CHAPTER 5

Servo and Audio Power Amplifiers

5.1 Introduction to Servo and Audio Power Amplifiers

The circuits in this chapter are low-distortion power amplifiers for servo and audio applications. All are economical to build, and are efficient for minimum power consumption. The input to each amplifier is assumed to be conditioned to perform the function desired and the level raised to greater than 1 volt rms. For audio amplifiers this would be accomplished in a pre-amplifier which is capable of providing the desired frequency equalization. In servo systems, the pre-amplifier generally performs the required control function rather than amplifying the signal since these systems work at high level.

To satisfy these requirements and to keep the design as simple as possible, the output stage of the amplifiers should operate in class AB. There are many possible approaches to the design of class-AB amplifiers: transformer-coupled, capacitor-coupled, direct-coupled, and various combinations of the three. All have their limitations. Transformer-coupled amplifiers can be made relatively simple, but wideband operation puts fairly stringent requirements on the transformer design. This makes them expensive to use in wideband audio amplifiers but ideally suited to use in narrowband servo systems. Feedback in transformer-coupled amplifiers can lead to ac instability.

Capacitor-coupled amplifiers offer fewer problems with ac stability than transformer-coupled amplifiers, but their design is generally more complicated since more components are required for biasing and more stages are usually needed to compensate for the power loss resulting from impedance mismatch between stages.

Direct-coupled stages have relatively good ac stability, but can cause some difficulty in the area of dc stability, resulting in circuit complexity. Amplifiers using a combination of coupling methods will enjoy the benefits and suffer the limitations of all three methods.

Bias

The bias voltage applied to the base-emitter junction of a transistor amplifier determines its collector current and sets the amplifier stage in class A, B, or C operation. A class-A transistor amplifier conducts 360° (full) of a sine wave. In class-B operation the transistor has approximately zero bias (near cutoff) and allows 180° conduction. In class-C operation the transistor is biased beyond cutoff and allows less than 180° conduction. A transistor which is biased just above cutoff is operating in class AB and allows small-signal class-A operation and large-signal class-B operation. These classes are illustrated in Figure 5-1. Class C amplifiers are excellent for narrowband, high-Q circuits encountered in rf work, but they are almost never used in servo and audio amplifiers because of their high distortion, and they will not be discussed further here.

The load for a class-A amplifier such as shown in Figure 5-2 is generally transformer-coupled to the amplifier since this provides dc isola-

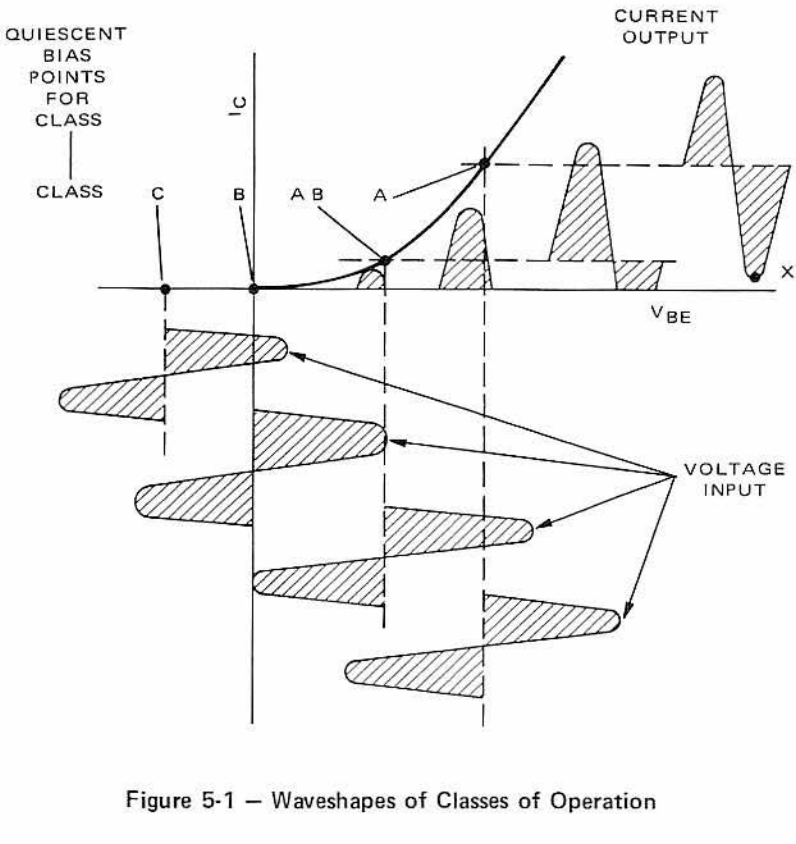


Figure 5-1 — Waveshapes of Classes of Operation

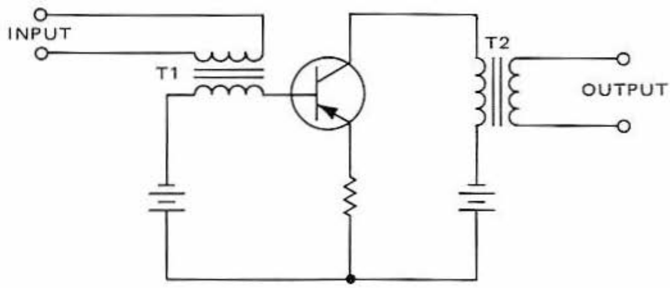


Figure 5-2 – Typical Class-A Amplifier

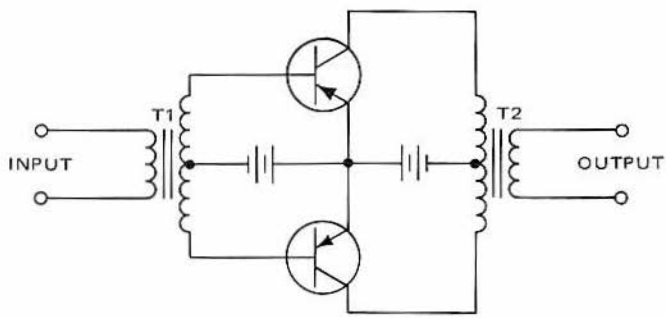


Figure 5-3 – Typical Class-A Push-Pull Amplifier

tion and, through the proper turns ratio, the optimum collector load impedance. The efficiency of the amplifier is defined as the output power at the fundamental frequency divided by the dc power supplied to the collector circuit. The maximum efficiency of class-A operation is obtained when the negative peak of the output signal approaches zero (point X in Figure 5-1) and the positive peak is close to twice the supply voltage. Theoretically the maximum efficiency is 50%, but practical amplifiers operate with less than 35% efficiency. As can be seen from Figure 5-1, the output voltage excursion swings over a nonlinear collector characteristic. The input signal must be adequate to drive the base over this range. Since the device is operated over a nonlinear range, distortion is produced; to minimize the distortion, the power output must be reduced from maximum. Careful selection of the quiescent operating point and the ac load line will give an adequate compromise between maximum power output and minimum distortion. A push-pull class-A amplifier (See Figure 5-3) may be used to give more power output per transistor for a given distortion. For this type of operation there is no dc saturation of the output transformer core and no current of signal frequency through the power source, but special driving and supply circuits must be used. The push-pull class-A amplifier is arranged so that identical transistors are driven 180° out of phase and the outputs combined through the use of a center-tapped output transformer. If transformer drive (T1) is not used, then R-C coupling can be employed through the use of a phase-inversion stage. For relatively low power amplifiers, class-A operation may be sufficient. However when the power level exceeds about 2 watts the power supply and operating costs can be reduced and the efficiency improved by going to class-B or class-AB operation.

A class-AB amplifier is similar to the push-pull class-A amplifier shown in Figure 5-3, but it is biased to a value between cut-off and that required for class-A operation. Class-AB amplifiers are used where higher power is required from a given set of transistors than is obtainable from class-A operation. The instantaneous collector current of each transistor exists for more than 180°, but becomes zero for a small portion of each cycle as can be seen from Figure 5-1. This results in considerable distortion in the individual collector current, but the mutual coupling in the output transformer greatly reduces the effect of the interrupted collector current. Since the quiescent collector current is lower than for class A, a higher supply voltage may be used to obtain the same quiescent collector dissipation. The average value of collector current rises as a base signal is applied, and this increases the power supplied by the collector power supply. Generally, this raises the collector dissipation. The output power is increased over that of class-A operation using the same devices, due to the higher collector voltage and the extended current excursion into the nonlinear region of the collector characteristics. However, for a given voltage and a given load, class A will deliver more output power. Class-AB amplifiers are used where higher output power is required than can be obtained using the same devices in class A, and lower distortion is required than can be obtained with (as will be shown) class B. The typical efficiency that can be obtained for a class-AB amplifier is about 55%.

Class-B push-pull amplifiers are used where higher power is required than can be achieved with the class-AB amplifier using the same devices and power supply. This higher output is achieved at the expense of an increase in the driving power and the distortion. However, the maximum efficiency is increased. The theoretical limit to the efficiency is 78-1/2%, but most practical amplifiers operate around 60% to 65% efficiency. A class-B amplifier is shown in Figure 5-4. For class-B operation the base bias on each transistor is set so that the individual collector currents are nearly

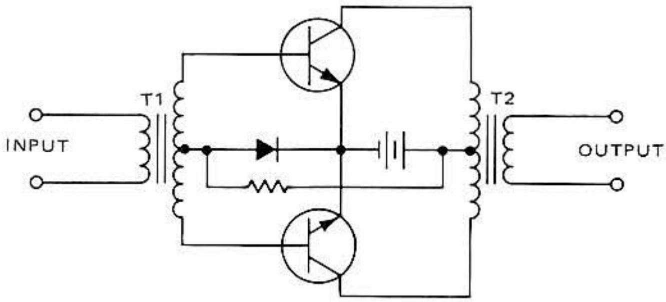


Figure 5-4 – Typical Class-B Push-Pull Amplifier

zero at quiescence as can be seen in Figure 5-1. The drive signal then causes collector current in either transistor only when the base-emitter junction of that device is forward biased. Since the driving signals at the bases are 180° out of phase, only one device is on at any given time. Thus the collector current is proportional to the driving voltage, and the power dissipated in the transistors is small with small signals, as opposed to the large dissipation for class-A amplifiers with low input. A relatively large driving power is required for the transistors when they are driven to achieve the maximum power output. This requires large collector currents and the power supply used must have good voltage regulation. If this is not so, the bias potential would vary with the input signal, leading to severe distortion. The frequency response for the class-B circuit shown in Figure 5-4 depends upon the transformers used, as does the response in the class-A and class-AB amplifiers of Figures 5-2 and 5-3.

Low-frequency rolloff is caused by the primary inductance, whereas high-frequency rolloff is caused by the leakage inductance and winding capacitance. In relatively narrowband servo systems, the use of transformers can greatly ease design problems. However, for audio amplifiers they create problems if wideband operation is desired. The advent of matched complementary transistors has made class-B operation very attractive since these transistors lend themselves readily to direct coupling, and, in addition, a phase inversion is not needed in the driver. This eliminates the need for transformers. One of the basic difficulties encountered with class-B amplifiers is the elimination of crossover distortion. One way to overcome this is to bias for class AB, but this results in higher quiescent power dissipation. With some transistors, it is sometimes useful to drive from a high-impedance, or current, source. This can give low crossover distortion with a minimum of quiescent current.

Feedback

The operating characteristics of a system can be improved through the use of negative feedback. Proper use of negative feedback can increase the bandwidth of an amplifier, improve its gain stability, reduce the noise generated in the stage, lower its output impedance and improve its linearity, which reduces intermodulation and harmonic distortion. Negative feedback must be used with care since it is possible to create regeneration at the band edges even though midband gain is stable. This is caused by excessive phase shift through both the amplifier and the feedback path. It generally occurs where the frequency roll-off characteristics are greatest. There is no simple way to relate the bandwidth of an amplifier with feedback to that without feedback. Therefore each system must be analyzed through the use of the feedback gain equation:

$$A_f = \frac{A}{1 - A\beta},$$

where, A_f is the gain with feedback,

A is the gain without feedback (load connected to the output), and

β is the feedback factor.

In the same manner, the distortion of a stage can be reduced with negative feedback as shown by the equation.

$$D_f = \frac{DA_f}{A} = \frac{D}{1 - A\beta}$$

where, D_f is the output distortion with feedback,

D is the output distortion without feedback, and

A, β and A_f are as defined above

The noise generated in the amplifier stage is reduced according to the following equation,

$$No = Ni\left(\frac{An}{1 - A\beta}\right),$$

where, No is the output noise amplitude,

Ni is the noise amplitude at point of introduction,

An is the gain without feedback from point of introduction of noise to amplifier output, and

A and β are as defined above.

The output impedance of an amplifier is reduced with negative voltage feedback. If negative current feedback is used the output impedance is increased. The equation for the output impedance of an amplifier with negative voltage feedback is

$$Z_f = \frac{Z}{1 - \left(A + \frac{AZ}{R_L}\right)\beta}$$

where, Z_f is the output impedance with feedback,

Z is the output impedance without feedback,

R_L is the load resistance, and

A and β are as defined above.

Another important parameter of an amplifier is its transient response. The transient response is measured by the degree of damping of the amplifier. The system can be either under-damped, critically damped or over-damped. These terms express whether the output waveform, with a step-function input, will overshoot the final output value, rise to the final value in a minimum time with little overshoot or require a long time to acquire the final value. An analytical determination of the transient response is quite difficult because of the varied nature of the amplifier's transfer function. Therefore this characteristic is generally measured by subjecting the amplifier to a step-function input voltage (square wave) and observing the degree to which the output follows the input.

Care must be taken to assure stability if a transformer is within the feedback loop, because of the phase shifts involved. Consider the low-frequency equivalent circuit of a transformer as shown in Figure 5-5, where E_g is the source voltage; R_g, the source resistance; L_p, the transformer primary inductance; and R'_L, the reflected load resistance. For this circuit, the input-output relationship is

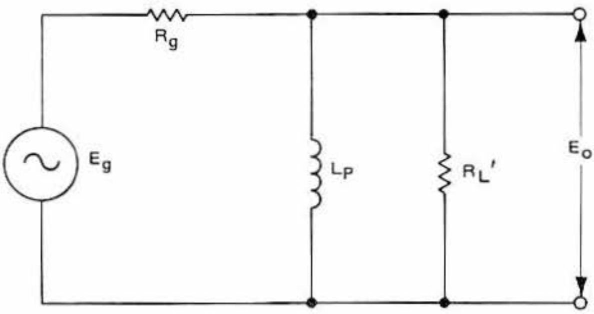


Figure 5-5 – Low-Frequency Equivalent Circuit of Transformer Coupling

$$\frac{E_O}{E_g} = \frac{R'_L\,(j\omega L_p)}{R_g R'_L + j\omega L_p\,(R_g + R'_L)}\,, \tag{1A}$$

and the phase shift is

$$\theta = 90^\circ - \text{ARCTAN}\frac{\omega L_p\,(R_g + R'_L)}{R_g R'_L} \tag{2A}$$

One obvious way to minimize phase shift is to make R_g small. Thus, it is desirable to drive from a low source impedance. Low impedance means an emitter-follower or a common-emitter stage, with heavy degenerative feedback. Neither offers high voltage gain, consequently, another stage would usually have to be added.

Making R_g very small compared to R'_L reduces equation (2A) to

$$\theta = 90^\circ - \text{ARCTAN}\frac{\omega L_p}{R_g} \tag{3A}$$

which indicates that, for minimum phase shift, ωL_p should be large with respect to R_g.

If it is not practical or economical to drive from a low-impedance source, the situation changes somewhat. Considering the case where the transformer is in a transistor collector circuit, the source resistance R_g will generally be much larger than R'_L. Now equation (2A) becomes

$$\theta = 90^\circ - \text{ARCTAN}\,\frac{\omega L_p}{R'_L}\,. \tag{4A}$$

Increasing the ωL_p-to-$R'L$ ratio will reduce the phase shift. If the load is constant (or nearly so), for example, when the feedback resistor is connected to a feedback winding, it is only necessary to make the transformer primary reactance large compared to the reflected load resistance to minimize phase shift. When the load is a transistor base-emitter junction, however, this may not be so easy. The designer does not always have complete limit curves on h_{ie} (input impedance), in which case he would have to make an educated guess or greatly over-design the transformer.

As a final note some factors pertaining to feedback should be mentioned. Negative feedback reduces the overall gain of an amplifier, so the open-loop gain must be high enough that the closed-loop gain will provide the desired results. If this is done, then the $A\beta$ product can be made large, with the result that the amplifier closed-loop gain is virtually independent of the gains of the devices used in the amplifier. This reduction in open-loop gain requires that the input voltage be increased to obtain the same output as obtained from open-loop operation, or in other words, the input must be a high-level signal. These few disadvantages of feedback are far outweighed by the increase in performance obtainable through the use of feedback.

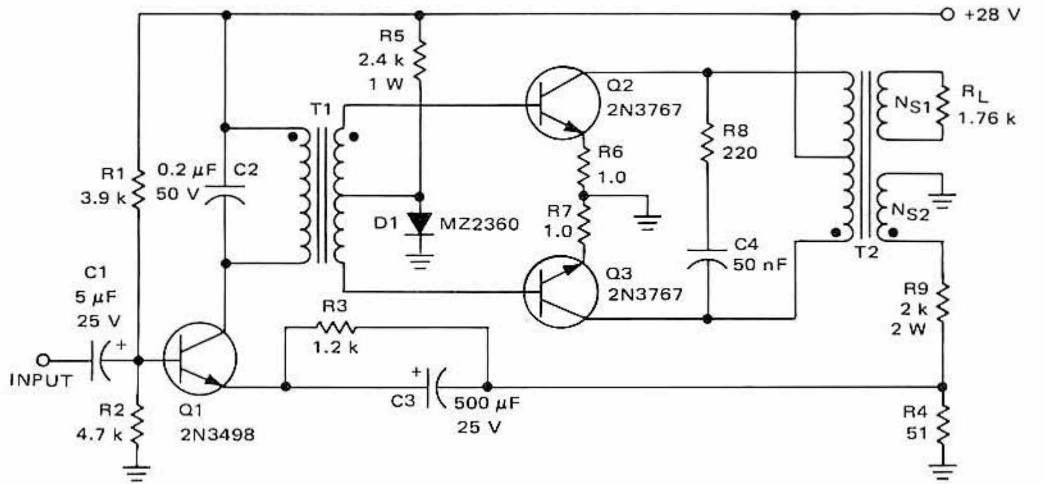
Stability

It is important to maintain a stable operating point that remains unchanged by transistor and temperature variations. An immediately noticeable effect of a substantial change in the quiescent operating point is an increase in distortion. This results from clipping of the peak of the output waveform. The more the operating point shifts, the more the waveform is clipped. If the operating point is not adequately stabilized, thermal runaway may result and this can be fatal to the transistors. The transistor parameters that can be affected by temperature are the base-emitter voltage, collector leakage current and current gain. For stability, the effects of changes in these parameters must be minimized. An effective way to achieve this is to place a small resistor in series with the emitter of each output transistor.

5.2 115 Vac, 7.5 W Transformer-Coupled Servo Amplifier

The transformer-coupled servo amplifiers shown in Figures 5-6 and 5-7 are designed to drive ac motors drawing 7.5 watts. Because excellent impedance matching is achieved with transformers, less power is lost between stages than with other coupling configurations. Thus only three transistors are required to provide a stable voltage gain of 100.

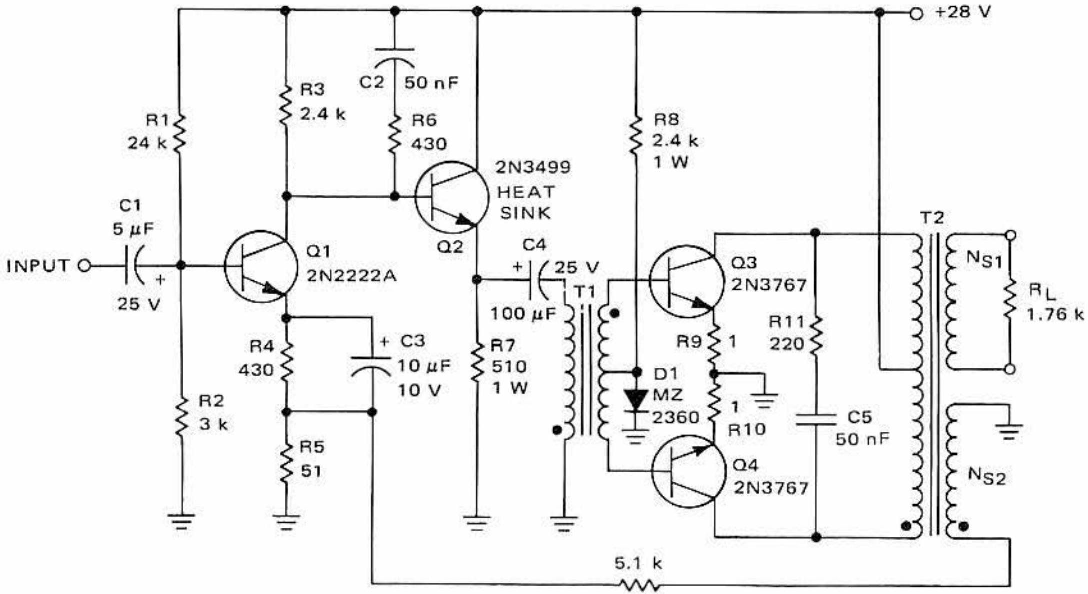
The common-emitter first, or driver, stage for Figure 5-6 contains



All resistors ±5%

Q2 and Q3 mounted on heat sinks with case-to-ambient thermal resistance ≤ 3.5°C/watt (each transistor)

Figure 5-6 — 115 Vrms Transformer-Coupled 7.5 Watt Servo Amplifier



T1, T2 – See Figure 5-8
All resistors ±5%
Q3 and Q4 mounted on heat sinks with case-to-ambient thermal resistance ≤ 3.5°C/watt (each transistor)
Q2 mounted on heat sink with case-to-ambient thermal resistance ≤ 25°C/watt

Figure 5-7 – 115 Vrms 7.5 W Modified Transformer-Coupled Servo Amplifier the driver transformer in its collector circuit. Normally it is undesirable to drive from a transistor collector when the transformer is driving a non-linear load such as a transistor base-emitter junction, but in this amplifier, this problem has been overcome. In order to minimize phase shift when driving from a high-impedance source, the primary shunt inductance of the transformer must be much larger than the reflected load impedance. In the case of a transistor base-emitter load, the highest value of the junction impedance must be considered. Fortunately, a servo amplifier generally works at a nearly constant frequency (or over a fairly narrow band). Consequently, in the amplifier circuit shown in Figure 5-6, a capacitor (C2) was used to tune the driver transformer primary. This minimizes the phase shift. This technique may be used as long as the low-frequency phase shift of the transformer does not cause frequency instability. If wideband operation of the servo amplifier were necessary, if the amplifier tended to be unstable when the feedback loop was closed, or if the amplifier phase characteristics created problems in the total servo loop operation, a common-collector (emitter-follower) stage could be added to provide low-source-impedance drive for the transformer (Figure 5-7). This would give low phase shift across a fairly wide band (less than 10 degrees from 50 Hz to 5 kHz).

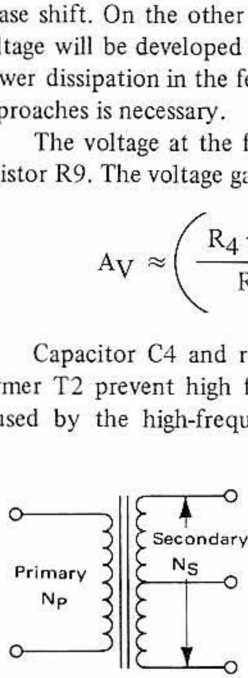
Again referring to Figure 5-6, it can be seen that the output stage is operating class B push-pull. Bias voltage for the two output transistors, Q2 and Q3, is provided by the drop across diode D1. A diode is used for biasing, since it will tend to compensate for variations in the base-emitter voltage of the transistor with temperature, thus increasing thermal stability. It is also important, for good thermal stability, that the secondary winding of the driver transformer have low dc resistance. In this circuit, dc secondary-winding resistance is about 58 ohms total (29 ohms per half). Emitter resistors R6 and R7 have been added to provide additional thermal stability.

As can be seen, feedback for the amplifier is derived from a separate feedback winding rather than from across the load or motor winding. One of the problems associated with transformer-coupled amplifiers is the proper use of feedback. It is rather difficult to carry any reasonable amount of feedback around two transformers. Also when the motor time constant is added, the effective use of feedback becomes even more difficult. One approach would be to add a network to compensate for the motor time constant. Since the time constant can vary quite widely from motor to motor, this would require an individually tailored network for each motor. A more practical solution is the use of a separate feedback winding. This eliminates the effect of the time constant of the motor on feedback loop stability. The difficulty with this approach is arriving at the proper turns ratio between the primary and feedback windings. For minimum phase shift – important with regard to frequency stability – it is desirable to keep the primary-to-feedback winding ratio as small as possible. This results in a lower reflected load impedance, and therefore, less phase shift. On the other hand, if this ratio becomes too low, then a large voltage will be developed across the feedback winding, leading to excessive power dissipation in the feedback resistor. A compromise between the two approaches is necessary.

The voltage at the feedback winding is carried to the input through resistor R9. The voltage gain of the amplifier is given approximately by

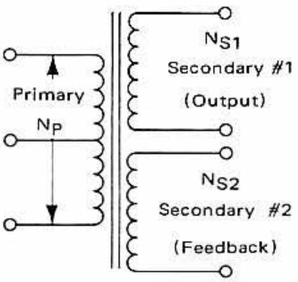
$$A_V \approx \left(\frac{R_4 + R_9}{R_4}\right)\left(\frac{N_{S1}}{N_{S2}}\right)$$

Capacitor C4 and resistor R8 across the primary winding of transformer T2 prevent high frequency instability. This instability is partially caused by the high-frequency characteristics of T2, and consequently,



N_P = 1000 turns No. 34 AWG
N_S = 500 turns No. 36 AWG, center tapped
Wind bifilar on magnetic metals EI-75 square stack core SL-14 material, 5 x 5 interleaved

(A) – DRIVER TRANSFORMER (T1)



N_P = 512 turns No. 24 AWG, center tapped
N_{S1} = 1070 turns No. 28 AWG
N_{S2} = 400 turns No. 28 AWG

Wind bifilar on Arnold AA-81 Silectron Core

(B) – OUTPUT TRANSFORMER (T2)

Figure 5-8 – Transformer Construction

depending upon the construction of T2, R8 and C4 may or may not be necessary.

Although data was taken only on the circuit illustrated in Figure 5-6, the other circuit (Figure 5-7) should perform similarly with one exception. This circuit (Figure 5-7) should have better phase-shift characteristics over a wider frequency range. Of course, it contains one more transistor, and is less efficient owing to the shunt drive of transformer T1.

The operating temperature range of this transformer-coupled servo amplifier is $-55^\circ C$ to $+125^\circ C$. If operation to only $100^\circ C$ is needed, the case-to-ambient thermal resistance of the heat sinks for the output transistors Q2 and Q3 can be increased from $3.5^\circ C$/watt (maximum) to $13.5^\circ C$/watt (maximum).

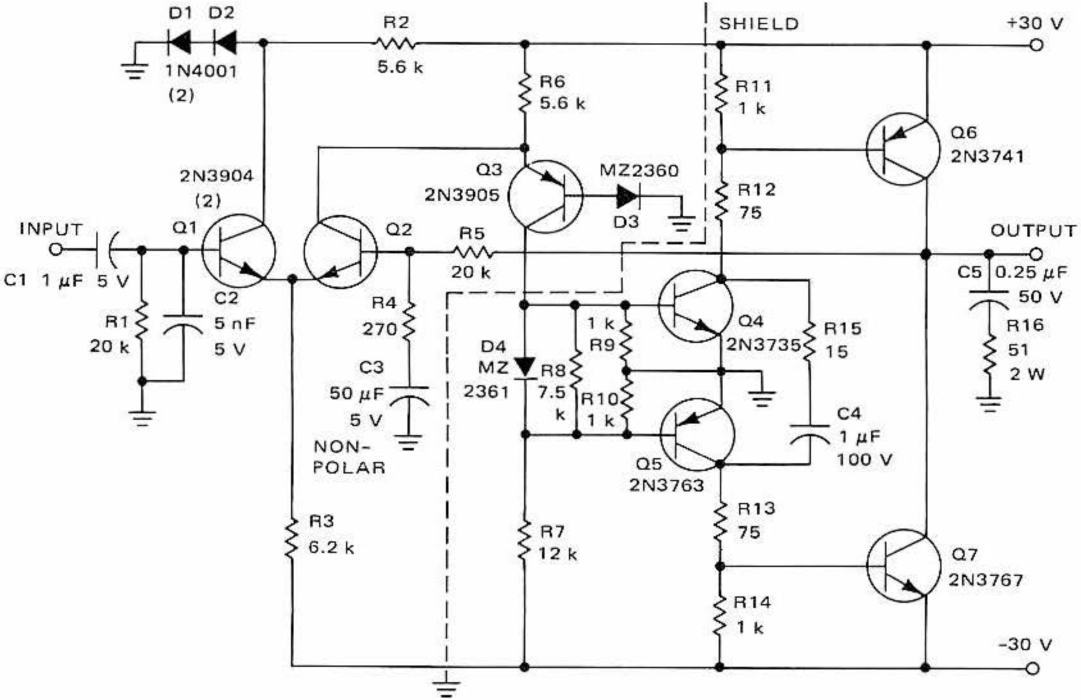
The voltage gain of the amplifier at $25^\circ C$ is 40 dB ± 1 dB, and the gain variation over the operating temperature range of $-55^\circ C$ to $+125^\circ C$ is within ± 2 dB. The power gain of the amplifier is a minimum of 37 dB. Maximum output is 115 volts rms into a 1.75 kΩ load, for an output of 7.5 watts.

Input impedance of the amplifier is 2 kΩ and the output impedance is approximately 200 Ω. Harmonic distortion is less than 5% up to the maximum output voltage of 115 volts rms.

5.3 20 V rms, 10 Watt Complementary Output AC Servo Amplifier

The complementary servo amplifier shown in Figure 5-9 can be used for both ac and dc loads, though several components must be changed for dc operation, as will be explained later. A disadvantage of complementary-amplifier design is that for output voltages of 115 volts rms, output transistors with breakdown voltages (BV_{CEO}) approaching 400 volts would be required. The input and driver transistors would need 200 volt breakdown voltages. The primary advantage of this type design is the elimination of transformers and the use of direct coupling throughout, thus permitting usage with both ac and dc loads. This circuit is designed for applications requiring 20 volts rms, thus it does not require relatively expensive, high-voltage transistors.

For the circuit shown in Figure 5-9, transistors Q1 and Q2 form a differential-amplifier input stage. Resistor R2 and diodes D1 and D2 in the collector circuit of Q1 act as a clamp so that a low voltage transistor can be used for Q1. In a similar manner, the collector of Q2 is clamped by the base-emitter junction of Q3 and diode D3. again allowing the use of a low-voltage transistor. Q3 is connected in a common-base configuration to give a high source-impedance drive for Q4 and Q5, which helps minimize



All resistors ±5%

Q6 and Q7 mounted on heat sinks with case-to-ambient thermal resistance ≤ 3.5°C/watt (each transistor)

Q4 and Q5 mounted on heat sinks with case-to-ambient thermal resistance ≤ 18°C/watt (each transistor)

Figure 5-9 – 20 Vrms 10 W Complementary Servo Amplifier crossover distortion. In addition, diode D4, a dual forward-reference diode, provides enough voltage to bias Q4 and Q5 into conduction. For low crossover distortion (low threshold or dead band for dc applications) the quiescent current through Q6 and Q7 should be between 3 and 5 mA. This value can be adjusted by R8, which shunts D4. For good thermal stability, diode D4 should be mounted directly on the heat sinks of Q4 and Q5, as close to the transistors as possible.

Although diode D3 helps clamp the collector voltage of Q3, it also serves a more useful purpose. Without D3, the base-emitter voltage drop of Q4 would tend to forward-bias the collector-base junction of Q3. Q3 would then operate in the saturation region, resulting in very nonlinear amplification. Diode D3 maintains the collector-base junction of Q3 in a reverse-biased condition.

Transistors Q4 and Q5 form a complementary driver pair which is direct coupled to the complementary output pair, Q6 and Q7. Through the use of positive and negative power supplies, the quiescent dc output voltage is approximately zero, so that the load can be direct coupled to the amplifier output.

Negative feedback is provided via resistor R5. R5 and resistor R4 set the ac voltage gain of the amplifier, which is given (approximately) by

$$A_V \approx \frac{R_4 + R_5}{R_4} \approx \frac{R_5}{R_4} .$$

As was mentioned, the complementary amplifier is capable of driving both ac and dc motors. The circuit in Figure 5-9 is for ac operation only. For dc, capacitors C1 and C3 must be shorted. The amplifier is then direct coupled from input to output. One point should be made, however; for ac operation, when R4 is returned to ground through C3, the dc closed-loop gain is approximately unity. This provides exceptionally good stability of the dc voltage at the output. With resistor R4 connected directly to ground, the dc stability will not be as good, since the closed-loop dc gain will now be equal to the ac gain. In addition, if the input is returned to ground through a low-impedance source, thereby effectively bypassing R1, there will be an offset voltage at the output due to the current flow through R5. (Ordinarily, in the ac amplifier, the offset voltage contributed by R5 is balanced out by the voltage drop across R1.) Increasing R4 and/or decreasing R5 will reduce the initial offset and increase the dc stability of the output for temperature changes. Of course, the gain of the amplifier will be reduced. Matching the two input transistors for current gain and base-emitter voltage drop, and matching the current-gain products of the Q4-Q6 and Q5-Q7 pairs will help reduce the offset voltage and improve the dc stability of the output.

As can be seen, most of the transistors are working into rather low impedances (usually base-emitter junctions). Because of this, the amplifier has a rather wide open-loop bandwidth, and exhibits a strong tendency to oscillate. Several networks may be added to stabilize the amplifier, namely, C2, C4 and R15, C5 and R16, but the most effective way to eliminate oscillation is careful construction. All leads should be kept as short as possible. The input stages should be isolated from the output. It may even be necessary to use a shield as shown in the figure.

The operating temperature range of the amplifier is -55°C to $+100^\circ$C. The upper temperature is limited by plastic transistors Q2 and Q3, which have an upper junction temperature limit of $+135^\circ$C. They were selected on the basis of economy. If $+125^\circ$C operation is required, Q1 and Q2 can be replaced by 2N3947's, and Q3 by a 2N3250. In addition, the case-to-ambient thermal resistance of the heat sinks for Q6 and Q7 must be decreased from a maximum of 3.5°C/watt to 2.0°C/watt.

The amplifier will drive 20 volts rms into a 40 Ω load, giving an output of 10 watts. The voltage gain of the amplifier at 25°C is 37 dB $\pm$ 1 dB. The gain variation over the -55°C to $+100^\circ$C ambient temperature range is less than ± 0.5 dB. Minimum power gain is 60 dB.

Input impedance of the amplifier is 15 kΩ and the output impedance is less than 1 Ω. Harmonic distortion is less than 5% at all levels up to 20 volt rms, and is typically less than 1% at 25°C.

One final note: The voltage at the base of Q2 should not exceed approximately 1.2 volts positive with respect to ground. If this voltage is exceeded, the collector-base junction of Q2 will be forward-biased and Q2 will saturate. This sets a maximum value for the feedback ratio, namely,

$$\frac{R_4}{R_4 + R_5} \leqslant \frac{1.2}{V_{OUT(peak)}} .$$

5.4 115 Vrms, 7.5 W Complementary-Output AC Servo Amplifier

The servo-amplifier circuit shown in Figure 5-10 uses high-voltage transistors to deliver a high output voltage without using an output transformer. This reduces phase shift and its accompanying problems and permits the circuit to be operated from the 117 V ac line without a power transformer. The direct-coupled driver consisting of Q1 and Q2 is used because of its simplicity. It is decoupled from the high voltage supply by

R7 and C2. The feedback is brought through R8 to the emitter of Q2 for stability. If the feedback were returned to the base of Q1, R5 would have to be bypassed to provide adequate open-loop gain, and this increased gain, in conjunction with the phase shift across coupling transformer T1, could cause oscillation. The feedback around T1 reduces the effect of its nonlinearities, and sets the amplifier gain as determined by the following equation,

$$A \approx \frac{R3}{R4} \frac{R8}{R5} .$$

The RC network across the primary of the transformer is used to minimize phase shift at 400 Hz. This narrows the bandwidth, and care must be exercised that additional phase shift at higher frequencies does not cause instability.

Capacitors C4 and C5 in parallel with diodes D1 and D2 provide a drive path for the base currents of Q3 and Q4 when the diodes turn off. For example, refer to Figure 5-10, and note the drive loop for the base drive current indicated for Q3. The base drive current flows through D1 in the reverse direction. This is possible initially because the forward diode current supplied by R9 is greater than the base current. As transistor Q3 is further turned on, however, the voltage at point P_L goes positive and the voltage across R9 will decrease. Therefore, the current supplied by R9 will decrease until the forward current through D1 is less than the base current. When this happens, the diode will turn off and the base current must be driven through some other path. When D1 turns off, the impedance of the drive loop would change from approximately 200 ohms to approximately 50 kΩ if capacitor C4 were removed. This would require a very large drive signal to supply sufficient base current to Q3. The capacitors have therefore been added to maintain a low-impedance drive path for Q3 and Q4. The addition of the capacitors has virtually no effect on the quiescent current.

The output stage is biased to prevent crossover distortion. Diodes were used to bias transistors Q3 and Q4 because the diode voltage drop tends to track the base-emitter drop, which simplifies the problem of thermal stability in the output stage. Due to the differences in the diodes and transistors, a value cannot be given for R9 and R10. They should be picked to produce a quiescent current of approximately 4 mA in Q3 and Q4, and will be approximately 20 kΩ to 50 kΩ. The quiescent currents I_{Q3} and I_{Q4} should be matched to minimize the dc offset voltage at the output. If I_{Q3} is not equal to I_{Q4}, the difference in the two currents will flow through R_L, causing a quiescent dc output voltage.

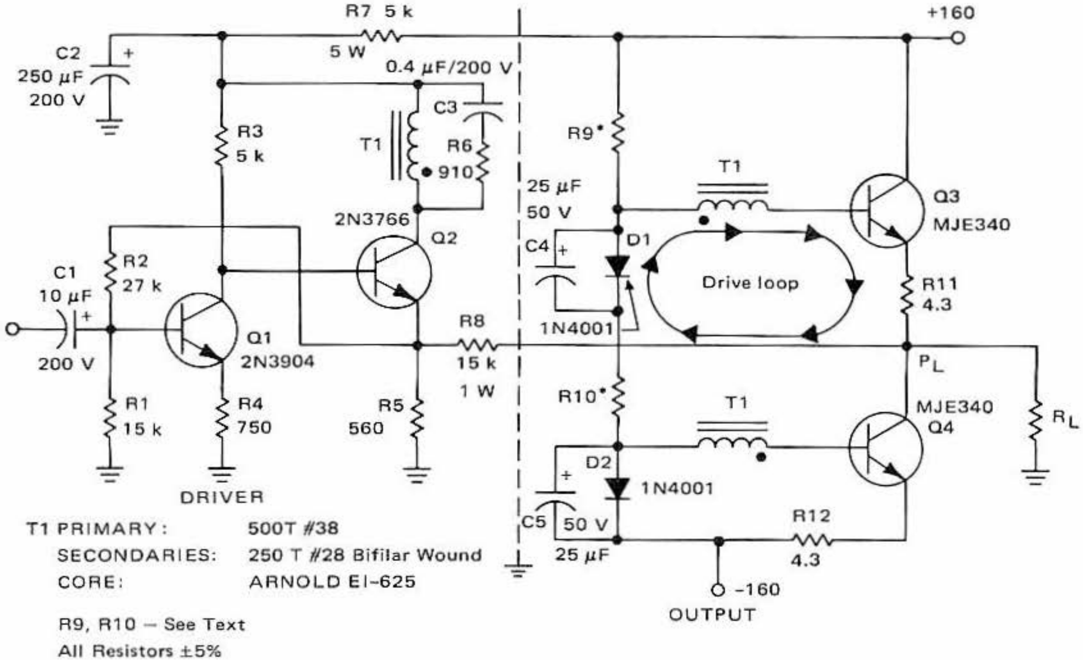


Figure 5-10 – 115 Vrms 7.5 W Complementary-Output Servo Amplifier

The frequency response is primarily limited by the transformer. The phase shift introduced by the transformer is critical because it limits the amount of feedback that can be used.

If a large amount of feedback is used and the transformer phase shift approaches 90°, there will be a strong tendency for the circuit to oscillate. If a wide bandwidth is required, three methods are available to reduce the phase shift of the transformer: (1) Increase the primary inductance of the transformer, (2) decrease the load reflected from the secondary, (3) decrease the impedance of the driving source.

Each biasing diode of the output stage should be mounted on the heat sink close to its associated output transistor so its voltage will track the transistor base-emitter voltage for thermal stability. The amplifier is designed to operate at 100°C if a heat sink of 3.5°C/watt is provided for each transistor. This heat sinking is adequate if R9 and R10 are adjusted so that quiescent current does not exceed 15 mA at 100°C.

The gain variation as a function of temperature is given in Figure 5-11. At −55°C the gain has dropped 1 dB, but from −20°C to +200°C the gain is within 0.25 dB. As can be seen from the plot of output voltage vs. input voltage in Figure 5-12, the amplifier has good linearity up to

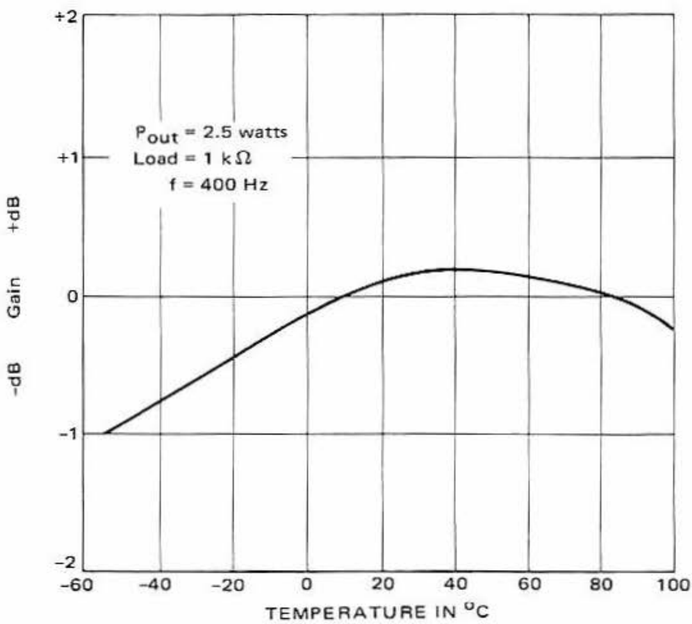


Figure 5-11 – Voltage Gain as a Function of Temperature

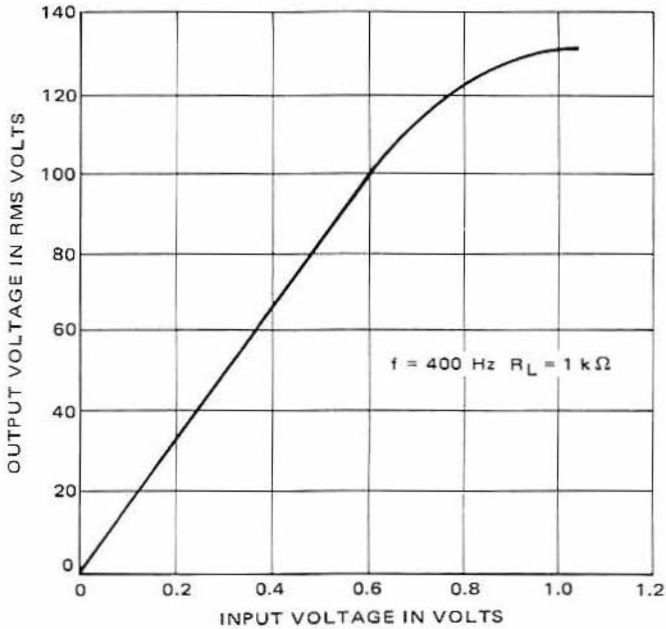
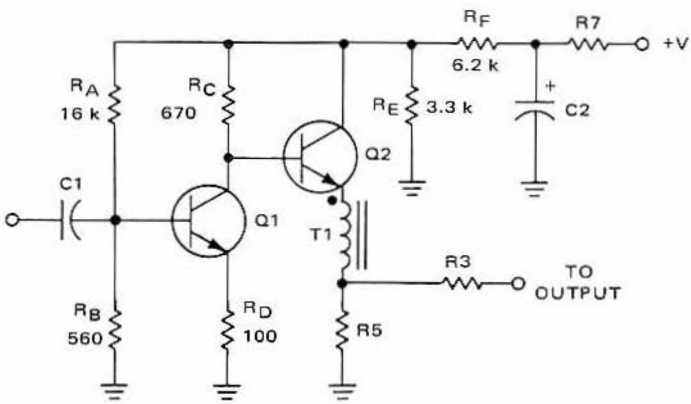


Figure 5-12 – Output Voltage versus Input Voltage



All Resistors ±5%.

Figure 5-13 – Emitter-Follower Driver Stage for Circuit Shown in Figure 5-10

110 volts rms output. When driving a 7.5 watt, 115 volt rms servo motor, the output voltage changed approximately 2 volts from no load to stall conditions. The overall efficiency at 400 Hz with 90 volt rms output is 54.4%.

The frequency response of this amplifier is +0, −1 dB from 95 Hz to 2 kHz. The response is limited primarily by the secondary interwinding capacitance of T1 which causes the gain of the output stage to roll off at high frequencies. To overcome the capacitive effects, a low driving-source impedance such as the emitter follower shown in Figure 5-13 can be used. This driving circuit extends the 1 dB frequency response to 30 Hz and 12.5 kHz. The other operating parameters for the circuit using the emitter-follower driver are the same.

5.5 28 V, 28 Watt DC Servo Amplifier

The schematic shown in Figure 5-14 is that of a four-stage, direct-coupled servo amplifier capable of driving a 1 ampere load. The input stage is a differential amplifier formed by transistors Q1 and Q2. One input to this stage is the amplifier input and is applied to the base of Q1 while the other input to the stage is obtained from the amplifier output, as will be explained. The input resistors (R1 and R2) are used to give the same driving impedance to Q1 as R7 and R8 present to Q2. The emitters are connected to the negative supply through a constant-current source comprised of Q3, D1, R6 and R5. This constant-current source provides stable operation of the first differential-amplifier stage. The differential output voltage of the first stage is applied to the second stage (Q4 and Q5) which is a differential-to-single-ended converter. High-frequency roll-off is provided by the series combination of C1 and R10 between the collectors of Q4 and Q5. This roll-off is used to increase the stability margin of the stage.

The drive for the driver stage (Q6 and Q7) is obtained from the collector circuit of Q5. This connection provides a relatively high source impedance to Q6 and Q7, the power drivers. A practical way of achieving a low threshold or dead band area at a minimum of quiescent power dissipation, is to drive from a high-impedance source such as this. Diode D2 provides enough voltage to bias Q6 and Q7 into conduction. This complementary pair is directly coupled to the output pair, Q8 and Q9. The driver and the output stage are complementary devices since this keeps the design simple. There is no need for an additional stage of phase inversion and the two stages are readily adapted to direct coupling. Two complementary stages are used since this achieves maximum utilization of devices.

The use of positive and negative power supplies sets the quiescent output voltage within a few millivolts of zero. Negative feedback is pro-

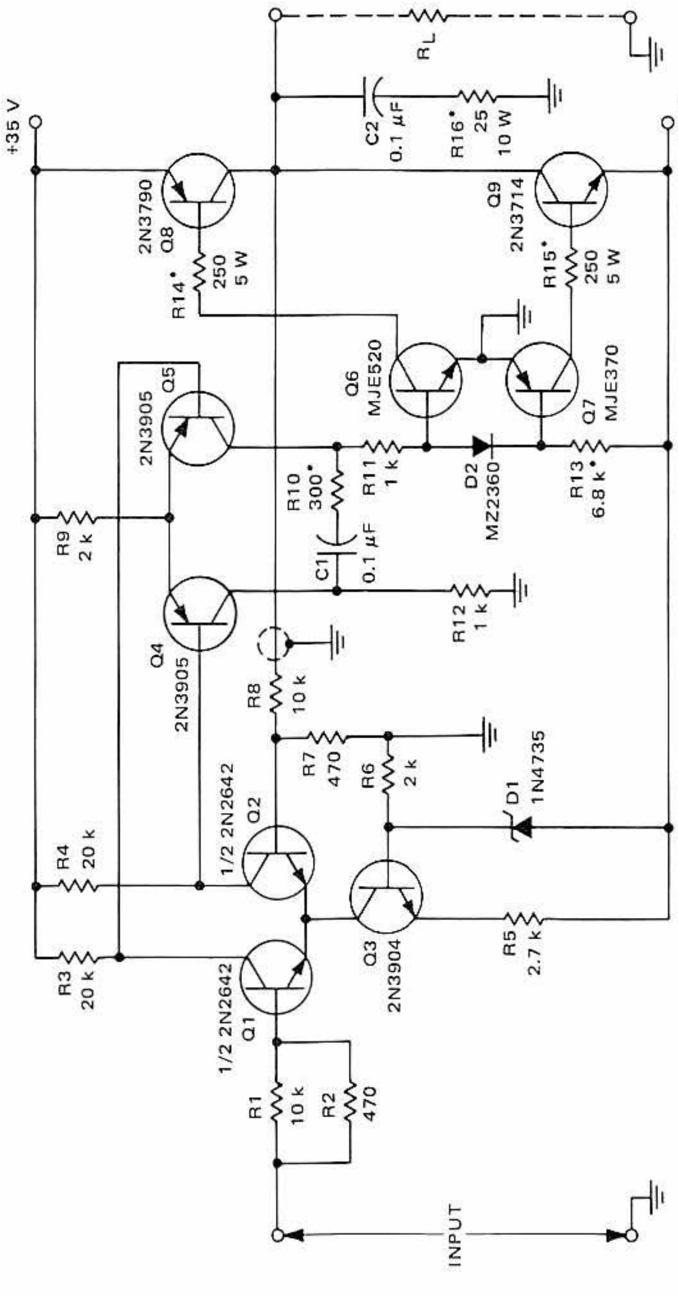


Figure 5-14 — 28 V 28 W DC Servo Amplifier vided via resistor R8. The combination of R7 and R8 sets the voltage gain of amplifier. This is approximately given by

$$A_V \approx \frac{R7 + R8}{R7} \approx \frac{R8}{R7} \; .$$

As a precaution against oscillation, resistor R8 should be mounted as close to Q2 as possible and the feedback from the output should be a shielded lead with the shield grounded at the end by R8 only.

Matching the two input transistors for current gain and base-emitter voltage drop, and matching the current-gain products of the Q6-Q8 and the Q7-Q9 pairs will help reduce the offset voltage and improve the dc stability of the output.

The gain of the amplifier is approximately 22 as shown by the gain characteristics plotted in Figure 5-15. The temperature characteristic of the offset voltage is shown in Figure 5-16. The change in offset is 26 mV for a temperature change from $-40°C$ to $+75°C$. The operating temperature range of the circuit is $-40°C$ to $+75°C$ with Q8 and Q9 mounted on a heat sink yielding a junction to ambient thermal resistance of $5°C$/Watt.

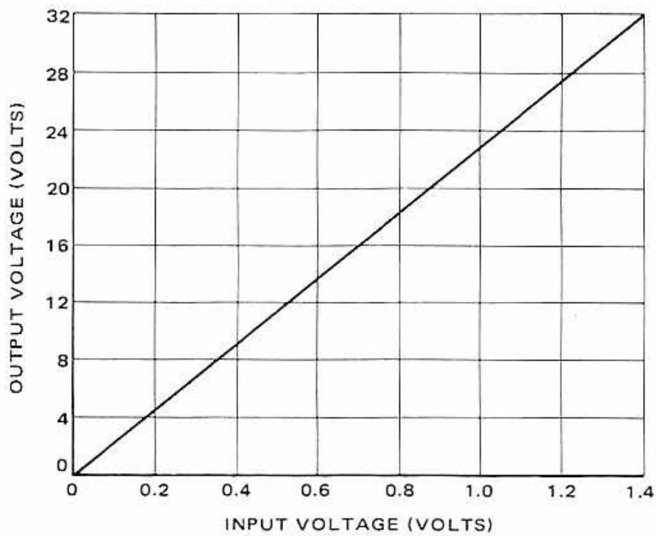


Figure 5-15 — DC Servo Amplifier Gain Characteristics

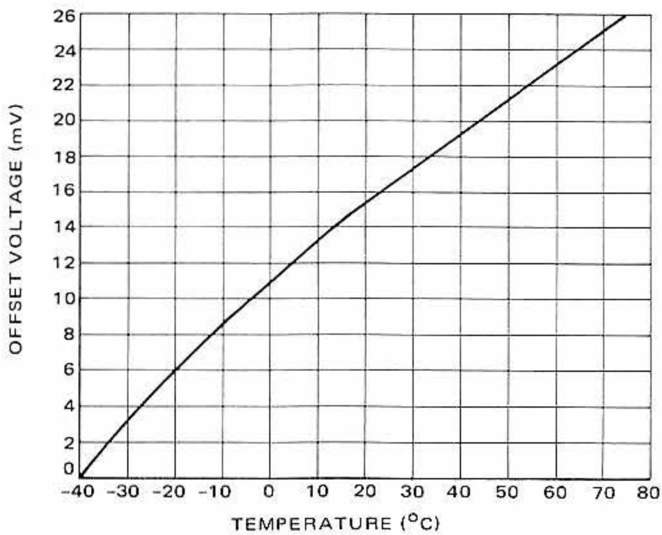


Figure 5-16 – DC Servo Amplifier Offset Voltage versus Temperature

5.6 Pulse-Width-Modulated DC Servo Amplifier

Figure 5-17 is a schematic of a pulse-width-modulated direct-coupled servo amplifier. The heart of the operation of this circuit is the input stage formed by Q1 and Q2. These two transistors and their associated resistors form a Schmitt trigger whose input voltage is that across capacitor C1. This capacitor is used with R1, R2 and R3 to integrate the voltages applied to these resistors. A reference voltage is applied to R2, the input voltage is applied to R1 and, as will be explained, the output voltage from the stage following the Schmitt is applied to R3. R2 is used to set the duty cycle of the output waveform at 50% when there is no input.

The output of the Schmitt drives Q3, which is used as a signal splitter to drive three stages. First of all, a feedback voltage is delivered to the input via D6, R7 and R3. The positive and negative excursions of the voltage are limited to the sum of D1, D2, D3, and D4. The net result of the integrator and the Schmitt trigger is a pulse-width-modulated signal at the output of Q3. The two other outputs of Q3 are the sources for the output driver stages. The output drivers and the output stages were split so the load voltage could easily be driven from the positive supply voltage. The signal splitting is accomplished by isolating diode D7. When Q3 is turned on, D7 is forward-biased and the voltage at the junction of R20 and R21 is

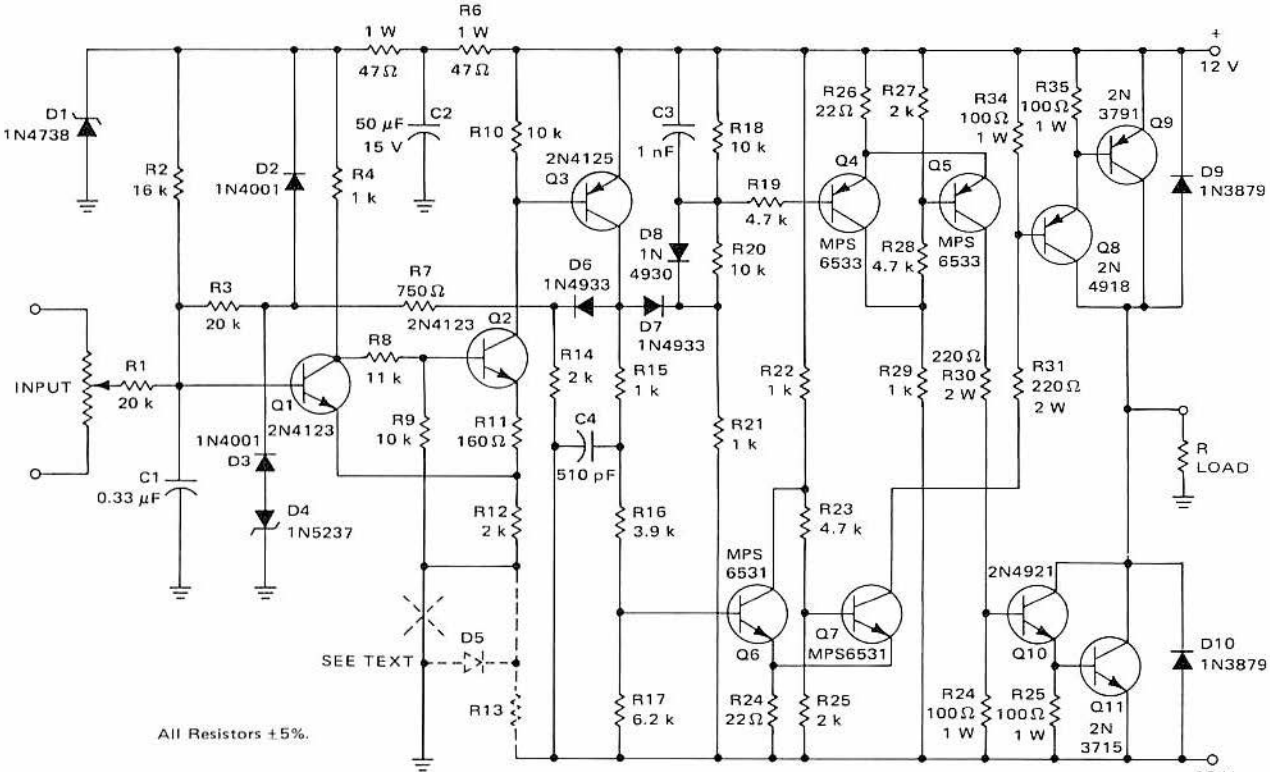


Figure 5-17 — Pulse-Width-Modulated DC Servo Amplifier pulled to the diode drop plus the saturation voltage of Q3, below the positive supply. This removes base drive from Q4 and turns it off. At the same time base drive is applied to Q6 and turns it on. When Q3 is turned off, drive is removed from Q6, turning it off, and diode D7 is reverse-biased, which allows Q4 to turn on. Thus the drivers for the negative output, Q4 and Q5, are on when the drivers for the positive output, Q6 and Q7, are off, and vice versa. Capacitors C3 and C4 are used to provide a slight delay in the turn on of Q4 and Q6, respectively, so that Q8 and Q9 turn off before Q10 and Q11 turn on, and Q10 and Q11 turn off before Q8 and Q9 turn on. This assures that the output transistors will not be on at the same time, thus preventing excessive current and probable damage to them. When Q4 turns on, base drive is removed from Q5, turning it off. This turns the negative output stage (Q10 and Q11) off. Likewise, when Q6 is turned on, Q7 is turned off, which turns off the positive output stage, Q8 and Q9. Diodes D9 and D10 across the two output transistors are used to keep these devices in their safe operating area by limiting voltage transients; therefore they must be fast-recovery rectifiers. This amplifier is capable of driving a 5 ampere load at temperatures as high as 75°C if the output devices are mounted on a heat sink with a thermal resistance less than 10°C/watt. The operating frequency is approximately 3 kHz.

One final note: If the hysteresis of the Schmitt trigger is reduced, some temperature compensation may be required for a stable operating frequency. This can be provided by adding D5 and R13, and removing the ground at the junction of R9 and R12. The value of R13 should be chosen such that the temperature characteristic of D5 compensates for the changes due to temperature effects in the firing levels of the Schmitt trigger.

5.7 1 Watt Integrated-Circuit Audio Amplifier

The only active device in the 1 watt audio amplifier shown in Figure 5-18 is an integrated-circuit power amplifier. The device is designed to amplify signals as high as 300 kHz and to deliver one watt to a directly or capacitively coupled load. Three distinct voltage gains can be obtained from the device by simply changing several pin connections as shown in Table 5-1. The typical voltage gain ratios available are 10, 18 and 36 as measured with a 16 ohm load and a 16 volt supply. The power bandwidth for an output with less than 5% total harmonic distortion is lowest at a voltage gain of 36 and is typically 210 kHz.

When this device is used, care must be exercised to lay out the circuit properly. Due to the large bandwidth of the amplifier, coupling

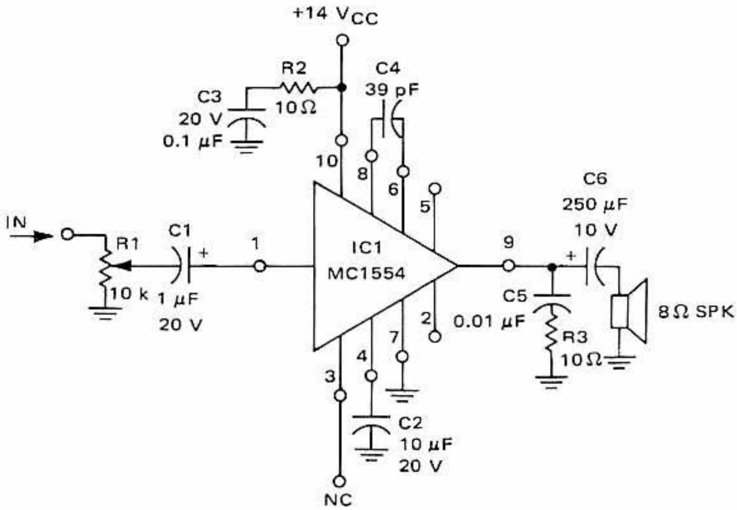


Figure 5-18 — 1 W Integrated Circuit Audio Amplifier

TABLE 5-1

VOLTAGE GAIN	PIN CONNECTIONS
V/V	
10	PIN 2 AND 4 OPEN PIN 5 A-C GROUND
18	PIN 2 AND 5 OPEN PIN 4 AC GROUND
36	PIN 2 CONNECTED TO PIN 5 PIN 4 AC GROUND

must be avoided between the output and input leads. This can be minimized by either (a) the use of short leads which are well isolated, (b) narrow-banding the overall amplifier by placing a capacitor from pin one to ground to form a low-pass filter in combination with the source impedance, or (c) use of a shielded input cable. In applications which require upper band-edge control, the input low-pass filter is recommended. Also to avoid oscillations, an RC stabilizing network (R3 and C5) must be placed from the output (pin nine) to ground, with short leads, to cancel the effects of lead inductance to the load. Inductance of the power supply leads can also cause instability, thereby creating the necessity for R2 and C3. These components must be connected directly from pin ten to ground with leads as short as possible.

The total harmonic distortion from 20 Hz to 20 kHz for 1 watt into a 16 ohm load, with a voltage gain of 10, is typically 0.4%. The frequency response is shown by the curves of Figure 5-19, and the total harmonic distortion by the curves of Figure 5-20. The component values shown give a total harmonic distortion at 1 kHz of less than 1% at 1 watt output power into an 8 ohm speaker. The speaker is capacitively coupled to the

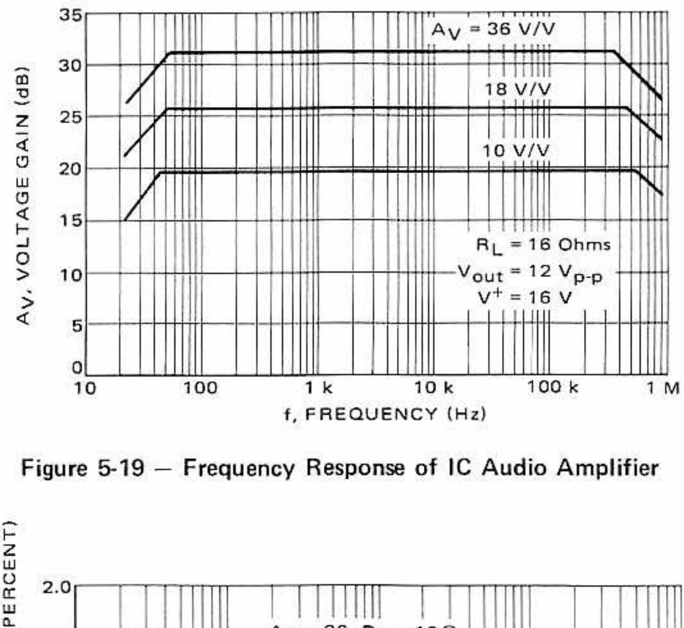


Figure 5-19 – Frequency Response of IC Audio Amplifier

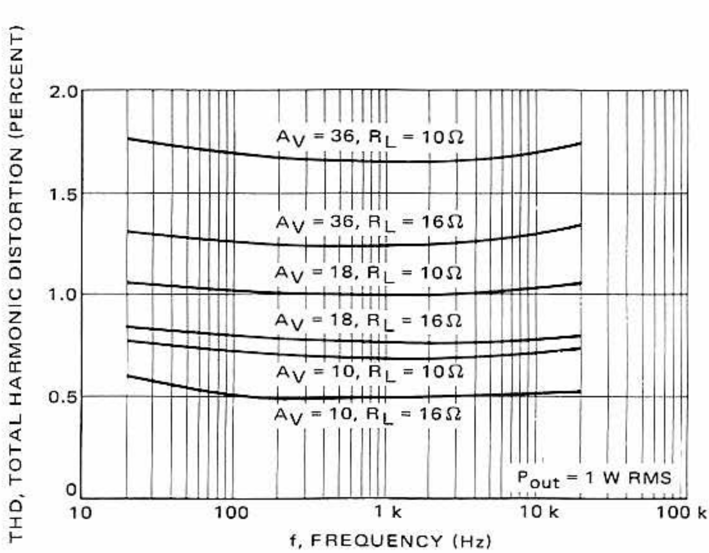


Figure 5-20 – Total Harmonic Distortion of IC Audio Amplifier

amplifier to eliminate quiescent dc in the speaker since a single-voltage power supply is used to drive the amplifier. The low harmonic distortion, low output impedance, excellent gain-temperature stability and selectable voltage gains makes this an excellent low-power audio amplifier.

5.8 4 Watt Wideband Amplifier

The amplifier whose circuit is shown in Figure 5-21 has a 100 kHz bandwidth and a minimum power output of 2 watts. The use of direct coupling and complementary transistors permits wide bandwidth and simplicity in design since there is no need for a phase-inversion stage. This also minimizes power requirements. Two complementary pairs are used, one as the driver and the other as the output stage.

The input stage, which consists of transistors Q1 and Q2, is a differential amplifier. The collector of Q2 drives the emitter of Q3, a common-base amplifier. Diode D1 in the base of Q3 overcomes the base-emitter drop of transistor Q4, thereby maintaining a reverse bias on the collector-base junction of Q3. This assures operation of Q3 in the linear region. Without D1, Q3 would operate in the saturation region resulting in very nonlinear amplification.

One of the basic difficulties encountered with class-B amplifiers is the elimination of crossover distortion. Generally the amplifier is biased

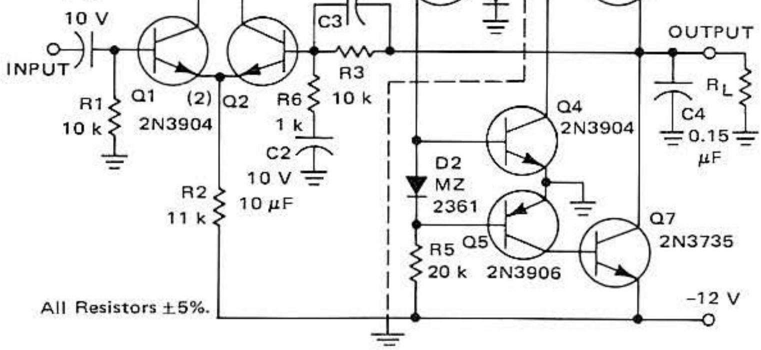


Figure 5-21 — 4 W Wide-Band Amplifier

somewhere between class A and class B (class AB), but this results in reduced efficiency due to the amount of quiescent power required. A more practical way is to drive from a high-impedance, or current, source. This gives low crossover distortion with a minimum of quiescent current. This amplifier offers a compromise between distortion and quiescent current. Dual forward diode D2 provides enough voltage to bias Q4 and Q5 just into conduction. In order to keep the quiescent current of Q6 and Q7 at a reasonable value (3-10 mA) from the standpoint of crossover distortion and efficiency, the voltage drop of D2 at 0.5 mA diode current should be between 1.05 and 1.15 volts. Diode D2 provides an added benefit since its variation over temperature changes will approximately match the base-emitter voltage changes of Q4 and Q5. The source impedance for Q4 and Q5 is effectively the value of R5, or 20 kΩ, since the output impedance of Q3, the common-base amplifier, is several megohms.

Transistors Q4 and Q5 form the complementary driver circuit and they are coupled directly to Q6 and Q7, the complementary output pair. Since the amplifier operates from positive and negative power supplies, and the quiescent output level is approximately 0 volts dc, the load can be directly coupled to the amplifier output.

Negative feedback is carried from the output by feedback resistor R3. This reduces the amount of distortion at the output, including that caused by crossover, resulting in a very clean output waveform. The ac gain of the amplifier is determined by resistor R6 which is decoupled from ground for dc by capacitor C2. The large amount of dc feedback gives a closed-loop dc gain of approximately one and results in exceptional dc stability of the output. This is an important consideration when driving speaker loads since any dc current in the speaker will cause an offset of the speaker cone, and may result in distorted output.

In order to assure a low dc offset at the amplifier output, and to keep distortion low, the beta product of the Q4-Q6 and Q5-Q7 pairs should be matched. The degree of matching required will depend upon the limits set for dc offset and distortion; i.e., the closer the beta-product match, the lower the offset and the distortion.

It should be noted at this point that certain precautions should be taken when building this amplifier. Because of the high gain-bandwidth product of the transistors used, there is a definite tendency for the amplifier to oscillate. To help minimize this tendency, all leads should be kept as short as possible. Also, an attempt should be made to isolate the input stages from the output with a shield. Capacitors C3 and C4 will also generally be necessary to provide sufficient ac stability to prevent oscillation.

The response of the amplifier is indicated in Figure 5-22. This shows that the amplifier is capable of delivering 4 watts of output within +0, −1 dB from 45 Hz to 100 kHz. Harmonic distortion from 35 Hz to 20 kHz (the limit of the test equipment used) at an output of 4 watts is less than 1 per cent.

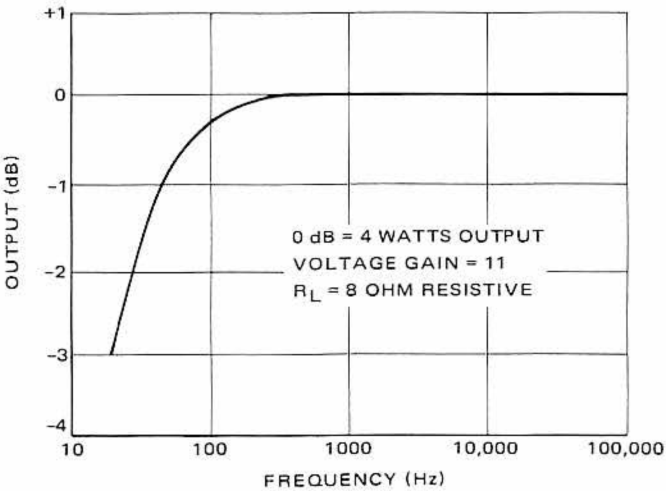


Figure 5-22 – Frequency Response Curve

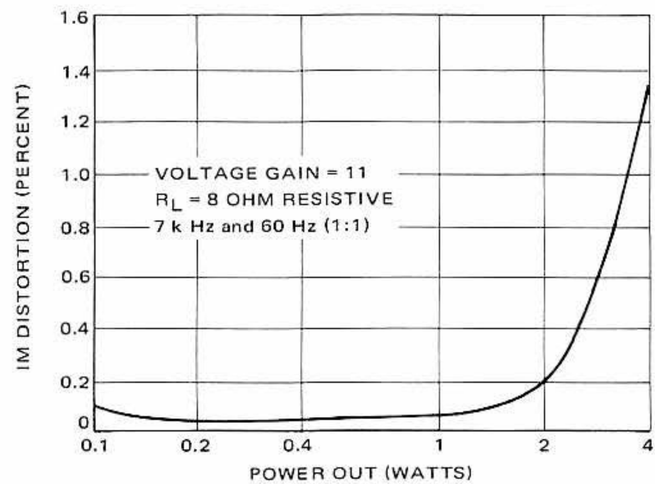


Figure 5-23 – IM Distortion versus Power Output

Intermodulation distortion (IM) for this amplifier is shown in Figure 5-23. From this, it can be seen that the amplifier is more than adequate for hi-fi purposes at a level of 4 watts of output.

The input impedance is approximately 10 kΩ from 35 Hz to 100 kHz. The output impedance is less than 0.25 ohm up to 20 kHz and approximately 0.5 ohm at 100 kHz.

The ambient operating temperature range of the amplifier at an output power of 4 watts, and based upon a junction-to-ambient thermal resistance for Q6 and Q7 of 70°C/watt, is −25°C to +50°C. The junction-to-case thermal resistance of the power output devices, Q6 and Q7, is approximately 5°C/Watt, putting the required case-to-ambient thermal resistance at 20°C/watt. No special heat sinking is necessary for transistors Q1 through Q5.

One final aspect of the amplifier should be mentioned. By removing capacitors C1 and C2, direct coupling to the input, and connecting the bottom of resistor R6 to ground, the amplifier can be used as a dc amplifier with excellent stability.

5.9 5 and 10 Watt Audio Power Amplifiers for 16 Ω Load

The circuit shown in Figure 5-24 is a simple, economical amplifier. It will drive a 16 ohm speaker to either 5 watts or 10 watts of music-power output depending on the parts used. The circuit operation has been separated into three parts for ease of explanation.

Q1 serves as a small-signal preamplifier whose function is to provide modest voltage gain and a fairly high input impedance. To accomplish the latter, the feedback bias network is boot strapped, and a transistor with fairly high h_{FE} is used for Q1. Some input impedance may be traded for higher sensitivity by adjusting the emitter resistance used in this stage. The collector load impedance is shunted with a small capacitor to provide a high-frequency roll off (which may be altered as desired) to prevent parasitic oscillation which may develop between the preamplifier stage and the basic power amplifier because of layout or other factors.

Q2 and Q3 work in cascade as individual common-emitter stages, although they may seem to be a complementary Darlington amplifier at first glance. Q2 actually operates as a common-emitter amplifier for the input signal, and as a common-base stage for the feedback signal. Its basic function is to transfer the dc bias provided by the voltage divider of R6, R7 and R8 to the feedback network of R9 and R10 so as to establish the quiescent voltage at the emitter of Q4 at about one-half the supply voltage. Since the ratio of R9 and R10 determines the closed-loop dc gain of

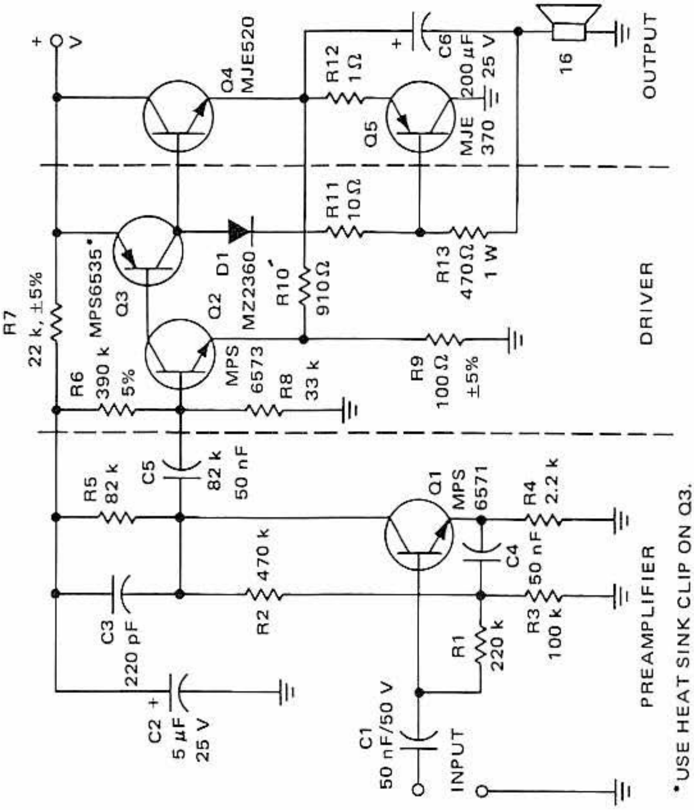


Figure 5-24 – 5 and 10 W Audio Power Amplifiers. Values Shown are for 5 W. For 10 W, Change Components to Those Indicated Below

C2	2 μF/30 V	Q5	MJE371
C6	200 μF/40 V	R7	51 k, ±5%
Q2	MPS6575	R10	1 k, ±5%
Q3	MPS6533	R13	680 Ω, 1 W
Q4	MJE521		

the amplifier, the selection and adjustment of parts values for R6, R7 and R8 is straightforward.

Q2 acts as the common-emitter driver and, in practice, provides the total open-loop voltage gain of the amplifier since the actual voltage gain of Q1 is near one, and the output devices are emitter followers (also with gains near unity). The dissipation of Q2 is inversely proportional to the rated h_{FE} of the output transistors at the peak load current seen. Thus a compromise is necessary between dissipation allowances for Q2 and the

h_{FE} specifications on Q4 and Q5. Bias for the complementary-symmetry output transistors is provided through D1 and its series resistor, R11. Because of the base-to-emitter voltage characteristics of the larger output devices, and the dc bias current passing through Q3 and its collector load network, it is not possible to use two silicon diodes for bias although this would be better. Therefore a diode-resistor combination is used, and is selected for threshold bias for the output devices. The dc collector load resistor for Q3 is bootstrapped by the output coupling capacitor so as to provide full ac current drive to the base of the PNP output transistor in peak signal excursions. This arrangement results in a small direct-current component through the loudspeaker voice coil, but its magnitude (about 2 or 3% of peak load current) is low enough that little or no displacement of the voice coil is experienced.

Q4 and Q5 are the complementary-symmetry emitter-follower output amplifiers. The feature of this complementary connection is that the drive signal is inherently split by the NPN-PNP combination to provide a push-pull output. The emitter-follower operation is obviously the easiest connection to use, and it is also the transistor configuration least affected by normal device parameter variations. The emitter-degenerative resistor required for thermal stability may be put in the emitter leg of the PNP transistor alone rather than being split between the output transistors since the effective dc path is identical, and the expected imbalance in ac performance is hardly measurable. The output coupling capacitor is the primary limitation on low-frequency response and may be changed as desired. The voltage rating for this component should be between half and full supply voltage since the normal dc bias (half supply) will see slow fluctuations of several volts due to the charge and discharge of stored energy when full output is being delivered at frequencies below about 100 Hz.

No specific load-fault protection has been incorporated in the circuit since the intended application is a sealed system. A shorted output will induce destruction of the circuit shown because the loss of feedback signal results in severe overdriving of driver and output stages, and an attempt to drive many times the normal peak load current into the short. The peak power will destroy both the output and driver stages. Fault protection can be incorporated at some increased expense by using a current-limiting scheme similar to that discussed in the 20, 35, and 60 watt amplifiers described in sections 5.11 and 5.12.

The operating characteristics of this amplifier are shown in Figures 5-25, 5-26, and 5-27. The input required for full rated output is 0.1 Vrms into 250 kΩ. The 16 ohm speaker was driven to 5 and 10 watts of output power for the circuit components given in the parts lists.

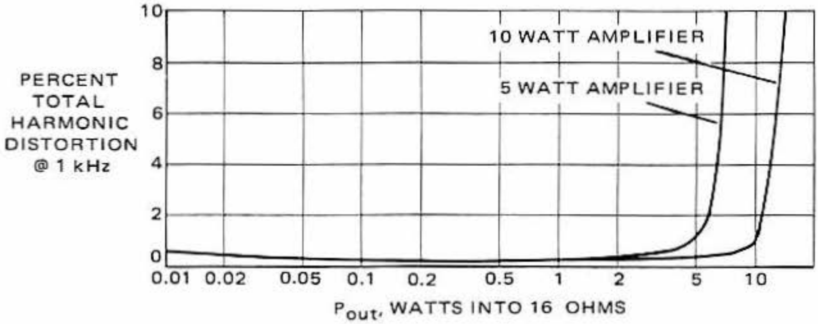


Figure 5-25 – Harmonic Distortion versus Power Output

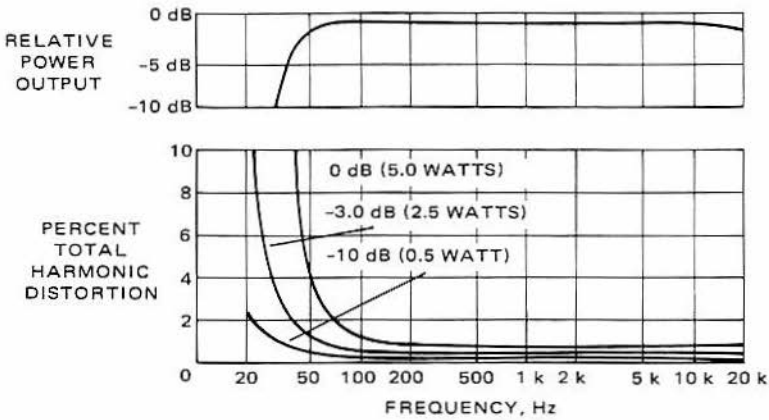


Figure 5-26 – 5 W-Amplifier Bandwidth and Distortion Characteristics

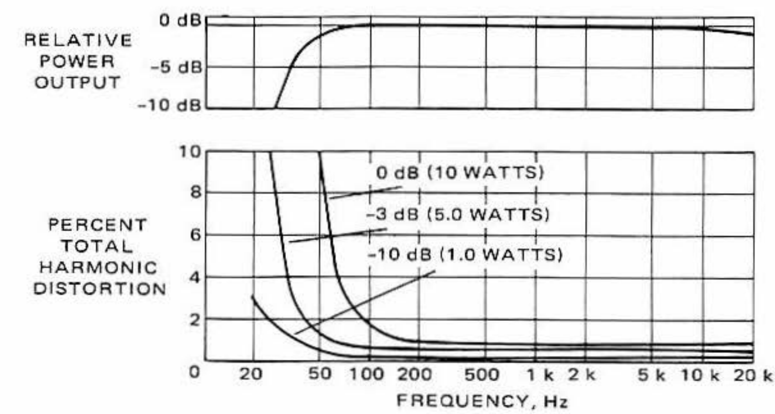


Figure 5-27 – 10 W Amplifier Bandwidth and Distortion Characteristics

5.10 High-Performance 10 Watt Audio Amplifier for 8Ω Load

The direct-coupled, class-B, complementary audio amplifier shown in Figure 5-28 provides 10 watts of music power output into an 8 ohm load with less than 1% distortion from 20 Hz to 20 kHz.

The input stage is a differential amplifier consisting of transistors Q1 and Q2. Biasing for this stage is achieved by resistor R2 (10 kΩ) and the feedback resistor R5 (10 kΩ). The collector of Q2 drives the emitter of Q3, which is connected as a common-base amplifier. The common-base configuration is used to provide high-source-impedance drive for the driver stage, which consists of Q5 and Q6. This high-impedance drive results in lower crossover distortion than low-impedance drive could provide. A common-base stage (Q4) is also connected to the base of Q6 to maintain the properties of the high-source-impedance drive. The base-emitter drop of Q5 sets the collector voltage of Q3 at approximately 0.6 volts. If the base of Q3 were grounded, the collector-base junction would be forward

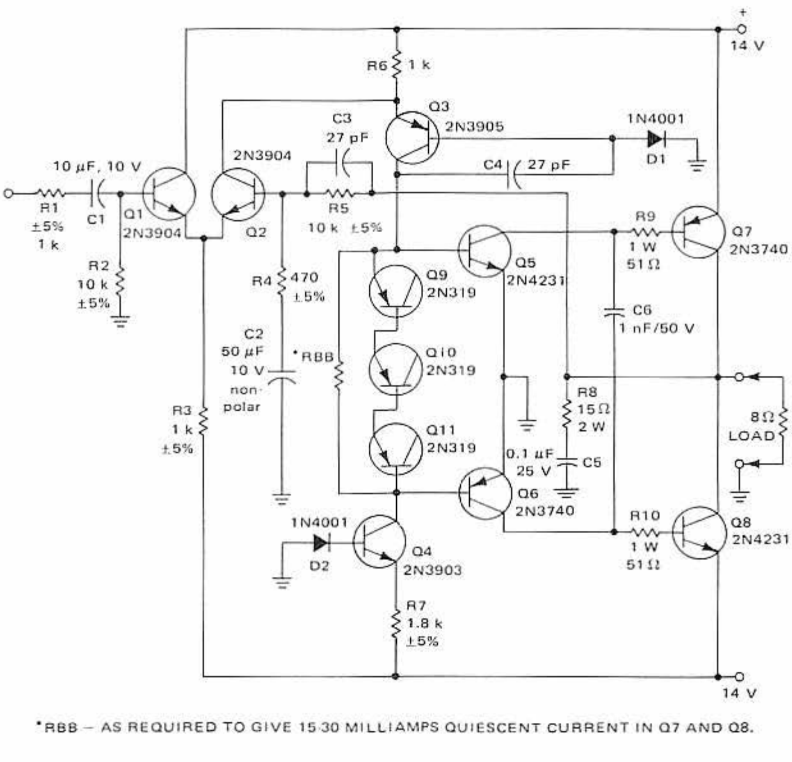


Figure 5-28 — High-Performance 10-Watt Amplifier biased, resulting in very nonlinear amplification. Diode D1 offsets the base-emitter drop of Q5. Diode D2 in the base circuit of Q4 accomplishes the same purpose.

The emitter-base junctions of the three germanium transistors (Q9, Q10, and Q11) connected between the bases of Q5 and Q6 provide just enough forward bias to establish about 15 mA of quiescent current in the output transistors, Q7 and Q8. Resistors can be added in the emitters of Q5 and Q6, but there is a significant loss in amplifier gain before they have any appreciable effect on thermal stability. Better results can be obtained by using the transistor base-emitter junctions as shown in Figure 5-28. These three junctions track the base-emitter junctions of Q5 and Q6 very closely over a wide temperature range. Resistor R_{BB} is used to establish the quiescent currents in Q7 and Q8 between 15 and 30 mA so that crossover distortion is minimized and the upper operating temperature is not limited. Generally 500 to 1000 ohms will be adequate.

The complementary driver transistors, Q5 and Q6, are direct coupled to the complementary output pair, Q7 and Q8. Since the amplifier uses positive and negative power supplies, the quiescent output voltage is approximately 0 volts. This allows the output to be coupled directly to the load.

Negative feedback for the amplifier is provided by R5. Approximately 35 dB of feedback is used, resulting in very low distortion. The closed-loop gain of the amplifier is given by

$$\mathrm{Av} \approx \frac{R_4 + R_5}{R_4} = 22.3$$

The gain can be varied by adjusting R_4.

Although the closed-loop ac gain of the amplifier is about 22, the closed-loop dc gain is approximately unity. The large amount of dc feedback results in exceptionally good dc stability of the output voltage.

Resistors R9 and R10 are used as current limiters for Q5 and Q6. Under normal operation, their effect is negligible. Capacitor C3 and the network consisting of C5 and R8 provide ac stability for the amplifier. It may also be necessary to add capacitor C4 from collector to base of Q3 to prevent spurious oscillations in the 50 to 70 MHz region. Obviously, any oscillation at this frequency could not be heard. In fact, it would encounter much difficulty in getting through the driver and output transistors. It can, however, contribute to distortion in the audible frequency region, and therefore is undesirable. Capacitor C6 was added to improve the transient

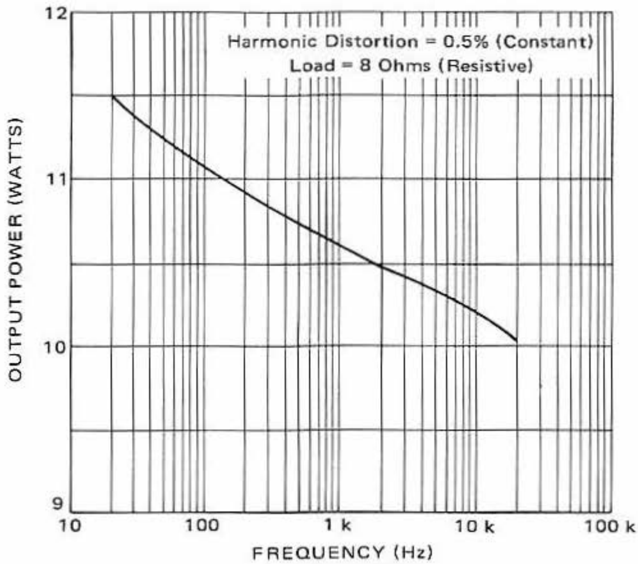


Figure 5-29 – Output Power versus Frequency

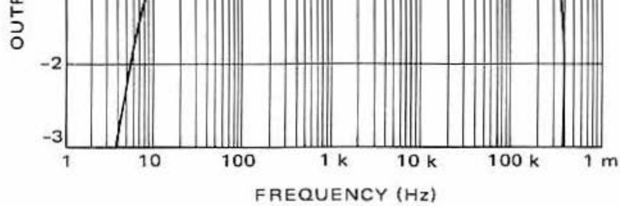


Figure 5-30 – Frequency Response

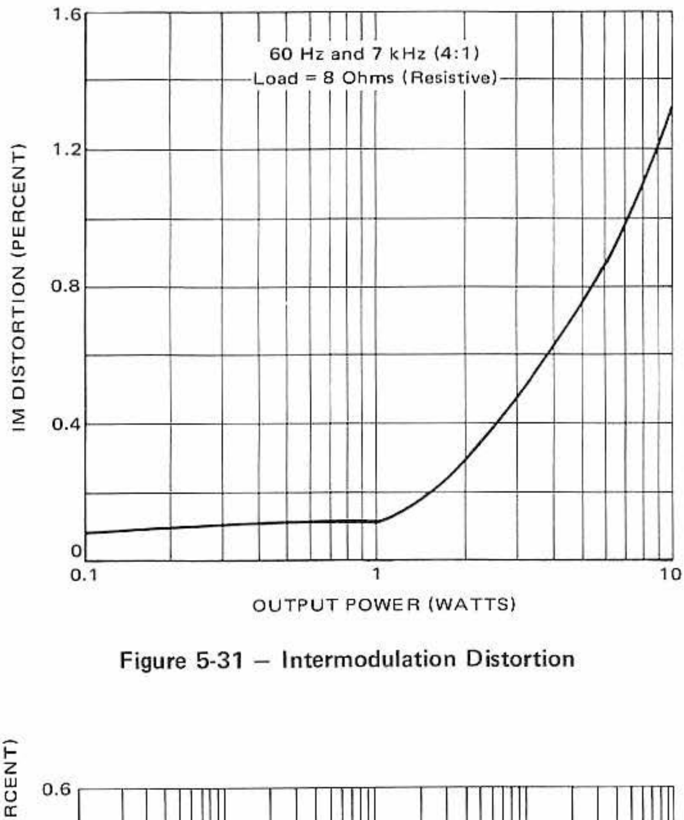


Figure 5-31 — Intermodulation Distortion

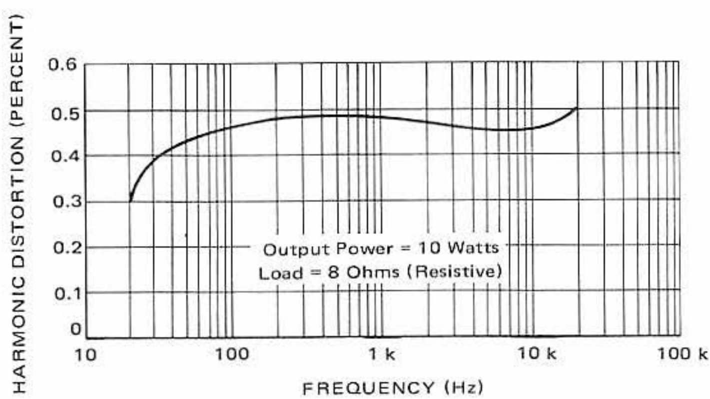


Figure 5-32 — Harmonic Distortion

response characteristics of the amplifier. Resistor R1 was also added for this purpose.

No special heat sinking is necessary for transistors Q1 through Q6. For 100°C operation, the maximum case-to-ambient ($\theta_{C\text{-}A}$) thermal resistance of Q7 and Q8 must be 10°C/watt per transistor.

As can be seen from Figure 5-29, the amplifier provides more than 10 watts at 0.5% total harmonic distortion (THD) from 20 Hz to 20 kHz. The frequency response at the 1 watt level (shown in Figure 5-30) is +0, -0.3 dB from 20 Hz to 20 kHz. It was down 1 dB at 9 Hz and 330 kHz. The -3 dB points were 4 Hz and 400 kHz. Phase shift was 16 degrees at 20 Hz and 4 degrees at 20 kHz.

Intermodulation distortion (IM) using 60 Hz and 7 kHz mixed 4:1 is shown in Figure 5-31. It is less than 0.5% at all levels up to 3 watts and increases to only 1.3% at 10 watts of output. Figures 5-32 and 5-33 show the harmonic distortion. Figure 5-32 is a plot of harmonic distortion versus frequency at a power output of 10 watts. It is less than 0.5% from 20 Hz to 20 kHz. Harmonic distortion versus output power at 1 kHz is shown in Figure 5-33.

Input impedance of the amplifier is approximately 10 kΩ, and is almost constant from 20 Hz to 20 kHz. The output impedance is 0.5 Ω at 1 kHz, 0.5 Ω at 20 Hz, and 0.25 Ω at 20 kHz.

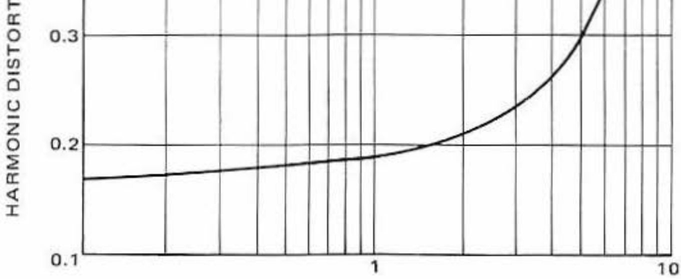
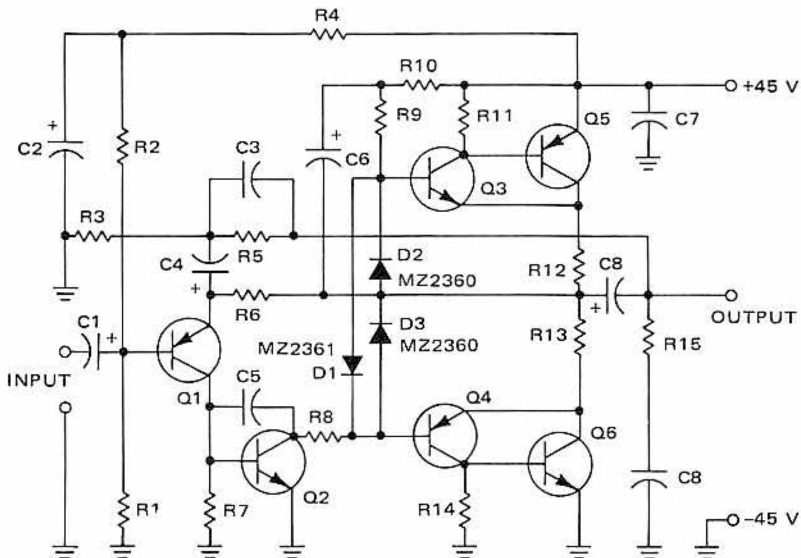


Figure 5-33 — Harmonic Distortion versus Output Power

5.11 20 Watt Audio Power Amplifier

The amplifier shown in Figure 5-34 will deliver 20 watts of music power to an 8 ohm speaker with a total harmonic distortion at rated output of less than 0.5% at 1 kHz. It has a typical 3 dB bandwidth of 40 Hz to 100 kHz. The input required for rated output is 0.3 volts rms into 100 kΩ.

The input transistor Q1 serves as a dc emitter follower through R6 to the output, and this makes the entire amplifier a dc emitter follower as



PARTS LIST:

C1	–	1 μF, 15 V	
C2	–	1 μF, 50 V	
C3	–	50 pF, 50 V	
C4	–	5 μF, 15 V	
C5	–	50 pF, 50 V	
C6	–	50 μF, 15 V	
C7, C8	–	0.1 μF, 50 V	
Q1	–	MPS6517	
Q2	–	MPS6575	
Q3	–	MPS6530	USE CLIP-ON
Q4	–	MPS6533	HEAT SINK
Q5	–	MJ-4931	
Q6	–	MJ-4831	

R1	–	150 k, 1/2 w, 5%
R2	–	560 k, 1/2 w, 5%
R3	–	100 Ω, 1/2 w, 10%
R4	–	560 k, 1/2 w, 5%
R5	–	5.6 k, 1/2 w, 10%
R6	–	33 k, 1/2 w, 10%
R7	–	1.2 k, 1/2 w, 10%
R8	–	100 Ω, 1/2 w, 10%
R9	–	2.2 k, 1/2 w, 10%
R10	–	2.2 k, 1/2 w, 10%
R11	–	220 Ω, 1/2 w, 10%
R12	–	0.43 Ω, 1/2 w, 5%
R13	–	0.47 Ω, 1/2 w, 5%
R14	–	220 Ω, 1/2 w, 10%
R15	–	10 Ω, 1/2 w, 10%

Figure 5-34 – 20 W Audio Amplifier

well. Thus the bias voltages inside the amplifier are very stable; the input bias circuit of R1, R2, and R4 form a voltage divider which sets the desired voltage at the output driving point at 1/2 the supply voltage, to insure symmetrical signal clipping. Because of bypass capacitor C4 in the emitter of Q1, however, the amplifier has very high ac gain. Ac feedback from the output is applied to emitter resistor R3 so that the closed-loop ac gain of the amplifier is determined by the ratio of resistors R3 and R5. In the suggested circuit design given, the closed-loop gain is 50. Some ac feedback is applied through R6 so that the parallel resistance of R6 and R5 form the total effective feedback resistor, although the contribution of R6 is small.

Q2 functions as a large-signal class-A driver. Since the output circuit is effectively an emitter follower, Q2 must swing the entire load voltage. As it operates with a voltage gain of some 60 dB or better, it also provides a convenient point for setting the dominant high-frequency pole in the amplifier's open-loop Bode response. A small capacitor from collector to base, reflected back to the base driving point as a high Miller capacitance, is sufficient to locate the pole conveniently far away from the basic turnover frequencies of the input and output circuits. Thus the amplifier has a dominant −6 dB/octave rolloff characteristic above about 50 kHz.

The collector load of Q2 consists primarily of the output load (the 8 ohm speaker) multiplied by the product of the h_{FE}'s of the transistors in the complementary Darlington circuit. Because of the bootstrapping by capacitor C6, the effective impedance presented to Q2 by R9 and R10 is much higher than the reflected speaker load. The main purpose of the bootstrapped output is to provide the drive to the upper Darlington amplifier (Q3 and Q5) on peak signal excursions.

Q3 and Q4 appear as emitter followers between the load and the driver. Q5 and Q6, the complements of Q3 and Q4, operate effectively with their drivers as Darlington pairs. One may analyze the output circuit as an amplifier with two cascaded common-emitter transistors (the NPN into the PNP, and vice versa) with total feedback for unity overall gain. Thus a basically linear, local amplifier is created: the output voltage equals the input voltage. The point to this complementary Darlington connection is to have the base-emitter junctions, which are biased through D1, control the response of the output circuit to the exclusion of the power transistor characteristics. Degenerative resistors R12 and R13 are located in the effective emitter of the complementary Darlington amplifier, and provide feedback for bias control purposes; this results in excellent thermal stability. Since the drivers are biased from a dual forward voltage reference diode, the chosen idling bias for the output circuit is fully compensated for ambient temperature variations. It needs no adjusting potentiometers and is largely insensitive to variations in supply voltage.

Diodes D2 and D3 provide fault protection. D1 and D3 effectively appear in series between the base of Q3 and the output on the positive half cycle of output signal, and act to limit shorted-output fault current to a value slightly above the normal peak load current. Similarly, D2 appears in series with D1 for the negative half cycle of output signal.

The low-frequency response is limited by the output coupling capacitor. A larger value will extend the low-frequency limit. Any high-frequency roll off should be set at the input since any attempt to place the roll off inside the feedback loop will increase the distortion and decrease the ac-stability margin of the amplifier.

Figures 5-35 and 5-36 show the harmonic and intermodulation-distortion characteristics of the amplifier.

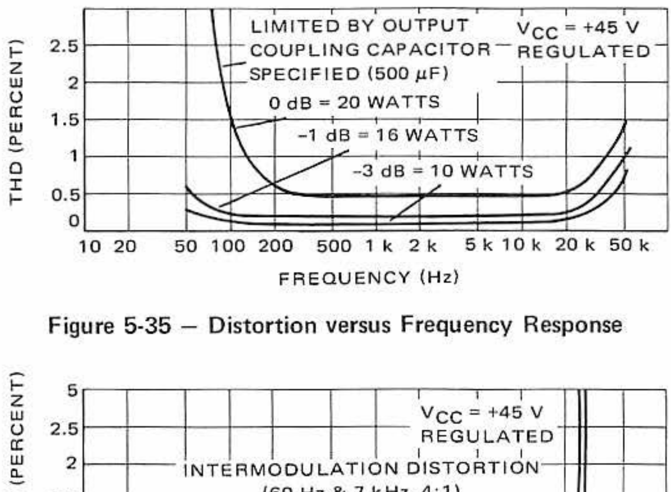


Figure 5-35 – Distortion versus Frequency Response

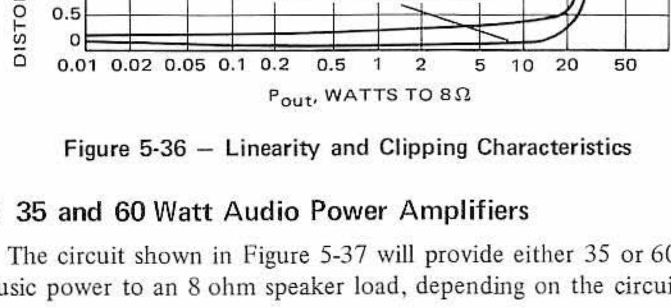


Figure 5-36 – Linearity and Clipping Characteristics

5.12 35 and 60 Watt Audio Power Amplifiers

The circuit shown in Figure 5-37 will provide either 35 or 60 watts of music power to an 8 ohm speaker load, depending on the circuit com-

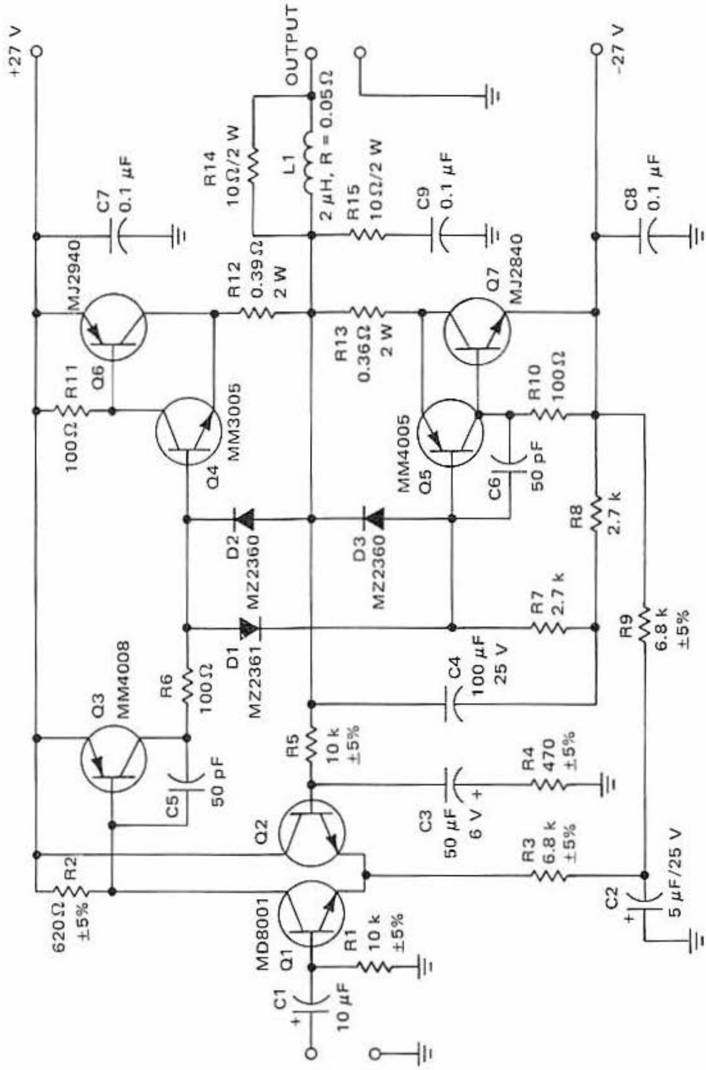


Figure 5-37 — 35 and 60 Watt Audio Power Amplifier. Values Above are for 35-W Amplifier. For 60 W, Change Components Listed Below.

D2, D3	—	MR2361	R3, R9	—	7.5 k ±5%
Q1, Q2	—	MD8002	R7, R8	—	3.9 k
Q3	—	MM4009	R12	—	0.47Ω, 2 W
Q4	—	MM3006	R13	—	0.43Ω, 2 W
Q5	—	MM4006	+V	—	+36 V
Q6	—	MJ2941	−V	—	−36 V
Q7	—	MJ2841			

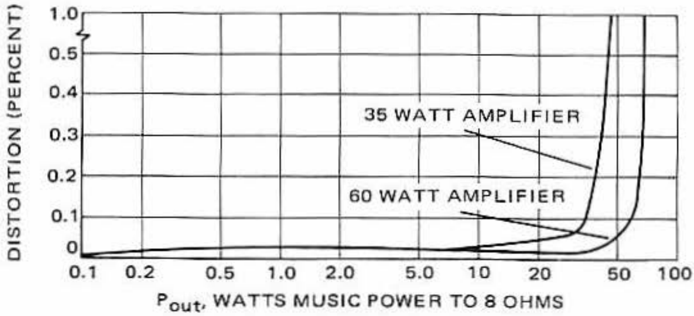


Figure 5-38 – Intermodulation Distortion (60 Hz & 7 kHz Mixed 4:1) versus Power Output

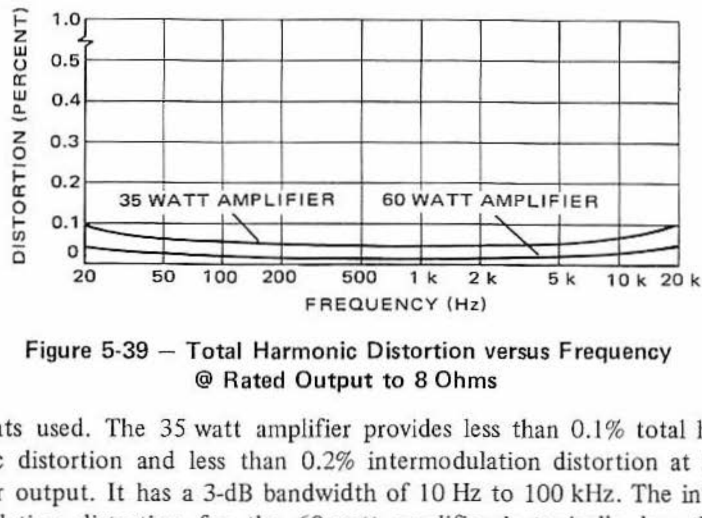


Figure 5-39 – Total Harmonic Distortion versus Frequency @ Rated Output to 8 Ohms

ponents used. The 35 watt amplifier provides less than 0.1% total harmonic distortion and less than 0.2% intermodulation distortion at full power output. It has a 3-dB bandwidth of 10 Hz to 100 kHz. The intermodulation distortion for the 60 watt amplifier is typically less than 0.15% with its other performance specifications the same as the 35 watt amplifier. Both amplifiers require an input of 1.0 volt rms into 10 kΩ for full power output. Figures 5-38 and 5-39 give the intermodulation and harmonic distortion.

Q1 and Q2 form a differential amplifier which is used because the inherent temperature-compensated balance of both base legs and the controlled symmetry of voltages in those legs provide excellent thermal stability. Q1 is biased through a resistor in its base leg to ground; Q2, biased through a similar resistor, will cause the output to be at equivalent ground because of the symmetry of the two halves of the local circuit. The error voltage appearing as an offset at the output consists of the differential base-emitter voltage between Q1 and Q2 and the differential voltage established by the base currents flowing through the bias resistors (R1 and R5). When Q1 and Q2 are supplied as a dual transistor, typical offsets of less than 10 mV are realized at the output.

Q1 also serves as a common-emitter amplifier to the input signal, and uses the common-base input impedance of Q2 as its effective emitter load. Constant-current emitter bias for Q1 and Q2 is supplied through the common-mode path (R3 and R9), and decoupling is provided in this path for optimum rejection of power-supply ripple.

Q3 functions as a common-emitter driver. Since the output circuit (as will be explained) is only an emitter follower, Q3 must swing the full load voltage plus the losses in saturation and transconductance of the output devices. This stage commonly operates with a voltage gain of 60 dB or better, and thus provides a convenient point for setting the single dominant high-frequency pole in the amplifier's open-loop Bode response. A small capacitor (C5) from collector to base, as multiplied by the Miller effect, appears back at the base driving point to set this pole conveniently far away from the basic turnover frequencies of the input and output circuits, so that the amplifier as a whole demonstrates a basic 6 dB/octave roll off above about 50 kHz. In this way, a generous amount of feedback can be applied to the circuit to optimize linearity as desired while still maintaining a comfortable stability margin.

The dc bias path for Q3 proceeds through a small collector resistor R6 (used to limit the power dissipation in Q3 under load-fault conditions), through biasing diode D1 and dc collector load resistor (R7 and R8). This latter resistor is split and bootstrapped through C4 so that full alternating current drive may be available to the lower half of the output circuit during peak signal excursion. Because of this bootstrapping, the effective ac load seen by Q3 is the reflected loudspeaker load impedance (i.e., load R_L multiplied by the product of the h_{FE}'s of the complementary Darlington output devices). R5, the dc base resistor for Q2, also serves as the loop feedback resistor by providing the feedback signal to the base of Q2. The ratio of R5 to R4 approximately determines the closed-loop voltage gain of the amplifier. Capacitor C3 in series with R4 is chosen to provide the desired low-frequency roll off and force the amplifier to demonstrate unity dc voltage gain for optimum bias stability.

The output circuit is a full complementary-symmetry Darlington emitter follower. The emitter-follower connection is established by Q4 and Q5, and is chosen because it provides the best basic linearity among the three possible transistor configurations, and is the least critical for normal variations in device parameters. The availability of complementary-symmetry silicon power transistors has allowed the practical development of the output circuit given, which is an improvement over the quasi-complementary connection with its many bias problems. The simplicity and advantage of a full complementary symmetry amplifier is seen in the way Q4 and Q5, biased with dual diode D1, completely control the output circuit. Transistors Q4 and Q6, and transistors Q5 and Q7 each form a local unity-gain dc operational amplifier. Complete control of the output is thus given to the drive voltage appearing at the bases of Q4 and Q5. Since bias is established by diode reference, excellent compensation and control through variations in line voltage and operating temperature are assured, and this represents a significant improvement over the bias schemes used with the quasi-complementary circuit.

Diodes D2 and D3, in conjunction with D1, provide load-fault protection. D1 and D3 will effectively appear in series between the base of Q4 and the output on the positive half cycle of output signal, and will act to limit the peak fault current to a chosen value (through R12 and R13) slightly above the normal peak-load current by limiting the voltage that can be developed between Q4's base and the output. Similarly, D2 in conjunction with D1, provides fault protection for the negative half cycle by operating between the output and the base of Q5. The peak fault current is kept to a value within the safe-area capability of the output devices (provided enough heat sinking is available) so that overload interruption may be accomplished by thermal cutouts or slow-blow fuses.

Some additional compensation networks appear at the output. Among these is a standard RC pad across the output, which inserts a stabilizing pole and zero into the amplifier's Bode response, and a choke in series with the load. The latter is a low-resistance rf choke used to provide a buffer between the amplifier and a capacitive load. A capacitor across the output of many amplifiers will upset stability and cause oscillation because it provides an uncompensated pole in the Bode response, usually at just the wrong place. A small inductance (about 2 μH) is used to buffer out this effect.

It is also advisable to provide power-supply bypass capacitors as close to the circuit as possible. These capacitors are used to provide very-high-frequency, return-path shorts to the amplifier. The high load currents flowing through long power-supply leads (consisting of series R and L) generate enough of a ground loop at very high frequencies to cause the amplifier to oscillate. Disc-ceramic bypass capacitors are chosen as needed to eliminate this source of parasitic oscillation.

5.13 50 Watt Audio Power Amplifier

Figure 5-40 is the circuit diagram of a 50 watt (music power) audio

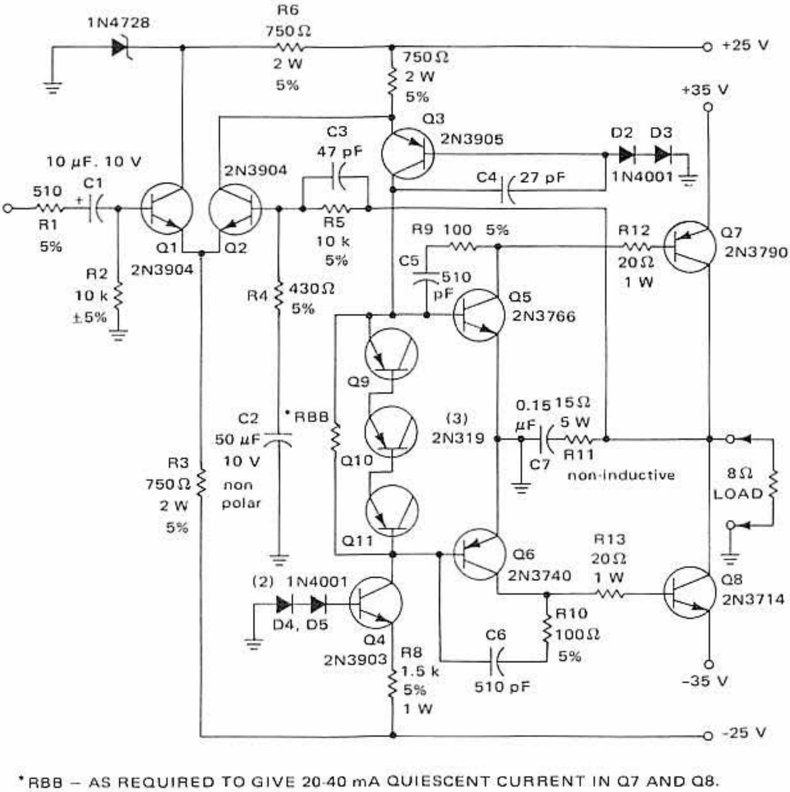


Figure 5-40 – 50 Watt Amplifier

power amplifier. Except for a few minor changes, it is the same circuit used for the 10-watt high performance audio amplifier discussed in section 5.10. A zener diode (D1) has been added in the collector of Q1. This is used to limit the voltage applied to Q1, thus allowing the use of a low-cost plastic-encapsulated transistor. The zener is also a low-cost, 1 watt Surmetic device. Two diodes are used in the base circuits of Q3 and Q4, instead of the one used in the 10 watt amplifier, because of the larger dynamic voltage swings of the base-emitter junctions of Q5 and Q6 as compared to those in the 10 watt amplifier. Also, it was found that a single capacitor from collector to collector of Q5 and Q6 was not adequate for good transient response, so a 510 pF capacitor in series with a 100 ohm resistor was added from each collector to base (C5 and R9, and C6 and R10).

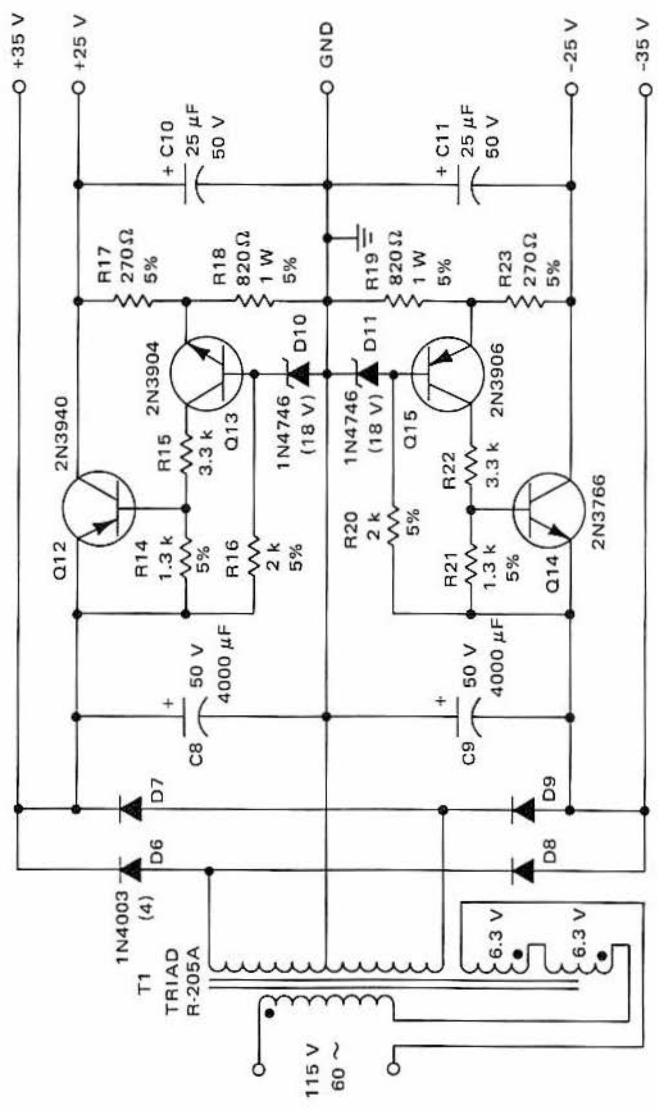


Figure 5-41 — 50 Watt Amplifier Power Supply

No heat sinks are required for transistors Q1 through Q4. The maximum case-to-ambient thermal resistance (θ_{CA}) of transistors Q5 and Q6 is 8°C/watt (per transistor) for 100°C operation. For Q7 and Q8 the maximum case-to-ambient thermal resistance is 2.5°C/watt (per transistor). In addition, for good tracking, transistors Q5 and Q6 should be mounted on one heat sink along with the three germanium transistors Q9, Q10, and Q11.

The power supply used for testing the 50 watt amplifier is shown in Figure 5-41. Filtered, but unregulated, voltage is supplied to the two output transistors. The zener-regulated supplies provide ±25 volts to the input stages. The power transformer used in this supply has a secondary voltage that is too high. This is the reason for using the two 6.3 volt windings as bucking voltages. The filtered voltage that results is ±38.5 volts at 0 watts out of the amplifier and ±34.5 volts at 50 watts output. If a power transformer were to be designed for the amplifier, better performance and lower distortion would be realized if the no-load output voltage were 39 volts, dropping to no less than 36 volts at 50 watts output. Higher voltages, though reducing distortion even further, could lead to breakdown in the output transistors. Also, power dissipation in the output transistors (and the drivers as well) would increase, thereby reducing the maximum operating temperature of the amplifier.

The ideal supply to use would be one that is completely regulated at approximately 36 volts. If ±36 volt regulated supplies are available, the

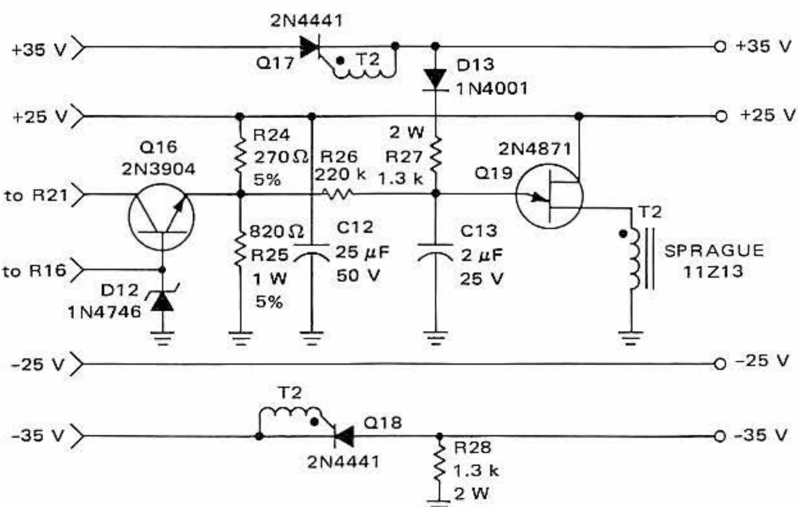


Figure 5-42 – 50 W Amplifier Delayed Switching Circuit

input and output stages can be operated from the same supplies by changing R3 and R7 to 1.3 kΩ, R6 to 1.1 kΩ, and R8 to 2.7 kΩ.

Two problems related to the power supplies arise in this amplifier. The first is that both the positive and negative supplies should come on at the same time. If power is applied to only one side, the driver-output pair associated with that side will conduct very high currents, possibly damaging one or both transistors due to high power dissipation.

The second problem arises primarily because of capacitors C1 and

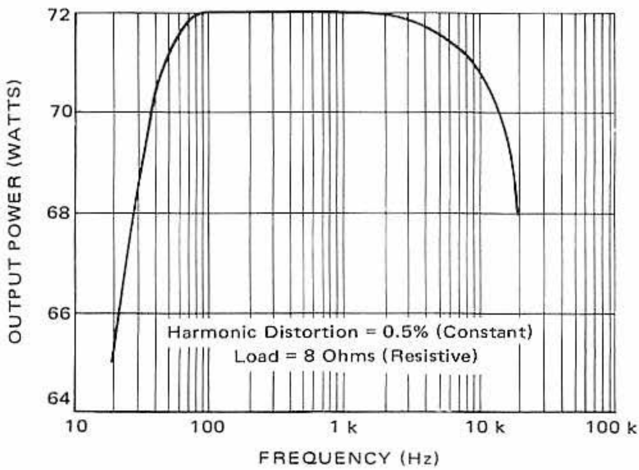


Figure 5-43 – 50-W-Amplifier Output Power versus Frequency

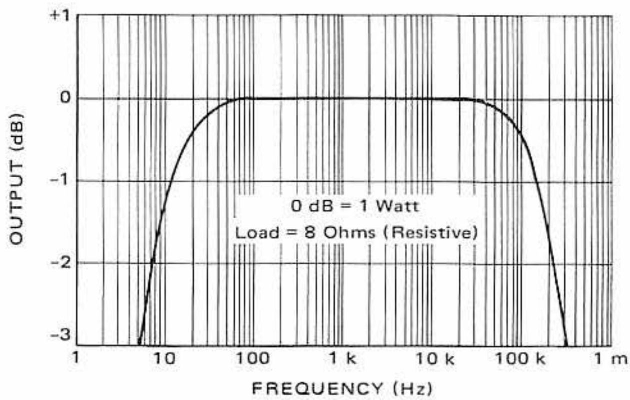


Figure 5-44 – 50-Watt-Amplifier Frequency Response

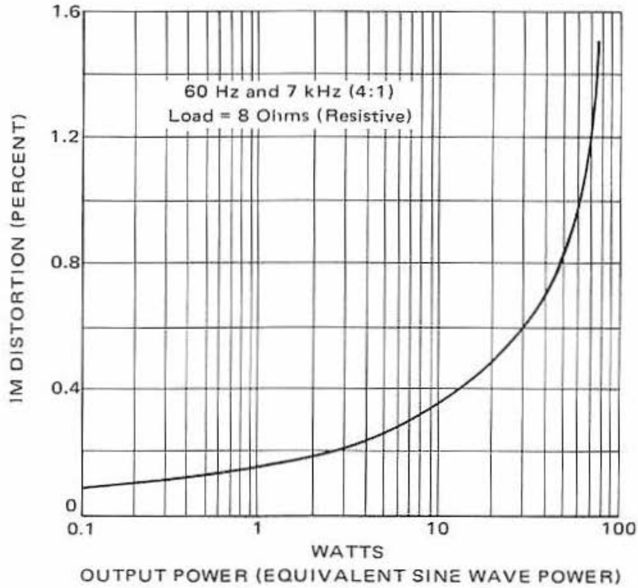


Figure 5-45 – 50-Watt-Amplifier Intermodulation Distortion

C2 in the amplifier. When power is applied, the charging of these capacitors causes some high-magnitude transients in the amplifier output. To eliminate or minimize this, the voltages to the input stages should be applied a few hundred milliseconds before the voltages to the output stage. This allows the capacitors to become fully charged. One way to do this would be with two switches or a two-position rotary switch. It could also be done electronically. One possible circuit is shown in Figure 5-42. The RC timing network (R26 and C13) is connected at the junction of the voltage divider, at the output of the positive regulated supply, thus assuring that the timing sequence won't begin until this supply is up to voltage. When the capacitor C13 has charged to the breakover voltage of the unijunction transistor, Q19, the unijunction conducts and discharges C13 through T2. This results in a pulse being applied, via T2, to the gates of SCRs Q17 and Q18, turning them on. The result is the required delay before voltage is applied to the output stage.

Output power capability of the amplifier at a constant total harmonic distortion of 0.5% is shown in Figure 5-43: the amplifier can provide 65 watts at 20 Hz, 72 watts at 1 kHz, and 68 watts at 20 kHz, at 0.5% THD.

Figure 5-44 is a plot of the frequency response at 1 watt output. It is almost flat from 100 Hz to 20 kHz, being down only 0.35 dB at 20 Hz.

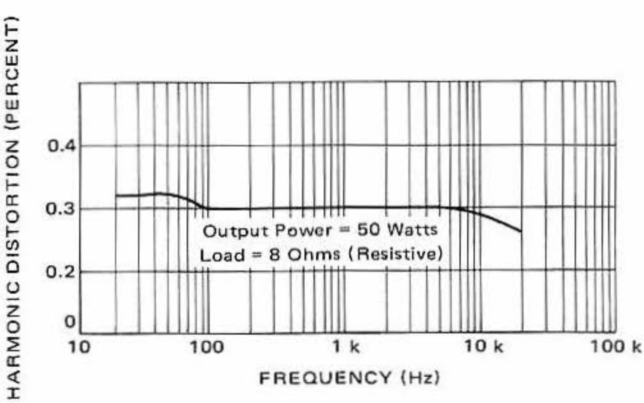


Figure 5-46 – 50-Watt-Amplifier Harmonic Distortion

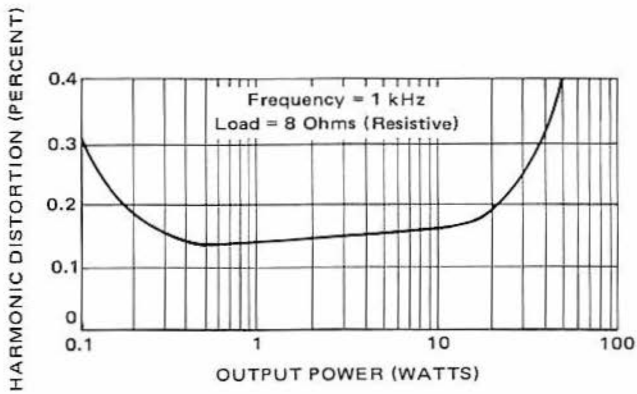


Figure 5-47 – 50-W-Amplifier Harmonic Distortion versus Output Power

The -1 dB points are at 11 Hz and 150 kHz, and -3 dB response is reached at 5 Hz and 320 kHz. Phase shift is 18 degrees at 20 Hz and 6 degrees at 20 kHz.

The results of intermodulation-distortion tests can be seen in Figure 5-45. IM is about 0.1% at 0.1 watt output, rising steadily to 0.5% at 20 watts output, and reaching 0.85% at 50 watts output. The output capability at 1.5% IM is approximately 65 watts.

Harmonic distortion is shown in Figures 5-46 and 5-47, Figure 5-46 being a plot of harmonic distortion versus frequency at 50 watts output and Figure 5-47 a plot of harmonic distortion versus power output at 1 kHz.

Input impedance of the 50 watt amplifier is approximately 10 kΩ from 20 Hz to 20 kHz. Output impedance is less than 0.1 ohm from 20 Hz to 20 kHz.

As a final note it should be mentioned that this amplifier can be used as a dc amplifier with excellent stability. Minor circuit modifications are required to achieve this. The resulting circuit is shown as the 25 volt, 28 watt, dc servo amplifier of Figure 5-14.

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