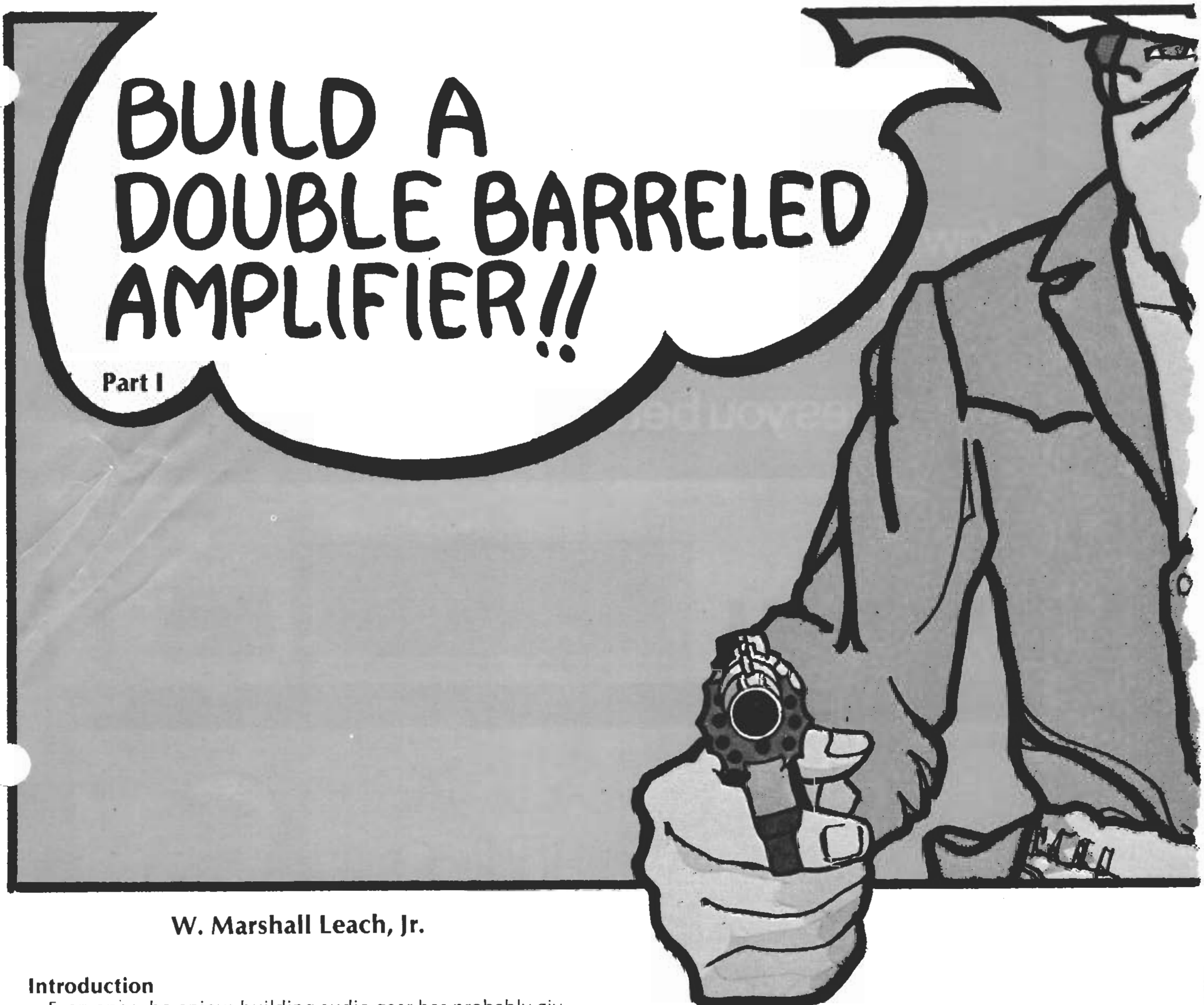

W. Marshall Leach, Jr.

Introduction

Everyone who enjoys building audio gear has probably given thought to constructing a high-power amplifier which will drive even the most inefficient loudspeakers to super loud levels. When it comes to power, most people agree that the higher the power reserve in an amplifier, the better the sound. There is a simple and logical reason for this. Audio signals can have very high peak power levels that run 10 to 20 times the average power level. Thus the peak power output capability of the amplifier becomes an important consideration if it is not to clip on peaks. As an example, let us consider an amplifier rated at 100 watts average power. (This is erroneously called "watts rms." The abbreviation "rms" stands for "root mean square" and is correctly used only in the specification of the amplitude of a.c. voltages or currents, e.g. volts rms or amperes rms, but not watts rms.) Because the power rating is measured with a sine-wave signal which has a peak-to-average power ratio of only 2, it follows that the peak power capability of our 100-watt amplifier is 200 watts. With audio signals having a peak-to-average power ratio of 10 to 20, the amplifier could be operated at an average power level of 10 to 20 watts if it is not to clip on peaks. With a typical loudspeaker efficiency of 0.5 percent, only 50 to 100 milliwatts of undistorted acoustic power could be obtained. This corresponds to a sound pressure level in the reverberant field of a normal listening room of 99 to 102 dB. Should this be insufficient or if the amplifier is to be used in large rooms or outdoors, a higher power rating will be required.

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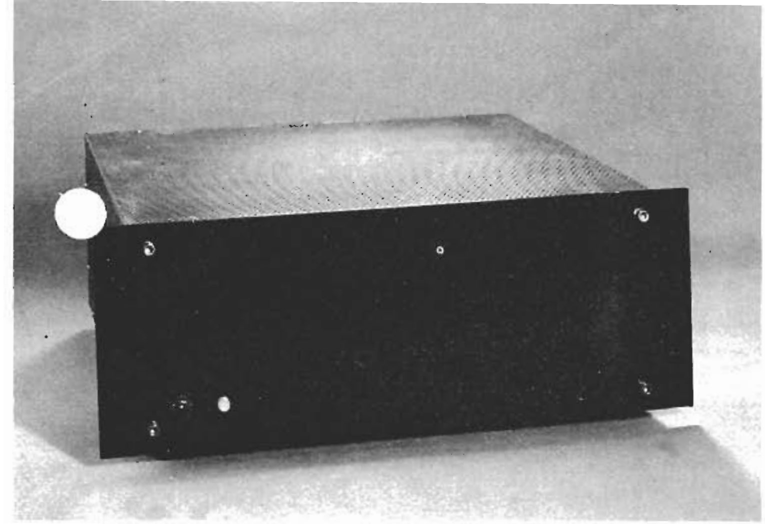
How much power is enough? This is a question which will certainly receive many answers, depending on who is asked. Someone with efficient loudspeakers who listens to quiet chamber music may say 10 watts. The audiophile with the latest inefficient loudspeakers who likes to demonstrate them at ear-splitting levels with the newest direct-to-disc recording may say the sky's the limit. However, most people probably consider a power rating of 250 watts per channel to be an adequate rating for a so-called high-power amplifier. From the standpoint of circuit design, 250 watts represents a power level which is about the maximum that can be achieved without compromises in the design, exorbitant costs, or both. The only method that can be used to increase the power rating of an amplifier is to increase the peak-voltage swing capability of its output signal. This can be done by designing it with a higher voltage power supply, strapping it,

or by adding a transformer between the amplifier and the load which steps up the output voltage. To increase the power supply voltage can require circuit design compromises such as the use of a quasi-complementary output stage rather than a fully complementary one. Strapping and a stepup transformer are also design compromises. Both effectively reduce the load impedance seen by the amplifier which can cause overload or excessive triggering of the amplifier protection circuit. Also, the transformer is incompatible with a direct coupled design, and it can cause intolerable phase shifts and distortion at very low frequencies.

Transistors are inherently low-voltage, high-current devices. Therefore, reliable circuit designs require that the voltage across transistors be held to a value that is low enough to permit them to safely deliver the load currents required for the maximum desired output power. This is especially important when reactive loads such as electrostatic loudspeakers or highly inductive crossover networks must be driven. Thus, there is a basic conflict between the requirement for a reliable design and the desire to increase the power rating by increasing the power-supply voltage. Fortunately, both of these objectives can be achieved by connecting transistor pairs in series, where normally only a single transistor would be used, so that each in the pair shares only one-half the total voltage that a single transistor would be required to drop. With this technique of using series-connected transistors, a power rating of 250 watts can easily be achieved without design compromises.

Dynamic distortions such as transient intermodulation distortion (TIM) and slewing induced distortion (SID) are important considerations in any design. Both are primarily a result of an insufficiently high-frequency overload margin in the critical low-level input stages combined with an overdose of lag compensation in the following stages that determine the gain-bandwidth product of the amplifier. Therefore, large-signal bandwidth and open-loop linearity become important design considerations in any amplifier if dynamic distortions are to be eliminated. The techniques for minimizing both dynamic distortions and static distortions (e.g. total harmonic distortion and intermodulation distortion) can be conflicting (1). A proper design should consider all distortion mechanisms equally without optimizing the circuit to specifically minimize any one at the expense of the other. For-



tunately, this is possible if the amplifier is designed to reject supersonic and inaudible input signals that can overload the critical low-level input stages and if sufficient local negative feedback is used in each stage to eliminate the need for excessive lag compensation for stability and freedom from oscillations.

Oscillation problems can certainly be one of the most perplexing problems that can plague a negative feedback circuit, particularly for the home builder. Therefore, frequency stability is a primary consideration in any amplifier design, and it, too, can require compromises in other design objectives. For example, an amplifier with two stages of voltage gain is probably the best-known circuit configuration. By operating both stages at a maximum gain, static distortions in the closed-loop amplifier can be minimized because this maximizes the negative feedback. However, when the gain of any stage is increased, its bandwidth is automatically reduced since the gain-bandwidth product of any stage is fixed by the particular transistor used in that stage. This decreased bandwidth can cause excessive phase shifts at high frequencies, requiring an overdose of lag compensation to prevent oscillations. The excessive lag compensation increases the amplifier susceptibility to dynamic distortions. In addition, the slewing rate can be degraded to an unacceptable level.

A good overall amplifier design philosophy is to use two stages of voltage gain followed by a current gain stage that is operated with a voltage gain of unity. The current gain stage is the power output stage which supplies all load current to the loudspeakers. Because high-current transistors have a lower gain-bandwidth product than low-current transistors, it is logical to operate the output stage at unity voltage gain so that it will have a maximum bandwidth. In this way, phase shifts in the output stage can be minimized, thereby reducing the amount of lag compensation required for stability. There are two commonly used ways to realize a unity-gain output stage. The first is to operate both the driver and the output transistors in the common-collector of emitter-follower configuration. This results in the highest gain-bandwidth product in the output stage. The second method is to operate the driver transistors in the common-collector configuration and the output transistors in the common-emitter configuration. In this case, the emitters of the driver transistors and the collectors of the output transistors are connected to the loudspeaker output, and the bases of the output transistors are driven from the collectors of the driver transistors. Because this connection forms a negative feedback path from the collectors of the output transistors back into the emitters of the driver transistors, a large reduction of static distortions in the output stage can be realized. It is felt, however, that the first connection results in a more stable and better sounding amplifier. This is because the output transistors in the second connection are operated in their slowest configuration. The driver transistors are forced to supply a higher and higher share of the load current as the frequency is increased, which in turn causes the high-frequency output impedance of the amplifier to increase, resulting in a reduced high-frequency damping factor. In addition, the feedback loops formed by the driver and output transistors involve 100 percent voltage feedback around a very high gain loop, and this makes such an output stage susceptible to oscillations which can be load-dependent and may not show up under normal testing. Although the first connection exhibits a higher static distortion, this can be overcome by operating the output stage at a higher bias current which causes the amplifier to operate Class-A over a larger signal swing.

The second stage of voltage gain drives the output stage. For a unity-gain output stage, this second stage must put out a signal equal in voltage amplitude to the loudspeaker output signal. Thus, it should be operated at a high gain in order to minimize the signal level that is required to drive it. When this is done, it follows that this stage will have the lowest bandwidth of any stage in the amplifier because of the limitations imposed by a fixed gain-bandwidth product. Therefore, it is this stage which determines the dominant pole in the amplifier and the stage which should be lag-compensated for frequency stability and freedom from oscillations.

The input stage to the amplifier also deserves special consideration. This stage must have two signal inputs, one to which the amplifier input signal is applied and the other to which the feedback signal is applied. The output signal from the input stage must be proportional to the difference between these two signals. A differential amplifier is a logical choice for this circuit. The design of the differential amplifier is very important, for it is this stage that primarily determines the susceptibility of the amplifier to dynamic distortions. First, it must have a bandwidth that is very much greater than that of the second stage to minimize the amount of lag compensation required for stability. This is achieved by operating the stage at a low gain and by lead compensating it for increased bandwidth. Second, it must have a sufficient bias current in order to drive the high-frequency capacitive input impedance of the second stage. If this bias current is insufficient, the slewing rate of the amplifier will be degraded. Third, the input stage should be designed so that it rejects any supersonic and inaudible input signals that lie outside the large-signal bandwidth of the second stage [2]. If these design objectives are properly achieved, the amplifier will be essentially free of dynamic distortion mechanisms: It cannot slew or produce transient intermodulation distortion before it clips.

The circuit architecture or topology is important in any power amplifier. To minimize static distortions, a fully complementary design is important. This is especially true for the critical output stage which must supply the full load current to the loudspeakers. A complementary circuit theoretically cancels the predominant even-order nonlinearities of the active devices, leaving only the odd-order nonlinearities to be cancelled by the negative feedback. To minimize these before overall negative feedback is applied to the circuit, it is important to use local negative feedback in each internal stage. In this way, the open-loop distortion can be made sufficiently low so that the overall negative feedback is not so high that an overdose of lag compensation is necessary for stability. With a careful balance between local feedback and overall feedback, it is possible to achieve a higher overall feedback ratio at high frequencies because of the reduced need for lag compensation, which reduces both static and dynamic distortions.

This article describes an amplifier which has been designed with these objectives and considerations in mind. The circuit has been designed so that transient intra-loop signal overload cannot occur, even with ultra-fast square-wave signals applied to the input. Because no internal stage is subject to transient overload problems, the amplifier is theoretically free of TIM distortion and it cannot slew. When properly constructed, the amplifier can be used with the finest associated equipment. It is highly stable with reactive loudspeaker loads such as electrostatic loudspeakers. Because it is capable of delivering large high-frequency transient current demands to the loudspeaker, its sound is clean and free from the so-called "transistor sound," even with music which contains loud high-frequency material and percussive sounds. In the critical midrange region, the feedback ratio is sufficiently high so that the midrange retrieval and definition are primarily determined by the source material. Although the circuit is d.c.-coupled from input to output, its low-frequency response is rolled off below 0.3 Hz. This low-frequency roll-off is accomplished by the feedback circuit rather than with a coupling capacitor in the forward signal path. By rolling off the d.c. response of the amplifier, the loudspeaker is protected from d.c. offset voltages and currents which could result from temperature effects or from a d.c. offset at the output of the preamplifier. However, the 0.3-Hz roll-off frequency is sufficiently low so that the phase shift at 20 Hz and higher is negligible [3].

Circuit Description

The complete circuit diagram of the amplifier is shown in Fig. 1. The circuit is a fully complementary design which is direct-coupled from its input to output to ensure full reproduction of even the lowest bass frequencies without phase shift distortion. A fully complementary circuit means that there is a pnp transitor for each npn transistor and vice versa. Although expensive, this approach assures low static distortion performance [1] [4] [5], especially for a single or dominant pole design that does not use maximum negative feedback, which is the present case.

The input stage consists of transistors Q1 through Q6. Q1 through Q4 are connected as a complementary differential amplifier. (As far as the author can determine, the use of the complementary differential amplifier in power amplifiers was pioneered independently in the late 1960s and early 1970s by John Curl, Bascom King, and Daniel Meyer [6].) To reduce the voltage across the transistors in the differential amplifiers to about one-half the power supply voltage, transistors Q5 and Q6 are added to the circuit and are connected in a common base configuration. By connecting the input transistors in series in this manner, each transistor in the input stage has a voltage across it that is no more than about one-half the power supply voltage. The bias current in the input stage is set at 4 mA by resistors R13 and R14. A constant current source in place of these resistors was not used to set this bias current because only the common-mode rejection ratio of the differential amplifiers would be improved. This is meaningless when it is not required to drive the differential amplifiers in the differential mode, as would be the case when driving an amplifier from a 600-ohm balanced line.

The addition of Q5 and Q6 to the input stage converts the differential amplifiers to what is called a cascode amplifier [7]. Transistor Q5 in cascode with Q1 acts like a single common-emitter transistor which has almost negligible internal feedback and a very small output conductance. Similarly, Q6 is in cascode with Q3. The gain of the input stage is set by the emitter degeneration resistors R9 through R12 and the collector resistors R15 and R16. It is approximately 6 dB. Capacitors C4 and C5 lead compensate this stage by cancelling a pole in its transfer function at 20 MHz.

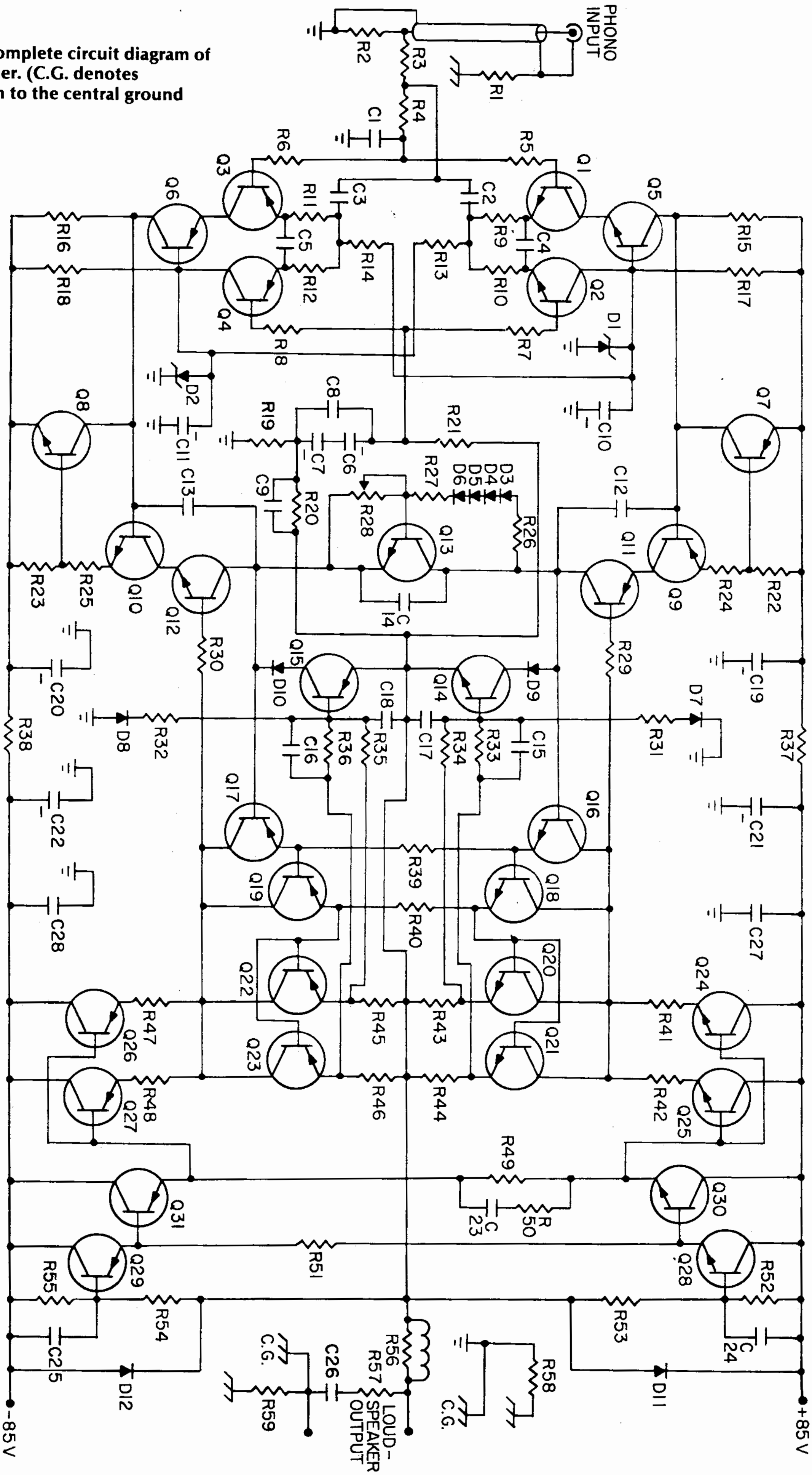
The connection of capacitors C1, C2, and C3 and resistors R3 and R4 to the input stage forms a second-order active low-pass filter [2] to protect the amplifier from inaudible and unintentional ultrasonic signals such as bias signals from a tape recorder, multiplex subcarrier signals from an FM tuner, or oscillations from a defective preamplifier. The upper -3 dB frequency of this filter is about 40 kHz. Above this frequency, it rolls off at 12 dB per octave. The filter alignment is the linear phase Bessel alignment, and it exhibits less than 0.11 degrees of phase shift distortion at 20 kHz. Rather than being a separate outboard active filter preceding the power amplifier, this filter is an integral part of the amplifier input stage itself. In addition to protection from possible damaging ultrasonic signals, the Bessel filter performs the important function of suppressing those mechanisms which cause dynamic distortions such as TIM and SID. With the filter, it is impossible for the amplifier to slew before it clips [3]. In addition, intra-loop current overshoots in the differential amplifiers are suppressed to a level much lower than that at which TIM and SID would occur. Thus, the amplifier is essentially free of dynamic distortion mechanisms.

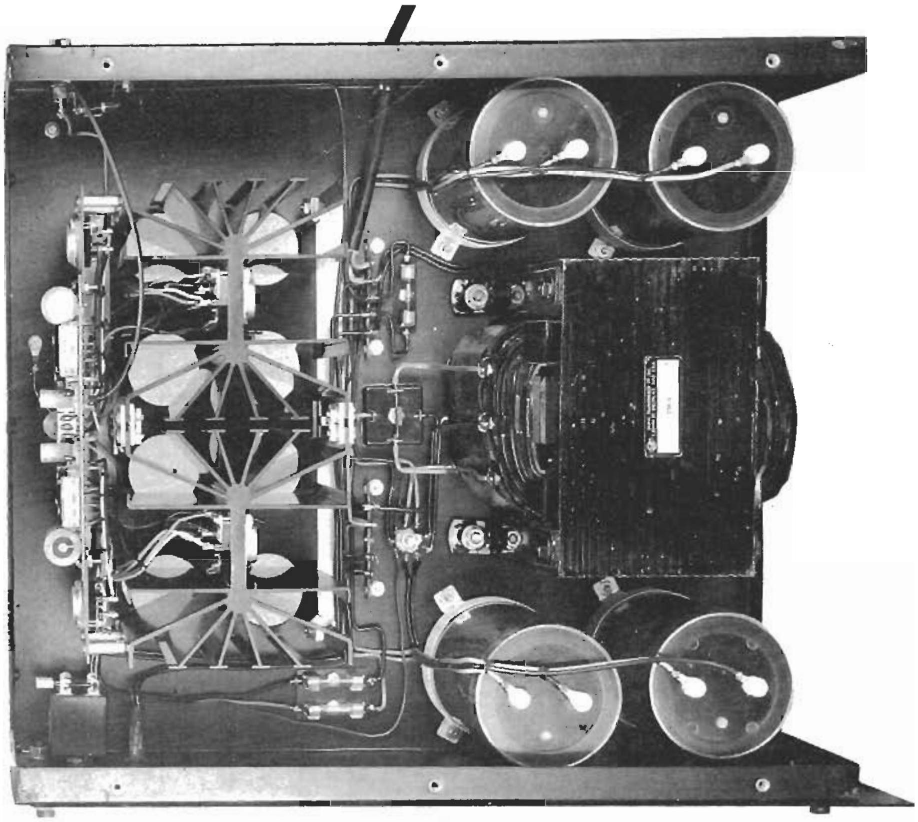
The second stage of voltage gain consists of transistors Q9 through Q12. To the author's knowledge, this is the first amplifier circuit to employ this particular cascode configuration. Transistors Q9 and Q10 act as complementary common-emitter amplifiers. Local negative feedbeck for intra-loop linearity is provided by the emitter-degeneration resistors R22 through R25. The collector output signals from Q9 and Q10, respectively, are connected to the emitter inputs of Q11 and Q12, respectively. Because Q11 and Q12 act as common-base amplifiers in the forward loop signal path, the addition of these two transistors to the circuit converts the second stage of the amplifier to a complementary cascode amplifier. However, unlike ordinary cascode amplifiers, the bases of Q11 and Q12 are not maintained at constant voltages. Instead, negative feedback signals derived from the output stage are fed back into the bases of Q11 and Q12 to cause these transistors to act as dynamically varying collector loads for Q9 and Q10. (The operation of this stage will be further explained after the output stage is described.)

Transistor Q13 is connected as a temperature compensated constant voltage source which regulates the bias current in the output transistors. Negative thermal feedback is provided by diodes D3 through D6 for thermal stability of the output stage. Mounted on the main heat sinks with the output transistors, these diodes sense the heat sink temperature to maintain constant quiescent bias current in these transistors when the heat sinks heat up under load. There are two main heat sinks, and two of the bias diodes are mounted in series on each. Potentiometer R28 is used to adjust the bias for minimum distortion.

The signal is coupled from the collectors of transistors Q11 and Q12 to the bases of the main output transistors by the Class-A discrete Darlington driver stage formed by transistors Q16 through Q19. The advantages of this Darlington driver configuration have been discussed in [4]. In brief, the very high input impedance to the drivers presents a load impedance to transistors Q11 and Q12 that is essentially independent of the loudspeaker load impedance. This makes it possible to operate the second stage at a very high gain in order to minimize static distortions in the amplifier. The extremely low output impedance of the Darlington drivers makes it possible to bias the output transistors for zero crossover distortion in the Class AB mode. Thus, biasing the amplifier in the inefficient Class A mode or the use of feedback to dynamically vary the bias voltage for a quasi-Class A mode of operation will result in no improvement in performance. The reason for this is that the output impedance of the driver stage is low enough so that the output transistors operate at their maximum bandwidth, which is equal to their gain-bandwidth product. Because this is greater than the unity-gain loop bandwidth of the amplifier, the speed of the ampli-

Fig. 1 — Complete circuit diagram of the amplifier. (C.G. denotes connection to the central ground point.)





fier is not set by the speed of the output transistors but by the stages that drive it.

The main output transistors in the output stage are transistors Q20 through Q23, which are operated in the emitter-follower or common-collector configuration for maximum bandwidth. They are biased in the Class AB mode for minimum distortion and minimum power dissipation. In the Class AB mode, all output transistors are conducting during no or small-signal inputs. However, as the input signal dynamically increases, two of the output transistors will progressively conduct more and the other two will progressively conduct less during any half cycle of the input signal until the latter two transistors cut off. During the signal transition through the zero signal or crossover region, all output transistors are conducting and the amplifier operates Class A. This eliminates all traces of the spike in the distortion waveform caused by crossover distortion.

The collector voltage across each output transistor is varied dynamically with the output signal to reduce the voltage across these transistors to one-half the voltage each would have to drop in a conventional design. This greatly improves the reliability of the output stage because high-power transistors can reliably deliver high load currents only at low collector to emitter voltage. This dynamic variation of the voltage is accomplished by transistors Q24 through Q31, which operate in the Class A mode. Because resistors R52 and R53 are equal, the voltage at the base of transistor Q28 is equal to the loudspeaker output voltage plus one-half the difference between the positive power supply voltage and the loudspeaker output voltage. Similarly, because R54 and R55 are equal, the voltage at the base of Q29 is equal to the loudspeaker output voltage plus one-half the difference between the negative power supply voltage and the loudspeaker output voltage. If the relatively small base-to-emitter voltage drops for transistors Q24 through Q31 are neglected, it follows that the voltage at the collectors of Q16, Q18, Q20, and Q21 is forced to vary so that it is always approximately halfway between the loudspeaker output voltage and the positive power supply voltage. Similarly, the collector voltage for transistors Q17, Q19, Q22, and Q23 is dynamically varied so that it is halfway between the loudspeaker output voltage and the negative power supply voltage. Thus, for the 85-volt positive and negative power supplies, the output transistors can never have more than 85 volts across them. This is 55 volts less than the rated open base breakdown voltage of 140 volts for these transistors. The output stage is therefore operated conservatively to prevent the transistors from operating outside their safe operating area. To the author's knowledge such a technique for dynamically varying the collector voltage of the output transistors in a power amplifier was first published in 1974 by James Bongiorno [8].

There have been other methods described recently for varying the supply voltages to the output stage in power amplifiers that involve diode or-gates which abruptly switch in which the output signal level exceeds the threshold of the or-gate, e.g. the so-called Class G circuits which typically use a poorly regulated high voltage power supply to provide transient peaks. Although these circuits are more efficient and thus can be designed with lighter heat sinks and transformers, the present technique is much preferred because it dynamically varies the collector supply voltages to the output transistors so that these voltages linearly track the output signal. Although this is a more expensive approach, it circumvents the problem of switching distortion which can be caused by the higher voltage power supplies switching in on high power peaks.

The dynamically varying collector voltages for transistors Q16 through Q23 are fed back to the bases of transistors Q11 and Q12 in the second stage of voltage gain. Because the base-to-emitter voltage drops for Q11 and Q12 are small, it follows that the voltage across Q9 and Q10 is approximately one-half of what the voltage would be in a conventional design. The other one-half is dropped by Q11 and Q12. The bases of Q11 and Q12 are the signal inputs to an internal negative feedback loop that encircles the output stage. This is a unique feature of the amplifier which greatly reduces static distortions produced in the output stage. Thus, the feedback signal to the differential amplifiers contains fewer distortion components, permitting the differential amplifier to be operated at a lower gain for lower dynamic distortions, greater frequency stability, and a higher slew rate.

The protection circuit consists of transistors Q7, Q8, Q14, and Q15, and diodes D7 through D12. The amplifier output voltage and current are sensed by Q14 and Q15 to automatically limit the output current in the event that the loudspeaker wires are accidentally short circuited or if a severely reactive load is driven which could damage the output transistors. With a short circuit on the output, the protection circuit will limit the peak output current to about 6 amperes. With a load resistance of about 2 ohms or greater, it will not limit. When the limiter circuit is tripped, Q14 or Q15 saturates, which short circuits the base drive signal for Q16 and Q17. Because this short circuits the collector outputs of Q11 and Q12 to the amplifier output terminal, these two transistors must be protected by limiting the maximum current they can deliver. This is accomplished by transistors Q7 and Q8, which sense the current drawn by Q9 through Q12 and limits it to approximately 30 to 35 mA. Diodes D11 and D12 are damper diodes that protect the output stage from inductive transients which can be caused by an excessively inductive load impedance. Diode D7 prevents Q14 from limiting on negative output signal swings while D8 prevents Q15 from limiting on positive swings. Diodes D9 and D10 protect Q14 and Q15 from inductive transients. All diodes in the protection circuit are fast-recovery, low-capacitance diode types, minimizing false triggering of the limiter circuit and assuring a quick recovery should limiting occur. Capacitors C17 and C18 slow the response of the limiter so that it will not trigger on fast transients. Capacitors C15 and C16 suppress oscillations that could occur in the limiter circuit if it is triggered.

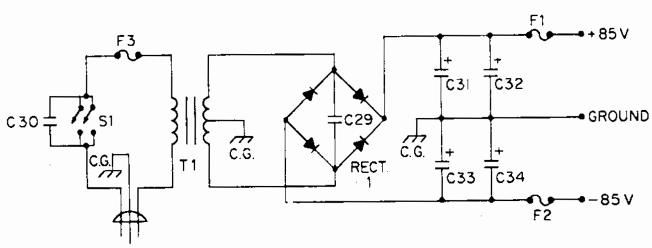


Fig. 2 — Circuit diagram of the amplifier power supply.

Although limiter circuits have a bad reputation among some audiophiles, they are a necessary evil in any application where the amplifier will be moved around or where the loudspeaker leads will be connected or disconnected during operation such as in sound reinforcement work. It is recommended that the amplifier be built with the limiter to protect the circuit in the event of construction errors. After the amplifier is operational, transistors Q14 and Q15 may be removed to disable it. However, this is not recommended. Without the limiter, the amplifier can be seriously damaged in the event that the loudspeaker output leads are accidentally short circuited. An excellent discussion of the design of amplifier protection circuits such as the voltage-current sensing limiter used in this design is given in [9].

The feedback network consists of resistors R19 through R21 and capacitors C6 through C9. This network has two feedback paths — an a.c. path and a d.c. path. The signal outputs from these two paths are summed to form the total feedback signal. The a.c. loop is formed by resistors R19 and R20, which set the closed-loop gain of the amplifier at approximately 26 dB. The d.c. feedback network continuously monitors the loudspeaker output voltage for a d.c. offset which could damage the loudspeaker. If a d.c. offset voltage appears at the output of the amplifier, it will be fed back with no attenuation through R21 into the differential amplifiers, where a correction voltage will be generated to cancel the offset. Capacitors C6 and C7 in series form a non-polar electrolytic that determines the lower -3 dB frequency of the amplifier which is 0.3 Hz. Below this frequency, these capacitors become open circuits and remove the a.c. feedback to the inverting input of the differential amplifiers. This gives 100 percent d.c. feedback for stability of all d.c. bias voltages and currents in the circuit. The phase shift associated with the gain roll-off below 0.3 Hz is less than one degree at 20 Hz. Capacitor C8 is a metalized polyester capacitor which bypasses the electrolytics at high frequencies.

One of the most important considerations in the initial design stages of an amplifier is the specification of the desired gain-bandwidth product. This is the upper frequency at which the open-loop gain has reduced to unity. When this frequency is divided by the desired closed-loop gain, the upper -3 dB frequency of the closed-loop amplifier is obtained. The gain-bandwidth product is another example of how the uncertainty principle affects almost everything in physics and engineering. The higher the gain-bandwidth product of an amplifier, the wider its bandwidth and the lower its distortion. However, the higher the gain-bandwidth product, the more susceptible the amplifier is to oscillations. This is especially true for the case of reactive loudspeaker loads, since they can be connected to the amplifier by so-called "high definition" cables that exhibit high capacitive load effects on the amplifier. (It is felt that some of these cables are marketed with erroneous and deceptive claims. The fact that a number of them have caused some amplifiers to self-destruct is evidence of their high capacitive loading effect. Some amplifiers oscillate with capacitive loads and can either self-destruct or sound different, particularly if the capacitive load trips the protection circuit. If the amplifier does not misbehave with these cables, it is doubtful that the loudspeaker will sound any different with them than with an equivalent gauge zip cord.) It was decided at the start of this project that stability margin would not be sacrificed for gain-bandwidth product and that the amplifier would be designed for unconditional stability in the sense that its square wave response, with the signal applied after the Bessel input filter, would exhibit no overshoots or ringing and that the amplifier would remain stable if its open-loop gain were reduced. Conditionally stable amplifiers can be designed for extremely low static distortion levels by staggering pole zero combinations in the open-loop gain function at frequencies below the unity loop-gain frequency, thus achieving very high levels of feedback. Such amplifiers are available on the commercial market and their design has been discussed in the literature [10]. However, it is felt that conditionally stable amplifiers are more susceptible to dynamic distortions and load-induced oscillations. In addition, their stability margin is reduced should the open-loop gain of the amplifier decrease with age.

During the experimental phase of this design, it was found that a 10-MHz gain-bandwidth product could be obtained with no square wave overshoots. The amplifier is stable with a higher value at the expense of a slight overshoot in the square wave response. However, it is felt that the 10-MHz figure is adequate, for it gives a closed-loop upper -3 dB frequency of 500 kHz when the amplifier is operated at a closed-loop gain of 20. Once the desired gain-bandwidth product has been specified, it is possible to design the input stage to set the slewing rate of the amplifier. This was chosen to be 80 volts per microsecond, a figure that is about 10 times that required to reproduce a full power sine wave at 20 kHz without slew-rate limiting. Conversely, the slew rate is high enough so that the amplifier can reproduce a full power sine wave up to a frequency of about 200 kHz. However, the builder is cautioned never to try this test, for the output transistors could be damaged. For a gain-bandwidth product of 10 MHz and a slewing rate of 80 volts per microsecond, the required ratio of the transconductance to bias current for the transistors in the input differential amplifiers can be calculated [11]. It is 0.79 Mho per ampere. Thus for the 4-mA bias current in the differential amplifiers, the required transconductance of this stage is 0.0031 Mho. This value is achieved by the addition of the 270-ohm emitter degeneration resistors R9 through R12. Capacitors C4 and C5 lead compensate this stage by cancelling a pole in its transfer function at 20 MHz. The gain-bandwidth product of the overall amplifier is set at 10 MHz by capacitors C12 and C13 in the second stage of gain. These capacitors also act as pole-splitting capacitors to increase the frequency of higher order poles in the second stage for a better stability margin. Capacitors C9, C24, and C25 complete the frequency compensation of the amplifier. C9 lead compensates in the feedback network to cancel a 2-MHz pole that occurs in the current gain of the output transistors. C24 and C25 stabilize the internal negative feedback loop that encircles the output stage.

By taking the present approach to design the amplifier for a specified slew rate and gain-bandwidth product, the use of nonlinear Class B slew-enhancement techniques in the input stage to increase slew rate as described in [10] can be avoided. Another technique which is used by some designers to achieve a high slew rate is to operate two identical amplifiers in a bridged or strapped configuration. The slew rate of the bridged combination is twice the slew rate of the unbridged amplifiers. However, the highest frequency at which the bridged combination can reproduce a full power sine wave without slew rate limiting is not doubled; it is equal to the large-signal bandwidth of the individual amplifiers. Thus bridging cannot be used to increase large-signal bandwidth as can be achieved by increasing the slew rate of the individual amplifiers.

The power supply is shown in Fig. 2. The transformer is a heavy duty, 12-ampere transformer weighing 38 pounds. Because the power supply voltage is plus and minus 85 volts, it is necessary to use 100-volt or higher capacitors in the power supply filter because the next lowest standard voltage for electrolytic filter capacitors is 75 volts. Unfortunately, it is very difficult to locate large-value 100-volt electrolytic capacitors, especially on the money saving surplus market. In the author's amplifier, two 8,600 - μF capacitors were connected in parallel for each power supply filter. This gives a total power supply capacitance of 34,400 μF. This value can be increased, but it is felt that any improvement will be minor because the amplifier is designed so that the output transistors cannot saturate when the amplifier clips. This prevents power supply ripple from being coupled into the loudspeaker load through saturated output transistors. Instead, when the amplifier clips, either Q9 or Q10 becomes saturated. Because these transistors are isolated from the power supply by low-pass filters, the ripple on the power supply lines cannot be coupled through Q9 and Q10 and into the loudspeaker.

Specifications and Measurements

The average sine wave power rating of the amplifier is 250 watts into an 8-ohm load, corresponding to a load voltage of 45 volts rms or 63 volts peak. Continuous sine wave power measurements load down the power supply more than audio signals because audio signals of the same peak level have a much lower average power level. Thus, a meaningful specification is the peak voltage output of the amplifier into an open circuit, i.e. the maximum transient peak that the amplifier could supply to any load impedance before the voltage on the power supply filter capacitors drops. This peak voltage for the amplifier is 78 volts, and it follows that, if the power supply were perfectly regulated, the average sine wave power rating into 8 ohms would be 380 watts. This is 1.8 dB higher than the average sine wave power rating and corresponds to the dynamic headroom of the amplifier. (Dynamic headroom can be a misleading specification. An amplifier with a perfectly regulated power supply will have a dynamic headroom of 0 dB. Therefore, the smaller this number, the better the power supply but the larger this number, the higher the transient peak power capability.)

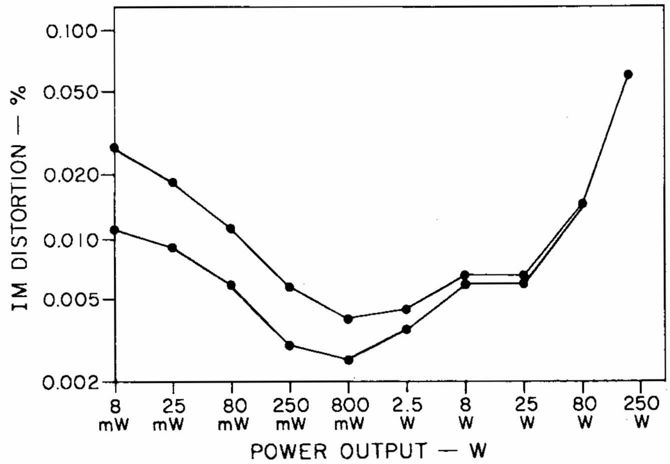


Fig. 3 — SMPTE IM distortion of the amplifier when driving an 8-ohm load. Points are connected by straight lines for plotting. The two curves show uncertainty in the low-level measurements caused by noise in the measurement system.

To test the amplifier for distortion, it was decided to abandon conventional total harmonic distortion (THD) measurements and investigate only the intermodulation distortion (IM) characteristics of the circuit. This is because THD measurements only identify distortion components which are harmonically related to the test frequency, and these components often lie outside the audible frequency band. Thus, some THD specifications can' be meaningless, particularly those made above 10 kHz. The most audible and annoying distortion is not harmonically related to the signal frequency. This type is called IM distortion, and it is generated when two or more signals with different frequencies interact inside the amplifier to produce what are called intermodulation products. The ear is most sensitive to IM distortion because, by definition, it is non-harmonically related to the signal, whereas THD is.

Two types of IM measurements were made on the amplifier. These were the SMPTE IM measurement and the DIM-100 measurement. The SMPTE IM standard, written by the Society of Motion Picture and Television Engineers, is a measurement technique that identifies static distortion mechanisms, i.e. those that are dependent only on the amplitude characteristics of the signal. Crossover distortion is an example of static distortion, and the SMPTE IM test is extremely sensitive to it. A good discussion of this test and its implementation is provided in [12]. The SMPTE IM distortion of the amplifier was measured with a Crown IMA intermodulation analyzer which uses two simultaneous sinusoidal test tones, one at 60 Hz and one at 7 kHz with an amplitude ratio of 4 to 1. The IM distortion is determined by measuring the percentage amplitude modulation on the 7-kHz tone caused by the larger amplitude 60-Hz tone. The measurement results are shown in Fig. 3. At 250 watts into 8 ohms, the IM level of the amplifier was 0.054 percent. At lower levels, it decreased to 0.004 percent at 800 milliwatts, increasing below that level. Noise in the measurement system had a strong affect on the low-level distortion measurements, and the data in Fig. 3 have been corrected for it. The upper curve in this figure is the rms difference between the distortion and the residual noise, and the lower curve is the algebraic difference between these; the actual distortion should lie somewhere between the two curves. The residual noise was determined by turning the level of the output signal on the IMA analyzer to

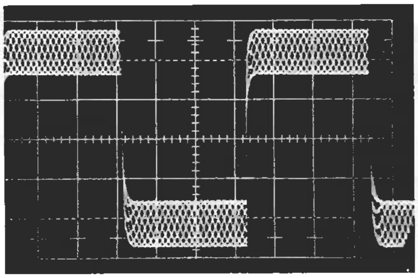


Fig. 4 — DIM-100 input test signal consisting of a 3.18-kHz square wave and a 15-kHz sine wave added with the ratio of 4 to 1. Vertical scale is 1 V per division; horizontal scale is 50 μS per division.

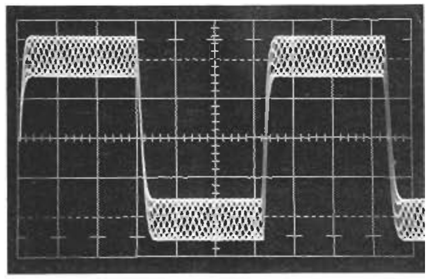


Fig. 5 — Output signal of the amplifier when driving an 8-ohm load with the DIM-100 input test signal. Vertical scale is 25 V per division; horizontal scale is 50 μS per division.

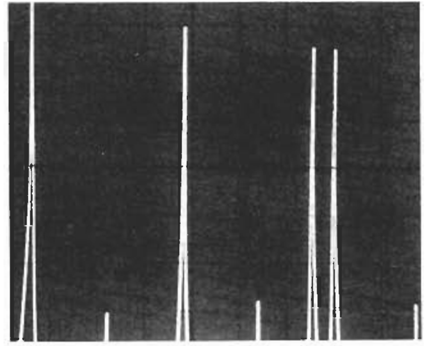


Fig. 6 — Spectral analysis of the amplifier output signal when driving an 8-ohm load with the DIM-100 input test signal. Vertical scale is 10 dB per division; horizontal scale is 2-kHz per division from 0 to 20 kHz.

zero and recording the percentage distortion reading on the analyzer meter. It was determined that the source of this noise was the analyzer, for the percentage distortion dropped to zero when the phono input jack on the amplifier was short circuited, indicating the noise was not being generated by the amplifier. There is no evidence of crossover distortion in the IM measurement data. This would have shown up as a large peak in the IM level in the critical power range from 10 milliwatts to 1 watt. (Author's Note: The noise was later traced to improper grounding of the IMA analyzer.)

The DIM-100 test is one that can be used to identify IM distortion components that fall in the audible range which are caused by distortion mechanisms dependent on both the amplitude and frequency characteristics of the signal. Examples of such distortions are TIM and SID, and this test is extremely sensitive to them. The DIM-100 measurement technique uses a spectrum analyzer to identify and measure the IM components when a square wave and a sine wave are simultaneously applied to the amplifier. Such a measurement technique was originally described by Schrock [13], although no method was given to quantify the measurements. In a later paper by Leinonen, et. al. [14], a method to do this was presented, and the test was called a dynamic intermodulation (DIM) test. It specifies a 3.18-kHz square wave and a 15-kHz sine waved added with the amplitude ratio of 4 to 1 for the amplifier test signal. The square wave is specified to be low-pass filtered with a single-pole filter. The upper -3 dB frequency of this filter is specified to be 30 kHz (DIM-30) for normal testing and 100 kHz (DIM-100) for measurements on the highest class of equipment. This latter test was the one chosen for the amplifier. The percentage of DIM-100 is specified as 100 percent times the ratio of the rms sum of the IM distortion components at the 9 IM frequencies which lie in the audio band to the rms amplitude of the 15-kHz sine wave. These 9 IM frequencies are 0.90, 2.28, 4.08, 5.46, 7.26, 8.64, 10.44, 11.82, and 13.62 kHz. A spectrum analyzer or frequency-selective voltmeter must be used to measure these.

It has been shown that high-frequency THD measurements can be used to identify dynamic distortion mechanisms in operational amplifiers [15]. These tests are also valid in testing power amplifiers but were not used for several reasons, the main one being that the THD must be measured with the amplifier operating near its slew-rate limit. A slew rate of 80 volts per microsecond would require full power THD measurements at 200 kHz and higher, which could damage the output transistors, and the accuracy of distortion analyzers above 200 kHz may not be good. Second, the measurements do not identify IM products that could fall back into the audio range, and these are the distortion components we hear. Finally, THD measurements above the 40-kHz cutoff frequency of the Bessel input filter would require unrealistically large input signals. For example, it would take 56 volts rms at the amplifier input for it to put out a full power sine wave of 45 volts rms at 200 kHz.

Two Hewlett Packard 3310A function generators were used to generate the DIM-100 test signal, and a Hewlett Packard 5381A frequency counter was used to adjust the frequency of each generator to correspond to those specified for the DIM-100 test. An oscillogram of this signal is shown in Fig. 4. Because there is no harmonic relationship between the frequencies of the sine wave and the square wave, the oscillogram shows several cycles of the sine wave superimposed on the square wave. The levels were chosen so that the square wave term produced a 100-volt peak-to-peak square wave at the amplifier output while the sine wave term produced a 25-volt peak-to-peak sine wave. The total peak-to-peak signal swing was thus 125 volts, or 1.5 volts less than that of a 250-watt sine wave into 8 ohms. The amplifier output signal when driving an 8-ohm load is shown in Fig. 5. The average power delivered to the load with this signal was 322 watts,

Fig. 7 — Square wave response of the amplifier at 50 Hz measured at 200 watts with an 8-ohm load. Vertical scale is 20 V per division; horizontal scale is 5 m S per division.

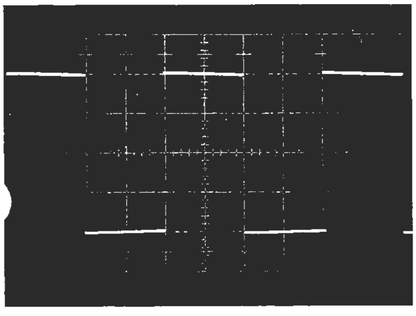


Fig. 8 — Square wave response of the amplifier at 10 kHz measured at 200 watts with an 8-ohm load. Rounded rise and fall portions of waveform are caused by Bessel low-pass input filter which rolls off amplifier response at 12 dB per octave above 40 kHz. Vertical scale is 20 V per division; horizontal scale is 20 μS per division.

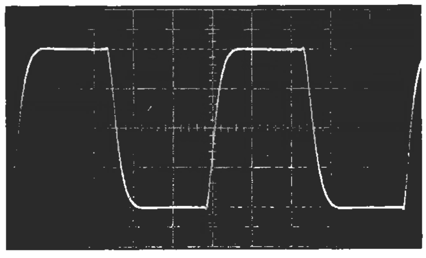
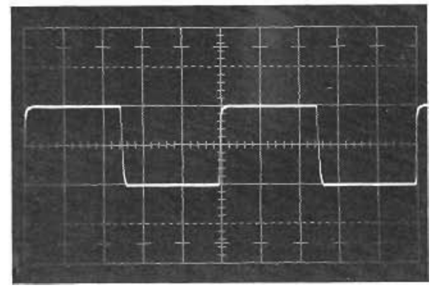


Fig. 9 — Square wave response of the amplifier at 20 kHz with the Bessel input filter bypassed. Measured at 20 V peak-to-peak with an 8-ohm load. Vertical scale is 10 V per division; horizontal scale is 10 μS per division.



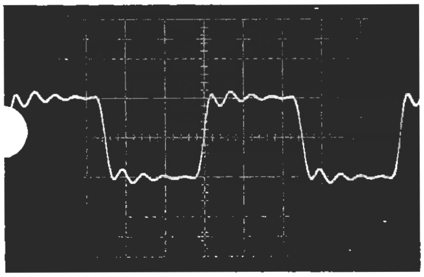


Fig. 10 — Square wave response of the amplifier at 10 kHz measured at 10 V peak-to-peak with a 2-µF capacitor load. Vertical scale is 10 V per division; horizontal scale is 20 µS per division.

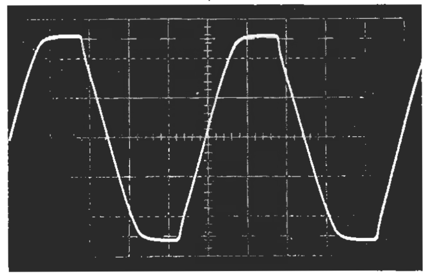


Fig. 11 — Sine wave response of the amplifier at 20 kHz with 2 dB of overdrive measured with an 8-ohm load. Vertical scale is 25 V per division; horizontal scale is 10 µS per division.

which was high enough to cause the load resistors to get quite hot during the test. (A sine wave with the same peak-to-peak signal level would correspond to an average power level of 244 watts.) To determine the DIM-100 distortion components in the output signal, the amplifier output was connected through a variable attenuator to the input of a Tektronix 5L4N spectrum analyzer, and its display is shown in Fig. 6.

Figure 6 shows a spectral analysis of the amplifier output signal displayed over a dynamic range of 80 dB from 0 to 20 kHz. There are seven frequency components in the figure, occurring at 3.18, 6.36, 9.54, 12.72, 15.00, 15.90, and 19.08 kHz. With the exception of the 15-kHz sine wave term of the test signal, all are harmonics of the square wave term. (The even-order harmonics were caused by an asymmetry in the 3.18-kHz square wave and would be absent with a perfect square wave.) There are no identifiable IM components at any of the nine specified frequencies for the DIM-100 test. The amplifier thus has no DIM-100 components that are greater than -80 dB below the fundamental frequency component of the 3.18-kHz square wave, and the same was true for tests at lower signal levels. It is important to point out that the DIM-100 test does not measure high-frequency IM products lying outside the audible band, so distortion mechanisms which could affect high-frequency THD measurements are not measured. This is because these mechanisms can only affect the sound quality of an amplifier if the distortion components they produce fall back into the audible frequency band — in which case the DIM-100 test will detect them.

Figures 7 through 11 summarize the waveform responses of the amplifier. The 50-Hz and10-kHz square wave responses at 200 watts into 8 ohms are shown in Figs. 7 and 8. The slight amount of tilt in Fig. 7 is caused by the use of 100 percent feedback at d.c. to stabilize the quiescent bias voltages and currents. The waveform in Fig. 8 exhibits the characteristic response of a second-order Bessel low-pass filter, and the upper -3 dB frequency of this filter is 40 kHz. This is low enough so that the maximum first derivative or time rate-of-change of the output signal can never approach the slew-rate limit of the amplifier when reproducing a worst-case signal, which is a full power square wave. The -3 dB frequency is high enough, however, so that deviations from ideal amplitude and phase response below 20 kHz are negligible [3]. Figure 9 shows the 20-kHz square wave response with the signal applied after the Bessel filter. The absence of overshoots or ringing in the waveform indicate unconditional stability with a phase margin that is close to 90 degrees. Thus, the amplifier is not subject to oscillations because of load effects or if the open-loop gain should decrease with age. The 10-kHz square wave response into a 2- µF capacitor is shown in Fig. 10. This figure reveals only a small amplitude ringing with very little overshoot, indicating outstanding stability of the amplifier into a reactive load. Figure 11 shows the 20-kHz sine wave clipping response into an 8-ohm load. The input signal overload for this test was 2 dB, and the figure shows a symmetrical clipping characteristic with very little evidence of "sticking." The tests reported in Figs. 10 and 11 are torture tests which, in general, should not be performed by the unexperienced unless the tester is aware of the consequences. The author has seen some amplifiers self-destruct during these tests.

The signal-to-noise ratio of the amplifier was measured with a Bruel and Kjaer 2609 measuring amplifier. With the phono input short circuited to ground, the output noise measured 0.67 millivolts unweighted and 0.174 millivolts "A"-weighted. These translate into signal-to-noise ratios referenced to 250 watts into 8 ohms of 96.5-dB unweighted and 108.2 dB "A"-weighted.

The damping factor was measured by driving the amplifier output terminal from the output terminal of a second amplifier, with an 8-ohm resistor connecting the two. The damping factor is one plus the ratio of the voltage at the output of the second amplifier to the voltage at the output of the amplifier under test. A Hewlett Packard 3575A gain phase meter was used to measure this ratio. At 20 Hz, the damping factor was found to be in the range of 300 to 500, the uncertainty caused by the effects of noise on the measurements. At 20 kHz, the damping factor was 60. It follows that the output impedance of the amplifier at 20 Hz is between 0.016 and 0.027 ohms, and at 20 kHz it is 0.13 ohms. The increase in the high-frequency output impedance is primarily caused by the inductor L1.

Construction details of the amplifier will be provided next issue in Part II.

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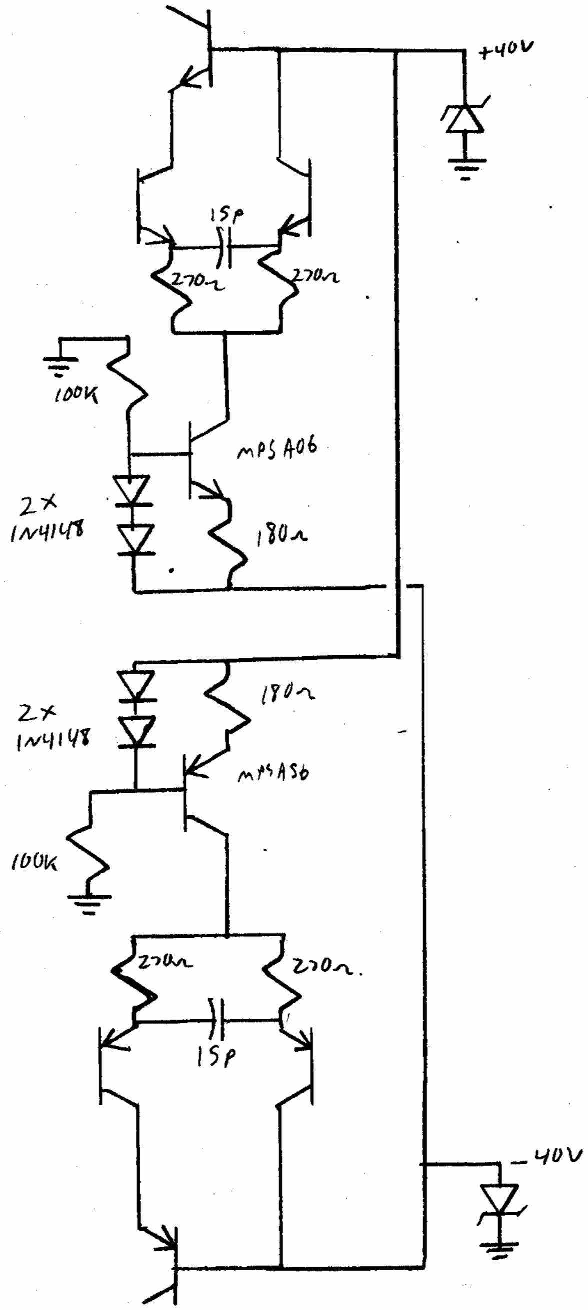


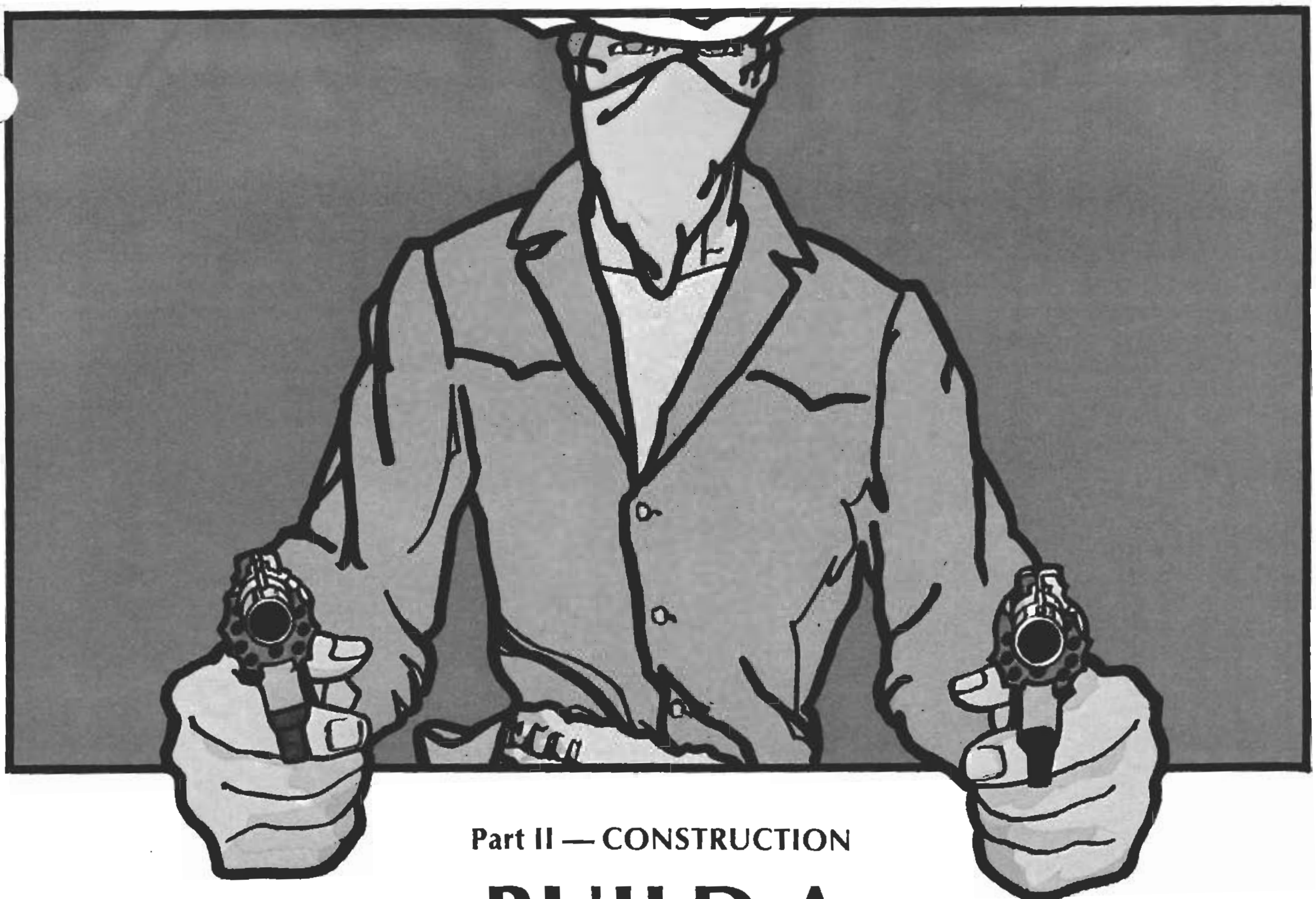
LEACH DOUBLE BARRELED AMP
CONSTANT CURRENT
SOURCE FRONT
END.

(WILL INCREASE POWER SUPPLY REJECTION RATIO SLIGHTLY)

JULY 9, 1983
Daniel Fraser
BURNABY B.C.

TO DOUBLE GAIN RAISE R20 FROM 5K6 1W TO
11K 1W

(WILL REDUCE S/N RATIO BY 3db AND MAY CAUSE
SLIGHT RINGING IN TOP END. WILL INCREASE DISTORTION
SLIGHTLY ON STATIC TEST AND MAY AFFECT SLEWING
INDUCED DISTORTION)



Part II — CONSTRUCTION

BUILD A DOUBLE BARRELED AMPLIFIER

W. Marshall Leach, Jr.*

This is an advanced construction project not recommended for those without experience in building home electronics projects. Power electronics must be respected, for it is unforgiving of even the smallest errors that could lead to the loss of expensive transistors. To minimize the chances of errors, test each part before assembly. Measure resistors with an ohmmeter to double-check the color code — otherwise, orange could be mistaken for red, blue for grey, etc. Check capacitors with an impedance bridge or capacitance meter, otherwise a dipped silver mica capacitor labeled 430 may be incorrectly thought to be 430 pF when the last digit is the multiplier and the capacitor is only 43 pF. The label codes vary with manufacturer, and 430 can be either 43 pF or 430 pF, depending on the code. When electrolytic capacitors are installed, their polarity should be double-checked. If incorrect, the capacitor will become a short circuit when power is applied.

Diodes and transistors should also be checked with an ohmmeter. Diodes should measure a high resistance with one polarity of the test leads and a low resistance with the other polarity. Because the resistance of the human body can affect these readings, the metal probes on the test leads should not be touched when making the measurements. Transistors are more complicated to test, for they require six resistance measurements. A low resistance should be measured with one polarity of the leads and a high resistance with the other polarity from base to emitter and from base to collector. A high resistance should be measured with both polarities from collector to emitter. Most transistors that have failed in a circuit will measure a short circuit on this last test. Neglecting to perform these simple tests at the start of construction can cause a lot of grief when the amplifier is first powered.

The construction details described here are broken into

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Illustration: Barbara D'Aprile

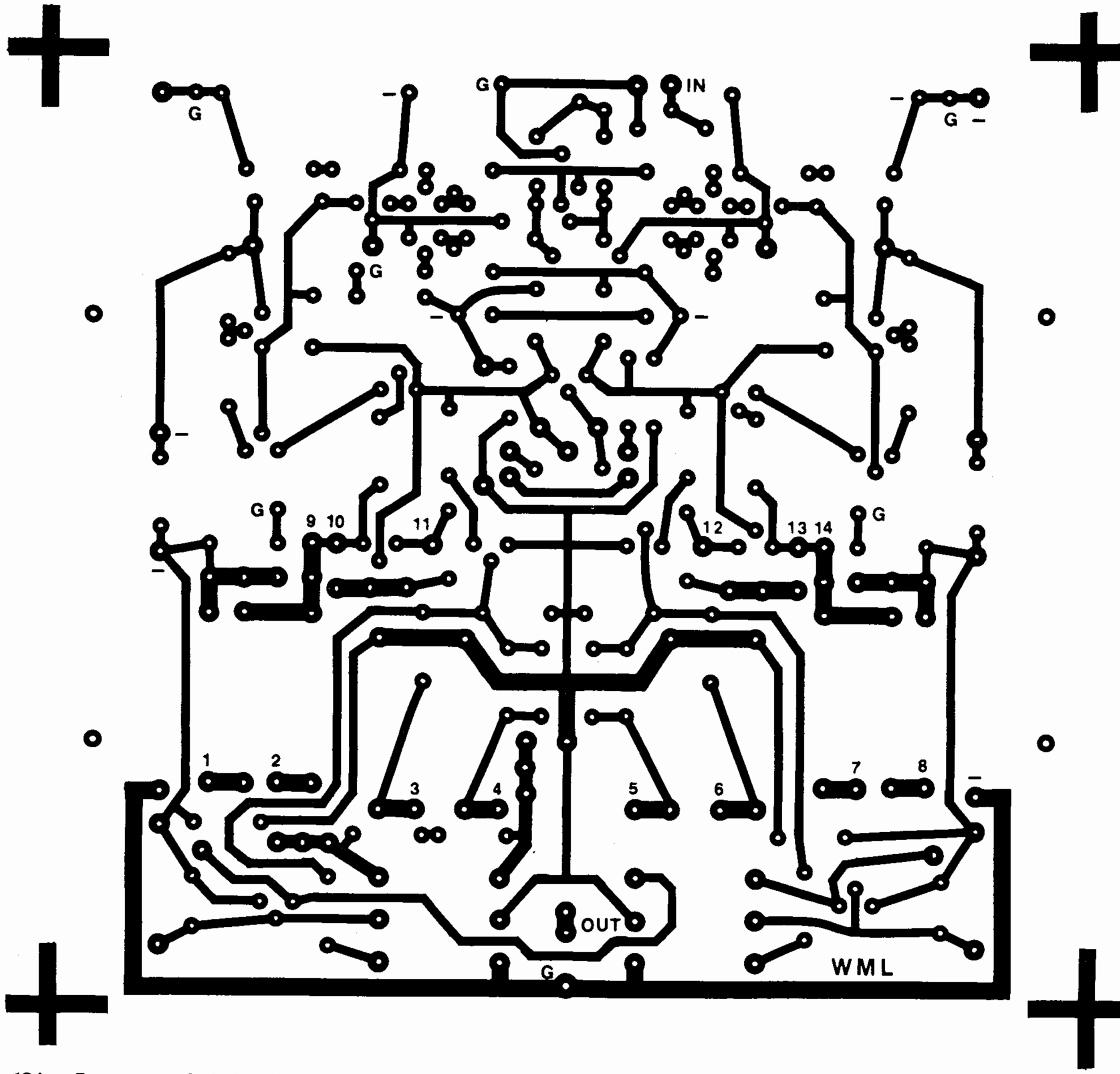


Fig. 12A.—Recommended circuit-board foil pattern for the copper foil or circuit side.

two parts. In this section, the circuit board and heat sink assembly are discussed. In the following section, the chassis wiring details are presented. Before construction is begun, all parts should be assembled so that the layouts can be modified to accommodate any parts having different dimensions from those used by the author. The recommended printed circuit board is a double-clad board, one side of which is used as a ground plane for the circuit. Ground plane construction is standard practice in r.f. circuits, and it is highly recommended for this amplifier. The front and back circuit-board foil patterns are shown in Fig. 12; parts locations are shown in Fig. 13.

Circuit-Board and Heat-Sink Assembly

For optimum results, match the differential amplifier transistors for equal current gains at a collector-to-emitter voltage of 40 V. This will ensure freedom from d.c. offset problems at the loudspeaker output terminal caused by unequal base currents in the input-stage transistors. A simple test circuit can be constructed to match these transistors, as shown in Fig. 14. Optimally, Q1 through Q4 should have equal or nearly equal current gains. Should it be difficult to find a match for all four transistors, it is sufficient to match Q1 to Q3 and Q2 to Q4. A third combination is to match Q1 to Q2 and Q3 to Q4, but this is less desirable.

The first step is to solder six of the seven ground connections on the board; ground connections are marked with a G in Fig. 12A. The one which is not soldered at this point is the one nearest the loudspeaker output connection. To solder these connections, insert a ¼-inch length of No. 22 solid wire through each hole and bend it down against the copper on each side of the board. On the ground plane side, it is best to bend these wires away from any adjacent component locations, especially those near capacitors C27 and C28. The connections can now be soldered. Sufficient heat must be used on the ground plane side to get a good solder joint — a good joint is one for which the solder has flowed smoothly onto the circuit board, and it appears shiny. Only a controlled heat soldering iron such as the Weller Soldering Station should be used to solder to the circuit boards. A 700-degree soldering

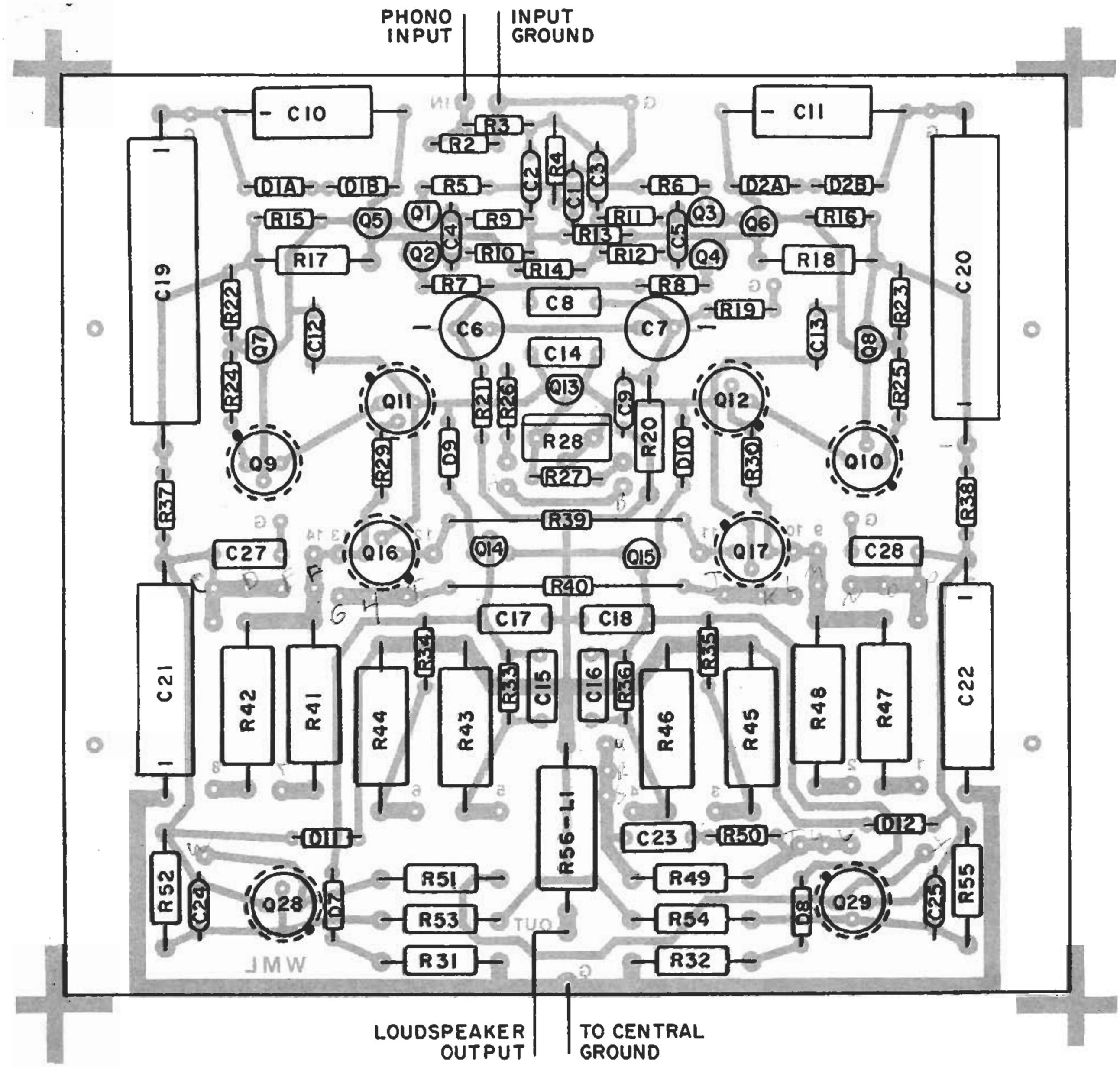


Fig. 13 — Parts layout for the recommended foil pattern.

height above the board. All require clip-on heat sinks which should be installed on the transistors *before* they are soldered to the circuit boards to prevent bending the transistor leads. It is not necessary to use any heat sink compound on these transistors.

The driver heat sinks can now be mounted to the circuit board as shown in Fig. 15. They are fabricated from 3/32-inch sheet aluminum, and the recommended dimensions are 5½ x 2⅛ inches. Mount them on the side of the circuit board opposite the ground plane with No. 4-40 x ⅜ inch screws with an inside star lockwasher under each nut. Insulate the driver heat sinks from the circuit board ground plane by installing a No. 4 fiber shoulder washer in each of the four mounting holes on the circuit board. The driver transistors can now be installed as shown in Fig. 15. A mica insulating wafer should be placed between each transistor and the heat sinks, and both sides of the mica wafers should be lightly coated with thermal compound. Use only molded plastic transistor sockets to install the drivers, and before connecting them to the circuit board, check with an ohmmeter for a possible collector short circuit to the heat sink or a short from a heat sink to the circuit board ground plane. If there are none, the driver leads can be connected and soldered as shown in Fig. 15; the connection codes are provided in Table I.

The final step in the circuit-board assembly is to solder the input coaxial cable, resistor R58, and the power supply, loudspeaker output, and power-supply ground leads to the board. With the exception of R58, these are all inserted from the side opposite the component side. Connect and solder appropriate lengths of color-coded No. 16 stranded wire to the plus and minus power-supply connections and to the loudspeaker output connection. To the ground connection nearest the loudspeaker output, insert one lead of resistor R58 over which 0.5-inch of insulation has been installed and one end of a length of No. 22 stranded wire. Solder these on *both sides* of the circuit board. Finally, connect and solder an appropriate length of miniature coaxial cable to the signal input and signal ground connections on the circuit board. This completes the circuit-board assembly.

Preparation of the main heat sinks is the next step. If they

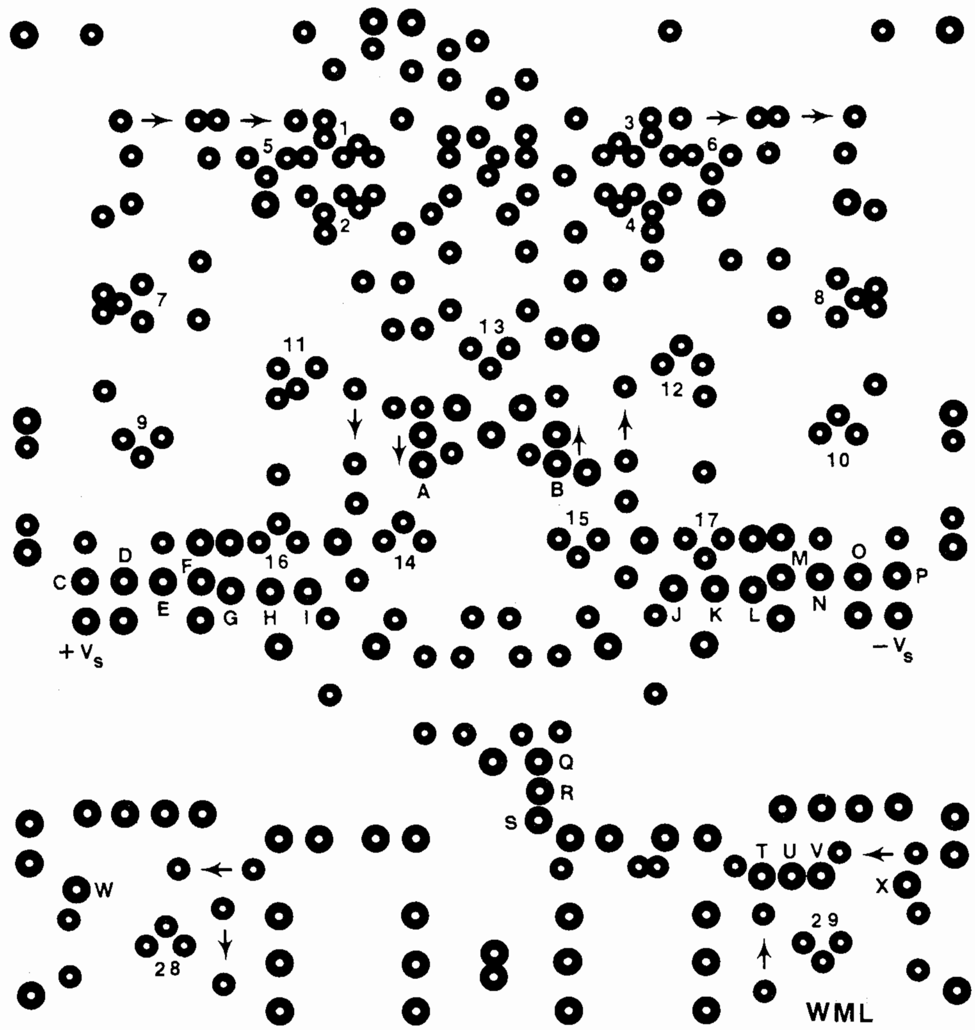


Fig. 12B.—Recommended circuit-board foil pattern for the ground-plane or component side.

tip is recommended, and the Weller PTM7 tip is an excellent choice. Only a high-quality solder such as Ersin SN62 multi-core No. 22 solder should be used. If difficulty is experienced in making the solder flow, use a pencil eraser to clean the area to be soldered and dip the tip of the solder into zinc chloride soldering paste before applying heat.

The next step is to mount and solder all ¼-watt resistors, with the exception of R1, R58 and R59, to the board. All of these should be bent on a resistor bender for a 0.4-inch hole spacing, except for resistors R4, R39 and R40. These should be bent to 0.5 inch for R4 and 1.4 inch for R39 and R40. A 0.5-inch length of insulation should be inserted over each lead of the latter two resistors before bending them. When inserting the resistors into the board, extra care should be taken to prevent peeling off a metal burr from the solder-plated resistor leads which could short to the ground plane. If this happens, the burr should be removed with a small blade screwdriver. Also, do not press the resistors down too hard against the ground plane, for their insulated coating could be broken and a short circuit could result.

With the exception of D3 through D6, the diodes should be installed next. They should be bent for a hole spacing of 0.4 inch and their proper direction is indicated by an arrow in Fig. 12B. The diodes should be installed such that the arrow points toward the banded end or cathode. All other resistors and capacitors should now be inserted. Each one-watt resistor should be bent for a 0.7-inch hole spacing. Pay particular attention to the polarity of the electrolytic capacitors. The negative terminals for these should be inserted into the holes labeled with a minus sign in Fig. 12B.

Next install all transistors which mount on the circuit board. The location of these transistors is labeled in Fig. 12A with the transistor number. Transistors Q1 through Q8 and Q13 through Q15 should be installed with their flat ends oriented as shown in Fig. 13. It is not necessary to bend the leads of these nine transistors to conform to the circuit-board hole spacings; simply spring the leads apart slightly and inser them until the transistors are about 5/16 inch off the board. All TO-5 case, i.e. metal can, transistors that mount on the board should be installed so that they are about this same

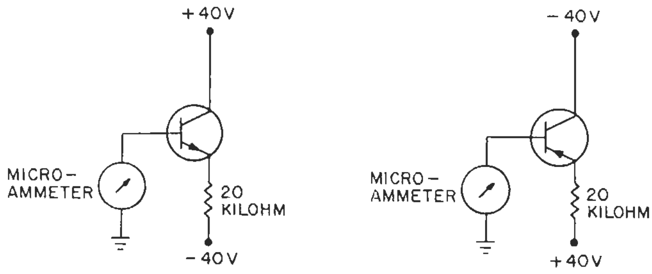


Fig. 14 — Test circuits for matching transistors Q1 through Q4. Select transistors for equal or nearly equal base currents. (A) Test circuit for matching NPN transistors Q1 and Q2; (B) test circuit for matching PNP transistors Q3 and Q4.

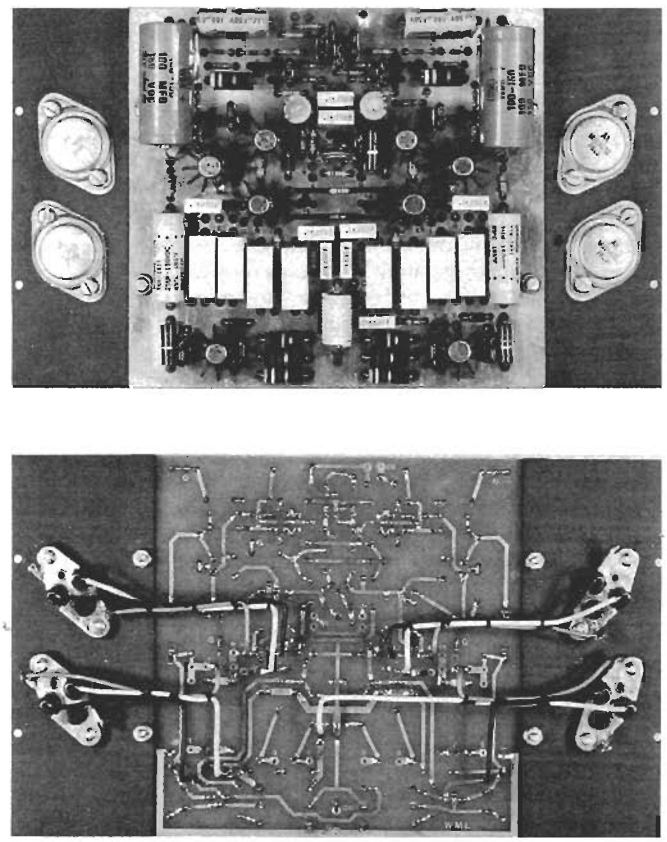
are not predrilled, each must be drilled for four TO-3 transistors. To mark the holes, arrange four mica insulating wafers in each heat sink channel in the position of the transistors and use a pencil to mark the position of the four holes on each wafer. After tapping the marked positions with a sharp pointed punch, drill all holes with a ¼-inch bit on a drill press. *Do not use a hand drill.* The holes must mate the transistors properly if short circuits to the heat sinks are to be avoided. In addition, each hole must be deburred to avoid a short circuit that could be caused by a punctured mica wafer. After the output transistor holes are drilled, the heat sinks can be drilled for the bias regulator diodes. Two of these diodes mount in holes drilled in the center of each heat sink. Drill the holes about ⅜ inch apart with a diameter that permits the diodes to fit snugly without binding. Finally, the holes for mounting the circuit boards to the heat sinks and any holes necessary for mounting the heat sinks to the chassis must be drilled. The circuit-board mounting holes should be marked and drilled for No. 4 screws (No. 33 drill bit) in the heat-sink outer flanges to mate with the holes on he outer edges of the driver transistor heat sinks. These should be marked carefully to ensure proper fit of the circuit-board assembly on the heat sinks. The method of mounting the heat sinks to the chassis may vary; I used bolts through the existing U-shaped cut-outs in the outer heat-sink flanges to attach aluminum brackets designed to hold the heat sinks ½ inch off the chassis bottom.

The output transistors can now be installed on the heat sinks. Use only high-quality molded plastic sockets. Because the heat sinks are so thick, it may be difficult to get a good connection between the socket pins and the base and emitter transistor leads. To solve this problem, I removed the pins from the transistor sockets and soldered them directly to the ends of the transistor leads. Care must be taken in positioning the pins on the transistor leads so that they will properly mate with the socket when installed on the heat sinks. Although this increases the difficulty of replacing an output transistor in the event of a failure, it does ensure good electrical connections for the high-current output transistor leads. The three wires to be soldered to the output transistor sockets should be color-coded No. 20 stranded wire about eight inches long. Next, install the transistors and sockets on the heat sinks. Each transistor should be insulated from the heat sink with a mica wafer that has been liberally coated on both sides with thermal compound. Transistors Q20, Q21, Q24, and Q25 mount on the heat sink that is to be closest to drivers Q18 and Q30, while Q22, Q23, Q26, and Q27 mount on the heat sink that will be closest to drivers Q19 and Q31. Use an ohmmeter to verify that no transistor lead is shorted to the heat sinks.

The diode bias assemblies are now installed into each heat nk and should be prepared as shown in Fig. 16. It is best to irst solder the wire that is used to connect each diode to the circuit board, and these wires should be color-coded No. 22 stranded wire about eight inches in length. Before soldering, tie-wrap the joints with a single strand from No. 22 stranded wire to form a tight mechanical connection. After installation into the heat sinks, insulate the solder joints with heat-shrink tubing, preferably shrunk with a heat gun. Take extreme care to avoid cracking any of the bias diodes during installation into the heat sinks. If a crack occurs, the diode will become an open circuit and the amplifier can be seriously damaged. Thus the diodes should not be forced into their mounting holes nor should there be tension on their leads. It is not necessary to use heat-sink compound on the diodes; if necessary, a drop of instant bonding glue on each may be used to bond them to the heat sinks.

Before attaching the circuit board to the heat sink, it is best to pretest it with a lab power supply. This can be done with dual 50-V supplies, preferably with a current-limit circuit. On the rear of the circuit board, temporarily tack-solder four 100-ohm 1/4-watt resistors, one from the emitter of Q30 to the collector of Q18, one from the emitter of Q31 to the collector

Fig. 15 — Photographs of each side of an assembled circuit board.



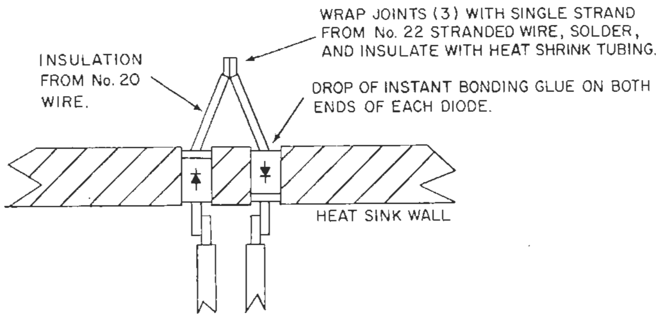


Fig. 16 — Recommended mounting of bias diodes in each heat sink.

of Q19, one from the emitter of Q18 to the junction of resistors R43 through R46, and one from the emitter of Q19 to this same junction. Solder a short-circuit jumper wire across C14. With clip leads, connect the negative output of one power supply to the circuit-board ground wire and the positive output to the circuit-board positive power-supply input. Connect the positive output of the other power supply to the circuit-board ground wire and the negative output to the circuit-board negative power-supply input. With the power supplies set to current limit at 100 mA, the voltages can be slowly turned up. If current limiting does not occur or if the current drawn does not exceed about 50 mA, turn the power supplies off, use clip leads to connect the output of a signal generator to the signal-input cable, and connect an oscilloscope probe to the loudspeaker output wire. The oscilloscope ground should be connected to the circuit-board ground wire. With a sine wave of 1 V rms at 1 kHz applied to the input, slowly turn up the power supply voltages. A clipped sine wave that looks like a square wave should appear on the oscilloscope before reaching plus and minus 10 V. At 50 V, the sine wave will appear unclipped with an amplitude of about 60 V peak-to-peak. Turn the power-supply voltages down and connect a 10-kHz square wave source to the circuit-board input. With power supplied to the board, the output signal should look like the waveform of Fig. 8. If it does not, check the values of R3, R4, C1, C2, and C3.

If the previous tests are negative, the circuit board should be inspected thoroughly. Check for solder bridges; cold solder joints; shorted components to the ground plane; backward electrolytic capacitors, transistors, and diodes; incorrect component values; unsoldered ground connections, etc. Correct all errors at this point — before the output transistors are connected. If components must be removed from the circuit board, desolder them with a desoldering braid (such as Solder Wick) and with as little heat as possible to prevent lifting the copper foil from the circuit board. When the circuit is operational, remove the jumper wire and four resistors from the rear of the board. The solder flux must now be removed from the boards with a soft-bristle brush and a good solvent (such as Stripper brand) spray-on circuit-board cleaner. When cleaned, each solder joint will be shiny; dull joints may be cold solder joints so make sure they are touched up.

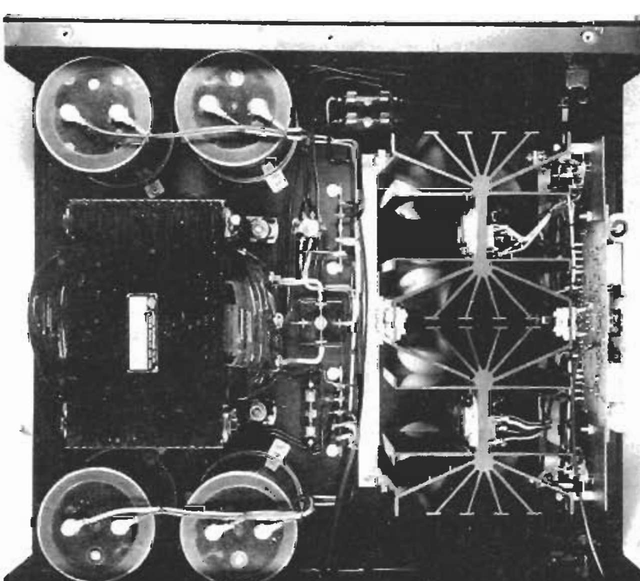
The circuit board is now ready to be mounted to the heat sinks. The wires from the power transistors and bias diodes should first be cut to length and soldered to the circuit board. These are connected from the circuit side of the board, i.e. the side opposite the components. To do this, lay the board in front of the heat-sink assembly with the ground plane side down and the loudspeaker output connection nearest the heat sinks. Cut the wires from the heat sinks just long enough so that after they are soldered to the board there will be no tension on them with the board in this position. If this is done properly, you will not have to desolder any wires from the board should it need servicing. However, to minimize undesirable coupling effects, the connecting wires should be no longer than necessary.

The proper connections of the wires from the heat sinks to the circuit board are given by the codes in Table I. After soldering these wires to the circuit board, use cable ties to separately tie the wires from each transistor and bias diode assembly into neat bundles. Carefully inspect the ground plane side of the board for any soldered wires that may protrude out too far and cause a short circuit. No wire should protrude through the board by more than 1/32 inch. When all wires are connected, double-check the wiring for errors. The bias diodes especially should be checked for correct connection to the board, for backward diodes can damage the output transistors when the amplifier is first powered. The circuit board can now be attached to the heat sinks. This is done with four No. 4-40 x ¾-inch threaded aluminum spacers and eight No. 4-40 x ⅜-inch screws with an inside star lockwasher under each. Neatly route the cabling between the heat sinks and circuit board after the circuit board is installed. This completes the circuit-board and heat-sink assembly.

Wiring of the Chassis

The amplifier can be built on any chassis, provided the general layout and wiring used by the author are followed. The chassis shown was bent from 16-gauge steel by a sheet-

Fig. 17 — Interior view of the amplifier chassis showing chassis wiring.



metal shop. Its dimensions are 17 inches W x 15 inches D x 6¾ inches H. A 7-inch x 19-inch steel panel is bolted to the front, and a perforated steel cover is placed over the chassis to make it mechanically rigid. The cover is screwed into No. 6-32 nutserts installed in the chassis outer flanges. An interior view of the chassis component placement and wiring is shown in Fig. 17. An aluminum chassis can be used, but it is difficult to paint and may not be strong enough to support the 38-pound transformer.

Before painting the chassis, punch and drill all holes, smooth sharp edges with a file, and remove all corrosion with a wire brush, sandpaper, and steel wool. For proper ventilation, vent the area beneath the heat sinks by punching eight 1⅜-inch circular holes under the heat-sink assembly with a chassis punch. After the chassis is completed, thoroughly wash it with a spray-and-wipe all-purpose cleaner to remove oil traces. Next, prime and paint the chassis (I used Rust-Oleum 960 spray primer and Rust-Oleum 7278 satin black spray paint). For best results, apply the paint in light duster coats and allow 30 to 60 seconds for the gloss to disappear between each spray. This will produce a professional-looking textured finish without runs. After one coat is built up, wait at least 48 hours before applying a second. This is important if the finish is to harden properly and not wrinkle.

The signal input jack must be insulated from the chassis, and a phono jack which mounts in a circular hole with a single nut from the back is recommended for this. It can be insulated from the chassis with a flat washer on one side and a shoulder washer on the other. The recommended loudspeaker output connector is a double five-way binding post. If two single five-way binding posts are used, drill the chassis for the standard ¾-inch spacing between them.

Before placing the circuit-board and heat-sink assembly into the chassis, install the power supply and associated a.c. chassis wiring. The power supply is wired as shown in Fig. 2. A single central chassis ground point is used, to which all high-current leads are connected. This central ground point should be a No. 10 machine screw on which are nine No. 10 soldering lugs separated by No. 10 flat washers. A No. 10 inside star lockwasher is installed under the bottom washer to ensure that the ground point makes good connection with the chassis ground through the paint. Firmly tighten the screw so that these lockwashers engage properly. The green a.c. safety ground from the power cord, the filter capacitor grounds, and the two transformer center taps are then connected to the ground point. The circuit board ground wire and the loudspeaker output ground terminal are also connected to this point after the circuit-board and heat-sink assembly has been installed. The latter connection should be made with a length of No. 16 stranded wire. It is important to carefully solder all the high-current leads in the power supply. Also, the the central ground point must make good electrical connection to the chassis. Before mounting the transformer, clean all paint and corrosion from its terminals, coat them lightly with zinc chloride soldering paste, and install a ¼-inch screw-size grommet into each hole in the transformer mounting brackets. The transformer is installed with four ¼-20 x ¾-inch bolts with two No. 10 flat washers between each grommet and the chassis — one washer between each grommet and nut, and one washer under the head of each bolt. Solder separate lengths of No. 16 stranded wire from the two secondary terminals labeled 57½ V to the central ground point. Next, solder separate lengths of No. 16 stranded wire from the two terminals labeled 0 V to one a.c. input on the bridge rectifier. Similarly, solder No. 16 wire from the terminals labeled 115 V to the other a.c. input on the bridge rectifier. Solder capacitor C29 directly across the two a.c. inputs on the rectifier. Wire the d.c. outputs from the bridge rectifier to a terminal strip from which separate lengths of No. 16 stranded wire are connected to each filter capacitor and to the d.c. power-supply fuse clips. The three wires on the a.c. power cord must be connected properly to minimize shock hazard. First, coat each lead with zinc chloride soldering paste and solder the leads to separate terminals on a terminal strip. Solder a length of No. 18 stranded wire from the terminal with the green or safety ground wire to the central ground point. Next, solder a length of No. 16 stranded wire from the white or neutral wire to the 0-V transformer primary tap. The black or "hot" power-cord wire should be connected with lengths of No. 16 stranded wire through the a.c. fuse clip and the a.c. power switch to the 115-V transformer primary tap. The pilot light is connected in parallel with the transformer primary.

After the power supply has been wired, the heat-sink and circuit-board assembly can be mounted on the chassis and connected to the power-supply fuse clips, the central ground point, and the loudspeaker output terminal. The No. 16 and 22 stranded wires for these connections should already have been soldered to the circuit board. Before soldering leads to the loudspeaker output terminals, install the series combination of resistor R57 and capacitor C26 directly across the ter-

Table I — Codes for external connections to the circuit board.

Circuit Board Label	Connected To
1	Emitter Q26
2	Emitter Q27
3	Emitter Q22
4	Emitter Q23
5	Emitter Q20
6	Emitter Q21
7	Emitter Q24
8	Emitter Q25
9	Collector Q22
10	Collector Q19
11	Base Q19
12	Base Q18
13	Collector Q18
14	Collector Q20
A	D3-D4
B	D5-D6
C	Collector Q24
D	Collector Q25
E	Collector Q30
F	Collector Q21
G	Base Q20
H	Base Q21
I	Emitter Q18
J	Emitter Q19
K	Base Q22
L	Base Q23
M	Collector Q23
N	Collector Q31
O	Collector Q26
P	Collector Q27
Q	Emitter Q30
R	Base Q24
S	Base Q25
T	Base Q26
U	Base Q27
V	Emitter Q31
W	Base Q30
X	Base Q31

Numbers are etched on the copper foil side of the circuit board and letters on the ground plane side.

minals and connect one end of resistor R59 to the loudspeaker ground terminal. Solder the other end of this resistor to a nearby grounding solder lug. The loudspeaker output wire from the circuit board and the speaker ground wire from the central ground point can now be connected and all connections soldered. Next, solder the free end of resistor R58 to a grounding solder lug that connects to the chassis ground near the circuit board ground (to which this resistor connects). The final step is to connect the input leads to the phono input jack. Solder resistor R1 between the insulated phono jack ground and a nearby grounding solder lug.

Proper grounding is a must if ground loop problems are to be avoided. To minimize these problems, separate chassis and signal grounds were used. The circuit-board ground plane was insulated from the driver transistor heat sinks so that there is only one direct connection between the circuit-board ground and the central ground point—through the wire soldered to the ground connection nearest the loudspeaker output connection on the circuit board. There are three small value resistors that complete the grounding — R1, R58, and R59. Resistor R1 solders between the insulated ground terminal on the phono input jack to a solder lug connected to the chassis ground adjacent to the jack. Resistor R59 solders between the loudspeaker output ground terminal and a solder lug connected to the chassis ground adjacent to the loudspeaker output terminal. Resistor R58 solders between the circuit-board ground closest to the loudspeaker output connection to a nearby grounding solder lug. Although complicated, this grounding procedure follows the practices recommended in [16] and should ensure freedom from oscillations and hum that can be caused by mutual coupling in the ground system.

These steps complete the amplifier assembly.

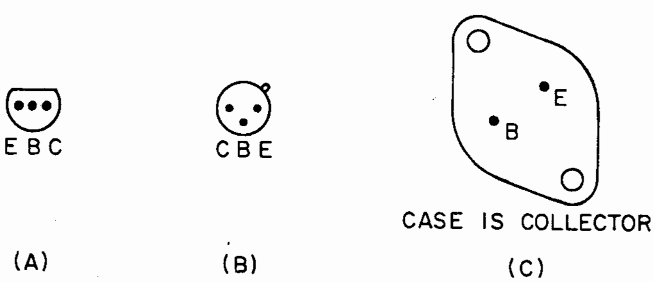


Fig. 18 — Transistor pin connections. Views are from the pin side of each transistor. (A) Small signal plastic case transistors; (B) TO-5 metal case transistors, and (C) TO-3 metal case transistors.

Turn-On Procedure

Before attempting to apply power to the amplifier, it is strongly recommended that the entire unit be checked for errors. If everything appears to be correct, the initial tests can be performed. First, install the a.c. power fuse; do not install the d.c. power fuses to the circuit board at this point. Plug the a.c. power cord into a Variac autotransformer, turn the amplifier on, and slowly increase the a.c. voltage until a d.c. voltage of about 5 V is measured with a voltmeter across the d.c. outputs of the bridge rectifier. The polarity of the voltages on each filter capacitor should now be checked with the voltmeter before increasing the voltage further. If the voltage polarity is incorrect on any capacitor, it will short-circuit if the voltage is increased. If all filter capacitors are polarized correctly, the a.c. voltage can be increased to 120 V. Do not increase the Variac above that value. The positive and negative d.c. power supplies should read within one or two volts of 85 V. The a.c. power can now be removed. Before further tests, discharge the filter capacitors with care, since approximately 110 joules of energy will be stored in them. It is best to do this by connecting two leads to a 16-ohm dummy speaker-load box and then touching the leads across the terminals of each capacitor for several seconds. If a dummy load box is not available, a 3.9-kilohm, 2-watt resistor can be used although the discharge will take several minutes. Do not hold a resistor by its body across the capacitor terminals to discharge them, for the resistor can become hot enough to burn the fingers.

In the next test, d.c. power is applied to the circuit board. Before proceeding, adjust potentiometer R28 for maximum resistance, i.e. 5 kilohms. This should be verified with an ohmmeter, for otherwise the power transistors could be damaged in the next test. Install a 10-ohm, 2-watt resistor with clip leads across each d.c. fuse clip to prevent the release of excessive current into the circuit if there is an error. If everything is normal, the current through the resistors in the next test will be less than 50 mA and the voltage drop across them negligible. Apply a 1-kHz sine-wave signal of amplitude 1 V rms to the amplifier input. With an oscilloscope, check the signal at the phono input jack to ensure there are no short circuits in the input cable from the jack to the circuit board. If there are no shorts, connect the oscilloscope to the loudspeaker output terminals. With the a.c. power switch on, slowly increase the a.c. voltage with the Variac until the output signal is observed. The signal will initially appear as a clipped sine wave that looks like a square wave. If no problems are encountered, the a.c. voltage can be slowly increased to 120 V. The amplitude of the sine wave output signal should be about 60 V peak-to-peak. The input signal should now be removed and the d.c. output voltage from the amplifier measured. This should be less than 100 mV, although a slightly higher value is acceptable. The magnitude of the d.c. offset depends on the matching of the transistors in the differential amplifiers, i.e. Q1 through Q4.

If the preceding tests are successful, the bias potentiometer R28 can be adjusted next. Remove the a.c. power and discharge the power-supply capacitors. Install a fuse in the negative power-supply fuse clip and a d.c. milliammeter across the positive power-supply fuse clip. With no input signal and no load connected to the loudspeaker output terminals, slowly turn the Variac up until the a.c. input voltage is 120 V. Adjust R28 until the milliammeter reads 250 mA. This will bias the amplifier so that it operates Class-A for a peak power level of 2 W or less with an 8-ohm load, thus eliminating all traces of crossover distortion. Let the amplifier idle for at least 30 minutes while readjusting R28 for 250 mA every few minutes. When the current stabilizes to this value, the heat sinks will be warm to the touch.

The amplifier is now ready to be used. Turn it off and let the power supply discharge. When the milliammeter reading drops to zero, remove it from the fuse clip and install the positive power-supply fuse. The amplifier will have no turn-on thump if it is switched on after the power supply has completely discharged. This takes several minutes after each turn-off. To prevent turn-on thumps from the preamplifier from reaching the loudspeaker, *always turn the preamplifier on for several seconds before turning on the power amplifier, and always turn the power amplifier off and wait until the sound stops before turning off the preamplifier.* This procedure should always be used with d.c.-coupled amplifiers to prevent d.c. transients generated in the preamplifier from reaching the loudspeakers.

Concluding Comments

The basic circuit topology of the amplifier was originally ɔnceived in 1975. With the enormous tasks of completing ıe circuit design details, translating them into a finished amplifier, and describing its construction for this article accomplished, the task of building a second amplifier to complete a stereo system will begin.

The original intent was to build a single stereo amplifier rather than two mono units. However, I recognized that most audiophiles who build the amplifier would probably prefer the two channels with separate power supplies. Thus, weight and size limitations dictated a separate chassis for each channel. A single chassis and power supply can be used if desired, although a special grounding procedure must be observed to prevent hum caused by ground loop problems. The two phono input jacks should be installed adjacent to each other. Connect their insulated grounds together and use a single resistor R1 from this connection to a nearby grounding solder lug. Omit the No. 22 ground wire that connects from the circuit-board ground of one channel back to the central ground. This circuit board will then be grounded back through the phono input cable grounds, the circuit-board ground plane of the other channel, and the No. 22 wire which connects that channel to the central ground point. Resistors R58 and R59 should be installed on both channels as described previously.

The problem of obtaining parts is not a trivial one. The transistors and diodes cannot be bought from consumer electronics stores or those that supply TV repair shops; they can only be ordered through industrial electronics suppliers. *Under no conditions should substitute transistors or diodes be used*, for the amplifier is frequency-compensated only for the types and manufacturers specified in the parts list. Many ırts, such as heat sinks and filter capacitors, can be obtained ɔm money-saving mail-order electronics surplus companies. There are also many mail-order firms that sell high-quality non-surplus items such as carbon film resistors, small electrolytic capacitors, small signal transistors, zener diodes, chassis hardware, etc. at bargain prices. The addresses of these companies can be found in the back pages of many consumer electronics magazines. (A list of sources for parts will be sent with all circuit board orders from the supplier listed in the accompanying parts list.)

The power transformer is the single most expensive item in the amplifier. If two surplus 230- to 115-V transformers can be found, their primaries can be connected in parallel and secondaries in series to form a 115- to 115-V transformer. If the transformers are not connected properly, the series secondaries will be out of phase and the voltages will cancel. Therefore, an a.c. voltmeter should be used to verify that the secondaries are phased properly. The current rating of the 115-V windings should be no less than 10 A. The specified transformer has a current rating of 12 A with its two 6-A secondaries connected in parallel for a 115 to 115-V rating. One measure of transformer quality is the weight. The specified one weighs 38 pounds, which should be comparable to the total weight of any substitute transformers. Under no conditions should more than 85 V be used on the amplifier power supply. With substitute transformers, it is sometimes possible to remove several turns from the secondary to reduce the output voltage in case it is too high. If the line voltage runs higher than 120 V a.c. with the specified transformer, the 125-V primary tap should be used. If a 230-W amplifier is sufficient, the use of the 125-V tap is highly commended. This will ensure protection of the amplifier in case the a.c. line voltage increases due to poor regulation by the power company. With a line voltage of 120 V a.c., the power supply voltages will be plus and minus 80 V.

Removal of the Bessel input filter to increase the amplifier bandwidth beyond 40 kHz is not recommended, for the frequency compensation will be affected. Capacitor C1 in this filter affects the high-frequency bandwidth of the input stage for signals applied to the feedback input. Its removal may cause oscillations. Also, because the amplifier slew rate is so high, the filter protects the amplifier and its loudspeaker load from accidental or unintentional ultrasonic input signals. Some advertisements have misled audiophiles to believe that very wide bandwidths are required for negligible phase distortion. Thus, some builders may be tempted to remove the filter to extend the amplifier bandwidth to 500 kHz. Do not do this, for the circuit still has a bandwidth of 500 kHz with the filter. The filter simply prevents signals with a frequency higher than 40 kHz from entering the circuit. The phase response of the Bessel filter has been carefully researched [3]. The filter phase response is such that it introduces a constant time delay in the audio band of 5.4 μS. This delay is equivalent to the listener moving 1/16 inch back from the loudspeaker. Further, it is minuscule compared to the time delay between an original recording session and any playback of the recording. As a point of interest, the nice phase characteristics of a Bessel low-pass filter are not shared by a Bessel high-pass filter. Thus, unfortunately, the low-frequency phase response of loudspeakers cannot be successfully linearized with the Bessel alignment.

As stated previously, removal of the limiter circuit is not recommended. Its threshold is high enough so that it would not be triggered by normal loudspeaker loads. Also, the limiter cricuit will not respond to fast transients so that it will not limit the slew rate of the amplifier with normal loads. Without the limiter, an accidental short circuit on the loudspeaker output can severely damage the amplifier, and I have heard from several builders of the low TIM amplifier [4], [5] who related their unfortunate experiences. In one case, the builder lost his amplifier when he sat on the bed; investigation showed that the loudspeaker wire ran under the carpet and beneath the bedpost.

Finally, there are some precautions which should be observed in testing the amplifier. Never attempt to apply continuous large-amplitude high-frequency sine waves or square waves to the input. Not only could the output transistors be damaged, but resistor R57 will surely be fried. This resistor must be noninductive, and thus a carbon composition resistor is specified. The highest wattage standard carbon-composition resistor is 2 W. If the amplifier is tested at 20 kHz with a full power sine wave, it will dissipate 3 W. With a square wave, it would be about double this figure. If the builder anticipates that the amplifier will be subject to continuous sine or square wave tests that could fry R57, it is recommended that four 39-ohm 2-W resistors be connected in parallel to form this resistor, which will increase its power rating to 8 W. During any tests, occasionally touch these resistors to check that they are not so hot that they could be damaged. With normal audio signals, R57 is safe at 2 W. If R57 is unknowingly fried during a test, the stability of the amplifier with reactive loads could be affected. This means that it may oscillate with reactive loads and possibly damage the output transistors.

For those readers who wish to ask questions or make comments about this construction article, I ask that you please not use the telephone. I must politely refuse all calls made to my office at Georgia Tech that pertain to this type of project. Instead, mail letters to me and enclose a self-addressed stamped envelope for return; I will make every effort to answer them. I sincerely hope that the ideas I have expressed in this article will spark the imagination of readers, and I wish the best of luck to those who build the amplifier. The best advice I can offer is to take your time! It has taken me four years to build my amplifier, and I estimate that anyone who takes less than four months to acquire the parts and build it will do a sloppy job. Happy listening!

\ddendum

Although not indicated on the circuit diagram, parts list, or parts layout, the protection circuit should be modified by the addition of two diodes and two resistors. The diodes should be type 1N4934. The resistors should be 680-ohm ¼ W 5 percent. Solder one diode and one resistor in parallel across the leads of capacitor C17 and the other diode and resistor in parallel across the leads of capacitor C18. The banded end or cathode of the diode which connects in parallel with C17 should be the end that connects back to the center lead or base of transistor Q14. The unbanded end or anode of the diode which connects in parallel with C18 should be the end that connects back to the center lead or base of transistor Q15. If the diodes are installed backward, the operation of the protection circuit will be seriously affected.

Parts List

All resistors are ¼-W 5 percent carbon film unless otherwise specified. Manufacturer codes for active devices: M—Motorola, R—RCA, I—ITT, N—National Semiconductor. *Do not substitute transistor or diode types or manufacturers.*

R1—4.7 ohm.
R2—43 kilohm.
R3, R4—6.2 kilohm.
R5 through R12 and R29, R30—270 ohm.
R13, R14—10 kilohm.
R15, R16—1.2 kilohm.
R17, R18—4.3 kilohm, 1 W.
R19, R40—240 ohm.
R20, R52, R53, R54, R55—5.6 kilohm, 1 W.
R21—56 kilohm.
R22, R23—20 ohm.
R24, R25—330 ohm.
R26, R27—1 kilohm.
R28—5 kilohm trimpot (CTS X201R).
R31, R32—3.9 kilohm, 1 W.
R33, R34, R35, R36—360 ohm.
R37, R38—100 ohm.
R39—470 ohm.
41 through R48—0.22 ohm, 5 W.
R49, R51—15 kilohm, 1 W.
R50—39 ohm.
R56—10 ohm, 2 W.
R57—10 ohm, 2 W or four 39-ohm, 2 W resistors in parallel (see text).
R58—3.3 ohm.
R59—1 ohm, ½ W.
C1—430 pF silver mica (Arco DM15-431J).
C2, C3, C24, C25—270 pF silver mica (Arco DM15-271J).
C4, C5—15 pF silver mica (Arco DM15-150J).
C6, C7—330 μF, 10-V radial electrolytic (Panasonic 330/10R).
C8, C14, C17, C18, C26, C27, C28—0.1 μF, 100- or 250-V metallized polyester (Plessey Minibox 0.1/100 or 0.1/250, C or D box).
C9—15 pF silver mica (Arco DM15-150J).
C10, C11—100 μF, 50-V axial electrolytic (Panasonic 100/50A).
C12, C13—10 pF silver mica (Arco DM15-100J).
C15, C16, C23—0.01 μF, 400- or 630-V metallized polyester (Plessey Minibox 0.01/400 or 0.01/630, B or C box).
C19, C20—100 μF, 100-V axial electrolytic (Mallory TC10101B).
C21, C22—25 μF, 100-V axial electrolytic (Mallory TC10250C).
C29—0.1 μF, 250-V metallized polyester (Panasonic 104K).
C30—0.01 μF, 630-V metallized polyester (Plessey Minibox 0.01/630, C box).
C31, C32, C33, C34—8,600 μF, 100-V electrolytic (Mallory CG832U100G1).
L1—10 turns No. 22 solid insulated wire wound tightly around R56 and soldered to the resistor leads.
D1, D2—Two 1N5250B 20-V zener diodes in series for each.
D3, D4, D5, D6 — 1N4001, 1N4002, 1N4003, or 1N4004.
D7, D8, D9, D10— 1N4934 (M) fast recovery rectifier.
D11, D12 — 1N4935 (M) fast recovery rectifier.
Rect 1 — MDA3504 (M) 400-V, 35-A bridge rectifier.
Q1, Q2, Q5, Q8, Q13, Q14 — MPS8099 (M) or 2N5210 (M, I, or N).
Q3, Q4, Q6, Q7, Q15 — MPS8599 (M) or 2N5087 (M, I, or N).
Q9, Q11, Q17, Q29 — MM5415 (M) or 2N5415 (R).
Q10, Q12, Q16, Q28 — 2N3439 (M or R).
Q18, Q30 — MJ15001 (M).
Q19, Q31 — MJ15002 (M).
Q20, Q21, Q24, Q25 — MJ15003 (M).
Q22, Q23, Q26, Q27 — MJ15004 (M).
T1 — Signal transformer 230-6, 115V C.T., 12A (Signal Transformer, 500 Bayview Ave., Inwood, N.Y. 11696).
F1, F2 — 8-A 3AG fast blow fuse (Littlefuse 312008).
F3 — 6.25-A, 3AG slow blow fuse (Littlefuse 3136.25).
S1 — 15-A 125 VAC two-pole single-throw circuit breaker (Airpax T21-2-15.0A-01031) with both poles wired in parallel, punch chassis for ½-in. hole.
One neon pilot lamp (Drake 6063-001-634), drill chassis for ⅜-in. hole.
One etched, plated, and drilled ground plane circuit board (see text).
Two heat sinks to mount on circuit boards for Q18, Q19, Q30, Q31 (see text).
One three-conductor UL-type SJ No. 16 a.c. line cord (Alpha 618) and line cord strain relief (Heyco SR-6P3-4), punch chassis for ⅝-in. hole.
One phono chassis jack (Switchcraft 3505F) insulated from chassis with ⅜-in. I.D. nylon shoulder washer on one side and flat nylon washer on other, punch chassis for ½-in. hole.
One double five-way binding post (Superior Electric DF30-2-8RC) for loudspeaker output terminals.
Three 3AG fuse mounting holders (Littlefuse 357001).
Four 4-40 x ¾ in. x ¼ in. O.D. aluminum threaded round spacers (Waldom 60477) for mounting circuit board to heat sinks.
Twelve TO-3 transistor sockets, molded types preferable.
Twelve TO-3 mica insulating wafers.
Eight TO-5 transistor finned heat-sink coolers (Wakefield 205CB).
Two Wakefield 441K transistor heat-sink coolers drilled for Q20 through Q27 and D3 through D6 (see text).
Misc. — Suitable heavy-duty steel chassis with cover; mounting brackets for circuit board and heat-sink assembly; silicone heat-sink compound; shielded phono cable; No. 16, 18, 20, and 22 stranded hook-up wire; No. 22 solid hook-up wire; five heavy duty ½-in. hard plastic feet; No. 6 and No. 10 soldering lugs; chassis hardware; rubber grommets for mounting transformer; etc.

An etched, drilled, and reflow solder-plated circuit board plus two mating heat sinks punched for the four driver transistors are available for $16 from Custom Components, P.O. Box 33193, Decatur, Ga. 30033. Shipping and handling charges for orders inside the continental USA are $1.00 plus five percent of the total price of each order. Georgia residents must add four-percent sales tax.

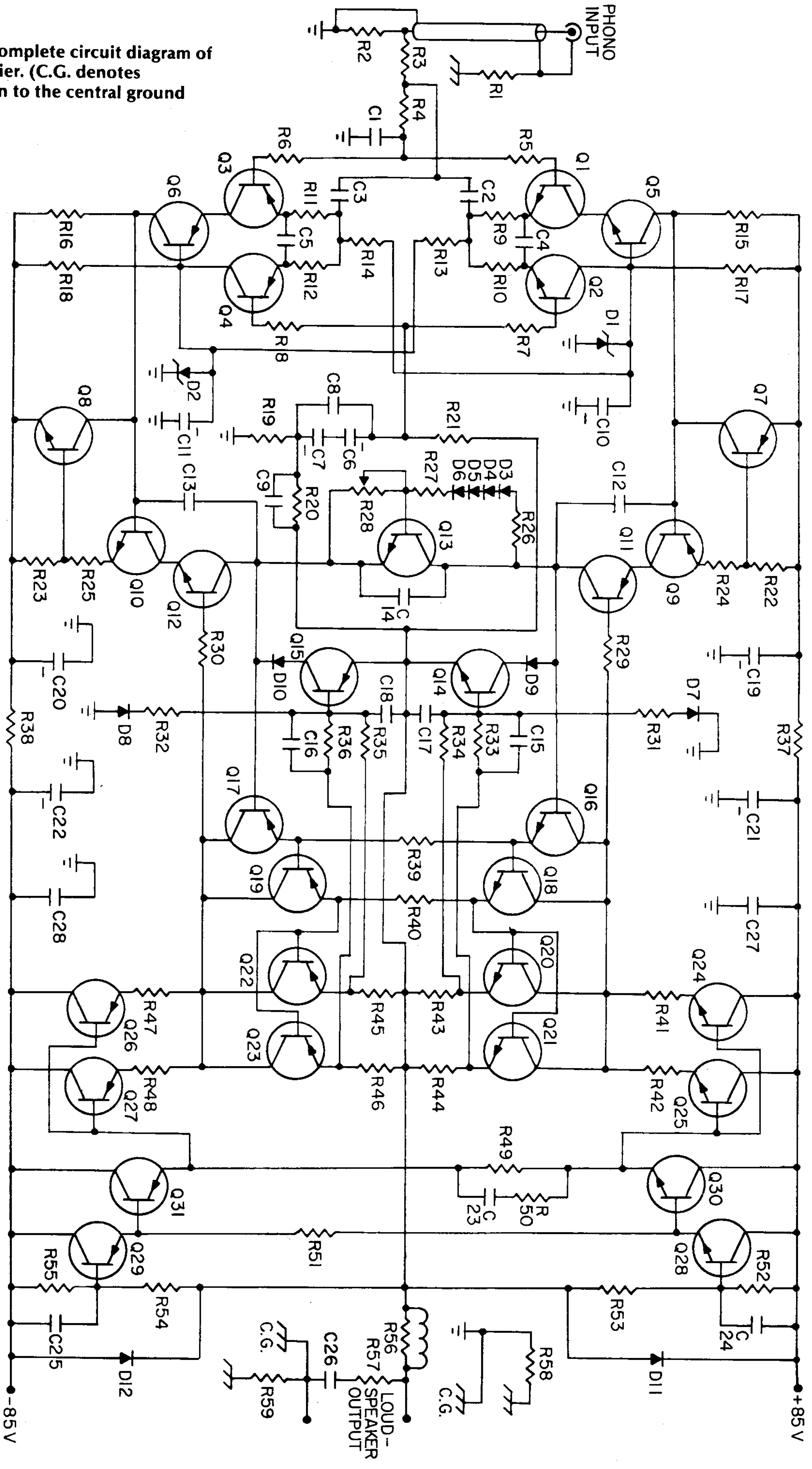


Fig. 1 — Complete circuit diagram of the amplifier. (C.G. denotes connection to the central ground point.)

MAY 1980 LEACH AMP

115V SEC. ON PWR XFMR.