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MOSFET power amp!

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Project 477

MOSFET power amplifier Part 1.

Employing recently released Hitachi MOSFETs, this power amplifier features a 'no compromise' design, is rated to deliver 150 W RMS maximum and features extremely low harmonic, transient and intermodulation distortion. As the circuit techniques and design problems will be unfamiliar to many readers, a thorough discussion of the theory and problems is included.

David Tilbrook

THE ENORMOUS SUCCESS of the series 4000/1 four-way loudspeaker has surprised even us. They were originally intended to be the 'flagship' of a range of loudspeakers and quite frankly we expected the biggest demand to be for the cheaper loudspeakers further down the range. This has proved not to be the case as sales of four-way kits continue to rise. It is evident that there is a big demand for the higher quality audio projects. We recognised this demand and eight months ago began the development of the Series 5000 power amp and preamp. The objective was to design an amplifier for home construction of the highest possible quality. The cost of the project was a secondary consideration, although in real terms the cost saving in "doing it yourself" is considerable.

discussion

Defining the problem

Of all the stages in the amplifier the output stage is subjected to the worst operating conditions: varying load impedance, heating due to the large power levels necessary to drive loudspeakers, and the occasional short circuit produced by the careless connection of loudspeaker cables or perhaps even loudspeaker failure.

The output stage is also the site of three distinct sources of gross non-linearity, that of amplitude overload (clipping), crossover distortion and slew rate limiting. All three generate a very large number of distortion products and are therefore particularly noticeable and fatiguing forms of distortion.

In order to understand the causes of these types of distortion it is helpful to look at the circuit shown in Figure 1. This is a very simple output stage using two transistors. The output to the loudspeaker normally sits at 0 volts, exactly half way between the positive (+V) and the negative supply (−V) rails. Now, if

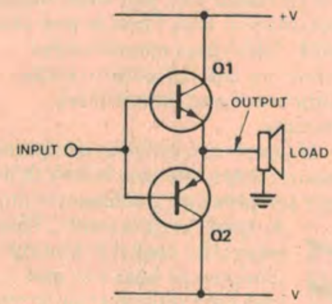


Figure 1. **Simplified circuit of a bipolar output stage.**

Q1 is turned on by a positive-going signal voltage the impedance between the output and the positive supply decreases and the output approaches +V. Similarly, if Q2 is turned on the impedance between the output and the negative supply rail decreases and the output approaches −V. When either output transistor is turned fully on, the output voltage will be equal to the supply voltage minus whatever voltage drop occurs across the output transistors. Any signal peak that exceeds this maximum output voltage will be amplitude limited or clipped (see Figure 2). It is possible to compress signal peaks that may otherwise cause clipping, but inevitably, the non-linearity still occurs. The large supply voltages associated with high powered amplifiers help reduce this problem and are one of the reasons that high power amps almost always sound better than low power ones . . . even at relatively low operating powers. In some respects

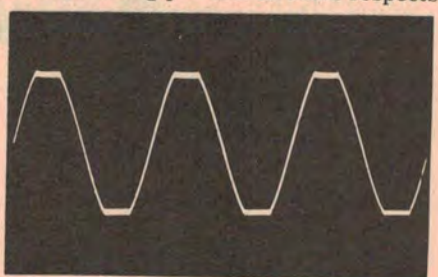


Figure 2. **An amplitude-limited waveform — "clipping".**

SPECIFICATIONS

Power output
100 W RMS into 8 ohms
(±55 V supply)

Frequency response
8 Hz to 20 kHz, +0 −0.4 dB
2.8 Hz to 65 kHz, +0 −3 dB
NOTE: These figures are determined solely by passive filters.

Input sensitivity
1 V RMS for 100W output

Hum
— 100 dB below full output (flat)

Noise
−116 dB below full output
(flat, 20 kHz bandwidth)

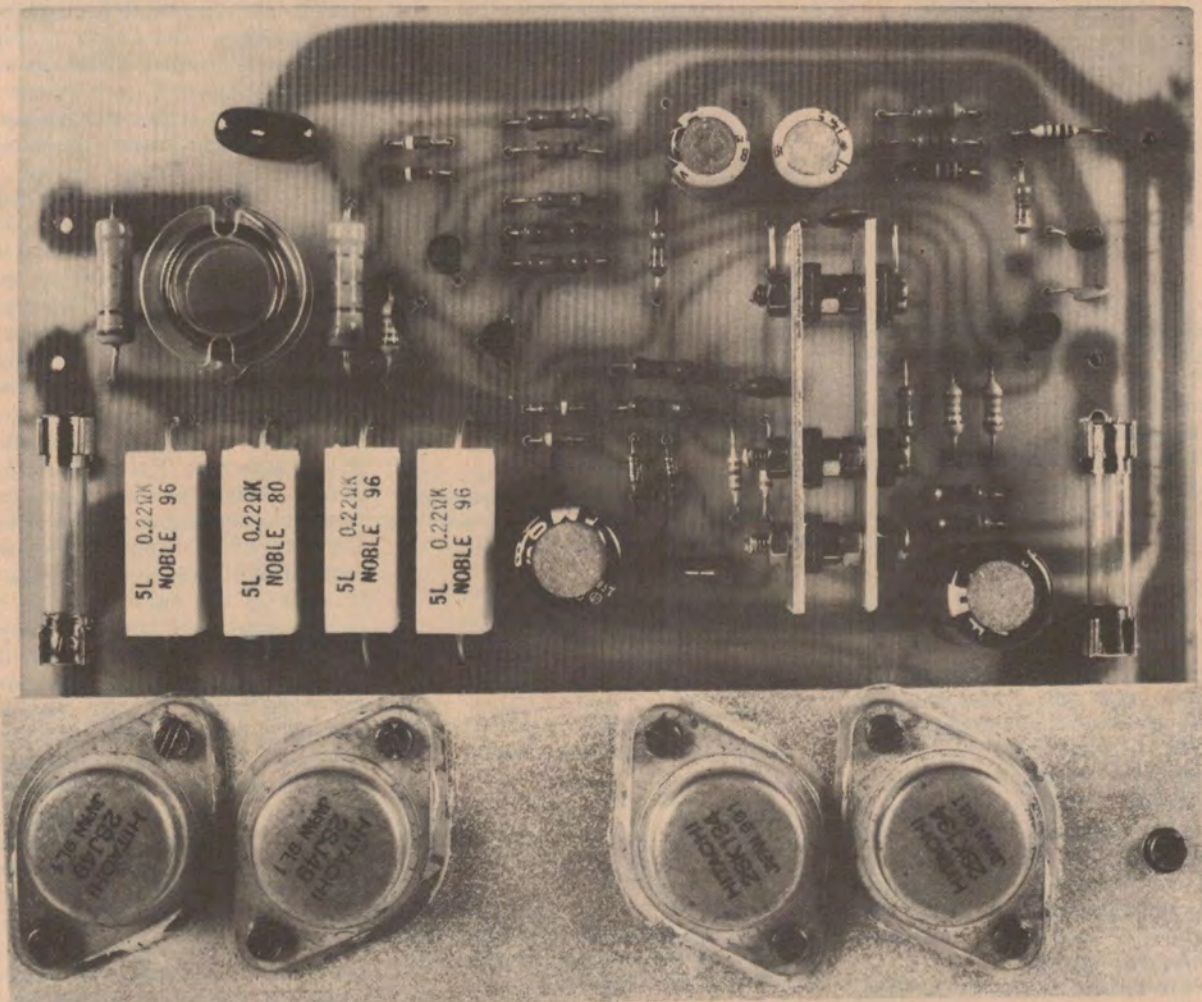
2nd harmonic distortion
< 0.001% at 1 kHz
(0.0007% on prototypes)
at 100 W output using a
±56 V supply rated at 4 A
continuous.
< 0.003% at 10 kHz and 100 W

3rd harmonic distortion
< 0.0003% for all frequencies
less than 10 kHz and all powers
below clipping.

Total harmonic distortion
Determined by 2nd harmonic distortion
(see above).

Intermodulation distortion
< 0.003% at 100 W.
(50 Hz and 7 kHz mixed 4:1)

Stability
Unconditional — see accompanying
oscilloscope photographs.



it is unfortunate that output power is measured using a continuous sine wave. This certainly tests the power amp power supply combination under continuous conditions but does not give any indication of the transient power capability. A modern, good quality record can easily cause transient signal peaks of the order of at least 20 dB above the average music level. A typical 50 watt power amplifier for example, with a supply voltage of approximately ±30 V unloaded could be driven into clipping by transients when the average music level is only 3 V RMS, i.e: slightly more than a watt into eight ohms. If on the other hand, the unloaded supply voltage is increased to ±50 V while keeping the loaded voltage the same as before (approx. 28 V) then the continuous power rating will still be 50 watts but the average music level before clipping is increased to 5 V RMS or 3 W into eight ohms. The difference between the continuous power output of an amplifier and its transient power capabilities is called *dynamic overload margin* or dynamic headroom and is given by the equation

$$\text{Dynamic Headroom (in dB)} = 10 \log \frac{P_T}{P_c} \quad \ldots\ldots\ldots\ldots (1)$$

where P_T is transient power (RMS) and P_c is the continuous power rating (RMS)

An amplifier with a good supply regulation like the first of the two amplifiers discussed above, will have a low dynamic headroom figure (approx 0.6 dB). The second of the two amplifiers with poorer supply regulation will have a higher dynamic headroom figure (approx. 4.4 dB), and could sound superior to the first amplifier. Of course, the poorer supply regulation would have to be taken into account when designing the amplifier. The supply rejection would have to be higher to ensure the same distortion characteristics as the first amplifier, and the output transistors must be capable of handling the higher supply voltage.

Crossover distortion

When a bipolar transistor is used as an emitter follower the relationship between the output and the input is a function of the load impedance and the forward transfer admittance of the output transistors. Specifically:

$$\frac{e_o}{e_i} = \frac{R_L}{(R_L + 1/y_{fs})} \quad \ldots\ldots\ldots\ldots (2)$$

where e_o is the output signal voltage
e_i is the input signal voltage
y_{fs} is the forward transfer admittance
and R_L is the load impedance.

It is the non-linear component of y_{fs} that causes distortion in the output stage. Equation (2) shows that if y_{fs} is large the value of $1/y_{fs}$ will be small and $(R_L + 1/y_{fs})$ will approach R_L. Therefore, for y_{fs} sufficiently large e_o/e_i will approach unity, and this is the ideal situation. ▶

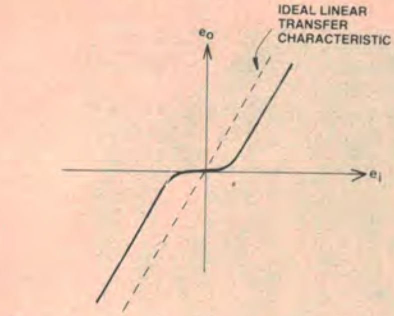


Figure 3. **Illustrating the relationship between e_i and e_o for a bipolar output pair operated without bias, as shown in Figure 1. The result is 'crossover distortion'.**

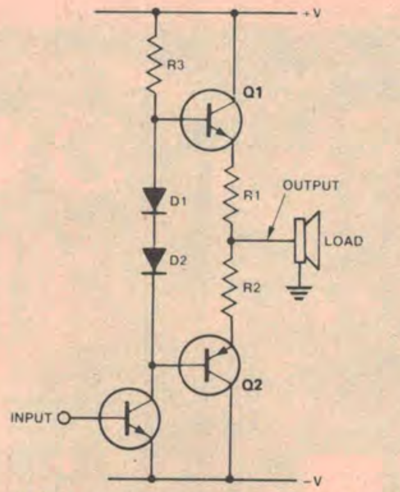


Figure 4. **A common method of linearising the relationship between e_i and e_o in an output stage is to apply bias using two diodes (D1 and D2).**

The problem with bipolar transistors is that, although their forward transfer admittance is high (approx. 40 Siemens for a typical output transistor and a current of 2 A) it drops dramatically if the base-emitter voltage drops below 0.6 V. In an output stage like that in Figure 1 the output signal voltage swings both positive and negative with respect to ground potential, with the transistor Q1 responsible for positive excursions and Q2 responsible for negative excursions. Whenever the voltage on the base of Q1 drops below 0.6 volts, or the voltage on the base of Q2 gets above −0.6 volts, (i.e: closer to 0 volts) the forward transfer admittance decreases rapidly, and the transfer characteristic of the output stage becomes grossly non-linear. This non-linearity produces crossover distortion (see Figure 3).

There are several methods commonly employed to overcome the problem of crossover distortion. Most make use of the concept of bias or quiescent current. With this technique, a fixed dc voltage of around 0.6 V is applied to the bases of the output transistors. In the output stage shown in Figure 4 this voltage is derived across the two diodes D1 and D2. If the diodes and the value of the resistor R3 are chosen correctly, then both output transistors are just turned on. With no signal voltage applied, the output of the stage is at 0 V so none of this dc current will flow in the load. Instead, this bias current flows directly from the positive to the negative rail and ac signal voltage is superimposed on this dc voltage. The base signal voltage must now reach −0.6 V to completely turn Q1 off. Since this region is now in the positive half cycle, Q2 has turned on and, with relatively high y_{fs}, will react essentially in a linear way to the input signal.

The same occurs when Q2 is turning off. It enters its low y_{fs} state in the region between 0 V and +0.6 V and being in the positive half cycle, Q1 will have high y_{fs} and maintain the linearity of the stage. The graph in Figure 5 illustrates the effect of bias current. The curves shown in Figure 3 have moved parallel to the e_i axis and are now closer to the ideal linear characteristics. Figure 6 shows actual CRO photographs of an amplifier with and without bias current applied. The bottom waveform is the distortion obtained simply by filtering out the fundamental frequency of the input sine wave. Note that the distortion waveform has peaks that correspond to points where the sine wave crosses 0 V.

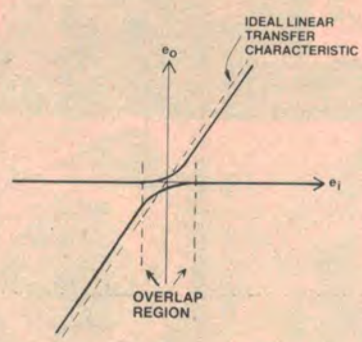


Figure 5. **How the application of bias to the output devices affects the relationship between e_i and e_o — the 'transfer characteristic'.**

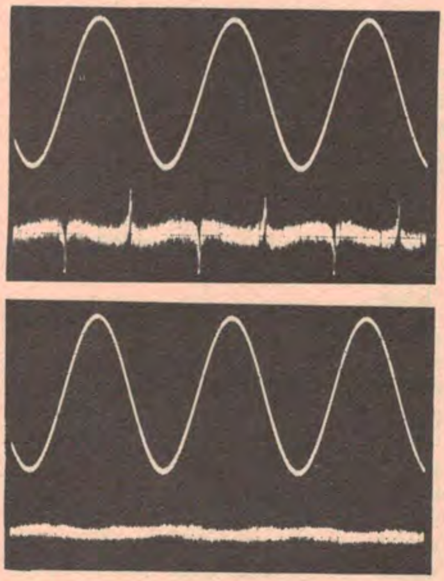


Figure 6. **Oscilloscope photographs taken from the ETI-477 module in operation: upper trace in each pic is output at 5 kHz, 10 V RMS; lower traces shows distortion analyser output. TOP: crossover distortion, reduced bias. BOTTOM: bias correctly set, distortion below resolution of analyser (0.003%!).**

The use of bias current to decrease crossover distortion has its disadvantages also. The dissipation in the output stage is increased, causing heating of the output devices. An amplifier with a 50 V supply and a quiescent current of 50 mA must dissipate 2.5 W in each of the output devices so the output stage will run warm even with no input signal. Furthermore, bipolar transistors have a positive temperature coefficient. If the base-emitter voltage is held constant but the output transistor temperature increases, then the bias current will increase due to a decrease in the emitter-collector resistance. This increase in bias current causes a further increase in temperature and consequently a further increase in bias current. This condition is called *thermal runaway* and if left unchecked will destroy the output devices. In practical power amplifier circuits the temperature is sensed by a temperature sensitive element, like another transistor or a diode, and the bias current is adjusted accordingly.

The positive temperature coefficient of bipolar transistors also causes another problem that limits the maximum power handling of the output transistors. Since it is impossible to ensure that the heating produced in the transistor chip is perfectly homogeneous, some areas of the chip will heat up more than others. These areas will decrease in resistance, conducting more current and heating further. This effect is called *secondary breakdown* and causes hot spots on the chip surface that can destroy the device.

Slew rate limit

The third source of non-linearity normally associated with the output stage is *slew rate limiting*. Just as the output stage is limited in its maximum output voltage it is also limited in the time taken to change from one voltage to another. The time taken for the output stage to swing over a certain voltage range is called the *slew rate* of the output stage. Furthermore since the output transistors have the biggest chip areas they are usually the slowest devices in the amplifier. If the signal slope (instantaneous rate of change of input signal voltage with respect to time) approaches the slew rate of the output transistors (or any other stage in the amplifier) distortion will be produced that is analogous to the distortion due to amplitude limiting. This distortion is sometimes called *transient intermodulation distortion* (TIM or TID) but it is important to realise that it is a slew rate limited phenomenon.

There are only two ways to eliminate this type of distortion, either by decreasing the signal slope of the input waveform or by increasing the slew rate of the output stage.

Decreasing the maximum signal slope implies decreasing the frequency response of the power amplifier. So if a good frequency response is to be obtained, the problem of slew induced distortions must ultimately be solved through the use of faster output transistors.

The MOSFET output transistor

The power MOSFET overcomes many of the problems discussed above. Hitachi are the first company to make available MOSFETs at a realistic price and with sufficient power handling for use in the output stage of audio power amplifiers. We have chosen the 2SK134 and the 2SJ49 devices for this project. These have a maximum power dissipation rating of 100 W, maximum drain to source voltage of 140 V and a maximum current of 7 A, which is a very formidable specification!

The first major advantage of MOSFETs over bipolar transistors is

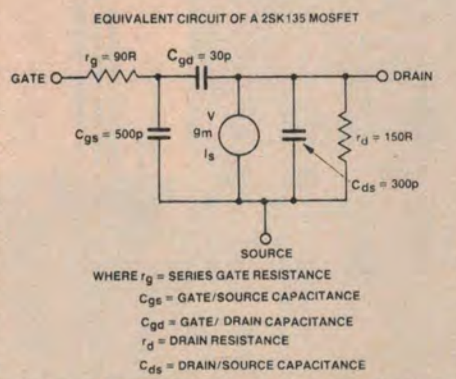


Figure 7. **Equivalent circuit of a typical power MOSFET (2SK135).**

their very high input impedance. Figure 7 shows an equivalent circuit for a typical MOSFET.

The gate appears as a 90 ohm resistance in series with a 30 pf capacitance to the drain and a 500 pf capacitance to the source. At dc, the input resistance is the resistance of the two effective capacitors — essentially an open circuit. The equivalent circuit also gives us insight into another of the MOSFET's great advantages. The combination of the series gate resistance and the total equivalent gate capacitance determines the cut-off frequency of the device at around 3 MHz! When driven correctly, the MOSFET is capable of excellent frequency response linearity and its slew rate is *unmatched* by any bipolar device of similar power. The speed of the MOSFET is attributable to the absence of an effect called *minority carrier storage* and it can therefore switch a current of 2 A in roughly 3×10^{-8} seconds or *30 nanoseconds!* This is around *100 times* the capability of most bipolar transistors.

This very fast response, coupled with the high input impedance and gate capacitances make the devices prone to oscillation, although they are not difficult to tame if care is taken with the pc board layout and a few fundamental precautions are taken. The best approach is to ensure that all gate wiring is kept as short as possible and to increase the value of the series gate resistance. This increases the $r_g C_{gs}$ time constant and limits the frequency response, greatly improving the device stability.

Figure 8 shows the frequency response of a typical power MOSFET and its relationship to the value of gate resistance. It is important that the distance between this resistance and the gate is kept to a minimum.

The extremely high slew rate of the MOSFET devices makes it possible to limit the maximum signal slope of the input signal while not affecting the frequency response of the amplifier inside the audio passband. In this way, the maximum signal slope cannot approach the slew rate of the output stage. Assuming no other stage in the amplifier slew rate limits this will overcome the problem of transient intermodulation distortion, but more about this later. ▶

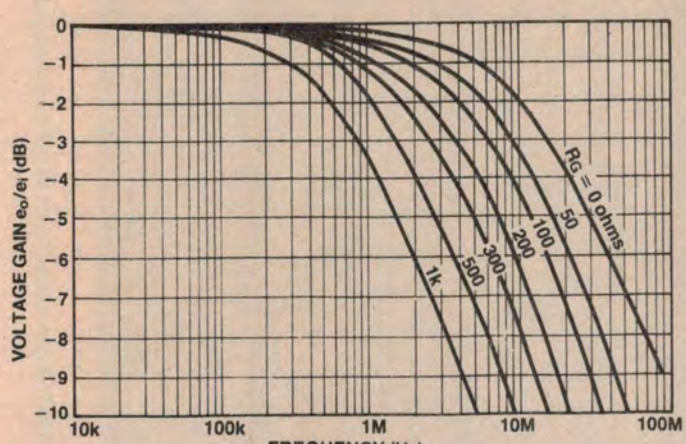
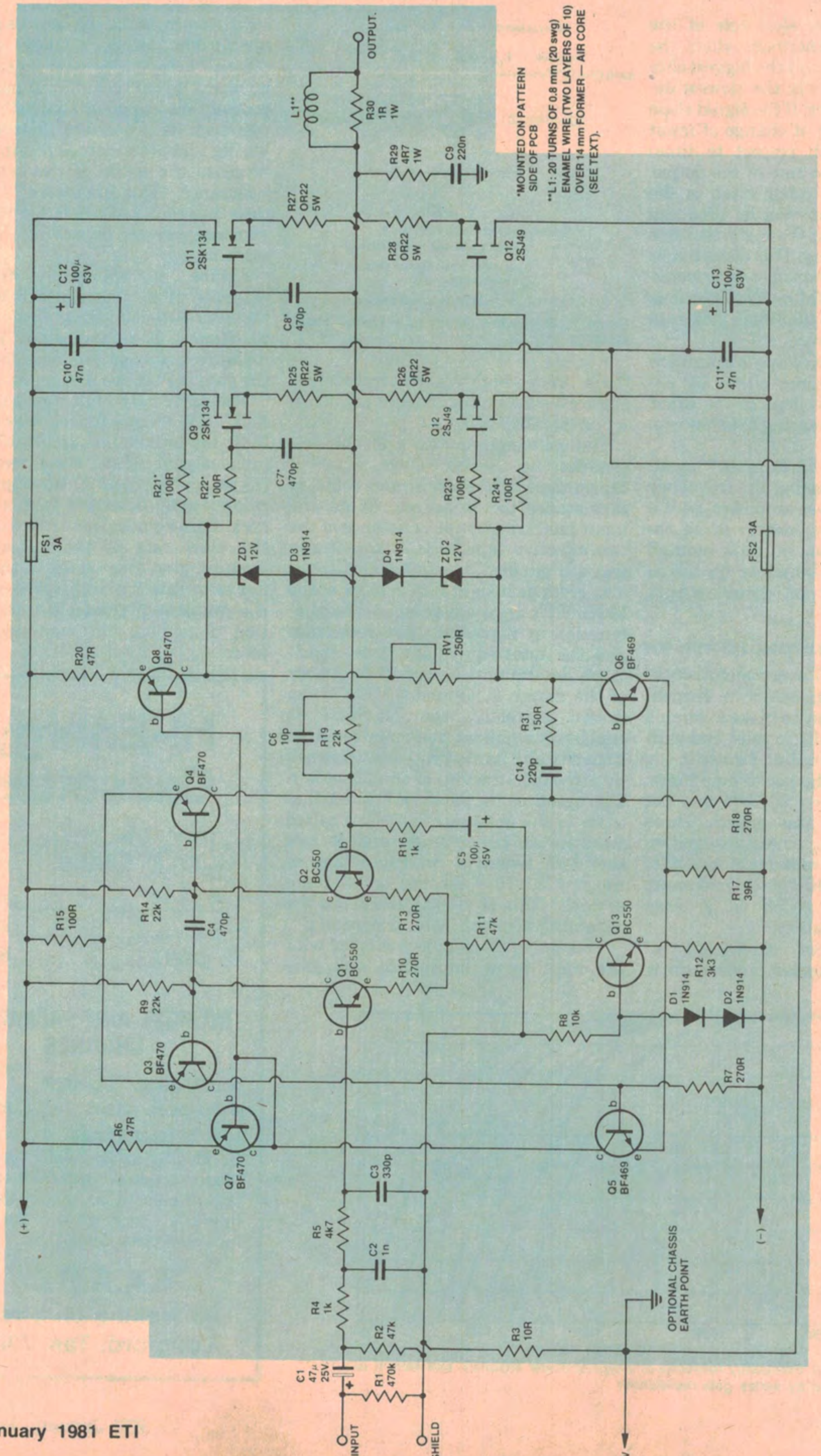


Figure 8. **Frequency response of a typical power MOSFET and how it is affected by series gate resistance.**



Circuit diagram of the ETI-477 MOSFET power amplifier module. A complete 'How It Works' description will be given next month.

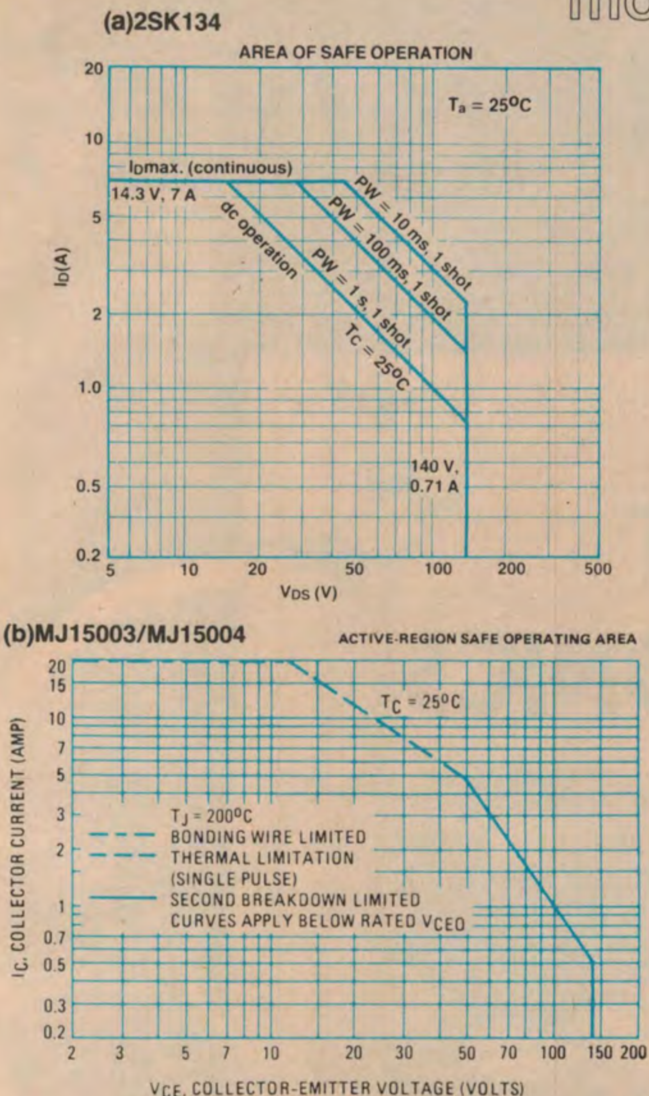


Figure 9. (a) SOAR curves for a 2SK134 power MOSFET. Compare with (b), the SOAR curves for a bipolar power transistor (MJ15003/MJ15004).

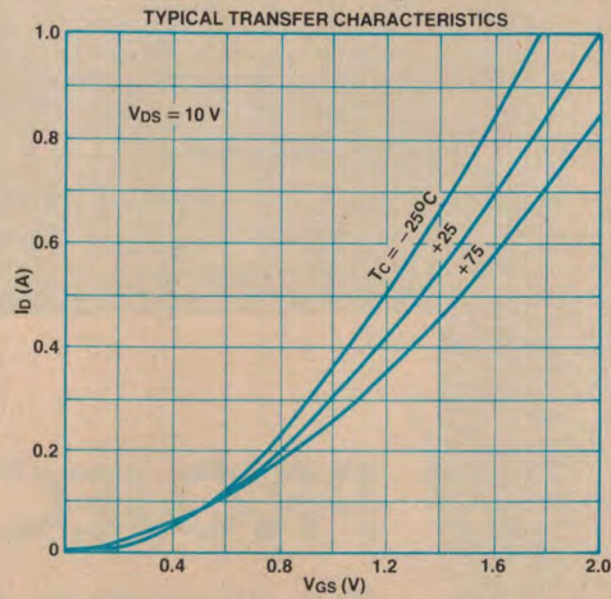


Figure 10. Typical transfer characteristics of a power MOSFET. Note that the greatest non-linearity occurs at low drain-source currents.

Another advantage of MOSFETs over bipolar transistors is their temperature characteristics. While the temperature coefficient of the bipolar device is *positive* the MOSFET has a *negative* temperature coefficient for drain source currents in excess of 100 mA. Heating of the devices causes an increase in the drain-source resistance and the current decreases. Furthermore, if one part of the chip surface heats more than any other, the increasing resistance in this area distributes current over the rest of the chip surface until the temperatures across the chip surface are equalised; so secondary breakdown is eliminated.

A look at the safe operating curves in Figure 9 shows a comparison between a MOSFET SOAR (Safe Operating ARea) and that of a good bipolar output transistor. Note that the bipolar has four limiting lines where the MOSFET has only three.

Crossover distortion and MOSFETs

It has been stated in a number of journals that one of the advantages of MOSFETs lies in the elimination of crossover distortion. Their argument relies on the fact that the variation in the forward transfer admittance of a bipolar transistor is exponential, while that of a MOSFET is more linear. The problem with this argument, as I see it, is that the MOSFET's greatest non-linearity still occurs for low drain-source current (see Figure 10) and certainly the Hitachi devices never achieve the high value of y_{fs} attainable with bipolar transistors. The specification for forward transfer admittance of the 2SK134 for example is approximately 1 Siemen, and this is only a fraction of the 40 S quoted earlier for bipolar devices. Remember that it is the non-linear component of y_{fs} that gives rise to distortion, and as a result, a MOSFET output stage with these characteristics could be expected to cause ten times the distortion of a bipolar design.

Although the bipolar turn-on characteristic is more severe, it is restricted to a smaller range of emitter current and once overcome by the application of bias current, the higher y_{fs} will actually yield a stage *with lower distortion*. The CRO photographs in Figure 6 were obtained using a MOSFET power amp and the crossover distortion is clearly evident.

In order to reduce crossover effects to satisfactory levels with these MOSFETs it is necessary to apply at least 100 mA of bias current, and for really good results approximately 200 - 300 mA would be needed. If the supply voltage is around ±50 V, each output device will dissipate five to ten watts with no input signal applied, substantially more than most bipolar output stages. This is not really a problem considering the MOSFET's negative temperature coefficient, but you should expect a MOSFET power stage to run warmer than bipolar output stages.

Another problem caused by the relatively low value of forward transfer admittance is the voltage drop between the gate and the source, which can be in the order of several volts increasing at high power levels; see Figure 10. The Hitachi devices have a maximum allowable gate to source voltage of 14 V and care must be taken in the design to ▶

ensure that this limit cannot be exceeded.

The minimum drain to source on-resistance for the Hitachi devices is around 1.7 ohms so that a drain current of 7 A continuous can be expected to cause a voltage drop between the drain and source of approximately 12 V. In order to get the same power as a bipolar stage a higher supply voltage is necessary to compensate for the higher voltage drop across the output devices. In order to make a power amplifier conservatively rated at 100 W into 8 ohms it is necessary to be able to deliver in excess of 28 V RMS to the load. This is equivalent to around 39 V peak. Adding the drain-source voltage drop of around 12 V gives 51 V, and allowing a margin for supply regulation of around 5% increases this to 56 V. Adding a further 20% for ac mains supply regulation implies that the output stage must be able to handle a supply voltage of around ±65 V. This is well within the maximum voltage specification of the 2SK134 and 2SJ49.

Examination of the SOAR characteristics of these devices reveals that it will be necessary to use two MOSFETs in parallel to achieve 100 W into 8 ohms and still not exceed the maximum power dissipation ratings of the devices. If we could guarantee that the amplifier would always be used with purely resistive loads the SOAR requirements could be relaxed substantially. If the amplifier had supply rails of ±50 V the maximum voltage swing across the load will be approximately ±40 V, giving a maximum load current swing of ±5 A, into an 8 ohm load. The maximum dissipation in the output devices will occur when the load current is around half the maximum current, i.e: 2.5 A and the voltage drop across the operating output transistor is approximately 30 V. So the power dissipation in the output devices would be less than 30 x 2.5 = 75 W.

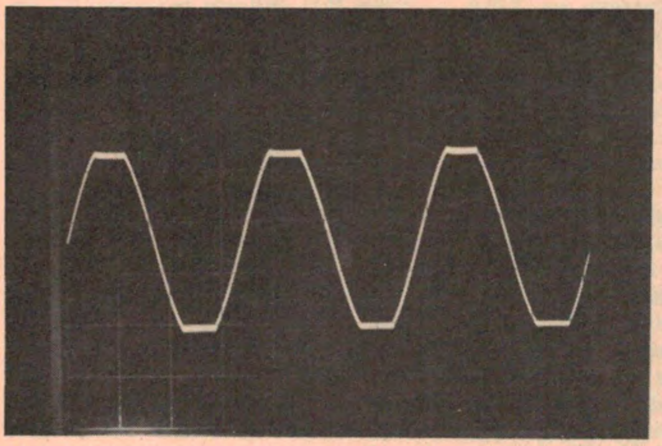
A single pair of output transistors would suffice.

Unfortunately, loudspeakers are not purely resistive loads. In some electrostatic loudspeakers for example, the amplifier load is actually the primary of a step up transformer, needed to supply the high signal voltage for the electrostatic elements. This can represent a highly inductive load and the output stage must be able to handle the associated phase shift. Similarly, it is not uncommon for the load to have a substantial capacitance, especially in loudspeakers with poorly designed crossovers. Under these conditions the charged capacitive or inductive reactance will supply energy back into the output stage. If, for example, in an amplifier with ±50 V rails an effective load capacitance is charged to the maximum negative voltage of, say, −40 V by a large negative going signal voltage, this potential will remain on the load when the output is subsequently driven to the maximum positive voltage of around +40 V. If the resistive component of the load impedance is not less than 8 ohms the maximum current in the load is now 10 amps. The worst case power dissipation in each half of the output stage will be around 5 A when the voltage drop across the operating output device is 40 V. The maximum power dissipation will therefore be around 200 W, so two pairs of output transistors will be necessary to ensure reliable operation.

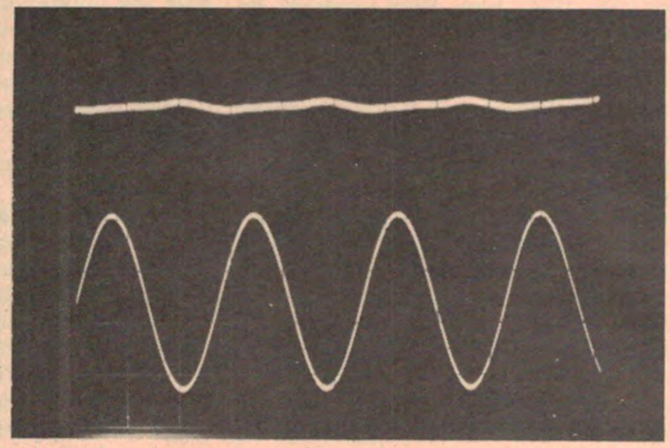
Since this problem is caused by the 'imaginary' (or reactive) component of the output load, these large signal currents will only exist momentarily while the load is charged or discharged to the new signal voltage. It is therefore possible for this load line to be marginally outside the dc safe operating area. Even taking this into account, a single pair of 2SK134/2SJ49s would not be sufficient. During the development of this power amplifier the output stage using two pairs of MOSFETs has been driven into hard overload, short circuits and even full power oscillation at over 10 MHz. Under these conditions the output device temperature was consistently measured in excess of 130°C. The MOSFETs are still performing perfectly, so these are extremely robust devices.

In summary, MOSFETs have both advantages and disadvantages when used in the output stage of an audio power amplifier. They are superior in speed and input impedance and are extremely robust. On the other hand, their higher distortion due to lower forward transconductance will necessitate an overall increase in the amount of negative feedback, so phase response will need to be carefully controlled to ensure stability. In general, the advantages outweigh the disadvantages, however, and it is for this reason that we have chosen these devices for the ETI-477 power module.

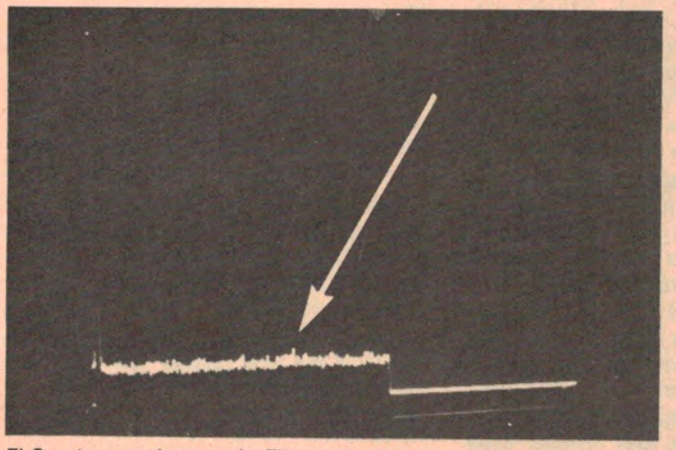
Discussion continued next month. Turn over for construction details for the module. ▶



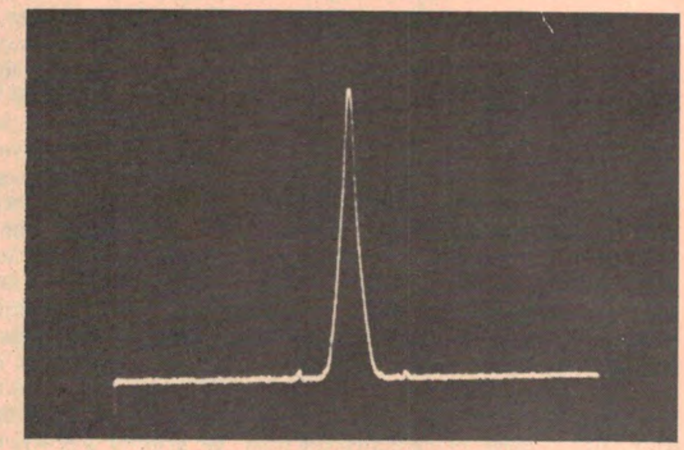
A) Overload recovery test. A 1 kHz sine wave input driving the module into amplitude overload (clipping). Note the amplifier remains stable when going into and coming out of overload.



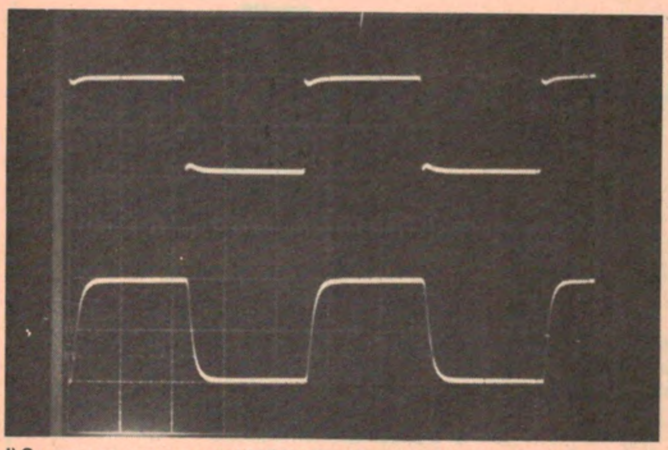
B) Total harmonic distortion measurement (AWA model F242A N. & D. meter). Lower trace is the 1 kHz, 10 W RMS output from the module. Upper trace shows output of the F242A, which in this case is at the limit of resolution (around 0.002% THD).



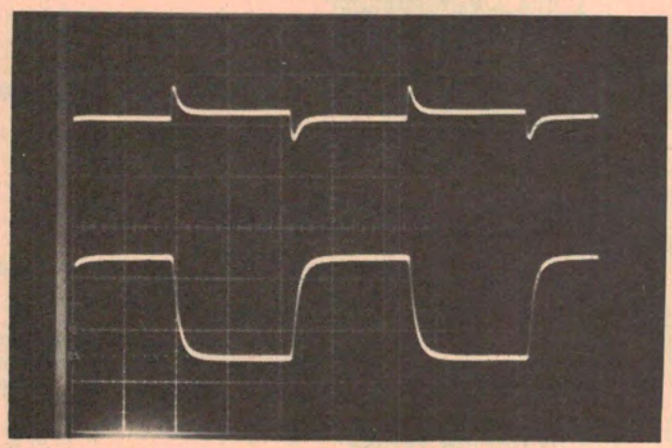
E) Spectrum analyser again. The peak on the left shows the fundamental: 20 kHz, 10 W RMS sinewave from the module. The second harmonic distortion is just visible above the noise (arrowed).



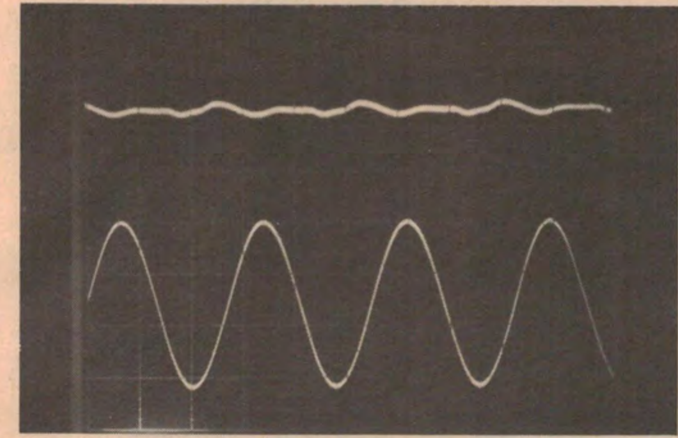
F) Intermodulation distortion proved as difficult to measure as THD, being below the resolution of most test equipment. A 50 Hz sine wave was mixed with a fundamental frequency in a 4:1 ratio. The fundamental was then varied over the audio range. Intermodulation products were not apparent for all frequencies below 7 kHz; i.e: less than 0.002%. This photo shows the IM products produced around a 7 kHz fundamental. Note they are just visible above the noise. This represents an IMD figure of around 0.004%.



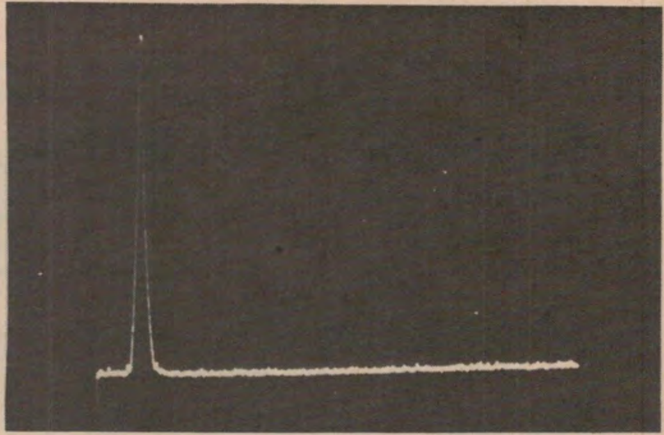
I) Square wave response at 10 kHz. Top trace is the input. The glitch after the rising and falling edges is due to a fault in the square wave generator. The harmonics produced, however, are well above the cutoff frequency of the input RC filter on the module. As a result, the output is a perfect band-limited square wave (lower trace).



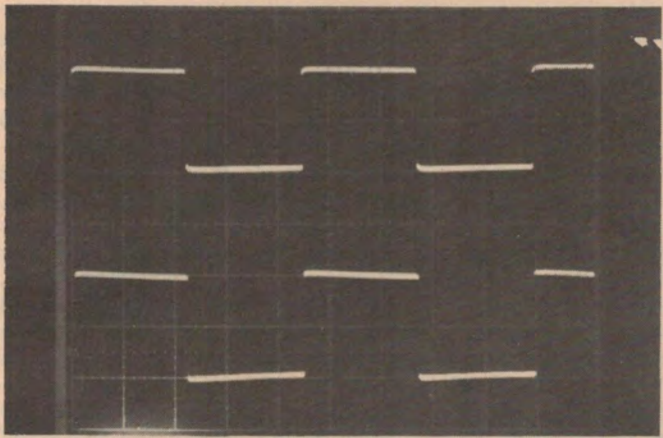
K) Oscilloscope photograph showing the error signal (top trace) in the negative feedback loop in response to a 10 kHz square wave drive producing 20 V p-p into an 8 ohm resistive load. Note that the error signal does not clip. This is a good qualitative indicator that the amplifier is free of transient-induced distortion. Scale for the error signal is 200 mV/division.



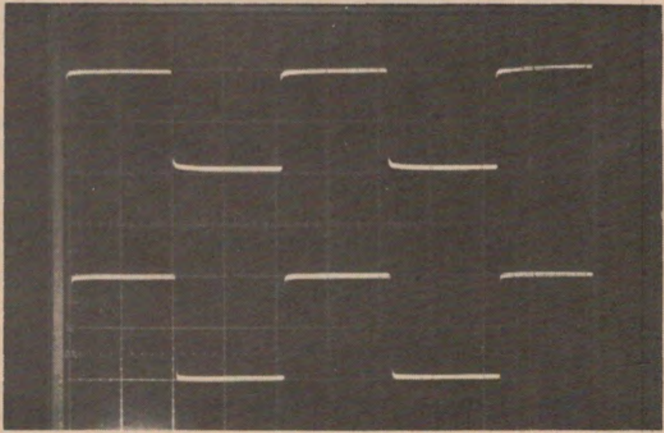
C) Total harmonic distortion, this time at 20 kHz, 10 W RMS output. The amplifier distortion is just becoming discernible above the resolution of the F242A. Note the difference between the distortion waveform shown here and that shown in B. THD here is around 0.004%.



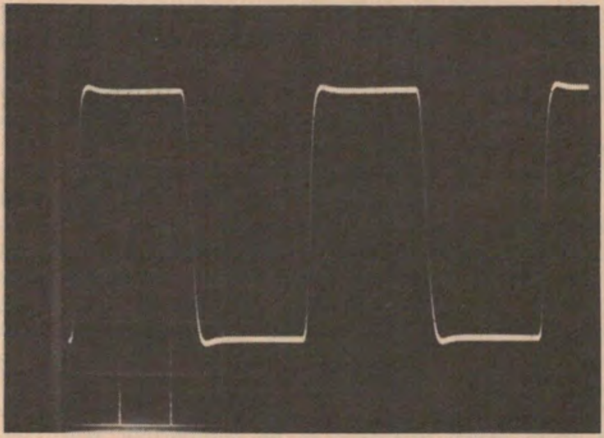
D) In order to measure the distortion products of the module it was necessary to use a Hewlett Packard 3580A spectrum analyser. This instrument can display a dynamic range of 90 dB on screen. The noise on the bottom of the trace here is around 0.002% of the fundamental. This photo shows the fundamental 1 kHz, 10 V RMS input from the module at far left. Notice that the distortion products are not visible above the noise. The THD/frequency curve shown elsewhere was obtained by fitting passive notch filters to the input of the 3580A analyser to increase its sensitivity. The limit of resolution of this technique obtained in ETI's laboratory is around 0.0003%, being the distortion generated by our AWA G233 sine wave oscillator!

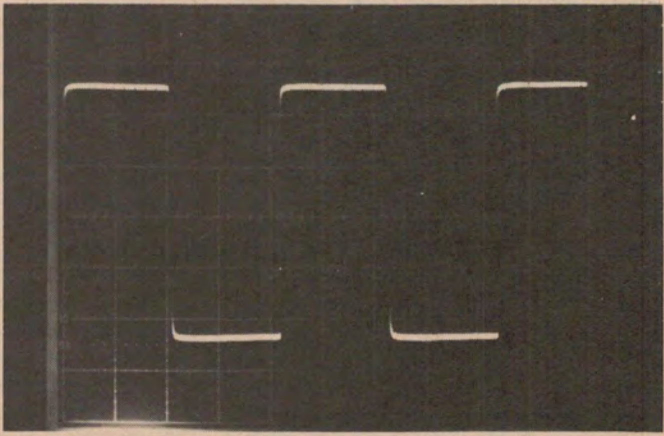


G) Square wave response of the ETI-477 module. Top trace is the 100 Hz input. Bottom trace is the resulting 20 V p-p output into an 8 ohm resistive load. The slight tilting of the output square wave occurs because of the high pass filter on the module's input and is therefore not a fault.



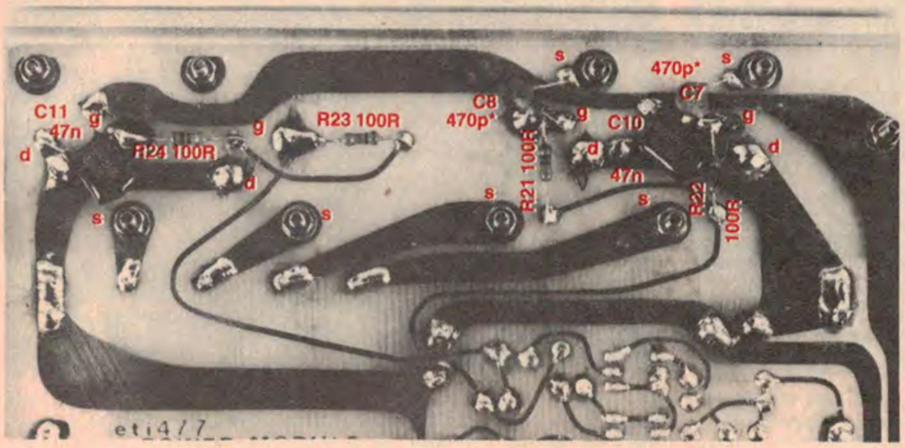
H) Square wave response of the ETI-477 with a 1 kHz input.



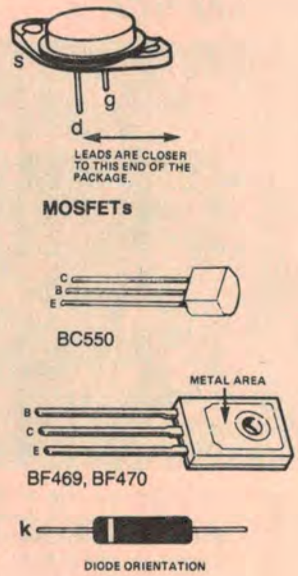


J) Oscilloscope photograph showing the module's performance into a reactive load. At left is the output waveform of the module, driven with a 10 kHz square wave. Output load is 2 uF in parallel with 8 ohms. Note that there is no sign of oscillation or instability. This is a very strenuous test as normally the reactive load would exhibit a series resistance which limits the charge and discharge times for the capacitance.

The photo on the right shows the output waveform from the module, again driven with a 10 kHz squarewave, the load this time being a 3 mH inductor. Again, the amplifier is totally stable. ▶



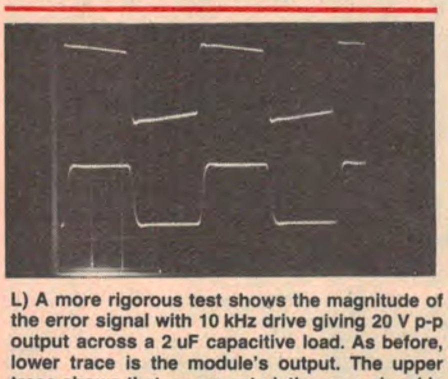
Overlay for the copper side of the pc board showing components mounted on this side.



Construction

The construction of the power amp module is not difficult since all the components are mounted on a single pc board. Since the design employs a fairly large amount of negative feedback, the pc board pattern is a critical factor in attaining the maximum theoretical performance. It would be virtually impossible to achieve the same performance if the pc board pattern were altered, without recourse to a distortion analyser with a sensitivity of at least 0.005% and a very good spectrum analyser. The pc board pattern shown ensures freedom from earth path interaction and therefore does not degrade the distortion performance of the design — but more about that next month.

Commence construction by soldering all the resistors onto the circuit board. The OR22 (0.22 ohm), 5 W source resistors in the output stage get warm if the amplifier is operated for extended periods at high power. They should never get hot enough to burn the circuit board, since any fault capable of causing this much power dissipation should blow the supply fuses first. Nevertheless, it is good construction practice to space these resistors a few millimetres off the surface of the board. The 4.7 ohm, 1 W resistor R29 should *definitely* be spaced off the board since it will overheat if a fault condition should cause oscillation of the amplifier at high frequencies. Do not mount the four 100 ohm resistors R21, R22, R23, R24 at this stage. These are mounted on the rear of the circuit board and are best left until after the MOSFETs are mounted.



L) A more rigorous test shows the magnitude of the error signal with 10 kHz drive giving 20 V p-p output across a 2 uF capacitive load. As before, lower trace is the module's output. The upper trace shows that, as expected, the error signal is much greater than with a resistive load, but still does not clip. This could safely be considered the worst realistic load from the point of view of TIM production. Scale for the error signal is again 200 mV/div.

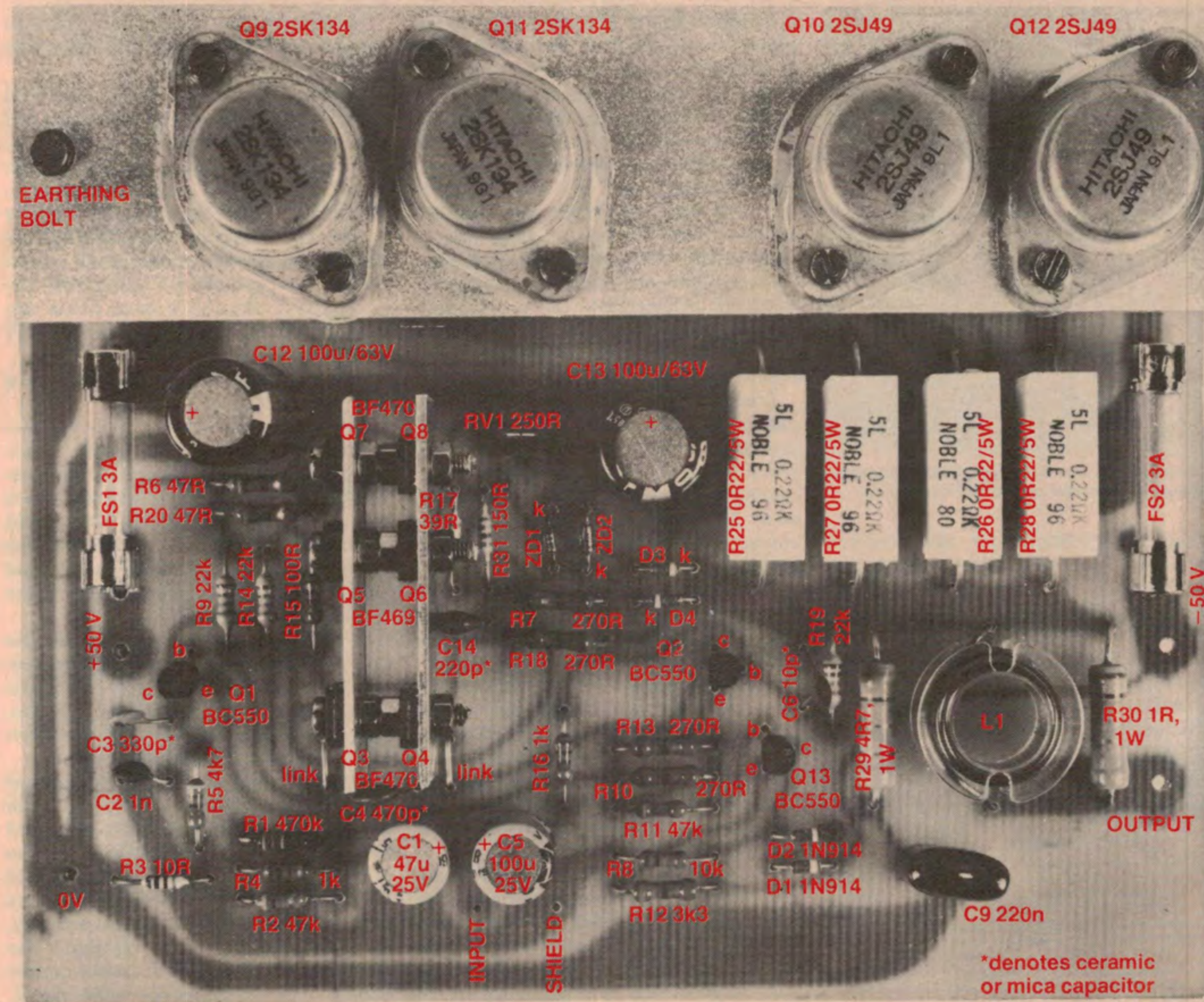
Solder the four pc board fuse clips into the board next. Now mount all of the capacitors, with the exception of C7, 8, 10 and 11. Once again, these mount on the rear of the board. Make sure the electrolytic capacitors C1, C5, C12 and C13 are inserted with the correct orientation as these are polarised components. Mount the 1N914s and zener diodes, taking care to orient them correctly. Solder the trimpot RV1 into place and then the small-signal transistors, Q1, Q2, and Q13.

Next step is to mount the six voltage amp transistors, Q3 through Q8. These are situated on the pc board in two parallel rows, each row with three transistors. In the prototype modules, these heatsinks were constructed from two pieces of aluminium, as can be seen from the photographs. The transistors are mounted using 6BA bolts, each passing through a pair of transistors. This forms a very strong assembly which can then be soldered onto the pc board. Insulating mica or plastic washers should be used between the metal side of the transistors and the heatsink strip, using a small quantity of heatsink compound between each mating surface. When this transistor-heatsink assembly is completed, but before soldering it into the circuit board, check that each transistor is effectively insulated from the heatsink. Using a multimeter on the resistance range, check for shorts between the centre lead (collector) of each transistor and the heatsink strip. Note that the bolts through the six transistors are automatically insulated from the metal rear of the transistor by the plastic body of the device so no additional insulation of the bolts should be necessary.

Before mounting the MOSFET output devices it is necessary to make the heatsink bracket. This is cut from a suitable aluminium extrusion. The pc board has been designed to suit extrusions with one of the sides at least 40 mm wide. The transistor mounting holes have been placed so that the heatsink brackets used in the ETI-466 300 W module are compatible, although there will be some unused holes.

If you are making your own heatsink bracket, drill the holes according to the drilling template and make sure that no aluminium chips or burrs remain around the holes. This is best done with the use of an oversize drill bit (about 13 mm). A couple of twists with the drill bit will put a slight chamfer around the hole and remove any rough spots.

The extrusion used really needs to be selected to be compatible with the particular heatsink that suits your application. Next month we will use two of these modules as the basis for a high quality stereo power amplifier with the ▶



Overlay for the component side of the pc board. Artwork for the pc board appears on page 113.

PARTS LIST — ETI 477

Resistors all ½ W, 5%

R1 .... 470k
R2, R11 .... 47k
R3 .... 10R
R4, R16 .... 1k

R5 .... 4k7
R6, R20 .... 47R
R7, 10, 13, 18 ...... 270R
R8 .... 10k

R9, 14, 19 .... 22k
R12 .... 3k3
R15, 21 - 24 .... 100R
R17 .... 39R

R25 - 28 .... OR22, 5 W
R29 .... 4R7, 1 W
R30 .... 1R, 1 W
R31 .... 150R
RV1 .... 250R trimpot

Capacitors

C1 .... 47u, 25 V electro
C2 .... 1n greencap
C3 .... 330p ceramic or mica
C4, 7, 8 .... 470p ceramic or mica
C5 .... 100u, 25 V electro
C6 .... 10p ceramic or mica
C9 .... 220n greencap
C10, 11 .... 47n greencap
C12, 13 .... 100u, 63 V electro
C14 .... 220p ceramic or mica

Semiconductors

D1, 2, 3, 4 .... 1N914 or similar
ZD1, ZD2 .... 12 V, 400 mW zener
Q1, 2, 13 .... BC550
Q3, 4, 7, 8 .... BF470
Q5, 6 .... BF469
Q9, 10 .... 2SK134
Q11, 12 .... 2SJ49

Miscellaneous

ETI-477 pc board; four pc mount fuse clips; two 3 A type 3AG fuses; one plastic bobbin (from P26/16 potcore, or similar); one metre of 0.8 mm dia. enamelled copper wire; two strips of 20g aluminium, each 15 mm wide by 47 mm long (for voltage amp heatsink — see text); 155 mm length of 40 x 12 mm aluminium extrusion for heatsink bracket (see text); assorted nuts, bolts, hookup cable etc.

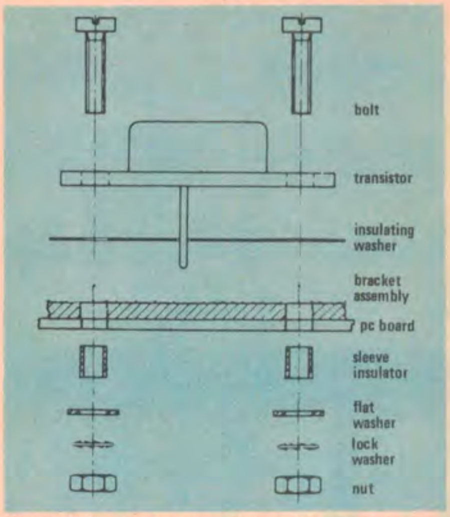
Price estimate

We estimate that the cost of purchasing all the components for this project will be in the range:

$62 - $68

(excluding heatsink and power supply)

Note that this is an **estimate** only and **not** a recommended price. A variety of factors may affect the actual price of a project, whether bought as separate components or made-up as a kit.



General diagram for mounting a TO3-cased device to a heatsink bracket and pc board assembly.

final specifications for the heatsink bracket. We will also discuss the problem of power supplies and the special precautions that should be taken to ensure good earthing to obtain maximum performance from the modules.

After the heatsink bracket has been drilled, the MOSFETs can be mounted onto the pc board. The bracket is held in place by the output devices and an 'earthing' bolt that connects the bracket to the 0 V rail (see overlay photo). The bolts holding the MOSFETs in place make the electrical connection to the source of each device, which is connected internally to the case. The bolts must be insulated from the heatsink bracket. Use a piece of spaghetti or heatshrink tubing cut to length such that the bolt will nowhere touch the heatsink bracket (see the accompanying TO-3 assembly diagram). Slip these into the holes in the heatsink bracket before assembling the MOSFETs.

Smear heatsink compound on one side of each of four mica or plastic TO-3 insulating washers and put them in place on the heatsink bracket. Smear heatsink compound on the under side of each MOSFET and put each in the correct place and secure them with bolts.

The output assembly should now be checked for shorts. Remove the earthing bolt first. The resistance between the case of each MOSFET and the bracket should be checked with a multimeter. If any device shows a short to the bracket it should be disassembled and the short found. Usually it is necessary to replace the TO-3 insulating washer as most faults of this type are the result of small metal burrs cutting through the washer when mounting the device.

Once the MOSFETs are mounted, the last passive components — resistors R21, R22, R23 and R24 plus capacitors C7, C8, C10 and C11 can be mounted on the rear of the circuit board. These are positioned on the rear of the board so that lead length is kept as short as possible. Cut the leads just short enough to mount the components in place. The accompanying photograph shows a close-up of these components on one of the prototype modules.

Set-up procedure

The recommended supply voltage for the modules is around +/− 55 V. With this voltage and reasonable supply regulation, the module will deliver around 100 W RMS into a nominal 8 ohm load. The power supply will be dealt with in more detail next month, but before applying power to the modules the following set-up procedure should be carried out.

First, re-check that the output devices are not shorted to the heatsink bracket. This is best done with the earthing bolt removed as mentioned earlier. If no shorts are found, replace the earthing bolt.

Do the same check for shorts between the six voltage amp transistor collectors and their heatsinks.

Check the polarity of all polarised components. It is often difficult to tell one end from the other on diodes since the markings are easily rubbed off. If in doubt, check these with a multimeter. Wind the wiper of the trimpot RV1 fully *counterclockwise* (least resistance). This ensures no bias is applied to the output stage. Now, remove the fuses from the pc board if they have been fitted and replace them with 10 ohm, ½ W resistors.

The module can now be connected to a power supply.

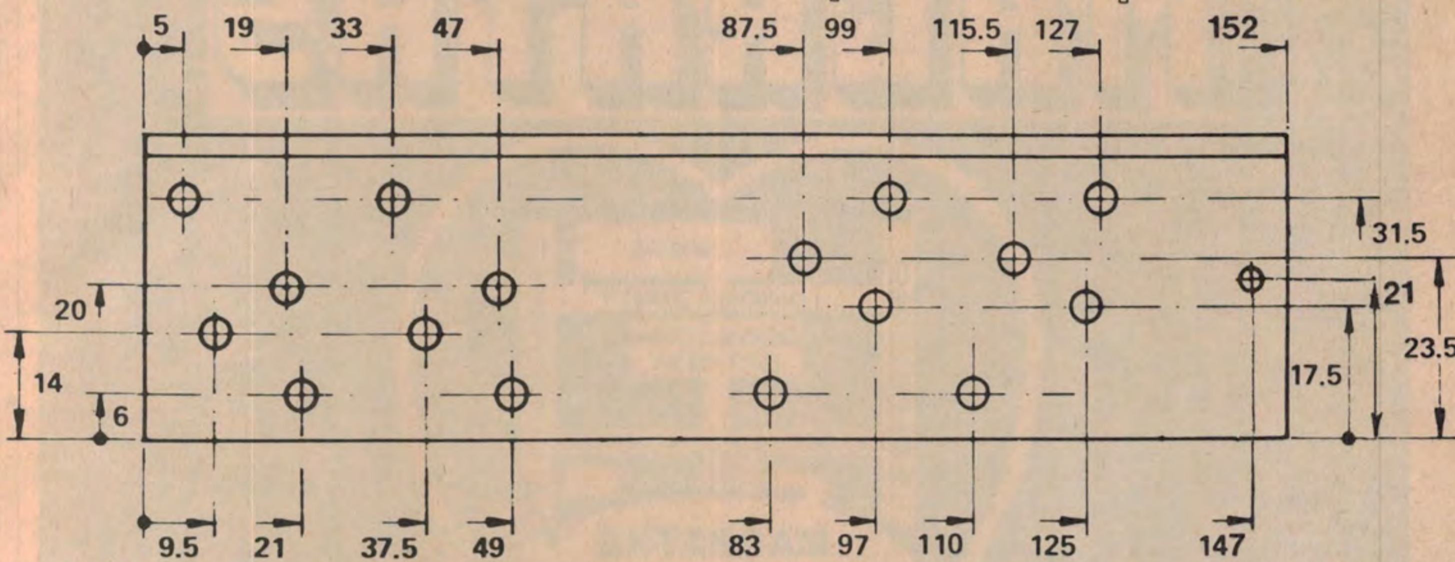
Make sure the power supply connections are sound, with good solder joints. If you have access to a current-limited bench supply it is best to connect the module to this for the set-up and initial test. If you can do this, set the current limit to around 200 mA. *Do not* connect a load to the output of the module at this stage.

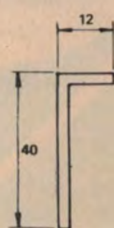
If the power is now turned on, the current through the two 10 ohm resistors replacing the fuses should be low. If these resistors start to smoke, this indicates a fault condition — turn the power off immediately.

If all is well, connect a multimeter across the 10 ohm resistor in the positive rail fuse holder and slowly wind the trimpot RV1 clockwise until the voltage measured is 1 V. This will set the bias current in the output stage at 100 mA. If the current sets up correctly, measure the voltage between the speaker output and 0 V on the power supply. You should see around ±25 mV. If you only have an analogue multimeter, this voltage may be too low to measure; in this case it is sufficient to show that the output is at 0 V.

If there is a fault in a direct-coupled amplifier like this, the output will usually be driven hard toward one of the supply rails and this is the reason the load should not be connected until these initial tests are done. Remember that 50 Vdc across an 8 ohm load equals a power dissipation in excess of 300 W, which would instantly destroy any loudspeaker!

If the module passes all these tests, it is safe to replace the fuses and connect a load. Make sure the power is off before removing the 10 ohm resistors from the fuse holder and allow time for the power supply electrolytics to discharge. There is 100 V between the fuses and *this is sufficient to cause electrocution.* Be *careful* when working with high power amplifiers. ●





ALL 4mm DIA

MATERIAL 40 x 12 x3 ALUMINIUM ANGLE EXTRUSION

Drilling details for the heatsink bracket assembly. All dimensions are in millimetres. Suitable aluminium angle stock is available from Alcan Handyman stores.

Next month we describe how to attach a power supply so as to achieve the performance we obtained along with a complete description of how to build a stereo power amplifier. This will use a heatsink as a front panel, manufactured exclusively for ETI — don't miss it.

View of the voltage amp transistors and heatsinking assembly. In the prototypes we used two 20g strips of aluminium, each 47 mm long by 15 mm high. This is the minimum size we would recommend and brackets measuring 50 mm long by 30 mm high are preferred. Centre-to-centre drilling dimensions can be taken from the pc board (page 113), measuring between the collector pins of each transistor.