Audio Power Amplifiers

The quality of an audio power amplifier is measured by its ability to provide high-fidelity reproduction of audio program material over the full range of audible frequencies. The amplifier is required to increase the power level of the input to a satisfactory output level with little distortion, and the sensitivity of its response to the input signals must remain essentially constant throughout the audio-frequency spectrum. Moreover, the input-impedance characteristics of the amplifier must be such that the unit does not load excessively and thus adversely affect the characteristics of the input-signal source.

Silicon power transistors offer many advantages when used in the power-output and driver stages of high-power audio amplifiers. These devices may be used, either as discrete components or as building-block elements in power hybrid circuits, over a wide range of ambient temperatures to develop up to hundreds of watts of audio-frequency power to drive a loudspeaker system. The following paragraphs describe the basic factors that must be considered and the important concepts and techniques employed in the design of transistor audio power amplifiers.

CLASSES OF OPERATION

A circuit designer may select any one of three classes of operation for transistors used in linear-amplifier applications. This selection is made on the basis of a combination of such factors as required power output, dissipation capability, efficiency, gain, and distortion characteristics.

The three basic classes of operation (class A, class B, and class C) for linear transistor amplifiers are defined by the operating point of the transistor. In class A operation, the active element conducts for the entire input cycle. In class B operation, the active element conducts for 180 degrees of an input cycle and is cut off during the remainder of the time. In class C operation, the active element conducts for some amount less than 180 degrees of an input cycle. The following paragraphs discuss the distinguishing features of class A and class B operation. In general, because of the high harmonic distortion introduced as a result of the short conduction angle, class C operation is used primarily in rf-amplifier applications in which it is practical to use tuned output circuits to eliminate the harmonic components. For this reason, class C operation is not discussed further.

Class A Operation

Class A amplifiers are used for linear service at low power levels. When power amplifiers are used in this class of operation, the amplifier output is usually transformer-coupled to the load circuit, as shown in Fig. 216. At low power levels, the class A amplifier can also be coupled to the load by resistor, capacitor, or direct coupling techniques.

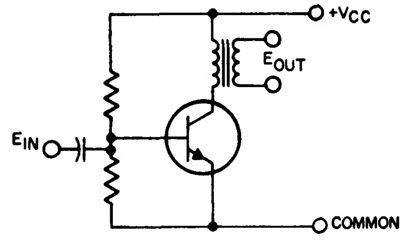

Fig.216 - Basic class A, transformer-coupled amplifier.

There is some distortion in a class A stage because of the nonlinearity of the active device and circuit components. The maximum efficiency of a class A amplifier is 50 per cent; in practice, however, this efficiency is not realized. The class A transistor amplifier is usually biased so that the quiescent collector current is midway between the maximum and minimum values of the output-current swing.

Collector current, therefore, flows at all times and imposes a constant drain on the power supply. The consistent drain is a distinct disadvantage when higher power levels are required or operation from a battery is desired.

Class B Operation

Class B power amplifiers are usually used in pairs in a push-pull circuit because conduction is not maintained over the complete cycle. A circuit of this type is shown in Fig. 217. If conduction in each device occurs during

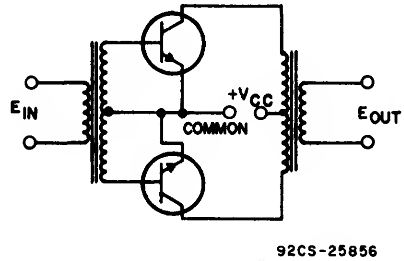

Fig.217 - Basic class B, push-pull transformer-coupled amplifier.

approximately 180 degrees of a cycle and the driving wave is split in phase, the class B stage can be used as a linear power amplifier. The maximum efficiency of the class B stage at full power output is 78.5 per cent when two transistors are used. In a class B amplifier, the maximum power dissipation is 0.203 times the maximum power output and occurs at 42 per cent of the maximum output.

Transistors are not usually used in true class B operation because of an inherent nonlinearity, called **cross-over distortion**, that produces a high degree of distortion at low power levels. The distortion results from the nonlinearities in the transistor characteristics at very low current levels. For this reason, most power stages operate in a biased condition somewhat between class A and class B.

This intermediate class is defined as **class AB**. Class AB transistor amplifiers operate with a small forward bias on the transistor to minimize the nonlinearity. The quiescent current level, however, is still low enough so that class AB amplifiers provide good efficiency. This advantage makes class AB amplifiers an almost universal choice for high-power linear amplification, especially in battery-operated equipment.

DRIVE REQUIREMENTS

In class A amplifiers, the output stage is usually connected in a common-emitter configuration. The relatively low input impedance that generally characterizes this type of configuration may result in a severe mismatch with the output impedance of the driver transistor. Usually, at low power levels, RC coupling is used and the loss is accepted. It may be advantageous in some circuits, however, to use an emitter-follower between the driver and the output stage to obtain an improved impedance match.

Class AB amplifiers have many types of output connections. One form is the **transformer-coupled output stage** illustrated in Fig. 218. Again, the common-emitter circuit is

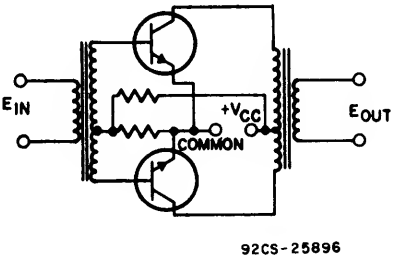

Fig.218 - Basic class AB, push-pull, transformer-coupled amplifier.

usually employed because it provides the highest power gain. The load circuit is never matched to the output impedance of the transistor, but rather is fixed by the available voltage swing and the required power output. The transformer is designed to reflect the proper impedance to the output transistors so that the desired power output can be achieved with a specific supply voltage.

The use of transformer coupling from the driver to the input of the power transistor assures that the phase split required for push-pull operation of the output stages and any necessary impedance transformation can be readily achieved. Output transformer coupling provides an easy method for matching several values of load impedance, including those encountered in sound-distribution systems. For paging service, servo motor drive, or other applications requiring a limited bandwidth, the transformer-coupled output stage is very useful. However, there are disadvantages to the use of transformer coupling. One disadvantage is the phase shift encountered at low- and high-frequency extremes, which may lead to unstable operation. In addition, the output transistors must be capable of handling twice the supply voltage because of the transformer requirements.

Another type of transistor output circuit is the **series-connected output stage**. With this type of circuit, the transistors are connected in series across the supply and the load circuit is coupled to the midpoint through a capacitor. There must be a 180-degree phase shift between the driving signals for the upper and lower transistors. A transformer can be used in this application provided that the secondary consists of two separate windings, as shown in Fig. 219. Other forms of phase splitting can be

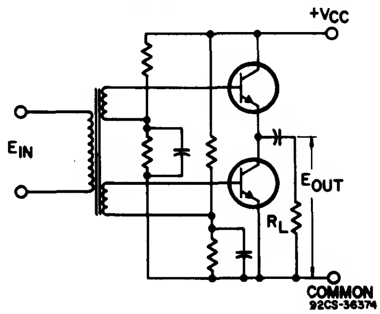

Fig.219 - Class AB, push-pull amplifier with series output connection.

used; all have problems such as insufficient swing or poor impedance matching. Capacitor output coupling also has disadvantages. A low-frequency phase shift is usually associated with the capacitor, and it is difficult to obtain a capacitor that is large enough to produce an acceptable low-frequency output. These disadvantages can be alleviated by use of a split supply and by connection of the load between the transistor midpoint and the supply midpoint with the return path through the power-supply capacitors. The power-supply capacitors must be large enough to prevent excessive ripple.

Complementary amplifiers are produced when p-n-p and n-p-n transistors are used in series. A capacitor can be used to couple the amplifier output when a single supply is used, or direct coupling can be employed when a split power supply is used, as shown in Fig. 220. Because no phase inversion is needed in the driving circuit for this output configuration, there are definite advantages in the simplicity of the design. One disadvantage of this type of amplifier is that the driver must be a class A stage which may have a high dissipation. This dissipation can be reduced, however, by use of a Darlington compound connection for the output stage. This compound connection reduces the driving-stage requirement. A method of overcoming this disadvantage completely is to use a quasi-complementary configuration. In this configuration, the output transistors are a pair of p-n-p or n-p-n transistors driven by a complementary pair. In this manner the n-p-n/ p-n-p drivers provide the necessary phase inversion. The driving transistors are connected directly to the bases of the output transistors, as illustrated in Fig. 221.

Adequate drive may be a problem with the transistor pair shown in the upper part of the quasi-complementary amplifier unless suitable techniques are used to assure that this pair saturates. Care must also be taken when split supplies are used to assure that any ripple on the lower supply is not introduced into the predriving stages by this technique. The advantage of a split supply is that it makes possible direct connection to the load and thus

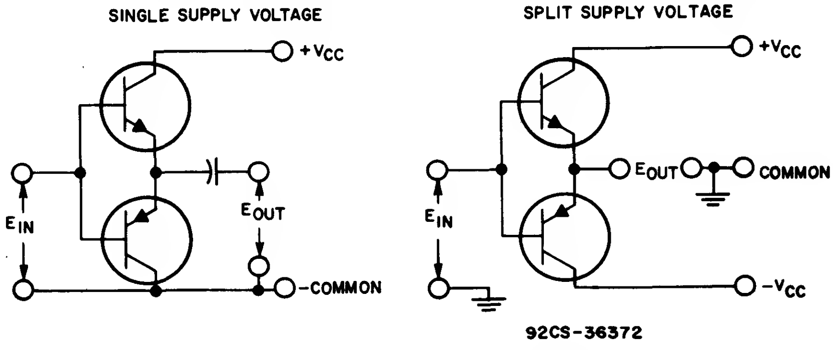

Fig.220 - Circuit arrangements for operation of complementary output stages (a) from single dc supply; (b) from symmetrical dual (positive and negative) supplies.

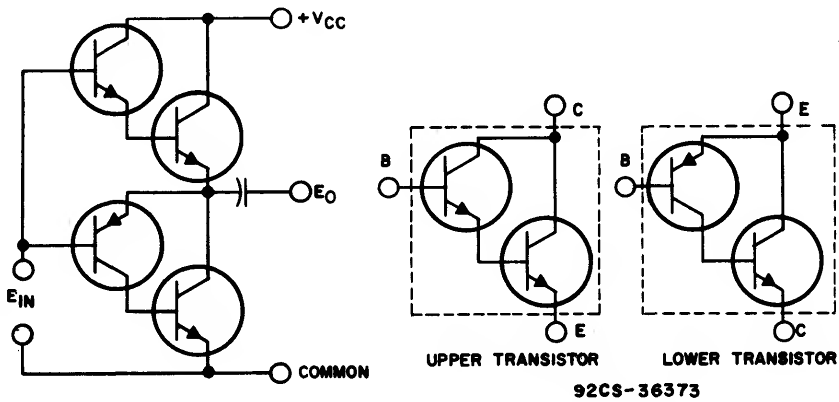

Fig.221 - Compound output stage in which output transistors are driven by complementary driver transistors: (a) over-all circuit; (b) upper transistor pair; (c) lower transistor pair.

improves low-frequency response.

To this point, **phase inversion** has been mentioned but not discussed. Phase inversion may be accomplished in many ways. The simplest electronic phase inverter is the single-stage configuration. This configuration can be used at low power levels or with high-gain devices when the limited drive capability is not a drawback. At higher power levels, some impedance transformation and gain may be required to supply the drive needed. There are several complex phase-splitting circuits; a few of them are shown in Fig. 222.

EFFECT OF OPERATING CONDITIONS ON CIRCUIT DESIGN

Some additional design problems involve the consideration of thermal stability, high line voltage, line-voltage transients, excessive drive, ambient temperature, load impedance,

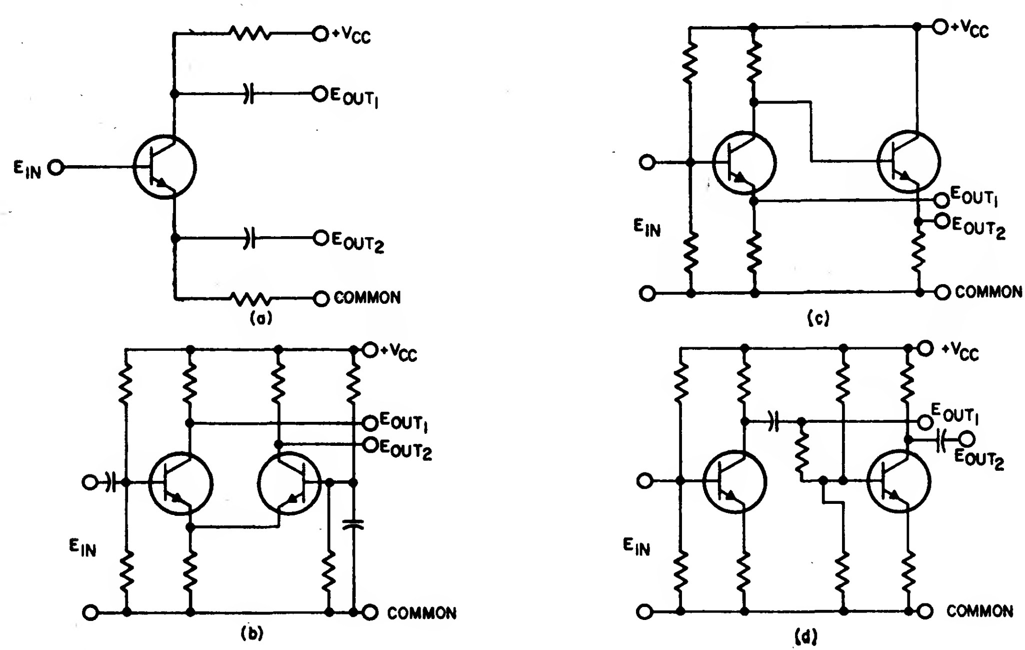

Fig.222 - Basic phase-inverter circuits: (a) single-stage phase-splitter type; (b) two-stage emitter-coupled type; (c) two-stage low-impedance type; (d) two-stage similar-amplifier type.

and other factors that may subject the transistors to abnormal high-stress conditions. A prime consideration is the maximum power dissipation at high supply voltage. Thermal stability is another problem that is often difficult to control. The problem is complex because the base-to-emitter voltage V_{BE} of a transistor decreases with an increase in junction temperature at a constant level of collector current. Therefore, if the V_{BE} of the transistor is held constant, the collector current I_C increases as the junction temperature rises. This process is regenerative because the dissipation increases with an increase in the value of I_C. One solution is to place a resistor in series with the emitter lead. This approach is not the best solution to the problem, however, because the use of the resistor increases circuit losses. A decrease in the loss may be obtained if the resistor is bypassed. Another approach is to use a thermistor or similar device which, when properly connected, reduces the base drive at high temperatures. This approach improves the stability without increasing the circuit loss.

The collector-to-base leakage current I_{CBO} can also be a problem because a fraction of this current is multiplied by the transistor h_{fe} and appears as a component of the collector-to-emitter current. In general, the value of I_{CBO} is in the order of microamperes in silicon devices and milliamperes in germanium devices. This leakage current is composed of two components. One component is caused by surface leakage and is unpredictable in its variations with temperature. It increases with voltage and may even decrease with increasing temperature. The other component is a function of the device material and geometry. This component approximately doubles with every 7° C temperature rise in silicon devices, and approximately doubles for every 10° C temperature increase in germanium devices. This component may also be voltage-dependent.

The total leakage is of interest to the circuit designer because it can be the mechanism for thermal-runaway problems. An increase in this leakage increases the total base current and thus causes an increase in collector current and dissipation. The increase in collector current and dissipation causes a rise in temperature which may produce a regenerative cycle that leads to thermal runaway. If an external resistor is connected between the base and emitter, some of this leakage current is shunted from the base, and the thermal-stability problem is reduced.

Another potential source of trouble in amplifiers is the feedback loop. Feedback is used to reduce distortion and extend the frequency range of the amplifier. The feedback loop usually encloses several if not all of the amplifier stages and can cause several problems. When transformer coupling is used, phase shifts may occur at the high- or low-frequency extremes; a positive voltage may then be fed back and cause oscillation. High-signal-level transients may cause the value of the transformer inductances and other components to change and become unstable so that they initiate oscillation. A similar condition can occur at low frequencies when capacitor-coupled transformerless designs are used.

Excessive drive levels at high frequencies can cause dissipation problems. An excessive drive level forces the output stages to saturate before the peak of the input signal is reached. This additional drive lengthens the storage time which, at high frequencies, may approach the period of the drive signal. Under this condition, two results occur: First, feedback does not increase after the point where the output stage saturates. This condition permits the drive signal to increase. Second, one transistor may not turn off until the second has been turned on. In series-type output stages, the second transistor is turned on with the full supply voltage present. This condition can lead to forward-bias second-breakdown problems.

Another potential source of difficulty with amplifiers occurs when the output is open- or short-circuited. Transformer-coupled output stages are particularly susceptible to operational problems with no load. Without a load, the transistors operate into a purely inductive load line and the probability of reverse-bias second breakdown must be considered. In series-type output stages, the major problem arises under short-circuit load conditions. As a result of the short circuit, feedback is removed and an open-loop gain condition exists together with the excessive-drive-condition problems previously mentioned. It is advisable to use some form of fast-acting overload protection for the power transistor; a fuse is usually not fast enough in this application.

Some frequency exists at which the gain of any transistor begins to decrease. This decrease in gain can be corrected over the required frequency range by use of feedback or a higher-frequency device. Roll-off of the frequency response of the preamplifier stages at some point prior to the limiting value of the frequency characteristics of the transistor is necessary. This technique assures that the drive is limited to a safe value by the input stage so that even the drivers are not affected by the high dissipation mentioned previously.

Several other factors that should be considered in the design of amplifiers for audio-frequency service include the frequency response desired, gain, optimum load, noise, and power output needed.

BASIC CIRCUIT CONFIGURATIONS

The selection of the basic circuit configuration for an audio power amplifier is dictated by the particular requirements of the intended application. The selection of the basic circuit configuration that provides the desired performance most efficiently and economically is based primarily upon the following factors: power output to be supplied, required sensitivity and frequency-response characteristics, maximum allowable distortion, and capabilities of available devices.

Class A Transformer-Coupled Amplifiers

Fig. 223 shows a three-stage class A transformer-coupled audio amplifier that uses dc feedback (coupled by R_1, R_2, R_3, R_4, and C_1) from the emitter of the output transistor to the base of the input transistor to obtain a stable operating point. An output capability of 5 watts with a total harmonic distortion of 3 per cent is typical for this type of circuit. In general, this output level is the upper limit for class A amplifiers because the power dissipated by the output transistor in such circuits is more than twice the output power. For this reason, it is economically impractical to use class A audio amplifiers to develop higher levels of output power. A circuit such as the one shown in Fig. 223 usually requires no over-all feedback unless extremely low distortion is required. Local feedback in each stage is adequate; amplifiers of this type, therefore, are usually very stable.

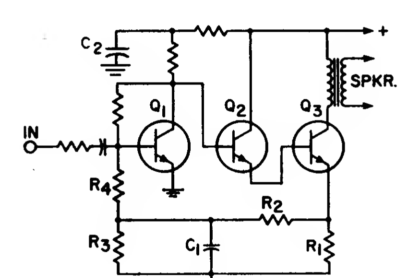

Fig.223 - Three-stage transformer-coupled, class A amplifier.

Class AB Push-Pull Transformer-Coupled Amplifiers

At power-output levels above 5 watts, the operating efficiency of the circuit becomes an important factor in the design of audio power amplifiers. The circuit designer may then consider a class AB push-pull amplifier for use as the audio-output stage.

Fig. 224 shows a class AB push-pull transformer-coupled audio-output stage. Resistors R_1, R_2, and R_3 form a voltage divider that provides the small amount of transistor forward bias required for class AB operation. The transformer type of output coupling used in the circuit is advantageous in that a suitable output transformer can be selected to match the audio system to any desired load impedance. This feature assures maximum transfer of the audio-output power to the load circuit, which is especially important in sound-distribution systems that use high-impedance transmission lines to reduce losses. A

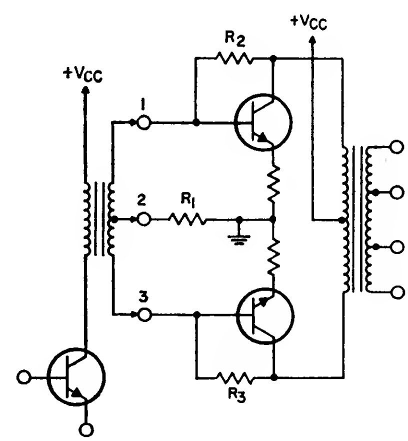

Fig.224 - Class AB, push-pull, transformer-coupled audio output stage.

major disadvantage of transformer output coupling is that it tends to limit the amplifier frequency response, particularly at the low-frequency end. Variations in transformer impedance with frequency may produce significant phase shifts in the signal at both frequency extremes of the amplifier response. Such phase shifts are potential causes of amplifier instability if they occur within the feedback loop. Open-circuit stability is always a problem in designs that use output transformers because the gain increases sharply when the load is removed. If too much over-all feedback is employed, the amplifier may oscillate. The local feedback caused by the bias arrangement of R_2 and R_3 helps to eliminate this problem.

Push-pull output stages, which use identical output transistors, require some form of phase inversion in the driver stage. In the circuit shown in Fig. 224, a center-tapped driver transformer is used for this purpose. The requirements of this transformer depend upon the power levels involved, the bandwidth required, and the distortion that can be tolerated. This transformer also introduces phase-shift problems that tend to cause instabilities in the circuit when high levels of feedback are employed. Phase-shift problems are substantially reduced when the output stage is designed to operate at low drive requirements. The reduced drive requirements can be achieved by use of the Darlington circuit shown in Fig. 225. Resistors R_1 and R_2

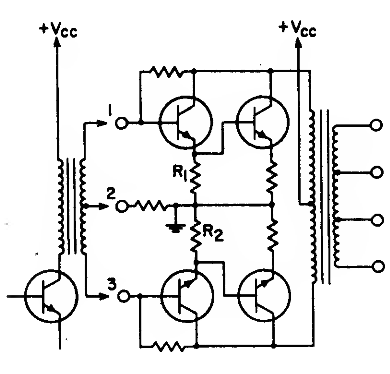

Fig.225 - Class AB, push-pull, transformer-coupled audio output stage in which Darlington pairs are used to reduce drive requirements of output transistors.

shunt the leakage of the driver and also permit the output transistors to turn off more rapidly. Impedance levels between the class A driver and the output stage can be easily matched by the use of an appropriate transformer turns ratio.

An alternative method of phase inversion is to use a transistor in a phase-splitter circuit, such as those shown in Fig. 222 and described later in the discussion on **Phase Inverters**. Unlike the center-tapped transformer method, impedance matching may be a problem because the collector of the driver, which has a relatively high impedance, operates into the low input impedance of the output stage. One solution is to reduce the output impedance of the driver stage by the use of smaller resistors. The resultant increase in collector current, however, also increases the dissipation. Moreover, very large coupling capacitors are necessary for the achievement of good low-frequency performance. The nonlinear impedance exhibited by the input of the output transistor causes a dc voltage to be produced across the capacitor under high signal levels. An alternate solution is to use a Darlington pair to increase the input impedance of the output stage.

Class AB Series-Output Amplifiers

For applications in which low distortion and wide frequency response are major requirements, a transformerless approach is usually employed in the design of audio power amplifiers. With this approach, the common type of circuit configuration used is the series-output amplifier.

The class-AB-operated n-p-n transistors used in the series-output circuits shown in Fig. 226 require some form of phase inversion of the drive signal for push-pull operation. A common approach is to use a driver transformer that has split secondary windings, as shown in Fig. 227. The split secondary windings are required because of the mode in which each of the series output transistors operates.

If ground were used as the drive reference for both secondary windings of the circuit shown in Fig. 227, transistor Q_1 would operate as an emitter-follower and would provide gain of somewhat less than unity. Transistor Q_2, however, is connected in a common-emitter configuration which can provide substantial voltage gain. For equal output-voltage swings in both directions, the drive input to transistor Q_1 is applied directly across the base and

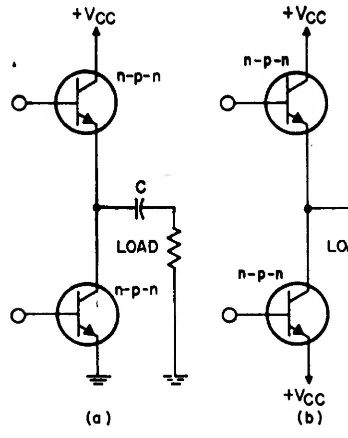

Fig.226 - Circuit arrangements for operation of series output circuit from (a) a single dc supply and (b) symmetrical dual supplies.

emitter terminals. Transistor Q_1 is then effectively operated in a common-emitter configuration (although there is no phase reversal from input to output) and has a voltage gain equal to that of transistor Q_2.

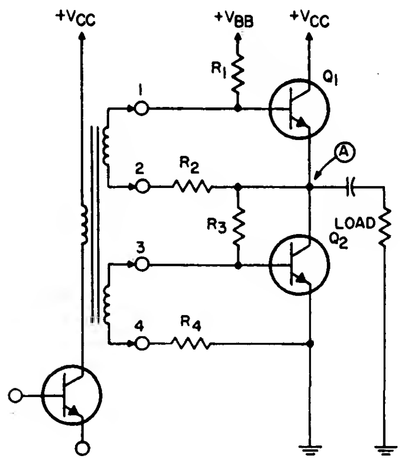

Fig.227 - Circuit using a driver transformer that has split secondary windings to provide phase inversion for push-pull operation of a series-output circuit.

The disadvantages of a driver transformer discussed previously also apply to the circuit shown in Fig. 227. In addition, coupling through interwinding capacitances can adversely affect the performance of the circuit. Such coupling is particularly serious because at both ends of the upper secondary (terminals 1 and 2) the ac voltage with respect to ground is approximately equal to the output voltage. During signal conditions, when output transistor Q_1 is turned on, this coupling provides an unwanted drive to Q_1. The forward transistor bias required to maintain class AB circuit operation is provided by the resistive voltage divider R_1, R_2, R_3, and R_4. These resistors also assure that the output point between the two transistors (point A) is maintained at one-half the dc supply voltage Vcc.

As in the case of the transformer-coupled output, phase inversion can be accomplished by use of an additional transistor. Fig. 228 shows a circuit in which the transistor phase inverter is used, together with a Darlington output stage to minimize loading on the phase inverter. It should be noted that capacitor C provides a drive reference back to the emitter of the upper output transistor. In effect, this arrangement duplicates the drive conditions of the split-winding transformer approach. A disadvantage of this circuit is the high-quiescent dissipation of the phase inverter Q_1 which is necessary to obtain adequate drive at full power output.

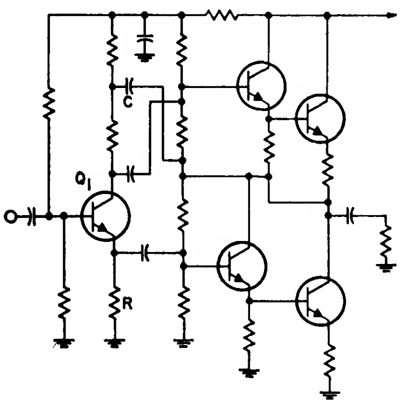

Fig.228 - Push-pull series-output amplifier in which driver and output transistors are connected as Darlington pairs and drive-signal phase inversion is provided by phase-splitter stage Q_1.

An unbypassed emitter resistor R is necessary because a signal is derived from this point to drive the lower output transistor. When transistor Q_1 is driven into saturation, the minimum collector-to-ground voltage that can be obtained is limited primarily by the peak emitter voltage under these conditions. To obtain the necessary voltage swing at this collector (a voltage swing that is also approximately equal to the output voltage swing), it is necessary to use a quiescent collector-to-emitter voltage higher than that required in a stage that uses a bypassed emitter resistor.

Complementary-Symmetry Amplifiers

When a complementary pair of output transistors (n-p-n and p-n-p) is used, it is possible to design a series-output type of audio power amplifier which does not require push-pull drive. Because phase inversion is unnecessary with this type of configuration, the drive circuit for the amplifier is simplified substantially. Fig. 229 shows a basic complementary type of series-output circuit together with a simple class A driver stage. The voltage drop across resistor R provides the small amount of forward bias required for class AB operation of the complementary pair of output transistors.

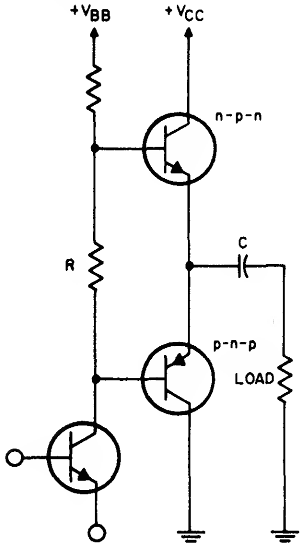

Fig.229 - Basic complementary type of series-output circuit with class A driver.

In practice, a diode is employed in place of resistor R. The purpose of the diode is to maintain the quiescent current at a reasonable value with variations in junction temperatures. It is usually thermally connected to one of the output transistors and tracks with the V_{BE} of the output transistors.

The complementary circuit is by far the most thermally stable output circuit. It places the output transistors in a V_{CES} mode because both transistors are operated with a low impedance between base and emitter. Therefore, the I_{CBO} leakage is the only component of concern in the stability criteria. At power-output levels from 3 to 20 watts, a complementary-symmetry amplifier offers advantages in terms of circuit simplicity. At higher power levels, however, the class A driver transistor is required to dissipate considerable heat, the quiescent power-supply current drain becomes significant, and excessively large filter capacitors are required to maintain a low hum level. This dissipation can be reduced, however, by use of a Darlington compound connection for the output stage. This compound connection reduces the driving-stage requirement.

There are two basic methods of overcoming this disadvantage entirely. The first is to use a quasi-complementary configuration; the second is to employ the compound true complementary-symmetry amplifier circuit shown in Fig. 230. Both methods replace the class A driver with a complementary driver stage. The circuit of Fig. 230 also employs a complementary grounded-base predriver stage which reduces static current drain even further. With this circuit it is practical to obtain power levels of over 100 watts with paralleled output transistors. For higher power levels, the quasi-complementary circuit is generally used because of the unavailability of higher power complementary devices.

Quasi-Complementary-Symmetry Amplifiers

In the quasi-complementary amplifier shown in Fig. 231, the driver transistors provide the necessary phase inversion. A simple but descriptive way to analyze the operation of a quasi-complementary amplifier is to consider the result of connecting a p-n-p transistor to a high-power n-p-n output transistor, as shown in Fig. 232. The collector current of the p-n-p transistor becomes the base current of the n-p-n transistor. The n-p-n transistor, which is operated as an emitter-follower, provides additional current gain without inversion. If the emitter of the n-p-n transistor is considered as the "effective" collector of the composite circuit, it becomes apparent that the circuit is equivalent to a high-gain, high-power p-n-p transistor.

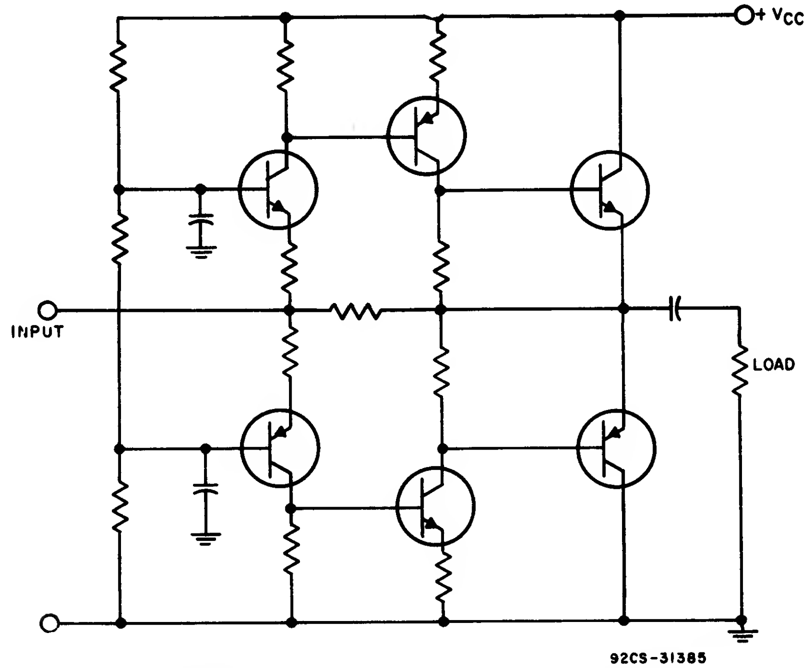

Fig.230 - Basic complementary type of series-output circuit with complementary type driver and predriver.

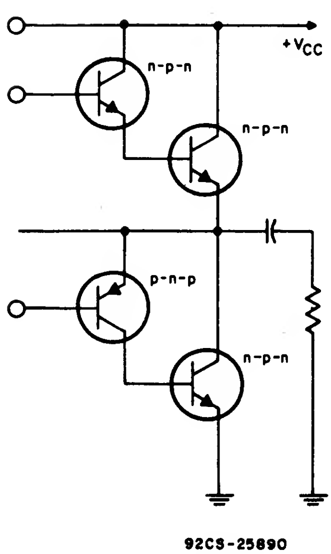

Fig.231 - Basic quasi-complementary type of series-output circuit.

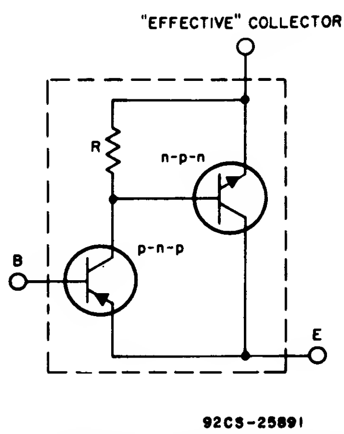

Fig.232 - Connection of p-n-p driver transistor to n-p-n output transistor.

The output characteristics of the p-n-p circuit shown in Fig. 232 and of a high-gain, high-power n-p-n circuit formed by the connection of the same type of n-p-n output transistor and an n-p-n driver transistor in a Darlington configuration, such as shown in Fig. 233, are compared in Fig. 234.

The saturation characteristics of the overall circuit in both cases are the combination of the base-to-emitter voltage V_{BE} of the output transistor and the collector saturation voltage of the driver transistor. Moreover, in both cases the current gain is the product of the individual betas of the transistors used. A quasi-complementary amplifier, therefore, is effectively the same as a simple complementary

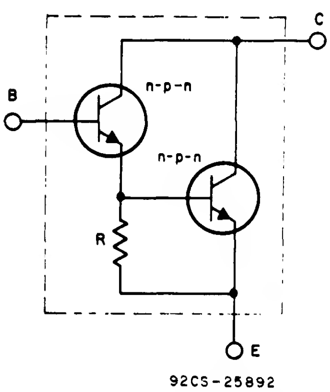

Fig.233 - Darlington connection of n-p-n driver transistor to n-p-n output transistor.

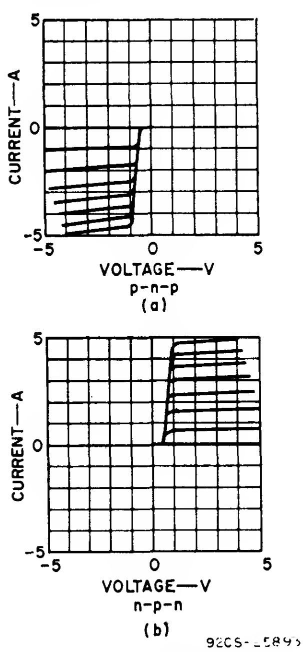

Fig.234 - Output characteristics for (a) p-n-p/n-p-n driver-output transistor pair shown in Fig. 232 and for (b) Darlington pair of n-p-n transistors shown in Fig. 233.

output circuit such as that shown in Fig. 229, and is formed by the use of high-gain, high-power n-p-n and p-n-p equivalent transistors. In both cases, the resistor R between the emitter and base of the output transistor places the device in a V_{CER} mode. This mode is not as stable as that of the complementary amplifier, but present no problem for silicon transistors.

A typical quasi-complementary amplifier is shown in Fig. 235. Capacitor C performs two functions essential to the successful operation of the circuit. First, it acts as a bypass to decouple any power-supply ripple from the driver and predriver stages. Second, it is connected as a "boot-strap" capacitor to provide the drive necessary to pull the upper Darlington pair of transistors into saturation. This latter function results from the fact that the stored voltage of the capacitor, with reference to the output point A, provides a higher voltage than the normal collector-supply voltage to drive transistor Q2. This higher voltage is necessary during the signal conditions that exist when the upper transistors are being turned on because the emitter voltage of transistor Q_2 then approaches the normal supply voltage. An increase in the base voltage to a point above this level is required to drive the transistor into saturation. Resistor R_1 provides the necessary dc feedback to maintain point A at approximately one-half the nominal supply voltage. Over-all ac feedback from output to input is coupled by resistor R_2 to reduce distortion and to improve low-frequency performance.

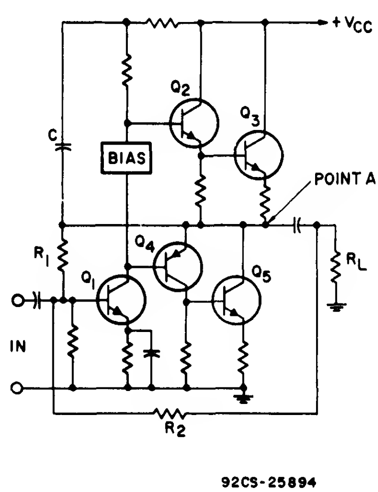

Fig.235 - Quasi-complementary audio power amplifier that operates from a single dc supply.

Series-output circuits can be employed with separate positive and negative supplies; no series output capacitor is then required. The elimination of this capacitor may result in an economic advantage, even though an additional power supply is used, because of the size of the series output capacitor necessary in the single-supply case to obtain good low-frequency performance (e.g., a 2000-microfarad capacitor is required to provide a 3-dB point at 20 Hz for a 4-ohm load impedance). Split supplies, however, pose certain problems which do not exist in the single-supply case. The output of the amplifier must be maintained at zero potential under quiescent conditions for all environmental conditions and device parameter variations. Also, the input ground reference can no longer be at the same point as that indicated in Fig. 235, because this point is at the negative supply potential in a split-supply system.

If the ground-point reference for the input signal were a common point between the split supplies, any ripple present on the negative supply would effectively drive the amplifier through transistor Q_1, with the result that this stage would operate as a common-base amplifier with its base grounded through the effective impedance of the input signal source. To avoid this condition, the amplifier must include an additional p-n-p transistor as shown in Fig. 236. This transistor (Q_6) reduces the drive effects of the negative supply ripple because of the high collector impedance (1 megohm or more) that it presents to the base of transistor Q_1, and effectively isolates the input source impedance from transistor Q_1. In

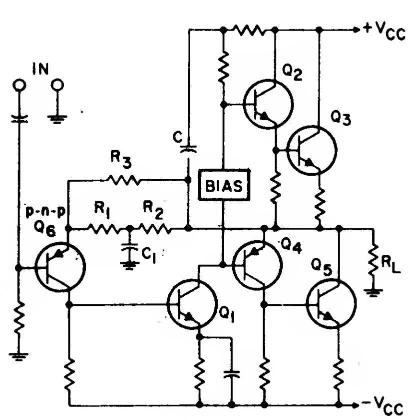

Fig.236 - Quasi-complementary audio power amplifier that operates from symmetrical dual dc power supplies. The p-n-p transistor input stage is required to prevent ripple component from driving amplifier.

practice, transistor Q_1 may be replaced by a Darlington pair to reduce the loading effects on the p-n-p predriver.

Negative dc feedback is applied from the output to the input stage by R_1, R_2, and C_1 so that the output is maintained at about zero potential. Actually, the output is maintained at approximately the forward-biased base-emitter voltage of transistor Q_6, which may be objectionable in a few cases, but which can be eliminated by a method discussed later. Capacitor C_1 effectively bypasses the negative dc feedback at all signal frequencies. Resistor R_3 provides ac feedback to reduce distortion in the amplifier.

True Complementary Symmetry Amplifiers

The true complementary symmetry amplifier shown in Fig. 237 has better thermal stability than other dc-coupled circuits, because transistors Q2 and Q5 are driven from the same low-impedance source. Forward bias for both transistors is provided by Q3, and is adjustable.

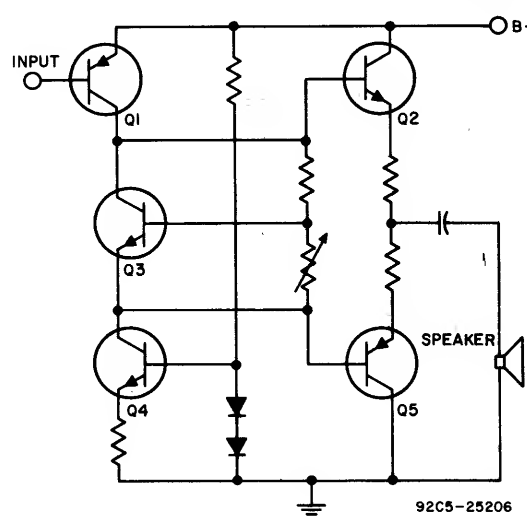

Fig.237 - True-complementary-symmetry amplifier.

Conjugate Complementary-Symmetry Amplifiers

Fig. 238 compares a transformer-coupled class AB amplifier to a **conjugate complementary amplifier**. The elimination of the transformer in the conjugate complementary amplifier, in the quasi-complementary amplifier and the true complementary amplifier shown in Fig. 237 permits a lighter-weight, less costly construction and eliminates the

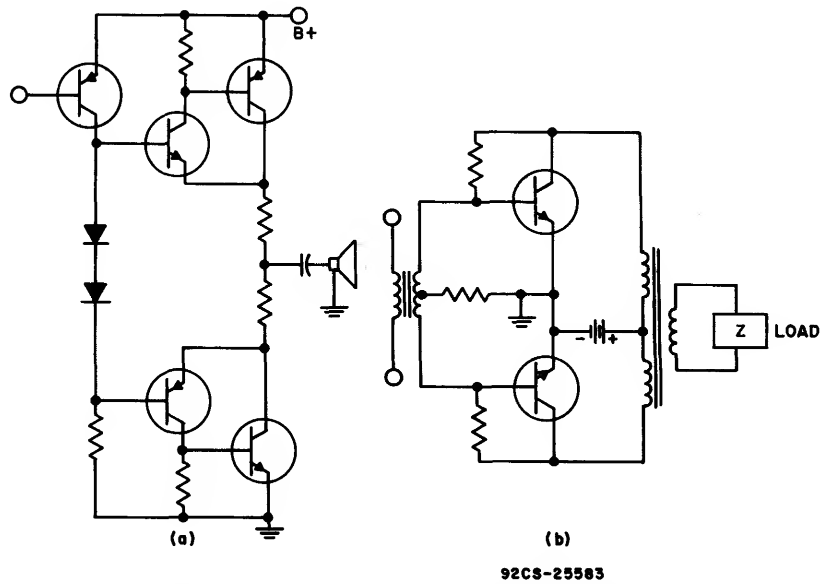

Fig.238 - (a) Conjugate-complementary-symmetry and (b) transformer-coupled amplifiers.

phase shifts and stability problems normally associated with transformers. The main advantage of a transformer-coupled circuit is easier matching of transistor volt-ampere capability to various load impedances.

Bridge Amplifiers

Fig. 239 shows the block diagram of an audio-amplifier configuration that, for a given dc supply voltage, transistor voltage-breakdown capability, and load, can provide four times the power output obtainable from a conventional push-pull audio-output stage. Alternatively, given power-output and load requirements may be achieved from this circuit

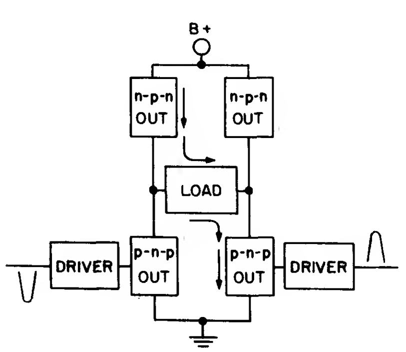

Fig.239 - Block diagram of bridge type of audio-amplifier circuit.

configuration with half the supply voltage and transistor voltage-breakdown capabilities required of conventional circuits. This performance is possible because the load can swing the full supply voltage on each half-cycle. The load is direct-coupled between the center point of two series-connected push-pull stages. This bridge type of arrangement eliminates the need for expensive coupling capacitors or transformers. These features are very attractive in applications for which the supply voltage is fixed, such as automotive or aircraft supplies.

The bridge-amplifier configuration consists essentially of two complementary-symmetry amplifiers with the load direct-coupled between the two center points. Each amplifier section is driven by a class A driver stage that uses a transistor Darlington pair. The amplifiers must be driven 180 degrees out of phase. This dual-phase drive is provided by a differential-amplifier type of input stage, which also provides the advantage of a high input impedance.

Fig. 240 shows the basic configuration of an experimental breadboard circuit designed to evaluate the bridge-amplifier approach to audio-amplifier design. The major difference between this type of circuit and the conventional complementary-symmetry circuit, besides the increased output power, is the higher current requirements of the class A driver

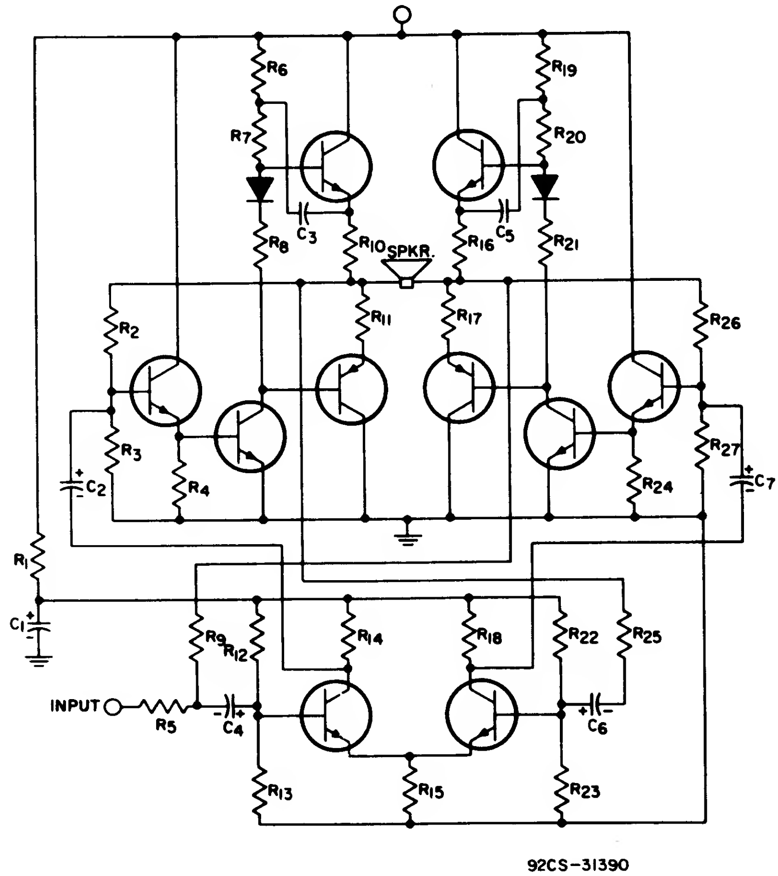

Fig.240 - Basic circuit configuration for a bridge type of audio amplifier.

stages. This current is twice the value normally required because the peak value of the output current is doubled. The feedback network from each complementary-symmetry output section back to the base of the corresponding class A driver stage, which establishes the center-point voltage in the output stage, also provides a minimum of 22 dB of ac feedback.

One problem encountered in the bridge amplifier is the achievement of a zero center-point (offset) voltage. The load circuit conducts a direct current proportional to the difference (offset) between the voltages at the two output stages. The dc dissipation in the load circuit is, of course, proportional to the square of the offset voltage. In this breadboard circuit, two potentiometers are used to balance the center-point voltage of the two output-stage sections.

The differential-amplifier input stage operates at ten times the required value of peak input current to assure linear operation. Balanced feedback is taken from each side of the load and coupled back to the separate bases of the differential-amplifier transistors. Fig. 241 shows curves of total harmonic distortion as a function of power output for operation of the bridge amplifier with 0 dB, 20 dB, and 28 dB of balanced feedback. Figs. 242 and 243 show total harmonic distortion and relative response as functions of frequency for the bridge amplifier operated with 20 dB of balanced feedback.

Phase Inverters

Phase inversion may be accomplished in many ways. The simplest electronic phase inverter is the single-stage configuration. This configuration can be used at low power levels or with high-gain devices when the limited drive capability is not a drawback. At higher power levels, some impedance transformation and gain may be required to supply the drive needed. There are several complex phase-

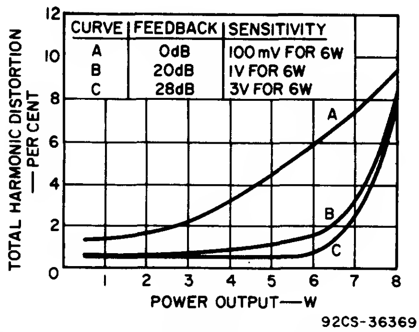

Fig.241 - Total harmonic distortion (at 1 kHz) of the bridge audio amplifier as a function of power output for different values of balanced loop feedback. (Distortion performance is comparable to that of a single-ended amplifier that provides one-quarter of the power output for the same dc supply voltage).

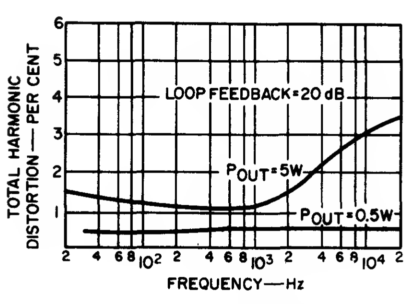

Fig.242 - Total harmonic distortion of the bridge audio amplifier as a function of frequency.

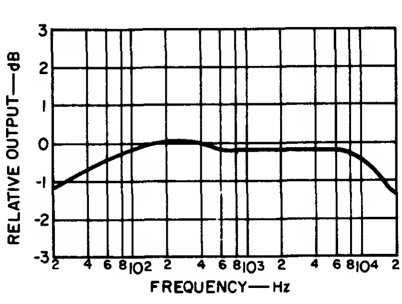

Fig.243 - Relative response of the bridge audio amplifier.

splitting circuits; a few of them are shown in Fig. 244.

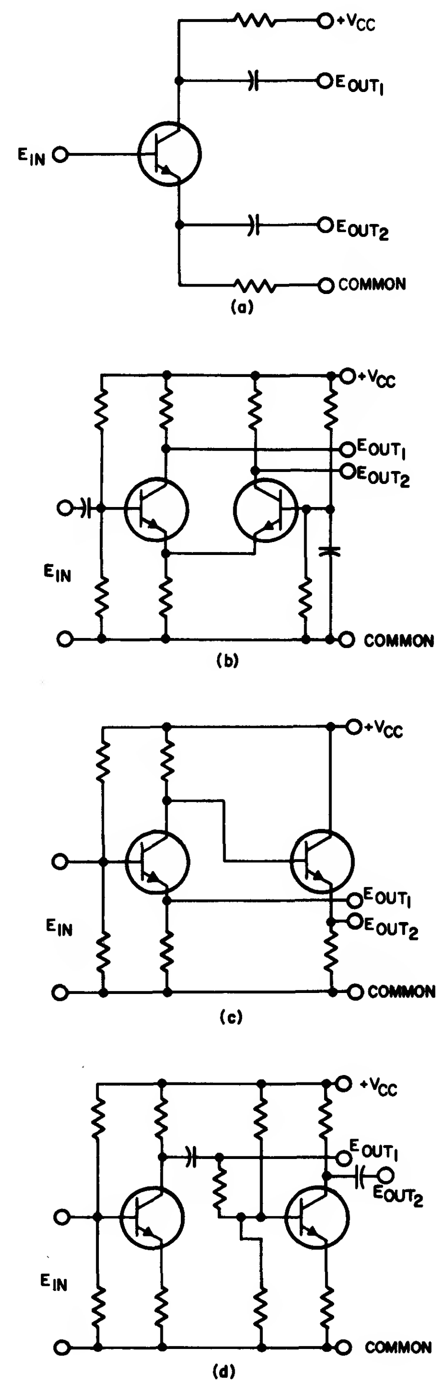

Fig.244 - Basic phase-inverter circuits: (a) single-stage phase-splitter type; (b) two-stage emitter-coupled type; (c) two-stage low-impedance type; (d) two-stage similar-amplifier type.

POWER OUTPUT IN CLASS B AUDIO AMPLIFIERS

For all cases of practical interest, the power output (P_o) of an audio amplifier is given by the following equation:

$$P_o = I(rms) \times E(rms) = (I_pE_p)/2$$
$$= (I_p^2R_L)/2 = E_p^2/2R_L$$

where I_p and E_p are the peak load current and voltage, respectively, and R_L is the load impedance presented to the transistor. Fig. 245 shows the relationship among these various factors in graphic form. Obviously, the peak load current is the peak transistor current, and the transistor breakdown-voltage rating must be at least twice the peak load voltage. The vertical lines that denote 4-ohm, 8-ohm, and 16-ohm resistances are particularly useful for transformerless designs in which the transistor operates directly into the loudspeaker.

Rating Methods

The **Institute of High Fidelity** (IHF) and the **Electronic Industries Association** (EIA) have attempted to standardize power-output ratings to establish a common reference of comparison and to provide a solid definition of the capabilities of audio power amplifiers. Obviously, an audio power amplifier using an unregulated supply can deliver more output power under transient conditions than under steady-state conditions. The rating methods which have been standardized for this type of operation are the **IHF Dynamic Output Rating (IHF-A-201)** and the **EIA Music Power Rating (EIA RS-234-A)**.

Both of these measurement methods allow the use of regulated supply voltage to simulate transient conditions. Because the regulated supply has no source impedance or ripple, the results do not completely represent the transient conditions, as will be explained later.

Measurement Techniques

The EIA standard is used primarily by manufacturers of packaged equipment, such

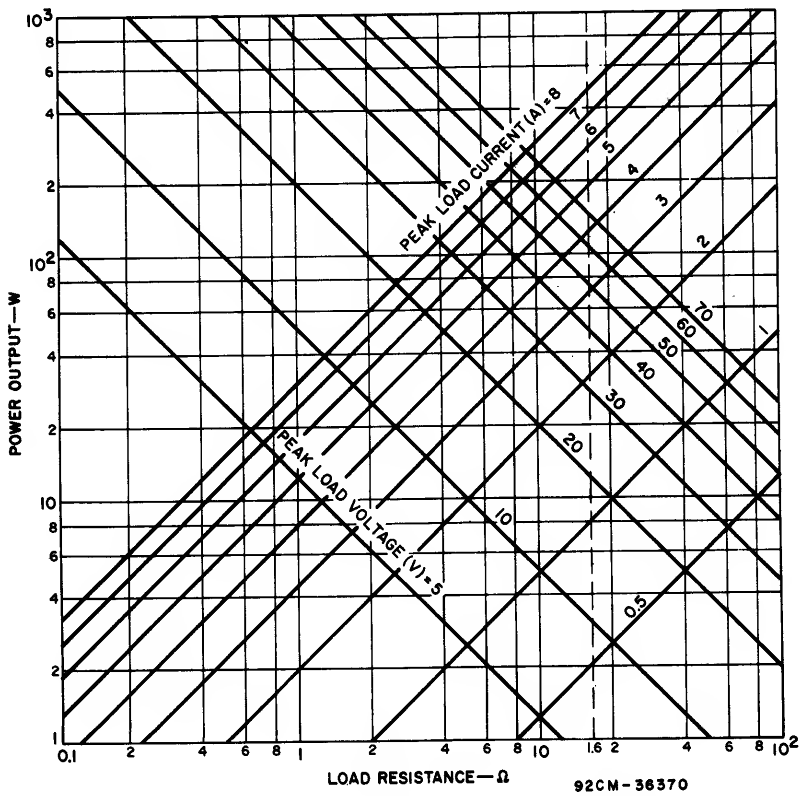

Fig.245 - Peak transistor currents and load voltages for various output powers and load resistances.

as portable phonographs, packaged stereo hi-fi consoles, and packaged home-entertainment consoles. The EIA **music power output** is defined as the power obtained at a total harmonic distortion of 5 per cent or less, measured after the "sudden application of a signal during a time interval so short that supply voltages have not changed from their no-signal values." The supply voltages are bypassed voltages. These definitions mean that the internal supply may be replaced with a regulated supply equal in voltage to the no-signal voltage of the internal supply. For a stereo amplifier, the music power rating is the sum of both channels, or twice the single-channel rating.

The IHF standard provides two methods to measure dynamic output. One is the **constant-supply method**. This method assumes that under music conditions the amplifier supply voltages undergo only insignificant changes. Unlike the EIA method, this measurement is made at a reference distortion. The constant-supply method is used by most high-fidelity component manufacturers. The reference distortion chosen is normally less than one per cent, or considerably lower than the EIA value of 5 per cent used by packaged-equipment manufacturers.

A second IHF method is called the **"transient distortion" test**. This method requires a complex setup including a low-distortion modulator with a prescribed output rise time and other equipment. The modulator output is required to have a rise time of 10 to 20 milliseconds to simulate the envelope rise time of music and speech. This measurement is made using the internal supply of the amplifier and, consequently, includes distortion caused by voltage decay, power-supply transients, and ripple. This method tends to be more realistic, and to yield lower power-output ratings than the constant-supply method. Actually, both IHF methods should be used, and the lowest power rating obtained at reference distortion with both channels operating, both in and out of phase, should be used as the power rating. (There is some question concerning unanimity among high-fidelity manufacturers on actually performing both IHF tests.)

Because music is not a continuous sine wave, and has average power levels much below peak power levels, it would appear that the music power or dynamic power ratings are true indications of a power amplifier's ability to reproduce music program material. The problem is that all three methods described have a common flaw. Even the transient-distortion method fails to account for the ability of the audio amplifier to reproduce power peaks while it is already delivering some average power. The amplifier is almost never delivering zero output when it is called on to deliver a transient. For every transient that occurs after an extremely quiet passage or zero signal, there are hundreds that are imposed on top of some low but non-zero average power level.

This condition can best be clarified by consideration of the power supply. Many amplifiers have regulated supplies for the front-end or low-level stages, but almost none provides a regulated supply for the power-output stages because regulation requires extra transistors or other devices; it becomes costly, especially at high power levels. The power supply for the output stages of power amplifiers is commonly a nonregulated rectifier supply having a capacitive input filter. The output voltage of such a supply is a function of the output current and, consequently, of the power output of the amplifier.

Effect of Power-Supply Regulation

Power-supply regulation is dependent on the amount of effective internal series resistance present in the power supply. The effective series resistance includes such things as the dc resistance of the transformer windings, the amount and type of iron used in the transformer, the amount of surge resistance present, the resistance of the rectifiers, and the amount of filtering. The internal series resistance causes the supply voltage to drop as current is drawn from the supply

Fig. 246 shows a typical regulation curve for a rectifier power supply that has a capacitive input filter. The voltage is a linear function of the average supply current over most of the useful range of the supply. However, a rapid change in slope occurs in the regions of both very small and very large currents. In class B amplifiers, the no-signal supply current normally occurs beyond the low-current knee, and the current required for the amplifier at the clipping level occurs before the high-current knee. The slope between these points is nearly linear and may be used as an approximation

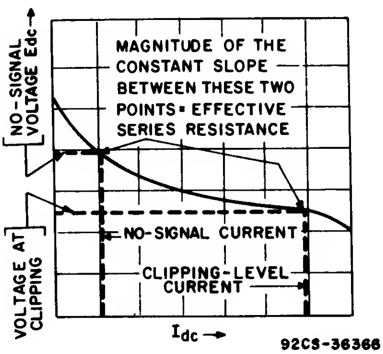

Fig.246 - Regulation curve for capacitive rectifier power supply.

of the equivalent series resistance of the supply.

The amount of power lost depends on the quality of the power supply used in the amplifier. Accordingly, rating amplifier power output with a superb **external** power supply (that is, not using the built-in amplifier power supply) provides false music power outputs. Under actual usage, the output is lower.

It should be emphasized that, while there is a discrepancy between the actual power measured under the EIA Music Power or the IHF Dynamic Power methods, these methods are not without merit. The IHF dynamic power rating, in conjunction with the continuous power rating, produces an excellent indication of how the amplifier will perform. The EIA music power rating, which is measured at a total harmonic distortion of 5 per cent with a regulated power supply, provides a less adequate indication of amplifier performance because there is no indication of how the amplifier power-supply voltage reacts to power output.

Some important factors considered by packaged-equipment manufacturers, the primary users of the EIA music power ratings, are mostly economic in nature and affect many aspects of the amplifier performance. Because there is no continuous power output rating required, two amplifiers may receive the same EIA music power rating but have different continuous power ratings. The ratio of music power to continuous power is, of course, a function of the regulation and effective series resistance of the supply.

One reason for the difference between ratings used by the console or the packaged-equipment manufacturer and those used by the hi-fi component manufacturer is that the latter does not always know just what will be required of the amplifier. The console manufacturer always designs an amplifier as part of a system, and consequently knows the speaker impedances and the power required for adequate sound output. The console manufacturer may use high-efficiency speakers requiring only a fraction of the power needed to drive many component-type acoustic-suspension systems. The difference may be such that the console may produce the same sound pressure level with an amplifier having one-tenth of the power output. High ratios of music-power to continuous-power capability are common in these consoles. A typical ratio of IHF music power to continuous power may be 1.2 to 1 in component amplifiers, whereas a typical ratio of EIA music power to continous power in a console system may be 2 to 1. Console manufacturers use the EIA music power rating to economic advantage as a result of the reduced regulation requirement of the power supply. A high ratio of music power to continuous power means higher effective series resistance in the power supply. This resistance, in turn, means less continuous dissipation on the output transistors, smaller heat sinks, and a lower-cost power supply.

Ratio of Music Power to Continuous Power

Some advantages of high values of the ratio R_S/R_L and correspondingly high ratios of music power output to transistor dissipation are as follows:

1. Reduced heat sink or transistor cost: Because the volt-ampere capacity of the transistor is determined by the music power output, it is not likely that reduced thermal-resistance requirements will produce significant cost reductions. Alternatively, the heat-sink requirements may be reduced.
2. Reduced power supply costs: Transformer and/or filter-capacitor specifications may be relaxed.
3. Reduced speaker cost: Continuous power-handling capability may be relaxed.

These cost reductions may be passed along to the consumer in the form of more music power per dollar.

The question arises as to how high the ratio R_S/R_L and the corresponding ratio of music power output to continuous power output may be before the capability of the amplifier to reproduce program material is impaired.

The objective is to provide the listener with a close approximation of an original live performance. Achievement of this objective requires the subjective equivalents of sound pressure levels that approach those of a concert hall. Although the peak sound pressure level of a live performance is about 100 dB, the average listener prefers to operate an audio system at a peak sound pressure level of about 80 dB. The amplifier, however, should also accommodate listeners who desire higher-than-average levels, perhaps to peaks of 100 dB.

A sound pressure level of 100 dB corresponds to about 0.4 watt of acoustic power for an average room of about 3,000 cubic feet.

If speaker efficiencies are considered to be in the order of 1 per cent, a stereophonic amplifier must be capable of delivering about 20 watts per channel. Higher power outputs are required for lower-efficiency speakers. The peak-to-average level for most program material is between 20 and 23 dB. A system capable of providing a continuous level of 77 dB and peaks of 100 dB would satisfy the power requirements of nearly all listeners. For this performance to be attained, the power-supply voltage cannot drop below the voltage required for 100 dB of acoustic power while delivering the average current required for 77 dB. Moreover, because sustained passages that are as much as 10 dB above the average may occur, the power-supply voltage cannot drop below the value required for 100 dB of acoustic power while delivering 87 dB of acoustic power (87 dB of acoustic power corresponds to about 1 watt per channel). This performance means that for 8-ohm loads, with output-circuit losses neglected, the power-supply voltage must not decrease to a value less than 36 volts, while delivering the average current required for 1 watt per channel (0.225 ampere dc).

It should be noted that the power-output capability for peaks while the amplifier is delivering a total of 2 watts is not the music power rating of the amplifier because the power-supply voltage is below its no-signal value by an amount depending on its effective series resistance.

THERMAL-STABILITY REQUIREMENTS

One serious problem that confronts the design engineer is the achievement of a circuit which is thermally stable at all temperatures to which the amplifier might be exposed. As previously discussed, thermal runaway may be a problem because the V_{BE} of all transistors decreases at low current. It should be noted, however, that at high current levels the base-to-emitter voltage of silicon transistors increases with a rise in junction temperature. This characteristic is the result of the increase in the base resistance that is produced by the rise in temperature. The increase in base resistance helps to stabilize the transistor against thermal runaway. In high-power amplifiers, the emitter resistors employed usually have a value of about 1 ohm or less. The size of the capacitor required to bypass the emitter adequately at all frequencies of interest makes this approach economically impractical. A more practical solution is to increase the value of the emitter resistor and shunt it with a diode. With this technique, sufficient degeneration is provided to improve the circuit stability; at low currents, however, the maximum voltage drop across the emitter resistor is limited to the forward voltage drop of the diode.

The quasi-complementary amplifier shown in Fig. 247 incorporates the stabilization

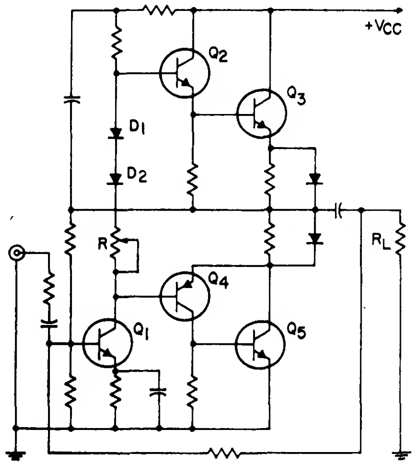

Fig.247 - Quasi-complementary amplifier that incorporates two stabilization networks.

techniques described. A resistor-diode network is used in the emitter of transistor Q_3, and another such network is used in the collector of transistor Q_6, with the emitter of transistor Q_4 returned to the collector of transistor Q_5. Previous discussion regarding the p-n-p driver and n-p-n output combination (Q_4 and Q_5) showed that the collector of the output device becomes the "effective" emitter of the high-gain, high-power p-n-p equivalent, and vice versa. For maximum operating-point stability, therefore, the diode-resistor network should be in the "effective" emitter of the p-n-p equivalent. Quasi-complementary circuits employing the stabilization resistor in the emitter of the lower output transistor do not improve the operating-point stability of the over-all circuit.

The circuit shown in Fig. 247 is biased for class AB operation by the voltage obtained from the forward drop of two diodes, CR_1 and CR_2, plus the voltage drop across potentiometer R, which affords a means for a slight adjustment in the value of the quiescent current. The current necessary to provide this voltage reference is the collector current of driver transistor Q_1. The diodes may be thermally connected to the heat sink of the output transistors so that thermal feedback is provided for further improvement of thermal stability. Because the forward voltage of the reference diodes decreases with increasing temperature, these diodes compensate for the decreasing V_{BE} of the output transistors by reducing the external bias applied. In this way, the quiescent current of the output stage can be held relatively constant over a wide range of operating temperatures.

EFFECTS OF LARGE PHASE SHIFTS

The amplifier frequency-response characteristic is an important factor with respect to the ability of the amplifier to withstand unusually severe electrical stress conditions. For example, under certain conditions of input-signal amplitude and frequency, the amplifier may break into high-frequency oscillations which can lead to destruction of the output transistors, the drivers, or both. This problem becomes quite acute in transformer-coupled amplifiers because the characteristics of transformers depart from the ideal at both low and high frequencies. The departure occurs at low frequencies because the inductive reactance of the transformer decreases, and at high frequencies because the effects of leakage inductance and transformer winding capacitance become appreciable. At both frequency extremes, the effect is to introduce a phase shift between input and output voltage.

Negative feedback is used almost universally in audio amplifiers; the voltage coupled back to the input through the feedback loop may cause the amplifier to be potentially unstable at some frequencies, especially if the additional phase shift is sufficient to make the feedback positive. Similar effects can occur in transformerless amplifiers because reactive elements (such as coupling and bypass capacitors, transistor junction capacitance, stray wiring capacitance, and inductance of the loudspeaker voice coil) are always present. The values of some of the reactive elements (e.g., transistor junction capacitance and transformer inductance as the core nears saturation) are functions of the signal level; coupling through wiring capacitance and unavoidable ground loops may also vary with the signal level. As a result, an amplifier that is stable under normal listening levels may break into oscillations when subjected to high-level signal transients.

A large phase shift is not only a potential cause of amplifier instability, but also results in additional transistor power dissipation and increases the susceptibility of the transistor to forward-bias second-breakdown failures. The effects of large-signal phase shifts at low frequencies are illustrated in Fig. 248, which shows the load-line characteristics of a transistor in a class AB push-pull circuit for signal frequencies of 1000 Hz and 10 Hz. The phase

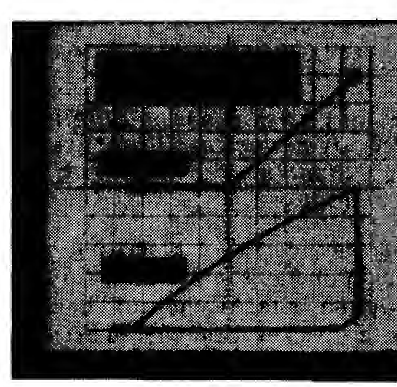

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Fig.248 - Effect of large signal phase shift on the load-line characteristics of a transistor at low frequencies.

shift is caused primarily by the output capacitor. In both cases, the amplifier is driven very strongly into saturation by a 5-volt input signal. The increased dissipation at 10 Hz, compared to that obtained at 1000 Hz, results from simultaneous high-current high-voltage operation. The transistor is required to handle safely a current of 0.75 ampere at a collector voltage of 40 volts for an equivalent pulse duration of about 10 milliseconds; it must be free from second-breakdown failures under these conditions.

EFFECT OF EXCESSIVE DRIVE

Simultaneous high-current high-voltage operation may also occur in class B amplifiers at high frequencies when the amplifier is overdriven to the point that the output signals are clipped. For example, if the input signal applied to the series-output push-pull circuit shown in Fig. 249(a) is large enough to drive the transistors into both saturation and cutoff, transistor A is driven into saturation, and transistor B is cut off during a portion of the input cycle. Fig. 249(b) shows the collector-current waveform for transistor A under these conditions.

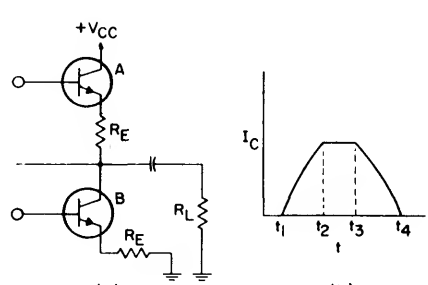


Fig.249 - (a) class B series-output stage, (b) collector-current waveform under overdrive (clipping) conditions.

During the interval of time from t_2 to t_3, transistor A operates in the saturation region and the output voltage is clipped. The effective negative feedback is then reduced because the output voltage does not follow the sinusoidal input signal. Transistor A, therefore, is driven even further into saturation by the unattenuated input signal. When transistor B starts to conduct, transistor A cannot be turned off immediately because the excessive drive results in a large storage time. As a result, transistor B is required to support almost the full supply voltage (less only the saturation voltage of transistor A and the voltage drop across the emitter resistors, if used) as its current is increased by the drive signal. For this condition to occur, a large input signal is required at a frequency high enough so that the storage time is greater than one-quarter cycle.

Because of the charging current through the output coupling capacitor, transistor A in Fig. 249(a) is also subject to forward-bias second-breakdown failure if the dc supply voltage and a large input signal are applied simultaneously.

All of these conditions point to the need for a good "safe area" of operation. Fig. 250 shows the safe area for the RCA-2N3055. In all cases, the load lines fall within the area guaranteed safe for this transistor.

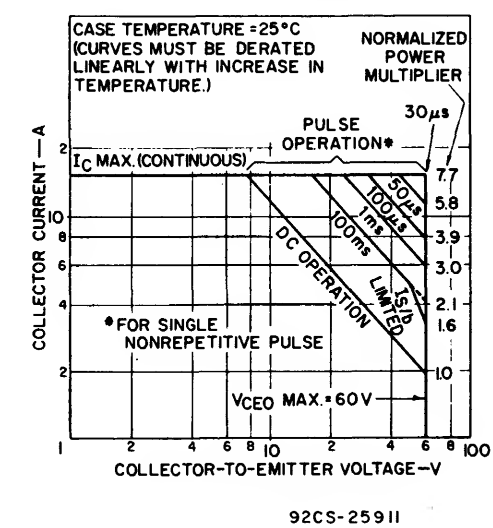


Fig.250 - Safe-area-of-operation rating chart for the RCA-2N3055 homotaxial-base transistor.

Short-Circuit Protection

Another important consideration in the design of high-power audio amplifiers is the ability of the circuit to withstand short-circuit conditions. As previously discussed, overdrive conditions may result in disastrously high currents and excessive dissipation in both driver and output stages. Obviously, some form of short-circuit protection is necessary. One such technique is shown in Fig. 251. A current-sampling resistor R is placed in the ground leg of the load. If any condition (including a short) exists such that higher-

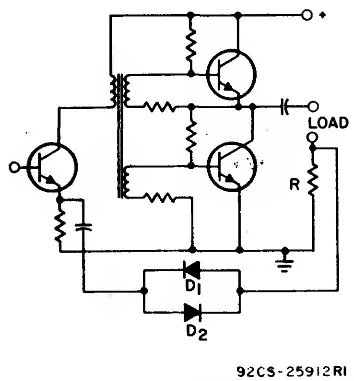


Fig.251 - Push-pull power amplifier with short-circuit protection.

than-normal load current flows, diodes CR_1 and CR_2 conduct on alternate half-cycles and thus provide a high negative feedback which effectively reduces the drive of the amplifiers. This feedback should not exceed the stability margin of the amplifier. This technique in no way affects the normal operation of the amplifier.

A second approach to current limiting is illustrated by the circuit shown in Fig. 252. In this circuit, a diode biasing network is used to establish a fixed current limit on the driver and output transistors. Under sustained short-circuit conditions, however, the output transistors are required to support this current limit and one-half the dc supply voltage.

The circuit shown in Fig. 253 illustrates a dissipation-limiting technique that provides

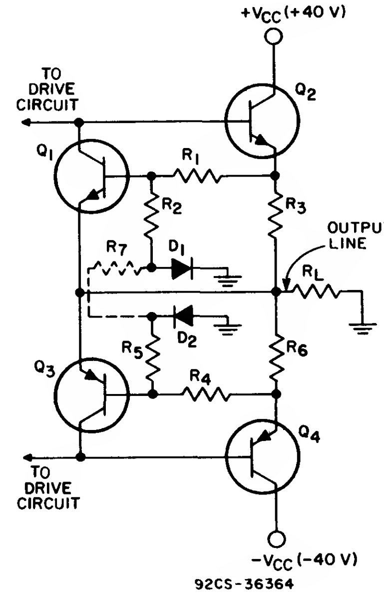


Fig.253 - Quasi-complementary audio output in which diode-resistor biasing network is used to prevent complementary transistors Q_1 and Q_2 from being forward-biased by the output voltage swing.

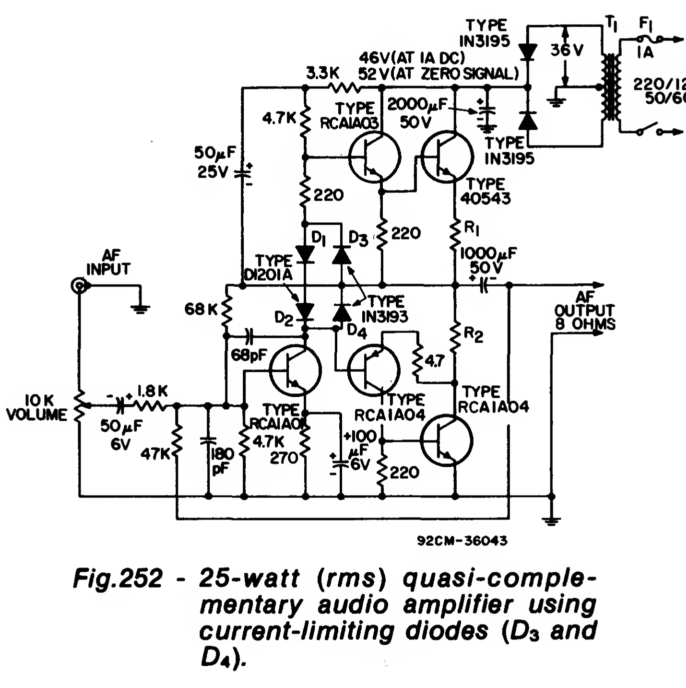


Fig.252 - 25-watt (rms) quasi-complementary audio amplifier using current-limiting diodes (D_3 and D_4).

positive protection under all loading conditions. The limiting action of this circuit is shown in Fig. 254. This safe-area limiting technique permits use of low-dissipation driver and output transistors and of smaller heat sinks in the output stages. The use of smaller heat sinks is possible because the worst-case dissipation is normal 4-ohm operation instead of short-circuit conditions. With this technique, highly inductive or capacitive loads are no

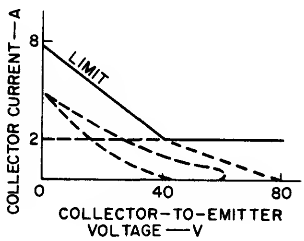


Fig.254 - Load lines for the circuit of Fig. 253. Load lines showing effect of the inclusion of high-resistance diode-resistor network in the forward-biasing path of Q_1 are shown dotted.

longer a problem, and thermal cut-outs are unnecessary. In addition, the technique is inexpensive.

V_{BE} MULTIPLIER BIASING CIRCUIT FOR POWER AMPLIFIER OUTPUT STAGES

The following paragraphs describe a biasing circuit for the output stage of a power amplifier. The biasing circuit is called a V_{BE} multiplier; its purpose is to provide proper bias for the output transistors of the amplifier under all operating conditions. The amount of forward bias provided determines the quiescent operating point of the output stage. The criteria for determining the proper quiescent collector current of the output transistors are the output-signal distortion level to be achieved and the need to minimize quiescent current because of dissipation in the output transistors. Fig. 255 shows the circuit of a typical complementary output stage for an audio amplifier. In this circuit, transistor Q3 serves as the biasing element for transistors Q4 and Q5.

Since all transistors are temperature sensitive, the bias circuit should change bias voltage in such a manner that the quiescent collector

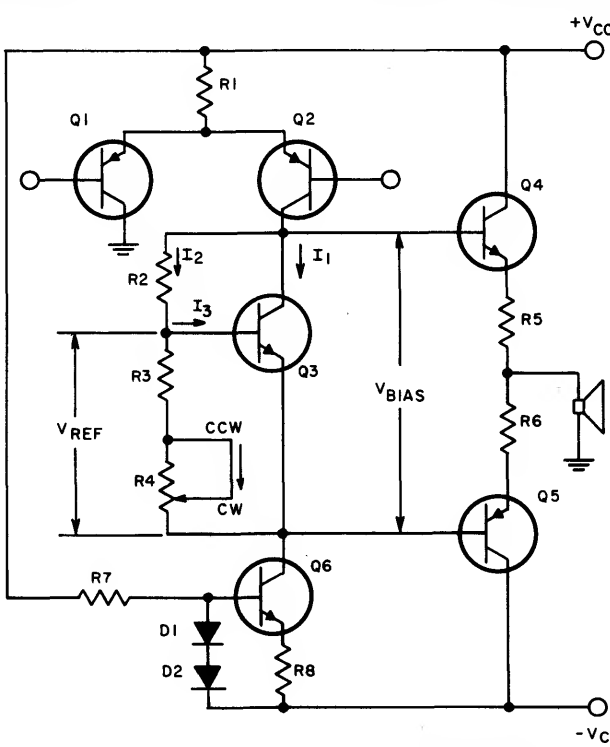


Fig.255 - Complementary output stage for an audio amplifier.

current of the output transistors remains constant. Typical temperature dependence of a silicon power transistor is shown in Fig. 256. The figure shows that the bias voltage must decrease approximately 2 mV/°C if the collector current is to be constant. Failure to provide thermal compensation will result in a current change of:

$$\frac{\Delta I_C}{\Delta T} \simeq 10\%/°C$$

A further examination of Fig. 256 shows that an error of 20 millivolts (3 per cent) in the bias voltage will result in a change in the collector current by a factor of 2.

Transistor Q3 in Fig. 255 varies the biasing voltage for the output transistors so that quiescent current does not change with temperature change. This constant-current condition is achieved by mounting Q3, Q4, and Q5 on the same heat sink so that a change in the junction temperature of the output transistors will change the heat-sink temperature proportionally and, therefore, the junction temperature of Q3. If, for example, temperature increases, the collector current of Q3 would tend to increase, but constant-current source Q6 keeps the collector current of Q3 constant. Under this condition, the V_{BE} of Q3 will decrease and V_{bias} will decrease proportionally. The net result will be the stabilization of the quiescent collector current of Q4 and Q5.

AUDIO AMPLIFIER CIRCUITS USING ALL DISCRETE DEVICES

A broad selection of power levels can be obtained from amplifiers using only discrete solid state devices. The following chart lists the type of circuit configuration and recommended output devices for power output levels ranging from 3 to 300 watts. A circuit diagram and performance data are shown for representative amplifiers. For information on the other audio amplifier circuits listed in the chart, refer to RCA "**Audio Amplifier Manual,**" APA-551 and the individual data sheets for the output devices.

25-Watt True-Complementary-Symmetry Audio Amplifier with Darlington Output Transistors

Fig. 257 shows a complementary amplifier rated at 25 watts output with an 8-ohm load, using Darlington transistors in the output stage. The amplifier also will supply 25 watts output with a 4-ohm load and 14-watts with a 16-ohm load. Thermal stability is provided by mounting the biasing transistor on the output heat-sink. Dissipation-limiting overload protection is incorporated in this circuit. A 70°C thermal cut-out should be used in the primary of the power supply.

Typical performance data are shown in Table XVI. Fig. 258 shows distortion as a function of power output.

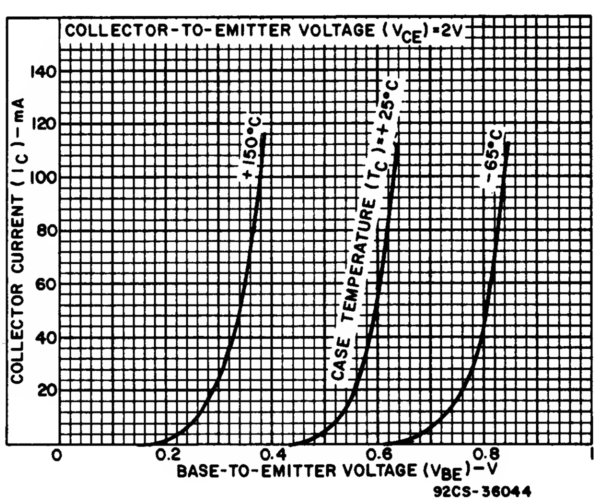


Fig.256 - Temperature dependence of a silicon power transistor.

Table XV - Selection Chart for Discrete Preamplifier Power Output Stage Amplifiers

Power Output (W)	Load Res. (Ω)	Supply Voltage (V)	Type of Circuit	Output Transistors NPN	Output Transistors PNP
3 12	8 4	20	True-Comp.	RCP703A	RCP702A
12 6.5	8 16	36	True-Comp.	RCA1C10	RCA1C11
45	4			RCA1C05	RCA1C06
25	8	52	True-Comp.	BD243A or RCA1C05	BD244A or RCA1C06
16	16			RCA1C05	RCA1C06
25	4	40		2N6387 or BDX33A	2N6667 or BDX34A
25	8	52	True-Comp. Darlington	2N6388 or BDX33B	2N6668 or BDX34B
14	16	52		2N6388 or BDX33B	2N6668 or BDX34B
40	4	46		BD501A or RCA1C07	BD500A or RCA1C08
40	8	64	True-Comp.	BD501B or RCA1C07	BD500B or RCA1C08
22	16	64		BD501B or RCA1C07	BD500B or RCA1C08
40 40	4 8	46 64	Quasi-Comp.	2-2N6101 or 2-RCA1C09	—
25 100	16 4	64		2-RCA1B06	—
70	8	90	Quasi-Comp.	2-BD550 or 2-RCA1B06	—
40	16			2-BD550 or 2-RCA1B06	—
100	4	84		2-RCA1B01	—
70	4	60		2-BD450	—
70	8	84	Quasi-Comp.	2-BD451 or 2-RCA1B01	—
38	16	84		2-BD451 or 2-RCA1B01	—
180	4	130		4-RCA1B04	—
120	4	90		4-BD550	—
120	8	130	Quasi-Comp.	4-BD550A or 4-RCA1B04	—
70	16	130		4-BD550A or 4-RCA1B04	—
300	4	160		6-RCA1B05	—
200	4	110		6-BD550A	—
200	8	160	Quasi-Comp.	6-BD550B or 6-RCA1B05	—
120	16	160		6-BD550B or 6-RCA1B05	—

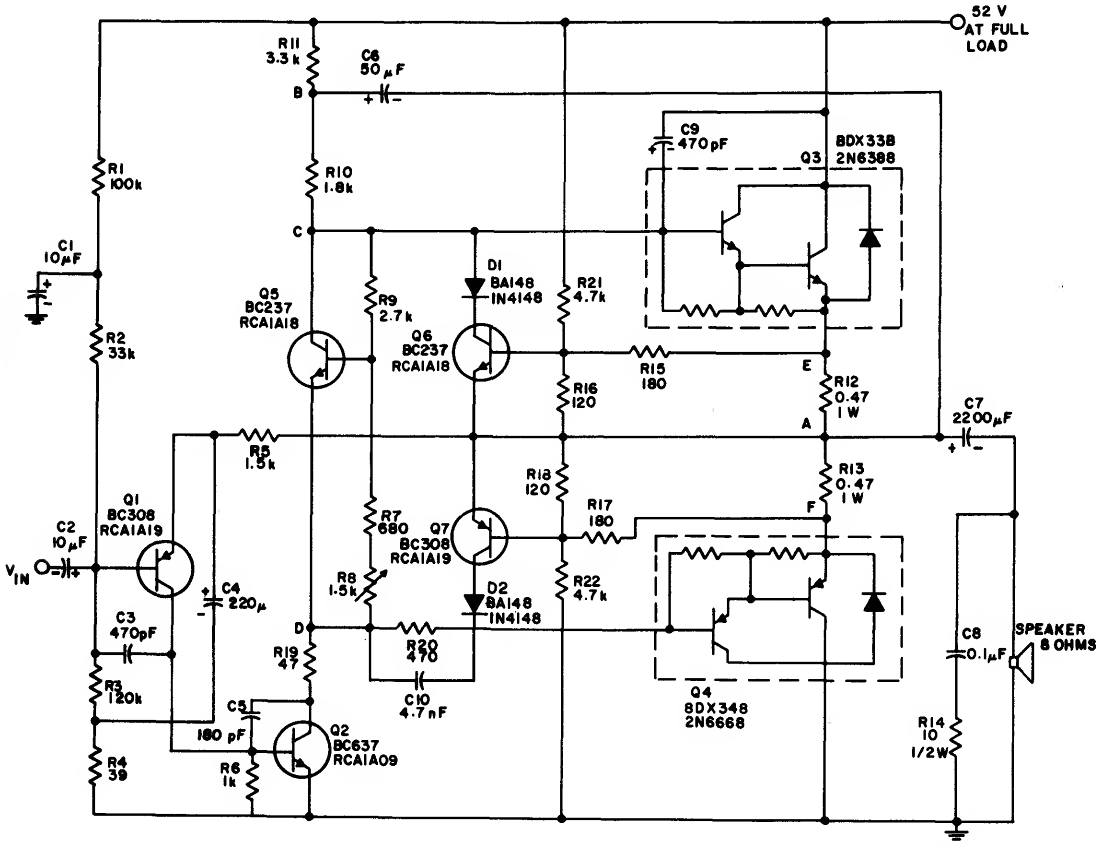


Fig.257 - 25-watt true-complementary-symmetry amplifier featuring Darlington output transistors.

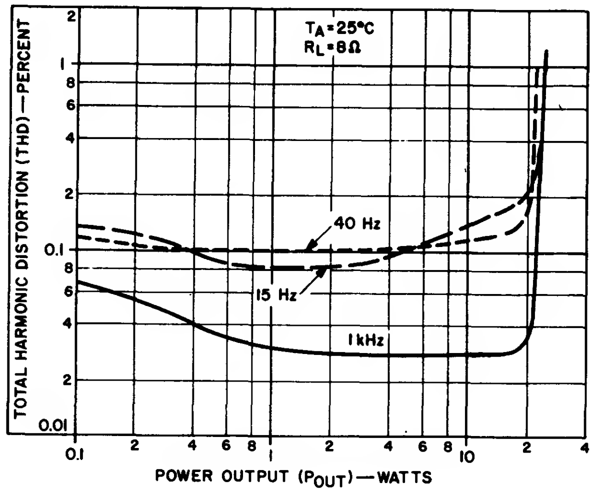


Fig.258 - Typical total harmonic distortion as a function of power output.

Table XVI - Typical Performance Data for 25-Watt Audio Amplifier

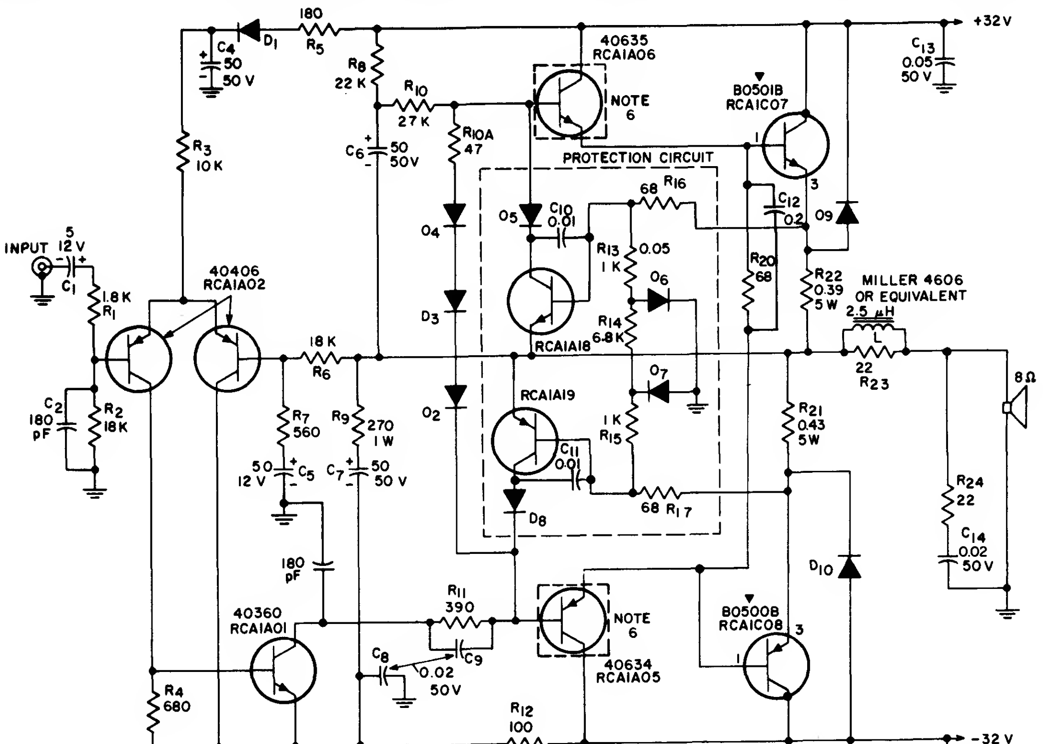
Measured at V_{CC}=52 V, T_A=25° C, and frequency of 1 kHz unless otherwise specified.

Power:	
Rated power (8 Ω load)	25 W
Typical power (4 Ω load)	25 W*
Typical power (16 Ω load)	14 W
Total Harmonic Distortion	See Fig. 258
Sensitivity:	
For 25-W output	360 mV

*With 40-V supply voltage and BDX33A, BDX34A substituted for BDX33B, BDX34B.

40-Watt True-Complementary-Symmetry Audio Amplifier

Fig. 259 shows a complementary amplifier using epitaxial base output transistors rated at 40 watts output with an 8-ohm load. A power supply intended to supply two identical amplifiers is shown in Fig. 260. The amplifiers also will supply 40 watts output with a 4-ohm load and 22 watts with 16-ohm load. This amplifier provides outstanding stability of the output transistors through the use of a unique turn-off drive circuit, which consists of a resistor and capacitor connected between the bases of the discrete output devices. This



NOTES:

1. D1-D10 - 1N4002.
2. Resistors are ½-watt, ±10%, unless otherwise specified; values are in ohms.
3. Non-inductive resistors.
4. Capacitances are in μF unless otherwise specified.
5. ▼ Provide heat sink of approx. 1.2° C/W per output device with a contact thermal resistance of 1.3° C/W max. and T_A=40° C max.
6. TO-39 case devices with heat radiator attached.

Fig.259 - 40-watt amplifier featuring true-complementary-symmetry output using load-line limiting.

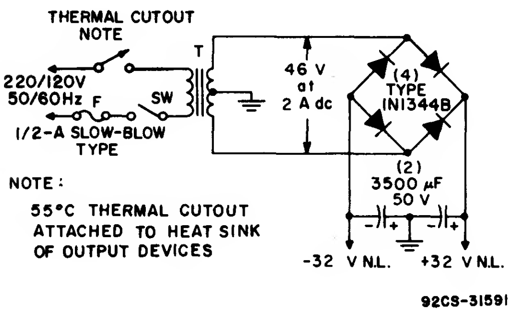


Fig.260 - Power supply for 40-watt amplifier.

technique allows the output transistors to operate with equal power-bandwidth performance. The circuit also incorporates dissipation-limiting overload protection.

Typical performance data are shown in Table XVII. Fig. 261 shows distortion as a function of frequency and Fig. 262 shows the frequency response.

70-Watt Quasi-Complementary-Symmetry Audio Amplifier

Fig. 263 shows a complementary amplifier using hometaxial-base output transistors, rated at 70 watts output with an 8-ohm load. A

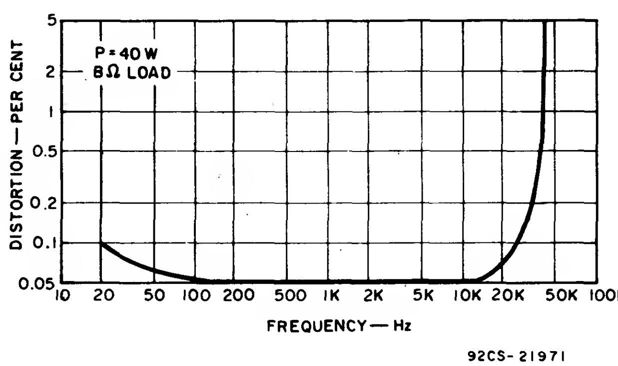


Fig.261 - Typical total harmonic distortion as a function of frequency.

Table XVII - Typical Performance Data for 40-Watt Audio Amplifier

Measured at V_{CC}=64 V, T_A=25° C, and a frequency of 1 kHz, unless otherwise specified.

Power:	
Rated power (8 Ω load, at rated distortion)	40 W
Typical power (4 Ω load)	75 W*
Typical power (16 Ω load)	22 W
Total Harmonic Distortion:	
Rated distortion	1%
Typical at 20 W	0.05%
IM Distortion:	
10 dB below continuous power output at 60 Hz and 7 kHz (4:1)	0.1%
IHF Power Bandwidth:	
3 dB below rated continuous power at rated distortion	80 kHz
Sensitivity:	
At continuous power-output rating	600 mV
Hum and Noise:	
Below continuous power output:	
Input shorted	80 dB
Input open	75 dB
Input Resistance	20 KΩ

*Typical power (4 Ω load) with 46-volt split power supply and BD500A, BD501A output .... 40 W
Typical power (4 Ω load) with 40-volt split power supply and BD500, BD501 output .... 25 W

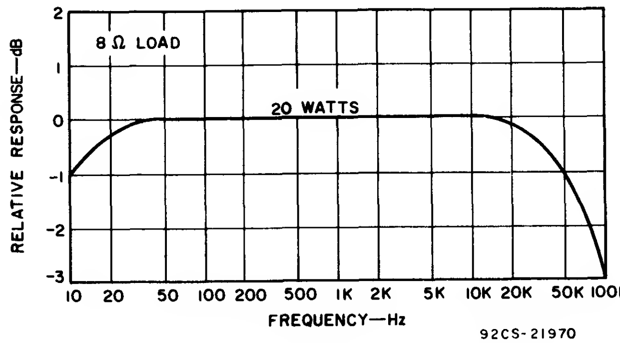
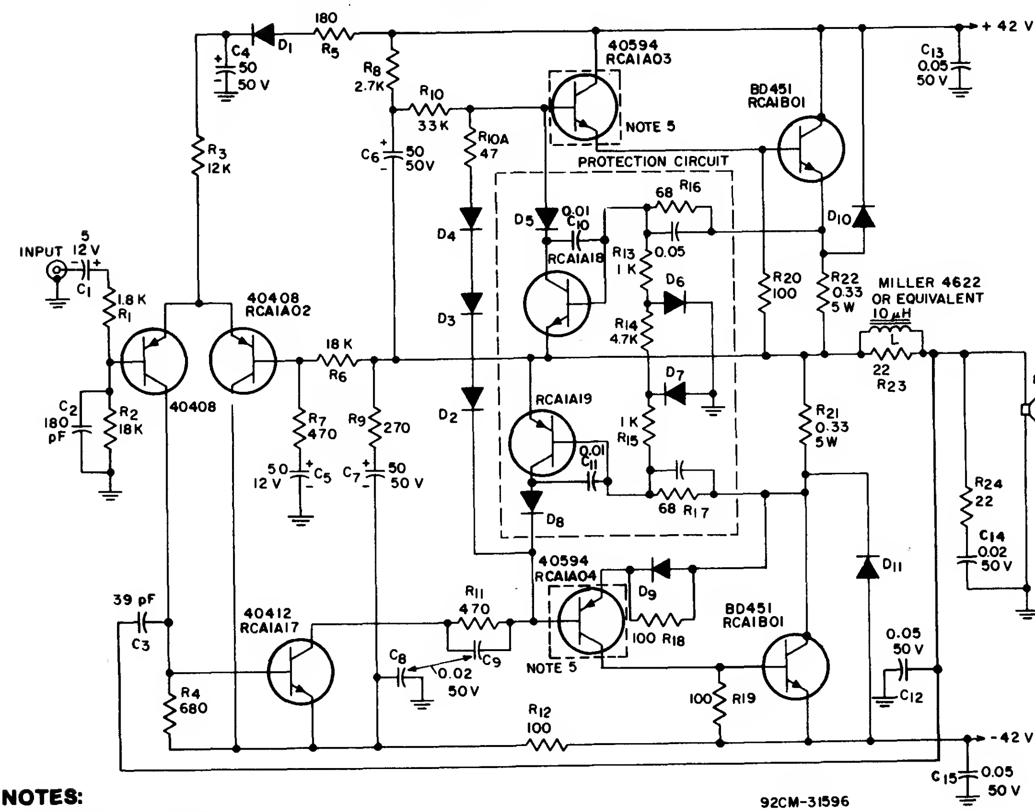


Fig.262 - Typical frequency response.



NOTES:

1. D1-D11 - 1N4002.
2. Resistors are ½-watt, ±10%, unless otherwise specified; values are in ohms.
3. Non-inductive resistors.
4. Capacitances are in μF unless otherwise specified.
5. Mount each device on TO-39 heat sink.
6. Provide heat sink of approx. 1.2° C/W per output device with a contact thermal resistance of 0.5° C/W max. and T_A=45° C max.

Fig.263 - 70-watt amplifier circuit featuring quasi-complementary-symmetry output employing hometaxial-base construction output transistors.

power supply intended to supply two identical amplifiers is shown in Fig. 264. The amplifier also will supply 100 watts with a 4-ohm load and 38 watts with a 16-ohm load. The circuit is unusually rugged in regard to overloads, but also incorporates dissipation-limiting overload protection.

Typical performance data are shown in Table XVIII. Fig. 265 shows distortion as a function of power output, and Fig. 266 shows the response curve.

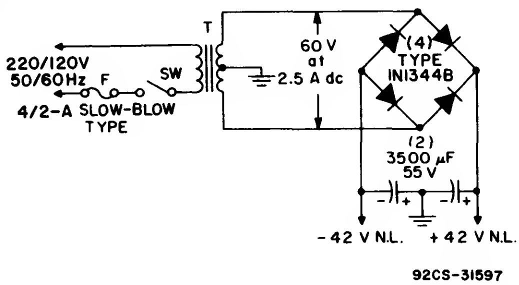


Fig.264 - Power supply for 70-watt amplifier.

120-Watt Quasi-Complementary-Symmetry Audio Amplifier

Fig. 267 shows an amplifier using two pairs of complementary output transistors in parallel rated at 120 watts output with an 8-ohm load. A power supply intended to supply two identical amplifiers is shown in Fig. 268. The amplifier also will supply 180 watts with a 4-ohm load and 70 watts with a 16-ohm load. Thermal stability is enhanced by mounting the biasing transistor on the output heat sink. The circuit incorporates dissipation-limiting overload protection.

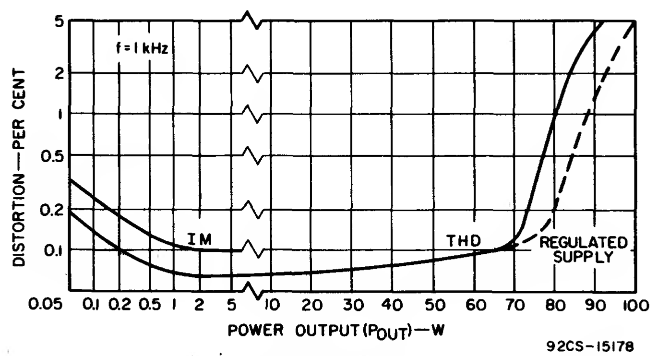


Fig.265 - Typical intermodulation and total harmonic distortion as a function of power output at 1 kHz.

Table XVIII - Typical Performance Data for 70-Watt Audio Amplifier

Measured at V_{CC}=84 V, T_A=25° C, and a frequency of 1 kHz unless otherwise specified.

Power:	
Rated power (8 Ω load, at rated distortion)	70 W
Typical power (4 Ω load)	100 W•
Typical power (16 Ω load)	38 W
Music power (8 Ω load, at 5% THD with regulated supply)	100 W
Dynamic power (8 Ω load, at 1% THD with regulated supply)	88 W
Total Harmonic Distortion:	
Rated distortion	1%
IM Distortion:	
10 dB below continuous power output at 60 Hz and 7 kHz (4:1)	0.1%
Sensitivity:	
At continuous power-output rating	700 mV
Hum and Noise:	
Below continuous power output:	
Input shorted	85 dB
Input open	80 dB
Input Resistance	20 KΩ

•With 2-RCA1B01 in output stage.
With 60-volt split power supply and 2-BD450 substituted for 2-BD451 .... 70 W

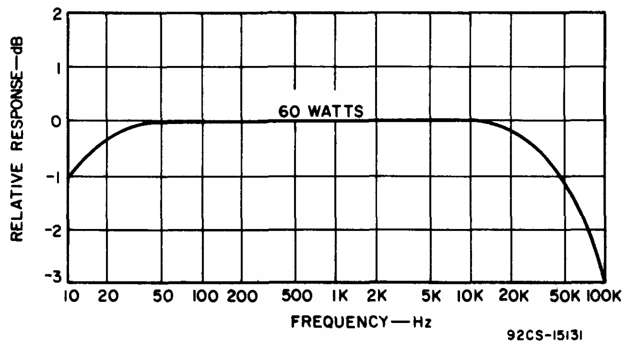
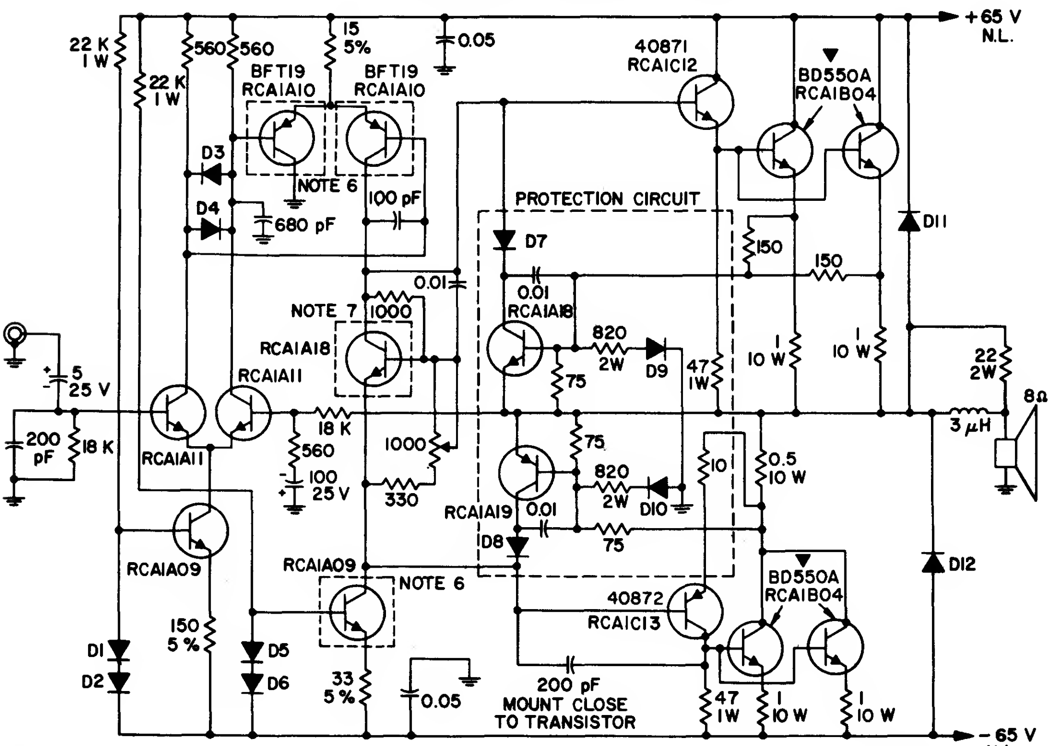


Fig.266 - Typical response as a function of frequency at 60-watt output.



NOTES:

1. D1-D8 - 1N5391; D9, D10 - 1N4148, D11-D12 -1N5393.
2. Resistors are ½-watt, ±10%, unless otherwise specified; values are in ohms.
3. Non-inductive resistors.
4. Capacitances are in μF unless otherwise specified.
5. ▼ Provide heat sink of approx. 1°C/W per output device with a contact thermal resistance of 0.5°C/W max. and T_A=45°C max.
6. Mount each device on TO-39 heat sink.
7. Attach TO-39 heat sink cap to device and mount on same heat sink with the output devices.

Fig.267 - 120-watt amplifier circuit featuring quasi-complementary-symmetry output circuit with parallel output transistors.

Typical performance data are shown in Table XIX. Fig. 269 shows distortion as a function of power output, and Fig. 270 as a function of frequency.

Table XIX - Typical Performance Data for 120-Watt Audio Amplifier

Measured at V_{CC}=130 V, T_A=25° C, and a frequency of 1 kHz, unless otherwise specified.

Power:	
Rated power (8 Ω load, at rated distortion)	120 W
Typical power (4 Ω load)	180 W*
Typical power (16 Ω load)	70 W
Total Harmonic Distortion:	
Rated Distortion	0.5%
IM Distortion:	
10 dB below continuous power output at 60 Hz and 7 kHz (4:1)	0.2%
Sensitivity:	
At continuous power output rating	900 mV
Input Resistance	18 KΩ
IHF Power Bandwidth:	
3 dB below rated continuous power at rated distortion	5 Hz to 50 kHz
Hum and Noise:	
Below continuous power output:	
Input shorted	104 dB
Input open	88 dB
With 2 KΩ resistance on 20-ft. cable on input	104 dB

*With 4-RCA1B04 in output stage.
With a 90-V split power supply and 4-BD550 substituted for 4-BD550A .... 120 W

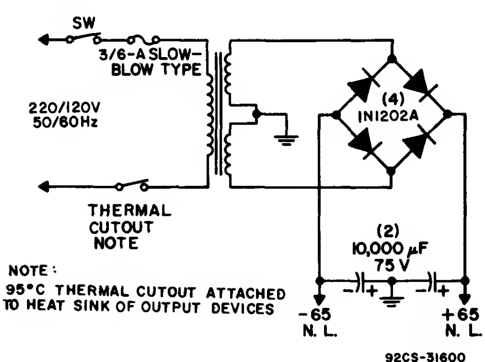


Fig.268 - Power supply for 120-watt amplifier.

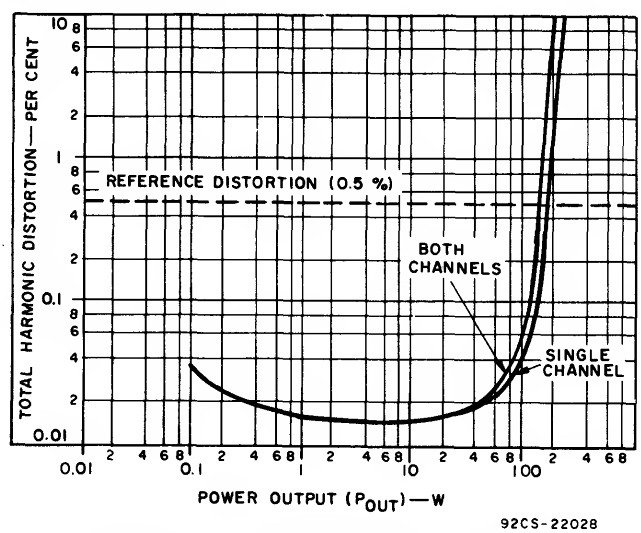


Fig.269 - Typical total harmonic distortion as a function of power output for single channel (8 Ω) and both channels driven at 1 kHz.

AUDIO AMPLIFIER CIRCUITS WITH IC PREAMPLIFIERS AND DISCRETE POWER OUTPUT STAGES

Many modern high-fidelity amplifiers use integrated circuits as preamplifiers and pre-drivers with power transistors in the driver and output stages. Integrated circuits usually offer some performance and cost advantages over discrete transistors for the low-power stages. Table XX is a selection chart for amplifiers in this classification with power output capability from a few watts to several hundred watts. A circuit diagram and performance data are shown for representative amplifiers. For information on the other audio amplifiers listed in the chart, refer to RCA **"Audio Amplifier Manual,"** APA-551, also the individual data sheets for the outut devices.

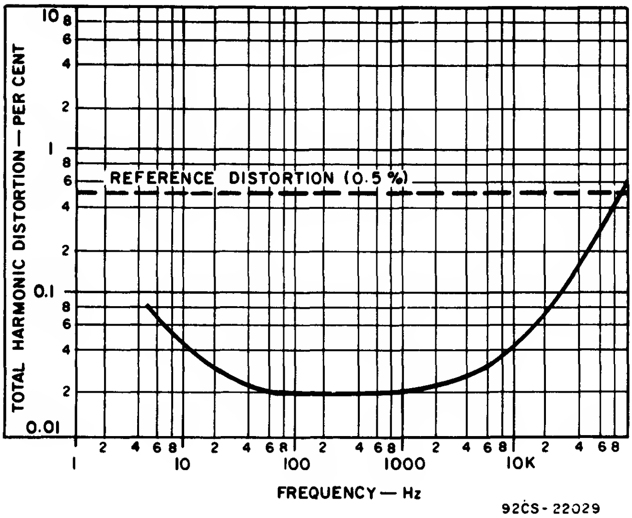
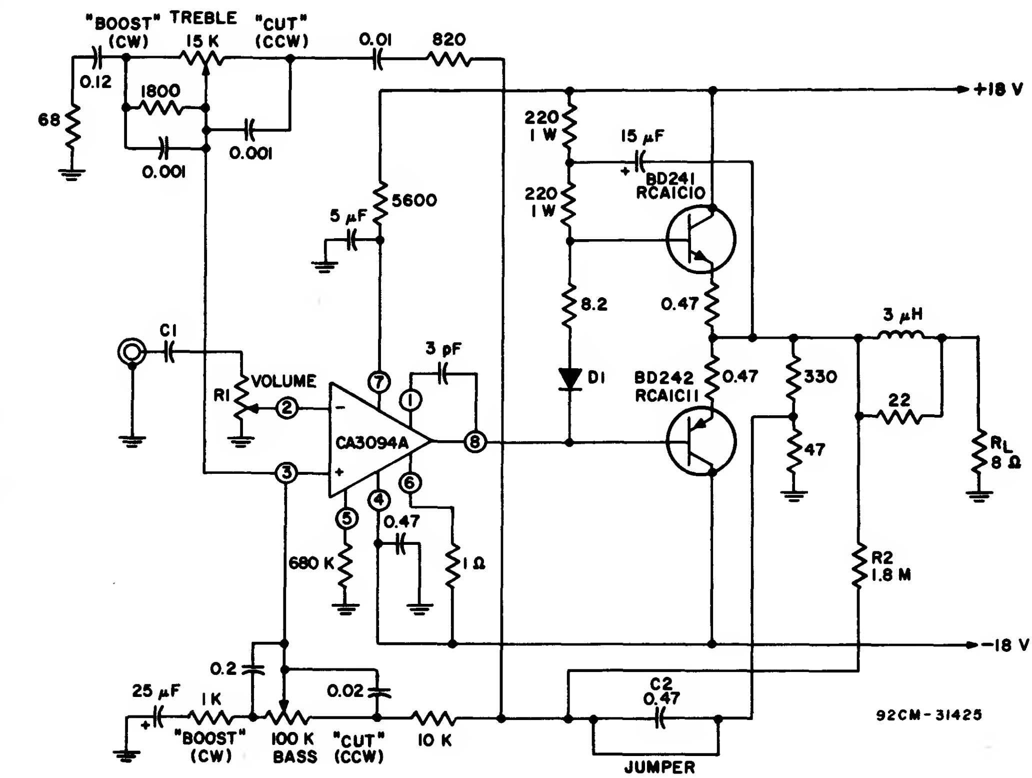
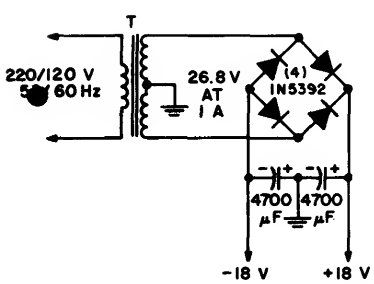


Fig.270 - Typical total harmonic distortion as a function of frequency for 60-watt output.

Table XX - Selection Chart for IC Preamplifiers with Discrete Power Output Stage Amplifiers

Power Output (W)	Load Res. (Ω)	Supply Voltage (V)	IC Type No.	Type of Output Circuit	Output Transistors NPN	Output Transistors PNP
4	8	12	CA3020	Single Class A	—	2N6107
9 12 6	4 8 16	36	CA3094A	True-Comp.	2N6290 or BD241	2N6109 or BD242
10	4	14.4	CA3094	Bridge	2-2N6288	2-2N6111
25 15 8	4 8 16	40	CA3094B	True-Comp. Darlington	2N6388 or BDX33A	2N6668 or BDX34A
30 60	8 4	60	CA3100	True-Comp. Darlington	2N6385	2N6650
40 20	8 16	14.4	2-CA2002	Push-Pull	2-2N6486	—
20	8	50	CA3140B	True-Comp. Darlington	BDX33B or RCA1C15	BDX34B or RCA1C16
50 100 150 100	8 4	72 108 120 90	CA3140A	True-Comp.	RCA8638 2-MJ15003 3-MJ15003 4-BD550 or 4-RCA8638D	RCA9116 2-MJ15004 3-MJ15004 —
100 60 300	8 16 4	114 114 120	CA3100	Quasi-Comp.	4-BD550A or 4-RCA1B04 4-BD550A or 4-RCA1B04 18-BD550A or 18-RCA1B05	— — —
300 160	8 16	172 172	CA3100	Quasi-Comp.	18-BD550B or 18-RCA1B05 18-BD550B or 18-RCA1B05	— —





NOTES:
1. FOR STANDARD INPUT: Short C_2; R_1=250K; C_1=0.047 μF; Remove R_2.
2. FOR CERAMIC-CARTRIDGE INPUT: C_1= 0.0047 μF; R_1=2.5 MΩ; Remove Jumper from C_2; Leave R_2.
3. D1 - 1N5392
4. Resistors are ½-watt unless otherwise specified; values are in ohms.
5. Capacitances are in μF unless other specified.
6. Non-inductive resistors.

Fig.271 - 12-watt amplifier circuit featuring an integrated-circuit driver and a true-complementary-symmetry output stage.

12-Watt True-Complementary-Symmetry Audio Power Amplifier

The CA3094-series IC power amplifiers have a configuration and characteristics well suited for driving complementary discrete power-output transistors. The circuit of Fig. 271 shows an amplifier of this type that can supply 12 watts output to an 8-ohm load from a 36-V split power supply with very low harmonic and intermodulation distortion. Fig. 272 shows IMD as a function of power output.

The large amount of loop gain and the flexibility of feedback arrangements with the CA3094 make it possible to incorporate the tone controls into a feedback network that is closed around the entire amplifier system. Fig. 273 shows voltage gain as a function of frequency with tone controls adjusted for "flat" response and for responses at the extremes of tone-control rotation. The use of tone controls in the feedback network results in excellent signal-to-noise ratio.

Diode D1 may be mounted on the output-transistor heat sink for improved thermal stability. Typical performance data are shown

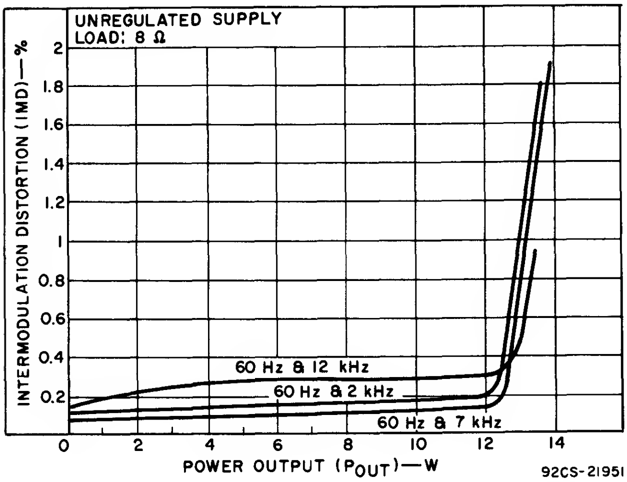


Fig.272 -
Intermodulation distortion as a function of power output.

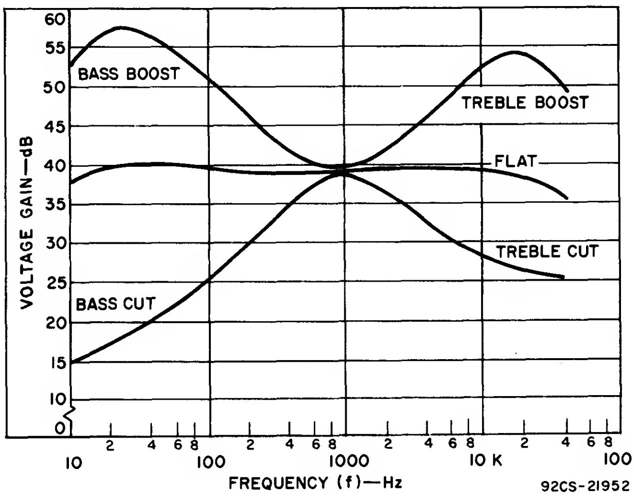
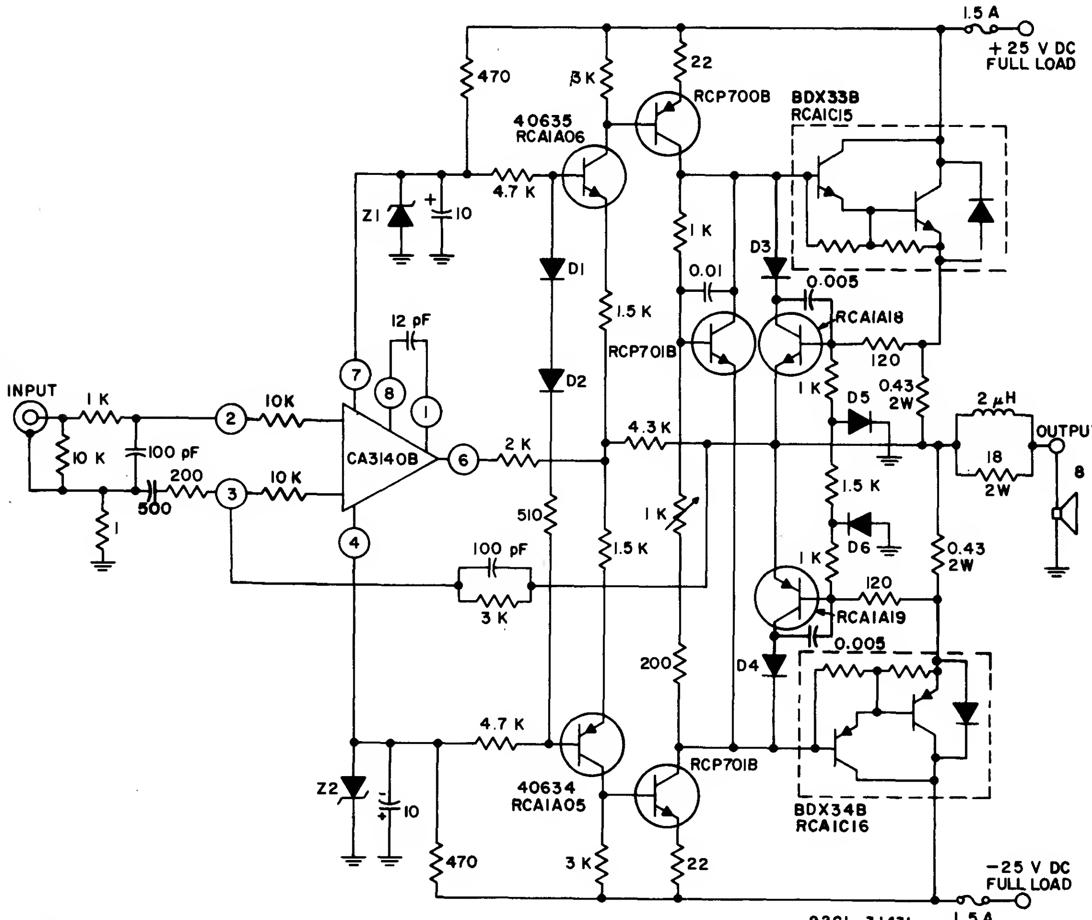


Fig.273 -
Voltage gain as a function of frequency.

Table XXI - Typical Performance Data for 12-Watt Audio Amplifier Circuit

Measured at V_{CC}=36 V, T_A=25° C, and a frequency of 1 kHz, unless otherwise specified.

Power:	
Rated power (8 Ω load, at rated distortion)	12 W
Typical power (4 Ω load)	9 W
Typical power (16 Ω load)	6 W
Music power (8 Ω load, at 5% THD with regulated supply)	15 W
Total Harmonic Distortion:	
Rated distortion	1%
Typical at 1 W	0.05%
IM Distortion:	
10 dB below continuous power output at 60 Hz and 2 kHz (4:1)	0.2%
Sensitivity:	
At continuous power-output rating (tone controls flat)	100 mV
Hum and Noise:	
Below continuous power output:	
Input open	83 dB
Input Resistance	250 KΩ
Voltage Gain	40 dB
Tone Control Range	See Fig. 273



NOTES:

1. D1-D2 - 1N3754; D3-D6 - 1N4148
2. Z1-Z2 - 1N4744.
3. Resistors are ½-watt, ±10%, unless otherwise specified; values are in ohms.
4. Capacitances are in μF unless otherwise specified.
5. Non-inductive resistors.
6. Heat sink per output transistor 1.3° C/W max. thermal resistance.

Fig.274 - 20-watt audio-amplifier circuit featuring full-complementary-symmetry with Darlington output transistors.

in Table XXI, for operation with 4-ohm, 8-ohm, and 16-ohm loads.

20-Watt True-Complementary-Symmetry Audio Amplifier

Fig. 274 shows a circuit with the CA3140B IC driving an amplifier with complementary Darlington transistors in the output stage. The CA3140B-series BiMOS IC's have a configuration and characteristics nearly ideal for driving complementary discrete power amplifier transistors.

The circuit of Fig. 274 is capable of supplying 20 watts output to an 8-ohm load with very low distortion using a 50-volt split power supply. Typical performance data are shown in Table XXII. Total harmonic distortion as a function of power output is shown in Fig. 275 and as a function of frequency in Fig. 276. Intermodulation distortion is shown in Fig. 277 and frequency response in Fig. 278.

Table XXII - Typical Performance Data for 20-Watt Audio Amplifier

Measured at V_{CC}=50 V, T_A=25° C, and a frequency of 1 kHz, unless otherwise specified.

Rated Power:	
8-Ohm Load	20 W
Total Harmonic Distortion:	
(THD)	See Figs. 275 and 276
Intermodulation Distortion:	
(IMD)	See Fig. 277
Sensitivity	0.85 V for 10 W
Input Impedance	10 KΩ
Hum and Noise:	
Below rated power output	
Open input	94 dB
Shorted input	97 dB
Phase Shift	+1.5° at 20 Hz -6° at 20 kHz
Slew Rate	30 V/μs
Rise Time	1.3 μs
Damping Factor	220

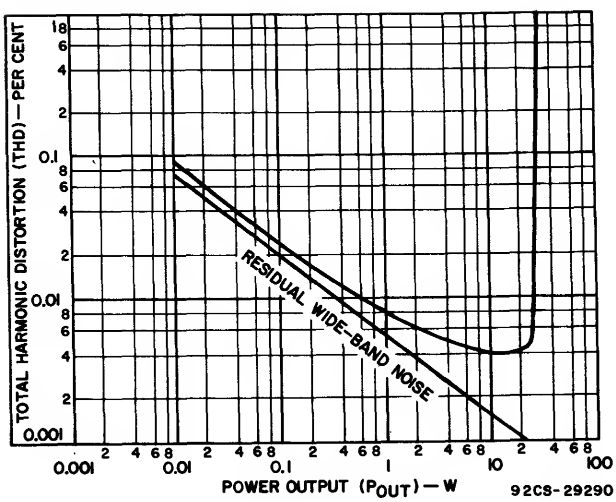


Fig.275 -
Typical total harmonic distortion as a function of power at 1 kHz, both channels driven.

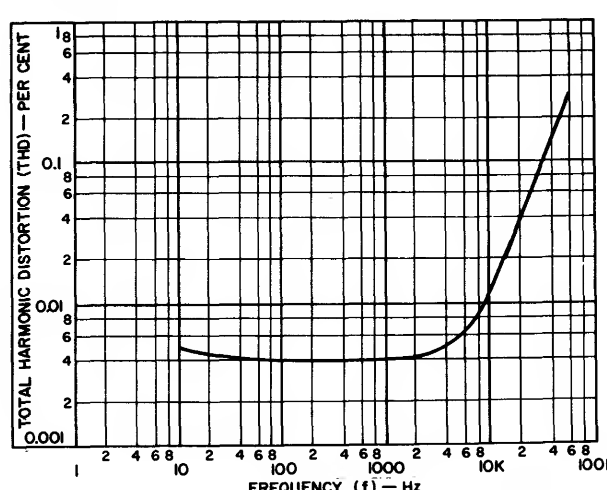


Fig.276 -
Typical total harmonic distortion as a function of frequency for 20-watt output.

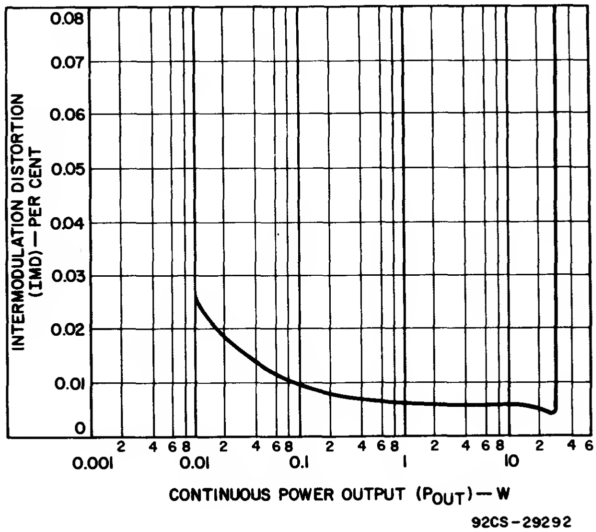


Fig.277 - Typical intermodulation distortion as a function of power at 60 Hz and 7 kHz, 4:1.

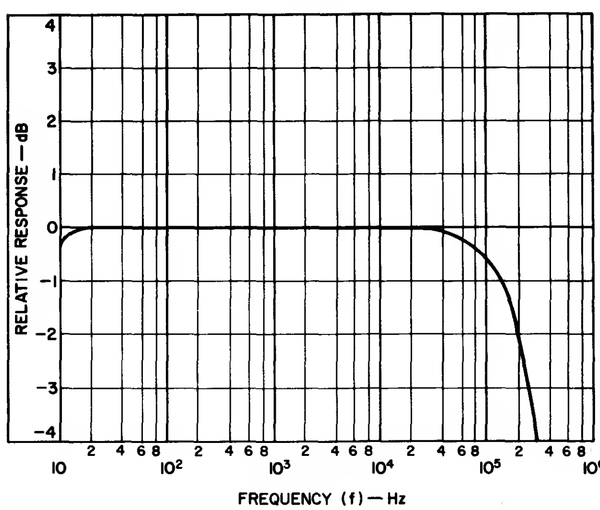


Fig.278 - Typical frequency response.

100-Watt True-Complementary-Symmetry Audio Amplifier

Figs. 279 and 280 show a circuit with the CA3140A IC driving an amplifier with two pairs of epitaxial transistors in parallel in a true-complementary-symmetry output stage. The amplifier is capable of supplying 100 watts output to an 8-ohm load, using a 108-V split powe. supply. Typical performance of this amplifier is shown in Table XXIII including operation with both 4-ohm and 16-ohm loads. The harmonic distortion as a function of power output is shown in Fig. 281 and as a function of frequency in Fig. 282. Additional features include thermal overload and reactive overload protection, and instant turn-on with no undesirable transients.

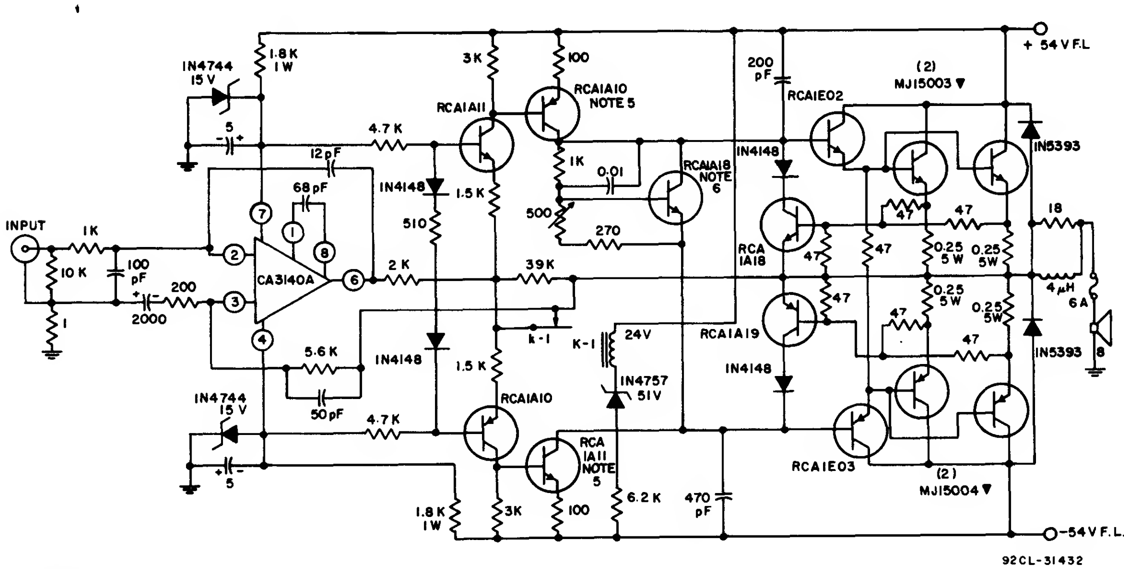
With a single pair of output transistors of the same type, or with the substitution of the RCA8638 and RCA9116 in the output stage, the amplifier is capable of 50 watts power output using a 72-V split power supply. With

Table XXIII - Typical Performance Data for 100-Watt Audio Amplifier

Measured at V_{CC}=108 V, T_A=25° C, and a frequency of 1 kHz unless otherwise specified.

Rated Power (8 Ω load at rated distortion)	100 W
Typical Power (4 Ω load)	150 W*
Typical Power (16 Ω load)	75 W
Total Harmonic Distortion: (THD)	See Figs. 281 and 282
Sensitivity	1 V for 100 W
Input Impedance	10 KΩ
Slew Rate	28 V/μs
Rise Time	1.1 μs
Damping Factor	140

*With 68-V split power supply, 100 W.



NOTES:

1. Resistors are ½-watt, ±5%, unless otherwise specified; values are in ohms.
2. Non-inductive resistors.
3. Capacitances are in μF unless otherwise specified.
4. K-1 relay, single-pole, single-throw, normally closed, with 24 V, 3 mA coil.
5. Mount each on heat sink, 5 sq. in. min. area.
6. Mount on same heat sink with the output devices.
7. ▼ Provide heat sink of approx. 1° C/W per output device with a contact thermal resistance of 0.5° C/W max. and T_A=45° C max.

Fig.279 - 100-watt audio amplifier with parallel output transistors.

three pairs in parallel, it is capable of 150 watts output using a 120-V split power supply, in both cases with an 8-ohm load.

100-Watt Quasi-Complementary-Symmetry Audio Power Amplifier

The circuit shown in Figs. 283 and 284 uses the CA3100 IC as a preamplifier and dual-Darlington-driven parallel output transistors to deliver 100 watts power output to an 8-ohm load from a 114-V split power supply. With the exception of the CA3100, this amplifier is entirely push-pull for improved high-frequency distortion and slew rate. Additional features include thermal overload and reactive overload protection, and instant turn-on with no undesirable transients.

Typical performance data for this amplifier are shown in Table XXIV. The harmonic distortion is shown in Fig. 285, and the response curve in Fig. 286.

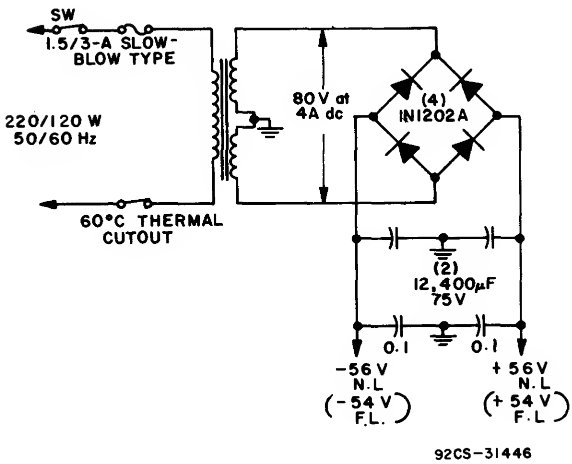


Fig.280 - Power supply for 100-watt amplifier.

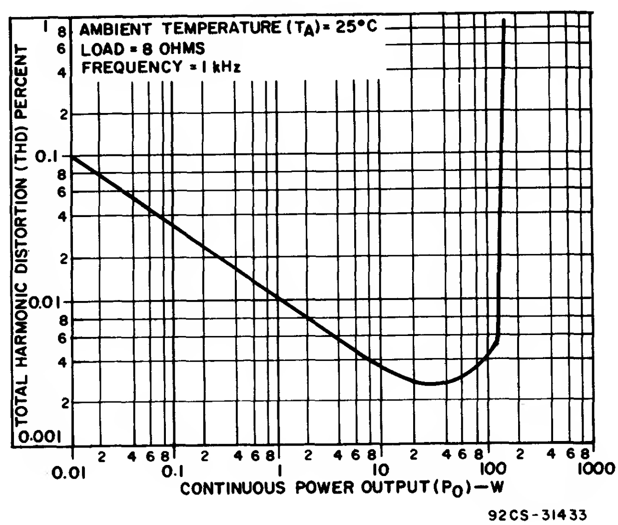


Fig.281 - Typical total harmonic distortion as a function of power output for 100-watt true-complementary amplifier shown in Fig. 279.

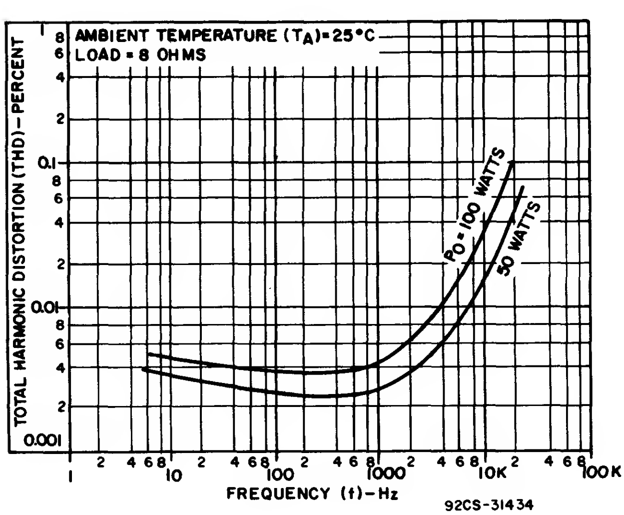
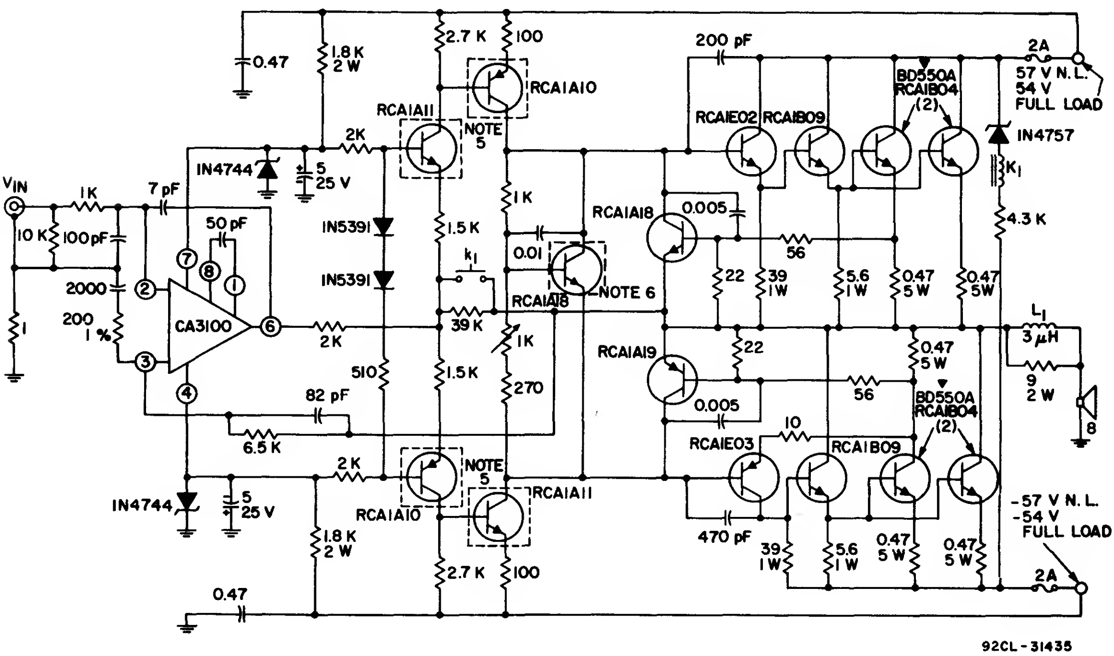


Fig.282 - Typical total harmonic distortion as a function of frequency for 100-watt true-complementary amplifier shown in Fig. 279.



NOTES:

1. Resistors are ½-watt, ±5%, unless otherwise specified; values are in ohms.
2. Non-inductive resistors.
3. Capacitances are in μF unless otherwise specified.
4. K-1 relay, single-pole, single-throw, normally closed, with 24 V, 3 mA coil.
5. Mount on common heat sink, 25 sq. in. min. area.
6. Mount on same heat sink with the output devices.
7. Provide heat sink of approx. 1° C/W per output device with a contact thermal resistance of 0.5° C/W max. and T_A=45° C max.

Fig.283 - 100-watt audio power amplifier featuring parallel output transistors.

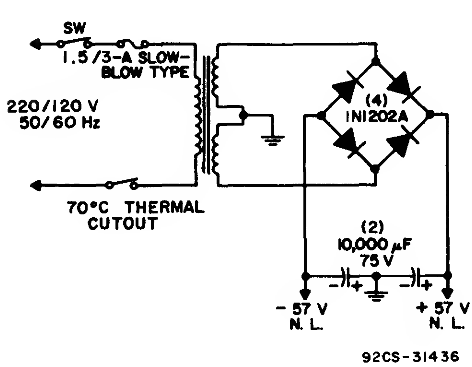


Fig.284 - Power supply for 100-watt audio amplifier.

Twenty-Five-Watt (RMS) True Complementary-Symmetry Audio Amplifier

The twenty-five-watt (rms) complementary-symmetry audio amplifier shown in block form in Fig. 287 uses the BDX33 and BDX34 in conjunction with five TO-92 transistors, two diodes, and a 52-volt (with eight-ohm load) or 40-volt (with four-ohm load) single power supply. The high-frequency performance of this amplifier will satisfy the most critical listener. Table XXV lists typical performance data.

The quiescent current in the class AB output stages (Q3 and Q4) of the amplifier, Fig. 291, has been fixed at 30 milliamperes, which places it above the knee of the h_{fe} characteristics of the BDX33 and BDX34 output devices. The bias that establishes this idling current is provided by the BC237 biasing transistor and can be adjusted by resistor R8. Because the biasing transistor is mounted on the heatsink with the output devices, excellent stabilization of the quiescent current with temperature increase is provided.

It is important to note that, as a result of the high unit-gain frequency of the BDX33 and

Table XXIV - Typical Performance Data for 100-Watt Audio Amplifier

Measured at V_{CC}=114 V, T_A=25° C, and a frequency of 1 kHz, unless otherwise specified.

Power:	
Rated power (8 Ω load)	100 W
Typical power (4 Ω load)	100 W*
Typical power (16 Ω load)	60 W
Total Harmonic Distortion:	
Rated Distortion	See Fig. 206
IM Distortion	< 0.05%
Sensitivity	0.9 V for 100 W
Input Impedance	10 KΩ
Hum and Noise:	
Below rated power output	
Open input	100 dB
Shorted input	106 dB
Phase Shift	+1° at 20 Hz -7° at 20 kHz
Slew Rate	46 V/μs
Rise Time	1.7 μs
Damping Factor	130

*With a 90-V split power supply and 4-BD550 substituted for 4-BD550A.

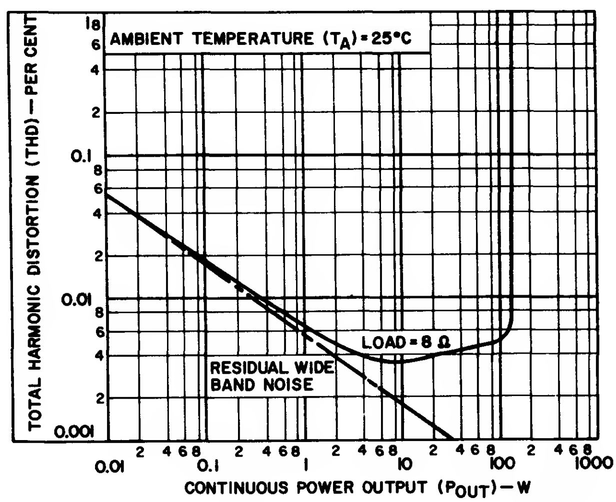


Fig.285 - Typical harmonic distortion as a function of power output for 100-watt amplifier shown in Fig. 283.

BDX34 Darlingtons, the high-frequency operation of the amplifier remains in the highly efficient class B mode, and dissipation and temperature are kept low.

The quiescent current in the driver stage (which must be at least equal to the maximum peak base current required by the n-p-n Darlington) is established by resistors R10 and R11, Fig. 291. The driver current is equal to the difference between the supply voltage and the center voltage divided by the sum of the series resistances (R10 + R11), and is approximately 5 milliamperes.

For proper operation of the circuit, the

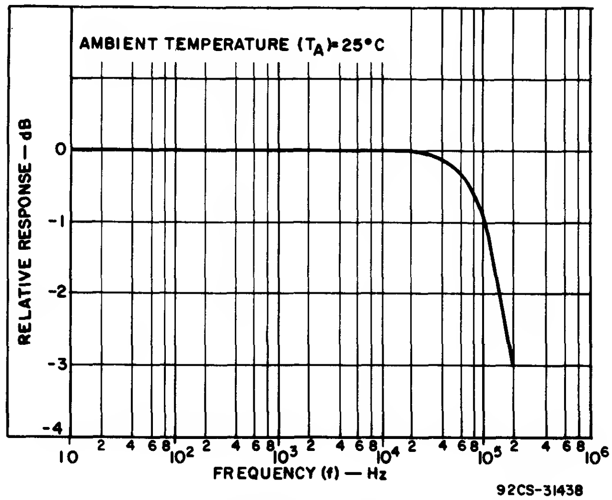


Fig.286 - Typical frequency response for 100-watt amplifier shown in Fig. 283.

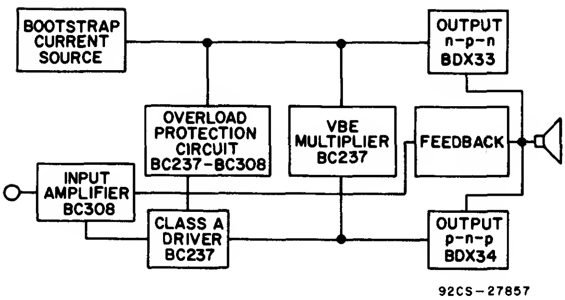


Fig.287 - Block diagram of the 25-watt, full complementary-symmetry, audio amplifier.

Table XXV - Typical Performance Data

All measurements made at an ac line voltage of 220 volts, T_A=25° C.		
Power Output	**Load**	
	8-Ohm	**4-Ohm**
At 1000 Hz for harmonic distortion=1%	26 W	28 W
At 1000 Hz for harmonic distortion < 0.1%	24 W	26 W
Total harmonic distortion as a function of power output at 1000 Hz	Figs. 288 and 289	
Total harmonic distortion as a function of power output at 40 Hz	Figs. 288 and 289	
Total harmonic distortion as a function of power output at 15 kHz	Figs. 288 and 289	
Frequency Response At an output of 15 W: 1 dB down 3 dB down	 40 Hz to 60 kHz 20 Hz to 80 kHz	 50 Hz to 50 kHz 25 Hz to 70 kHz
Sensitivity For power output of 30 W	360 mV	260 mV
Electrical Stability 20 kHz square wave 1 kHz square wave 100 Hz square wave	 Fig. 290(a) Fig. 290(b) Fig. 290(c)	

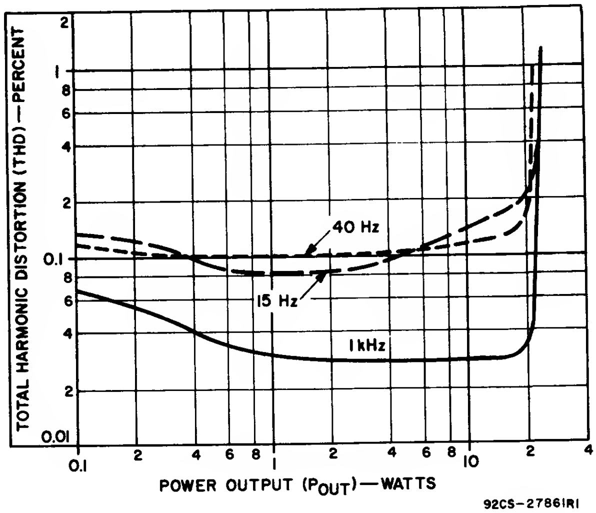


Fig.288 - Total harmonic distortion as a function of power output for the 25-watt amplifier with an eight-ohm load.

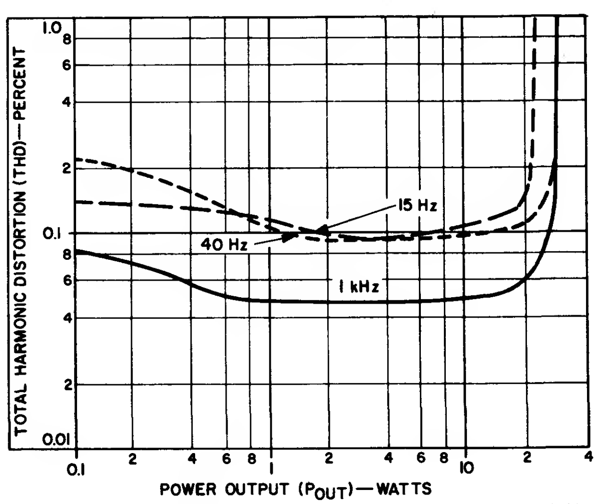


Fig.289 - Total harmonic distortion as a function of power output for the 25-watt amplifier with a four-ohm load.

current I_1 flowing through resistor R10 must remain essentially constant during any excursion of the ac output voltage, Fig. 291. For this reason, a 50-microfarad bootstrap capacitor, C6, is connected between the bias resistors and point A. As the voltage across C6 does not change during ac output-voltage excursions, the change in voltage at point B is the same as at point A. The change in voltage at point C is essentially the same as that at point A; it differs only by the small change in the base-to-emitter voltage of Q3. Therefore, the voltage at points B and C changes by essentially the same amount, and the voltage across resistor R10 remains constant, as does the current I_1.

DC and ac voltages are fed back to the emitter of Q1 to keep the center voltage constant and to assure symmetrical levels of clipping.

20 kHz SQUARE WAVE 2 V / DIV

(a)

1 kHz SQUARE WAVE 2 V / DIV

(b)

Fig.290 - Oscilloscope curves demonstrating amplifier stability: (a) 20-kHz square wave, (b) 1-kHz square wave. Scale on all photos is 2 volts per division.

100 Hz SQUARE WAVE 2 V / DIV

(c) 92CS-28535

Fig.290 - Oscilloscope curves demonstrating amplifier stability: (c) 100-Hz square wave. Scale on all photos is 2 volts per division.

40-Watt Automotive Audio-Power Booster

In recent years, there has been a growing demand for higher power-output capability in automotive tape and audio systems. One of the factors limiting output capability is the 12-volt automotive-system voltage. The following text and illustrations describe the combination of a dc-to-dc regulated up-converter and a simple and economical output amplifier that will deliver 40 watts into a 4-ohm load.

Power Converter

The converter, shown schematically in Fig. 292, is externally excited by an RCA CD4047 integrated-circuit multivibrator operating in the astable mode. The period of the multivibrator is determined by the selection of the values of R and C through the use of the formula: period=4.4 RC. Using the values of R_1 and C_1 indicated in Fig. 292, the period for the circuit shown is 48.4 microseconds, or roughly 20 kilohertz. The Q and $\overline{Q}$ outputs are used to sink base drive for Q_1 and Q_2, which provide primary current for base-drive transformer T_1. Because Q_1 and Q_2 are Darlington transistors, they are able to provide a direct interface with the transformer with minimal loading and high current gain; R_2 limits primary current to 400 milliamperes.

T_1 is a 4-to-1 step-down transformer whose secondary is coupled to the bases of inverter-output transistors Q_3 and Q_4. The center tap of the base-drive transformer secondary is coupled to the common emitter leads of these transistors through the bias network consisting of D_1, D_2, R_3, C_2. This network eliminates common-mode conduction in Q_3 and Q_4, thereby increasing the efficiency and improving the thermal stability of the converter. Note that the bias network does not provide forward bias for starting, but just the opposite, and that transistors Q_3 and Q_4 are operated in a class C manner, with negative bias. When the base-drive signals are near the zero-voltage cross-over point, the negative voltage developed across C_2 is the predominant base signal, and the transistor that was on, and is in the process of turning off, experiences a back bias with an energy content sufficient to

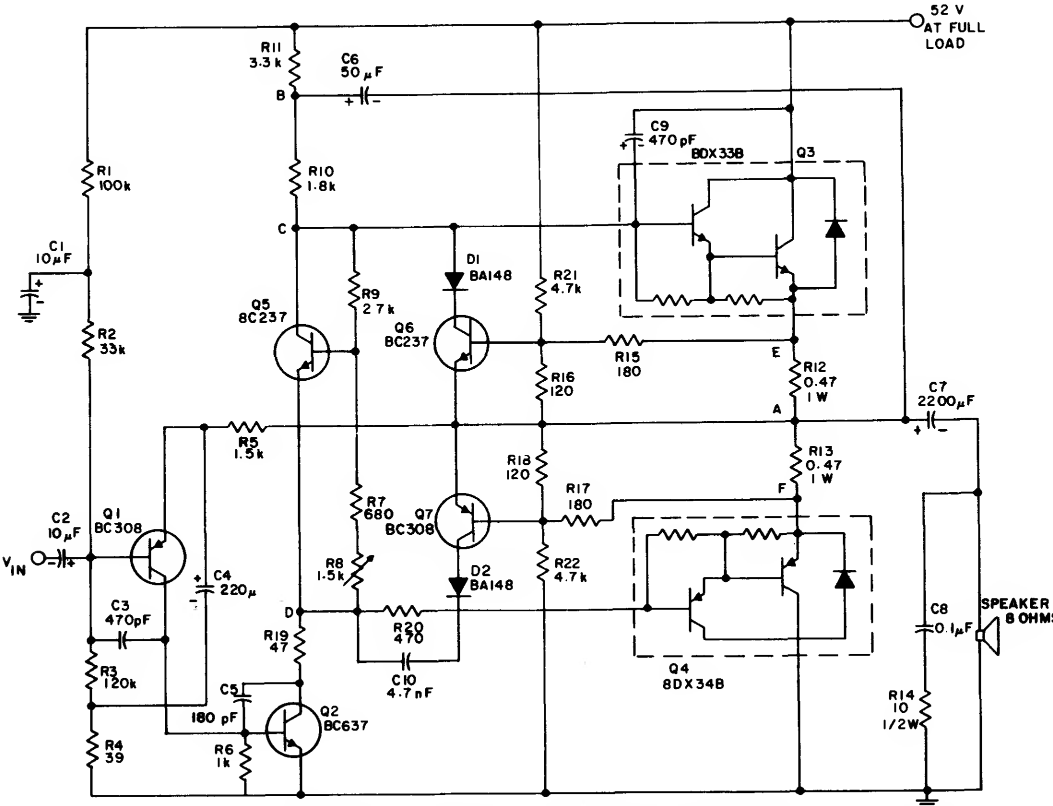


Fig.291 - Schematic diagram of the 25-watt amplifier. Values are given for 8-ohm load.

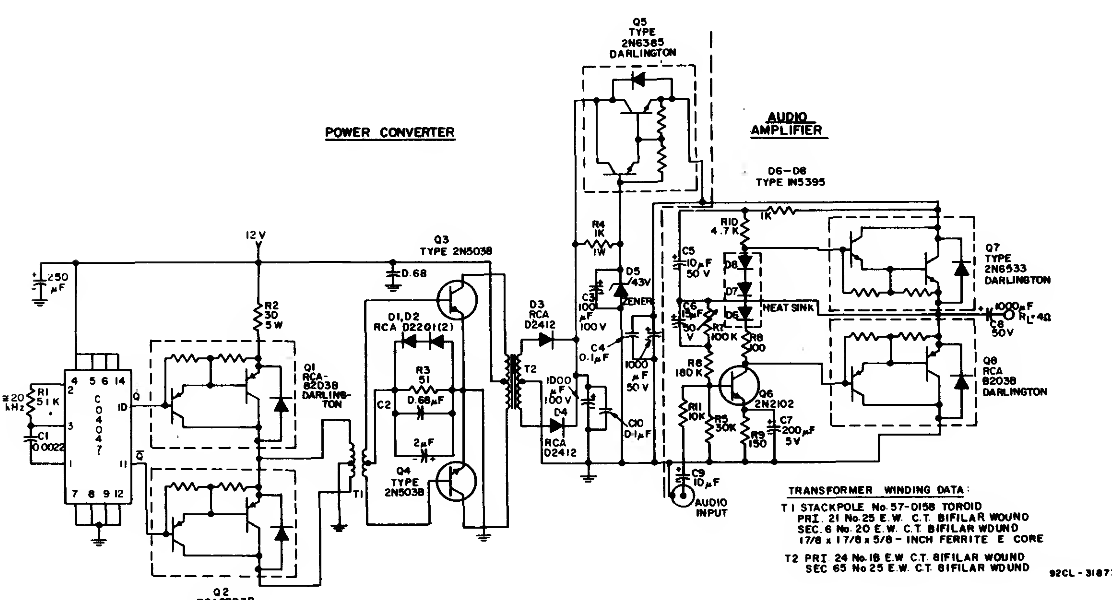


Fig.292 - Schematic diagram of the 40-watt amplifier.

compensate for a worst-case stored-charge condition, thereby reducing switching dissipation.

T_2, a step-up transformer with a turns ratio of 1 to 5.5, provides an unregulated output voltage of 66 volts across filter-capacitor C_{10} via high-speed rectifier-diodes D_3 and D_4. Since, in the automotive environment, the battery voltage varies depending on loading and engine rpm, a series voltage regulator is provided, with a base control voltage determined by zener-diode D_5, to maintain a constant audio-amplifier operating voltage. A Darlington transistor regulator was selected because of its high current gain and minimal base-drive requirements. Resistor R_4 provides bleeder current through the high side of the unregulated supply; C_3 is placed across D_5 to provide zener stability during peak current loading. As a result of the placement of C_3 in the base circuit of Q_5, a capacitor multiplier circuit is formed that reflects an emitter output capacitance of βC_3, thereby further increasing output-voltage stability.

The output voltage of the converter, which is of greater magnitude than the available system voltage, is now applied to the audio-power-amplifier section of the circuit.

The audio amplifier, also shown schematically in Fig. 292, consists of three transistors that provide an ac gain of approximately sixteen. The output devices labeled Q_7 and Q_8 are Darlington transistors configured in a complementary push-pull output arrangement and using a common power supply. A quiescent voltage equal to one-half the supply voltage exists at the common-emitter junction point, and is maintained by the values of R_5, R_6, and R_7, which provide dc feedback. Resistor R_7 is variable and makes possible a fine adjustment of this feedback voltage.

When a complementary pair of output transistors is used (n-p-n and p-n-p), it is possible to design a series-output type of audio amplifier for which the drive circuitry is substantially simplified relative to other amplifier designs; the series-output does not require push-pull drive because phase inversion is unnecessary. The drive in the series-output circuit of Fig. 292 is developed across R_8 and provides the small amount of forward bias required for class AB operation of the complementary pair of output transistors.

Diodes D_6, D_7, and D_8 are also part of the drive circuit; their purpose is to maintain the quiescent current at a reasonable value with variations in junction temperature. To assure thermal tracking, the diodes are mounted on the output heatsink and track with the V_{BE} of the output transistors, thereby providing thermal stability. Transistor Q_6, operated class A, is dc stabilized by R_9; its high ac gain is assured by bypassing the 150-ohm resistor with a 200-microfarad capacitor. In the past, for output power greater than 20 watts, a quasicomplementary output stage was used; this stage used more costly phase-inverter drivers. These phase-inverter drivers were necessary due to the power dissipation in class A operation of the base-driver stage. The design described in Fig. 292 overcomes the need for the phase-inverter drivers due to the high beta of the Darlington transistors. Consequently, the loading of Q6, the base driver, is greatly reduced to the point that a TO-5 package provides ample output base-drive requirements. This design enhances the thermal stability of the base-driver stage over previous designs using class A stages.

The input resistance of the amplifier is determined by R_{11}, and the gain is set by the ratio of R_6 to R_{11}. Capacitor C_5 provides two functions essential to circuit operation. First, it acts as a bypass to decouple power-supply ripple. Second, it is connected as a "bootstrap" capacitor to provide the drive necessary to pull the upper Darlington transistor into saturation. This latter function results from the fact that the stored voltage of the capacitor, with reference to the common output point, provides a higher voltage than the normal collector-supply voltage required to drive transistor Q_7. This higher voltage is necessary during the signal conditions that exist when the upper transistor is being turned on because the emitter voltage of transistor Q_7 then approaches the normal supply voltage. An increase in the base voltage to a point above this level is required to drive Q_7 into saturation. C_9 provides ac coupling for the audio signal while blocking dc that could upset the biasing of transistor Q_6.

The frequency response for the circuit described is well within hi-fi specifications; the circuit provides a response from 40 hertz to 28 kilohertz within a 3-dB tolerance (Fig. 293). The percentage of total harmonic distortion is less than 1 percent over 75 percent of its rated power output, (Fig. 294).

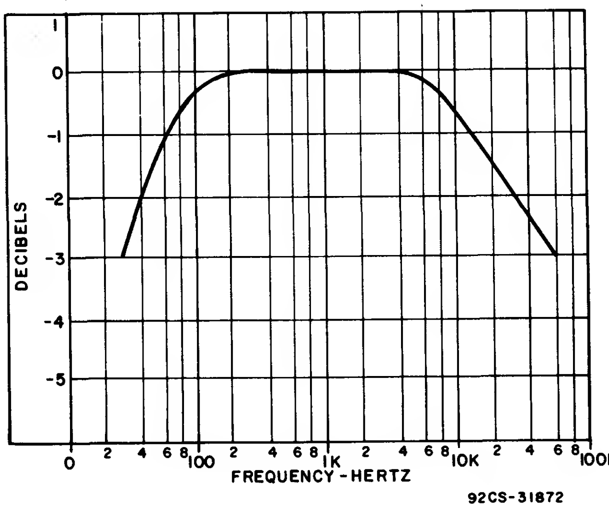


Fig.293 - Frequency response of the 40-watt amplifier with a 4-ohm load.

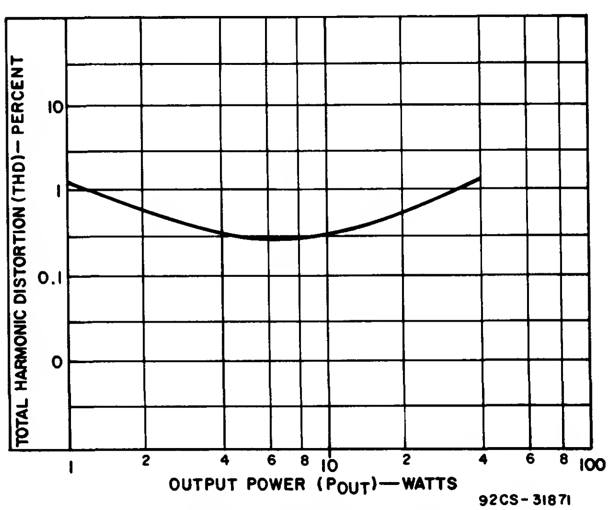


Fig.294 - Total harmonic distortion at 1 kilohertz with a 4-ohm load.

One-Hundred-Watt True-Complementary-Symmetry Audio Amplifier

The BD750 and BD751 series of power transistors are complementary p-n-p and n-p-n series, respectively, selected from the ballasted epitaxial-base silicon transistor families, RCA8638 and RCA9116. They feature high-dissipation capability, low saturation voltage, maximum safe-operating area, a gain-bandwidth product (f_T) higher than 4 MHz, and high gain at high current levels. The transistors are especially suitable for use in the output stage of true-complementary high-power audio amplifiers.

Table XXVI shows the peak load voltages and currents ($\hat{V}_L$ and $\hat{I}$, respectively) required for the various output power levels by the output stage of a 100-watt power amplifier with an 8 or 4-ohm load. The table also shows the supply voltage for a typical design and the required V_{CER} capability of the output transistors.

Table XXVI - Characteristics of 80 and 100-Watt Audio Amplifiers

	Output Power				Units
	80 W		100 W		
Load impedance: R_L	4	8	4	8	Ω
R_E	0.27	0.68	0.27	0.68	Ω
Peak load current - $\hat{I}$	6.4	4.5	7.1	5	A
Peak load voltage - $\hat{V}_L$	25.3	35.8	28.3	40	V
Typical design supply voltage - V_S	70	94	78	104	V
Output device min. V_{CER} required	90	120	100	130	V
Output device max. dissipation under unclipped sine-wave conditions - P_T	29	26	36	32	W
ϕ at clipping for rated output power - ϕ_M	29	28	29	27	°
Protection circuit (limitation line): R_1	180	470	180	470	Ω
(refers to Fig. 295) R_2	3.9	8.2	3.9	8.2	KΩ
R_3	56	75	50	68	Ω
Short circuited output conditions: I_C (peak)	3.5	2.5	3.7	2.65	A
(f=20 Hz, duty cycle 50%) V_{CE} (peak)	34	45	37.7	50	V
P (max)	119	113	140	133	W
Suggested types: n-p-n	BD751	BD751A	BD751B	BD751C	
p-n-p	BD750	BD750A	BD750B	BD750C	
Suggested types: $I_{S/b}$ capability (DC) V_{CE}	35	45	40	50	V
I_C	5.71	4.44	6.25	5	A
P_T	200	200	250	250	W
$R_{\theta(j-c)}$	0.875	0.875	0.7	0.7	°C/W
$\Delta T_{(j-c)}$ (max)	73	69	68	65	°C
$\Delta T_{(c-h)}$ (max)	24	23	28	27	°C
T_h (max)	103	108	104	108	°C
T_{cutout} (max)	95	95	95	95	°C
Heatsink thermal resistance per output device $R_{\theta ha}$ (B) (T_A (max)=45°C)	1.5	1.7	1.2	1.4	°C/W
Max. working case temperature under unclipped sine-wave conditions at T_A= 45°C - T_C (max)	100	100	103	103	°C

The circuit diagram shown in Fig. 295 consists of an integrated circuit input stage and a power stage composed of discrete transistors; it can be treated as two cascaded gain blocks with one common feedback loop. The discrete gain block has its own local feedback provided by R_{17}, R_{10} and C_8. The integrated circuit for the input stage, A_1, is the CA3100, which offers a high unity-gain crossover frequency, wide power bandwidth, a high slew rate, low noise, and low offset. A parts list, parts layout, and printed-circuit board template for the amplifier are provided in the Appendix.

The input stage of the discrete section is a common base stage (Q_{12}, Q_{13}), which serves as a voltage translator and is operated in the class A mode. The next stage (Q_1, Q_2) is also class A operated; its main purpose is voltage amplification. The top and bottom portions of this stage are connected to the V_{BE} multiplier (Q_3), which provides the bias for the output section (driver Q_8, Q_9 and output Q_{10} and Q_{11}). The quiescent current can be adjusted by means of R_{22}; in the practical amplifier under discussion, it was fixed at 200 milliamperes. The driver and output stages are the emitter-follower stages that achieve needed current gain.

A load-line-limiting circuit (Q_6, Q_4 and Q_7, Q_5) is connected across the inputs of the driver stage. As explained above, this load-line

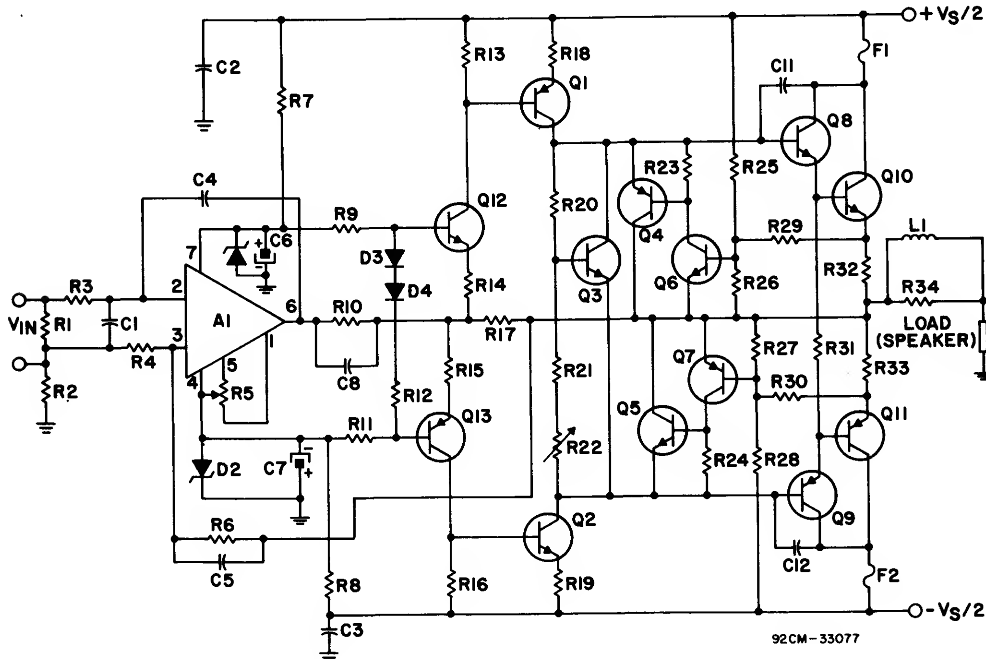


Fig.295 - Circuit of a dc-coupled 100-watt audio amplifier.

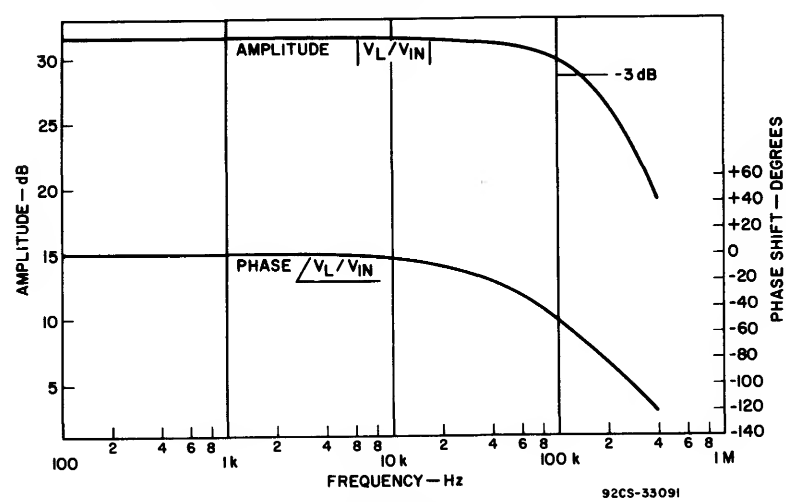


Fig.296 - Typical frequency response of the complete amplifier of Fig. 295.

limiting is necessary to protect the amplifier against excessive dissipation and possible destruction under overload or short-circuited output conditions. In the practical amplifier, two transistors are used for each half of the circuit.

With the component values given in the Appendix, the cut-off frequency at -3 dB of the open-loop response of the discrete section of the 100-watt amplifier is approximately 1 kHz. Fig. 296 gives the typical frequency response of the complete amplifier shown in Fig. 295. Typical performance data for 100-watt audio amplifiers with 4 and 8-ohm loads is provided in Table XXVII and Figs. 296, 297, 298 and 299.

Table XXVII - Typical Performance Data for 100-Watt, 4 and 8-Ohm Audio Amplifiers

Rated Power	100 W	100 W
Load Impedance	4 Ω	8 Ω
Sensitivity	530 mV	750 mV
Input Impedance	10 KΩ	10 KΩ
Slew Rate	25 V/μs	25 V/μs
Frequency Response	See Fig. 296	
Square-wave Response	See Fig. 297	
Total Harmonic Distortion	See Figs. 298 and 299	

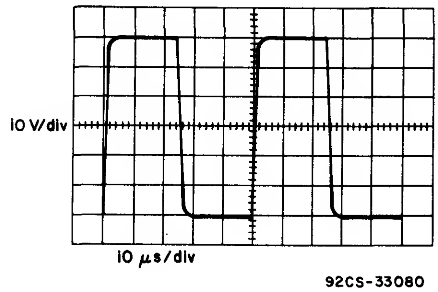


Fig.297 - 20-kHz square-wave output waveform.

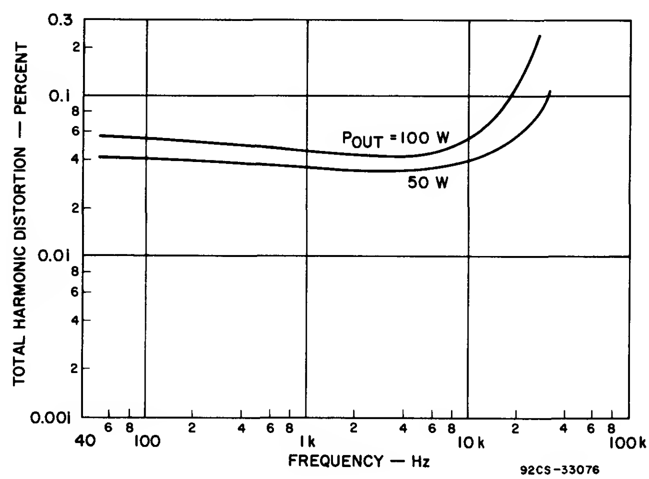


Fig.298 - Total harmonic distortion as a function of frequency.

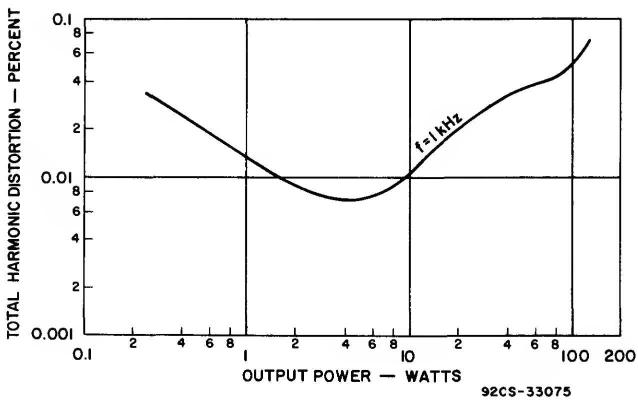


Fig.299 - Total harmonic distortion as a function of output power.

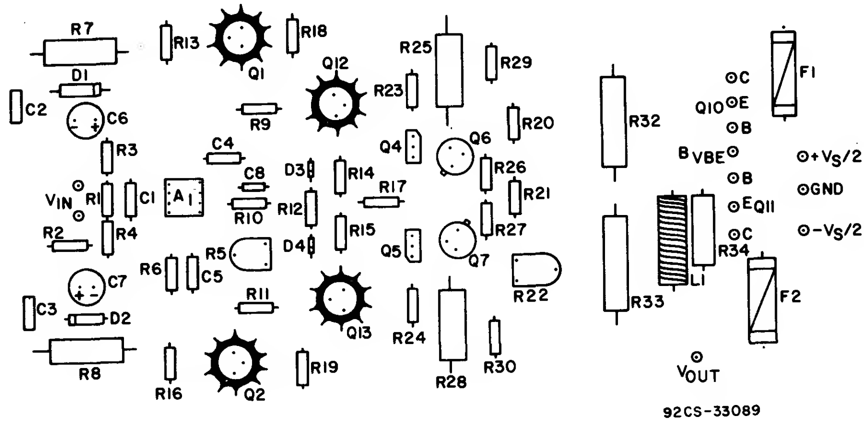


Fig.300 - Components side of PC board.

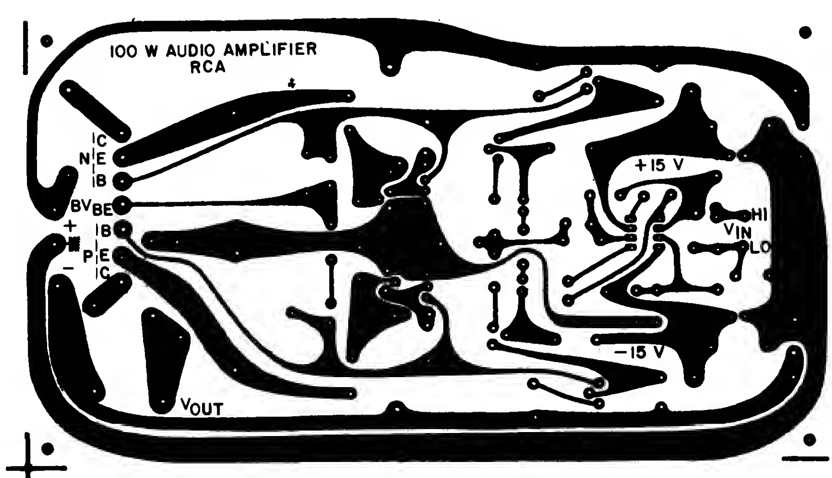


Fig.301 - Copper side of PC board.

Parts List for Amplifier of Fig. 295 with 4 or 8-Ohm Load

Components	4 Ohms	8 Ohms
R1 (Note 1)	10 K	10 K
R2	1	1
R3	1 K	1 K
R4	220	220
R5 (Note 2)	Pot, 10 K	Pot, 10 K
R6	8.2 K	8.2 K
R7	1 K, 1 W	1.8 K, 1 W
R8	1 K, 1 W	1.8 K, 1 W
R9	1.8 K	1.8 K
R10	2.2 K	2.2 K
R11	1.8 K	1.8 K
R12	220	220
R13	4.7 K	1.8 K
R14	820	820
R15	820	820
R16	4.7 K	1.8 K
R17	39 K	39 K
R18	47	47
R19	47	47
R20	390	1 K
R21	56	56
R22 (Note 3)	Pot, 1 K	Pot, 1 K
R23	100	100
R24	100	100
R25	3.9 K, 1 W	8.2 K, 1 W
R26	50	68
R27	50	68
R28	3.9 K, 1 W	8.2 K, 1 W
R29	180	470
R30	180	470
R31 (Note 7)	100	100
R32	0.27, 7 W	0.68, 7 W
R33	0.27, 7 W	0.68, 7 W
R34	4.7, 1 W	10, 1 W

Components	4 Ohms	8 Ohms
C1	100 pF	100 pF
C2	0.47 μF, 50 V	0.47 μF, 50 V
C3	0.47 μF, 50 V	0.47 μF, 50 V
C4	12 pF	12 pF
C5	100 pF	100 pF
C6	22 μF, 25 V	22 μF, 25 V
C7	22 μF, 25 V	22 μF, 25 V
C8	10 nF	10 nF
C11 (Note 7)	3.9 nF	3.9 nF
C12 (Note 7)	3.9 nF	3.9 nF
D1	Zener, 15 V	Zener, 15 V
D2	Zener, 15 V	Zener, 15 V
D3	1N4148	1N4148
D4	1N4148	1N4148
Q1 (Note 4)	RCA1A10	RCA1A10
Q2 (Note 4)	RCA1A11	RCA1A11
Q3 (Note 5)	RCA1A18	RCA1A18
Q4	RCP700A	RCP700A
Q5	RCP701A	RCP701A
Q6	RCA1A18	RCA1A18
Q7	RCA1A19	RCA1A19
Q8 (Note 6)	RCA1C03	2N6474
Q9 (Note 6)	RCA1C04	2N6476
Q10 (Note 6)	BD751B	BD751C
Q11 (Note 6)	BD750B	BD750C
Q12 (Note 4)	RCA1A11	RCA1A11
Q13 (Note 4)	RCA1A10	RCA1A10
A1	CA3100	CA3100
F1	4 A	3 A
F2	4 A	3 A
L1	2 μH	4 μH
V_S	78 V	104 V

Notes for Parts List:

1. All resistors are non-inductive.
2. Adjust for an output of zero volts with zero volts at the input.
3. Adjust for a quiescent current of 200 mA.
4. Mount each device on heatsink of 30 cm^2 minimum area.
5. Mount on same heatsink as driver and output devices Q_8, Q_9, Q_{10} and Q_{11}.
6. Provide heatsinking as described in text.
7. These components cannot be found on the components layout of Fig. 300. They are to be mounted directly on the driver-device sockets that are fixed on the heatsink.