe}	Small-signal output admittance, a-c input shorted, for common base, collector, and emitter, respectively
y_{rb}, y_{rc}, y_{re}	Small-signal reverse-transfer admittance, a-c input shorted, for common base, collector, and emitter, respectively
Z	Impedance, general
z	Open-circuit impedance parameter
Z_F	Large-signal forward impedance
z_f	Small-signal forward impedance
z_{fb}, z_{fc}, z_{fe}	Small-signal forward-transfer impedance, a-c output open, for common base, collector, and emitter, respectively
z_{ib}, z_{ic}, z_{ie}	Small-signal input impedance, a-c output open, for common base, collector, and emitter, respectively
Z_L	Load impedance
z_{ob}, z_{oc}, z_{oe}	Small-signal output impedance, a-c input open, for common base, collector, and emitter, respectively
z_{rb}, z_{rc}, z_{re}	Small-signal reverse-transfer impedance, a-c input open, for common base, collector, and emitter, respectively

3.2 DIODE CIRCUITS.

Semiconductor diodes are employed for rectification and detection similarly to electron-tube diodes; in addition, they have special properties that make them particularly useful for bias and voltage stabilization. Since junction diodes can be made of the same material as the transistor and have the same temperature coefficient and resistance, they will **track** better over the same temperature range, providing nearly ideal thermal compensation. Likewise, application of the avalanche breakdown phenomena provides a special voltage-stabilizing (Zener) diode.

3.2.1 **Junction Diode Theory.** When P-type and N-type germanium are combined in manufacture, the result is a P-N junction diode, which has characteristics similar to that of the electron-tube diode. If properly biased, the junction diode will conduct heavily in one direction and very lightly or practically not at all in the other direction. The P and N sections of the diode are analogous to the plate and cathode of the electron-tube diode. The direction of heavy current flow is in the forward, or easy current, direction; the flow of light current (back current) is in the reverse direction. To produce a forward current flow, it is necessary *to bias the junction diode properly.* Figure 3-1 illustrates proper bias connections for forward and reverse currents. The triangle in the graphical symbol (sometimes called an arrowhead) points against the direction of electron current flow. It is evident that the polarities and electron current flow of the junction diode are identical with those of the vacuum-tube diode and the crystal semiconductor (point-contact) diode. Because of the reverse (back) current flow (which is not present in any appreciable amount in a vacuum-tube) and the resistivity of the semiconductor, the operation and theory of a junction diode differ somewhat from that of the electron tube diode.

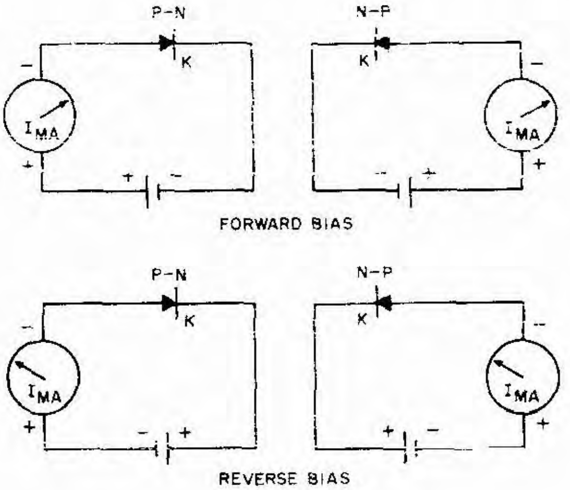


Figure 3-1.
Diode Biasing Circuits

A discussion of the operation of the PN (and the NP) junction and its application to the transistor is included at this point, since an understanding of current flow through the junction and how it varies with external applied bias is essential to circuit operation in later discussions. In semiconductor theory two types of current carriers are encountered, namely, electrons and holes. At room temperature heat *energy imparted to a semiconductor* causes some electrons to be released from their valence bands with sufficient energy to place them in the conduction band. The resultant vacancy created in the valence band possesses a positive charge and is called a **hole.** When holes are present in the **valence** band, electrons can change their energy state and conduction is possible by **hole** movement. Likewise, electrons in the **conduction** band can also change their state and conduction is also possible by **electron** movement. Thus, conduction within the semiconductor is caused by the movement of positive (hole) and negative *(electron) carriers.* Although the movement of holes is the result of the movement of electrons, the charge which moves is positive; therefore, it is common to speak of hole movement in the valence band rather than electron movement. In contrast, in a pure conductor such as copper the conduction band and valence bands overlap and are not separated by a forbidden region; thus there is an excess of electrons available, and conduction is spoken of only in terms of electron movement.

An intrinsic semiconductor is one to which no impurities have been added, and in which an equal number of electron and hole carriers exist. An extrinsic semiconductor is one to which an impurity has been added, and in which conduction takes place primiarily by one type of carrier. Although the amount of impurity is extremely small (on the order of one part in one million or less), the effect upon the conductivity *of the semiconductor* is profound. The addition of an impurity which creates a majority of electron carriers is known as a **donor** (because it donates electrons), and the extrinsic semiconductor which results is called N-type. Likewise, the addition of an impurity which creates *a majority* of hole carriers produces a P-type semiconductor, and is referred to as an **acceptor** impurity (because it will accept electrons).

A PN junction is a single *crystal consisting* of P and N types of semiconductors formed by an alloying or growing process. To facilitate an understanding of its operation, it is assumed that if the P and N materials are brought together externally the junction will function normally, *although actually* it will not. Each type of material is considered to be electrically neutral. When the P and N materials are brought into contact to form the PN junction, a concentration gradient exists for electrons and holes. Holes diffuse from the P material into the N material, and electrons diffuse from the N material into the P material. This process continues until the donor and acceptor sites near the junction barrier lose their compensating carriers and a *potential gradient is built* up which opposes the tendency for further diffusion. Eventually a condition of balance is reached where the current across the junction becomes zero. Figure 3-2 shows graphically the relationships across the junction and the final charge dipole which results from the diffusion process.

The region containing the uncompensated donor (negative) and acceptor (positive) ions is commonly referred to as the **depletion** region. (Since the acceptor and donor ions are fixed and are charged electrically, the depletion region is sometime called the **space charge** region.) The electric field between the acceptor (positive) and donor (negative) ions is called a **barrier**, and the effect of the barrier is considered to be represented by a space-charge equivalent battery (commonly called a **potential hill** battery). In the absence of an external field, the magnitude of the difference in *potential across* the space-charge **equivalent battery** (this potential is not available for external use as a battery) is on the order of tenths of a volt.

When the negative terminal of an external battery is connected to the P material and the positive terminal is connected to the N material, the junction is said to be

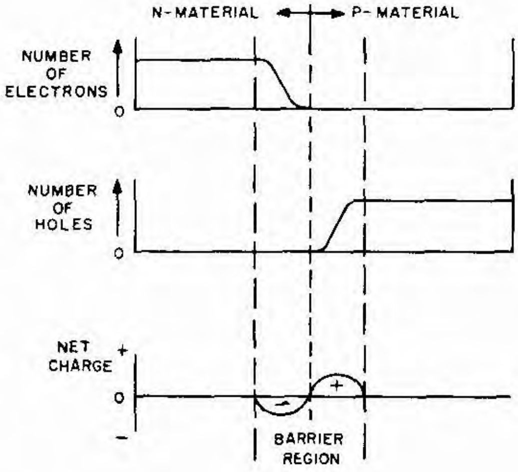


Figure 3-2.
Electric Field Relationships

reverse-biased. In this condition, the external battery polarity is the same as that of the potential-hill battery as shown in figure 3-3. Therefore, the bias battery aids the potential-hill battery, and very little or no forward current passes across the junction. This action occurs because the holes are attracted to the negative terminal of the external battery and away from the junction. Similarly, the electrons are attracted to the positive terminal of the battery and away from the junction. Thus, the depletion area is effectively widened, and the potential across the junction is effectively increased, making it more difficult for normal current to flow. With the majority carriers effectively blocked, the only current that can flow is that caused by the minority carriers, and it is in the opposite (or reverse current) direction. This reverse current is called **back current**, and is substantially independent of reverse-bias values until a certain voltage level is reached. At this voltage the covalent bond structure begins to break down, and a sharp rise in reverse current occurs because of **avalanche** breakdown. The breakdown voltage is popularly called the **Zener voltage**, although there is some doubt as to the manner in which it occurs. Once the crystal breaks down, there is a heavy reverse (back) current flow, which, if not controlled, can overheat the crystal and cause permanent damage. If the current is kept at a safe value, the crystal will return to normal operation when the reverse bias is again reduced to the proper value. The construction of the junction determines the type of back current flow. A crystal with more N-type material than P-type material will have a back current due to electron flow; conversely, a crystal with predominantly P-type material will have a back current due to hole flow.

Back current exists solely because the depletion area, although depleted of majority carriers, is never entirely free of minority carriers, and, since they are effectively polarized opposite to the majority carriers, the external reverse-bias polarity is actually a forward bias for the minority carriers.

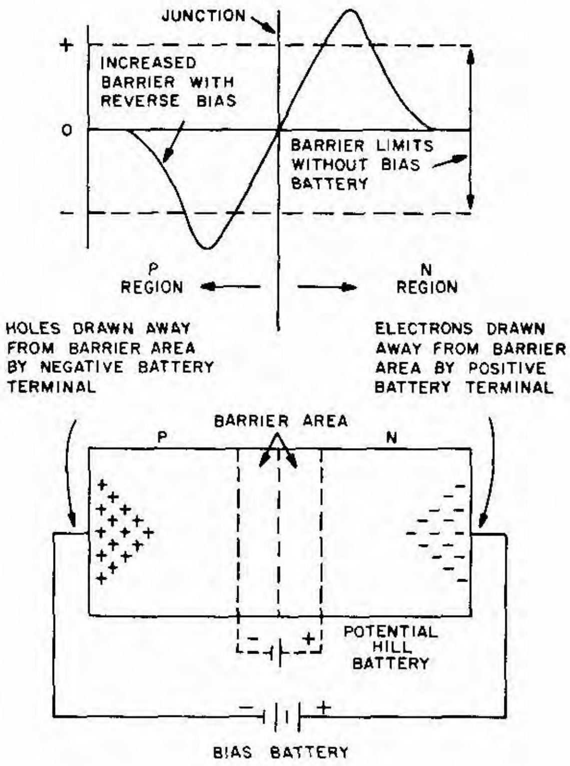


Figure 3-3.
Reverse Bias Conditions

When the external bias battery is connected so that it is oppositely polarized to the potential-hill battery (positive to P region and negative to N region), the barrier voltage is reduced, and a heavy forward current flows; this bias condition is called **forward bias**. Forward current flow is heavy because the electrons of the N region are repelled from the negative battery terminal and driven toward the junction, and the holes in the P region are forced toward the junction by the positive terminal. Depending upon the battery potential, a number of electrons and holes cross the barrier region of the junction and combine. Simultaneously, two other actions take place. Near the positive terminal of the P material the covalent bonds of the atoms are broken, and electrons are freed, to enter the positive terminal. Each free electron which enters the positive terminal produces a new hole, and the new hole is attracted toward the N material (toward the junction). At the same time an electron enters the negative terminal of the N material and moves toward the junction, heading for the positive terminal of the P material. This action reduces the effective value of the potential hill so that it no longer prohibits the flow of current across the barrier, or junction, as shown by the upper portion of figure 3-4. Internal current flow occurs in the P region by holes (the majority carriers) and in the N region by electrons (also majority carriers). Externally, the current consists of electron flow and is dependent upon the bias battery potential.

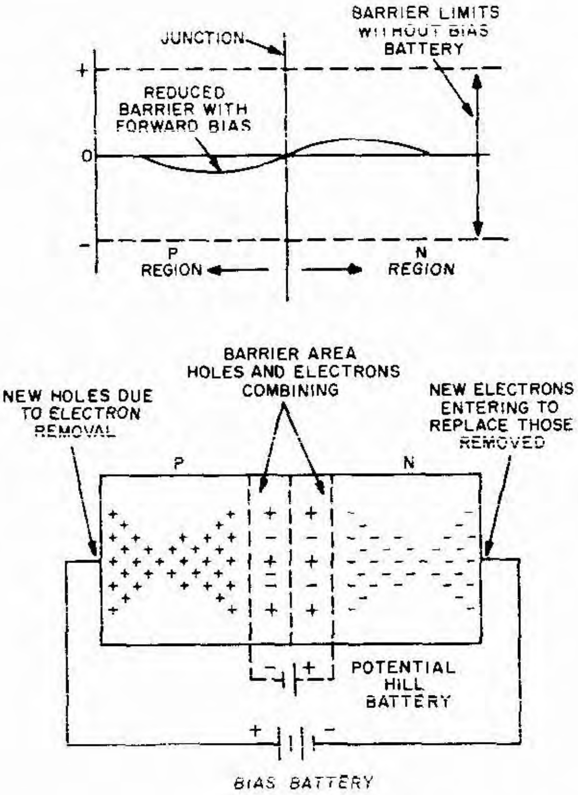


Figure 3-4.
Forward-Bias Conditions

If the forward bias is increased, the current through the junction likewise increases, and causes a reduced barrier potential. If the forward bias were increased sufficiently to reduce the barrier potential to zero, a very heavy forward current would flow and possibly damage the junction because of heating effects. Therefore, the forward bias is usually kept at a low value. Although the initial junction barrier potential is on the order of tenths of a volt, the material comprising the junction is a semiconductor and has resistance. Thus the applied bias must be sufficient to overcome the resistive drop in the semiconductor; as a rule, one or two volts is usually required to produce a satisfactory current flow.

Even though the depletion region is less depleted with forward bias, the minority carriers still exist. The flow of reverse leakage current is practically negligible, however, because the forward bias is in effect a reverse bias to the minority carriers and reduces the back current practically to zero.

The dynamic transfer characteristic curve of the junction diode, illustrated in figure 3-5, shows how the conduction varies with the applied voltage. Observe that as the reverse bias is increased, a point is reached where the back current suddenly starts to increase. If the reverse bias is increased still further, avalanche breakdown occurs and a heavy reverse current flows as a result of crystal breakdown (sometimes called the **Zener effect**). The minimum breakdown voltage of the junction diode corresponds to the maximum inverse peak voltage of an electron-tube diode.

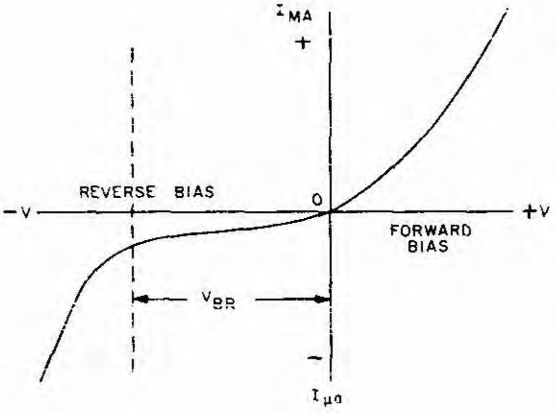
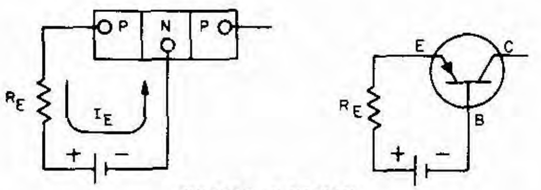


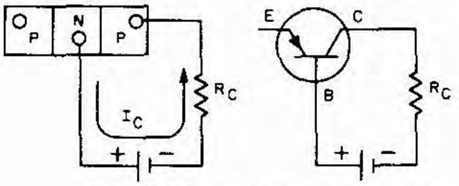
Figure 3-5.
Diode Transfer Characteristic Curve

Basically, a transistor consists of two junction diodes placed back-to-back with the center element being common to both junctions. Connecting the common elements of two junctions together externally will not produce proper results, but when manufactured as one piece, the PNP junction transistor can be considered as such for ease of understanding. The diode junctions of the transistor are biased with d-c potentials, emitter to base in a forward direction, and collector to base in a reverse direction. See figure 3-6.

Transistors are made of NPN materials, as well as PNP materials. Current flow in one type is in exactly the opposite direction to that in the other type, and biasing polarities are reversed. Otherwise, they operate identically except that the internal current flow in the transistor is considered to be the result of hole conduction for the PNP type and electron conduction for the NPN type. All external flow in a transistor circuit is electron flow as in the electron tube. Refer to paragraph 3.3 for a complete discussion of



EMITTER JUNCTION
FORWARD BIASED



COLLECTOR JUNCTION
REVERSE BIASED

Figure 3-6.
Forward- and Reverse-Biased Junctions

transistor action.

3.2.2 **Forward-Biased Diode Stabilization.** The circuit of figure 3-7 employs junction diode CR1 as a forward-biased diode to compensate for transistor emitter-base resistance variations with temperature. This type of circuit compensation is effective over a range of from 10 to 50 degrees Centigrade (usually no compensation is needed below 10 degrees). Higher temperature ranges require additional compensation (see paragraphs 3.2.3 and 3.2.4).

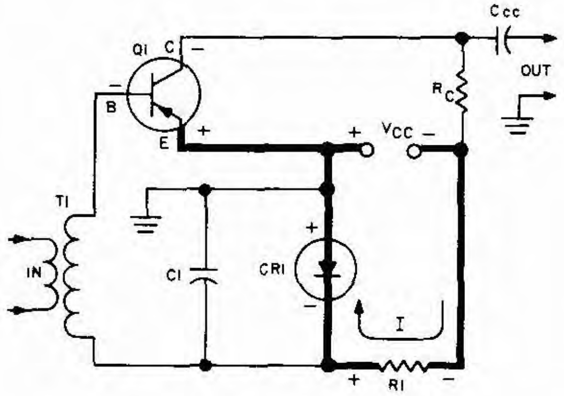


Figure 3-7.
Forward-Biased Diode Stabilization, CE Circuit

Under static conditions, the emitter of transistor Q1 is biased positive with respect to the base (forward biased), and current flows through R1 and external diode CR1 connected across the source voltage, V_{cc}. In this condition, CR1 is forward-biased also, and the polarities around the circuit are as shown in figure 3-7. Current flow is light (on the order of 100μa or less), and junction diode CR1 can be considered as a resistor (r_e about 25000 ohms). Thus the transistor emitter-base junction bias consists of the voltage drop across diode CR1. Since every junction diode has a negative temperature coefficient of resistance, an increase in temperature causes the transistor emitter-base junction resistance to decrease, and would normally produce an increased collector current in the transistor. However, the resistivity of junction diode CR1 also decreases with an increase of temperature, the diode voltage drop is lower, and increased diode current flow causes a larger voltage drop across R1 (which opposes the diode-developed bias); therefore, less actual bias is developed. The net effect is to reduce the total forward bias on the transistor and thus lower the collector current sufficiently to compensate for the increase of collector current with temperature.

The d-c secondary resistance of T1 does not offset the operation since the transistor base current flow is negligible. However, considering the collector-to-base reverse-bias (saturation) current, I_{CBO}, which flows from the base through T1, CR1, V_{CC}, and R_C to the collector, we find no compensation is provided by this circuit. The normal current through forward-biased diode CR1 is so heavy that it effectively swamps the small I_{CBO} current (on the order of 2 or [illegible] as compared to 75--200 μa for the normal current). As far as signal variations are concerned, capacitor C1 effectively bypasses diode CR1 and the output voltage developed across R_C is applied to the next stage, through coupling capacitor C_{cc}, in the conventional manner.

3.2.3 **Reverse-Biased Diode Stabilization.** The circuit of figure 3-8 employs external junction diode CR1 as a reverse-biased diode to compensate for transistor collector-base saturation current variations with temperature. This type of circuit is effective over a wide range of temperatures when the diode is selected to have the same reverse-bias (saturation) current as the transistor. The reverse-biased diode provides a high input resistance, which is particularly advantageous when the preceding stage is resistance-capacitance coupled.

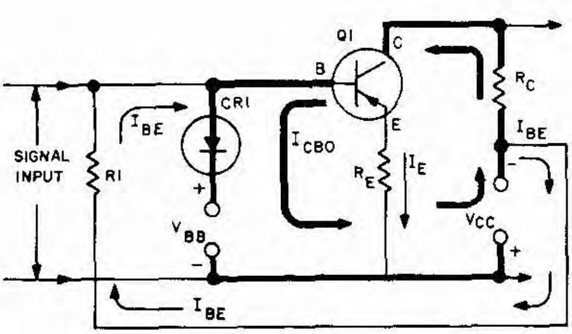


Figure 3-8.
Reverse-Biased Diode Stabilization, CE Circuit

Two current paths are provided by the circuit shown in figure 3-8. The base-emitter current (I_{BE}) flows internally from the base to the emitter, then externally through R_E and V_{CC}, and through resistor R1 back to the base, and is not materially affected by diode CR1 because of its large resistance in the reverse direction. The other path provides for the saturation current from collector to base, *through* CR1, V_{BB}, and V_{CC}, and through collector load resistor R_C to the collector. Temperature variations in the *emitter-base junction resistance* are compensated for in the first path by swamping resistor R_E (see paragraphs 3.3.1 and 3.4.2). *Variations of saturation* current with temperature are compensated for by the second path, using diode CR1. When a temperature *increase causes the transistor* junction saturation current to rise, the diode saturation current also increases, so that there is no chance for *the* I_{CBO} *current* carriers in the junction to pile up, increase the transistor forward bias, and cause a consequent *rise in emitter current.* As a result, although the saturation current may increase, the emitter current does not, and there is *only a negligible* change in the total collector current. In effect, diode CR1 operates similarly to a variable grid-leak in an electron tube circuit. The reverse bias permits only a few microamperes of current to flow in the base-to-emitter circuit, and maintains a high input resistance, while simultaneously compensating for changes in I_{CBO} with temperature.

3.2.4 **Double-Diode Stabilization.** The circuit of figure 3-9 utilizes two junction diodes in a back-to-back arrangement. Junction diode CR1 is forward-biased and compensates for emitter-base junction resistance changes with temperatures below 50 degrees Centigrade, as described in paragraph 3.2.2 above. Diode CR2 is reverse-biased and compensates for higher temperatures as discussed below.

Forward-biased diode CR1 operates in the same manner as discussed in paragraph 3.2.2, and the parts are labelled exactly as in figure 3-7.

Reversed-biased diode CR2 can be considered inoperative at room temperatures and below. When the junction temperature reaches the point where saturation current flows, CR2 conducts and current (I_1) flows through R2, producing a voltage drop with the polarity as shown in figure 3-9. This voltage drop is in the proper direction to reduce the forward bias set up by diode CR1 and R1; its net effect is to reduce the total collector current, *to compensate* for the increase in transistor I_{CBO} due to temperature increase.

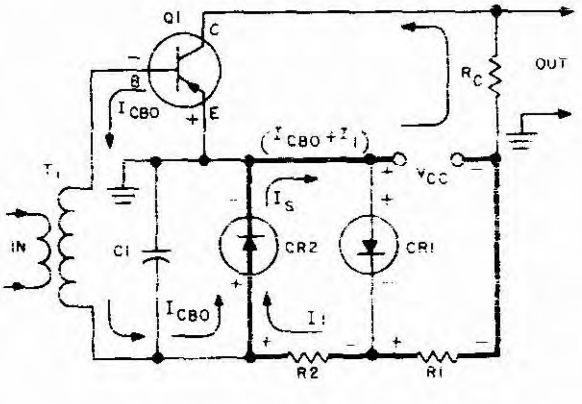


Figure 3-9.
Double-Diode Stabilization, CE Circuit

The reverse-biased diode is selected to have a larger saturation current (I_S) than the transistor it stabilizes, since I_S consists of the transistor saturation current plus the current through R2 ($I_S = I_{CBO} + I_1$). Thus the diode saturation current controls the transistor at all times, effectively reducing the forward bias as the temperature increases and stabilizing the collector current. Capacitor C1 bypasses both diodes for ac so the bias circuit is not affected by signal variations.

3.2.5 **Diode Voltage Stabilization.** If a junction diode is reverse-biased, current flow does not entirely cease, but continues at a low rate (a few microamperes), until the bias is increased to the point where it reaches the breakdown voltage. If the reverse-bias is increased *beyond the breakdown point,* the diode reverse-saturation current suddenly increases, because of the avalanche effect, and the *applied voltage remains* practically constant. In most cases this will destroy the diode because of overheating (except *in the special case of the* Zener diode). Reduction of the bias below the breakdown voltage level returns the junction to *its normal operation again* (provided that no damage has occurred). Application of the avalanche phenomenon has resulted in the development of a voltage-stabilizing diode known as the breakdown diode, often called a Zener diode.

The breakdown, or Zener, diode is a PN junction modified in the manufacturing process to produce a breakdown voltage level which is closely controlled over a range of from 2.5 to 200 volts or more. Each Zener diode has a specific breakdown voltage (and operates over a small voltage range), depending upon its design characteristics, and must be selected for the desired operating voltage.

Because of its unique properties, this diode has many uses other than the basic voltage-regulating application. For example, it may be used for surge protection, as an arc reducer (across contact points), as a d-c coupler in an amplifier, as a reverse polarity gate, or as a biasing element. Its basic properties will be discussed in the following paragraphs for proper understanding, and special applications will be treated as the need arises in other parts of this manual.

Figure 3-10 shows a typical [illegible] voltage-regulating circuit of the most elementary type. Resistor R1 is selected to produce the proper breakdown voltage to main-

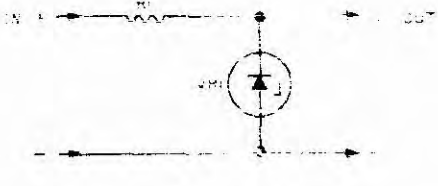


Figure 3-10.
Breakdown Diode Voltage Regulator

tain diode VR1 at the correct operating point. When the input voltage rises, current through the diode increases, and the drop across R1 becomes greater so that the output voltage remains the same. Conversely, when the input voltage decreases, current through the diode decreases, and the drop across R1 becomes less so that the output voltage again remains the same. Should the load resistance decrease, and more current be required, the current is divided between the load and the diode to ground path, so that no more current is drawn through R1 and the voltage across VR1 remains the same. When the load resistance increases so that less current is required, the additional current is absorbed in the diode current flow to ground, again dividing so that no additional or reduced current passes through R1, and the voltage across the diode remains the same. Thus through the avalanche effect, the breakdown diode operates similarly to the electron tube type of glow discharge voltage regulator.

3.2.5.1 **Temperature Compensation.** Forward-biased junction diodes have a negative temperature coefficient of resistance, and so do reverse-biased junctions until the breakdown voltage is reached. Once the diode is operating in the avalanche-effect region, the temperature coefficient becomes positive and of a larger value. An uncompensated diode can vary as much as 5 percent (of the maximum rated voltage); with temperature compensation, however, it is possible to reduce this figure to .0005 percent (a few millivolts) or better.

Because a forward-biased junction changes resistance with temperature in exactly the opposite direction from the breakdown diode, it is possible to use one or more forward-biased diodes for temperature compensation. Figure 3-11 shows a basic compensation circuit, in which CR1 is a forward-biased diode and VR2 is the breakdown diode. The CR1 diode is selected to have a temperature characteristic which is the exact inverse of the breakdown diode's temperature characteristic (if necessary more than one diode is used in series, but usually not more than three). Thus the combined resistance of both diodes (CR1 and VR2) in series remains constant over a wide range of temperatures and voltages to produce the desired compensation. The compensating diode must be able to pass the current taken by the breakdown diode, and should not introduce any appreciable voltage drop in the circuit (across the diode itself). Fortunately, the use of forward bias produces only a very small voltage drop, depending on the material forming the junction (about 0.6 to 1.5 volts for silicon), and units with adequate current ratings are available to produce the desired regulation. A particular advantage of the breakdown diode over the electron tube type of voltage regulator is that breakdown voltages are easily manufacturable over a wide range of approximately 2.5 to 200 volts, whereas the lowest tube voltage available is about 70 volts. Adequate voltage stabilization for quite a wide range of circuit operations and temperatures is therefore available. Unique applications of the preceding principles will be discussed in connection with applicable circuits in other sections of this manual.

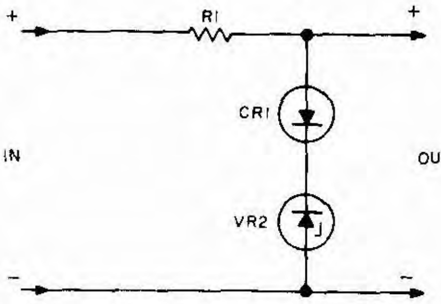


Figure 3-11.
Breakdown Diode Temperature Compensation

3.2.6 **Shunt-Limiting Diode.** The circuit of figure 3-12 utilizes a junction diode connected in shunt between the base and emitter of the transistor as a protective peak limiter for transient voltages.

Under normal circuit conditions diode CR1 is inoperative, being reverse-biased by the potential across resistors R1 and R2, and the transistor is forward-biased by the drop across R1. When an oppositely polarized input signal (or noise transient) exceeds the bias voltage across R1, diode CR1 becomes forward-biased and conducts, effectively shunting the transistor base and emitter terminals. Thus the base-emitter junction of the transistor is prevented from becoming reverse-biased by excessive signal swing.

This peak limiting action is particularly applicable to transformer-coupled transistor stages because they develop transient voltages when the collector current is suddenly cut off by bias reversal. The resulting high collector-emitter voltage with base-emitter circuit reversed-biased can then produce strong internal oscillations and cause excessive power dissipation, which could destroy the transistor. In this circuit C1 is a low-resistance bypass capacitor to shunt R1; it is not used to resonate with the secondary of T1.

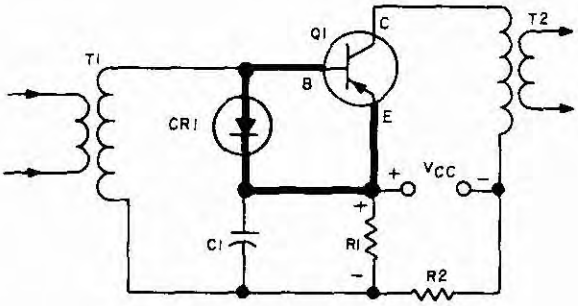


Figure 3-12.
Basic Shunt-Limiting Diode Circuit

3.2.7 **Diode AGC Circuit.** Figure 3-13 shows a typical diode AGC circuit which uses the variation of collector current in the second stage to control the diode. Proper base-emitter bias of Q2 is achieved through the voltage divider consisting of R3, R4, and R5. (Note that at audio frequencies detector diode CR2 is in parallel with R3, since the reactance of the transformer secondary is negligible.) The potential at the base of Q2 is negative with respect to ground and is numerically equal to V_{CC} minus the drop across R5; it is just slightly negative with respect to the emitter. This bias, in the absence of any input signal, establishes the static collector current through Q2, which is of such magnitude that the voltage across R2 is slightly greater than that across R1, thereby keeping diode CR1 cut off.

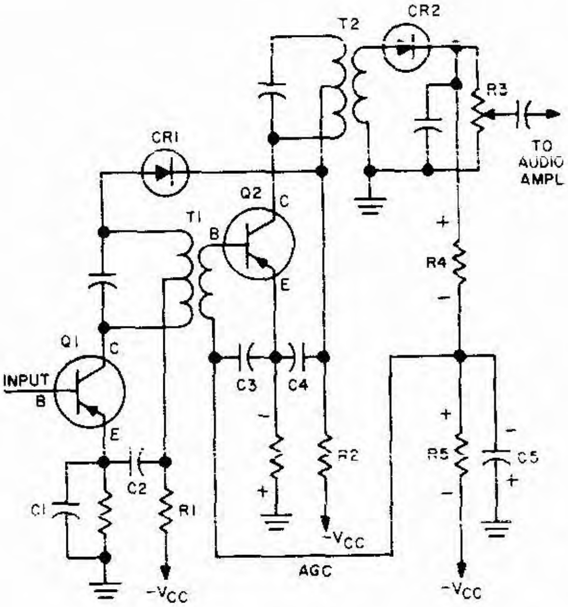


Figure 3-13.
Basic Diode AGC Circuit

As the carrier level of the incoming signal increases, the increased conduction through the detector diode, CR2, results in an increase through R4 and R5. The increased voltage across R5 reduces the negative potential at the base of Q2 (V_{CC}—E_{R5}), therely reducing the forward bias and hence the collector current also. This in itself may be sufficient to reduce the gain of Q2 somewhat because of a change in transistor parameters.

However, another effect is also realized through the action of CR1. As the I_C of Q2 decreases, the voltage across R2 decreases, and the **reverse bias** of CR1 changes to forward bias. When the diode is forward-biased it conducts, and, being connected across the output tank of Q1, it effectively shunts or damps the signal. (CR1 shunts the collector tank of Q1 since the reactance of the bypass capacitors is small.) As a result, the input of Q2 is diminished. Diode conduction never becomes so heavy that it acts as a short circuit across the tank circuit. Instead, the conduction is light, and the internal resistance of the semiconductor diode (which can be appreciable at light currents) acts as a variable resistance shunt to maintain the output of Q2 substantially constant. Since R1 is shunted by C1 and C2, and R2 is shunted by C3, C4, and C5, the diode is not affected by the a-c signal variations, but is controlled only by the applied d-c bias which is indirectly determined by the AGC voltage across R5.

3.2.8 **PNPN Switching (Four-Layer Diode).** The four-layer diode is a two-terminal device which operates in either of two states; an open, or high-resistance, state or a closed, or low-resistance, state. It is effectively an on-off switch which can be employed as such in switching circuits, or it may be used as a relaxation oscillator or multivibrator. In the non-conducting state it presents a resistance on the order of megohms, and in the conducting state it presents a low resistance of about 20 ohms.

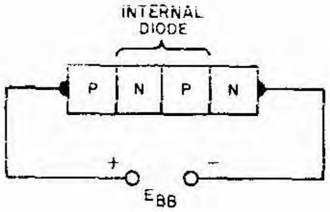


Figure 3-14.
PNPN Diode Bias Conditions

Figure 3-14 shows typical biasing polarity with the P-end connected to positive and the N-end connected to negative, a forward-biased condition. The internal NP junction, then, is reverse-biased and held in an almost nonconducting condition; this is the resting open-circuit, or high-resistance, condition. As the bias voltage is increased, it reaches a point which produces a condition similar to avalanche breakdown in the internal junction, and heavy current flows. Since there is no connection to the internal junction except through the other semiconductor material, there is a small minimum resistance at full current flow; this is the closed-circuit, or low-resistance, condition.

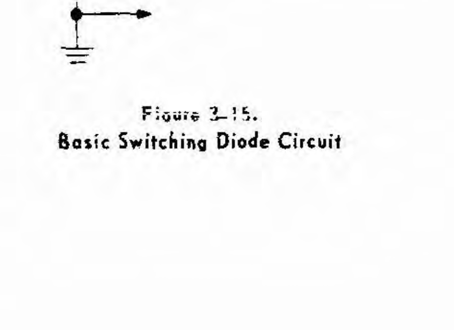


Figure 3-15.
Basic Switching Diode Circuit

Figure 3-15 shows an elementary switching diode circuit in which the diode allows the capacitor to alternately charge and discharge at a rate determined by the capacitor and resistor time constant. When the capacitor charging voltage reaches the diode breakdown voltage, the diode discharges the capacitor until the discharge voltage is unable to substain the breakdown condition; the diode then ceases to conduct, and the charging cycle begins again.

3.2.9 **Photodiodes.** The photodiode is a specially constructed junction diode arranged so that it is possible to utilize light variations impinging on the junction to produce output current variations.

Both PN and NP junctions may be used. Operation is based upon the fact that photons striking the junction produce electron-hole pairs in the junction which cause variations in current when the junction is reverse-biased. Figure 3-16 shows a typical photodiode circuit. It is essentially a d-c biased junction connected in series with a load resistance. Either current or voltage variations produced in the load resistance may be utilized.

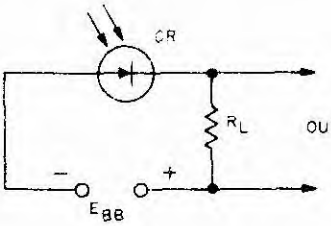


Figure 3-16.
Photodiode Input Circuit

Normally the photodiode junction produces a slight reverse current in the absence of light. When light is focussed on the PN (or NP) junction, the current through the junction is increased. The amount of current is in proportion to the light; as the light increases, the junction current increases, and as the light decreases the junction current decreases. The response of the photodiode is greatest when the light is directed exactly at the PN junction and drops off on either side (±0.5 millimeter) of the junction rather rapidly. The small physical size, high efficiency, low power consumption, low noise level, and simple circuitry of the photodiode makes its use particularly advantageous. It is affected adversely by temperature changes and humidity.

3.2.10 **Tunnel Diodes.** Basically the tunnel diode is a PN junction formed of a semiconductor mixture which is impregnated with more than the normal amount of impurities. The resulting crystal has current-voltage transfer properties which are different from those of the normal diode junction.

Figure 3-17 shows a typical junction diode transfer characteristic plotted as a dashed line, with the tunnel diode characteristic superimposed in the form of a solid line. Note that diode forward current increases with applied forward-bias voltage until point A is reached, then it reverses itself and decreases with increase of applied voltage to point B; between points B and C the curve is similar to that between 0 and point A. The region of operation between points A and B is a negative-resistance region over which the diode may be used an an amplifier or oscillator, since negative-resistance devices are capable of supplying power to the source instead of absorbing power from it.

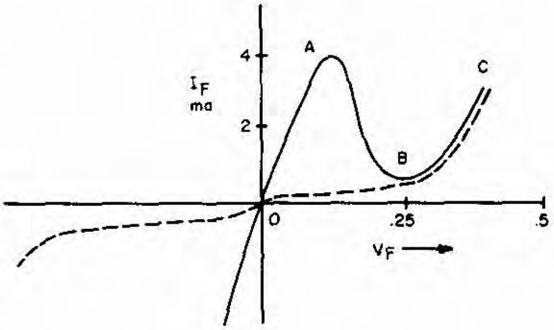


Figure 3-17.
Tunnel Diode Transfer Characteristics

The tunnel diode is also of value in switching circuits. Because three or more regions of operation are possible, it is only necessary to select the proper bias and load line to achieve different types of operation with the same circuit. Figure 3-18 shows a basic switching circuit utilizing the tunnel diode for three different functions. Under condition A the load line is adjusted so that it intersects the characteristic curve at three points; two are in the positive-resistance region and the third is in the negative-resistance region. Application of a positive pulse turns the diode on, and application of a negative pulse turns it off; thus the circuit operates as an off-on multivibrator. Under condition B the diode is operated in the positive-resistance region and functions as a one-shot multivibrator, producing a rectangular pulse for each trigger pulse. Under condition C the diode load line is set so that it never intersects the positive regions of the characteristic curve, and the circuit operates in an astable condition as a free-running multivibrator. The tunnel diode can operate at frequencies up to 1000 megacycles at high switching speeds and appears particularly adapted for computer applications.

Because of its newness, tunnel diode applications have been treated only lightly in this paragraph; they will be discussed more fully, in other sections of this techincal manual, in conjunction with the circuits in which they are used.

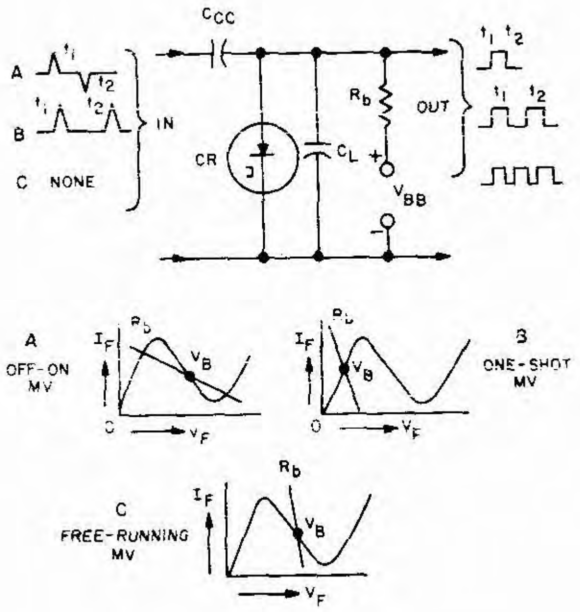


Figure 3-18.
Tunnel Diode Operating Characteristics

3.3 TRIODE COMMON-BASE CIRCUITS.

In the common-base circuit, the input signal is injected into the emitter-base circuit, and the output signal is taken from the collector-base circuit, with the base element being common to both. The common-base circuit is equivalent to the electron tube grounded grid circuit. It has a low input resistance (30 to 160 ohms) and a high output resistance (250K to 550K), and is mostly used to match a low-impedance circuit to a high-impedance circuit. It has a maximum voltage gain of about 1500 and a current gain of less than 1, with a power gain of 20 to 30 db. There is no phase reversal between input and output signals; both signals are in-phase and of the same polarity. (That is both the input and output voltages move in the same direction, starting from zero at approximately the same time, throughout the complete positive and negative alternations. When the input signal is positive, so is the output signal, and vice versa.)

It is common practice to speak of a change of phase between an input and an output signal in both electron tube and semiconductor discussions when what actually occurs is only a change of polarity. Actually, in the semiconductor, when a long transit time occurs (with respect to the frequencies being amplified), the output signal is delayed in starting because of the finite time taken for the input signal to reach the output terminal. This delay (transit time) produces an actual phase (time) difference between the input and output signals even though the polarities may be identical, or even opposite.

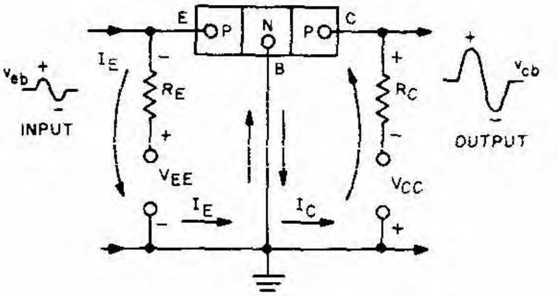


Figure 3-19.
PNP Common Base Circuit

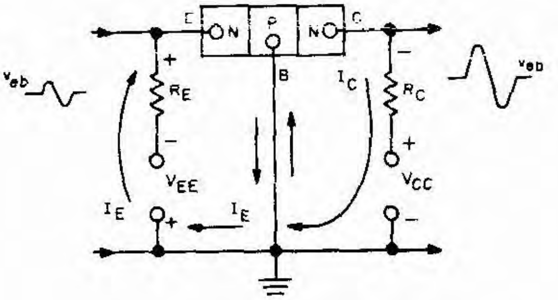


Figure 3-20.
NPN Common Base Circuit

The common-base connections for PNP and NPN transistors are shown in figures 3-19 and 3-20, respectively, together with polarities and external current paths. Emitter-base bias and collector-base bias are obtained from separate sources so that two voltage sources are required. With this type of connection, it is possible to ground the base directly for both ac and dc, if desired. The circuit for a single voltage source will be discussed later.

The semiconductor device symbol is used in figures 3-19 and 3-20 rather than the graphical transistor symbol for ease of presentation and understanding, since it better indicates the functioning and construction of the device. Both the PNP and NPN circuits are shown together so their operation can be more easily compared. Both circuits are forward-biased from emitter to base, and reverse-biased from collector to base. The polarities of the two circuits are opposite because the transistors are of opposite composition, and the currents in the external circuit flow in opposite directions. Note that in the PNP circuit the external current (electron) flows from emitter to collector, while in the NPN circuit it flows from collector to emitter. Internally, however, the flow is *always from emitter to* collector through the base region. (Current theory explains the internal flow through the medium of "holes" for the PNP transistor and "electrons" for the NPN transistor.) There is one current from emitter to base and another from collector to base; in some instances these currents combine and in other instances they oppose each other, depending upon circuit configuration, biasing, and applied signal voltages. The emitter current is always greater than the collector current, and the algebraic difference between the two is the base current. The base current will be discussed when it is relevant to circuit action; otherwise, it will be ignored since it is a very small portion of the total current involved. In the junction type transistor, about 95 percent of the emitter current always reaches the collector. Transistor action is dependent entirely upon the fact that the signal applied to the low-impedance input (emitter) circuit causes a current flow change which, when transferred to the high-impedance output (collector) circuit, produces a voltage gain. The current gain (α) of the common-base circuit

is defined as $\alpha = \frac{\Delta I_C}{\Delta I_E}$, or the ratio of a small change

in emitter current to a small change in collector current produced by the emitter current change. It is always less than one.

A brief discussion of the basic current flow in a transistor is included at this point because the flow of current, both internally and externally, are important to a complete understanding of circuit action. Figure 3-21 illustrates the current flow in each junction separately and in the complete transistor. In part A, current flow is shown for the emitter-base junction with forward bias applied and the collector open-circuited. The flow externally is by electrons from the emitter to the base. Internally the flow is from emitter to base through the majority hole carriers and from base to emitter through the minority carriers. Assuming a value of 1 milliampere for the emitter current, I_E, the base current, I_B, is approximately the same because the collector is open (neglecting any external or internal leakage currents). The hole current predominates, there being on the order of 200 hole carriers to 1 electron carrier because of the P-doping effect. Since one milliampere of current amounts to 6.28×10^{15} carriers per second, it can be said that the internal emitter current flow consists of 6.24×10^{15} holes per second plus 3.14×10^{13} electrons per second, and the external flow is 6.28×10^{15} electrons per second.

In part B of figure 3-21, the emitter is open-circuited and the collector-base junction is reverse-biased. Since reverse bias reduces current flow to a mimimum, the collector current, I_C, is actually I_{CO}, the reverse leakage current from collector to base, plus any surface leakage effects, which are neglected in this discussion. Internally, there is a hole current from base to collector (minority carrier) plus an electron flow from collector to base (also a minority carrier). The minority carriers are the cause of current flow because of the reverse bias; the actual current flow is very small, being on the order of 20 microamperes (or less) for a typical transistor. Since the emitter is open, there is no flow from emitter to collector, and the base current, I_B, is approximately the same as I_{CO}.

In part C of figure 3-21, both the emitter and collector circuits are completed, forward bias is applied to the emitter, calling for strong current flow, and reverse bias is applied to the collector, calling for minimum current flow (considering junction biasing only). However, because of the reverse bias on the collector, the collector is connected to the negative supply source. Therefore, the potential hill across the collector junction is reduced, providing an attraction for the hole current diffusing through the base from the emitter. Thus, an easy flow of hole current is permitted and it accounts for most of the emitter current transfer from emitter to collector. As a result of base injection, however, there is an electron current flow from the base to emitter which cannot be collected through the collector because the collector junction barrier polarity opposes conduction of negative charges. There is also a recombination current which consists of holes that flow into the base; these holes

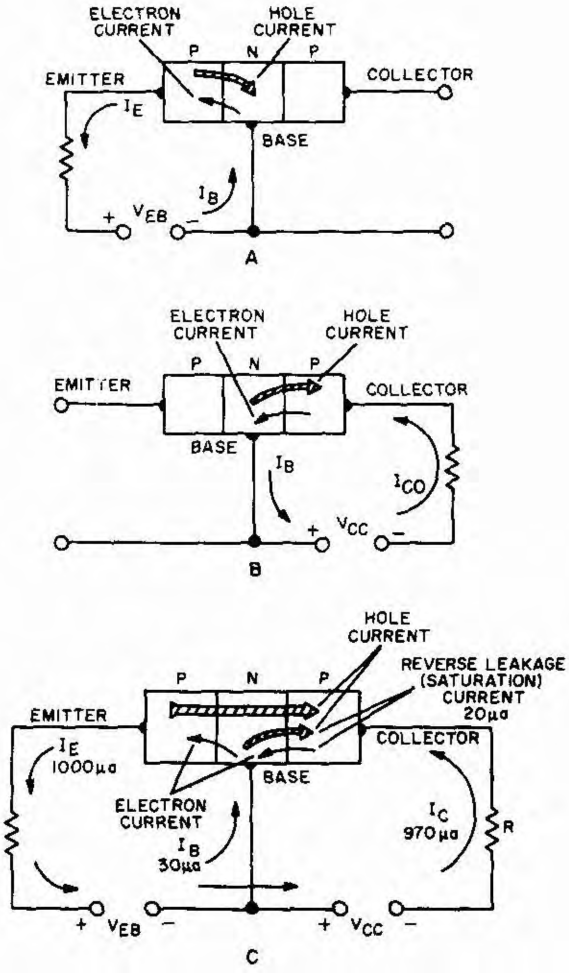


Figure 3-21. Transistor Current Flow

combine with electrons before reaching the collector and thereby cause an electron flow in the base lead and into the base. This current represents a loss, and reduces the amount of emitter current that can reach the collector. The current through the collector junction consists of the emitter current which is permitted to reach the collector, plus the reverse leakage current, I_{CO}. Assuming that 95 percent of the emitter current reaches the collector, the total current can be represented mathematically by a coefficient α (alpha) times I_E plus the saturation current I_{CO}; that is,

$$I_C = \alpha I_E + I_{CO} \qquad (1)$$

The base current is the difference between the portion of the emitter hole current that does not reach the collector (the group of holes which recombine with electrons in the base) and the saturation current. Therefore the expression for base current is:

$$I_B = (I_E - \alpha I_E) - I_{CO} \text{ or } I_B = I_E (1-\alpha) - I_{CO}. \qquad (2)$$

Using the values of base current (20 μa), emitter current (1000 μa), and alpha (.95) assumed previously, and substituting them into formula (1) gives the following result:

$$I_c = .95\,(1000) + 20 = 950 + 20 = 970\ \mu a$$

Using formula (2) for base current and substituting:

$$I_B = 1000\,(1-.95) - 20 = 50 - 20 = 30\ \mu a$$

It can be seen from the numerical example that the value of the recombination current is 50 μa, that the base current is the difference between the recombination current and I_{CO}, that the collector current is the sum of I_{co} and αI_E and that the various internal currents can be made equal to the external currents. In the example above, the external current is that indicated in figure 3-21, and is an electron flow from emitter to collector with a small amount also flowing into the base. The discussion has assumed small-signal conditions and the use of a PNP transistor in the common-base connection. When other configurations such as common-emitter and common-collector circuits are used, the values of current change somewhat because of the differences in input and output connections and the current paths between them. When the NPN transistor is employed, operation is the inverse of that for the PNP transistor, with electrons acting as majority carriers and holes as minority carriers. For large-signal conditions, operation is slightly different and will be discussed at the appropriate point in the circuit discussions in other sections of this technical manual.

Figure 3-19 (shown previously) shows a small input signal (v_{eb}) applied to the emitter-base junction of the PNP transistor. The circuit is considered to be biased so that it operates over the linear portion of its dynamic transfer characteristic, and is resting in a quiescent state in accordance with the d-c potentials applied (similar to electron tube class A operation). Assuming a sine-wave input, it is apparent that as v_{eb} increases to its maximum positive value the forward-bias between emitter and base is increased, causing the emitter-base junction to produce an increased current flow, which passes through the external circuit and eventually through collector load resistor R_C. The increase of current through R_C causes an increased voltage drop in the positive direction, so that as the input signal reaches its positive maximum so does the output signal. Therefore, both signals are always in phase and of the same polarity.

On the negative swing of the input signal, as the negative signal voltage is added to the positive forward bias, the result is a reduction of the bias and less emitter current flows. In turn, a reduced collector current flow through R_C results in a decreased voltage drop across R_C and produces a reduced output voltage; the negative going output signal remains effectively in phase with the same polarity as the input signal.

The NPN circuit shown in figure 3-20 functions similarly but inversely to the PNP circuit; that is, on the positive swing of the input voltage, the collector current is reduced, and the drop across R_C is less negative (or more positive). On the negative input cycle forward bias is increased, and the collector current produces a greater negative drop across R_C; thus the output voltage also follows the input voltage in phase and polarity.

3.3.1 **Bias (Common Base).** Transistors are normally biased by placing a forward voltage on the emitter-base junction (increasing the bias voltage causes increased emitter and collector current flow), and a reverse polarity voltage on the collector junction. The operating (or bias) point is determined by specifying the d-c, no-signal (quiescent) values of collector voltage and emitter current. Biasing circuits and arrangements are varied; separate supplies such as batteries are commonly used, as well as so called self-bias arrangements, voltage dividers across the collector supply, and other transistors or diodes.

The common-base circuit is usually restricted to the use of separate bias for the emitter-base junction or to a voltage-divider arrangement using a single voltage supply which serves as the collector-base supply also. Figure 3-22 illustrates a typical single-supply type of arrangement. Resistors R1 and R2 form a voltage divider across the collector supply, with the base connected at their junction and the emitter connected to the high side of the supply. Thus the emitter is always at the highest potential, the base is at a lower potential because of the voltage drop across R1 due to the current from the supply source flowing through the voltage divider, and the collector is at the lowest potential. The difference in potential between the emitter and base represents the forward bias applied to the emitter-base junction. For a PNP transistor, as shown in figure 3-22, forward bias is achieved by making the emitter positive with respect to the base; for an NPN transistor, the polarity of the source is reversed, and the emitter is

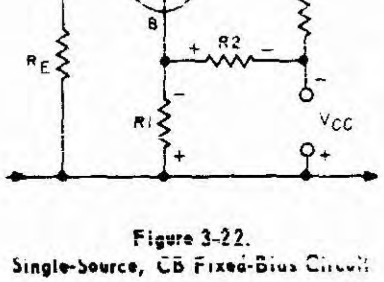


Figure 3-22.
Single-Source, CB Fixed-Bias Circuit

negative with respect to the base. Although the value of R1 is normally low, it may be necessary in some cases for R1 to be bypassed with a very-low-reactance capacitor, to assure that the base is well grounded for ac.

The common-base configuration offers almost ideal thermal compensation, since the input resistance (R_E) in the emitter circuit acts as a swamping resistor, and changes in collector current with temperature are minimized by low base-to-emitter resistance. See paragraph 3-2 for a discussion of diode circuits for stabilization of bias, and paragraph 3.4.2 for a discussion of emitter swamping resistors.

3.4 TRIODE COMMON-EMITTER CIRCUITS.

In the common-emitter circuit, the input signal is injected into the base-emitter circuit and the output signal is taken from the collector-emitter circuit, with the emitter element being common to both. The common emitter is equivalent to the electron tube grounded-cathode (conventional) amplifier circuit. It has a somewhat low input resistance (500 to 1500 ohms) and a moderately high output resistance (30 K to 50 K or more), and is the most commonly used transistor circuit configuration. It is widely used for a number of reasons. Because the input signal is applied to the base rather than the emitter, a considerably higher input impedance is obtained than in the common-base circuit. High power gains are obtainable (25 or 40 db), and an actual current gain is possible (from 25 to 60 or better). The actual voltage gain is slightly less than that of the common-base circuit because of the higher input impedance, but this is partially off-set by the current gain; in practice, voltage gain values of 300 to 1000 (or better) are obtained. Because the signal is applied to the base, a polarity reversal takes place, making the output signal of opposite polarity to the input signal, as in the conventional electron tube amplifier.

The common-emitter connection for PNP and NPN transistors is shown in figures 3-23 and 3-24, respectively, with polarities and **external** current paths. Base-emitter bias is obtained from a separate supply than that of the collector-base junction so that two voltage sources are required. With this type of supply connection it is possible to directly ground the emitter both for ac and dc, if desired. Single voltage supply circuits will be discussed later. The PNP and NPN circuits are shown together for ease of comparison of operation. Both circuits are forward-biased from emitter to base and reverse-biased from collector to emitter. Current flow and polarities in both circuits are opposite because of the difference in material from which the transistors are manufactured.

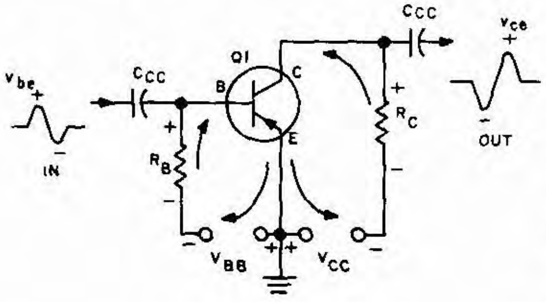


Figure 3-23.
PNP Common-Emitter Circuit

Current flow is from the emitter to the collector through the base region as in the common-base connection, and, likewise, only a small amount of current is diverted in the base-to-emitter circuit. Transistor action also depends on the fact that a small change of current in the low-resistance input circuit produces a voltage gain when applied to the high-impedance output circuit. But unlike the common-base circuit, the current gain is not based on the emitter-to-collector current ratio alpha (α); instead it is based on the base-to-collector current ratio beta (β) because the signal is injected into the base, not the emitter. Therefore, since a small change of base current controls a large change in collector current, it is possible to obtain considerable current gain (a value of 60 is not unusual). Since the collector load resistance, R_C, is less than the load resistance of the CB circuit, less voltage gain might be expected. However, the increased current in the collector produced by current gain off-sets the loss of output resistance, so that the voltage gain is nearly comparable to that of the CB circuit. By manipulation of circuit constants and selection of transistors, the voltage gain can be made to exceed that of the CB circuit.

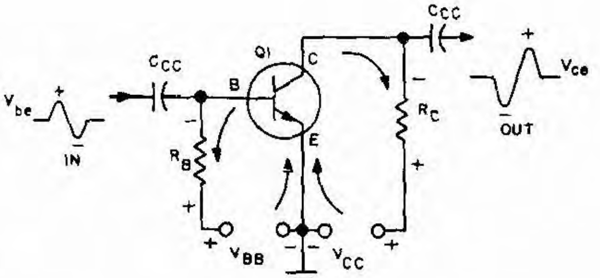


Figure 3-24.
NPN Common-Emitter Circuit

Beta is defined as $\beta = \frac{\Delta I_C}{\Delta I_B}$ with V_C constant, and

is related to alpha of the CB circuit by $\beta = \frac{\alpha}{1-\alpha}$

thus the closer α is to 1, the larger is β; as α approaches 1, β approaches infinity.

Figure 3-23 shows a small input signal (v_{eb}) applied to the base of the PNP transistor. The circuit is biased to operate over the linear portion of its dynamic transfer characteristic, and rests in a quiescent state determined by the static d-c potentials applied (similar to electron tube class A operation). Assuming a sine-wave input, it is apparent that, as v_{eb} increases to its maximum positive value, the forward bias is reduced, less current flows in the emitter and collector circuits, and the drop across the collector output resistor (R_C) becomes less, producing a negative-going voltage. Conversely, as the input signal swings negative, the forward bias is increased and more emitter and collector current flows. The increased voltage drop across R_C is in a positive-going direction, and the output signal reaches a positive maximum. It is evident that, since the output signal is at a positive maximum when the input signal is at a negative maximum and vice versa, the input and output polarities are exactly opposite. Therefore, the common-emitter circuit is similar to the vacuum-tube common-cathode circuit, producing a polarity reversal of the input signal (although not strictly accurate, this polarity reversal is commonly spoken of as a phase difference).

The NPN circuit shown in figure 3-24 functions similarly but inversely to the PNP circuit. That is, when the input signal is positive, the forward bias is increased, and the collector current increases and produces a negative-going output across R_C. On the negative input cycle, the collector output is positive; therefore, this circuit also produces an out-of-phase signal of opposite polarity. It is evident, then, that the common-emitter circuit always produces a polarity reversal of the input signal.

3.4.1 **Bias Circuits (Common-Emitter).** Because the common-emitter circuit is more frequently used, it has a greater variety of biasing schemes than the other configurations. The basic principles though remain the same' that is, the emitter-base junction must be forward-biased while the collector-base junction is reverse-biased, and the d-c no-signal values of base current and collector voltage specify the operating point.

Figure 3-25 illustrates a method of using two supply sources to produce a PNP emitter-base bias arrangement which series-aids the collector supply. It is evident that the emitter-base bias voltage is the voltage of the emitter-connected source, while the collector-emitter voltage is the total voltage between emitter and collector, or both sources in series. For an NPN transistor the polarities are reversed.

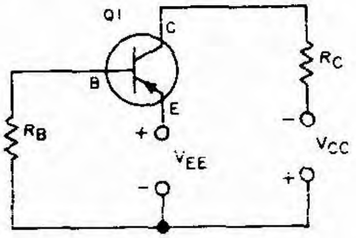


Figure 3-25.
Series-Aiding Bias Circuit, PNP

The connections for a single voltage supply source biasing arrangement are shown in figures 3-26 and 3-27.

The circuit of figure 3-26 is a voltage divider, fixed-bias arrangement, with R1 and R2 connected across the collector supply, and the base connected at their common connection. Thus the base is kept at a lower positive potential than the emitter and is therefore negative with respect to the emitter.

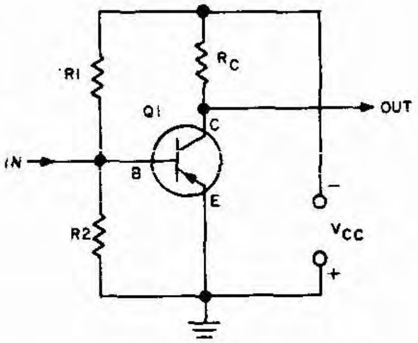


Figure 3-26.
Voltage-Divider (Fixed) Bias, CE Circuit

A self-biasing arrangement is shown in figure 3-27 which involves the internal resistance inherent in the transistor junctions and is similar to the contact bias of the electron tube. The emitter is at the highest positive potential, and the collector is at the lowest negative potential, so the emitter-collector relationships are correct. Since the base is sandwiched between these elements and floating, it is at some intermediate value, determined by the internal resistance parameters and the internal current flow. The potential between the emitter and base must be kept to a small value compared to that between the collector and base. This is achieved internally by the high-resistance action of collector-base junction and the low-resistance action of the emitter-base junction, which provide the desired voltage relationship. The supply voltage polarity is reversed for NPN transistors. By placing resistor R_B (shown dotted in figure 3-27) from base to emitter, the base is effectively biased off and less base current (I_B) flows. The collector current is reduced by $\frac{1}{1-\alpha}$ for each I_B microampere, and more economical operation is achieved by reducing the battery drain.

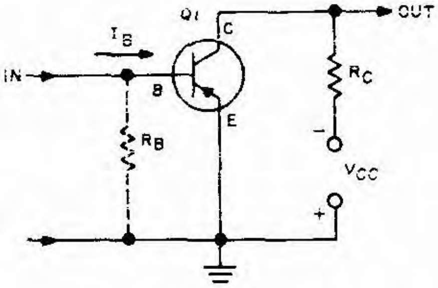


Figure 3-27.
PNP Internal Self-Bias CE Circuit

A variation of the self-bias arrangement using external resistors is shown in figure 3-28. Here the bias supply is connected in series with the emitter, and the voltage divider made up of R_B in series with the internal emitter-base resistance, determines the proper bias potentials. The voltage across R_B is subtracted from that of the bias supply to determine the actual input bias. Collector resistor R_C is chosen to produce the desired operating collector voltage. The polarity of the supply voltage is reversed for NPN transistors.

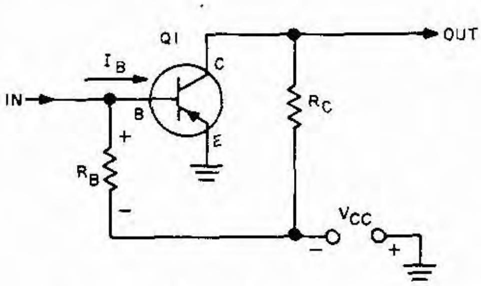


Figure 3-28.
PNP External Self-Bias CE Circuit

3.4.2 **Bias Stabilization.** The discussion of the diode stabilization circuits paragraphs 3.2.2 through 3.2.9, together with the discussion in this paragraph, covers basic thermal stabilization circuits. Since a number of variations of the circuits discussed below are possible, any other special circuits will be discussed in the sections of this manual as they appear. Stabilization as discussed here will be confined to thermal considerations; voltage stabilization is discussed in paragraph 3.2.5.

The no-signal, d-c values of collector voltage and emitter current are determined by the applied bias, which sets the operating point of the transistor. Under ideal conditions temperature would not affect the bias and the circuit would be thermally stable. Actually, however, a temperature increase causes an increase in the flow of reverse-bias collector (saturation) current (I_{CBO}), and the increase in I_{CBO} causes the temperature of the collector-base junction to increase with a consequent increase in saturation current. As this action continues, distortion occurs, and the transistor is rendered inoperative or it destroys itself. To reduce thermal instability (runaway), low values of resistance, rather than high values, must be employed in the base circuit. Refer to paragraph 3.2.3 for the discussion of a reverse-biased diode which decreases its resistance with an increase in temperature.

Another consideration is that the emitter-base junction of a transistor (or a diode) has a negative temperature coefficient. That is, as the temperature increases the emitter-base resistance decreases, causing a larger flow of collector current in addition to the flow of saturation current discussed above. To correct this condition R_E, a large-value resistor (swamping resistor), is placed in the emitter circuit, where it produces a resistance stabilizing effect (see figure 3-29). Actually, in this case the variation of emitter-base resistance with temperature is such a small portion of the over-all emitter series resistance that it exerts little effect on the over-all operation of the circuit.

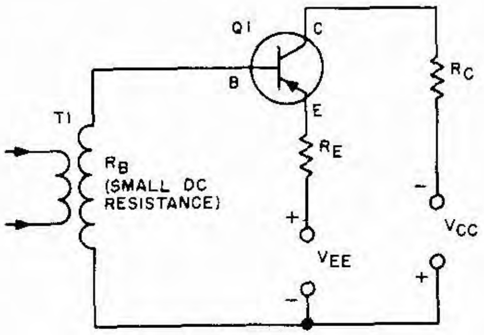


Figure 3-29.
Emitter Swamping Circuit

In figure 3-30 resistors R_1 and R_B operate as a voltage divider across the collector voltage supply source to supply negative (forward) bias to the base-emitter circuit. This arrangement allows a single voltage supply to be used for both bias and collector voltages. When the value of R_1 together with R_B in parallel is less than R_E, the effects of voltage-divider voltage stabilization and the thermal compensation provided by the swamping effect of R_E combine to offer a more stable bias circuit. Unless the proper ratio is maintained, no compensation is achieved. The stabilization is improved as the quantity

$$\frac{R_B \cdot R_1}{R_B + R_1} \Big/ R_E \text{ approaches zero.}$$

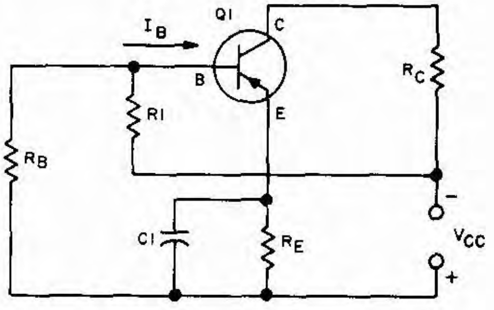


Figure 3-30.
Voltage-Divider Bias and Emitter Swamping Stabilization

Capacitor C1 bypasses R_E for signal variations; otherwise, degeneration would be produced by the swamping resistor. The capacitor is chosen to have a reactance about one-tenth that of the swamping resistor at the lowest frequency to be passed.

The circuit of figure 3-31 is a variation of that of figure 3-30; it uses only voltage-divider stabilization, the emitter swamping resistor being omitted. In this circuit, R1 and R2 form a voltage divider across the collector supply, and the effective bias is essentially the voltage existing at the junction of R1 and R2. A large resistor, R_B, is connected from the divider junction point to the base to provide a higher input resistance and avoid the shunting effect of R2 (R2 is small in value because the base-to-emitter bias is only a fraction of a volt). The stabilizing action in this circuit is provided by the voltage divider alone, since the divider is less affected by variations in element currents or voltages than are self-bias arrangements. The disadvantage of this circuit is that it consumes more d-c power because of the voltage divider; the circuit of figure 3-30 is preferable because of its increased stability.

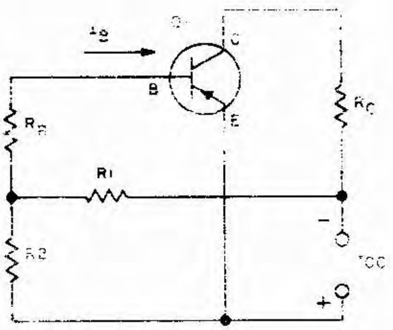


Figure 3-31.
Voltage-Divider Bias

Another method of compensating for emitter-base resistance change with temperature is to feed back an opposing voltage which is proportional to the temperature change per degree. An equivalent method is to correspondingly reduce the forward bias applied the circuit. See figure 3-32.

In figure 3-32, the circuit of part A represents both a-c and d-c feedback. When resistor R_F is divided into two parts and bypassed by capacitor C as shown in part B, the feedback loop is shunted, and only d-c bias variations affect operation.

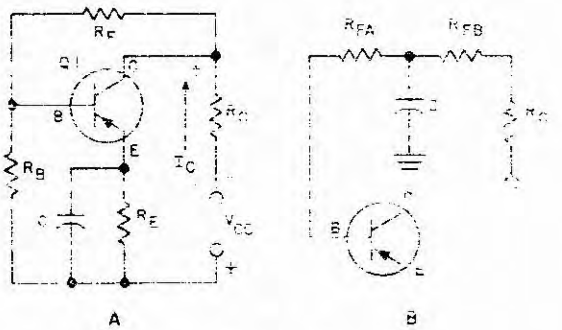


Figure 3-32.
Negative Feedback Bias Circuit

Compensation is achieved as follows: when the collector current (I_C) increases because of a temperature increase, the collector becomes less negative because of the larger positive voltage drop in resistor R_C. Since the drop across R_C opposes the initial bias, less forward bias is applied to the base through feedback resistor R_F, and the collector current automatically decreases to the original value (provided that the proper feedback ratio is maintained). There are two other types of compensation for the circuit in part A of figure 3-32; voltage-divider stabilization through R_B and emitter current feedback through R_E.

In figure 3-33, three variations of the voltage feedback circuit are shown. The circuit of part A represents voltage feedback alone. Actually for d-c biasing conditions, R_F and R_C can be considered as one resistor having a value equal to that of the external self-bias resistor, R_B, in figure 3-28. It is seen that any change of current through R_C, therefore, will either increase or decrease the bias applied through R_F.

In part B of figure 3-33, the addition of resistor R_B produces a voltage divider across the bias supply so that in addition to voltage feedback, the effect of voltage divider stability is offered. In 3-33C, current feedback through emitter resistor R_E is added, and when the resistor R_B shown dotted is also added, a combination of voltage and current feedback together with voltage-divider stabilization is obtained, and the circuit is identical to that of figure 3-32.

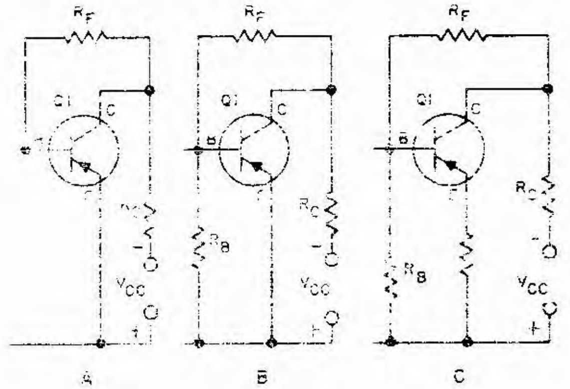


Figure 3-33.
Voltage Feedback Circuits

In the above discussion of stabilization, it has been assumed that only the d-c, no-signal operation is of interest, [illegible] that signal operation does not materially affect the bias.

Figure 3-34 shows a voltage-divider base-bias arrangement using a thermistor to compensate for emitter current changes with temperature. When the emitter current tends to rise with temperature, the thermistor, having a negative temperature coefficient of resistance, reduces in value as the temperature increases. This reduction in resistance increases the current flow from the V_{CC} supply and causes an increased voltage drop across R1. The base bias is reduced correspondingly, lowering the emitter current and compensating for the temperature change. Since the thermistor is constructed of a material different from that of the transistor, it does not change resistance in exact proportion to the emitter current change; therefore, [illegible] compensation occurs at only a few points over the operable range. In this respect, semiconductor diodes provide much more ideal compensation. This method of stabilization is identical in concept to that described in paragraph 3.2.2.

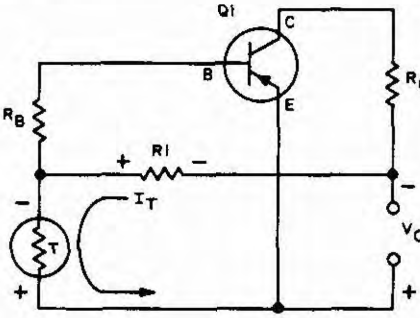


Figure 3-34.
Thermistor Base-Bias Compensating Circuit

A number of thermistor compensation circuits have been developed, but they all use the same principle of changing bias inversely with temperature to compensate for the change. Figure 3-35 shows an emitter bias compensator, in which the base bias is provided by a voltage divider consisting of R1 and R2, and compensating emitter bias is provided by R3 and the thermistor. The drop across R3 applies a reverse bias to the emitter as the temperature increases, reducing the emitter current correspondingly.

Normally ideal thermal compensation, as well as a reduction of the number of parts required, can be achieved by the use of crossconnected transistors arranged so that the element voltages or currents of one transistor compensate for thermal variations by producing correction voltage or currents in the other, while both transistors operate as amplifiers. For example, it is possible to use the variations of the emitter-base junction resistance with temperature of one transistor to control the emitter-base bias of a second transistor, or to stabilize the emitter-collector current of one transistor with the stabilized emitter-collector current of another transistor. Since these circuit arrangements become rather involved, they will be discussed as special circuits at appropriate points in other sections of this technical manual. However, since thermal compensation is an important part of the d-c amplifier, its use in a two-stage stabilized unit is discussed below. See figure 3-36.

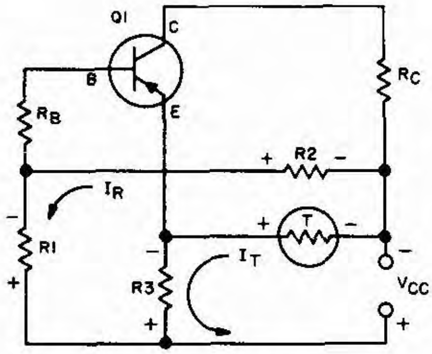


Figure 3-35.
Thermistor Emitter-Bias Compensating Circuit

In the circuit of figure 3-36, an increase in collector current produced by a temperature rise in transistor Q1 reduces the forward bias of transistor Q2. Transistor Q1 is connected as a CB amplifier, which basically has an ideal stability factor. Nevertheless, a very slight variation of collector current will occur with temperature variation. This slight temperature variation is the result of reverse leakage current (I_{CBO}) caused by the internal flow of minority carriers (electrons) from collector to base. The reverse leakage current is substantially independent of collector voltage and mainly dependent upon temperature, being constant for a specific temperature, and increasing with temperature. The effects of temperature-caused variations of the base-emitter resistance of Q1 are minimized by the relatively large swamping resistance offered by R_E, and will not have any appreciable effect on the collector current, I_C. Thus, while the emitter junction is essentially stabilized, the collector junction is not. Although the collector junction is reverse-biased for normal forward current, this bias is actually a forward bias for reverse current. Therefore, the reverse current flow can reach values as high as 5 milliamperes. Although this high reverse current does not lead to thermal runaway, it does change the parameters, causing a change in the operating point and resulting in improper circuit operation. In addition, since the two transistors are direct-coupled, current changes in Q1 will be amplified by Q2, causing a much greater shift. While the total base current, I_B, is the net result of I_{CBO} plus the base-emitter current, the current of interest is the very small (incremental) changes of reverse leakage current, ΔI_{CBO}, with temperature; for this explanation then, the absolute values of base current may be disregarded.

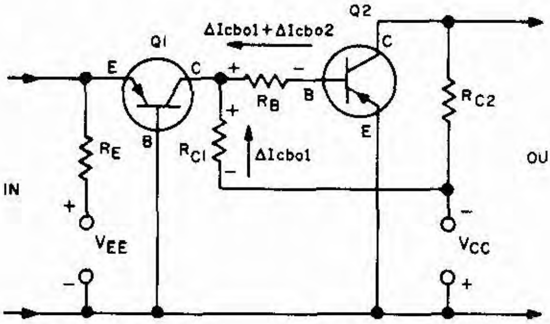


Figure 3-36.
Temperature-Stabilized D-C Amplifier

Referring again to figure 3-36, it is evident that the main current path for I_{CBO1} is through R_{C1}, the collector-base junction of Q1, and V_{cc}. An additional current path which is also the path of I_{CBO2}, is provided through R_{C2}, the collector-base junction of Q2, R_B, the collector-base junction of Q1, and V_{CC}. Any incremental change of I_{CBO1} and of $I_{CBO2} + I_{CBO1}$ will produce an incremental change in the voltage drops across R_{C1} and R_B, respectively. The change in voltage across R_{C1} will be in a direction to decrease the forward bias of Q2 (see polarity indicated in figure 3-36), while the change in voltage across R_B will be in a direction which increases the forward bias of Q2. If the values of R_{C1} and R_B are chosen so that the incremental change of voltage across R_{C1} is slightly greater than the incremental change across R_B, then a thermally caused increase of collector current will be compensated for by a reduction of the forward bias of transistor Q2.

The discussion above considers only the very small changes in current produced by temperature variation in the collector junction of Q1; it does not consider the static operating conditions nor signal variations. Normally, Q1 operates as a conventional CB amplifier thermally stabilized by series emitter swamping resistor R_E, and biased by supply V_{EE}. The input signal is applied across R_E and amplified by Q1, appearing across R_{C1} as a direct-coupled input to the base of Q2, a conventional CE amplifier. The output of the two stages is developed across collector load resistor R_{C2}. The collector supply for both stages is taken from the single V_{CC} source. For a complete discussion of d-c amplifiers, see D-C Amplifier Circuits in Section 6 of this technical manual.

3.5 TRIODE COMMON-COLLECTOR CIRCUITS.

In the common-collector circuit, the input signal is injected into the base, and the output signal is taken from the emitter, with the collector being common to both circuits. The common-collector circuit is equivalent to the electron tube cathode-follower circuit. It has a high input resistance (2K to 500K) and a low output resistance (50 to 1500 ohms). It has a current gain similar to that of the common-emitter circuit, but a lower power gain than either the CB or CE circuits (10 to 20db). The output signal is in phase and of the same polarity as the input signal, and the voltage gain is always less than unity. This circuit is used mostly for impedance-matching and isolation of output stages; thus its function is similar to that of the electron tube cathode follower. It has the ability to pass signals in either direction (bilateral operation), a feature which is particularly useful in switching circuitry.

Figures 3-37 and 3-38 show the PNP and NPN common-collector connections, together with polarities and external current paths. Base-emitter bias and collector-base junction voltages are obtained from separate supplies so that two voltage sources are required. Signal voltage supply circuits are discussed under Biasing Arrangements (paragraph 3.5.1). As in the CE and CB discussions, the PNP and NPN circuits are shown together for ease of comparison of operation. Both circuits are forward-biased from emitter to base and reverse-biased from collector to emitter. The currents and polarities in both circuits are opposite because of the different types of germanium used.

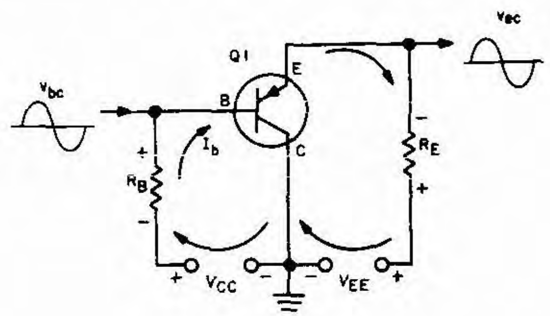


Figure 3-37.
PNP Common-Collector Circuit

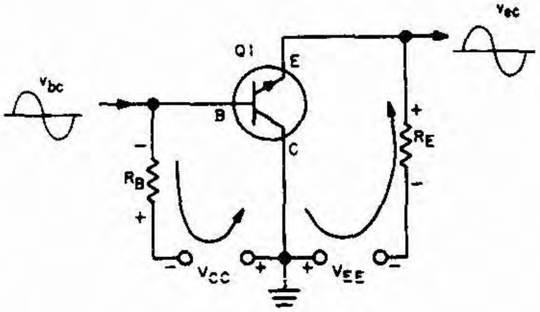


Figure 3-38.
NPN Common-Collector Circuit

The current flow and transistor action of the CC circuit are as explained for the common-base connection, but the current gain is not based on the emitter-to-collector current ratio, alpha (α). Instead, it is based on the emitter-to-base current ratio, gamma (γ), because the output is taken from the emitter circuit. Since a small change in base current controls a large change in emitter (and collector) current, it is still possible to obtain considerable current gain. However, since the emitter current gain is offset by the low output resistance, the voltage gain is always less than unity, exactly as in the electron tube cathode-follower circuit.

Common-collector current gain, gamma (γ), is defined as $\gamma = \frac{\Delta I_E}{\Delta I_B}$ with collector voltage constant, and is related to collector-to-emitter current gain, alpha (α), of the CB circuit by the formula: $\gamma = \frac{1}{1 - \alpha}$

In the PNP circuit of figure 3-37, a small input signal (v_{bc}) is shown applied between base and collector of the transistor. The circuit is biased to operate over the linear portion of its dynamic transfer characteristic, and rests in a quiescent state determined by the static d-c potentials applied (similar to electron tube class A operation). Assuming a sine-wave input, it is apparent that as v_{bc} increases to its maximum positive value the forward bias is reduced. Thus, the emitter and collector currents are reduced, producing a decreased voltage drop across the emitter output resistor, R_E, and producing a positive-going output voltage. Conversely, as the input signal swings negative the forward bias is increased and more emitter-collector current flows. The increased voltage drop across R_E is in the negative-going direction, and the output signal reaches a negative maximum. Since the output signal varies in the same direction as the input voltage, both reaching their positive and negative maximums simultaneously, it is evident that these signals are in phase. Therefore, the output of the common-collector circuit is of the same phase and polarity as the input signal and no phase reversal is produced, just as in common-base operation.

The functioning of the NPN circuit shown in figure 3-38 is similar to, but the inverse of, the functioning of the PNP circuit. When the input signal is positive, the forward bias is increased (its polarity is opposite to that of the PNP circuit bias), and the emitter current increases and produces a positive-going output across R_E. On the negative input cycle, the emitter output is negative; therefore, the output of this circuit is also of the same phase and polarity as the input signal. Thus, the common-collector circuit always produces an in-phase output signal, regardless of the type of transistor used.

3.5.1 **Bias (Common-Collector).** Common-collector biasing schemes are similar to those of the CB and CE configurations, and the basic principles are the same. That is, the base-emitter junction is forward-biased, the base-collector junction is reverse-biased, and the d-c, no signal values of base current and collector voltage specify the operating point.

Figure 3-39 shows a series-aiding bias arrangement in which two voltage supplies are used. This arrangement is similar to that for the CE circuit shown in figure 3-25, and the operation is also similar. In figure 3-40 a single voltage source bias arrangement is shown. Note that the flow of current through R_E is in a direction which produces a voltage drop that opposes the applied bias and collector voltage. The actual bias is the algebraic sum of the two voltages. Polarities and current flow are opposite for NPN circuits.

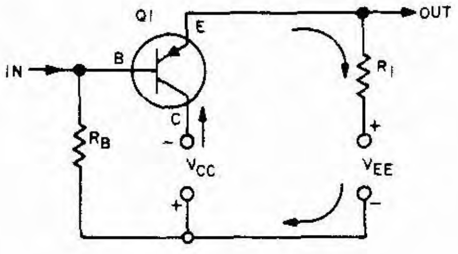


Figure 3-39.
Series-Aiding Bias Circuit

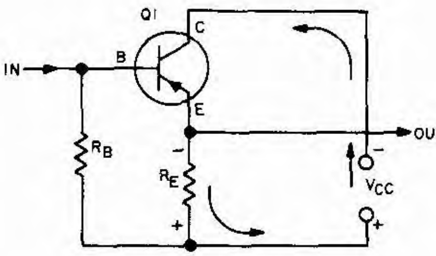


Figure 3-40.
Self-Bias Circuit

3.6 TETRODE, POWER, AND SPECIAL PURPOSE CIRCUITS

In this paragraph, various types of tetrodes, power considerations and power transistors, and special purpose transistors and their basic circuitry are presented. While it is possible that many of the special purpose items will not be encountered in Naval equipment, information on these items has been included because of the rapid advances in the state of the art. It is anticipated that, in later revisions to this publication, the information concerning those items which have the greatest application will be expanded and the information concerning those which have little use will be deleted or will indicate their limited application.

3.6.1 **Tetrodes.** The addition of a fourth element to a transistor produces a tetrode transistor. Both junction and point-contact transistors can be formed into tetrodes. In the junction transistor the fourth electrode is essentially another base electrode (B2), whereas in the point-contact type it is essentially another emitter (E2).

3.6.1.1 **Junction Tetrode (Double-Based Transistor).** The junction tetrode consists of a conventional junction transistor with another base electrode (B2) added on the side opposite the B1 connection. The addition of proper bias between the base electrodes decreases the collector capacitance and the base resistance, and thus improves the high-frequency response. As compared to a conventional junction transistor which has a high-frequency cutoff of approximately 30 mc, a tetrode will have good response up to 200 mc.

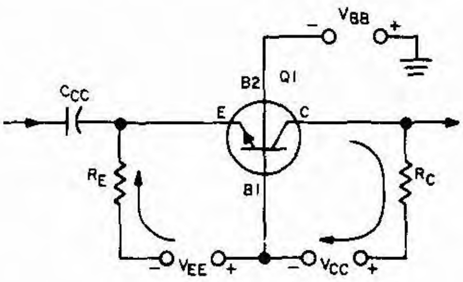


Figure 3-41.
Junction Tetrode Circuit

Figure 3-41 shows the basic tetrode circuit. The NPN transistor is used as an illustration for ease of explaining the operation. The tetrode is connected in the same manner as a triode, with forward emitter-junction bias and reverse collector-junction bias. Between base 1 and base 2, however, *a large value of voltage is* connected (on the order of volts as compared with tenths of a volt), usually about 6 volts. Since the base material is a semiconductor and the points of application are on opposite sides, there is a definite resistivity between B1 and B2 which produces a uniform drop across the base (if the base were ohmic, a short circuit would ensue). Since the applied base bias is *negative* and large, it blocks electron current flow through all parts of the base region, except for a small volume near the B1 connection. Thus the current is restricted to the small controlled space around the B1 terminal (as shown in figure 3-42). This effective reduction of the volume of the base region reduces the resistance to base current and also the base-collector capacitance through the collector junction. The over-all result is to improve the high-frequency response of the transistor.

Since the entire base area is not available for current passage, the over-all current gain of the tetrode transistor is less than that of the triode type. At the higher frequencies (above 30 mc), the performance of the tetrode is superior to that of the triode, and the improvement in frequency response offsets the slight loss of stage gain.

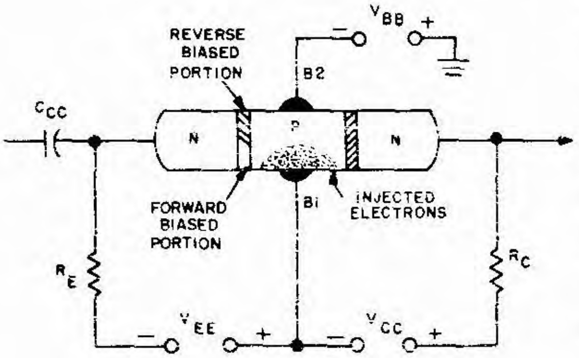


Figure 3-42.
Tetrode Action

3.6.1.2 Spacistor Tetrode. The spacistor tetrode provides a high input impedance (up to 30 megohms) and a high output impedance (on the order of megohms). Its reduced input and output capacitance and a very short transit time. It is capable of power amplification of 40 db and a voltage gain as high as 3000. Operation is possible up to several thousand megacycles, as compared to an upper limit of 300 to 400 megacycles for the other types of high-frequency transistors.

Figure 3-43 shows the basic spacistor circuit. The spacistor basic PN junction is reverse-biased, and is manufactured so that the P-region contains a smaller depletion area than that of the N-region. The P-region acts as the base and the N-region as the collector. By the use of a large reverse bias (on the order of 100 volts), a strong electric field is produced across the junction, but because of the reverse bias only a small reverse current flows. An emitter connection (called the **injector**) is made on the large N-depletion area of the junction, and it is *also reverse*-biased. The emitter bias is chosen so that it is less than the collector bias, *and an electron* current flows from the emitter to the collector electrode. The magnitude of this electron current is proportional to the difference of voltage between the emitter and collector, and, because of the intense electric field through which it flows, the transit time is very small. A fourth electrode (called the **modulator**) is connected to a small piece of P-type material which forms another PN junction with the large N-depletion area, and is located very close to the emitter connection. While the modulator is positively biased, it is less than the collector bias, so that the modulator is effectively reverse-biased with respect to the collector. Therefore, practically no modulator current flows, and a high input resistance (serveral megohms) is obtained. When the modulator bias is varied, as by an input signal, the current flow between the emitter and collector is varied accordingly. Thus the modulator electrode acts similarly to the grid of an electron tube, and the emitter acts similarly to the cathode.

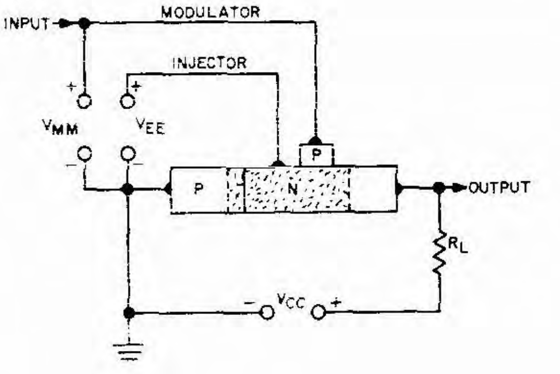


Figure 3-43.
Spacistor Circuit

Because the emitter and modulator electrodes cover only a small area, the input capacitance is low. The collector-to-base (output) capacitance is also low because of the effect of the large N-depletion area, which reduces the total capacitance by effectively providing greater separation between the collector and base. As a result of the reduction of capacitance and the *short transit time* from emitter to collector, the high-frequency response is greatly improved.

3.6.1.3 **Crystal-Mixer Tetrode.** The crystal-mixer tetrode is a special four-element, point-contact device specially developed and constructed for use as a mixer. Although it has the same number of elements as the point-contact tetrode, it differs in construction and operation. This unit has approximately the same conversion gain as a conventional vacuum-tube mixer, but it operates better at high frequencies. It is superior to the conventional crystal-diode or triode mixer.

The mixer tetrode is constructed with two emitters and one collector as shown in figure 3-44. The emitters are located at equal distances on opposite sides of the collector so that both emitters have the same effect on the collector. The collector output is equal to the sum of the outputs it would have if each emitter were operated separately, provided that the collector is not given into saturation.

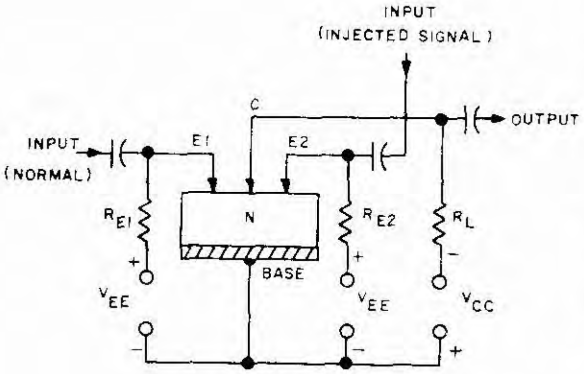


Figure 3-44.
Crystal Mixer Tetrode

3.6.1.4 **Point-Contact Tetrode.** The conventional point-contact four-element (tetrode) transistor differs from the specially constructed crystal-mixer tetrode discussed above in both operation and construction. Emitter No. 2 in the point-contact tetrode is spaced a greater distance from the collector than emitter No. 1. Both emitters are forward-biased and the collector is reverse-biased, as in a junction transistor. When emitter No. 1 is connected and emitter No. 2 is left open, the transistor operates like a conventional triode with a current gain of 1.5 to 3. When emitter No. 2 alone is used, because of the greater spacing from the collector, the current gain is small (about 0.2). When both emitters are used together, the current gain of emitter No. 1 is enhanced considerably, from 4 to 8, which is roughly 2-1/2 times the original value. Emitter No. 2 may be used as an input control element, if desired, and the current gain will vary according to the input signal.

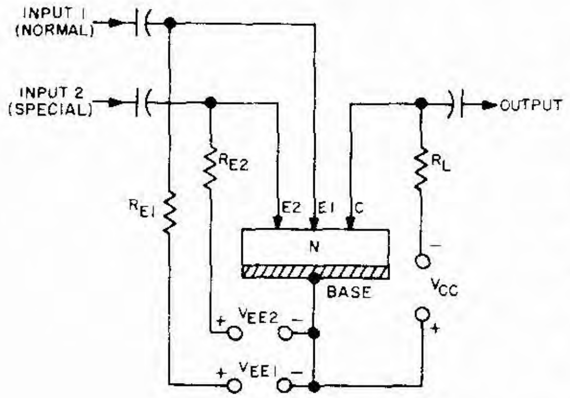


Figure 3-45.
Basic Point-Contact Tetrode Circuit

The increase in current gain when both emitters are properly biased and operated simultaneously results from the fact that the concentration of holes injected by the first emitter attracts electrons away from the base region, between the second emitter and the collector. This reduces the possibility of electron-hole recombinations and enables many of the second-emitter injected holes to reach the collector, thereby increasing the flow of collector current.

3.6.2 **Power Transistors and Considerations.** Power transistors use special construction and design considerations to achieve rated output. The conventional transistor is usually operated at low voltages and low values of current, whereas the power transistor is operated at relatively high voltages and high currents to produce power outputs. The classification of power transistors, however, is somewhat arbitrary, and does not include the same order of values as used for electron tubes. For example, transistors of from 2 to 50 milliwatts are classed as non-power types, from 50 milliwatts to 500 milliwatts (1/2 watt) as low-power types, from 500 to 1000 milliwatts (1-watt) as medium-power types, and all over 1 watt as high-power types.

Because the transistor is a nonlinear device, it does not respond to large signals (where variations in collector voltage and current are a significant fraction of the total range of of operation) in the same manner as it responds to small signals; hence, small-signal parameters are used to define non-power (linear) operation, and large-signal parameters are used to define power (nonlinear) operation. Small signals can arbitrarily be defined as those which are less than 1 volt, and large signals as those which are greater than 1 volt. Since it is possible to have a small signal driving a power amplifier, it can be seen that sometimes either parameter can be used to predict circuit performance with good approximations.

While both point-contact and junction transistors can be used as power amplifiers, the point-contact type is usually limited to values not greater than 1 watt. This limitation is due to the fact that the point contact is unable to carry a heavy current without excessive heating and consequent damage to the transistor.

The power transistor must be able to dissipate the internally generated heat while operating at the increased temperature resulting from its own heat. It must also be able to operate at high currents and voltages without breaking down or causing excessive non-linearity (distortion). Since the junction transistor does not concentrate the heat around a point source, but spreads it throughout the junction, it has a definite advantage for power use. Both NPN and PNP junctions may be used, and they operate fundamentally the same regardless of the method of manufacture (grown junction, diffused-junction, alloy junction, etc).

The current rating of a power transistor is the maximum collector current that can safely be carried by the transistor, without exceeding the power rating, causing internal damage, or producing an excessive loss of current gain at higher emitter currents.

The voltage rating of a power transistor is the voltage for which a specific leakage current occurs for a specific circuit configuration and operating current.

The power rating of the transistor is the maximum permissible power which may be safely dissipated by the unit without exceeding the maximum junction temperature and causing damage instantaneously or over a period of time.

The saturation voltage is the value below which the collector voltage cannot be further reduced, even by increasing the input current.

The thermal resistance of a transistor is the ratio of the difference in actual power rating with respect to the rise of temperature of the transistor. It is commonly expressed in degrees centigrade/milliwatt, or watt.

Thermal runaway is the condition whereby a small increase of collector leakage current occurs, because of an increase in the ambient temperature of the junction, and causes an increase in junction temperature, which, in turn, causes another increase of leakage current. This action builds up in an exponential manner, until complete thermal runaway occurs, permanently damaging the transistor. The power transistor is most sensitive to thermal runaway when it is operating near its maximum collector dissipation value and no thermal compensation circuitry is used. See paragraph 3-2 for a discussion of thermal stabilization circuits.

Either of two general methods is used to provide for heat dissipation and improve the power-handling capability of a power transistor—the use of a so-called infinite heat sink or liquid cooling of the transistor. The heat sink consists of an integrally constructed base mount, usually made of copper, plus a physical connection from the transistor shell to the collector. The base mount usually is directly connected to the chassis, which serves as the infinite heat sink. In cases involving high power, a special heat sink is provided, including fins for heat radiation into the surrounding air, and sometimes the transistor shell is provided with fins. Where the transistor cannot be grounded directly to the chassis, it is usually insulated from the chassis by a thin mica sheet. Liquid cooling is accomplished by sealing a cooling agent in the transistor case and allowing liquid convection to the metal case to provide the transfer of the heat to the air. In extreme cases, a circulating system is provided and the transistor is immersed in the cooling liquid similar to a water-cooled electron tube.

Medium power transistors usually employ Class A amplification, whereas higher-powered units use Class AB or B amplification because of its increased efficiency. See paragraph 3-7 for an explanation of classes of operation. Class C is seldom employed because of the distortion produced, except for transmitting circuits which use the flywheel effect of the tank circuit to overcome the distortion, as in electron tube operation.

Complementary-symmetry circuits are also used to provide additional power output with a reduction in the number of components required and the over-all circuit cost. Figure 3-46 shows a typical complementary-symmetry, push-pull circuit using an NPN transistor for one half of the circuit and a PNP transistor for the other half. Since the polarities and currents in these transistors are opposite and equal (for matched units), it is evident that one transistor works on one half of the input cycle and that the other transistor works on the other half of the cycle. The out-of-phase outputs are added at the proper time (in-phase) to produce an output from the load resistor which is equal to the combined effect of the collector currents.

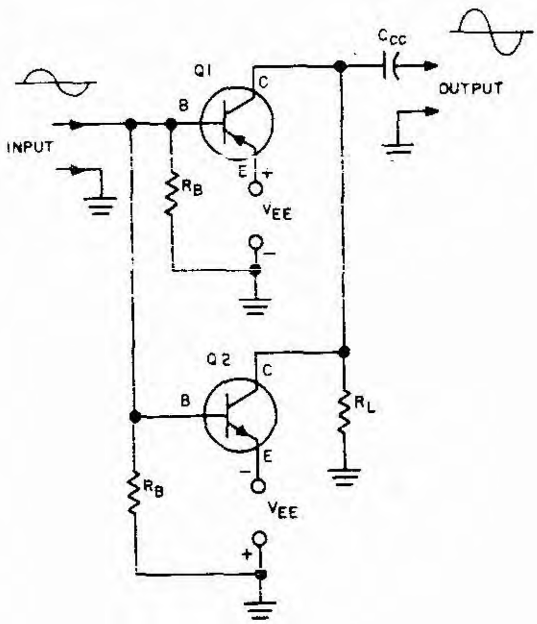


Figure 3-46.
Complementary-Symmetry Push-Pull Circuit

Power transistors are used in audio, video, and r-f applications where outputs are required to operate electromagnetic or electrostatic output devices, in switching circuits, and in power circuits as substitutes for relays. They are especially advantageous when used as power converters to supply high voltage dc or ac from low-voltage d-c supplies to mobile, aircraft, or marine equipment. They provide increased efficiency of conversion and eliminate maintenance, mechanical, and interference problems encountered with vibrators, generators, or dynamotors.

3.6.3 **Special-Purpose Transistors and Circuits.** The special-purpose transistors and circuits discussed in the following paragraphs are representative of the present state of the art. Discussion is limited to the salient points considered necessary for the user of this technical manual in the event some of the special-purpose devices are encountered. Emphasis has been placed on presenting functionally different types, rather than manufacturer's claims for proprietary construction and materials.

3.6.3.1 **PNPN Triode (Hook Collector).** The PNPN triode transistor is a four-layer semiconductor device with three junctions, and exhibits a high current gain in the common-base connection. This transistor has been used more in switching circuits than in other applications because of its higher noise level, leakage currents, and limited frequency response.

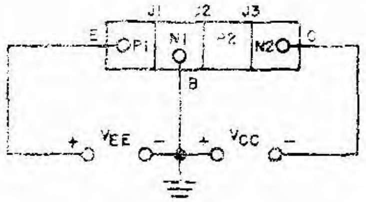


Figure 3-47.
PNPN Triode

Figure 3-47 shows typical bias connections and the arrangement of the layers for this transistor. The transistor essentially consists of a conventional triode two-junction PNP unit with an added N-layer at the collector end. The emitter P-area (J1) and the collector N-area (J3) are forward-biased, the base N-area is grounded, and the second P-area is left floating between the base and collector. As a result, the middle NP junction (J2) is effectively reverse-biased. Hence, the P1, N1, and P2 sections are comparable in action to an ordinary triode transistor. The floating P2 region, however, causes a few unusual effects. The potential hill due to the depletion region at junction J3 has a great retarding effect on the holes from the emitter, causing a "space charge" of holes to build up on the P2 side of J3. The result of this is a reduction of the J3 junction resistance, thereby allowing increased electron movement from the N2 section into the P2 section. The complete path for this current (except the small part which recombines with some of the holes) is from the negative terminal of Vcc to N2, P2, N1, base lead, and back to Vcc.

Large values of current amplification may be obtained since a relatively small accumulation of holes in the P2 section gives rise to a much larger number of electrons from the N2 region. Control of the collector current is obtained since it is determined by the number of "trapped" holes in the P2 region, which in turn is controlled by the emitter current; the emitter current, of course, is determined by the input signal to the transistor. The name **hook collector** is derived from the transistor energy diagram, which resembles a hook. NPNP transistors have also been manufactured.

The hook transistor operates as an amplifier and should not be confused with the PNPN switch or with the silicon-controlled rectifier which operates similarly to a thyratron tube. In the PNPN triode switch or controlled rectifier, the second P-area is always the gated region, whereas in the hook collector it is left floating.

3.5.3.2 **Unipolar (Field-Effect Transistor).** The unipolar transistor uses a unique construction to utilize the effect of an electric field to control the passage of current carriers through the basic semiconductor bar. It offers a high input resistance (about one megohm) with a relatively high output resistance and good high-frequency response.

Figure 3-48 shows the basic unipolar circuit. The basic N-germanium bar has a potential applied between the source-end and drain end in series with load resistor R_L. The circular P-alloy (called the gate) which encircles the bar is reverse-biased to the source-end of the bar. The field effect produced by the reverse bias depletes the area beneath the P-electrode of current carriers. The input signal varies the basic conductivity of the N-bar and effectively controls the current supplied by the source. Thus, current flow through the load resistor produces a corresponding but amplified output voltage.

Since the predominant carrier in N-type material is electrons and they are controlled by the electric field between the gate and source-end of the bar, and since holes are not involved, the term **unipolar** was derived to indicate that only one carrier is involved.

The advantage of this type of device is that a small signal controls a much greater output, which is limited essentially only by the size of the source supply and the resistivity of the basic bar material. Applications are somewhat limited by a rather high noise figure.

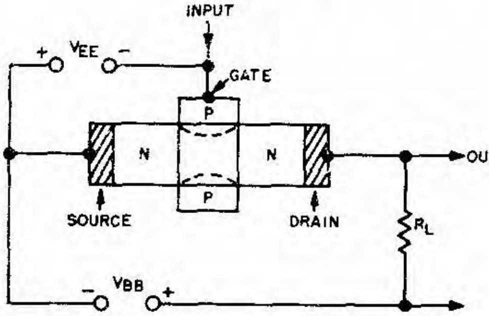


Figure 3-48.
Unipolar (Field-Effect) Transistor

3.6.3.3 **Unijunction Transistor (Double-Based Diode.)** The silicon unijunction transistor is a three-terminal semiconductor device, sometimes called a double-based diode, which is unique in that it can be triggered on by, or an output can be taken from, each of the three terminals. Once the unit is triggered, the emitter current increases

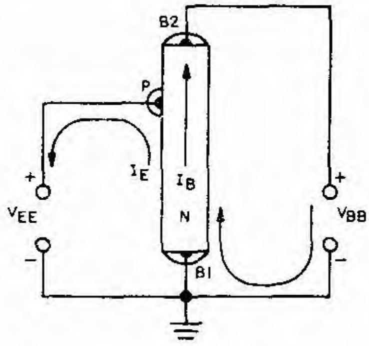


Figure 3-49.
Unijunction Transistor

regeneratively until it is limited by the power supply, thus, the action of the transistor is similar to that of the gas thyratron tube. It can be employed in a variety of circuits, but it finds its greatest usefulness in the switching and pulse fields.

Figure 3-49 shows the basic unijunction bias connections. The unijunction transistor consists of an N-type silicon bar with two ohmic base contacts at the ends and a P-type emitter (PN junction) near base No. 2 (B2). A base biasing potential, applied between the two base contacts, establishes a voltage gradient along the bar; the emitter is located nearer B2, so that more than half of the base bias along the bar appears between the emitter and base No. 1. If an external potential is applied between base No. 1 and the emitter greater than the internal voltage gradient between the same points, the junction is forward-biased; if the external potential is less than the internal voltage a reverse-bias is produced. Normally, reverse bias is applied between the emitter and base No. 1 (B1) so that in the off condition the emitter current is at cutoff. When a positive trigger pulse of voltage is applied to the emitter (or a negative trigger to B1 or B2), the emitter is forward-biased. An increased hole current causes a reduction in the resistance, and a reduction in the internal voltage drop between the emitter and base one. As a result, the emitter current increases regeneratively until it is limited by the power supply. This action is spoken of as conductivity modulation of the interbase current. The unijunction is returned to the off state by a negative trigger at the emitter. A typical circuit showing input and output points is shown in figure 3-50.

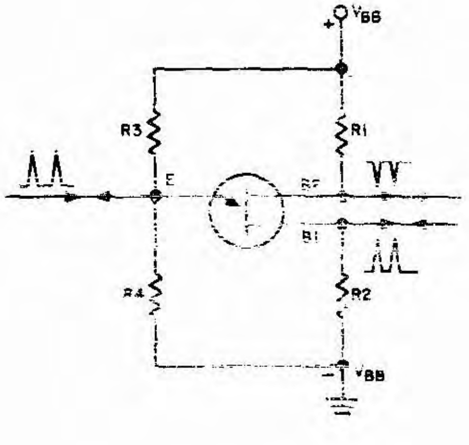


Figure 3-50.
Unijunction Input and Output Points

Resistors R3 and R4 in figure 3-50 form a bias voltage divider for the emitter, which normally holds the circuit at cutoff. An input applied across R4 will produce a relatively high-voltage output across R1 and a relatively low-voltage output across R2. An input across R2 will produce an output across R1 and R4. Likewise, an input applied across R1 will produce outputs across R2 and R4.

Since the silicon base bar is temperature sensitive, this device can also be used in temperature-sensitive control applications by utilizing the variation of resistivity of the base with temperature. The interbase resistance increases with temperature at a practically constant rate. A typical value is [illegible] percent per degree C.

3.6.3.4 **Surface-Barrier Transistor.** The surface-barrier transistor is a specially constructed germanium device which is quite similar to the PNP junction transistor in operation. It has a lower noise figure an excellent high-frequency response (about [illegible] times that of the alloy junction), and will operate at extremely low currents.

Manufacturing is by the chemical (electrolytic) etching process, which produces cavities on opposite sides of a thin, pure N-type germanium wafer. Then the same process is reversed to electro-deposit metallic electrodes in these cavities. The thin base region (about .0002 inch) helps produce the excellent high-frequency response, and the etching and plating processes provide excellent control of transistor characteristics. Junction alloy transistors use ohmic, non-rectifying contacts to the semiconductor. The surface-barrier transistor uses metal electrodes which are excellent rectifiers and operate efficiently as collectors or emitters for the minority carriers. The collector electrode is made larger than the emitter electrode to provide a high current gain. Operation depends upon the surface-barrier effect; that is, the electric field which exists at the surface of the germanium excludes both holes and electrons from a thin region near the surface. The addition of the metal electrode to the germanium produces a concentration of holes (minority carriers) directly under the surface. Forward-biasing of the emitter attracts the holes to the region between the electrodes, and reverse-biasing of the collector provides a means of collecting the holes as they diffuse across the thin base region.

Figure 3-51 shows a typical CB amplifier arrangement using SB transistor and indicates the operating voltages and gains. A voltage gain of 190 is realized, with an overall power gain of [illegible].

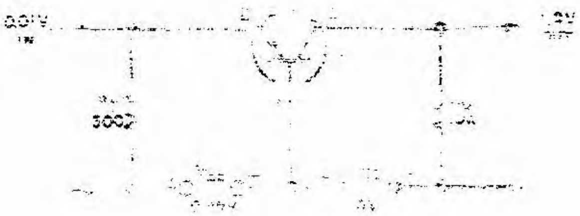


Figure 3-51.
Typical SB, Common-Base Circuit

3.6.3.5 [illegible] PNP Transistors. These transistors [illegible] type transistor. Their basic difference is in construction; an intrinsic area is inserted between the base and collector areas. The over-all result is to reduce the base-to-collector capacitance and permit higher-frequency operation (on the order of 900 mc maximum).

3.6.3.6 **Drift Transistor.** This transistor uses a construction which gradually changes the resistivity of the semiconductor from a highly conducting material at the emitter to a more resistive material (nearly pure germanium) at the collector. When a potential is applied between the emitter and collector, an electric field effect is produced which causes the internal carriers to drift across the junctions at high velocity, instead of relying upon diffusion effects. Thus, the drift transistor can be operated at higher frequencies than the normal junction transistor. Figure 3-52 shows the impurity concentration gradient for a typical drift transistor. It is also representative of the resistivity gradient and the electric field produced, which vary in the same manner. Because of the variable impurity distribution in the base region, less of the depletion area extends into the base, and the total effect is that of widening the depletion area. Shorting of the transistor by punch-through effect as the collector voltage is raised is eliminated, because the depletion area will gradually extend into the base and collector areas, and be blocked by the heavy impurity concentration near the emitter region, limiting further spread to the collector region. At the same time, the strong electric field produced by the varying resistivity gradient from one end to the other of the transistor causes an attraction for the minority carriers and thus urges injected holes across the base region in the same direction as the diffusion currents. The result is to provide a shorter transit time for the injected carriers than would normally occur for the same base width if only the diffusion process were acting as a transport medium. With a shorter transit time, the high-frequency response is extended above that of the normal transistor.

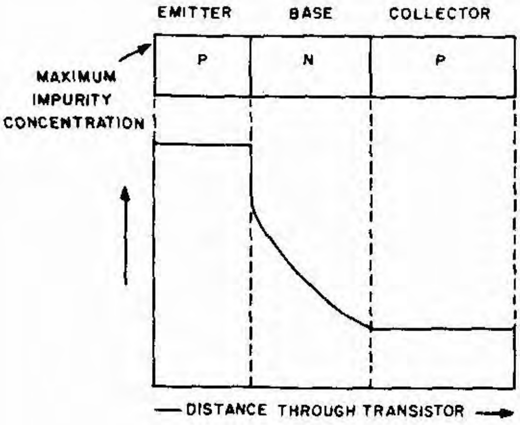


Figure 3-52. Drift Transistor Construction

3.6.3.7 **Diffused-Base (MESA) Transistor.** The diffused-base transistor utilizes manufacturing processes to produce a physically thin diffused-base alloy on a basic germanium bar. Thus the collector-base capacitance is reduced, and the slowness of the diffusion process through the base region is minimized. Better high-frequency response results.

3.6.3.8 **Silicon Controlled Rectifier.** The silicon controlled rectifier is a silicon 3-junction, 3-terminal device. It is the semiconductor equivalent of the thyratron tube. It can be either a PNPN or a NPNP unit. In the PNPN unit, the anode is the P terminal and the cathode is the N terminal. The internal N region is not connected externally, but floats between the anode and the second P region or external gate terminal (see figure 3-53).

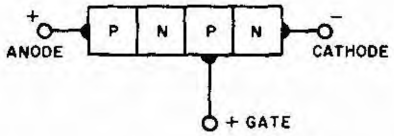


Figure 3-53. Silicon Controlled Rectifier

The controlled rectifier is connected as a conventional rectifier with anode to positive and cathode to negative. In this condition, both end junctions are forward-biased, but the middle junction is reversed-biased, and only a small reverse current flows (conduction is effectively blocked). When a positive gate is applied to the gate electrode, the middle junction is forward-biased and heavy current flows; or, when a specific blocking voltage is exceeded, the silicon controlled rectifier also breaks down and operates exactly as if gated. Once conduction is initiated, it continues until either the current or the voltage drops below a small holding value or until the external circuit is interrupted. Figure 3-54 shows a typical circuit utilizing two controlled rectifiers in a full-wave rectifier circuit. The output level is determined by the control circuit. These units are operable over ranges of from 20 to 600 volts blocking and currents of 1 to over 100 amperes, under control of gates from less than 1 volt at 1/4 milliampere to 3 to 4 volts at 10 milliamperes, with turn-on time of 1 to 5 microseconds and turn-off times of 10 to 20 microseconds.

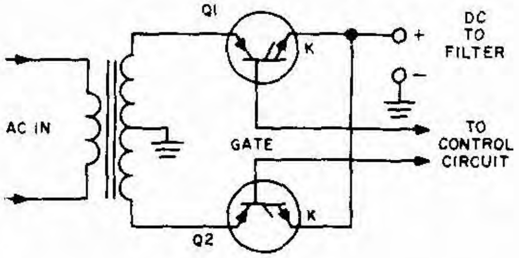


Figure 3-54. Controlled Rectifier Circuit

3.6.3.9 **Phototransistors.** The phototransistor is a combination of two junction diodes arranged as a conventional transistor, for *example, in a PNP configuration,* with only the two end leads brought out. The mechanical *arrangement* is such that light is focussed on either one or both junctions to vary the conductivity of the unit. This unit is identical in operation to the photodiode, except that it is much more sensitive, from 50 to 500 times, because of transistor action.

It is evident from figure 3-55 that the connections are the same as for the photodiode and that the emitter junction is reverse-biased, with the collector junction forward biased and the base floating. Biasing is achieved through the internal resistance of the junctions. The *emitter is more* negative than the collector, and the base floats somewhere in between, *being at a lower* positive potential than the collector, so that it is effectively reverse-biased (assuming a PNP unit). Since the base is truly floating (it is not connected to any input or returned to ground except through the internal base-emitter resistance), it is extremely susceptible to any light impinging on the junction. Variations in light intensity cause the junction conductivity to vary, and thus act similarly to an input signal applied to a conventional emitter-base junction. Since the collector junction is forward-biased, the changing emitter conductance causes corresponding and amplified changes of collector current, developing an output across load *resistor* R_L.

Because of the large collector current control offered by the phototransistor, *it may be used directly* to control a relay connected to turn power on or off, or to operate a switching circuit; see figure 3-56.

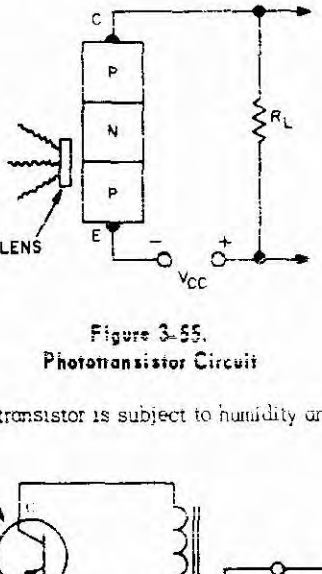


Figure 3-55.
Phototransistor Circuit

The phototransistor is subject to humidity and temperature changes like the photodiode. Humidity effects are dependent upon the construction and encapsulation processes, but temperature variations may be compensated for by the use of a bridge circuit and a thermistor of equal but opposite characteristics; see figure 3-57.

Figure 3-56.
Phototransistor Relay

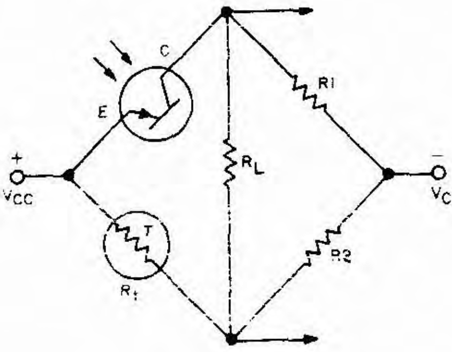


Figure 3-57.
Thermistor-Compensated Bridge Circuit

3.6.3.1 **Thermistor.** A thermistor is a special semiconductor device which functions as a thermally sensitive resistor whose resistance varies with temperature. Thermistors have large negative temperature coefficients; that is, as the temperature rises their resistance decreases, and as the temperature drops their resistance increases. The resistance of a thermistor is varied not only by ambient temperature changes but also by heat generated internally by the passage of current.

Since the termistor is basically a variable resistor, it is usually constructed from semiconductor material of greater resistivity than is used in transistors or semiconductor diodes. *Therefore,* its response to ambient temperature variations does not track equally with that of the transistor semiconductor, so that compensation is achieved only at a few points of correspondence. As a result, its greatest usage is in the field of temperature controls and measurements, and power-measuring equipment based on heating effect, such as r-f measuring microwave equipment. Although it is a semiconductor, it has no intrinsic amplification capability like the transistor, and is mentioned in this technical manual only because it is used in thermal compensating circuits for transistor stabilization.

3.7 CLASSES OF AMPLIFIER OPERATION.

Since *transistors are analogous* to vacuum tubes, the same general classes of amplification, input and output *parameters,* distortion, and efficiency are applicable. Thus the transistor can be operated as a Class A, Class B, Class AB, or Class C amplifier. Operating conditions and results for the ideal case are used in the following discussion to define the differences and relationships between the classes of operation.

3.7.1 **Class A.** The Class A amplifier is biased so that it operates on the linear portion of the collector characteristic, providing for equal swings above and below the bias point. Collector current flows continuously (with or without signal) for 360 degrees of the operating cycle, and the transistor is operated so that the maximum collector dissipation is never exceeded (other classes momentarily exceed this rating).

The Class A amplifier is basically a small-signal amplifier, although it can be used as a large-signal amplifier provided that the quiescent current does not exceed the maximum transistor ratings. Usually, large-signal amplifiers of the power type are operated as Class B amplifiers. Class A amplifiers may be operated in push-pull or as single-ended stages.

While the efficiency of a vacuum-tube amplifier operated Class A averages around 30 percent, the efficiency of a transistor operated Class A varies considerably, depending upon the circuit configuration and the parameters used. Considering the ideal case, it can be demonstrated mathematically that the direct-coupled Class A amplifier produces a theoretical maximum of 25 percent in either the CB or CE circuits. On the other hand, a resistance-coupled circuit will produce a maximum collector efficiency of 17 percent. The highest possible efficiency, 50 percent, is obtained with transformer-coupled configuration (assuming a perfect transformer, with no losses), or by use of a shunt collector feed.

For Class A operation, the transistor must be capable of dissipating more than the desired power output.

Figure 3-58 shows typical Class A operation for a CB configuration, and figure 3-59, for a CE configuration. Note that operation does not extend into the saturation region since the knee of the curve makes operation here very nonlinear. Likewise, operation in the cutoff region is not permitted, because current would flow for less than the entire cycle. Comparing the graphs of figures 3-58 and 3-59, it is seen that the CB circuit is inherently more linear since the constant-current curves are more equally spaced than those of the CE circuit. Therefore, the distortion is lower in the CB circuit than in the CE circuit. Other parameters which affect the distortion produced are: the amplitude of the input signal (if too large, it will be clipped), the value and linearity of the input resistance, and appreciable variation of the bias with temperature. To minimize distortion and produce maximum gain, the CB circuit uses an input resistance of about two times the source impedance, while the CE circuit uses an input resistance one to three times the source impedance, to minimize the over-all input resistance variations. Although the CB circuit produces less distortion and more power output for the same percentage distortion as the CE circuit, the CE circuit is usually preferred for all around use because it is easily cascaded and has a high power gain.

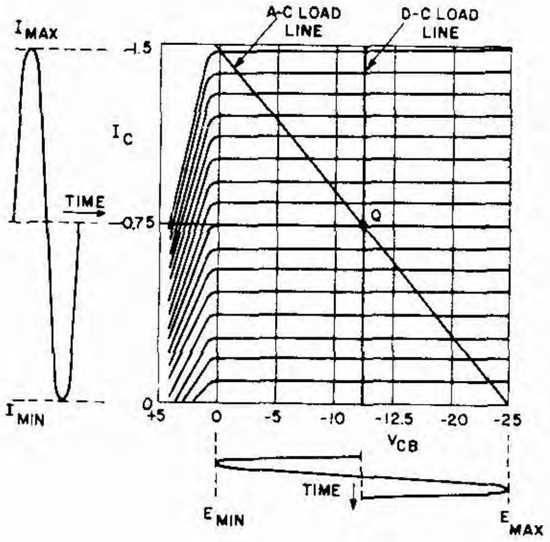


Figure 3-58.
CB Class A Graphical Operation

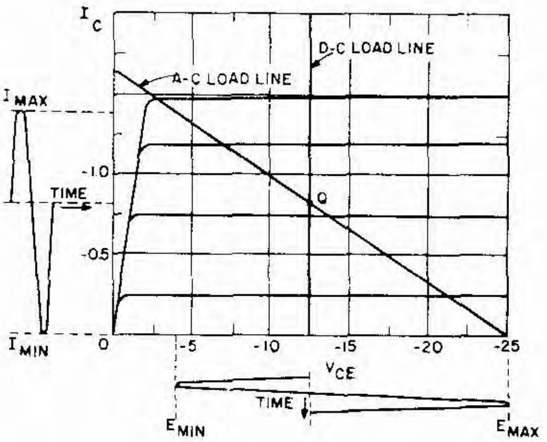


Figure 3-59.
CE Class A Graphical Operation

3.7.2 **Class B.** Class B amplifier operation is obtained when the collector current flows for one half of the operating cycle, and is entirely cutoff during the other half. The bias is set at the cutoff point (zero bias), and during the positive input signal swing an NPN transistor will amplify in a normal fashion; on the negative swing the transistor is cut off and does not operate. A PNP transistor will conduct during the negative half cycle of the input signal and remain cut off during the positive swing. To produce the full input signal, two transistors must be employed, operating back-to-back in a push-pull arrangement. On one half of the cycle one unit operates, on the other half cycle the other unit operates, and the halves are combined and added in the load. Thus each transistor operates for only half the operating period.

Figure 3-60 shows a graph of Class B operation with the input and output signals projected from the transfer characteristic. Note that while collector cutoff is assumed there is a small flow of reverse leakage current, I_{CEO}, which reduces the total efficiency of the circuit. Ideal efficiency is 78 percent, which is quite an improvement over Class A operation. Distortion components are the same as for Class A plus an additional type, known as **crossover distortion.** Since two transistors are employed, even though operating only half the time, the distortion is greater than for Class A depending on the circuit design. At the present state of the art, no general figures to indicate the possible range of values of distortion are available, since to obtain the necessary gain or output it may be necessary to accept more distortion with one transistor and design than with another.

Because each transistor operates for only half the time and the power conversion efficiency is high, the Class B operated transistor is required to dissipate only about 35 percent of the total output power desired. Hence, much greater power output is possible with lower rated transistors for the Class B amplifier. Since the ideal case is usually not obtained, the percentage to be dissipated should be calculated in each instance, but the value given above is a rough approximation for comparison purposes.

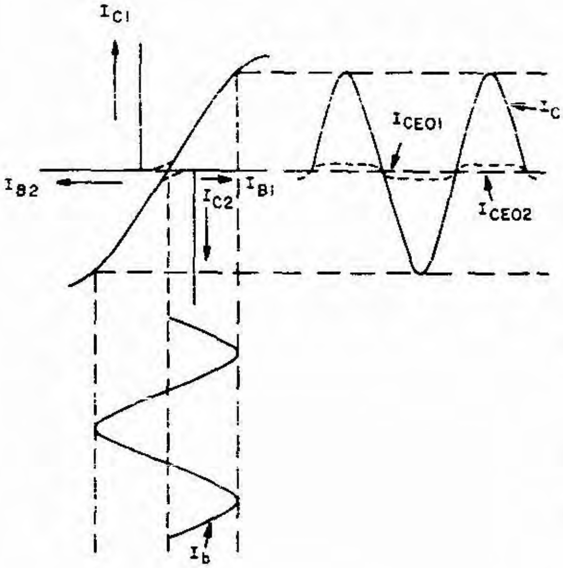


Figure 3-60.
Typical Class B Operation

Class B operation is usually used for audio power amplifier stages, but is seldom employed single-ended. However, single-ended operation is applicable to transmitters using tank circuits to fill in the missing half of the output signal, as in vacuum-tube operation.

Crossover distortion is caused basically by the nonlinearity of the transistor characteristics. At small input voltages the current change is small and varies exponentially, but at higher input voltages the transistor conduction is heavier and heavier. This action produces an inward belly, as shown in figure 3-61, and increased distortion.

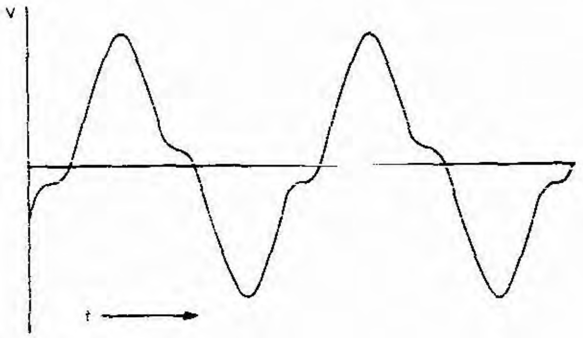


Figure 3-61.
Crossover Distortion

Compensation for crossover distortion is usually achieved by placing a slight forward bias on the base of the transistor, to move the bias point to a more linear portion of the transfer characteristic. While feedback can be used or the source resistance can be increased to minimize crossover effects, circuit design complications, together with the loss in the increased source resistance, make these types of compensation generally unsatisfactory for general use. Actually, the *biasing-off type of compensation places* the amplifier in the Class AB range of operation.

3.7.3 **Class AB.** Operation in the Class AB region is obtained by biasing to the point where collector current flows for more than a half cycle, but not for the entire cycle as in Class A operation. By arranging the bias properly, efficiencies between 50 and 78 percent are obtainable, *with an average of 55 percent representing* typical Class AB operation.

Figure 3-62 shows a graph of typical Class AB operation. Distortion is less than that of Class B and more than that of Class A. The circuit arrangement is usually push-pull. However, for those applications which can tolerate the increased distortion, it is possible to use single-ended operation with an increase of output over Class A operation. For push-pull operation, Class AB represents greater power output and slightly more distortion than Class A, but less power output and slightly less distortion than Class B.

3.7.4 **Class C.** Class C operation is obtained by biasing to the point *where collector current flows* for less than a half cycle, with the transistor remaining in the cutoff condition and with a slight cutoff (reverse) current flowing during the inoperative portion of the cycle. Class C operation is not used for audio amplification because of the severe distortion it produces. It is used for tank circuit applications where the distortion is smoothed out and minimized by the *flywheel effect*, as in vacuum-tube operation; also, it is used in single-ended or push-pull configurations. Switching circuits are usually operated as Class C amplifiers.

To achieve Class C operation with a bias point considerably below cutoff, it is necessary to reverse-bias the emitter (*assuming a common-emitter circuit*), as opposed to forward-bias for normal operation in the other classes.

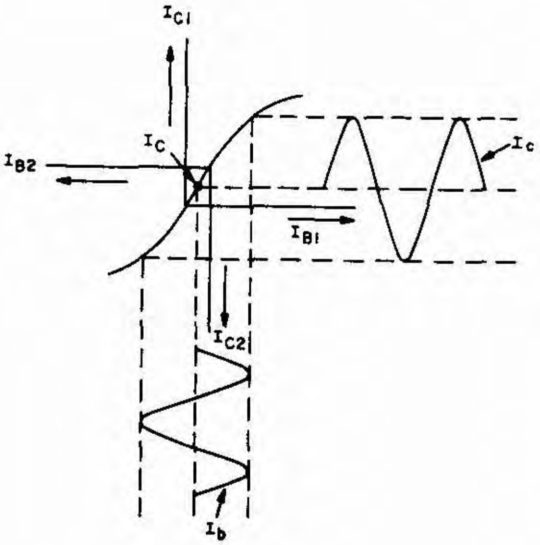


Figure 3-62.
Typical Class AB Push-Pull Operation

The Class C amplifier can be considered to operate as a pulsed oscillator for r-f energy, where the input pulses cause conduction for a small portion of the operating cycle, and the transistor rests with a small steady current during the remainder of the cycle, in a cutoff condition, with the r-f oscillations being sustained by the parallel resonant tank circuit. Assuming an input sine wave, it is only necessary to supply the losses in the tuned inductor by transistor conduction, and obtain an amplified sine wave output produced by the voltage gain factor. Naturally this type of operation is restricted to essentially sine-wave oscillations produced at the frequency to which the tuned load circuit is resonant. Figure 3-63 shows a typical Class C amplifier circuit with input and output waveforms.

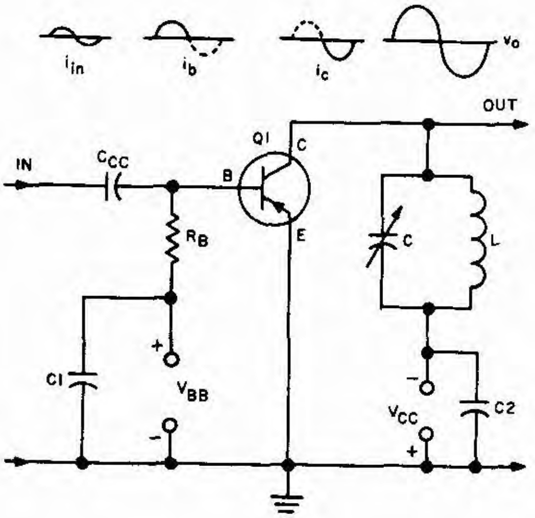


Figure 3-63.
Class C Amplifier, CE Circuit

3.8 COUPLING METHODS.

The transistor, like the vacuum tube, is usually connected in cascaded stages to amplify the feeble input signal to the large output value needed. Coupling is accomplished by using resistance-capacitance networks, impedance networks, or transformers, or directly, by connecting the output element to the input element of the succeeding stage, as described in the following paragraphs. The discussion in this section will be limited to the basic circuit and important considerations involved for audio or relatively low-frequency circuits. Where special combinations or design considerations are required to achieve a particular result (for example, r-f or i-f coupling), they will be discussed in the proper section with the special circuit with which they are used. Since all coupling networks are frequency responsive to a certain extent, some coupling methods afford better results than others for a particular circuit configuration. Generally speaking, resistance coupling affords a wide frequency response with economy of parts and full transistor gain capabilities, impedance and transformer coupling provide a more efficient power matching capability with moderate frequency response, while direct coupling provides the maximum economy of parts with excellent low-frequency response and d-c amplification.

3.8.1 **R-C Coupling.** The R-C coupler utilizes two resistors and a capacitor to form an interstage coupling device which provides a broad frequency response, with high gain, an economy of parts, and small physical size. It is used extensively in audio amplifiers, particularly in the low-level stages. Because of its poor input-output power conversion efficiency (17 percent for the ideal case), it is seldom used in power output stages.

Figure 3-64 shows a typical resistance coupler. Resistor R_L is the collector load resistor for the first stage, capacitor C_{CC} is the d-c voltage-blocking and a-c signal-coupling capacitor, and R_B is the input-load and dc-return resistor for the base-emitter junction of the second stage.

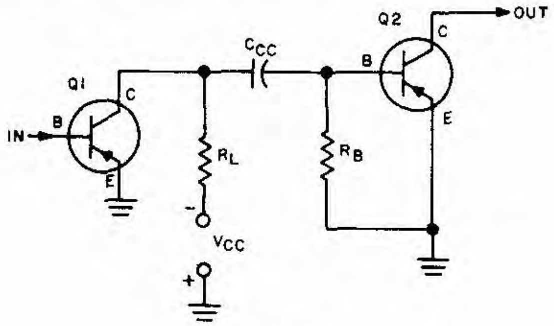


Figure 3-64.
Resistance Coupling

Since the input resistance of the second stage is low (on the order of 1000 ohms for a CE circuit) and the reactance of the coupling capacitor is in series with the base-emitter internal input resistance, C_{CC} must have a low reactance to minimize low-frequency attenuation due to a large signal drop across the coupling capacitor. This is achieved by using a high value of capacitance; thus, for low audio frequencies, values of 10 to 100 microfarads or more are employed (compare this with the vacuum-tube coupling capacitance of less than 1 microfarad).

To prevent shunting the input signal around the low base-emitter input resistance, the base dc return resistor, R_B, is made as large as practical with respect to the transistor input resistance. Since increasing the base series resistance deteriorates the temperature stability of the base junction (see discussion of bias stabilization in paragraph 3.4.2), the value selected for the input resistor is a compromise between reducing the effective shunting of the input resistance and maintaining sufficient thermal stability over the the desired temperature range of operation.

The high-frequency response is normally limited by the stray circuit capacitance plus the input and output capacitance; hence, the transistor itself is usually the limiting factor. The low-frequency response is normally limited by the time constant of the coupling capacitor, C_{CC}, and the base return (input) resistance, R_B. For good low-frequency response, the time constant must be long in comparison to the lowest frequency to be amplified.

Like the vacuum-tube coupling networks, transistor coupling networks may also be compensated to increase frequency response. Figure 3-65 shows the basic equivalent circuits for three types of compensation: (A) shunt peaking; (B) series peaking, and (C) combined shunt-series peaking. Insertion of series inductor L_1 produces a parallel resonant effect with output capacitance C_{oe} and input capacitance C_{ie} and improves the high-frequency response about 50 percent. Insertion of inductor L_2 in series with C_{CC} produces a series resonant circuit with input capacitance C_{ie} and further increases the high-frequency response about 50 percent over that of shunt peaking. Using both series- and shunt-peaking effects provides a gain about 80 percent greater than that of the series-peaking circuit alone.

Since the response to low frequencies is limited only by the coupling network, low-frequency compensation can be provided as in vacuum-tube circuits. Figure 3-66 shows a typical low-frequency compensation circuit. With resistor R_1 inserted in series with R_L, the collector load is increased at those frequencies for which the resistance of R_1 is effective. Since capacitor C_1 parallels or shunts R_1, it is evident that the higher frequencies are bypassed around it, but, since the capacitive reactance of C_1 increases with a decrease of frequency, the lower frequencies pass through R_1. Thus, the load resistance for low frequencies is increased and so is the output at these frequencies. The combination of C_1 and R_1 is chosen to provide the desired frequency compensation. This type of compensation also corrects for phase distortion, which is usually more prevalent at the lower frequencies.

3.8.2 **Impedance Coupling.** The impedance coupler is used extensively in the transistor field. Here the increased

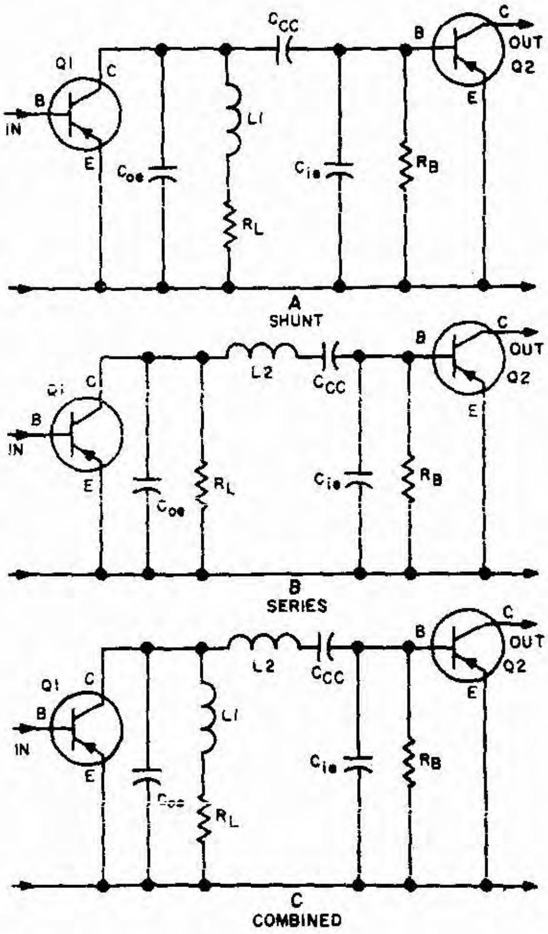


Figure 3-65.
Shunt, Series, and Combined Peaking Circuits

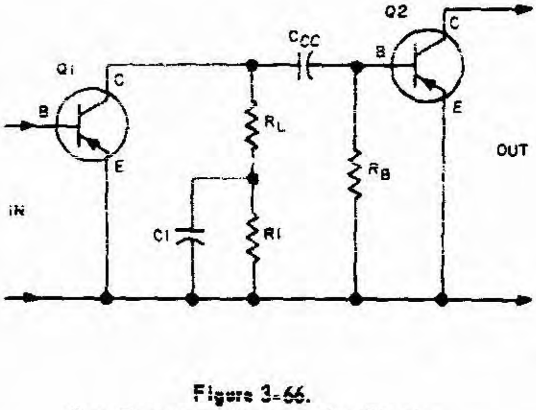
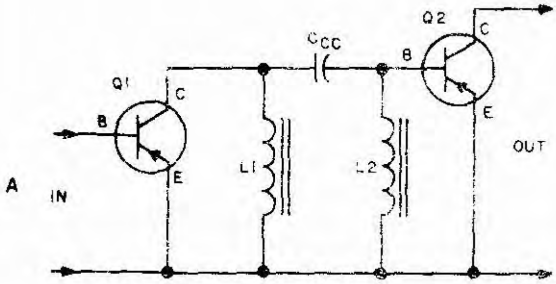


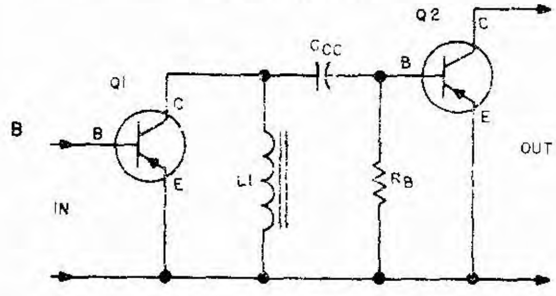
Figure 3-66.
Low-Frequency Compensation Circuit

power-handling (and matching) capabilities of the inductor provide more output than the load resistor. While the overall frequency response of impedance coupling is not as good as that of resistance coupling, it is much better than that of transformer coupling, because there are no leakage reactance effects to deteriorate the high-frequency response.

Part A of figure 3-67 shows the basic impedance-coupling circuit, and Part B shows a typical variation. The high-frequency response of the impedance coupler is limited mainly by the collector output capacitance, and the low-frequency response is limited by the shunt reactance of the inductor, L1. The efficiency of the impedance coupler is approximately the same as that of the transformer-coupled circuit (50 percent for the ideal case).



L-C-L COUPLING



L-C-R COUPLING

Figure 3-67.
Impedance-Coupling Circuits

3.8.3 **Transformer Coupling.** Transformer coupling is used extensively in cascaded transistor stages and power output stages. It provides good frequency response and proper matching of input and output resistances with good power conversion efficiency. It is relatively much more costly and occupies more space than the simple R-C circuit components, but it compares favorably in these respects with the impedance coupler. Its frequency response is less than that of the resistance- or impedance-coupled circuit.

Figure 3-68 shows a typical transformer coupler. Coupling between stages is achieved through the mutual inductive coupling of primary and secondary windings. Since these windings are separated physically, the input and output circuits are isolated for d-c biasing, yet coupled for a-c signal transfer. The primary winding presents a low d-c resistance, minimizing collector current losses and allowing a lower applied collector voltage for the same gain as other coupling methods, and it presents an a-c load impedance which includes the reflected input (base-emitter) impedance of the following stage. The secondary winding also completes the base d-c return path and provides better thermal stability because of the low d-c (winding) resistance. Since the transistor input and output impedance can be matched by using the proper turns ratio, maximum available gain can be obtained from the transistor.

As in the impedance coupler, the shunt reactance of the transformer windings causes the low-frequency response to drop off, while high-frequency response is limited by the leakage reactance between the primary and secondary windings, in addition to the effect of collector capacitance. Because of the low d-c resistance in the primary winding, no excess power is dissipated, and the power efficiency approaches the maximum theoretical value of 50 percent.

3.8.4 **Direct Coupling.** Direct coupling is used for amplification of dc and very low frequencies. As in vacuum-tube circuits, this method of coupling is limited to a few stages since all signals are amplified, including noise, and it is extremely susceptible to instability because of shift of operating point, cumulative d-c drift, and thermal changes. Its use in power output stages is limited because of the low conversion efficiency (about 25 percent). It does offer an economy of parts, and it lends itself to the use of complementary-symmetry circuitry.

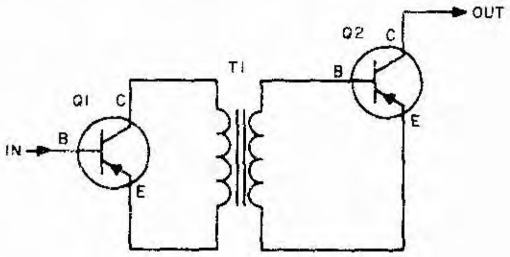


Figure 3-68.
Transformer Coupling

Figure 3-69 shows a basic d-c amplifier utilizing two PNP transistors and two power sources. When a signal is applied to the base of Q1, the amplified output is directly applied to the base of Q2 from the collector of Q1. The output is taken from load resistor R_L of Q2. Since the base bias of Q2 is applied through R_{B2}, the amplified signal on the collector of Q1 must not drive the base of Q2 positive; that is, it must not exceed the negative bias.

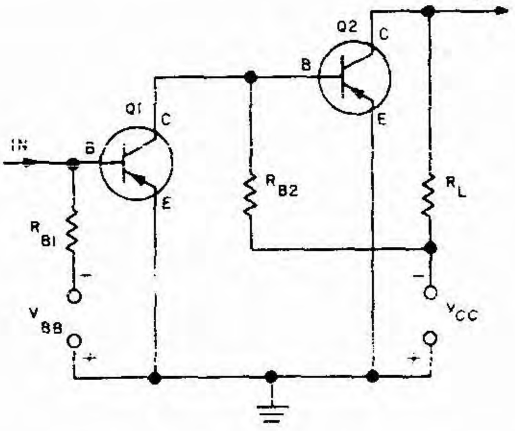


Figure 3-69.
D-C Amplifier

Figure 3-70 shows a basic connection not possible with electron-tube amplifiers. The grounded-base circuit of Q1 is direct-connected to the grounded-emitter circuit of Q2. Thus the input circuit of Q2 is the load for Q1, and collector bias for Q1 is obtained through the collector-to-base junction of Q2. Since Q2 biases Q1, only one power source is needed.

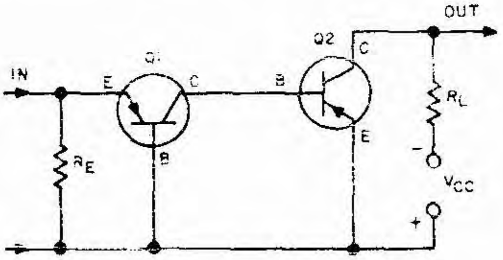


Figure 3-70.
CB to CE D-C Amplifier

Figure 3-71 shows a typical complementary-symmetry circuit using an NPN and a PNP transistor. Since the currents flow in opposite directions, thermal effects oppose and stabilize each other. As in figure 3-70, the collector bias for Q1 is obtained from the base-collector junction of Q2.

From the circuit of figure 3-71, it is clearly seen that if another stage were added an additional and larger collector bias supply would be required to maintain the collector-to-base potential negative for each stage. This limitation is analogous to that of the d-c supply for the vacuum-tube amplifier. It is also evident that a shift of d-c bias potential would be amplified and passed along to the second amplifier, whereas in the a-c coupled (resistance-capacitance) amplifier such d-c shift would be blocked by the coupling capacitor.

Note the use of complementary symmetry or the use of one transistor to bias another with d-c coupling affords the minimum of component parts possible, and represents an economic advantage that is possible only with transistors.

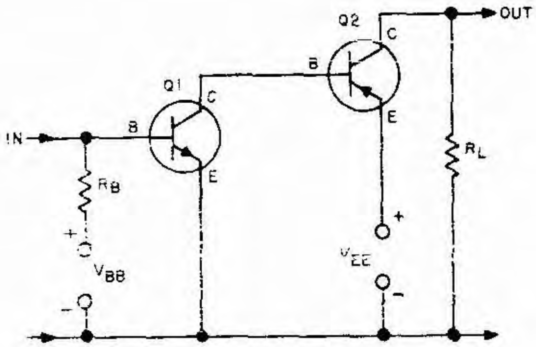


Figure 3-71.
Complementary-Symmetry D-C Amplifier

Special bias circuits are used with d-c amplifiers to reduce thermal effects; see bias stabilization discussion in paragraphs 3.2 and 3.4 for basic circuitry.

3.9 TIME CONSTANTS.

The time constant as defined in paragraph 2.5 for electron tubes is the same for transistors. It is important to realize that transistors differ primarily from electron tubes in their input and output resistances and their capacitances. Where time constants are used in the grid circuit of an electron tube with essentially infinite input impedance, it is not necessary to be concerned with the input-impedance effect on the time constant. Since the transistor has a finite and relatively low input impedance, however, it may affect a time-constant circuit. Also, the capacitance to base of the transistor is usually larger than the electron tube plate-to-ground or grid capacitance; hence, when it shunts the time-constant circuit, it must be considered.

Thus it can be seen that R-C and L-R circuits used for time-delay, coupling, and frequency-response effects and for the shaping of pulses and the controlling of switching circuits are practically interchangeable (except for shunting effects) between electron-tube circuits and transistor circuits.

Transient conditions which produce large overshoots must be avoided in transistor circuits because the transistor is more easily damaged than the electron tube by potentials which exceed the maximum rated voltages. Generally speaking, therefore, the R-L circuit is used less than the R-C circuit.

On the other hand, because of the shunting effects of the internal transistor parameters, the high-frequency response is degraded, even in the audio range. Consequently, when high frequency response is desired, more high-frequency compensating circuits are used with the transistor than with the electron tube. Therefore, the use of time-

769-978 O-65—5

constant circuitry will be completely discussed in the applicable circuit analysis in other sections of this technical manual, since it depends primarily on the type of transistor selected.

3.10 HYBRID PARAMETERS.

The hybrid parameters, or h-parameters, of the transistor are mostly of use to the circuit designer. However, manufacturers have found them convenient for use in defining the performance of their products. Therefore, a working knowledge of the use and meaning of h-parameters is essential to the technician.

The h-parameters appear in two forms which are interchangeable; one form employs letter subscripts, and the other, numerical subscripts. At present, the numerical subscripts are commonly used for general circuit analysis, and the letter subscripts are used for specifiying characteristics of transistors. Industry standardization accepts both forms, but the present trend in usage indicates that the numerical subscripts may be preferred. In addition, there are other systems of parameters, such as "a", "b", "y", "g", "r", and "z" parameters, which have somewhat more specialized uses. All of these systems, which may seem confusing to anyone other than a trained engineer, represent different methods of mathematically defining transistor action, as well as the parameters limiting that action, for design purposes. Although the triode transistor is a 3-terminal device, it may be treated as a "black box" with two input and two output connections (a basic 4-terminal network). By utilizing the electrical characteristics (h-parameters) of this black box, it is possible to calculate performance of various circuits when various input signals are applied and various loads are connected. Basically, these h-parameters are limited to frequencies sufficiently low (270—1000 cps) that the capacitive and inductive effects of the transistor can be neglected. The h-parameters for common-base connection are usually shown in specification sheets because the emitter current and collector voltage can be maintained more precisely than for other configurations. Because of the trend toward the use of common-emitter circuits and the relative ease of measurement, some CE h-parameters will also be observed. In any event, formulas are available for conversion from one configuration to the the other in most text books (and transistor manuals), so only the common-base connection will be discussed in this manual.

The simple common-base configuration shown in figure 3-72A can be represented by the four-terminal h-parameter equivalent circuit of figure 3-72B. In using the h-parameters, it is customary to ignore the bias and consider only the instantaneous a-c values involved. Thus, the h-parameters represent operating conditions for a small signal close to the operating point. Therefore, the equivalent circuits do not show bias supplies and polarities. Conventional currents (not electron flow) and voltage polarities are assumed, and if erroneously assigned, will result in a negative answer when the problem is solved. Since the circuit of figure 3-72 is essentially a four-terminal network (two input and two output terminals), it can be described by two simultaneous equations in which the h-parameters are the coefficients, namely:

$$E_1 = h_{11}\, i_1 + h_{12}\, E_2$$
$$I_2 = h_{21}\, i_1 + h_{22}\, E_2$$

To conform with Kirchhoff's laws, h_{11} in figure 3-72B must be an impedance and h_{22} an admittance, while h_{12} and h_{21} are essentially dimensionless ratios. For low frequencies h_{11} and h_{22} are resistive. Since these parameters involve two opposites—impedance and admittance—the term **hybrid** is used to describe them.

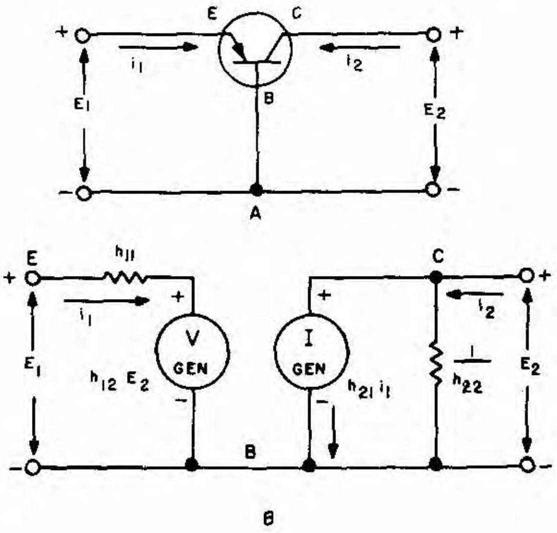


Figure 3-72.
Transistor and Four-Terminal Network Equivalent Circuit

If the output terminals are short-circuited for ac (by a large capacitor), the output voltage (e_2) is zero, and the following simple formulas (like Ohm's law) show the relationships between input voltage and current and between input current and output current.

$$h_{11} = \frac{E_1}{I_1}$$

$$h_{21} = \frac{i_2}{i_1}$$

If the a-c input circuit is opened by the insertion of a large inductance in series with the bias, thereby reducing i_1 to zero, the following additional formulas will be obtained:

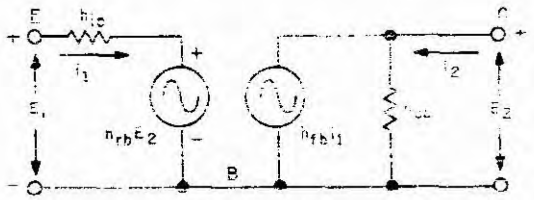
$$h_{12} = \frac{E_1}{E_2} \text{ and } h_{22} = \frac{i_2}{E_2}$$

Thus, h_{11} is effectively the transistor input impedance (with the output short-circuited), and h_{22} is the output admittance (with the input open-circuited). Similarly, h_{12} is the voltage feedback ratio with the input open-circuited (this represents the internal feedback due to reverse-current effects and common impedance coupling within the transistor). Parameter h_{21} is the forward *current amplification* ratio with the output short-circuited.

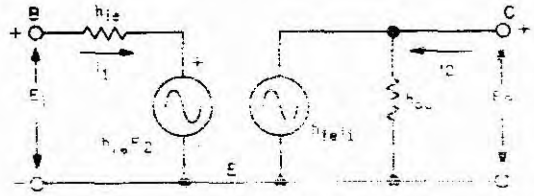
The functioning of the equivalent circuit and the meaning of the h-parameters can perhaps be more clearly understood by considering the following intuitive reasoning. Consider the circuit of figure 3-72B, and for the moment neglect the voltage produced by the voltage generator (V_{GEN}) in the input circuit. When the input signal voltage (E_1) is applied to the input terminals, current i_1 flows through resistor h_{11} and causes current h_{21} i_1 *to flow in the output current* source (current generator I_{GEN} in the figure). Thus, h_{21} is the current gain of the equivalent circuit. Current h_{21} i_1 divides between the output resistor, which is equal to $1/h_{22}$ (h_{22} is an admittance), and the external circuitry connected across the transistor output terminals. The voltage developed across the output as a result of this current is the output voltage (E_2). When voltage E_2 appears across the output, it causes an internal feedback voltage to be fed back to the input; *this voltage is* h_{12}, E_2, *represented by voltage generator* V_{GEN} at the source. This feedback voltage opposes the initial signal input (E_1) and effectively subtracts from it.

Figure 3-73 shows the hybrid equivalent circuits for the three basic transistor configurations. Note that the circuits are all the same because the defining equations must be satisfied. The parameter notation, however, is different for each configuration; it consists of the general *parameter designations* h_i, h_r, h_f, and h_o with the additional subscript b, e, or c added to represent the common base, emitter, or collector configuration, respectively. The letter and numerical forms of the parameters are related as follows: h_i is h_{11}, h_r is h_{12}, h_f is h_{21}, and h_o is h_{22}.

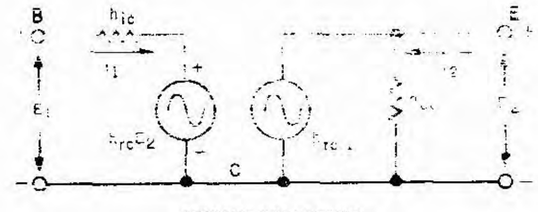
From the preceding discussion, it is obvious that with a few external measurements the h-parameters of the typical "black box" can be determined, and that substitution of these values in the proper formulas will allow circuit performance to be approximated so that the proper matching values of external circuitry can be chosen by the designer.



COMMON BASE



COMMON EMITTER



COMMON COLLECTOR

Figure 3-73.
Basic Hybrid Parameter Equivalent Circuits