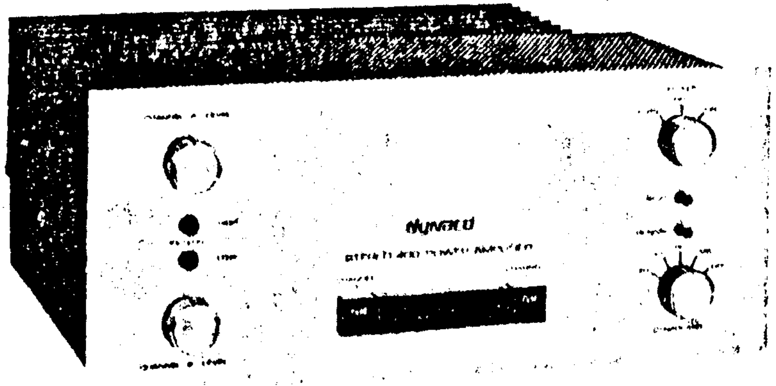


Figure 5-1. *(Continued)*

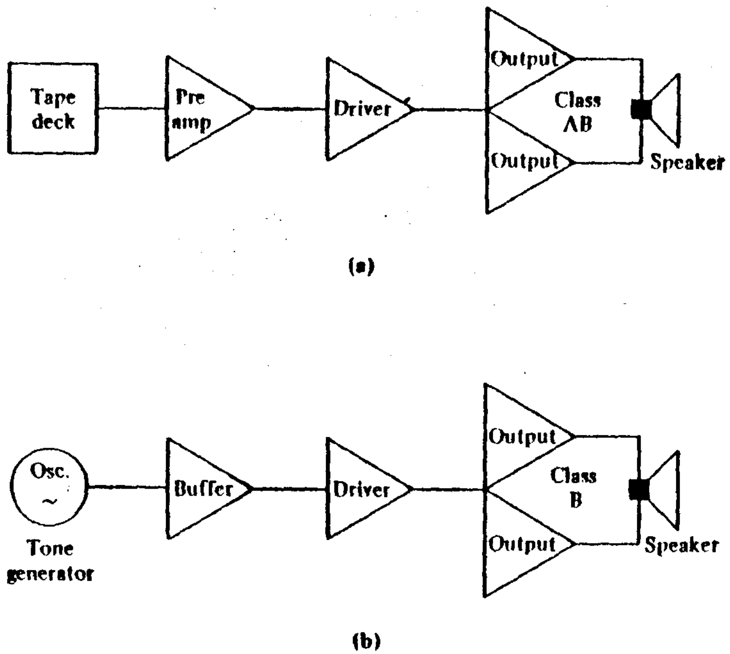


Figure 5-2. Examples of power-amplifier and driver arrangements: (a) tape deck, preamplifier, driver amplifier, and output power amplifier; (b) oscillator, buffer amplifier, driver amplifier, and output power amplifier.

less (OTL) designs. The arrangement depicted in Fig. 5-2a is representative of high-fidelity design; the arrangement shown in Fig. 5-2b is typical of electronic organs. Some form of negative feedback is always employed

in an audio output configuration to minimize distortion. A power amplifier may be combined with a preamplifier in the same cabinet, or it may be enclosed in an individual cabinet. In the case of public-address (PA) systems, the power amplifier may be housed with the speaker in a common enclosure. There is an indicated trend to the mounting of high-fidelity power amplifiers in the same enclosure as the speaker system. This design facilitates voice-coil negative feedback and provides unusually good transient response.

5-2 Output-Transformerless Power Amplifiers

It was previously noted that audio transformers impose undesirable limitations on the design of high-fidelity amplifiers, particularly power amplifiers. Costly and bulky transformers are required for full high-fidelity frequency response. Phase shifts in the low- and high-frequency cutoff regions also impose difficulty in the design of negative-feedback loops. Therefore, output-transformerless (OTL) configurations are favored by hi-fi audio circuit designers. Three basic types of circuitry are employed. One approach utilizes a driver transformer to obtain phase inversion. In this function, the transformer is not labored as much as when it is utilized as an output component. In turn, its size and cost are comparatively low.

A second OTL configuration in general use is called the quasi-complementary circuit; it uses two identical output devices driven by two lower-powered complementary transistors that provide the equivalent of phase inversion. A third configuration, termed the fully complementary amplifier, employs a complementary pair of devices in its output section so that phase inversion is accomplished in the power output transistors, or in the combination of the output transistors and their drivers. We will consider the basic characteristics of these three design approaches.

5-3 Transformer Phase Inverters

A typical configuration that uses a transformer for phase inversion is shown in Fig. 5-3. Q1 and Q2 may be regarded as a two-stage voltage amplifier that drives the output transistors Q3 and Q4 through the driver transformer. Each transistor stage can be considered a power amplifier. In other words, Q1 develops a small amount of signal power to drive a larger transistor (Q2), which in turn develops sufficient power to drive the high-power output transistors Q3 and Q4. In theory, at least, Q3 and Q4 are larger devices than Q2, and Q2 is a larger device than Q1.

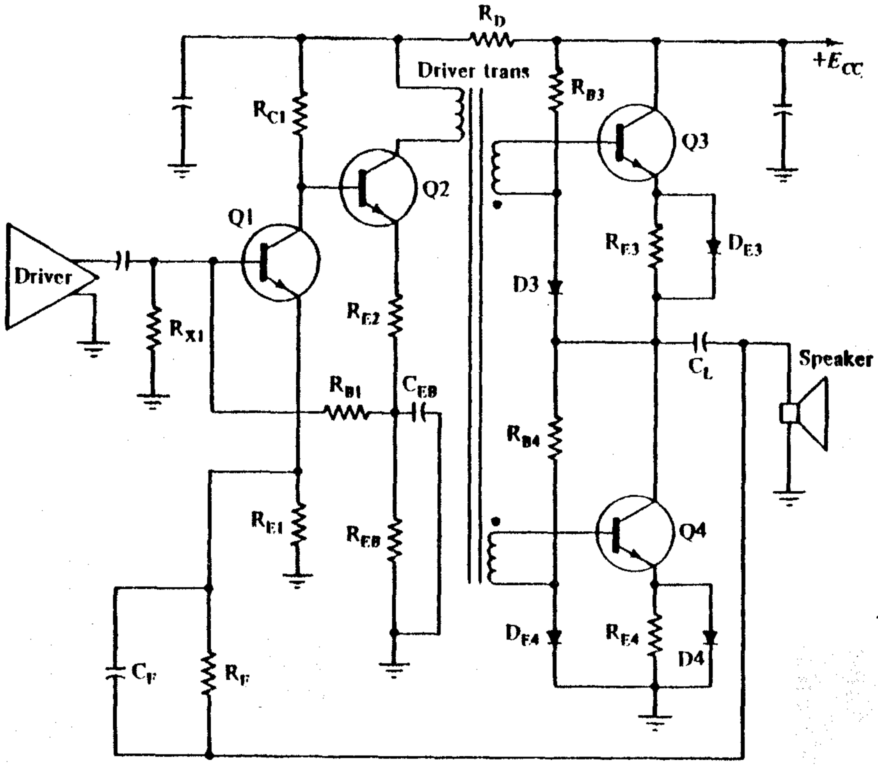


Figure 5-3. Driver transformer provides phase inversion for the push-pull output stage; may be used in PA equipment.

Q1 is direct coupled to Q2, and Q1 obtains its bias voltage from Q2's emitter circuit. Inasmuch as the dc feedback through R_{B1} from the junction of R_{E2} and R_{EB} is appreciable, adequate temperature stabilization is realized. R_{EB} in Q2's emitter circuit is bypassed by C_{EB} to prevent any ac signal voltage from being fed back from this point along with the dc bias voltage. Next, the output from Q2 is fed to a transformer that has two identical secondary windings; the audio circuit designer often specifies bifilar windings. The phase relationship between these two windings is indicated by dots. Dots placed at the ends of the two windings indicate that these ends are in phase with respect to the unmarked ends.

When the instantaneous voltage in a cycle is such that the unmarked ends of the windings are positive with respect to the dotted ends, Q3 is forward biased, and this transistor conducts while Q4 is reverse biased and therefore cut off. In the following portion of the cycle, opposite polarity relations exist at the transistor bases; Q4 conducts while Q3 is cut off. Finally, the composite signal wave form is reconstituted across R_L. The impedance ratio of the transformer selected by the audio circuit designer is based upon considerations of presentation of an ideal load to the driver

transistor Q2. Typical designs utilize an impedance ratio of approximately 9 to 1. The audio circuit designer customarily optimizes the winding ratio for minimum distortion in his breadboard model.

If we assume that adequate transistors are chosen and are mounted on efficient heat sinks, the amount of audio signal power that the amplifier can deliver is based on the value of the supply voltage E_{CC}, upon the collector-to-emitter saturation voltage, and upon the voltage drops across emitter resistors R_{E3} or R_{E4}. Output power is related to the output load resistance by the familiar terms V^2_{rms}/R_L or $I^2_{rms}R_L$. Peak voltages and currents are 1.41 times greater than the rms values, and peak-to-peak voltages and currents are 2.83 times greater than the rms values. The power supply must have the capability of swinging the peak-to-peak output voltage across the load resistance, in addition to the peak-to-peak voltage swing that occurs across one of the emitter resistors R_{E3} or R_{E4}.

Note that collector-to-emitter saturation voltage limits the swing of the voltage across the load. Inasmuch as two transistors are involved in this consideration, the sum of both of the saturation voltages at the peak of the collector current swing must be added to $V_{p\text{-}p} + I_{p\text{-}p}R_{E3}$ to estimate the minimum supply voltage that will be required if the amplifier is to deliver a specified amount of power. Restriction of the path of operation to within the linear region requires that the specified saturation voltage be multiplied by a factor of at least 3 before it is added to the other values in the relationship, in order to determine the minimum E_{CC} supply voltage value that will be required if the amplifier is to deliver a specified amount of power without objectionable distortion.

We observe that diodes D3 and D4 in the output section are forward biased and are included by the designer to stabilize the quiescent current against variations of V_{BE} versus temperature. Idling current is established by the voltage dropped across the diodes, in addition to the voltage dropped across the other resistors in the dc circuit. Resistors R_{E3} and R_{E4} in the emitter circuits are employed primarily to supply ac and dc negative feedback, and thereby reduce inherent distortion as well as contribute to temperature stabilization. As an auxiliary benefit, the emitter resistors assist in limiting the emitter current and thereby provide some measure of protection against overdissipation in the output transistors in case that R_L is accidentally short-circuited. In class A amplifiers, approximately 0.5 to 1.5 V is developed across an emitter resistor; similar drops are desirable in class AB or class B operation on peak-current points of the signal cycle.

Diodes D_{E3} and D_{E4} may be omitted; however, they permit the use of higher-valued emitter resistors for improved temperature stabilization. In this situation, the diodes are required to bypass the resistors for obtaining large current swings. The amplified signal voltage is capacitively coupled to R_L; a large value of capacitance must be used to obtain low-frequency output. R_F and C_F, in combination with R_{E1}, are the essential components of the feedback circuit. The circuit designer generally determines the value of C_F experimentally for best transient response. Circuits with driver transformers are necessary when germanium-type output transistors are utilized. Germanium transistors have comparatively high leakage currents, and complete isolation of the output transistors by a driver transformer is desirable. Although it is possible to employ quasi-complementary circuitry with germanium transistors, it is best adapted to operation with silicon-type output transistors.

5-4 Quasi-Complementary Power Amplifiers

A complementary-symmetry power amplifier employs a pair of NPN and PNP output transistors to provide phase inversion and push–pull output with simplified and economical circuitry. The chief disadvantage of a complementary-symmetry output configuration is the relatively low power capability of the PNP transistor, compared with that of the NPN transistor. Therefore, circuit designers often utilize a quasi-complementary configuration, in which a low-current PNP transistor is directly coupled to a high-current NPN transistor in order to simulate the circuit action of a high-current PNP transistor. (See Chart 5-1.)

A basic configuration for the quasi-complementary arrangement is shown in Fig. 5-4. This circuitry is direct coupled throughout. The incoming signal voltage is amplified by Q1 and is then applied to the complementary pair Q2 and Q3. Over the positive portion of the signal cycle, the bases of the complementary pair are driven positive with respect to their emitters; the NPN-type transistor Q2 conducts, while the PNP-type transistor Q3 is turned off. On the next half-cycle, the transistor roles are reversed. Half-cycles of signal voltage are applied to the output transistors Q4 and Q5 after amplification by the complementary pair. Both portions of the signal cycle are fed to R_L through C_L, and the input signal waveform is reconstituted across the load resistor. Negative feedback proceeds via the parallel combination of R_F–C_F.

Dc conditions in this configuration are such that half of the supply voltage must be present at point $E_{CC}/2$ in the diagram. The bias current is determined by resistors R_{B1} and R_{X1}, and the bias current through Q1 essentially establishes the quiescent condition. The collector load for Q1 consists basically of R_{B2} + R_{T2} + D1 + D2. These diodes are utilized to set and to maintain the idling current in the output circuit at the original quiescent value, despite temperature changes. However, the audio circuit designer may choose other types of temperature-sensitive devices. If temperature compensation is not essential, they may be replaced by resistors. Observe that capacitor C2 is included in a positive-feedback bootstrapping

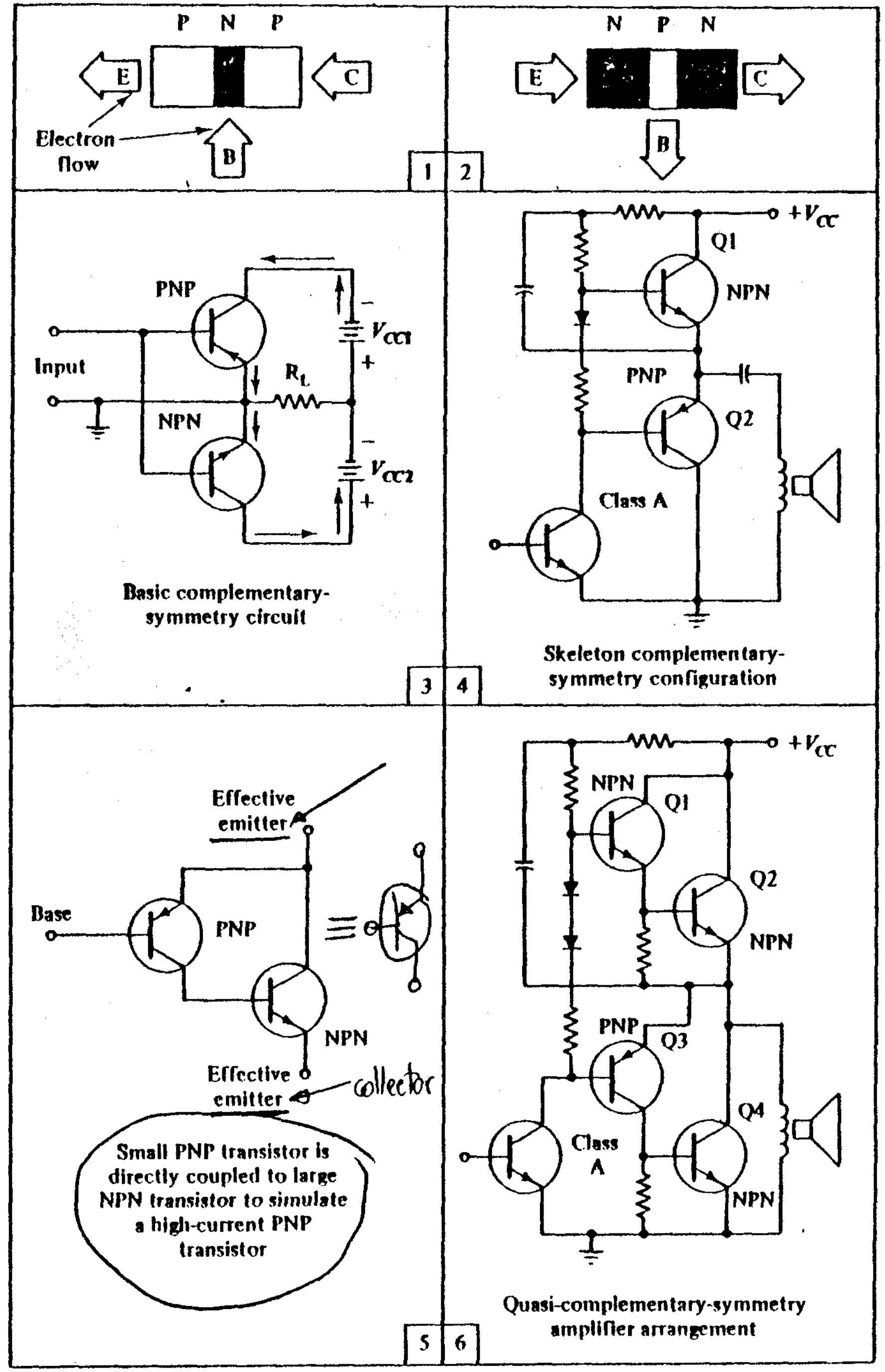


Chart 5-1.

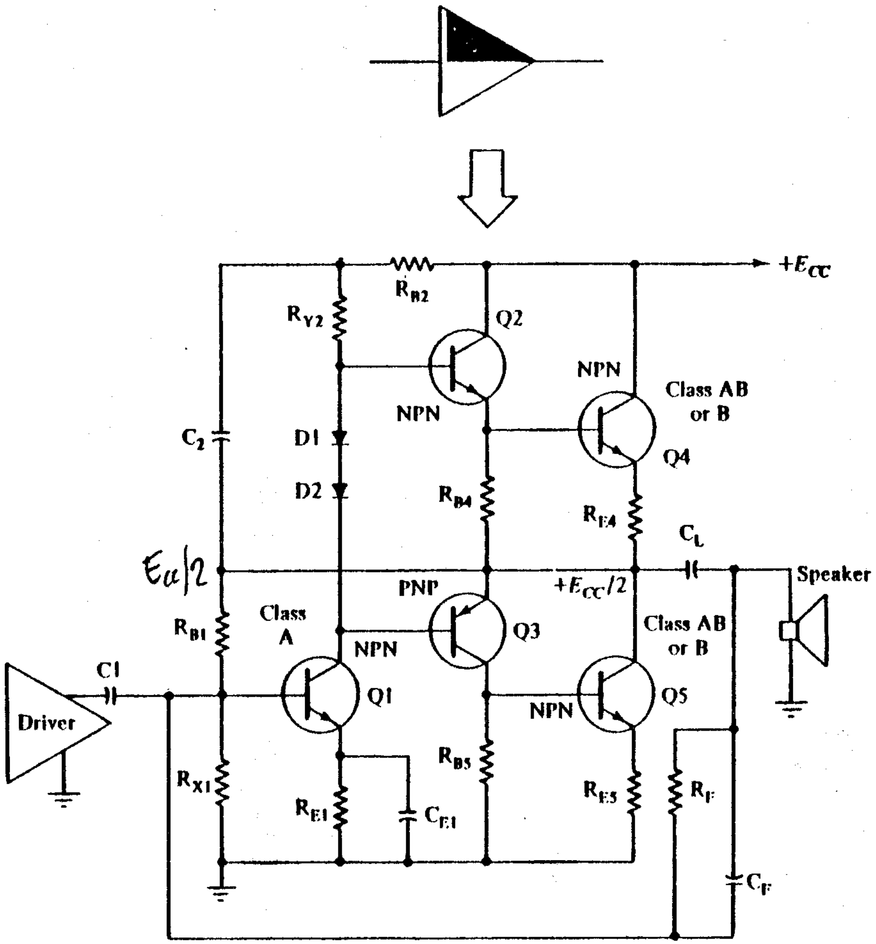


Figure 5-4. Basic quasi-complementary amplifier configuration.

circuit branch. Also, the resistors in the bias circuit for Q2 are split into components R_{B2} and R_{Y2}, thereby providing a junction for connection of C2.

When the input signal has a comparatively high amplitude, the bias on the driver transistors tends to shift the operating point into class B, thereby introducing crossover distortion. To compensate for this shift in operating point, large amounts of feedback must be placed around the circuit. The gain must be appreciable to accommodate the amount of feedback that is required. Positive feedback via C2 increases the load impedance that the complementary pair presents to Q1 and increases the gain of the circuit. High-amplitude positive peaks in the signal tend to cut off Q2 by bringing the base and emitter terminals to $+V_{CC}$ potential. However, there

is a voltage across C2 as a result of its being charged while the circuit is idling. This voltage maintains the base at a positive potential with respect to the emitter, so that Q2 continues conducting over the complete signal cycle.

Design of the bootstrap circuit is comparatively simple. Inasmuch as R_{B2} and R_{Y2} are essentially connected across the load via C2, these resistors are chosen as large as possible consistent with the base current requirement of Q2. Equal values of resistance are utilized. Under quiescent conditions, $E_{CC}/2$ drops across the series circuit formed by R_{B2} and R_{Y2} as well as across the circuit formed by R_{B2} and C2. Since $R_{B2} = R_{Y2}$, the voltage across C2 is one half of $E_{CC}/2$, or it is equal to $E_{CC}/4$. C2 charges to this voltage level and maintains a constant current through R_{Y2} and Q2's base–emitter junction. C2 must be chosen sufficiently large that it can maintain its charge during low-frequency operation.

Another bootstrap circuit designed to operate without capacitor C2 and resistors R_{B2} and R_{Y2} is depicted in Fig. 5-5. Note that a resistor is connected from the junction of C_L and R_L to the base of Q2. Thus C_L doubles as a bootstrap capacitor in addition to coupling the signal to the output load resistor or speaker. Load resistor R_L is connected to $+E_{CC}$, which is at ac ground potential. Observe that this circuit has a drawback, in

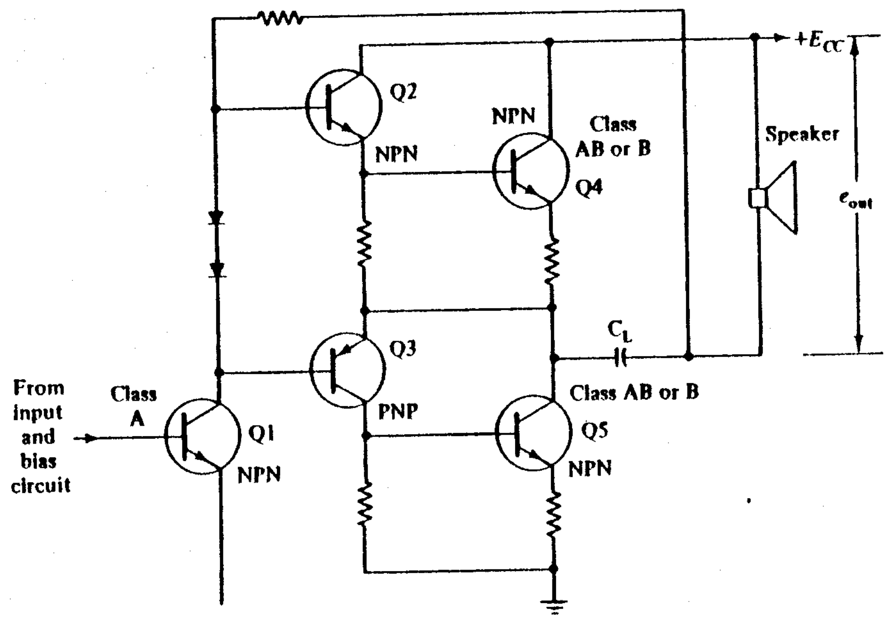


Figure 5-5. This arrangement eliminates the bootstrap capacitor by combining its function with that of the output blocking capacitor.

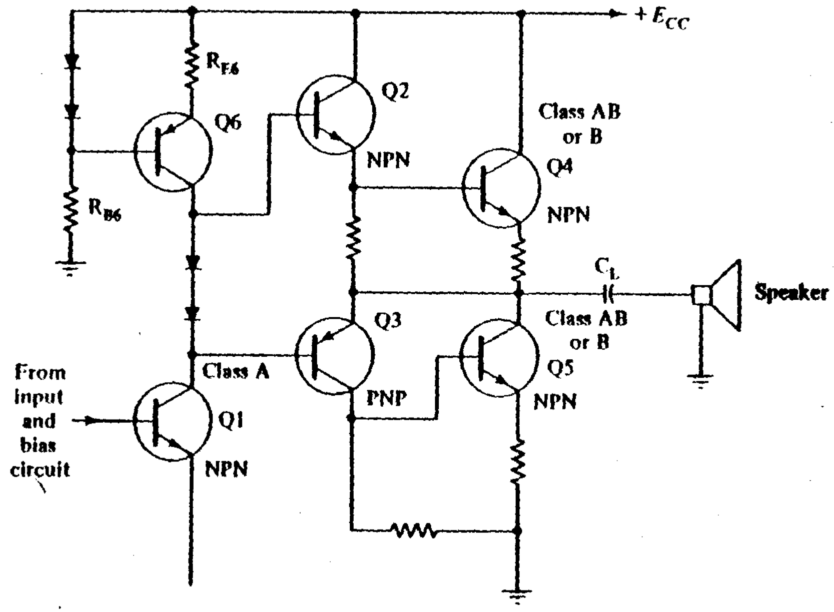


Figure 5-6. Transistor Q6 provides a constant-current source for the drivers, eliminates a bootstrap capacitor, and reduces low-frequency distortion.

that the dc base current for the drivers flows through R_L. If this current has a small value, it will not adversely affect a loudspeaker used as load R_L. A constant-current source for the bases of the complementary drivers can eliminate the requirement of a bootstrap capacitor. Refer to Fig. 5-6. This arrangement applies a constant current to the drivers and presents a high impedance to Q1, the voltage amplifier transistor. The voltage drop between the base of Q6 and $+E_{CC}$ is made as small as possible in order to avoid limiting the output voltage swing. Accordingly, low forward-voltage dropping diodes are ordinarily chosen for the constant-current circuit, instead of comparatively high voltage zener diodes. Advantages of this configuration include lower distortion in low-frequency operation and more symmetrical clipping of peaks on both half-cycles.

5-5 Directly Coupled Load

A large electrolytic capacitor is utilized in the foregoing configurations to couple the load to the output transistors. This component has some disadvantages, among which its nonlinear characteristic, low-frequency rolloff in combination with R_L, and the corner frequency created by th... come into consideration. Instabili...

feedback is applied around the circuit. Moreover, this coupling capacitor must be charged through the output transistors. This can lead to transistor damage in the event that the power–time product happens to exceed the transistor's maximum rating. Refer to Fig. 5-4. One end of R_L is connected to ground. While the circuit is idling; the other end of R_L must be at the same ground potential in order to avoid any dc current flow through the loudspeaker load. A coupling capacitor fulfills this requirement.

In the absence of a coupling capacitor C_L, dc current flow must be eliminated by placing the junction of Q4 and Q5 at a zero potential with respect to ground while the circuit idles. To accomplish this circuit action, a positive voltage with respect to ground, $+E_{CC}$, is applied to Q4's collector and a similar negative voltage, $-E_{CC}$, is applied at Q5's emitter or, more precisely, at the lower end of resistor R_{E5}. If both of these transistors conduct equal values of current during the idling period, there is 0 V at the junction of the two devices to which the load is connected (or at the junction of Q5 and the lower end of R_{E4}). With a signal voltage applied, the positive-going portion of the wave form swings the signal voltage across R_L from zero toward $+E_{CC}$, while the negative-going portion of the wave form swings the signal voltage from zero toward $-E_{CC}$.

One problem is to maintain a constant quiescent current at all times, so that the voltage across R_L remains zero. However, the current through Q1 will drift as the temperature changes. Any change in Q1's collector current will upset the balance at the output more than will current drift in any transistor farther up the chain. Current drift due to the driver and output stages is minimized by maintaining the upper and corresponding lower devices at equal temperatures on heat sinks or in free air. In turn, drift in one half of the driver and output circuit is balanced out by the drift in the other half of the configuration. Note that Q1 has no corresponding transistor to overcome or balance collector-current changes resulting from drift. Therefore, a balancing device must be added to the circuit.

A differential amplifier circuit employs two transistors. The audio-circuit designer should mount the two transistors in proximity so that temperature variations affect each equally. A basic configuration is depicted in Fig. 5-7. Q1 and Q2 operate in a differential circuit that prevents collector-current changes resulting from temperature variation. This differential pair drives the second differential pair Q3 and Q4. When a signal voltage is applied at the input of Q1, it appears in amplified form at the collectors of Q1 and Q2. The input signal is in phase with the signal at Q2's collector, whereas the input signal is out of phase with the signal at Q1's collector. As the signal is again amplified ... verted once ...

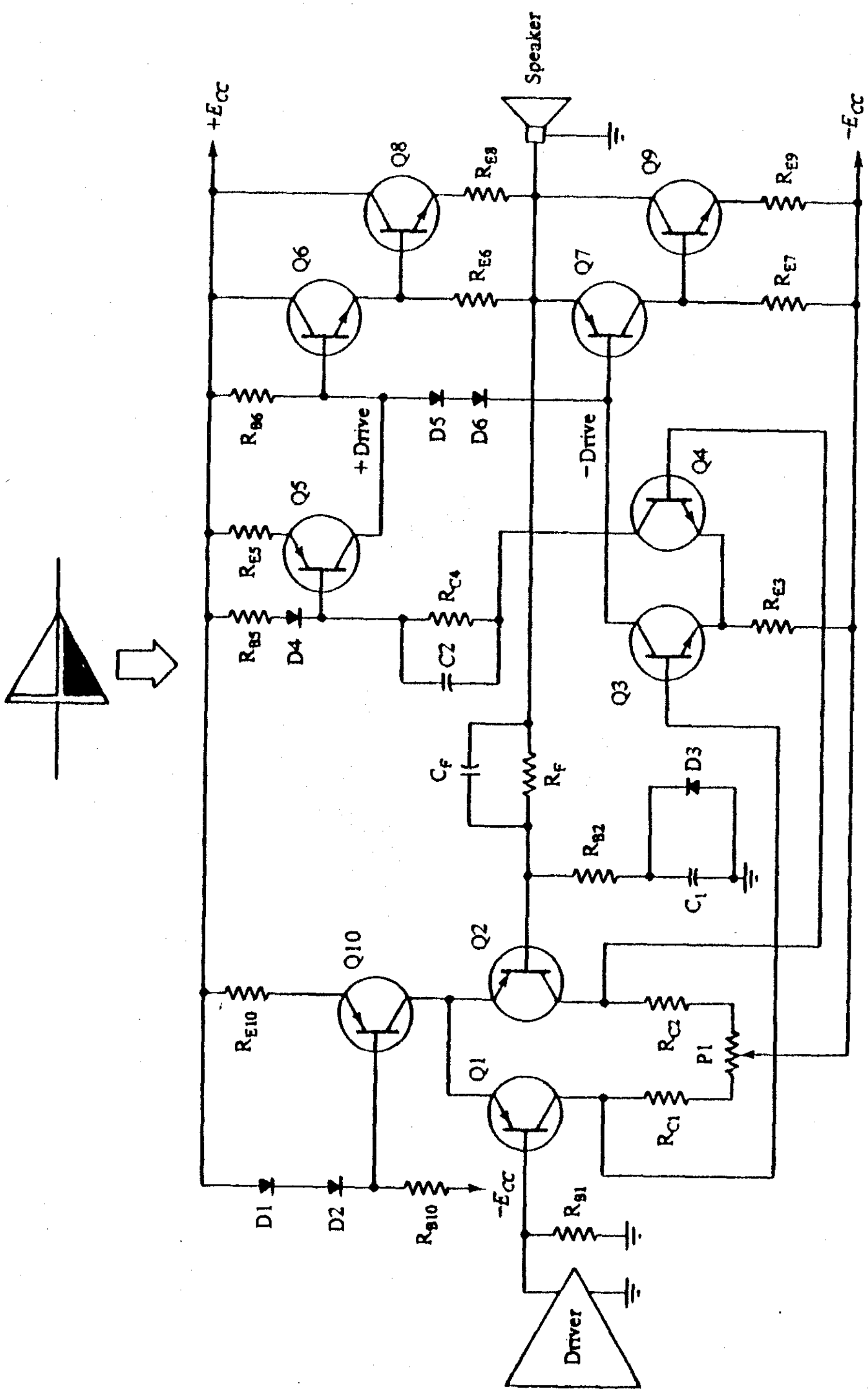


Figure 5-7. Differential-amplifier input circuit minimizes changes in quiescent current relations versus temperature.

Next, from Q4 the signal voltage is fed to the base of Q5; this transistor operates as a unity-gain amplifier, and it inverts the signal phase. The applied signal voltage drives the collector of Q5 to $+E_{CC}$, and thereby provides the positive drive voltage. Phase reversal through Q5 makes the positive drive signal identical in phase with the negative drive signal at Q3. In turn, two signals with identical phase characteristics are fed to the quasi-complementary output circuit comprising Q6 through Q9. Suitable phase relationships are achieved here to provide a reconstituted and amplified version of the input signal. The positive drive causes the upper pair of devices to swing to $+E_{CC}$, while the lower pair of devices swings from $-E_{CC}$ as a result of the negative drive. Equal swings of both halves of the output signal waveform to $+E_{CC}$ and to $-E_{CC}$ makes bootstrapping unnecessary.

Negative feedback is applied from the output of the amplifier via R_F and C_F. Both ac and dc feedback voltage are developed across R_{B2}. This resistor is bypassed to ground by C1 to provide a return for the ac feedback voltage, and diode D3 is connected to ground to provide a dc feedback voltage return path. The audio circuit designer should not assume that this circuit is in finished form. It is unlikely to operate satisfactorily unless all component values and the tolerances on both components and devices are carefully assigned in view of the stipulated performance specifications. Note that instability is likely to be encountered unless the frequency rolloff is carefully controlled at both the low- and the high-frequency ends of the audio band. It is instructive to analyze various sections of the circuitry, as follows.

Refer to Fig. 5-7. Q10 is a constant-current amplifier transistor; its circuit is designed to establish and to maintain the sum of the idling currents through Q1 and Q2. If the circuit designer employs silicon devices, the voltages across the base–emitter junction and across D2 are identical. Similarly, the voltages across D1 and R_{B10} are identical and have a value of approximately 0.7 V. This is in the range of the normal voltage drop across a silicon diode. Suppose that the sum of the idling currents through Q1 and Q2 is adjusted to a value of 2 mA. In turn, a current flow of 2 mA occurs through R_{E10}. In turn, the resistor will have a value of 0.7/0.002, or 350 Ω. Note that R_{B10} is utilized to establish a current level through and consequently to fix the voltage dropped across D1 and D2.

To determine the idling current for D1 and D2, first note the maximum value of current required in the output load at the signal peak. This current value is divided by the product of the current gains of all the stages, with the exception of the first stage. In turn, the designer knows the minimum value of collector current required from each of the two output devices. Observe that if Q3 and Q4 are to be capable of swinging almost to $-E_{CC}$, then the voltage across R_{E3} must be small. Assume that this range is 1.5 to 2 V. In turn, this value of voltage must also be present at the collectors of Q1 and Q2. During the idling intervals, a current of 1 mA is to flow through each collector. Accordingly, a simple Ohm's-law calculation will assist the designer in determination of the resistance values for the collector circuits. A portion of each collector resistance is subtracted from R_{C1} and from R_{C2} to provide for the resistance of potentiometer P1. This control serves to balance the relative quiescent currents through the transistors; it is adjusted for 0 V across R_L while the circuit is idling. As a result of dc negative feedback, this adjustment is stabilized.

It follows from our previous discussion that the positive drive voltage must be equal to the negative drive voltage. In turn, R_{B5} is assigned an equal value to R_{E5}. The voltage drops across D4 and the base–emitter junction are equal. Consequently, the currents through R_{B5} and R_{E5} are equal, threby producing a negative drive current in the collector circuit equal to the current flow through R_{B5}. Since the current through R_{B5} is almost identical to the collector current that flows through Q3 and Q4, the negative drive voltage is equal to the positive drive voltage. Power that is available at the collector of Q4 and that is not required at the base of Q5 is dissipated by R_{C4}. The level of quiescent idling current through the output devices is set by action of D5 and D6. If the audio circuit designer desires adjustment flexibility, one of the diodes may be replaced by a potentiometer. Better temperature compensation can be realized if the diodes are replaced with transistor constant-current circuits, as depicted in Fig. 5-8.

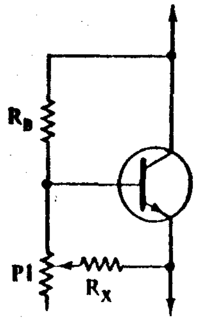


Figure 5-8. Transistor adjustable constant-current circuit.

Refer to Fig. 5-9. This is a simplified version of the circuit that was shown in Fig. 5-7. In this variation, the differential output is taken from one transistor of the pair Q1–Q2. R_{E1} is chosen sufficiently large that a constant-current source is not required; this eliminates the cost of one transistor. The amplified signal voltage is applied to Q3 and is then fed to the quasi-complementary circuit comprising Q4 through Q7. Bootstrapping is provided by capacitor C1 so that Q4 can swing to saturation. Independent

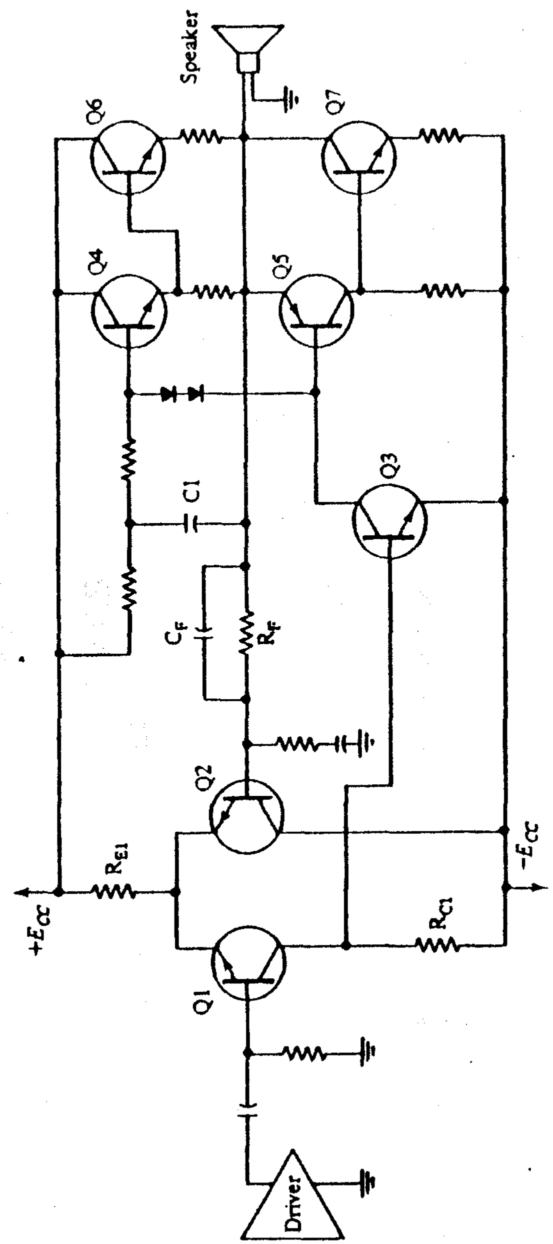


Figure 5-9. Simplified version of the circuit depicted in Fig. 5-7.

drivers with the capability of swinging the signal voltage to the limits of the power-supply voltage function to make bootstrapping unnecessary in the version of Fig. 5-7.

5-6 Full Complementary Output Arrangement

Analysis of the quasi-complementary configuration reveals that the upper two transistors operate as a Darlington pair, whereas the lower two transistors operate as a complementary beta-multiplier pair. Although these are quite similar arrangements, less distortion will result if the audio circuit designer employs identical pairs. He may choose either a complementary pair or a Darlington pair, as exemplified in Fig. 5-10. Either of these circuits can be substituted directly into any of the quasi-complementary arrangements that were illustrated. Either of the circuits can be employed to re-

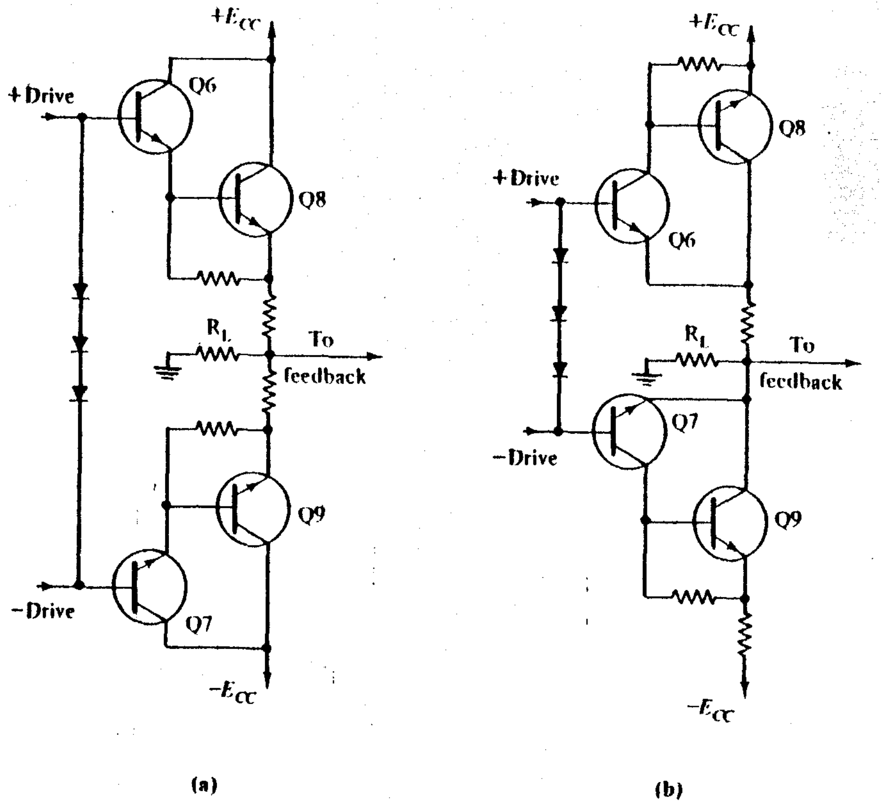


Figure 5-10. Symmetrical output circuit arrangements: (a) dual Darlington; (b) dual beta multiplier.

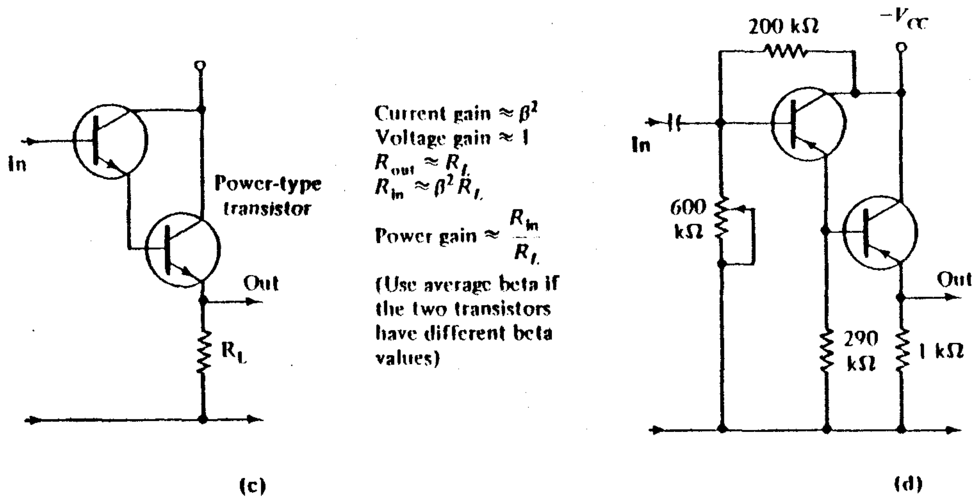
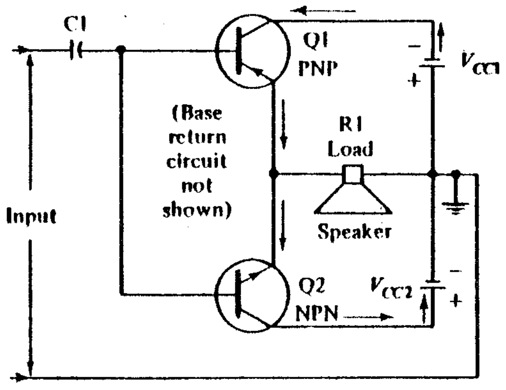


Figure 5-10. (*Continued*) (c) basic Darlington compound (Darlington) circuit; (d) practical arrangement with bias resistors.

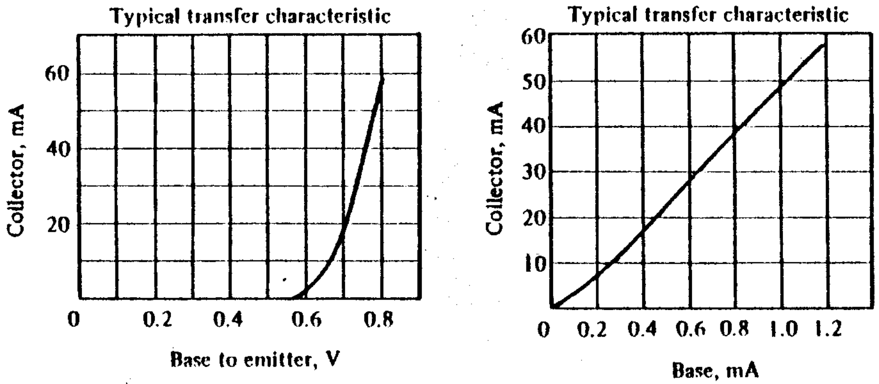
place transistors Q6 through Q8 in the configuration of Fig. 5-7. The same general design principles are observed that have been described.

Various circuit designers prefer the *complementary-symmetry* type of audio power amplifier. It offers the advantages of reduced circuit complexity, eliminates the need for a separate phase-inverter stage, and provides extended frequency response as a result of reduced common-mode conduction. A skeleton circuit diagram for a complementary-symmetry amplifier is shown in Fig. 5-11. Note that Q1 and Q2 operate in the CC mode. Each transistor conducts over one half of an input cycle, because Q1 is a PNP-type, whereas Q2 is an NPN-type transistor. Resultant output circuit action can be followed by analysis of the simplified arrangement depicted in Fig. 5-12. Note that the internal emitter–collector circuit of Q1 is represented by the variable resistor R1, and that of Q2 is represented by the variable resistor R2. While the arm of one variable resistor is in its "off" position, the other resistor arm varies through its range. Next, these relations reverse. Observe that Q1 and Q2 must have closely matched characteristics or the output waveform will contain amplitude distortion.

Power amplifiers usually require transistor operation at power levels that are near thermal runaway conditions. This hazard is aggravated by biasing networks that have marginal stability, but which may be chosen by the circuit designer because of increased operating efficiency. Inasmuch as thermal runaway in a power stage is almost certain to damage or destroy the transistors, the audio circuit designer must give careful attention to worst-case principles in order to eliminate, or at least to minimize, the possibility of thermal runaway. Worst-case conditions include the onset of



(a)



(b)

Figure 5-11. Zero-bias complementary-symmetry configuration: (a) skeleton circuit diagram; (b) typical power-transistor transfer characteristics.

indefinite increase in current gain (h_{fe}), zero base–emitter voltage, minimal load impedance, and saturation current (I_{co}) at a maximum value. In a class B amplifier, the maximum transistor power dissipation occurs at the time when the signal power output is 40 percent of its maximum value. At this time, the power dissipated by each transistor is 20 percent of the maximum power output. For purposes of comparison, in a class A amplifier, maximum power dissipation occurs in the absence of an input signal. The arrangement shown in Fig. 5-11 is one of the basic OTL types of audio-

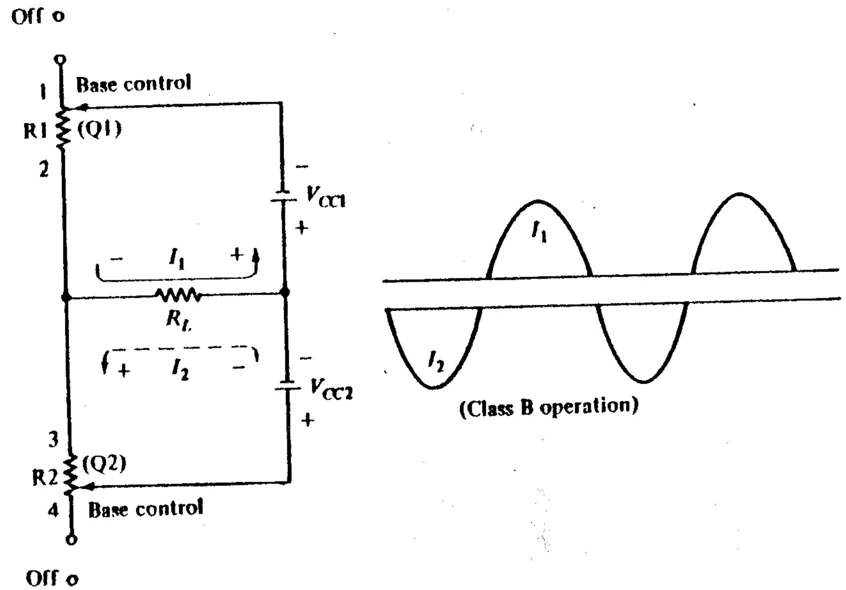


Figure 5-12. Simplified version of a complementary-symmetry output circuit.

output configurations. Most complementary-symmetry amplifiers are operated in class AB. Sufficient forward bias is applied to the transistors to avoid crossover distortion. With reference to Fig. 5-13, a small forward bias is applied; this is an emitter-follower configuration in which R_L, R2, and R3 provide forward bias for Q1 and Q2. Bias resistor R1 is connected in series with the base–emitter junctions of Q1 and Q2. If the two junctions have equal resistance values, the voltage drop across each junction will be one half of the total voltage drop across R1. In practice, the value of R1 is very small, and its unbalancing effect on the input signal to Q2 is negligible. Optimum class AB operation will be obtained only if the characteristics of Q1 and Q2 are reasonably well matched. Mismatch of these transistors can result in a combination of crossover distortion and stretching distortion. If the transistors are well matched and too much forward bias is employed, both of the transistors will develop stretching distortion.

Cascaded Complementary-Symmetry Audio Power Amplifiers

Many audio power amplifiers are designed with complementary-symmetry circuitry in the common-emitter mode. Direct coupling is generally utilized, with an OTL configuration as exemplified in Fig. 5-14. One CE comple-

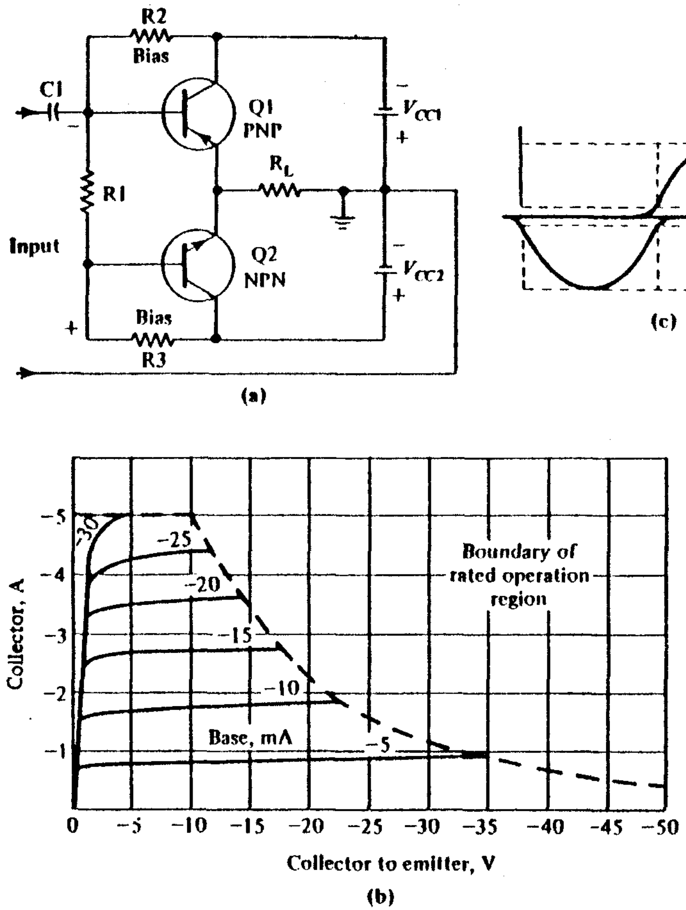


Figure 5-13. Forward-biased CE complementary-symmetry configuration: (a) schematic diagram; (b) maximum power dissipation rating for a typical power transistor; (c) small forward bias minimizes crossover distortion.

mentary-symmetry stage (Q3–Q4) is directly driven by another CE complementary-symmetry stage (Q1–Q2). A single-ended input signal is utilized. When the input signal goes positive, Q1 conducts and Q2 remains nonconducting. Because of the 180-degree phase reversal that occurs from input to output in the CE configuration, Q1's collector is negative going; this causes Q3 to conduct, and Q3's collector is positive going. When the input signal goes negative, Q2 conducts and its collector is positive going. This causes Q4 to conduct and its collector is negative going. Transistors Q1 and Q3 are nonconducting during this interval. Battery V_{BB1} supplies

the required biasing voltages for Q1 and Q3; battery V_{BB2} supplies the required biasing voltages for Q2 and Q4.

Observe that the base–emitter junction of Q3 is connected in series with the collector–emitter circuit of Q1 and V_{BB1}. Accordingly, Q3's emitter is positive with respect to its base (forward bias), and Q1's collector is positive with respect to its emitter, as required for electron flow through Q1. Similar circuit action occurs in the arrangement of Fig. 5-14b, with a different supply-voltage location; the audio signal developed across the speaker provides negative feedback in the input branch of Q1 and Q2, thereby providing a high value of input impedance. Transistors Q3 and Q4 are connected in a CE configuration to match the output resistance of Q1 and Q2. As in amplifier stages with a single transistor, the CE configuration in the cascaded complementary-symmetry arrangement provides higher powet gain to a low-impedance load.

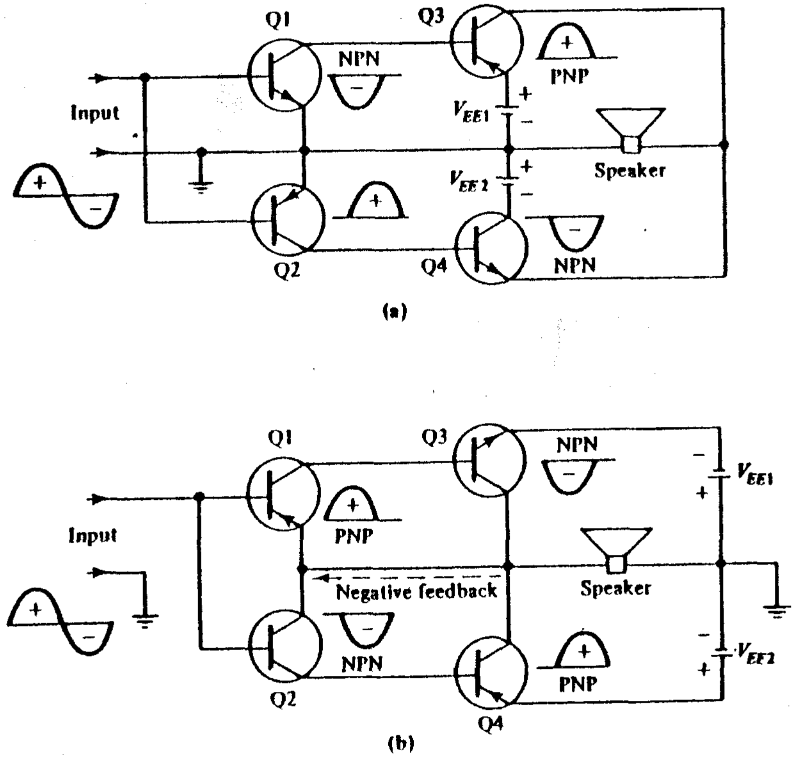


Figure 5-14. Examples of direct-coupled complementary-symmetry stages.

Compound-Connected Complementary-Symmetry Configuration

The current gain, voltage gain, and power gain of a transistor are directly proportional to its short-circuit forward-current amplification factor. This factor is defined as the ratio of the output current to the input current, with the output load equal to zero (short circuited). To obtain maximum gain in an amplifier stage, it is necessary to utilize a transistor that has a high value of short-circuit forward-amplification factor. Most power transistors have an α_{fb} value ranging from 0.940 to 0.985, with an average value of 0.960. Note, however, that no matter what the value of α_{fb} may be, the amplification factor will decrease as the emitter current increases. Hence, when an amplifier stage employs a single transistor, there is inevitably a nonlinear relation of emitter current to collector current. This nonlinearity becomes most prominent at high current levels (peak power output).

This nonlinear relationship results in a reduction of the current amplification factor for a single transistor at high values of emitter current. In a power amplifier that draws heavy emitter current and which is operated near the maximum rated output of the transistor, this variation is aggravated. On the other hand, if two transistors are compound connected, as exemplified in Fig. 5-15, this nonlinear relationship can be minimized. The dashed lines in the diagram enclose the pair of compound-connected transistors. Observe that the base of Q1 is connected to the emitter of Q2, and that the two collectors are connected together. Both of the transistors operate in the CB configuration. The current gain for the compound-

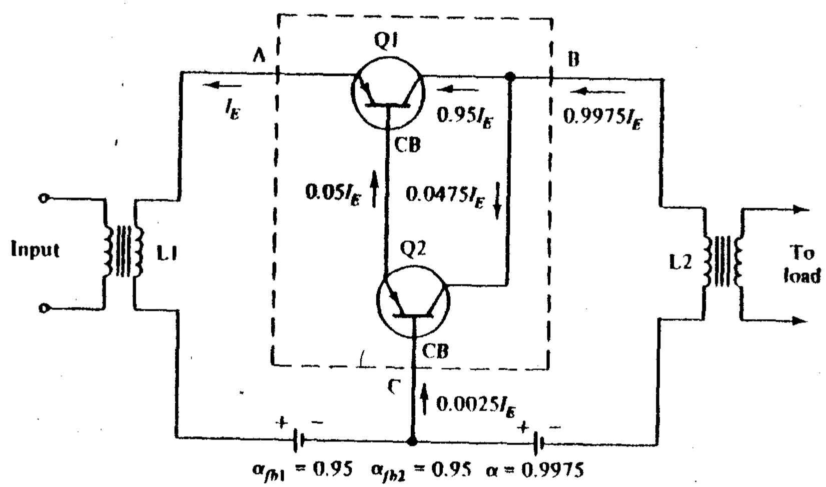


Figure 5-15. Flow diagram for compound-connected transistors.

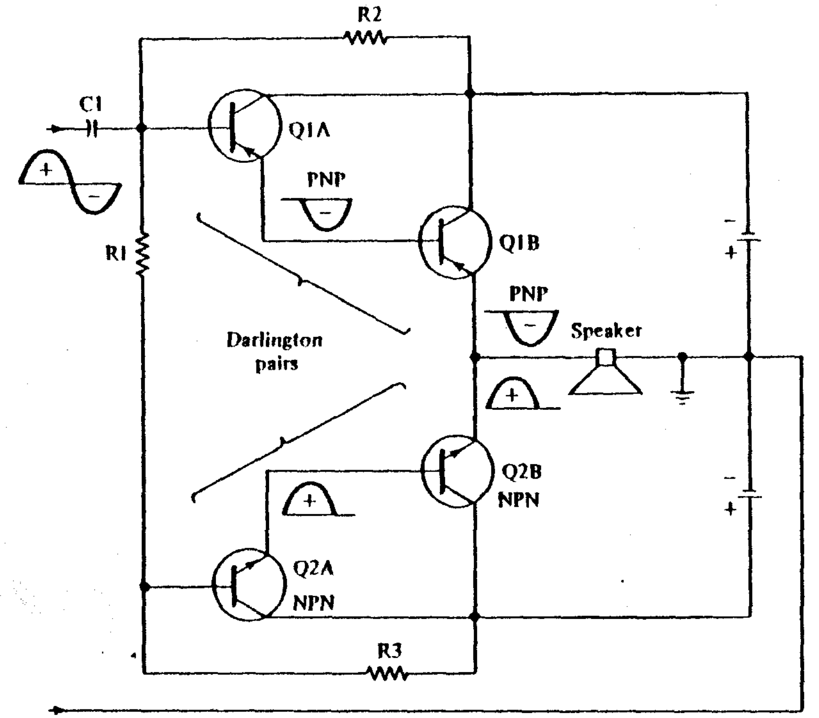
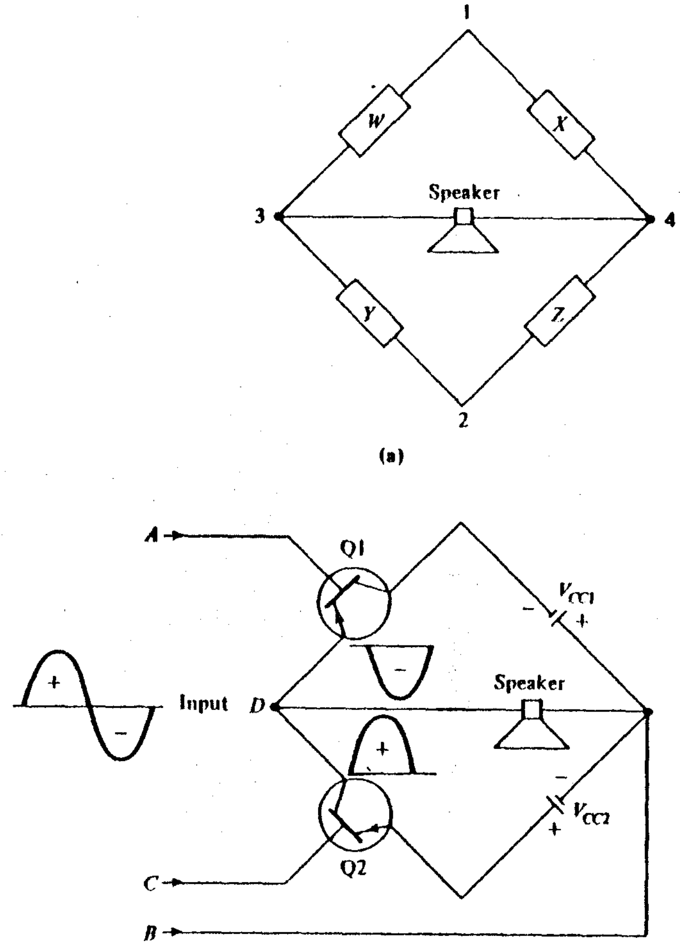


Figure 5-16. Complementary-symmetry compound-connected (Darlington) output amplifier arrangement.

connected transistors is greater than for a single transistor, in addition to providing greatly improved linearity of operation. If we assume that each transistor has a current-gain value of 0.95, their combined current-gain value becomes 0.9975 in the compound connection. This increase corresponds to a beta increase from 19 to 399.

It is not essential that the transistors in Fig. 5-15 have equal current-amplification factors. Also, compound-connected transistors are employed to advantage in single-ended amplifiers, in conventional push–pull amplifiers, and in complementary-symmetry amplifiers. An example of compound-connected (Darlington) transistors in a complementary-symmetry power-amplifier arrangement is shown in Fig. 5-16. This configuration is basically similar to the complementary-symmetry arrangement depicted in Fig. 5-13. However, Q1 is replaced by the compound connection of Q1A and Q1B in Fig. 5-16. Similarly, Q2 is replaced by the compound connection of Q2A and Q2B. Since the pairs of transistors are connected as Darlington pairs, this arrangement is also called a Darlington-pair complementary-symmetry configuration. Inasmuch as the transistors operate



(a)

(b)

Figure 5-17. Basic bridge arrangement, and a bridge with two transistors and two batteries: (a) basic bridge circuit; (b) transistors and batteries operate in bridge arms.

basically in the CC mode, a large amount of negative feedback occurs, and the percentage of distortion is very low.

Complementary-Symmetry Bridge Arrangement

Audio power amplifiers are also configured in the complementary-symmetry bridge arrangement, as exemplified in Fig. 5-17. The arms of the bridge, *W*, *X*, *Y*, *Z*, can be arranged for balance, whether the circuit elements are resistors, capacitors, transistors, or batteries. Bridge balance occurs when, regardless of the voltage applied at points 1 and 2, the voltage drops across the bridge arms are such that zero voltage is developed across points 3 and 4. In other words, no current flows through the speaker. Thus the resistors in the basic bridge circuit can be replaced by transistors Q1 and Q2, and by batteries V_{CC1} and V_{CC2}, as depicted in Fig. 5-17b. If, in turn, the transistors are biased to draw equal emitter currents (or to draw no emitter–collector current) under quiescent (idling) conditions, the bridge is balanced and zero current flows through the speaker. On the other hand, if an input signal causes either of the transistors to conduct more current than the other, the bridge becomes unbalanced and current flows through the speaker. Sine-wave signals applied to points *A–B* and *C–D* that are 180 degrees out of phase with each other will result in an amplified sine-wave signal across the speaker. If the transistors are operated in class A or B, no dc current will flow through the speaker.

It is undesirable to have a dc current flow through a speaker voice coil because this produces cone offset, which tends to distort the reproduced sound. Observe that, if a conventional push–pull power amplifier were utilized, the speaker voice coil would need to be center tapped, and only one half of the coil could be used for each half-cycle of the input signal. Such center-tapped operation entails reduced efficiency in conversion of electrical energy into sound energy. Observe the bridge configuration depicted in Fig. 5-18. All four arms of the bridge consist of transistors. Q1 and Q3 are PNP types, whereas Q2 and Q4 are NPN types. This configuration has an advantage in that no part of the input circuit or of the output circuit need be operated at ground potential. Moreover, a complementary-symmetry bridge amplifier can be energized from a single-ended driver.

Assume under idling conditions that all the transistors in Fig. 5-18 are zero biased and that they draw no current (class B operation). If the transistors draw no current, there is no completed circuit across V_{CC} and no current flows through the speaker. Next, if an input signal drives point *A* negative with respect to point *B*, Q1 and Q4 will become forward biased, and Q2 and Q3 will become reverse biased. Therefore, Q1 and Q4 will conduct; electrons will flow from the negative battery terminal through Q1 collector-to-emitter, through the speaker, through Q4 emitter-to-collector, and back to the positive battery terminal. This path for current flow is indicated by the solid-line arrows, and a voltage of the indicated polarity is dropped across the speaker.

Next, if an input signal causes point *A* to become positive with respect to point *B*, Q3 and Q2 will conduct, and Q1 and Q4 will be reverse biased. In turn, the path of electron current is indicated by the dashed-line arrows. A voltage of the indicated polarity is dropped across the speaker. Assume

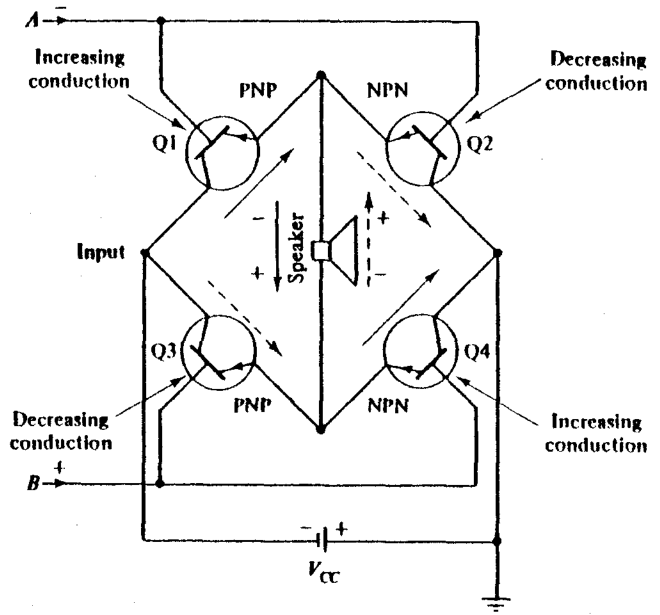


Figure 5-18. Complementary-symmetry bridge arrangement.

under idling conditions that the transistors are biased to draw equal currents in class A operation. (This bias circuit is not shown in the diagram.) Under this condition of operation, electrons emerge from the negative battery terminal. One half of the electron current flows through Q1 collector-to-emitter and through Q2 emitter-to-collector, and thence into the positive battery terminal. The other half of the electron current flows through Q3 collector-to-emitter and through Q4 emitter-to-collector, and thence back to the positive battery terminal. Thus the bridge is balanced and there is no current flow through the speaker.

Assume next that an input signal causes point A to become more negative with respect to point B. Transistors Q1 and Q4 become more forward biased and draw more collector current. Transistors Q2 and Q3 become less forward biased and draw less collector current. Accordingly, the bridge becomes unbalanced, and the difference in current between Q1 and Q2 flows through the speaker in the direction of the solid-line arrow and through Q4 into the positive battery terminal. A voltage is accordingly developed across the speaker with the indicated polarity. If the input signal causes point A to become positive with respect to point B, Q2 and Q3 become forward biased and draw more collector current. Transistors Q1 and Q4 become less forward biased and draw less collector current. The difference between the Q3 and Q4 currents flows through the speaker in

the direction of the dashed-line arrow, and develops a voltage drop with the indicated polarity.

Summary of Power-Amplifier Characteristics

1. The noise factor of an amplifier is defined as the quotient of the signal-to-noise ratio at the output of the amplifier and the signal-to-noise ratio at the input of the amplifier.
2. The noise factor of a transistor increases as its collector voltage is increased. The noise factor, or noise figure, is stated for a given bandwidth and is equal to the ratio of total noise at the output to the noise at the input.
3. A CE amplifier with degeneration develops a comparatively high value of input impedance.
4. Zero-biased class B push–pull amplifiers produce crossover distortion.
5. Crossover distortion can be minimized or eliminated by operating a push–pull amplifier in class AB.
6. Stretching distortion results from application of excessive forward bias in class AB operation.
7. Complementary-symmetry push–pull amplifiers have numerous advantages over related configurations and are widely used. However, transistors must have closely matched characteristics to minimize distortion.
8. Compound-connected or Darlington-connected transistors provide a comparatively high amplification factor.
9. Darlington-connected transistors are frequently used in complementary-symmetry amplifier arrangements to obtain high gain and high power output with minimum circuit complexity.

5-7 Examples of Innovative Circuit Design

It is instructive to observe some examples of innovation in audio power amplifier circuit design. As an illustration, class D high-efficiency amplifiers have been devised. The basic principle of class D amplification is shown in Fig. 5-19. Essentially, the input wave form is converted into an amplitude-modulated pulse train that drives the power-output stage. After power amplification of the signal, the pulses are integrated to reconstitute the original input wave form. Higher efficiency is obtained because a power transistor can be driven to considerably higher peak levels when a signal that consists of comparatively narrow pulses is processed.

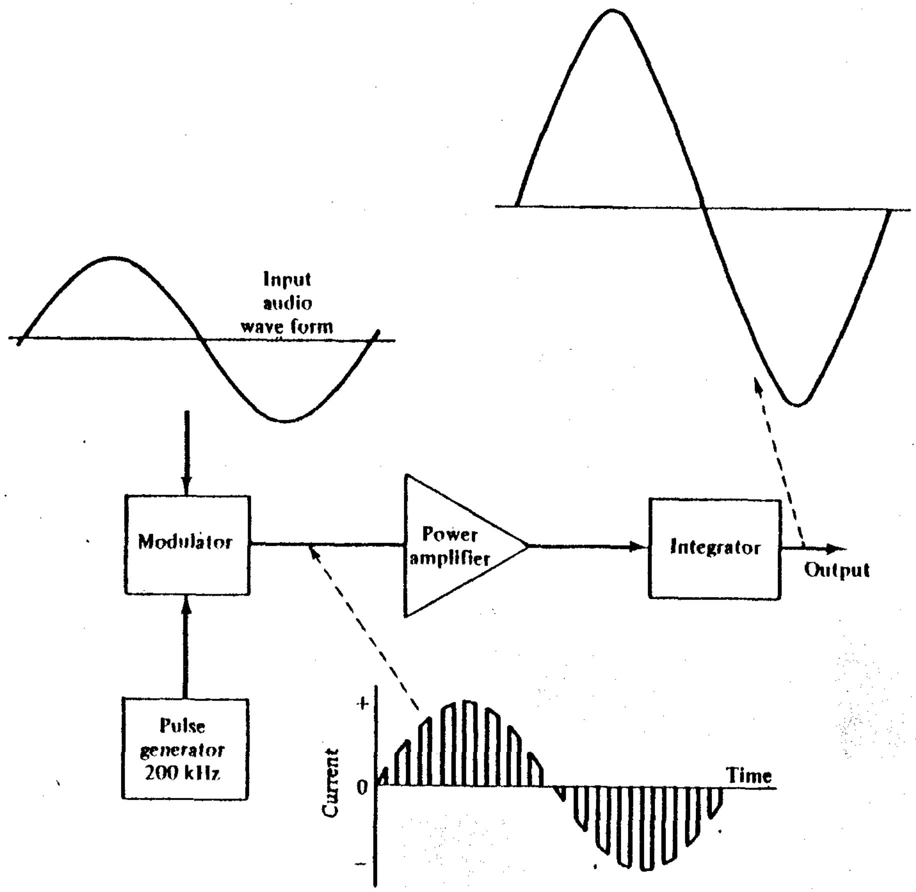


Figure 5-19. Principle of a class-D amplifier.

As another example, a pulse-width modulated amplifier also operates at higher efficiency than does a conventional power amplifier. A pulse-width modulated (PWM) wave form is exemplified in Fig. 5-20. This operating technique employs pulses of uniform amplitude and varying width. A narrow pulse corresponds to a low-amplitude input level, whereas a wide pulse corresponds to a high-amplitude input level. A method of PWM wave-form generation is depicted in Fig. 5-21. After power amplification of the PWM waveform, it is integrated to reconstitute the original input waveform. Operating efficiency is comparatively high, because the power transistors can be driven to a relatively high peak level by the pulse waveform.

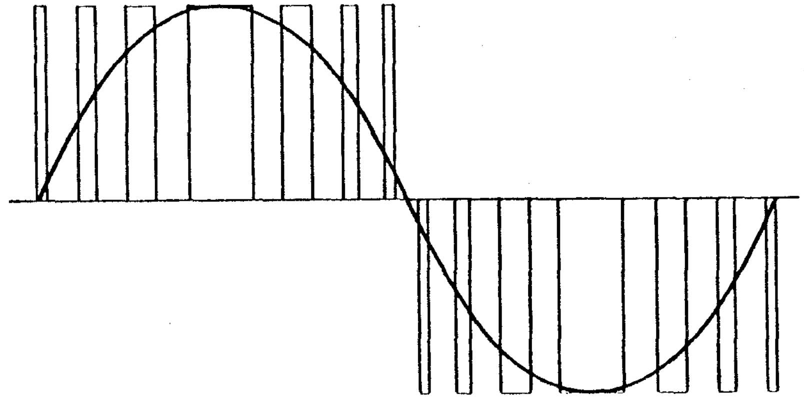


Figure 5-20. Example of pulse-width modulated (PWM) wave form.

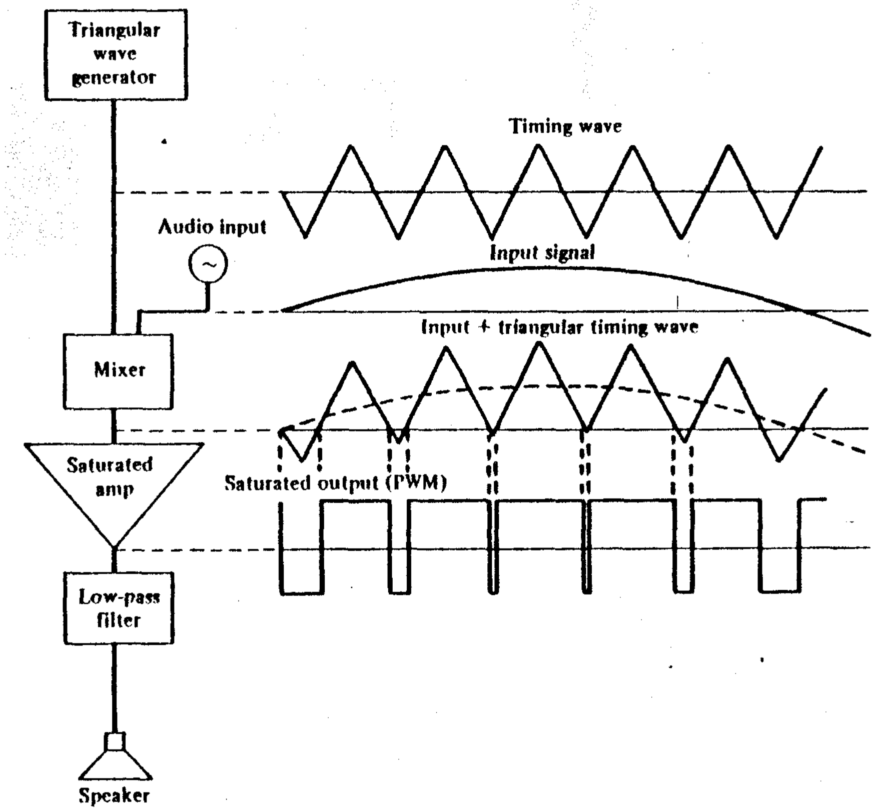


Figure 5-21. Method of generating a PWM wave form.

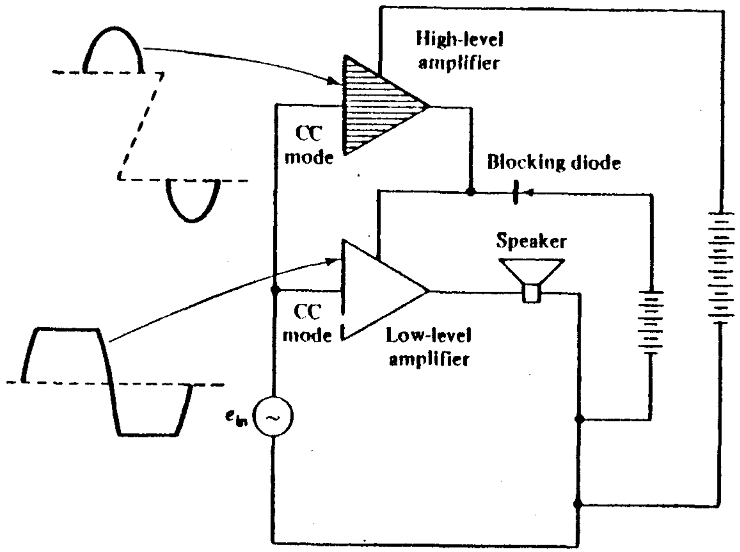


Figure 5-22. Basic plan of a class G amplifier.

A new design of high-efficiency high-fidelity amplifier is termed the class G configuration (Fig. 5-22). A skeleton circuit for this form of amplifier is shown in Fig. 5-23. V_{in} denotes the input audio signal; the output is developed across R_L (usually a speaker load). The input signal voltage is applied to the bases of transistors Q1 and Q2. Supply voltage V_1 is applied through diode D1 to the emitter of Q1 and to the collector of Q2. In turn, the collector of Q1 is connected to a higher value of supply voltage, V_{CC}. When the input signal voltage V_{in} is less than V_1, Q1 has a reverse

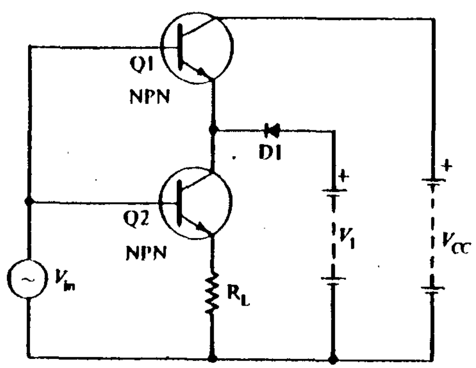


Figure 5-23. Skeleton circuit for a class G amplifier.

base–emitter bias, ...llector current is cut off. Current flowing through R_L is obta... ... via diode D1.

Next, if the sig... ...ge increases to a value greater than V_1, but less than V_{CC}, Q1 becom... forward biased. Its collector current is turned on. Now current flowing through R_L is obtained from V_{CC} via Q1. Diode D1 al... has the function of preventing current flow from V_{CC} back into source V_1. The operating efficiency of the class G configuration is considerably greater than that of a class B arrangement. However, in its most basic form, a class G amplifier develops an objectionable amount of distortion. This distortion is seen in the waveform depicted in Fig. 5-24. This distortion results from the circumstance that Q1 is not turned on until the amplitude of the input signal voltage exceeds the collector voltage of Q2 by a value equal to the base–emitter voltage of Q1. In turn, Q2 starts to saturate before Q1 begins conduction, and an irregularity is introduced into the output signal waveform.

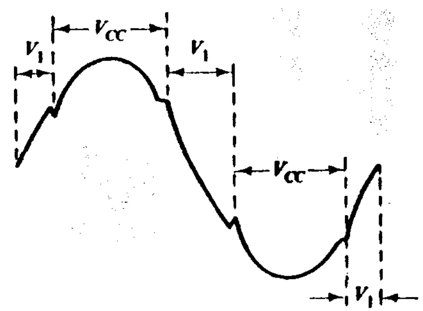


Figure 5-24. Distorted output wave form (changeover distortion) developed by a simplified class G amplifier configuration.

This *changeover* distortion can be reduced by adding another diode, D2, as shown in Fig. 5-25. In turn, when V_1 has a smaller value than the input signal voltage, the potential between the collector and emitter of Q2 is lower than the saturation level by an amount equal to the threshold value of D2, and Q2 remains unsaturated. Diode D3 also serves an essential function. Inasmuch as a reverse bias voltage is applied between base and emitter of Q1 when the input signal voltage is lower than V_1, the base–emitter junction of Q1 must be able to withstand a reverse voltage that is greater than V1. Thus D3 protects the junction against breakdown. A skeleton class G push–pull amplifier configuration is seen in Fig. 5-26. Diodes have been omitted from the base circuits to avoid unnecessary detail. Note that Q3 and Q4 are PNP transistors, whereas Q1 and Q2 are

NPN transistors. In other words, this is a complementary configuration. Its efficiency is considerably greater than that of a conventional complementary arrangement.

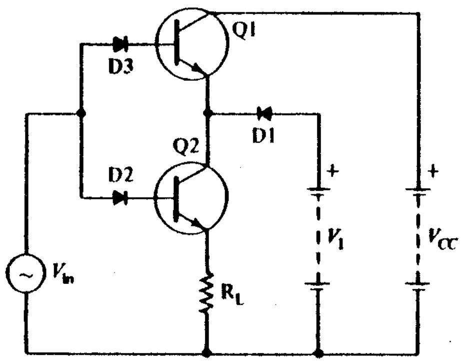


Figure 5-25. Distortion is reduced by action of diode D2.

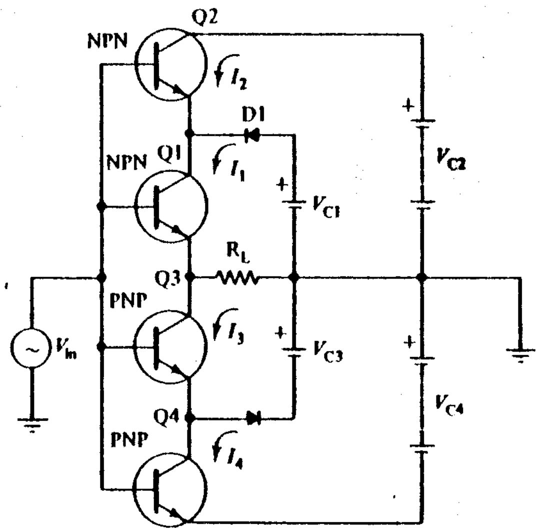


Figure 5-26. Skeleton class G push-pull amplifier configuration.