

High power high quality amplifier using mosfets

Though power-mosfets are superior to bipolars because of wide frequency response, high switching speed and absence of secondary breakdown, input capacitance can cause non-linearity problems at high frequencies if you don't have enough drive current. This is especially important when paralleling mosfets for higher output power.

The amplification of very low level signals delivered by a pickup to the high power level needed to drive inefficient loudspeakers presents unique problems which have attracted the attention of many engineers, as witnessed by the many articles in WW. Some of these are closely related to the devices used: bipolar transistors. The relatively new power devices in m.o.s. technology, power mosfets, are capable of solving most of these problems.

Audio power amplifiers using bipolar devices tend to be overloaded protection. The first one is needed because of the positive temperature coefficient of the collector current versus base-emitter voltage. With increasing power dissipation we get an increasing junction temperature, which increases with the collector current and therefore power dissipation. Eventually, this will lead to thermal runaway, causing destruction of the device. To avoid it a sophisticated circuit is needed to sense the temperature on the output devices and regulate the quiescent current through them accordingly.

Bipolar power transistors are also well known for current concentration under high current conditions. This causes local

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hot-spots which can eventually lead to secondary breakdown. This is especially dangerous at combinations of high voltage and high current. Consequently bipolar power amplifiers use sophisticated voltage limiting to get the maximum from the devices without damaging them.

The Hitachi 2SK134/2SJ49 power mosfets used in this design do not have such problems. Their most important advantage is the negative temperature coefficient of the drain current versus gate-source voltage. As Fig. 1 shows, the drain current is decreasing with increasing temperature above approximately 100mA. This means that thermal runaway cannot occur and complicated temperature-tracking circuits are therefore not required.

Another advantage is the uniform current distribution across the whole silicon

die, which does not deteriorate in the high current, high voltage area. The result is that there are no hot spots and no second breakdown. Sophisticated V-A limiters are

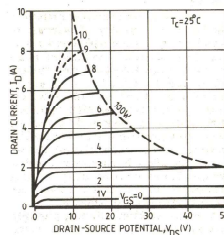


Fig. 3. Typical I_D - V_{DS} curves for 2SK134 show that maximum drain current is limited to seven amps.

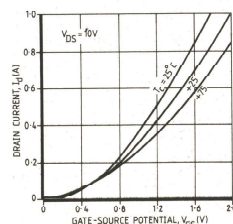


Fig. 1. Negative temperature coefficient of the drain current above 100mA prevents thermal runaway with Hitachi 2SK134/2SJ49 type mosfets and avoids use of temperature-tracking circuits.

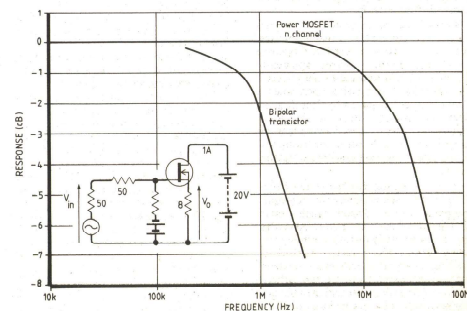


Fig. 2. This n-channel mos source follower has ten times the bandwidth of an emitter follower.

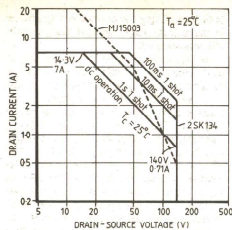


Fig. 4. Compared to a giant bipolar transistor, the mosfet has limited current capability at low voltages, but outperforms the bipolar at combinations of high current and high voltage.

therefore not required for protection. In fact most amplifiers do not need any protection at all.

There are several more advantages with mosfets that are important in audio applications. One is that they are majority carrier devices. Because the charge carriers are controlled by electric fields and not by injection of minority carriers in the active region, there is no stored charge in the gate region. The result is wide frequency response and fast switching, even when coming out of clipping.

Like bipolar transistors, mosfets are usually used in the follower configuration in audio amplifiers. Fig. 2 shows the frequency response of an n-channel mosfet in source-follower configuration. For comparison, the response of an emitter follower is also shown: bandwidth of the source follower is ten times wider than that of a bipolar device.

Another important advantage is their high input impedance. Together with the uniform current distribution, this allows paralleling of mosfets without undue problems. However the spread in the gate-source turn-on voltage has to be considered when doing this, as we will see later. Although the statement "high impedance" is correct at very low frequencies, it is by no means applicable for the whole audio frequency range. The input looks like a pure capacitor with a value of about $1nF$. This can cause non-linearities at high frequencies if you don't have enough drive current available. The input capacitance therefore becomes the most important factor in selecting the driver circuit.

Disadvantages of power mosfets

Although having a number of advantages compared to bipolars, mosfets are by no means ideal. The most important disadvantage with the 2SK134/2SJ40s is the high on-resistance. Worst-case calculations show that it can go up to 1.7Ω ; a current of 5 amps through the device gives a voltage drop of 8.5 volt across it. Clearly this would increase power dissipation and decrease efficiency. Power supply design

has to be based on this worst-case on-resistance. Typical on-resistance is around one ohm, so there is some margin built-in if we calculate with the worst-case value. Other manufacturers are offering devices with much lower on-resistance but they are limited in other parameters such as breakdown voltage and/or in offering complementary devices. A wider selection of mosfets is expected to be available on the market in the next few years.

When looking at the I_D - V_{GS} characteristics of the Hitachi mosfets, Fig. 3, notice that the curve is not shown at higher drain currents than about 7A — the maximum rated drain current. Compared to some of the bipolar transistors rated at 20-30A this might look meagre. But taking into consideration the device's complete freedom from secondary breakdown, one can usually use more of the available current capability of the mosfets than of the bipolars. Fig. 4 shows the safe operating areas for the Hitachi 2SK134 and, for comparison, of a representative bipolar power device, the MJ15003, being a 20A device and superior at low voltages. At high voltages the mosfet is equal to, or in one area it is even better than, the bipolar device.

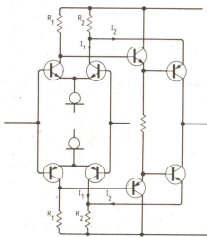


Fig. 6. This offers very good linearity and easy control of the second stage current, I_s .

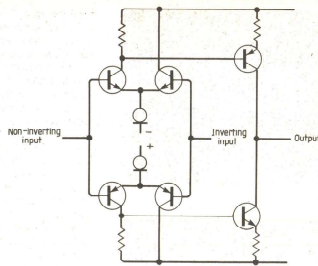


Fig. 5. Many top-quality amplifiers have used a fully balanced, complementary, dual-differential topology, known for very low static distortion at low levels and for good power supply ripple rejection, but it's difficult to get good linearity at high levels, because the second-stage transistors are working with extreme V_{GS} excursions, causing large gain variations.

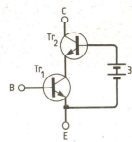


Fig. 7. The cascode configuration is one of the best for large signal handling capability.

The maximum output power one can get with power mosfets is limited only by the thermal capability of the die-package combination. A 2SK134/2SJ49 pair can easily deliver 60W into 8Ω and more than 75W into 4Ω . Two pairs in parallel make a 120W/ 8Ω or a 150W/ 4Ω amplifier, as in this design.

Driver design

The most important considerations when selecting the topology for the driver circuit are

- linearity
- open-loop bandwidth
- drive capability.

Although there are some three-stage driver circuits in commercial amplifiers of Japanese origin, most European and American designers prefer the two-stage design which can achieve both good linearity and wide bandwidth.

The differential input is universally accepted as the input stage. The second stage, which has to supply all of the drive to the output stage, is either a common-emitter stage with a current source as a load, or a differential amplifier using a current mirror for differential to single-ended conversion.

Many top-quality amplifiers have been using a fully-balanced complementary dual-differential topology in recent years, known for very low static distortion at low levels and for good power supply ripple rejection, Fig. 5.

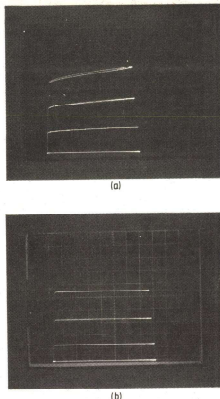


Fig. 8(a) shows the I_C - V_{CE} characteristics of a typical high-voltage transistor, the MJE340. Although limited in terms of voltage and current, the curves clearly indicate gain variations with changing collector-emitter voltage. (b) shows the characteristics of cascode circuit where T_2 is a MJE340, and T_1 is MPSA55. The composite characteristics of this circuit are extremely linear, approaching that of an ideal transistor. Voltage variations across the transistors do not cause significant gain variations, consequently these can handle large voltage swings with very little distortion.

It is more difficult to get good linearity at high levels because the second-stage transistors are working with extreme V_{CE} excursions, causing large gain variations. There are a number of ways to improve large-signal handling capability; I have found the circuit shown in Fig. 6 excellent in this respect. This fully symmetrical circuit also allows easy control of the current in the second stage; it is fixed by the ratio of the two collector resistors in the input stages:

$$R_1/R_2 = (I_2 + I_1)/I_1$$

I have found that this current transformation from first stage to second stage works well up to a ratio of about 30 and, with the particular transistors used, up to a second stage current of 30mA. This current was used to drive a pair of 25K134/23J49s in a 60W amplifier with good linearity and slew rate. However, if we want to drive two pairs of devices to increase the output power to over 100W, this current is not sufficient any more (see calculations below). To be able to use higher current in the second stage with

good linearity, I redesigned it using a cascode configuration.

Fig. 7 shows the basic cascode connection of two transistors. Used alone, a high-voltage transistor such as the MJE340 clearly suffers from gain variations with changing collector-emitter voltage; see the I_C - V_{CE} characteristics in Fig. 8(a). The composite characteristics of a cascode circuit (MPSA55, MJE340) are extremely linear, approaching that of an ideal transistor, Fig. 8(b). Voltage variations across the transistors do not cause significant gain variations, consequently these can handle large voltage swings with very little distortion.

An additional advantage of the cascode circuit is its wide bandwidth. It was established many years ago¹ that we need an open-loop frequency response up to 20kHz to avoid transient overload inside the feed-

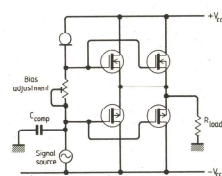


Fig. 9. Paralleling mosfets is easy, thanks to the high-input impedance and the uniform current distribution across the die.

back loop. Using the cascode configuration in the second stage, we can easily satisfy this requirement.

As the collector-base voltage of T_1 is held constant in this circuit there is minimum charging of its collector-base capacitance, effectively eliminating the influence of this capacitor on the frequency response. Transistor T_2 is operating in common-base mode, which is inherently a wideband configuration. Together, they offer a very linear operation over a wide frequency range. A slight disadvantage is the increased voltage loss across the two devices. This is not a problem in our application because the voltage loss across the output devices due to the on-resistance will dominate. Alternatively the driver circuit could be supplied from a higher voltage.

Finally, look at the drive requirements. Fig. 9 shows the output stage, consisting of two pairs of n and p-channel devices operating in source-follower mode. The input capacitance of the p channel fet is around 900pF, the n-channel around 500pF. Used in the source-follower mode, the input capacitance is reduced by the local feedback. As the transconductance of the devices is fairly low ($g_m = 0.7$ to $1.4S$), this reduction is moderate. Calculations based on the published figures in the Hitachi data sheets show that we will end up with 100 to 200 pF input capacitance for each of the devices using an 8Ω load.

The input of the output stage is normally used as the main roll-off point for the amplifier. Because input capacitance varies with operating conditions, it is necessary to add a real capacitor here (C_{comp} in Fig. 9), to come up with the total value needed to stabilize the amplifier. Assuming C_{comp} 400pF, and 150pF each

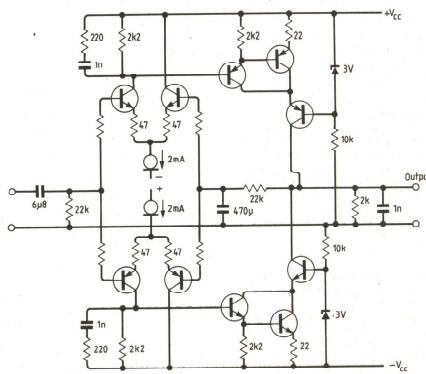


Fig. 10. Driver circuit chosen as it was for open-loop measurements: gain without the load was too high; the 2kΩ load reduced it to about 66dB. In parallel with 1nF the distortion was 0.1%, both at 1 and 10kHz. Response was down 1dB at 20kHz, 2dB at 30kHz.

for the mosfets, the total load capacitance seen by the driver circuit is around 1nF.

To avoid non-linearities caused by this input capacitance, we have to have enough current available in the driver stage to charge and discharge it under all conditions. I was especially interested in very low distortion across the whole audio band with a reasonably high internal slew rate as a secondary requirement. This was to be higher than the slew rate of any signal capable of entering the amplifier, through the low-pass filter used at the input. Using the well-known formula for the slew rate

$$dV/dt = i/C$$

calculate the necessary current to charge and discharge the input capacitance at the rate required. For a slew rate of 100V/ μ s, and assuming an input capacitance of 1nF, 100mA is needed in the driver stage.

The final driver circuit selected consists of a balanced complementary differential input stage and complementary cascode second stage, Fig. 10. The differential input stages are working with 2mA each (1mA per transistor) while the second stage, for reasons of linearity and power dissipation, ended up with approximately 50mA.

Open-loop distortion for the whole amplifier is approximately 0.5% measured at 1kHz and 100W into 8 Ω . If all of the feedback (40dB in this case) reduced the distortion effectively, we would end up having 0.005% in the final amplifier. This is more than adequate for any high quality amplifier. However, the 2k Ω resistor is shunting too much of the available current away from the load capacitor (coupled to the input capacitor of the output stage), resulting in a little higher distortion at high frequencies than I was looking for. To avoid this, I removed the 2k Ω load and replaced it with 10k Ω (actually 2 \times 2k Ω , see final circuit). This increased the open-loop gain and the feedback, and the final amplifier shows very low distortion across the whole audio range.

It has been said that due to the low g_{m0} of the mosfets, special circuit configurations like the composite source-follower or extremely high open-loop gain in the driver stage are necessary to get an acceptable linearity. It is my experience that the only precaution necessary is to have enough current available to charge and discharge the mosfet input capacitances. The design described shows that very good linearity can be achieved with a moderate amount of feedback (30 and 40dB, respectively) and normal source-follower configuration.

Circuit discussion

In the complete circuit diagram, p.73, the blocking capacitor C_1 together with R_1 - R_2 and the output impedance of the preamplifier determines one of the low-frequency roll-offs of the amplifier. Assuming that the impedance feeding the amplifier is low, the -3dB point will be about 1Hz. The non-polar capacitor C_1 should be polypropylene, polycarbonate or, if these are not available, a polyester capacitor. Worst case, a non-polar electrolytic or two non-

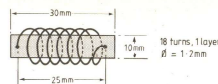


Fig. 11. A suitable output coil can be made on a plastics spacer with parallel resistor to dampen ringing caused by capacitive load.

mal electrolytics connected to form a non-polar capacitor can also be used. The layout gives a number of possibilities as far as type and size of the capacitor(s) are concerned.

Components R_1 and C_2 form a low-pass network to prevent high frequency signals with slew rates higher than the internal slew rate of the amplifier from reaching the input. As the output impedance of the preamplifier forms part of this filter, the value of C_2 should be adjusted for the particular system it is used in. Assuming that the output impedance is very low, the values shown give a -3dB point of over 200kHz, so C_2 should be changed to 1 to 1.2nF, bringing the -3dB point down to about 60kHz. The value shown can be used with preamplifiers with output impedance up to about 10k Ω . C_2 should be a polystyrene or polypropylene capacitor. The d.c. path for the input transistors goes through R_2 which for minimum d.c. offset at the output of the amplifier should be equal to R_{22} .

For minimum d.c. offset we also need high-gain transistors in the input stages but it is difficult to find them with high breakdown voltage. The best compromise I have found is the BC346B and BC350B complementary pair; h_{FE} is specified as 180-450 at 2mA collector current and breakdown voltage 65V. A better choice as far as breakdown voltage is concerned is the MPS8099 and MPS8599 pair (80V) which gives the possibility of higher supply voltages; h_{FE} is 100-300 at 1mA, but does not create any problems with the relatively low-impedance d.c. path used.

The current sources, which supply 2mA to the differential stages, require a stable voltage reference. This is normally obtained from low-voltage zener diodes or a string of small-signal silicon diodes. 1N4148 can be used in this circuit: they show a good repeatability and are very inexpensive. Operating the 1N4148s at around 1mA, the voltage drop is very close to 0.6V; two in series with 2mA through them give an acceptable performance (R_{11} , R_{13} =22k, R_{10} , R_{12} =300 Ω).

A zener diode solution, shown on p.73, needs zener diodes operated at a higher current than the 1N4148s, hence the change of R_{11} , R_{13} to 10k Ω . Although not shown in the schematic, a good temperature compensation can be achieved by connecting a silicon diode in series with the 4.7V zener, which follows the V_{BE} variations of the current-source transistor. This is recommended when operating the amplifier at very high ambient temperatures.

The proposed layout accepts all three solutions: two silicon diodes in series, a

single zener diode or a zener diode and a silicon diode in series.

The 3V reference for the cascode circuit (formed from TR_7 , TR_8 & TR_9 as the p-n-p part, and TR_{11} , TR_{12} and TR_{13} as the n-p-n part) is derived from a string of 1N4148s. This might seem extravagant, but the only alternative, a 3V zener diode, required a much higher operating current. Operating at 5mA, the average forward drop is 0.64V, five of them in series supply 3.2V for the cascode circuit.

Transistor 10 and associated circuitry is for bias adjustment. Hitachi recommend a regular potentiometer for this but as the relatively high current in the second stage is 50mA I didn't find it satisfactory to pass all of this through the wiper of a small trimpot. Instead, I use a normal bias adjustment circuit (V_{BE} -multiplier) found in practically all bipolar amplifiers. The difference is that this circuit is not supposed to have any temperature compensation: TR_{10} should therefore not be mounted on or near the output heatsink.

When switching the amplifier on the first time, the potentiometer should be set to its anti-clockwise position. While monitoring the current through the whole amplifier (for example by removing one of the fuses and connecting an ammeter across its terminals), the quiescent current should be adjusted to 260mA (100mA in each of the output devices, plus driver). This ensures an optimum operation from the point of view of temperature stability, and low-level distortion.

The mosfets are guaranteed for a minimum gate-source breakdown of \pm 14V. The devices have built-in protection diodes, so it should not be necessary to use external protection unless the amplifier is overloaded at very high frequencies. For testing I usually drive the amplifier to maximum output in the frequency range 30 to 50kHz with an 8 Ω load and then connect a large capacitor (around 4 μ F) across the output. The amplifier should work either normally if current handling capability permits or should automatically limit the output.

I have seen the Hitachi devices act a number of ways, depending on the circuit configuration. They might latch, conducting a very high current; they might break up, delivering a very distorted wave form; or they might go into oscillation. Although none of these conditions ever produced a destruction, the devices should not operate under such conditions for an extended period of time. Diodes D_{13} - D_{16} serve that purpose; no matter what current is demanded from the output stage, the diodes prevent the driver stage from delivering more than \pm 10.6V peak gate-to-source voltage. This protects the output from out-of-audio-band overstress and just about any sort of abuse is tolerated, until we reach the thermal limitation of the die/package combination.

There are three more blocks of components on the p.c.-board. One is the feedback network, in which d.c. feedback is applied through R_{22} for minimum offset this is equal to R_2 and a c.-feedback, given by R_{23} and R_{21} , isolated from the d.c.

feedback by C_5 and C_6 . The d.c. gain is unity, and a.c. gain is approximately

$$(R_{23} \parallel R_{22} + R_{21}) / R_{21}$$

which is close to 20 times or 26dB. Capacitor 5 is a non-polar electrolytic and since electrolytics have an increasing impedance at higher frequencies, I connected a 0.1µF capacitor in parallel with it.

Components C_7 and R_{24} provide lead compensation and optimize square-wave response; C_7 should be polystyrene, polypropylene or a dipped-mica type.

The next block is the output network, consisting of L in parallel with R_{26} , and the RC network R_{17} - C_{15} . To prevent short circuit of the output with capacitive load and high frequencies, an inductance is in series with the output. This has an increasing impedance with frequency, thus preventing the short circuit. This can cause sustained ringing with a capacitive load, so we dampen the inductance with R_{26} . A suitable coil is shown in Fig. 11 (I used a plastics spacer as a coil former).

Components R_{17} - C_{15} terminate the amplifier resistively at very high frequencies, necessary because loudspeakers present a very high or uncontrolled impedance at h.f., causing oscillation that can destroy tweeters.

Finally, there are RC networks in each of the supply connections. Although the amplifier has a good inherent ripple rejection, additional filtering is an advantage from the point of view of isolating the two channels. This is especially important at high frequencies, hence the paralleling of the electrolytic with a 0.1µF polyester capacitor.

Heatsink assembly

The rest of the components are mounted directly with the output devices on the heatsink assembly.

The power mosfets are high speed devices, and require special mounting and

wiring precautions. The first one is a proper decoupling of the power supply connections directly at the device packages. Again, a combination of electrolytic and polyester capacitors is used for this. The second one is the usual source follower problem of parasitic oscillation with capacitive load: a resistor in series with the gates to prevent this. Resistors R_{29} , R_{30} , R_{35} and R_{36} should be wired directly to the gate pins of the devices.

The uniform current distribution across the die means hot-spots cannot occur and together with the fact that they have a high input impedance enables operation in parallel mode without too much trouble. However, some precautions have to be taken because of the V_{GS} spread. We want to operate them at a quiescent current of 100mA. Unfortunately, the V_{GS} voltage necessary to turn on the mosfet to this drain current varies from device to device.

Ideally, the devices should be matched to within ±10mA at 100mA drain current. But unfortunately, this requires a large number of devices to select from and can only be done in large-scale production. Happily, devices coming from the same production batch (bought at the same time from a distributor and marked with the same date code), seem to be close enough to work satisfactorily in the amplifier. Resistors R_{41} , R_{42} , R_{43} and R_{44} are also helping in equalizing the current at the same time, as they help linearize the characteristics of the mosfets. These resistors should be non-inductive types.

Although the n and p-channel devices are said to be complementary, some parameters are different. One is input capacitance, specified as 900pF for the p-channel and 500pF for the n-channel ones. This difference can influence rise and fall times in the amplifier, make square-wave response asymmetrical, and made stabilizing more difficult.

As we usually do not know the exact value of the input capacitances, we can

Measurements on practical amplifier

Gain	26dB
Input impedance	24kΩ
Output power	120W into 8Ω 150W into 4Ω
Harmonic 1kHz	0.002% at 120W into 8Ω
distortion 10kHz	0.0065% at 120W into 8Ω
Slew rate	80V/µs without input filter
Rise time	1.2µs without input filter 2µs with input filter

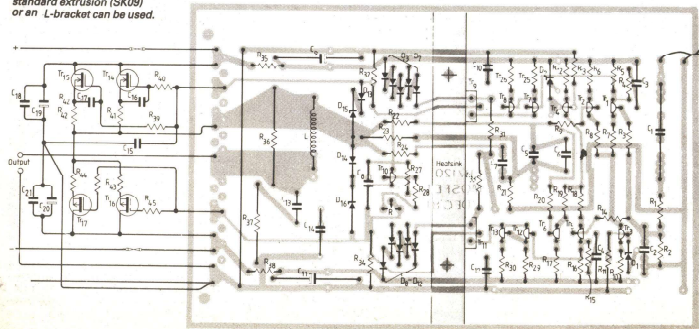
only make an approximate balancing of these by adding extra capacitance to the n-channel devices, the optimum value found by experiment. Capacitors C_{16} and C_{17} should be polystyrene or dipped mica, and soldered directly to the gate-source pins of the n-channel fet's.

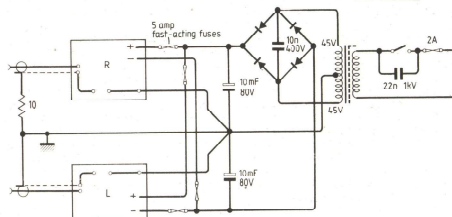
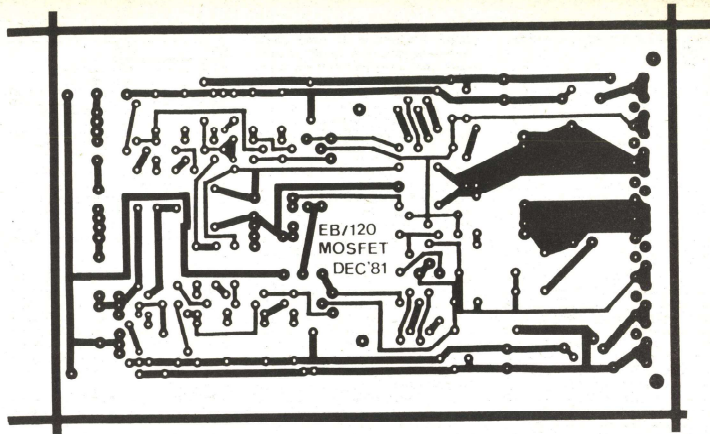
Capacitor C_{15} , together with C_{14} on the p.c. board, make up the compensation shown in Fig. 9. The reason for dividing it into two is to accommodate a wide variety of wiring schemes. A minimum value of around 100pF is needed on the board, the rest does a better job when connected directly at the input of the mosfets. Make them polystyrene types, polypropylene or dipped-mica.

Although not intended primarily for plug-in systems, the p.c. board is made in a standard-size Eurocard format (100x160mm), including a 31pin connector. For simplicity, the input is available at the other end of the board. This makes the layout very straightforward; except for the ground path, the layout follows the circuit diagram very closely. There are two jumpers on the board, both indicated on the component layout.

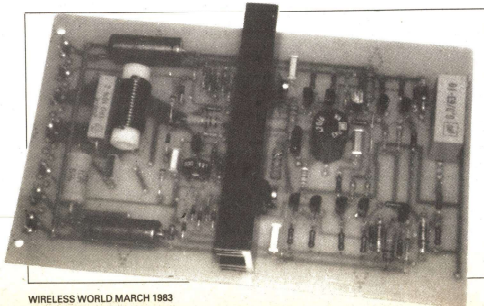
Thermal resistance should be around 0.5deg C/W per channel, which allows the amplifier to operate at a very high average output power continuously. A 150mm piece of a standard SK47 extrusion, with a T or L-bracket to mount the mosfets on, works fine. If the possibility for continuous operation with very low impedance loads exists (less than four ohms), it is a

Transistors 9 & 11 need heat dissipators - either a standard extrusion (SK47) or an L-bracket can be used.





The grounding scheme is flexible in that output ground and input ground are not connected together on the board. This makes it very simple to adopt a number of wiring schemes: a proposed wiring diagram for a stereo amplifier using a common power supply is shown. Using the boards with separate supplies, output and input ground can be connected together, which gives the best signal/hum ratio in a system.



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
The author

In 1973, Erno Borbely joined Motorola in Geneva as a senior applications engineer, responsible for audio and radio. He worked mostly on low-noise circuit design and on power amplifier circuits, and some of the ideas formulated there later found their way into the products of the David Hatter Co., which he joined in 1978 and where he designed the DH-200 mosfet power amplifier.

He'd previously got a degree in electronic engineering from the Technical University of Norway in 1961 and for seven years worked for the Norwegian Broadcasting Corporation designing professional audio equipment. In 1969 he moved to the U.S. to work for David Hatter at Dynaco.

He now works in Furstenfeldbruck as National Semiconductors european training manager.

good idea to put thermal breakers on the heatsinks with a cut-out temperature of about 75°C.

Plus and minus 57V supplies are needed for 120W into an eight ohm load. Assuming a 10% transformer regulation, the no-load voltage should be $\pm 63V$. This translates into a transformer with a $2 \times 45V$ secondary, and rated at 500VA, toroidal for minimum hum. The filter capacitors must be at least 10mF each, rated at 80V. The bridge has a rating of 30A with a surge rating of 300A. 

References

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2. J. Linsley Hood. 80-100 watt mosfet audio amplifier. *Wireless World*, June 1982.