A 60W MOSFET Power Amplifier

by Erno Borbely

IN THE LAST COUPLE OF YEARS I, like many of my friends and colleagues, have been trying to improve the weakest link in the audio chain—the loudspeaker. As I do not design drivers, I have to rely on what is available on the market; all I can do is utilize these drivers in the best possible way. This work usually involves testing large numbers of speakers, often using active crossovers, active drivers, etc. In other words, I need a lot of amplifiers.

To alleviate the problem I set out to design a medium-power amplifier for use in such systems. I aimed for a simple design, easy to make, failsafe in operation, and providing excellent performance under all operating conditions. I believe the design presented here fulfills all these requirements, and consequently it may be useful to you also.

USING MOSFETS

Over the last 20 years I have had experience with bipolar power devices as both a user and a manufacturer. During these years their power handling, speed, and other characteristics have steadily improved. However, the inherent problem with these devices is still there: the positive temperature coefficient of the collector current, increasing with increasing temperature, limits the maximum collector current by local current concentration and thermal runaway.

Of course, over the years people have learned to live with this problem and produce some very high quality amplifiers. For reliability the name of the game was to use the power devices more conservatively than the manufacturers claimed was necessary.

An entirely new power device emerged from the research labs in the mid 70's: the power Mosfet. Introduced by several companies, they were expensive and hard to come by. Hitachi produced the first reasonably-priced ones, which I read of in 1977.[1] My practical experience dates from 1978, when I joined the David Hafler Co. and started to develop the DH-200. I have been using the Hitachi 2SK134/2SJ49 power Mosfets ever since.

Power Mosfets have a number of advantages over bipolar devices. Most important, of course, is that the drain current has a negative temperature coefficient: it decreases with increasing temperature (see *Fig. 1*, transfer characteristics at three different temperatures).

This means first of all that power Mosfets have no thermal stability problem, so we don't have to temperature-compensate the bias voltage. Secondly, they have no local current concentration and consequently no safe operating area limitation. This translates into the practical advantage of not having to use elaborate protection circuits (V-I limiters) in our amplifiers.

Some other relevant advantages for audio use are high switching speed due

FIG. 1

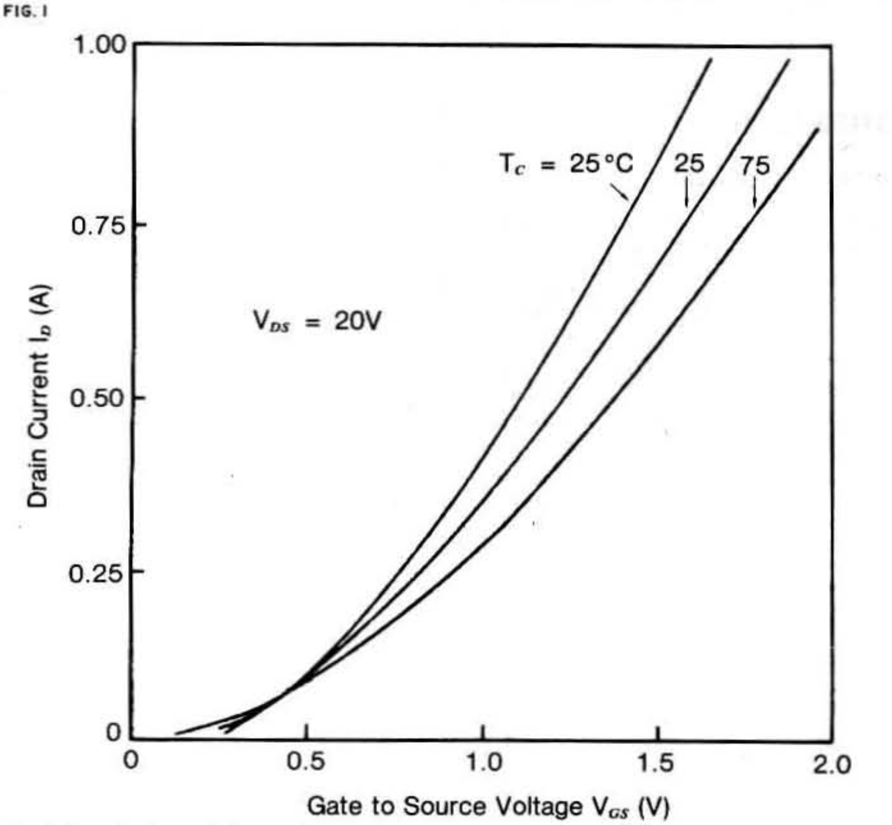

Fig. 1. Transfer characteristics at various temperatures.

to absence of minority carrier storage, good frequency response because of fast carrier speed, and high power gain because of high input impedance.

All this sounds very impressive, you may say; but surely there must be some disadvantages too, or everybody would switch over to these devices. Yes, there are but fortunately they are not serious. The major problem until recently was they were hard to obtain in small quantities for amateur use. However, they are now available through even small distributors at reasonable prices; I assume Old Colony will also carry them.

One technical problem is the relatively high on-resistance—approximately 1Ω—of the 2SK134/2SJ49 devices. If you drive 5A through them, you get a 5V voltage drop, which obviously increases power dissipation and reduces efficiency. Bipolars have a much lower saturation voltage, even at higher currents.

Another limitation is the 7A maximum drain current for the 2SK134/2SJ49's. This may sound small compared to bipolars, but when you start designing your amplifier you will find this is not the limiting factor in terms of output power. With proper heatsinks you can get over 70W into 8Ω and close to 100W into 4Ω from one pair of Mosfets. For higher power you have to connect several in parallel. Alternatively, you could have higher current units, but I don't know whether Hitachi sells them in small quantities. In my application, one pair of these devices is more than adequate.

SOURCE-FOLLOWER OPERATION

The simplest way to use Hitachi's power Mosfets in an audio output stage is in source-follower mode (see *Fig. 2*), which is equivalent to the emitter-follower connection with bipolar transistors. *Figure 3* shows the frequency response for a source-follower, using an N-channel device[1]; for comparison the frequency response of an emitter follower is also shown. Note that the Mosfet's −3dB point is more than 10 times higher than that of an equivalent bipolar device.

FIG. 3

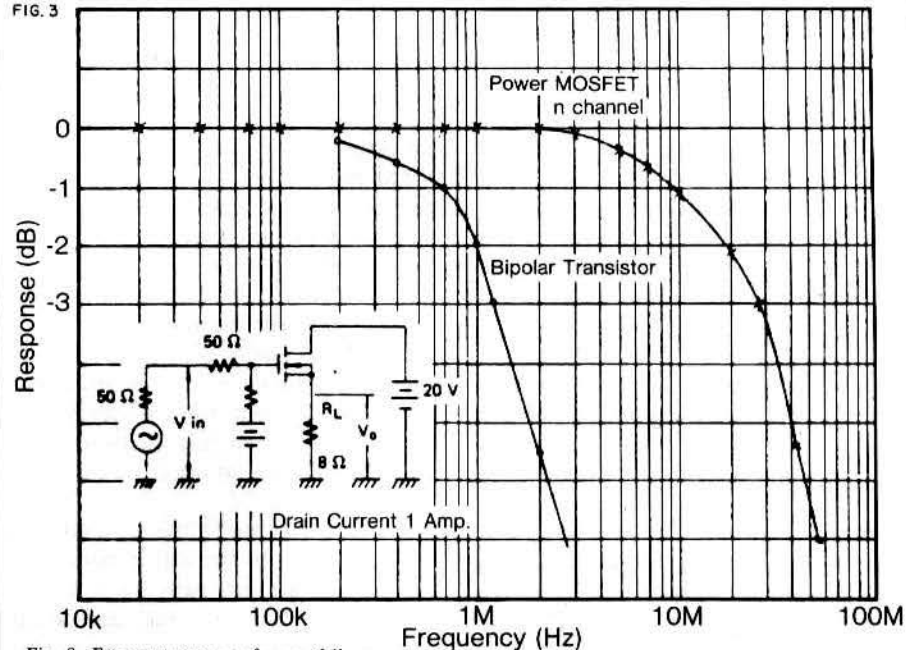

Fig. 3. Frequency response of source follower.

Bipolars are not used alone in power amplifiers with more than a few watts output, because of the high current needed to drive them. A Darlington connection is preferable because of the reduced drive requirements from the previous stage, which will of course further reduce the frequency response shown earlier.

We need no extra source follower to drive Mosfets since the input-impedance is very high. Unfortunately, however, this input-impedance is all capacitive and the capacitance has a fairly high value—somewhere between 600 and 1000pF. This in itself would cause no problem as long as we could drive it from a low impedance source. But if we look more closely at a source follower's equivalent circuit (see *Fig. 4*), we find a small inductance, the lead inductance of the gate contact, in series with the input. This inductance can team up with C_L, consisting of drain-to-source capacitance, stray wire capacitance, etc., and cause a very high-frequency oscillation. Increasing the source impedance will prevent this oscillation.

So first we said use a low source impedance drive for wide frequency response. Now we're saying increase the source impedance to avoid an inherent instability. Clearly, both requirements cannot be satisfied simultaneously.

FIG. 4

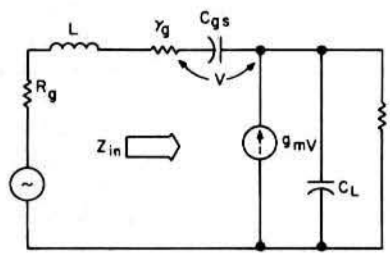

Fig. 4. Equivalent circuit of source follower.

Fortunately, there is a way out of this dilemma. If we look at the drive requirements for the purely capacitive input-impedance, we realize we don't have to have a low impedance source. We just have to make sure we have enough current available from the driver circuit to be able to charge and discharge the input capacitance.

Let's assume the total capacitance to be driven is 1000pF and that we want to drive this capacitor to the full voltage swing of 30V peak (equivalent to 60W into 8Ω) at 40kHz. We need a drive current of:

$$i = S \cdot C \qquad (1)$$

where S is slew rate and is equal to:

$$S = 2 \cdot \pi \cdot f \cdot V_{peak} \qquad (2)$$

and C is the total capacitance to be driven with a slew rate equal to S.

Calculating the slew rate first:

$$S = 2 \cdot \pi \cdot 40 \cdot 10^3 \cdot 30 = 7.5\text{V}/\mu\text{S} \qquad (3)$$

FIG. 2

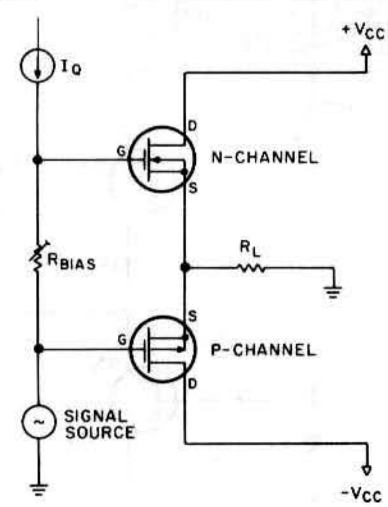

Fig. 2. Source follower output stage.

and the necessary drive current becomes:

$$i = 7.5 \cdot 10^6 \cdot 1000 \cdot 10^{-12} = 7.5\text{mA} \quad (4)$$

This calculation does not take internal compensation or external low-pass filters into consideration. To make sure that we end up with a reasonable high frequency distortion in the final amplifier, we start with a much higher slew rate. My rule-of-thumb value is five times the calculated value. This will bring us up to 30-40mA in the driver stage, which should give us plenty of margin for both slew rate and low distortion.

DRIVER CIRCUIT

My idea of a driver circuit is one that converts the 1V or so input voltage coming from the preamplifier to the voltage swing needed to drive the output devices, without influencing the signal in any way. In other words, the conversion process must be linear, and this linearity should be inherent in the circuit before feedback is applied.

Talking about feedback, I would like to make a couple of comments, as this subject seems to be perennial in *TAA* and other audio publications. First of all, treat feedback like a friend, not an enemy. Just as with friendship, rely on it when you need it, but never strain it. Applying the right amount of feedback in a well-controlled manner cannot hurt anybody; but when you try to cover up a bad design with a lot of feedback, it may turn sour on you.

My second point: feedback does not have to produce TIM, SID, or whatever the current name happens to be for internal overload in a feedback amplifier. Those engaged in feedback amplifier design (see for example Peter Baxandall's and R. Greiner's excellent articles on the subject) have been aware of these problems for many years, and you can avoid them by following simple design rules.

Some people even demand no feedback at all in audio amplifiers. Let's face it: it's possible to design audio amplifiers without overall feedback (though not without local feedback), but I simply don't see the need for it. Properly designed amplifiers with overall feedback sound just as good as those without it, and provide much simpler circuitry combined with long-term stability.

Now I've finished and will return to the driver circuit!

To my mind inherent linearity is the most important factor in driver circuit topology. Assuming that the output stage has no voltage gain, which is the case in practically all power amplifiers on the market, the driver stage immediately preceding the output must be able to swing the whole output voltage and some more to compensate for any output stage voltage loss.

The most linear stage for large voltage swings is the grounded-base or common-base amplifier. Unfortunately, interfacing this stage with others is not easy because of biasing problems. Lender[2] presents the most elegant solution to this problem I have come across. His circuit has a single differential input stage and a current mirror transfers the output back to single-ended output. A simplified version of his circuit is shown in *Fig. 5*.

The AC voltage at the Q_4 emitter is very small due to the low impedance at this point. We can call this a current coupling between first stage and second stage. Of course, a one-to-one coupling is out of the question: the first stage works with 1mA, let us say, while the second stage (Q_4) is supposed to work with some 30mA. The current shifts from 1 to 30mA with the help of emitter follower Q_3 and with resistors R_1 and R_2. Let's see how this works.

The voltage drop across R_1 and R_2 is equal within the V_{BE} difference of Q_3 and Q_4. Consequently we can write:

$$V_1 = \frac{I_O}{2} R_1 \quad (5)$$

$$V_2 = (I_2 + \frac{I_O}{2}) \cdot R_2 \quad (6)$$

But $V_1 = V_2$, and we can write:

$$\frac{I_O}{2} \cdot R_1 = (I_2 + \frac{I_O}{2}) \cdot R_2 \quad (7)$$

The resistor ratio then will be:

$$\frac{R_1}{R_2} = \frac{I_2 + \frac{I_o}{2}}{\frac{I_o}{2}} \quad (8)$$

Assuming that $I_O = 2\text{mA}$, which is a practical value, we get:

$$\frac{R_1}{R_2} = \frac{I_2 + 1}{1} = I_2 + 1 \quad (9)$$

If we want, for example, $I_2 = 10\text{mA}$, then the ratio of the two resistors has to be 11. Using $R_1 = 2.2\text{k}$, R_2 must be 200Ω. Lender's circuit is very linear, but we can further improve it by making it completely symmetrical, as shown in *Fig. 6*.

Looking at this circuit, you can see that in the first stage we lose about three volts on each side to second stage interface. Theoretically, this circuit should have a separate power supply ±5V higher than the supply of the output stage. For simplicity, I use the same supply for driver and output stage, with an RC filter in between.

COMPLETE CIRCUIT DIAGRAM

And now let's look at *Fig. 7*, the complete circuit diagram.

Q_1 and Q_2 form the NPN part and Q_5 and Q_6 the PNP part of the double differential input amplifier. These transistors work with approximately 50V collector-emitter voltage; V_{CEO} should therefore be at least 60V. In addition, h_{FE} should be as high as possible, so as to cause a minimum of offset due to the

FIG. 5

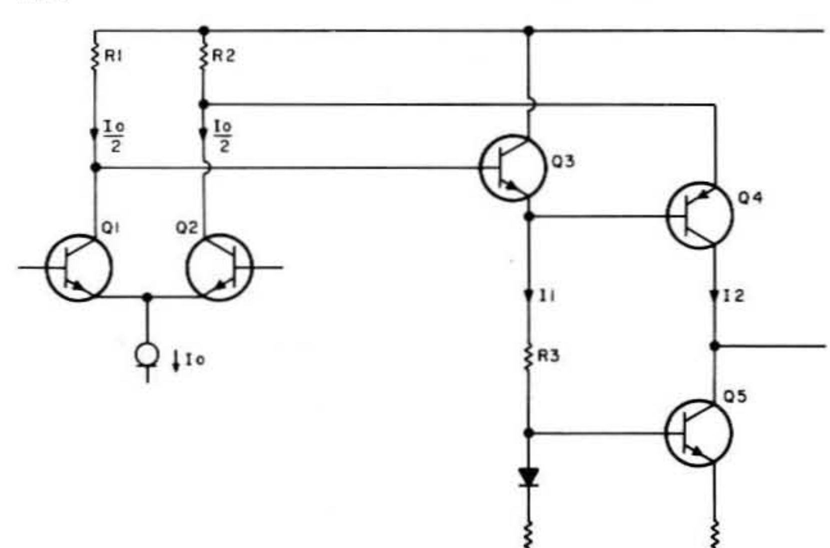

Fig. 5. Lender's driver circuit.

bias current. Strictly speaking, matching is not necessary as long as the NPN and PNP transistors are complementary devices coming from the same h_{FE} group, and as long as this h_{FE} is high enough. Typical devices applicable in this position are the MPS 8099 (NPN) and 8599 (PNP) transistors. h_{FE} at 1mA is specified as 100-300, which is rather low but will cause no problem as long as the DC resistance between base and ground is held relatively low.

A better choice is the BC 546B (NPN) and 556B (PNP) complementary pair, with V_{CEO} = 65V and h_{FE} = 180 – 450, specified at 2mA. These numbers are of the European type, but the devices are manufactured and sold by a number of American companies as well.

If you have a well-stocked distributor in your area, you may find matched dual transistors, which are suitable as input devices. I am using a 2N 2920 on the NPN side and a 2N 3811 on the PNP side. These are 60V devices, with h_{FE} = 300-600 at 1mA.

Figure 8 shows pin connections for all three-types of devices. Note that they are all shown "bottom view," which is the way you will find them in semiconductor catalogs.

Q_3 and Q_4 supply the necessary 2mA constant current to the NPN and the PNP input stages respectively. These transistors are not critical; you can use either the MPS 8099/8599 or the NC

FIG. 6

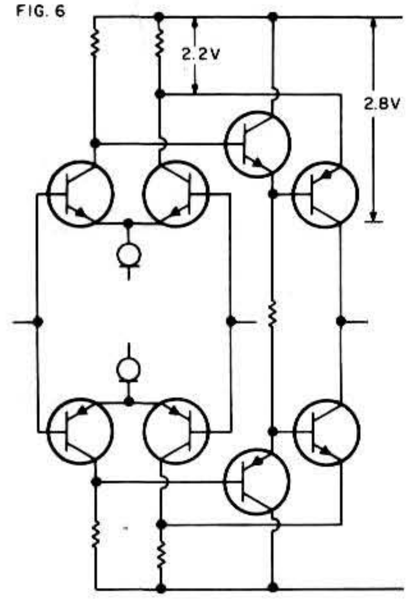

Fig. 6. Symmetrical driver circuit.

FIG. 8

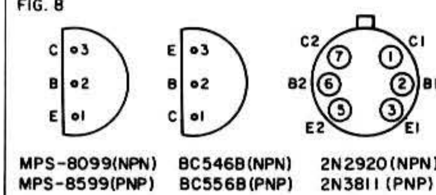

Fig. 8. Input devices.

FIG. 7

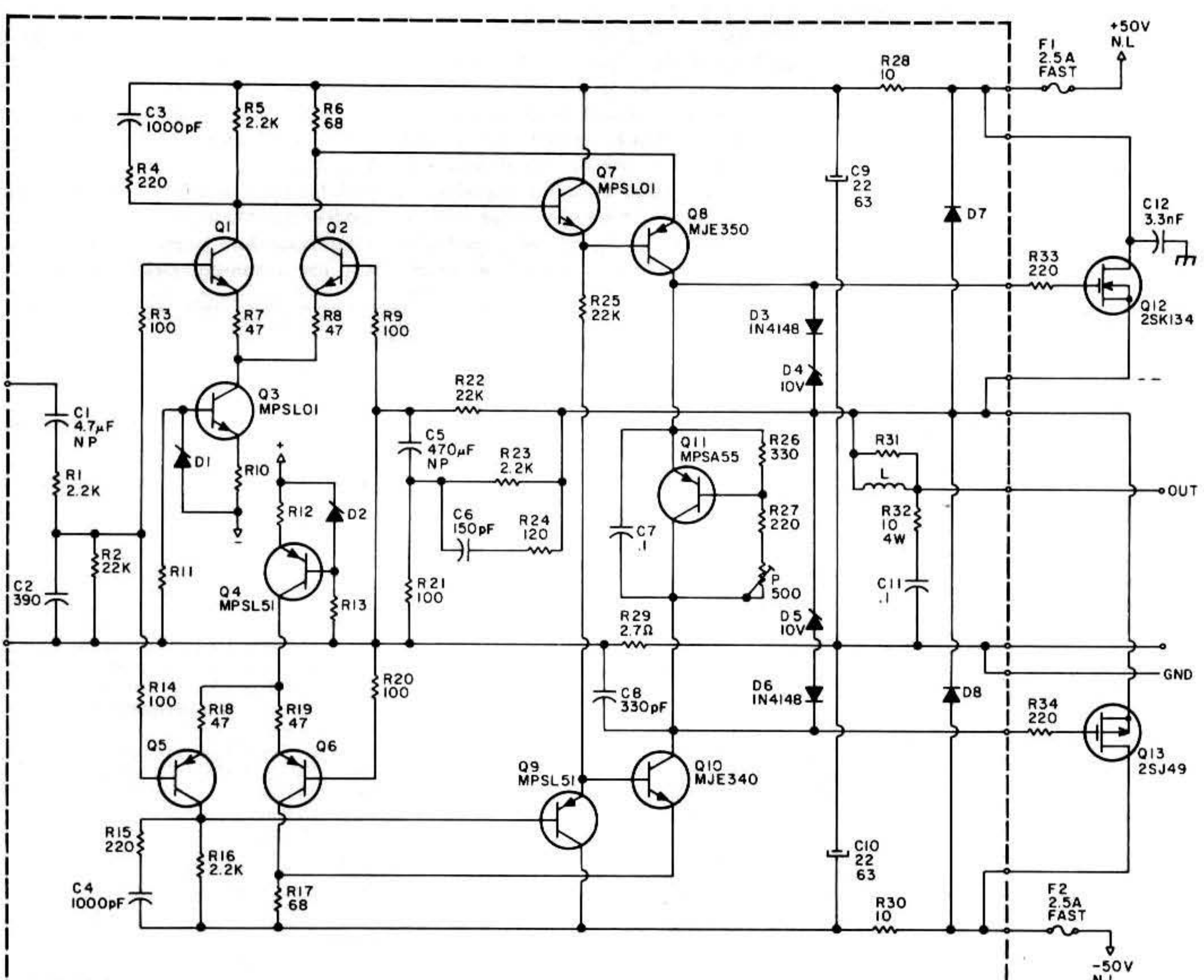

Fig. 7. Complete circuit diagram.

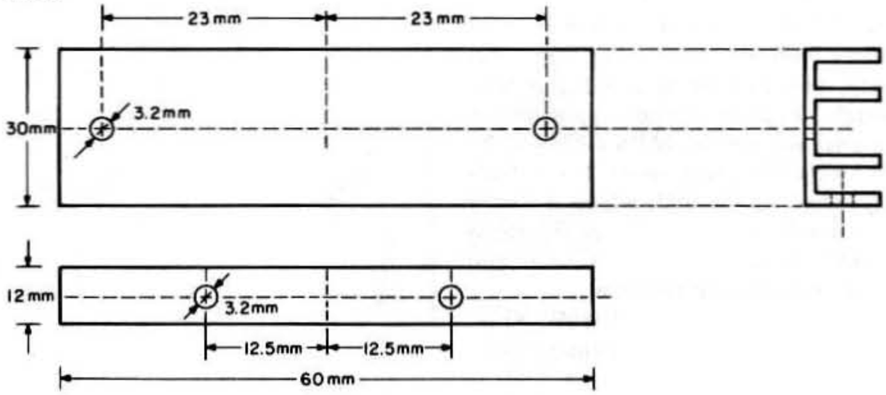

Fig. 10. Heatsink for Q_9 and Q_{10}. (Not drawn to scale)

546B/556B. Since h_{FE} does not have to be very high, the MPS L01 L51 transistors will also serve. As for the zener diodes D_1 and D_2 indicated on the schematic, I recommend the excellent and inexpensive voltage reference diodes from National Semiconductor. National have a variety of active devices operating over a wide current range and offering a much lower dynamic impedance than a normal zener diode. The LM 385Z-1,2, with a specified reverse breakdown voltage of 1.23V is suitable; its guaranteed tolerance is ±30mV. Specify the LM385Z in an inexpensive plastic package. The pin configuration appears in *Fig. 9*. Pin 2 is connected to the most positive voltage.

If you have trouble locating this reference diode, you can use a regular zener diode. A 4.7V, ½W diode such as the 1N5230B is suitable. In this case, however, you must change R_{10} and R_{12} to 2kΩ . (Not 2.2k!)

The input network, C_1, R_1, C_2, and R_2, has a number of functions. C_1 blocks the DC voltage from the preamplifier. Its minimum value should be 2.2µF, giving a 3Hz rolloff. I use a 4.7µF for 1.5Hz rolloff. Ideally C_1 should be a polypropylene capacitor, but they can be hard to find. The next best is polycarbonate. If you fail to find these, use either a 4.7µF non-polar electrolytic or two 10µF normal electrolytics to make up a non-polar capacitor. All these options are possible on the PC board.

The R_1-C_2 network forms a low-pass filter with a 185kHz rolloff. The actual rolloff in your system will depend on your preamplifier's output impedance, since it acts as though connected in series with R_1. If the output impedance is low, the above value is valid; with output impedance in the range of 10kΩ, including R_1, the −3dB point will move down to approximately 40kHz, which is the lowest acceptable. Don't use a preamp with higher output impedance without modifying the value of C_2, which should be a polypropylene or polystyrene capacitor.

R_2 makes up the DC path for the input transistors' bias current. For minimum DC offset at the output, it should have the same value as R_{22}.

The feedback network consists of R_{21}, R_{22}, R_{23}, R_{24}, C_5, and C_6. R_{22} is the DC path for the feedback; C_5 isolates it from the R_{23}-R_{21} AC feedback. C_5 should be a 470µF tantalum capacitor, but you can also use a non-polar electrolytic or two

FIG. 9

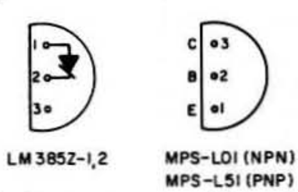

Fig. 9. Constant current source devices.

PARTS LIST (See *Fig. 7*)

Resistors

R_1	2.2k
R_2	22k
R_3	100
R_4	220
R_5	2.2k
R_6	68
R_7	47
R_8	47
R_9	100
R_{10}	300
R_{11}	10k ½W 5%
R_{12}	300
R_{13}	10k ½W 5%
R_{14}	100
R_{15}	220
R_{16}	2.2k
R_{17}	68
R_{18}	47
R_{19}	47
R_{20}	100
R_{21}	100 (95.3 if 1%)
R_{22}	22k
R_{23}	2.2k 1W 2% metaloxide
R_{24}	120
R_{25}	22k ½W 5%
R_{26}	330
R_{27}	220
R_{28}	10
R_{29}	2.7 1W 5% metaloxide
R_{30}	10
R_{31}	1 5W 5% wirewound
R_{32}	10 4W 5% metaloxide
R_{33}	220
R_{34}	220

All resistors ¼W, 5% carbonfilm, unless otherwise noted. The ¼W, 5% resistors can be replaced with 1% Metalfilm. Use nearest value.

Potentiometer

P	500 ¼W linear

Capacitors

C_1	4.7µF/100V non-polar
C_2	390pF/100V polypropylene or polystyrene
C_3	1000pF/100V polypropylene or polystyrene
C_4	1000pF/100V polypropylene or polystyrene
C_5	470µF/6.3V tantalum or non-polar electrolytic
C_6	150pF/100V polypropylene or polystyrene
C_7	.1µF/250V polycarbonate or mylar
C_8	330pF/250V polystyrene or dipped mica
C_9	22µF/63V electrolytic
C_{10}	22µF/63V electrolytic
C_{11}	.1µF/250V polycarbonate or ceramic
C_{12}	3300pF/250V dipped mica, polystyrene, ceramic

Transistors

Q_1	MPS 8099 } or 2N2920 dual transistor
Q_2	MPS 8099 }
Q_3	MPS L01
Q_4	MPS L51
Q_5	MPS 8599 } or 2N3811 dual transistor
Q_6	MPS 8599 }
Q_7	MPS L01
Q_8	MJE 350
Q_9	MPS L51
Q_{10}	MJE 340
Q_{11}	MPS A55
Q_{12}	2SK 134
Q_{13}	2SJ 49

Diodes

D_1	LM 385Z-1,2 1.23V ref. diode (National Semiconductor)
D_2	LM 385Z-1,2 1.23V ref. diode (National Semiconductor)
D_3	1N4148
D_4	1N 5240B 10V ½W zener
D_5	1N 5240B 10V, ½W zener
D_6	1N4148
D_7	1N4004
D_8	1N4004

Coil

L	See Text

Fuses

F_1	2.5A Fast
F_2	2.5A Fast

1000µF normal electrolytics connected for non-polar operation. Again, all these options are possible with the proposed PC board layout.

Total AC-gain is given by the formula:

$$\frac{R_{22} \| R_{23} + R_{21}}{R_{21}} \qquad (10)$$

where $R_{22} \| R_{23}$ means R_{22} in parallel with R_{23}. With the value indicated, the gain is equal to 21 times or 26.4dB. Taking into consideration the voltage drop across R_1, which amounts to about .8dB, we come up with a total gain of 25.6dB. If you are using 1% resistors in your amplifier, I suggest that you change R_{21} to 95.3Ω, which will bring you closer to 26dB. However, don't forget that R_{22} and R_{23} can cause error if their tolerance is not accurate.

Q_8 and Q_{10} are the amplifier driver transistors. Operating at about 31mA, they dissipate approximately 1.5W each and require a heatsink. High voltage transistors for this application are hard to find, except in the TO-5 package which unfortunately is not easy to heatsink. I chose an NPN/PNP pair in TO-126 plastic package: the MJE 340 (NPN) and MJE 350 (PNP). A suitable size heatsink for the two devices is shown in *Fig. 10*. If you cannot find an extrusion of suitable size, you can make one from an L-bracket. Don't forget to use mica and silicone grease under the transistor. *Figure 11* shows pin configuration for MJE 340/350.

FIG. 11

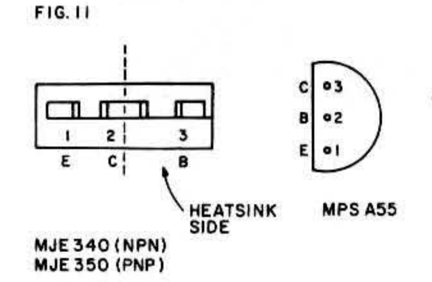

Fig. 11. Second stage devices.

Emitter followers Q_7 and Q_9 operate at less than 5mA, supplying a worst case base current of approximately 1mA to Q_8 and Q_{10}. Suitable devices are MPS L01/MPS L51.

Q_{11} and its associated components, R_{26}, R_{27}, and P, supply bias. Hitachi recommend a simple potentiometer to adjust the bias, as shown in *Fig. 2*. This is not suitable for us because of the relatively high second stage current. A simple way to relieve the potentiometer from conducting all this current is to use it with a transistor, just as in a normal bias adjustment circuit in bipolar amplifiers. The difference is that this circuit has no temperature compensation function. Therefore do *not* mount transistor Q_{11} on or near the output heatsink: it should operate at ambient temperature. Q_{11} should be a high current transistor; an MPS A55 works well.

Before you switch on the amplifier, turn potentiometer P to its CCW position, to make sure there is no excessive current through the output stage. The simplest way to adjust the bias is to take out fuse F_1 (or F_2) and connect an ammeter across the fuse terminals. When you so apply supply voltage you will be able to monitor the current through the whole amplifier.

With no audio signal at the input (short the input if you have a shorting plug) adjust the amplifier's quiescent current to approximately 150mA, as measured by the ammeter across the fuse holder. This gives approximately 100mA quiescent current through the Mosfets, which is optimum for thermal stability and low level distortion, and is the only amplifier adjustment you have to make.

Diodes D_3, D_4, D_5, and D_6 protect the Mosfets from gate-to-source breakdown. The 2SK134/2SJ49 devices are specified at 14V. If you load your amplifier with a very low impedance (for example, a large capacitive load at very high frequencies) the output voltage can move very little. However, the feedback tries to correct the situation by pushing Q_8-Q_{10} to deliver more voltage swing to the output devices to compensate for the low output. This will eventually overload the second stage, delivering high-voltage spikes to the Mosfets. Under these abnormal conditions, diodes D_3-D_6 will prevent voltage spikes greater than 10V from reaching the Mosfets.

Network L-R_{31} isolates the amplifier from capacitive loads. L's increasing impedance at high frequencies prevents a capacitive load from shorting amplifier output. R_{31}, which equals 1Ω, damps resonant circuit Q so sustained ringing will not occur with capacitive loads.

You can make the coil L yourself. I made two versions: one on a round form used in ferrite cores, and the other on a long form. The latter is a plastic spacer, in which I drilled holes 20mm. apart through which to thread the ends of the coil. I give dimensions for both types in *Fig. 12*.

FIG. 12

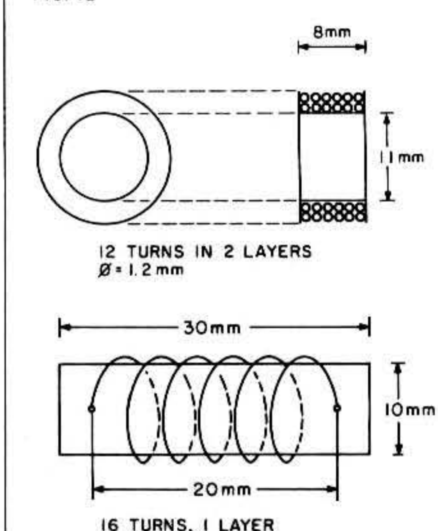

Fig. 12. Output coil. (Not drawn to scale.)

R_{32} and C_{11} terminate the amplifier with a resistive load at very high frequencies. We need this because most speaker systems on the market represent a very high and usually uncontrolled impedance at these frequencies. By providing a resistive load we ensure that the amplifier is working under controlled conditions at all times.

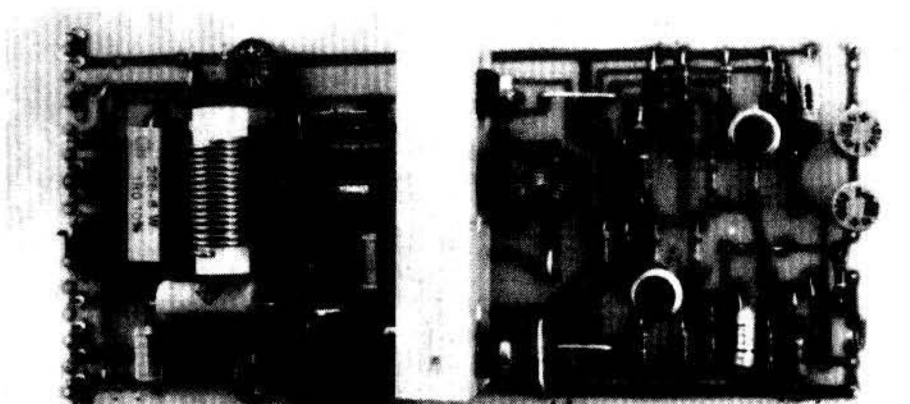

Photo A. One channel of the amplifier prototype. Note the piece of ½" aluminum angle fabricated into a heatsink for driver transistors Q_8 and Q_{10}. Note the three jumpers to the right of the heatsink and the one jumper at the output (left) end of the board. The output choke is wound on a small dowel drilled to secure the wire at each end. The small pins on the output end are part of a European breadboard system, making convenient tie points for connections to the output Mosfets.

D_7 and D_8, which are 1N4004 type diodes, protect the amplifier from kick-back energy from inductive loads. Resistors R_{33} and R_{34} are not on the PC board. Wire them as close as possible to the Mosfet gate-pins. I usually use sockets for power devices and solder the 220Ω resistors directly to the socket.

I also solder the capacitor C_{12} directly to the socket drain pin. The other side of the capacitor must be connected to a ground lug close to the Mosfet. Make sure that this lug makes good contact with the heatsink: if the heatsink is black-anodized you must scrape off the finish, or the capacitor will "hang in the air."

Mount the output devices on a heat-sink with a thermal resistance no worse than 1°C/W. Don't forget the mica insulator for the Mosfets, and be generous with silicone grease on both sides of the mica. Pin configuration for the Mosfets is shown in *Fig. 13*.

A 1:1 copy of the copper side of the PC board layout appears in *Fig. 14*. *Figure 15* shows the board from the component side.

As you probably noticed, this amplifier has no protection circuit except for the gate-source diodes. Nevertheless, the amplifier is short-circuit-proof, at least for a brief time. If you mount a thermal breaker on the heatsink, you will also take care of the time-factor; I didn't because thermal breakers cost a lot.

You will also notice that I use no DC sensing and protection circuit. First of all, the amplifier has no DC thumps to speak of; secondly, I found power Mosfets so reliable that I felt such protection

FIG. 13

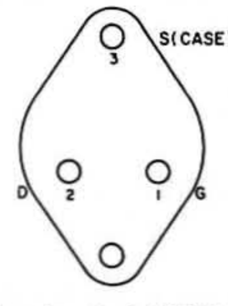

Fig. 13. Pin configuration for 2SK134 and 2SJ49.

FIG. 14

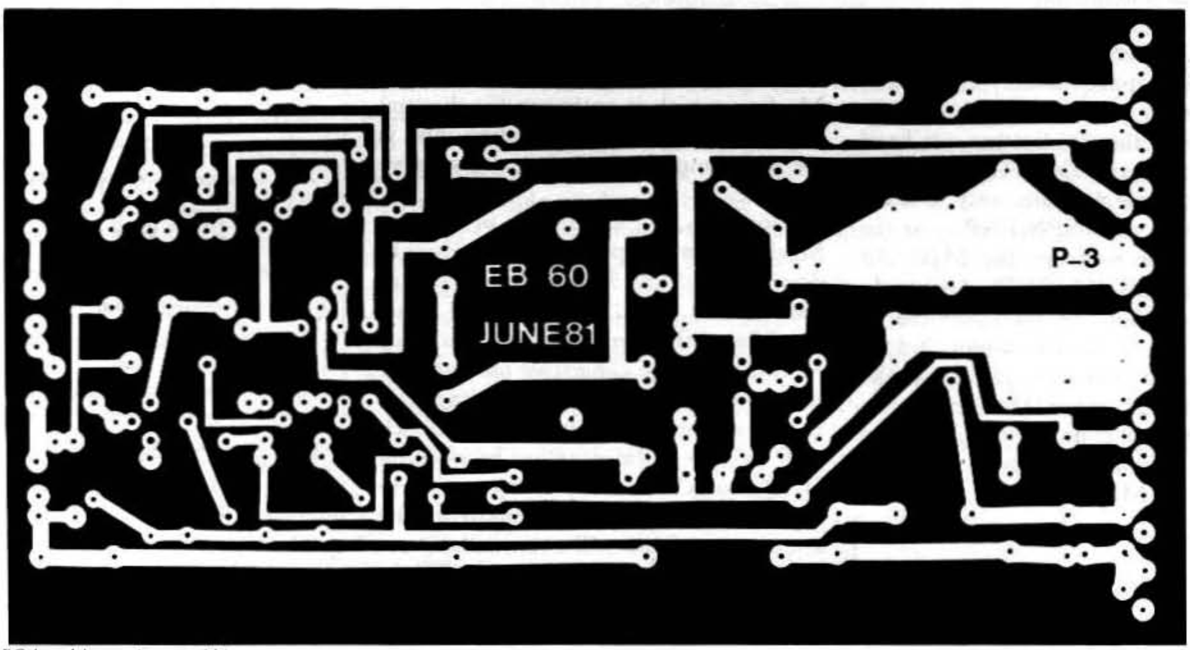

Fig. 14. PC board layout (copper side).

FIG. 15

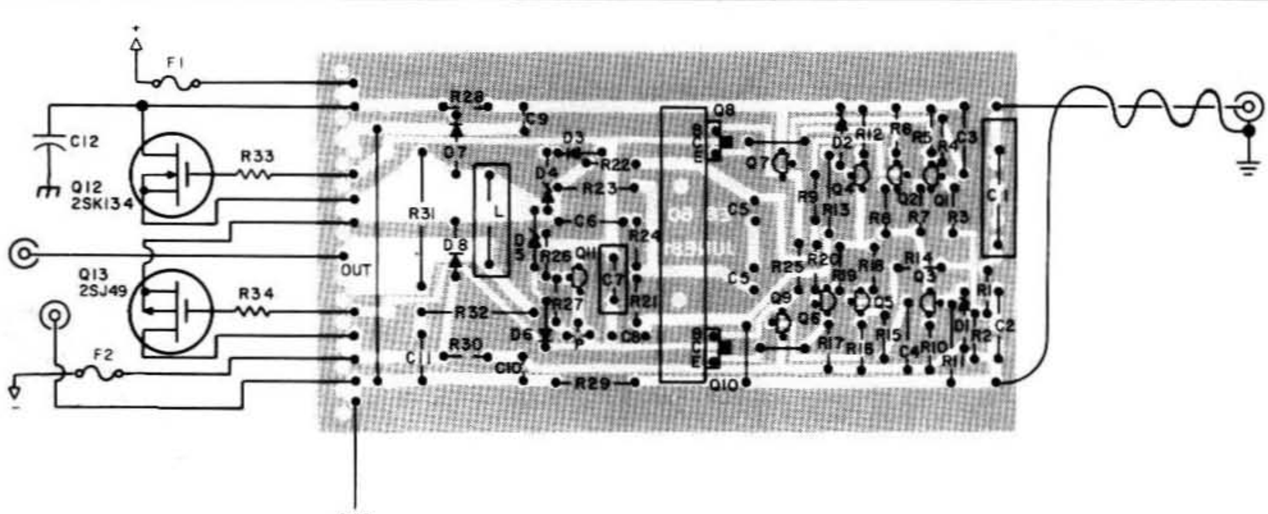

Fig. 15. Component placement on PC board. (Component side.)

was unnecessary. However, if you feel uneasy about this, don't hesitate to build in a DC sensing and protection circuit; I am sure you can find dozens of such circuits on the market.

POWER SUPPLY

We need (+) and (−) 50V no-load supply voltage to operate this amplifier. The supply should not drop below ±45V at full output, i.e., when measured at 60W in 8Ω. Naturally, if you make a stereo amplifier and feed both channels from the same supply, the above specifications apply when driving both channels.

The transformer I use for the stereo version is rated at 200VA, and has a 2x35V secondary winding. This is just a bit low on the voltage side; if you have a choice, try to get a 2x36V version.

The rectifier bridge should be rated at a minimum of 15A, with a surge current rating of at least 150A. A 200V bridge will do.

The filter capacitors are 10,000μF each, rated at 63V. I consider this adequate for a 2x60W amplifier, but if you prefer very large storage capacitors, don't hesitate to design your own power station.

Figure 16 shows the power supply and wiring diagram for a stereo version of the amplifier. Note that one of the input jacks is isolated from the chassis and a 2.7Ω, ½W resistor is connected between the ground side of the jack and the chassis.

CONCLUSION

The simple amplifier circuit presented in this paper offers very high performance (.002% THD at 1kHz, internal slew-rate of 40V/μS) combined with high reliability. I have made a number of these amplifiers in the last couple of years and none of them has quit working even with the most difficult loudspeaker loads. As for the power Mosfets, I haven't managed to destroy a single one since I started to work with them in 1978.

This circuit is specifically designed for a single pair of output devices. Should you require more power, I would propose a different driver circuit with more current capability. Tell the Editor your requirements: maybe he will ask me to come up with a high power version. In the meantime, have fun with your new project!

Key Specifications

Gain:	26dB
Feedback: approx.	30dB
Input Impedance:	24kΩ
THD: 1kHz:	.002% at 60W/8Ω
10kHz:	.015% at 60W/8Ω
Slew-rate:	40V/μS (without input LP filter)
Rise-time:	2μS (with input LP filter)
Output:	60W/8Ω 75W/4Ω

REFERENCES

1. Sampei, Ohashi, Ochi: "100 Watt Super Audio Amplifier Using New MOS Devices." IEEE *Transactions* on Consumer Electronics. Vol. CE-23. No. 3. August 1977.
2. R. Lender: "Power Amplifier With Darlington Output Stage." Motorola Inc., Geneva, Switzerland, 7.9.1974. Internal Report.

Erno Borbely *received a degree in Electronic Engineering from the Technical University of Norway in 1961. For seven years he worked for the Norwegian Broadcasting Corporation designing professional audio equipment. Moving to the U.S. in 1969, he started to work for David Hafler at Dynaco as a design engineer, becoming Director of Engineering in 1972. While at Dynaco, he designed the "dynatune" circuit of the FM-5 tuner, for which he received a patent. He also designed the high power output stage using series-coupled power transistors for the Stereo 400 power amplifier.*

In 1973 he joined Motorola in Geneva, Switzerland as a Senior Applications Engineer, responsible for audio and radio, working mostly on low-noise circuit design and on power amplifier circuits. Some of the ideas formulated there later found their way into the products of the David Hafler Co. which he joined in 1978. The phono stage and pre-preamp in the DH-101 for example, was first shown at the Winter CES in January 1979.

Author Borbely is presently employed by National Semiconductor as Training Manager for Europe. His spare time activities include loudspeaker systems design and audio circuit design.

FIG. 16

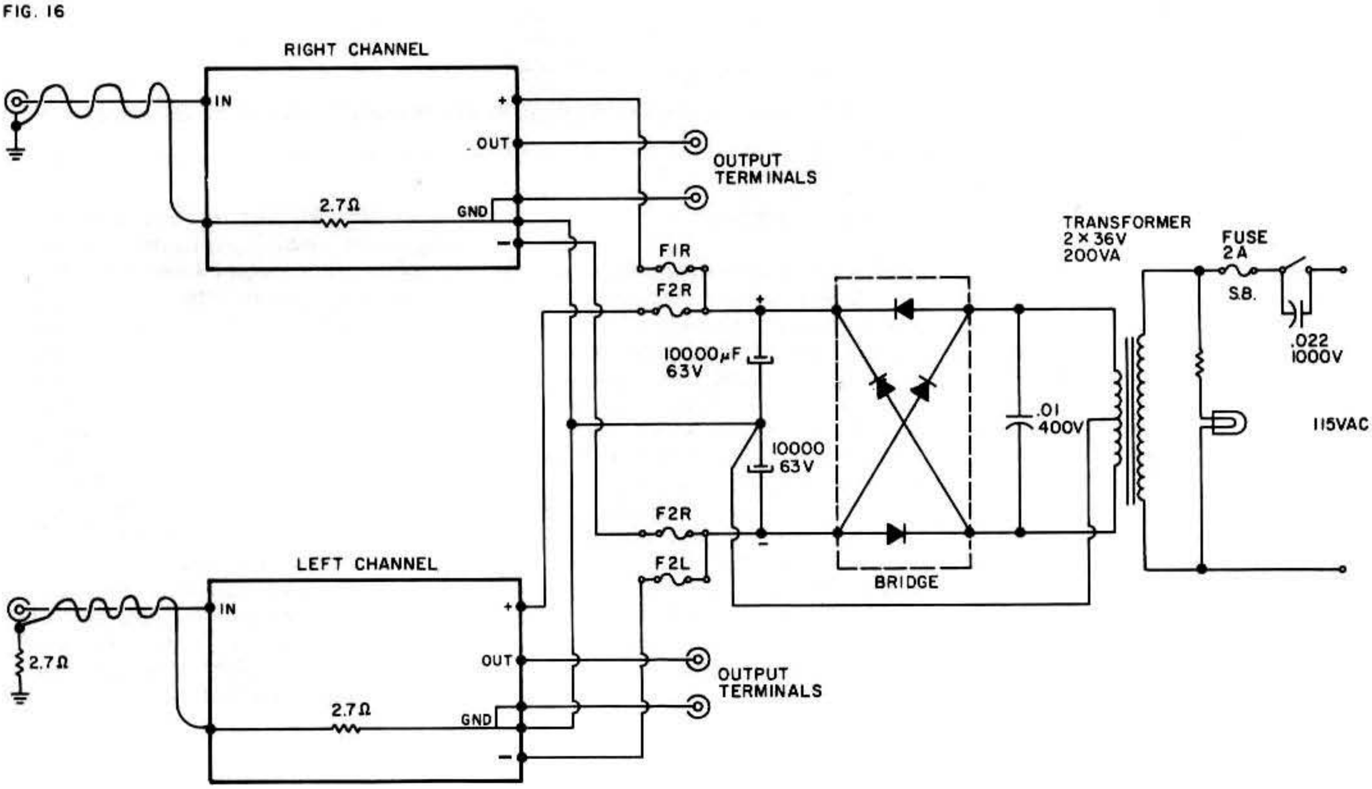

Fig. 16. Power supply and wiring diagram for a stereo amplifier.