80-100 WATT MOSFET AUDIO AMPLIFIER

A three-part article on the design and construction of a modern, high-power amplifier begins with a description of the problems of amplifier design in relation to the characteristics of 'vertical' power mosfets. A matching, modular preamplifier design will follow.

by John L. Linsley Hood

The problem with designing audio amplifiers is that there are a number of design requirements which are impossible to satisfy completely: such things as freedom from harmonic and intermodulation distortion; independence, in terms of distortion or transient response, on the nature of the load reactance; freedom from spurious (amplifier-generated) signals over the whole range of signal inputs and likely load characteristics; rapid settling time and freedom from 'hang-up' on step-input or overload, particularly under reactive load conditions; and complete absence of input-signal or load-induced instability.

Not only are these requirements impossible to achieve absolutely, but the work needed to improve one of these may simultaneously bring about a worsening in other respects, so part of the task of the designer is to choose, within the appropriate limits of cost and complexity, between conflicting possibilities and requirements. No two designers (or their commercial or advertising managers) are likely to come to the same balance of compromises in these respects, and this leads to subtle differences in the tonal characteristics of the designs.

A characteristic of commercial trends in the last twenty years, which I view with regret, is an overwhelming concentration on the attainment of very low harmonic-distortion figures over the whole of the audible spectrum, to the extent that many modern commercial designs attain steady-state t.h.d. figures fifty or more times better than possible, under any conditions, from the signal sources which feed the amplifiers. A similar amount of effort is expended commercially in achieving very high signal-to-noise ratios – which would be valuable if it were matched in the handling of programme material by the programme producers.

The reason for this commercial interest is a simple one. the major emphasis in most equipment reviews is placed on t.h.d. and s/n ratio, coupled with, in the case of power amplifiers, power output in watts/pound (sterling) or, occasionally, watts/pound (avoirdupois). This trend would be wholly praiseworthy if it could be achieved without impairment in other desirable characteristics of the equipment: unfortunately, it cannot. If one wants some quality very good, one must accept some others relatively bad! If only one knew which ones were important to the listener, this choice would be easy, but one doesn't. Quite a lot of work has been done in the field of psycho-acoustics to try to characterize the effects on the listener of specific electrical defects, but this work is far from complete and impaired, from the point of view of the designer, by the omission of most of the minor performance defects practical amplifier designs are heir to.

Nevertheless, a predictable result of the accumulation of experimental findings on acoustic effects, coupled with a greater awareness on the part of designers of the existence of residual performance shortcomings, is that there is a keen interest in new developments in components and circuit techniques, as a possible route to improved performance.

Of these new component developments, one of the most interesting, in the field of active devices, has been the growing availability of rugged, reliable and reasonably priced power mosfets (metal-oxide-silicon field-effect transistors). These devices have a very much better h.f. response – almost embarrassingly so – than the normal audio power transistor, and allow a considerable extra freedom in solving h.f. loop-stability problems, where some compromise must always be reached in a feedback amplifier design between the conflicting requirements of gain (or phase) margins and the need to retain a high loop gain at the upper end of the audio spectrum to achieve a high degree of steady-state linearity. In addition, these devices are almost completely free of the charge-storage effects found in junction transistors, which tend to impair complex-signals transfer.

Unfortunately, power mosfets have electrical characteristics and circuit requirements which are very different from those of the junction power transistor, so that they cannot be used as a direct replacement for junction transistors in existing designs. One must reappraise the circuitry.

Power mosfet

Insulated-gate field-effect transistors, of the type shown in outline in Fig. 1(a), and which operate by means of a mobile layer of charge induced in an otherwise non-conducting region of a semiconductor, have been known and used in small-signal applications for many years – particularly in v.h.f. circuitry, where their very fast response times are of great value. However, the conducting path in these devices is, by the nature of their method of construction, parallel to the surface of the semiconductor element. It is difficult, although some semiconductor manufacturers have achieved this in an endeavour to avoid restricting patents, to make this conducting path sufficiently short to achieve a low enough resistance for larger signal use.

The technical breakthrough in this type of device came about when it was appreciated that a 'V' or 'U' groove etched through the junctions in a fairly conventional transistor gave the possibility of an insulated-gate, induced-charge f.e.t., in which the current flow would be 'vertical. (as in the conventional junction transistor) rather than 'lateral. (in relation to the surface of the chip) as in the normal insulated-gate component. This gave a method of manufacture of 'mosfets', as

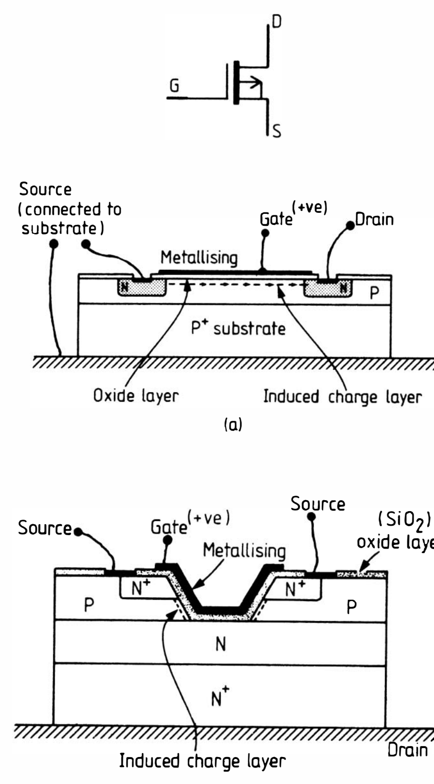


Fig. 1. *Small-signal, n-channel, insulated-gate f.e.t. of 'lateral' construction is shown at (a), while at (b) is the vertical power mosfet, in which the conducting path is a great deal shorter.*

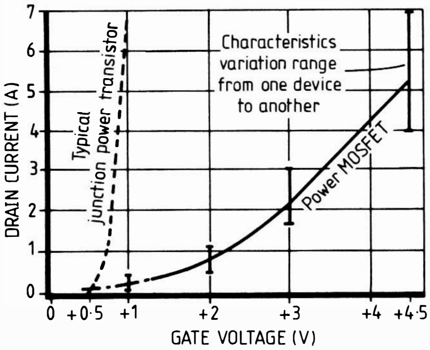


Fig. 2. *D.c. operating conditions of typical power mosfet compared with those of a junction power transistor.*

these devices are now almost universally known, which was open to any manufacturer of epitaxial planar junction transistors with the necessary skills in mask manufacture to fabricate the large number of parallel-connected igfet gates on a single chip, which are needed to lower the conducting resistance and increase the effective mutual conductance (g_m). A typical construction for such a vertical – or power – mosfet is shown in Fig. 1(b), though the proliferation of such designs within the past few years makes the concept of a 'typical' construction progressively less tenable. However, they do all have in common the parallel connexion of a large number of elements, which makes the mask design more complex, and the potential manufacturing reject rate and cost relatively high in comparison with the larger power junction transistors.

The electrical performance, under d.c. conditions, of a power mosfet, is shown in Fig. 2, with a superimposed curve from a junction power transistor added to the graph to draw attention to the differences in performance. Two features are immediately obvious from this graph – that a significantly higher forward voltage applied to the gate of the mosfet is necessary to obtain an adequate, and adequately linear, operating current, and that the mutal conductance of the power mosfet (about 2A/V in its linear region) is very much lower than that of the bipolar junction transistor (which can be in excess of 15A/V, or many hundreds of amps/volt in the case of Darlington-connected pairs).

In conventional audio-amplifier design, as it has become established over the past 20 years, the 'architecture' normally employed in the circuit is a low-power voltage amplifier, usually operated in class A, with as high an a.c. gain as is practicable without the use of an inconvenient number of gain stages, followed by an impedance-converter stage – usually a push-pull pair of compound emitter followers, forward biassed into AB operation, with an operating point chosen so that the mutual conductance of the pair of emitter followers is close to that which will be given by one half, alone, when operating in its linear region. Negative feedback is then applied from output to input to improve the overall linearity and other operating characteristics of the amplifier.

This configuration gives satisfactory bandwidth and linearity, and allows high power outputs with low quiescent thermal dissipation. The main drawback in this system is that there are invariably some low-level residues of crossover distortion, which increase in magnitude at higher frequencies, as the open-loop gain of the class A amplifier decreases – mainly as a result of the added h.f. loop-stabilizing components. This problem is worsened by the loss in loop gain through the output emitter-following stage, in which the gain is always less than unity.

To a first approximation, the output impedance of an emitter follower is $1/g_m$, which would give a gain in the output stage of $Z_L/(1/g_m+Z_L)$ at low frequencies. However, there is a further loss of gain at higher frequencies due to the limitations in turn-on and turn-off times of the transistors, so it is customary to use two or more transistors in a compound configuration in each half of the emitter follower, partly to sustain a low output impedance, and partly to allow the 100% negative feedback within the emitter-follower group to force improvements in the internal h.f. characteristics.

Inevitably, therefore, a complementary pair of output source followers using power mosfets, having a maximum g_m of some 2A/V, will perform less well in terms of linearity with a comparable class A amplifier stage than a pair of compound emitter followers with an effective (overall) g_m of, at least, some 100A/V.

Two solutions are available to this problem, of which the first is simply to incorporate the power mosfets in a compound source-follower circuit, with one or two small-signal bipolar transistors to increase the internal loop gain and reduce the dynamic output resistance and non-linearity of this stage. This provides a very simple answer to the difficulties introduced by the low g_m of the power mosfet, while allowing full advantage to be taken of the advantages of these devices (freedom from 'secondary breakdown', simpler output stage overload protection, and much better h.f. characteristics leading to lower overall t.h.d. and better transient response). I have shown a practical audio amplifier circuit using mosfets in the output stage configuration of Fig. 3(a), in another place[1].

For relatively low-power use, up to say 50 watts, this type of output stage is entirely satisfactory, and gives a good performance without calling for unconventional circuit design. However, there is a practical limitation in higher power use in that the forward voltage drop in each compound half cannot be less than the forward gate voltage (ref. Fig. 2) added

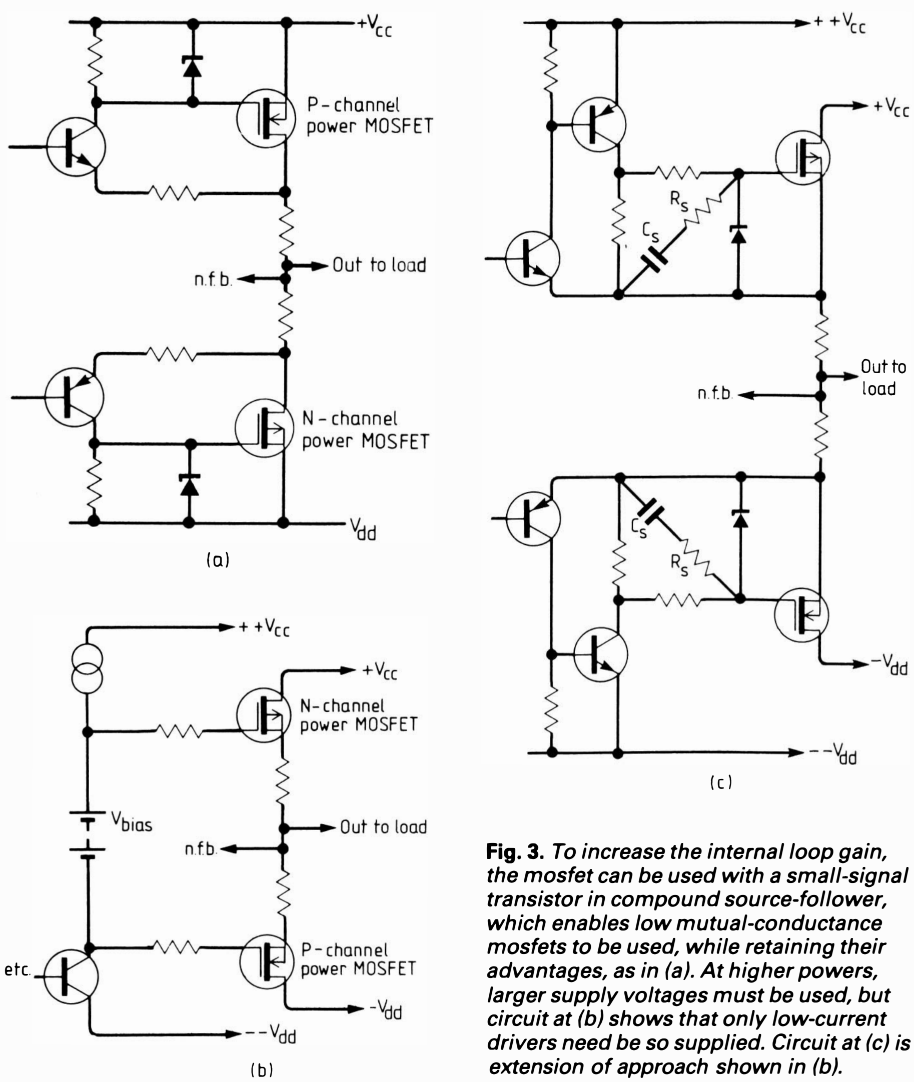


Fig. 3. *To increase the internal loop gain, the mosfet can be used with a small-signal transistor in compound source-follower, which enables low mutual-conductance mosfets to be used, while retaining their advantages, as in (a). At higher powers, larger supply voltages must be used, but circuit at (b) shows that only low-current drivers need be so supplied. Circuit at (c) is extension of approach shown in (b).*

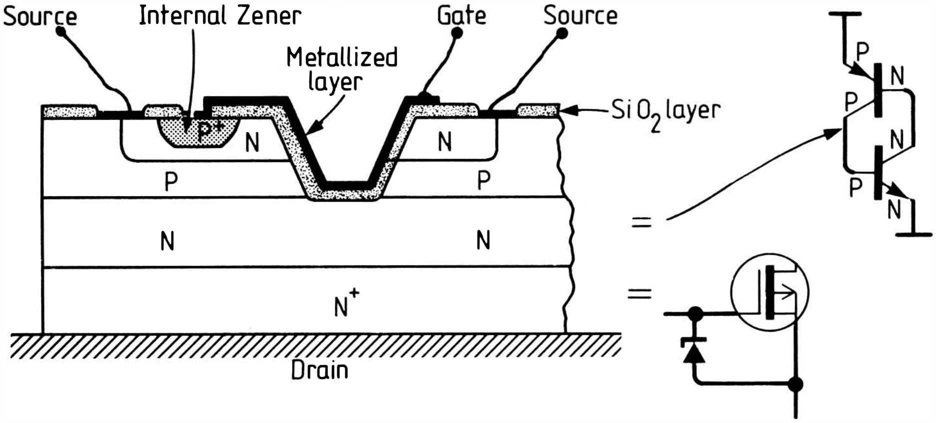


Fig. 6. *Internal protective Zener diode can cause inadvertent thyristor action, in which gate loses control.*

to the saturation voltage (V_{ce})of the driver transistor, even though the necessary drain-source voltage of the mosfet for this output current may be less than this. This leads to the need for higher supply-line voltages, with a consequent increase in the cost of mains transformer and smoothing capacitors.

In the very simple complementary-mosfet output stage of Fig. 3(b), the driver stages can be supplied from a higher voltage line without so much of a cost penalty, since the supply currents required by the driver stages are comparatively small.

This advantage can be retained by the use of the circuit arrangement of Fig. 3(c), while still allowing a very high effective g_m of the compound emitter follower, and a high level of internal negative feedback. The problem with this circuit is that it is no longer unconditionally stable within its internal feedback loop, and h.f. stabilizing components such as C_s and R_s need to be added to achieve the desired overall gain and phase margins – an elaboration which is unnecessary in the simpler arrangement of Fig. 3(a).

The other solution to the difficulty of the lower effective transfer ratio of the simple mosfet source follower of Fig. $2_{(b)}$ is to increase the gain of the class A amplifier stage, and this approach is explored below. Meanwhile, there are some other potential pitfalls in the use of power mosfets which need consideration if a workable and reliable design is to be put together.

Specific problems with mosfets

Although the power mosfet is, in its normal method of construction, equivalent to a bipolar junction transistor with its base and emitter joined together, and is therefore immune from the problem of 'secondary breakdown' (the funnelling of emitter current through diminishing areas of the base-emitter junction and consequent localized overheating and damage) it does suffer from other problems which are unique to itself. Of these, the first and most immediate is that the gate insulation layer, an oxide film formed on the surface of the silica, is less than 0.0001in (2.5 microns) thick, and will break down if the voltage between the gate and the source exceeds some 10-20 volts – depending on the device manufacture. Since the time delay involved in this breakdown, which will destroy the device, is likely to be very short, the circuit must be designed to protect the gate against even very brief voltage excursions beyond this limit.

This difficulty can be lessened, in the construction of the device, by incorporating a Zener diode between source and gate, as shown in Fig. 4. However, this technique in its turn leads to the problem that the device must then be protected against a reverse bias – of the order of 0.6 volts – which would cause this internal Zener to conduct, since this can sometimes lead to the triggering of a thyristor-type action within the mosfet, in which the gate is irrelevant. This may not destroy the device, but may damage associated circuit elements. The simplest form of protection is the use of an external germanium diode, connected in parallel with the gate/source Zener, and arranged to conduct before the internal diode. This is not a preferred solution, however, since the reverse insulaton resistance of the Ge diode is much poorer than the unmodified input resistance of the mosfet, and is non-linear with voltage. The circuit of Fig. 3(a) is immune from this problem.

The second difficulty in the use of power mosfets arises from the very high operating frequencies possible with these components. This leads to an effective circuit element of the form shown in Fig.

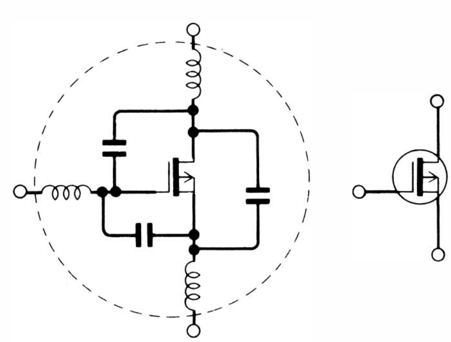


Fig. 5. *At high frequencies, stray capacitances and inductances, shown at (a), turn mosfet into an oscillator, with damaging effect.*

5(a), when the user expects the device to behave as in Fig. 5(b)! This causes immediate high-frequency oscillation, with frequently destructive effects, when such mosfets are incorporated into apparently sensible circuit configurations, and since the resultant burst of oscillation probably occurs in the 200-1000MHz range – and is brief anyway – it is unlikely that it will be seen on any monitoring instrument. The unhappy experimenter is then left contemplating a defunct device, thinking that its sensitivity to static electrification is so great as to render it unusable.

Happily, the internal gate-source capacitance is sufficiently high, typically in the range 600-1500pF, that stray static charge is unlikely to induce an electrical breakdown of the gate insulation. This internal capacitance, which must not be overlooked in circuit design considerations, also provides a convenient means for taming the h.f. behaviour of the transistor, sicne an external 'gate-stopper' resistor can then cause a predictable roll-off in h.f. response, to bring the unity gain transition frequency down to a more manageable level. An external resistor in the range 470R-4k7 is normally adequate.

Given these precautions, my experience is that power mosfets are at least as durable as normal bipolar power transistors, and allow a substantial improvement in circuit performance for a relatively small extra cost.

References

1. Linsley Hood., J. L., *Hi-Fi News and Record Review*, Vol. 25, No. 12 (Dec. 1980) pp. 83-85.
www.keith-snook.info

To be continued

LITERATURE RECEIVED

Power semiconductors and d.c. power supplies are detailed in the 1982 edition of the Lambda catalogue. It includes some application notes and dimensional drawings which makes it useful as a handbook. Lambda Electronics Co, Abbey Barn Road, High Wycombe, Bucks. **WW 400**

Some literature that you would like to receive may be out of print. A service to help you find that elusive book is provided by **The Out of Print Book Service,** 17 Fairwater Grove East, Cardiff, CF5 2JS. They do not charge a fee but ask that all requests for the service should be accompanied with a postage stamp at the current first class letter rate and full details of the book required. **WW 401**

Ceramic chip capacitors for high frequency applications are described in a bulletin from Hy-Comp Ltd, 7 Shield Road, Ashford Industrial Estate, Middlesex TW15 1AV. The bulletin lists their stability, specifications and terminations. **WW 402**

80-100W MOSFET AUDIO AMPLIFIER

One solution to the problem of low mosfet g_m – using the mosfet with bipolar small-signal devices – was described in the first part of this article. Here, John Linsley Hood presents an alternative, which is to improve the gain of the voltage-amplifier stage.

by J. L. Linsley Hood

A considerable amount of development work has been done over the past decade in the design of high gain class A amplifier stages, mainly in the evolution of integrated circuit operational amplifiers, in which the design requirements in this application are high gain, good immunity from supply line breakthrough, and wide bandwidth. In addition, some attention has been paid to cost-effectiveness which, in discrete component terms, means the use of active elements to provide the best performance for the lowest number of components.

Although this work has been done, largely, by the designers of i.c. op-amps, it is profitable for those working on discrete-component circuitry to look over their shoulders to see what they are up to.

The Class 'A' amplifier stage

It is a fundamental requirement of any feedback amplifier design that it should be absolutely and unconditionally stable, and this requirement applies with as much force to audio amplifiers of this type as to closed-loop servo-mechanisms. This consideration, coupled with the need for audio-amplifier designs to operate satisfactorily with a wide range of load reactances, leads to designs in which there is a substantial gain or phase margin at the unity-gain point on the Bode plot.

This requirement is easy to satisfy in a single-stage amplifier, of the type shown in Fig. 6, where it is unlikely that the internal phase shift, without feedback, will exceed 90° until frequencies are reached at which the gain is very much less than unity. However, the stage gain from such an amplifier will only lie in the range 50-250, with a conventional resistive collector load. It is true that gains of several thousand can be obtained from such a stage if the collector load resistor is replaced by a constant-current source, as in the 'Liniac' configuration[2] and this figure can be increased towards the hundred thousand mark if the amplifier transistor is operated in cascode with a device such as a junction fet, to increase its output resistance. Unfortunately, such high gains are only obtainable at very low collector currents, which imply very high output impedances and a relatively poor h.f. performance, which would not allow an adequate loop gain at the upper end of the audio band.

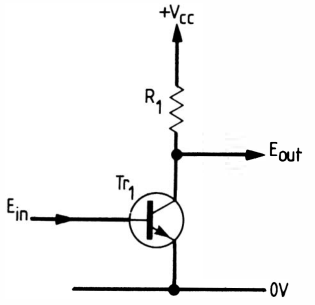


Fig. 6. Basic single-stage amplifier. At reasonable collector currents, stage gain is modest, but phase shift is less than 90° at frequencies below f_1.

The use of a two-stage voltage amplifier gives a much greater degree of design freedom, and while such an amplifier may not automatically guarantee, under all load conditions, that the internal phase shift does not approach 180° until the open-loop gain is negligible, the necessary conditions for an adequate phase margin, at unity gain, are much easier to contrive in this type of design than in circuits with more amplifying stages. For this reason, the possibilities of two-stage voltage amplifiers have attracted the attention of many designers working in the audio and operational-amplifier fields, where large amounts of negative feedback are deliberately employed to improve bandwidth, transient response, and linearity.

Because of the relatively low input impedance of Tr_2, a simple two-stage amplifier of the type shown in Fig. 7(a) will not give a gain which is that of a single stage, squared, but nevertheless a stage gain in the range 2000-3000 can be obtained, and this can be increased by a further factor of ten if a constant-current source is used as the load for Tr_2 instead of the load resistor R_2. Moreover, this order of stage gain can be achieved with collector currents in Tr_2 which are high enough to allow a relatively low output impedance at Tr_2 collector. In practice, the input transistor will most commonly be replaced by a long-tailed pair, as in Fig. 7(b), which reduces the odd harmonic distortion introduced by the input stage, as shown by Taylor[3,4]. This arrangement also facilitates the design of direct-coupled amplifier systems, in which the long-tailed pair input cancels the voltage offset otherwise introduced by Tr_1.

Circuit configurations of this type, using one or other of the arrangements of Fig. 7, with the constant-current source sometimes replaced by a bootstrapped load

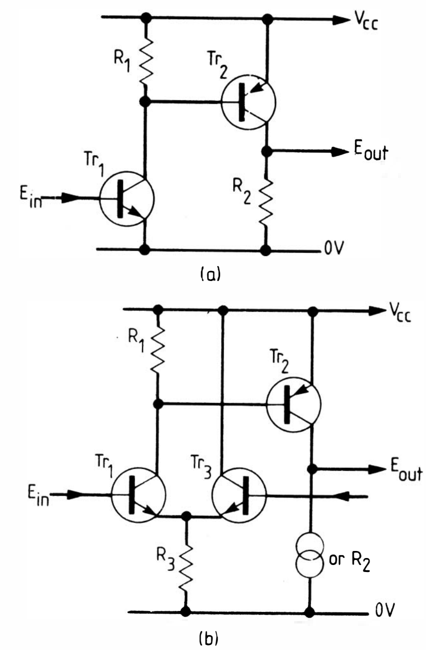


Fig. 7. Much higher gain is obtained from two-stage circuit at relatively high Tr_2 collector currents. Long-tailed pair input at (b) reduces odd harmonics.

resistor, formed the bulk of class A gain stages used in commercial audio amplifiers up to a few years ago. In these designs, the output-impedance transformation was accomplished by a complementary, or quasi-complementary, compound emitter-follower pair, using bipolar junction transistors, and overall negative feedback would be taken from the output to the inverting input connexion. With careful design, circuits of this type give overall harmonic distortion figures of 0.01-0.02%, as measured at 1kHz, and at a few dB below clipping point. Although the t.h.d. figures will worsen somewhat at lower power output levels, the signal-to-noise ratio of such an amplifier is only some 80-90dB, which means that the residual distortion is soon swamped by circuit noise.

The approach has changed somewhat in the last few years with the marketing of amplifiers, mainly of Japanese origin, having distortion levels in the 0.002-0.008% bracket, apparently in an attempt to obtain more impressive consumer reviews. In these designs, changes have been made in both the output stage biassing arrangements to minimize crossover discontinuities, and also in the number and complexity of the preceding gain stages. Where the improvement in

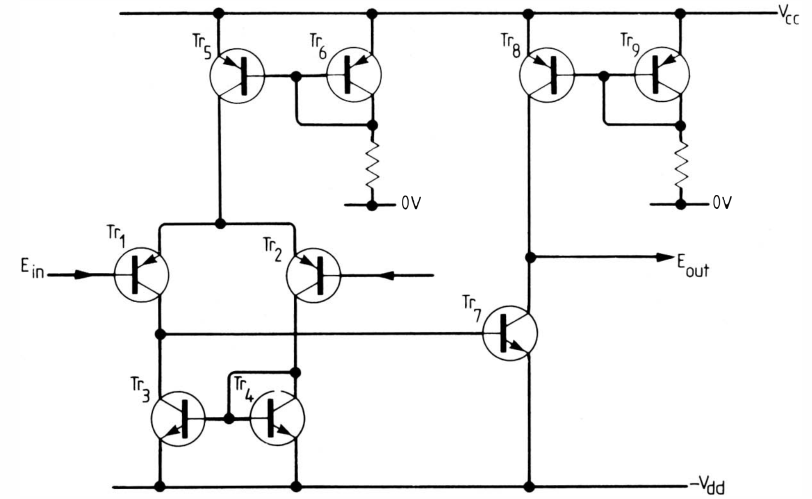


***Fig. 8.** Use of current mirror as load for Tr_1 offsets lower gain of long-tailed pair over single transistor. Process is repeated at output stage.*

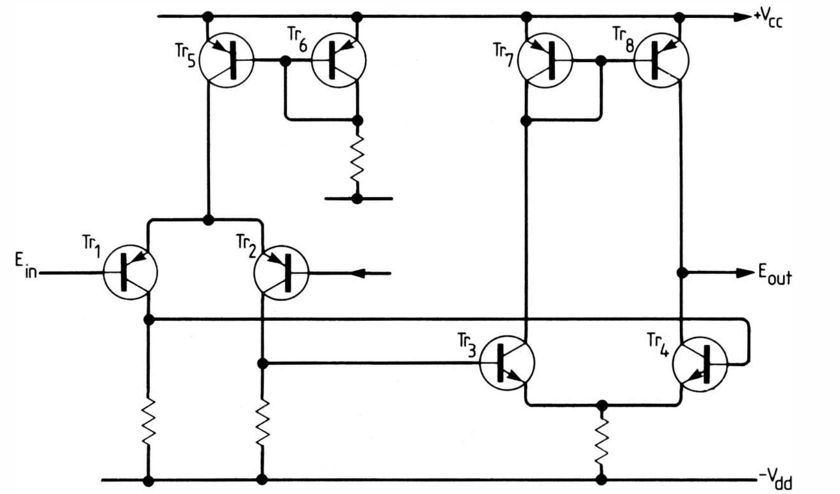


***Fig. 9.** Symmetrical, push-pull version of Fig. 7(a) circuit, with current-mirror load for Tr_4.*

linearity has been made at the expense of stability margin, or has led to a less ideal response to a step function input, the final result may be less good in the ear of the listener. Nevertheless, there is one trend amongst those currently being explored which is of interest, particularly since it follows techniques developed earlier for operational-amplifier integrated circuits, and I propose to consider this in greater detail.

Higher-gain configurations

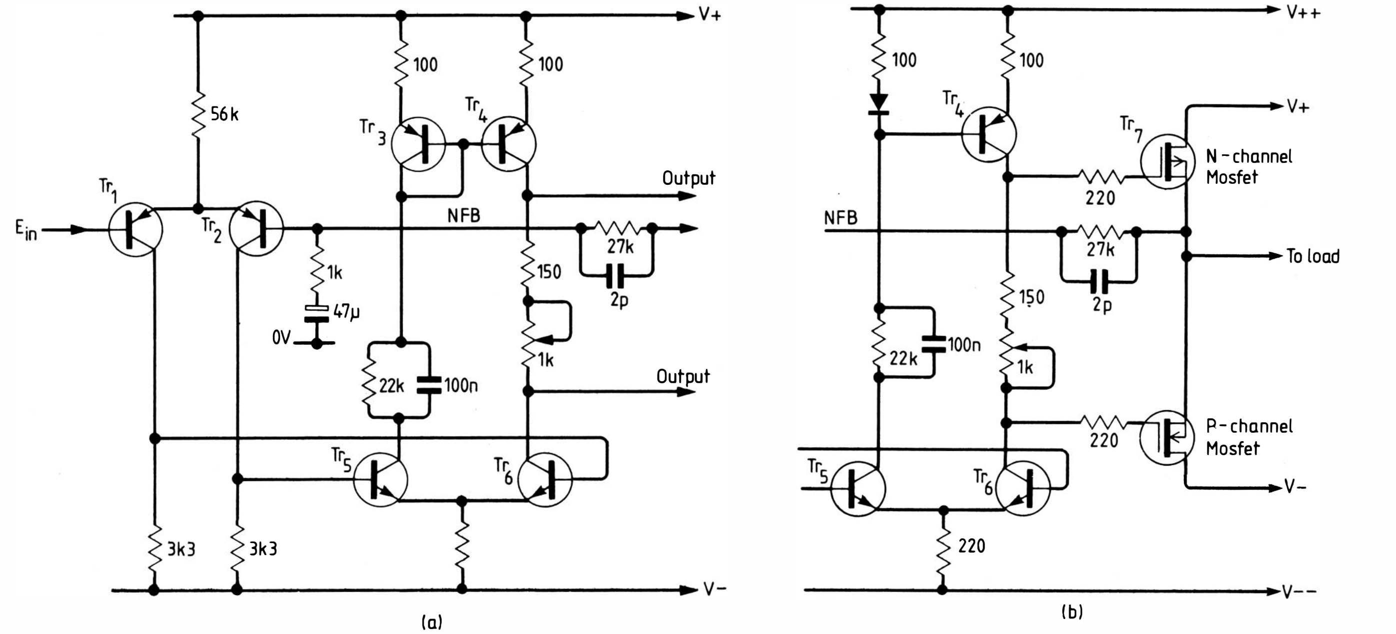
One of the penalties in the use of a symmetrical-input long-tailed pair is that the g_m of the input device is halved, with a consequent reduction in the gain in comparison with a similar amplifer having only a single input resistor. This loss of gain can be minimized by the use of an asymmetrical input stage, with the feedback transistor operated at a higher current than the gain transistor, as in the earlier design of my own[5], provided that the loss of the inherent d.c. balance is within tolerable limits.

A better method of avoiding this loss of gain, and very commonly used in i.c. operational amplifiers such as the 741, is to use a current mirror as the load for the amplifying transistor, with the operating current being set by the feedback one. This type of arrangement is shown in Fig. 8, and when used with a constant current load – normally yet another current mirror – on the output of Tr_7, will give stage gains of the 50,000-100,000 range at low frequencies. Although this approach is easier to adopt in 'single-chip' fabrications, where the simultaneous fabrication of all the transistors will ensure identity of the base-emitter turn-on characteristics necessary for proper current-mirror operation, it has also been used in discrete-component designs.

This circuit employs nine transistors, and while the actual transistor count may be of little moment in an integrated circuit where the fabrication of active components presents no difficulty and little extra cost, it is nevertheless interesting to observe that a better stage gain can be obtained using the arrangement of Fig. 9, which uses one less transistor. In this circuit, the two-stage amplifier remains a symmetrical push-pull version of that shown in 7(a), but with the current-mirror active load moved back to the output of the second stage. This particular version of the two-stage amplifier was used first, so far as I can discover, in the National Semiconductors LH0001 low-power operational amplifier. Low-frequency small-signal voltage gains of up to 200,000 are possible using this layout.

The good gain and phase characteristics of this particular circuit arrangement have

***Fig. 10.** Two commercial circuits using form of Fig. 9.*



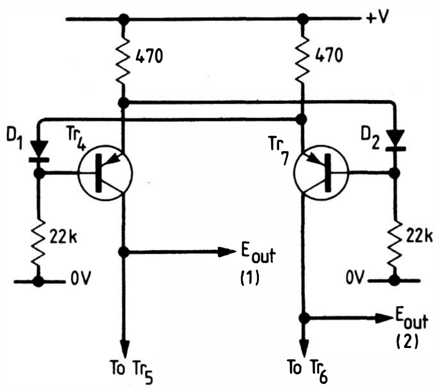


Fig. 11. *'Double-mirror' allows high gain, but with push-pull operation.*

attracted the attention of Japanese designers seeking a circuit to provide enough gain to compensate for the relatively poor dynamic transfer characteristics of a simple, complementary-pair, power-mosfet source follower. The circuit design of Fig. 10(a) with emitter resistors in $Tr_{3,4}$ helping to minimize V_{be} differences, is normally simplified to that in Fig. 10(b), which is recommended by Hitachi for use with their 2SJ and 2SK-series complementary devices.

Because of the low impedance presented at the collector of Tr_5 (Fig. 10) by the input to the current-mirror active load, this circuit provides only a 'single-ended' output. This restriction may be avoided, with a further improvement in voltage gain, if the simple current mirror of Figs 9 and 10 is replaced by the 'double-mirror' circuit shown, on its own, in Fig. 11. This permits an l.f. voltage gain in excess of 500,000, with a balanced push-pull output, which would be of advantage in circuits employing identical rather than complementary output devices.

There is, however, a point which must be borne in mind in this usage, that in any straightforward embodiment of this circuit arrangement, the d.c. output potential at the collector of only one of the two second stage amplifier transistors can be controlled by the use of the internal d.c. negative feedback path; the output potential at the collector of the other could lie at any point between the +V and −V supply limits, which would preclude any significant undistorted output swing being obtained from the uncontrolled output. This difficulty can be removed, however, by using the output from the uncontrolled amplifier to regulate the input current supplied to the initial long-tailed pair, as shown in Fig. 12. This effectively stabilizes the output potential of this point as well, and leads to the interesting possibility of a truly symmetrical, very high-gain stage with antiphase outputs, both of which can swing within nearly the total voltage range of the supply.

Power mosfet amplifiers using the general circuit structure of Fig. 3(b) (Part 1), and any of the gain blocks of Figs. 9-12, will give very satisfactory steady-state t.h.d. and transient (step function) performance, but there is a fairly substantial snag in respect of the output stage quiescent current. This is normally controlled by some circuit arrangement such as the variable resistance in the collector circuit of Tr_6 in Fig. 10, which produces a suitable voltage drop to maintain an appropriate forward voltage bias on the output devices, and which may be adjusted to set the output stage quiescent current to a suitable value. This chosen value of forward bias must be stable, and maintained at the set value throughout the life of the amplifier, and affected as little as possible by changes in ambient temperature, supply line voltages or ageing of components.

Current mirrors, while providing excellent active loads to gain stages, are not, in any sense, constant-current sources, but merely circuit arrangements which reflect into the output limb the current fed into the input limb. In the circuit configuration of Fig. 10, the current through Tr_4 and Tr_6 is determined by the potential applied across the 'tail' resistor of Tr_1 and Tr_2, and the current gains and forward V_{be} potentials of Tr_5 and Tr_6: in the circuit of Fig. 11, this current is determined mainly by the V_{be} characteristics of Tr_4 and Tr_7 and

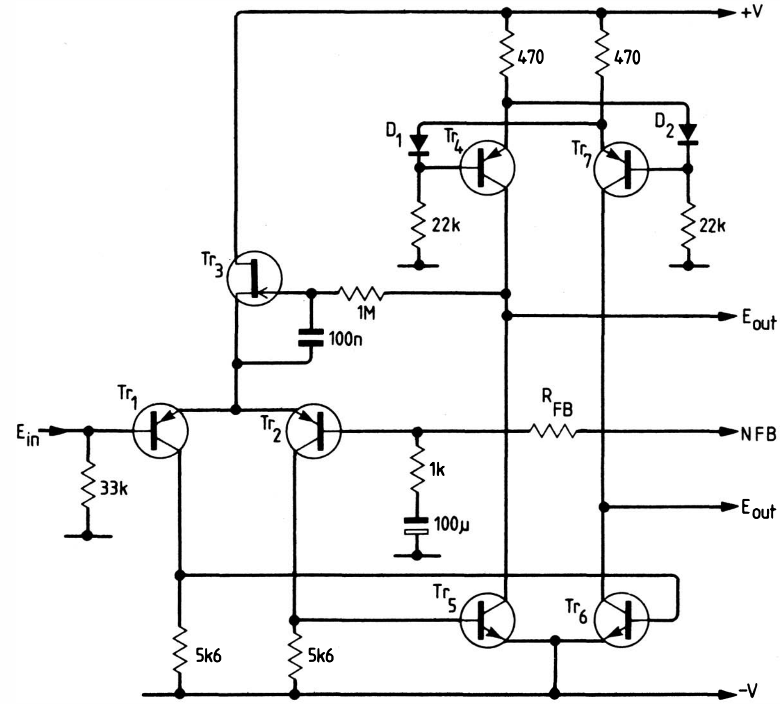


Fig. 12. *Amplifier using circuit of Fig. 11, d.c. stabilized by connexion from Tr_5 collector to gate of Tr_3. Provides symmetrical output, very high gain and large voltage swings.*

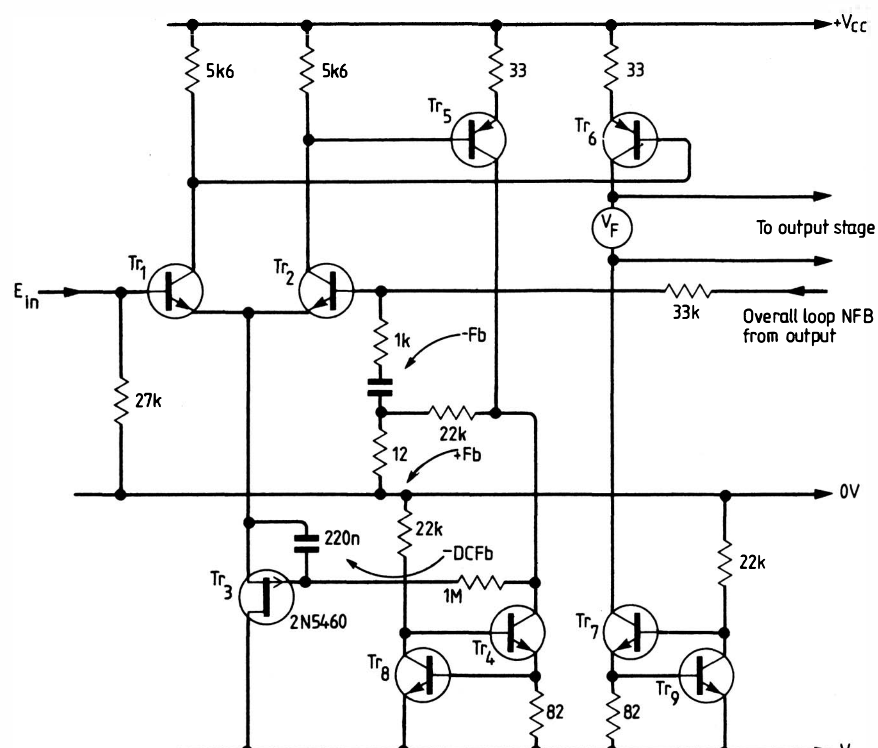


Fig. 13. *Ordinary constant-current loads used in conjunction with positive feedback give same results as current mirrors.*

the supply potential across the 22k feed resistors.

In such circuit configurations it is difficult to achieve quiescent current stability in the output stage over the expected range of supply voltage variations and fluctuations in ambient temperature. So, with regret, I concluded that a more formal constant-current source was necessary as the load element in Fig. 12, at least for Tr_6, if a straightforward voltage dropper circuit in this path was to be used to generate the forward bias of the output devices.

While it is difficult to envisage any standard constant-current source which would have the same dynamic advantage as a current-mirror active load, an analysis of the behaviour of such an active load (which is that of positive feed-forward from the parallel antiphase limb) in an amplifier having an overall negative-feedback loop, shows that it is substantially identical in its characteristics to any arrangement giving a similar gain increase by 'positive feedback within the negative-feedback loop'. This is a circuit technique of some antiquity and more widely used, deliberately or inadvertently, than one might guess.

This realization, in addition to focusing attention upon the possibilities and penalties of the technique, opens the door to the use of such a symmetrical amplifier configuration, with each half operating against a standard constant-current load, and an open-loop gain equal to or greater than that of the same circuit using a current mirror as the second-stage load by the use of a suitable positive-feedback path, as shown in the circuit of Fig. 13. In this, the polarities of the transistors have been reversed to take advantage of the slightly more favourable characteristics of the p-channel junction fet constant-current source in the 'tail' of the input long-tailed pair.

This use of positive feedback meets the main requirements of such an application, in that the positive feedback path is derived from a stage operated under more linear conditions than the main amplifier loop, and with a wider gain/bandwidth characteristic. Since the positive feedback signal is applied to the same point as the n.f.b. one, it can be visualized as cancelling part of the distorted signal normally fed back, thus leaving the distortion components as a larger part of the negative-feedback input, which facilitates their reduction.

This type of circuit, as an application of positive feedback, may be contrasted with the very commonly employed output-stage 'bootstrap', in which the positive-feedback signal is derived from a stage having worse linearity, and less good phase linearity, than the overall amplifier loop.

The practical power mosfet audio amplifier based on this design of voltage-amplifier stage, will be described in the next part of this article. This circuit has an output power of up to 100 watts, depending on supply line voltage, and has very good stability of d.c. operating conditions, gain and phase margins, and offers a significant advantage in audible quality over conventional bipolar power transistor designs.

References

2. Linsley Hood. J. L., *Wireless World*. Sept. 1971 pp 437-441.
3. Taylor. E. F., *Wireless World*. Aug. 1977 pp 28-32. www.keith-snook.info
4. Taylor. E. F., *Wireless World*. Sept. 1977 pp 55-59. www.keith-snook.info
5. Linsley Hood. J. L., *Hi-Fi News and Record Review*, Nov. 1972 pp 2120-2123.

To be concluded

EVENTS

June 19
RSGB H.f. Convention 1982 to be held at the Belfry Hotel and conference centre, Milton Common, Oxford. One day exhibition and lecture programme organised by the Radio Society of Great Britain, 35 Doughty Street, London WC1N 2AE.

June 21-22
Cable-casting in Europe – The commercial future. A conference on cable tv to be held at the International Press Centre, Shoe Lane, London EC1. Further information from International Broadcasting, 3-5, St John St. London EC1M 4AE.

June 22-24
Lab design '82, Joint Meeting of the Laboratory of Government Chemists and the IEE, at Church House, Westminster, London SW1. Details from IEE, Savoy Place, London WC2R 0BL.

June 29-July 1
The influence of microelectronics on measurements, instruments and transducer design. Conference to be held at UMIST. Details from the Conference Registrar, IERE, 99 Gower Street, London WC1E 6AZ.

July 4-9
Electronics systems. Summer school for teachers. Department of Electrical Engineering Science, University of Essex, Wivenhoe Park, Colchester CO4 3SQ.

July 5-7
Distributed computing systems. A Conference sponsored by the SERC, to be held at (and details from) Department of Computer Science, University of Strathclyde, Livingstone Tower, 26 Richmond Street, Glasgow G1 1XH.

July 6-9
Man/machine systems. IEE international conference to be held at University of Manchester Institute of Science and Technology. Details from IEE.

July 6-8
Reliability '83. Fourth national conference organised by the National Centre of Systems Reliability, and the Institute of Quality Assurance. To be held at the Metropole Hotel. National Exhibition Centre, Birmingham. Details from National Centre of Systems Reliability, UKAEA, Wigshaw Lane, Culcheth, Warrington, WA3 4NE.

July 12-15
The Video Revolution. A residential symposium organised by the Society of Electronic and Radio Technicians, 57-61 Newington Causeway, London SE1 6BL, to be held at the University of Reading.

July 12-16
Electronic application for teachers. Details from the Administrative Assistant, Room 110, Registrar's Department, University of Salford, Salford M5 4WT.

July 12-16
Integrated circuit engineering. Short course presented by the George Washington University, at Imperial College, London. Details from the UK/European office of the George Washington University, Hanover Square, London W1R 9DE.

July 12-16
Spectrum Management. Short course presented by The George Washington University. Details as above.

July 18-23
Optical fibre telecommunications. IEE Vacation school at the University of Essex.

July 19-23
Optical engineering including electro-optics. A short course presented by the George Washington University at Imperial College, London.

July 19-23
Satellite systems for navigation, traffic control and surveillance. A short course presented by the George Washington University at Imperial College, London.

July 19-22
Acoustical imaging '82. Twelfth international symposium to be held at the IEE, Savoy Place, London.

July 21
IEEIE Annual General Meeting. Savoy Hill, London WC2R 0BS.

July 26-28
Electronic Image Processing. IEE international conference at the University of York.

July 26-30
Data Communications Systems and Networks. Short course presented by the George Washington University at Imperial College, London.

July 26-30
Communications satellite engineering. Short course presented by the George Washington University at Imperial College, London.

July 26-28
Electromagnetic pulse and its effects on systems. Short course presented by the George Washington University at Imperial College, London.

A light pen for microcomputers.
For this project, described on p.30 of this issue, glass fibre printed circuit boards and a fibre optic cable assembly is available from M. R. Sagin, Nancarras Mill, The Level, Constantine, Falmouth, Cornwall.

80-100W MOSFET AUDIO AMPLIFIER

The final section of this three-part article describes the complete amplifier circuit in detail, with the addition of a loudspeaker protection circuit.

by J. L. Linsley Hood

In the earlier parts of this article I discussed some of the design requirements of power mosfet audio amplifiers and described the evolution of a high-gain, symmetrical, class 'A' driver stage suitable for use with a power mosfet output. Inevitably, the final design of the gain stage, as shown in the completed power amplifier circuit of Fig. 14, shows some minor differences in comparison with the basic voltage-amplifier circuit, which underlines the point that any final design represents only the small tip of a large submerged iceberg of design effort. Unless one is lucky, or one's target performance is relatively modest, or one has considerable experience with closely similar designs, there is always a large amount of work necessary to convert a reasonably satisfactory basic design into a final version having, as nearly as possible, a blameless performance under all conceivable test conditions.

Design considerations

The choice of output power rating for any power amplifier is, inevitably, somewhat arbitrary and depends on the voltage ratings of the available components, and on the cost of the power transformer, smoothing capacitors and heat sinks which one is prepared to afford. However, in practical terms, the major considerations which limit the possible output power are the voltage ratings of the output devices, and of the available electrolytic reservoir capacitors.

The output power mosfets I decided to use are the complementary n-channel and p-channel devices from Hitachi, since they are readily available, are reasonably inexpensive, appear to be adequately rugged, and have useful power ratings. These particular mosfets are available in peak working voltages up to 160V. However, there are other similar devices, either available now or promised in the near future, from Fairchild, Motorola, Ferranti, Supertex, International Rectifier and Intersil, so it seems likely that a design based on complementary power mosfets will not restrict the user to a single source of components.

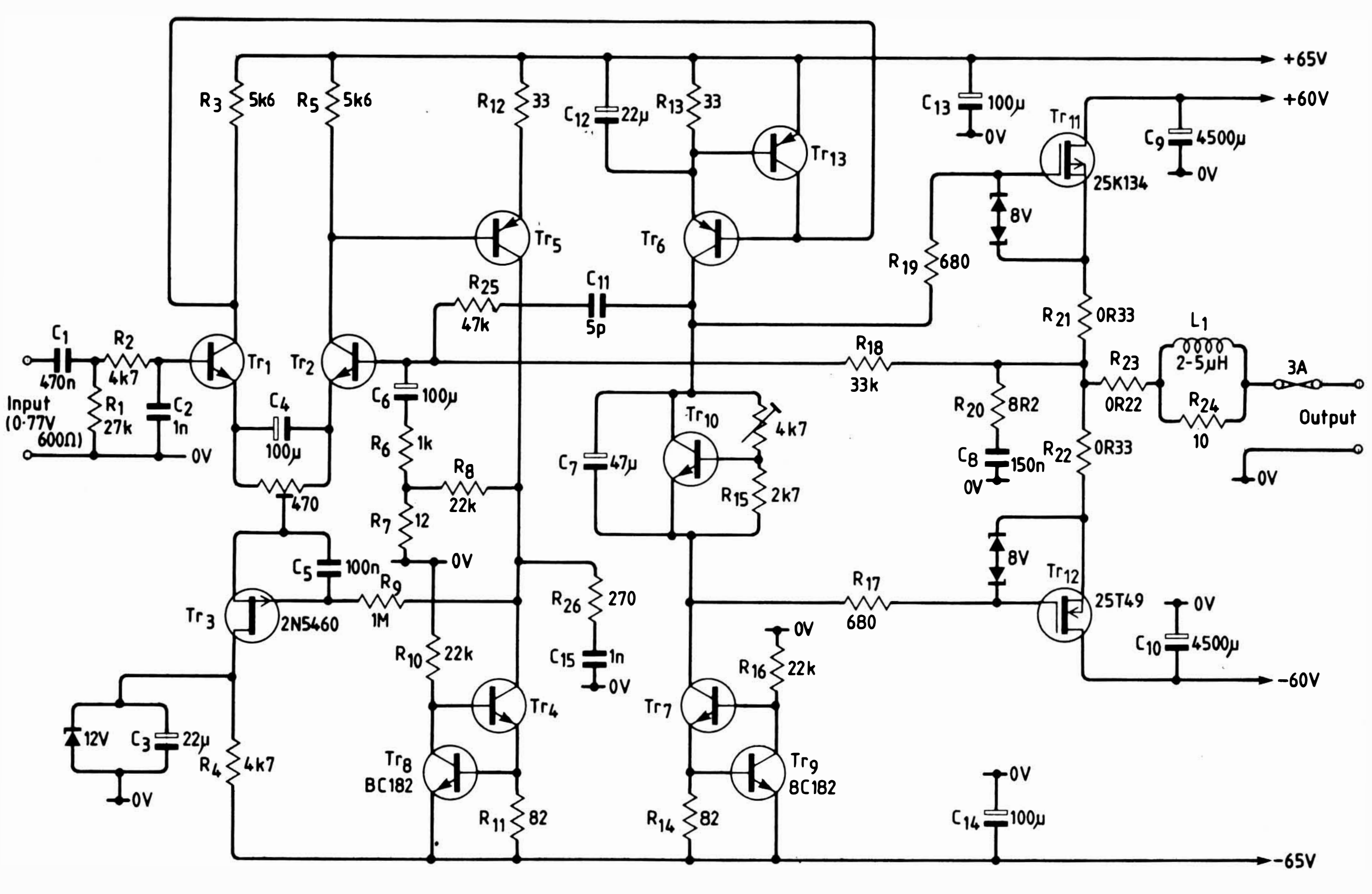
Some earlier experiments with mosfet-output audio amplifiers had shown that the r.m.s power output could be related to the available supply voltage in the manner shown in Fig. 15, over the range 25-100 watts. Since it had been decided, for various reasons, to use a symmetrical positive and negative supply, 63V electrolytic capacitors on each half would allow a safe working voltage, overall, of 120 volts, equivalent to a ±60V supply. In practice, the limited regulation of a simple rectifier/-capacitor power supply is likely to reduce this, on load, to some ±55V, giving an overall power output of 80 watts.

This output power requires a voltage swing of 25.3V r.m.s. across an eight ohm load, and if it is desired to drive this from an input voltage of '0VU' – which in audio-engineering terms implies 0.775V r.m.s. at a 600 ohm source impedence – the gain will require to be 32.6, which gives a suitable feedback resistor combination of 33k and 1041 ohms – though, in the event, for other considerations, it was decided to make this 1012, made up from a 1k and an 12 ohm series chain.

In the interests of d.c. symmetry, the input-circuit resistance should be also of the order of 33k. The values suggested are adequately close to this.

The performance of any feedback ampli-

Fig. 14. *Complete circuit diagram of the 100W amplifier.*



fier under transient (step-function or square wave) input conditions is helped if the input rise time can be limited. This can be done most easily by an input RC integrating network, R_2C_2, which gives a −3dB point at about 30kHz, allowing an adequate bandwidth for audio use.

A 470 ohm trimmer potentiometer in the emitter circuit of the input long-tailed pair allows accurate d.c. balance to be obtained with transistors having normal commercial spreads in V_{be} values and current gain. This is bypassed by a 100μF tantalum bead capacitor to avoid loss of open-loop a.c. gain. The output d.c. potential may be adjusted by means of this potentiometer to 0V, ±20mV.

Circuit performance depends strongly on the characteristics of the 'tail' of the 'long-tailed pair'. For correct operation of any such circuit, the dynamic impedance of the tail should be very large in comparison with the impedance as seen at the emitters of Tr_1 and Tr_2. Also, ideally, to minimize common-mode problems, the current from this source should be largely independent of the dynamic emitter potentials. Finally, the tail circuit should provide an adequate isolation from unwanted signal components on the supply line. A junction fet satisfies all these requirements very fully, and also allows, as explained above, control of the operating current in the second-stage class A amplifier. To allow a wider range of negative supply-line voltages, the negative-line supply to this fet is derived from a Zener-diode-stabilized −12 volt source. The use of a separate power supply for the driver stages is of considerable assistance in avoiding the performance degradation which can occur due to the intrusion of distorted signal potentials from the high-current output stage.

The second stage, class 'A', voltage amplifier is similar to that shown earlier in Fig. 13, except that conventional, two-transistor, constant-current sources are used as the loads for each half, and that a small amount of a.c. positive feedback is derived from the output of Tr_5, through R_8 and R_7, in addition to the current stabilizing d.c. negative feedback path through R_9 to Tr_3. The positive feedback restores the open loop a.c. gain to the 500,000 figure, over the frequency range 100Hz-3kHz, obtainable from the less d.c. stable configuration of Fig. 12.

The output power mosfets require a quiescent current value of 100mA for optimum performance – although it is difficult, because of the efficient operation of the n.f.b. loop, to see any significant change in the distortion residues, as this is adjusted, at any frequency below 10kHz – and this quiescent current is largely independent of the output device temperature. The 'amplified diode' circuit of Tr_{10} is not, therefore, used to sense the output transistor temperature, but used simply to generate a reasonably constant voltage drop.

Although the output devices present a very high l.f. input impedance, the effect of the 1200pF total gate-source capacitance cannot be ignored, and the current through Tr_6–Tr_7 must be enough to avoid any slew-rate limiting within the rise-time levels allowed by the input CR network, (R_2C_2). A current of 7mA is adequate for this, and permits worst-case dissipations of 900mW for $Tr_{4,7}$ and 450mW for $Tr_{5,6}$, which are within their limits.

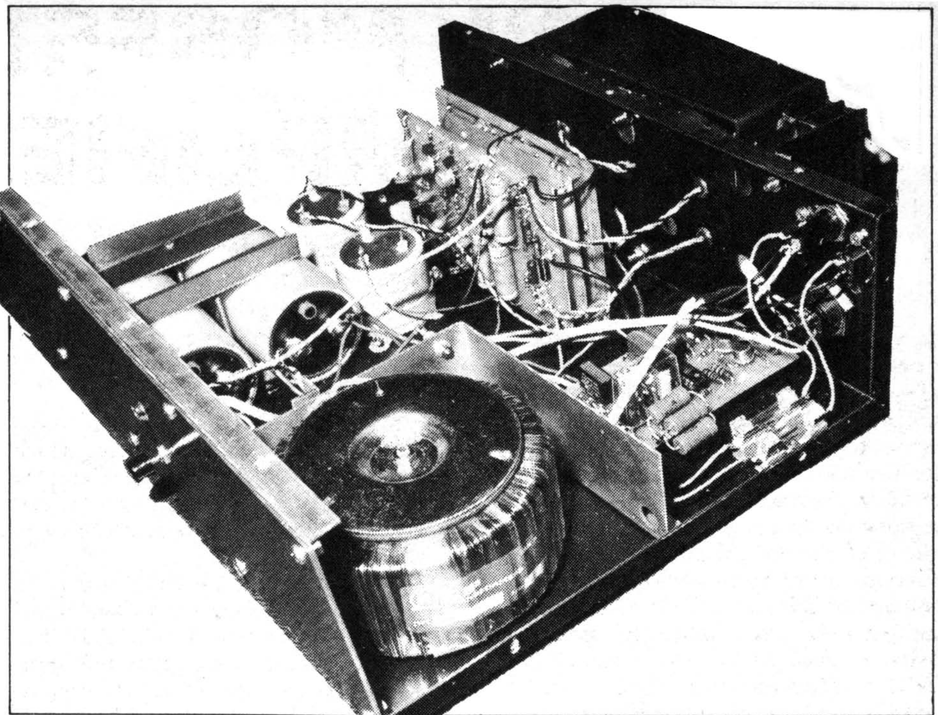
Since the Hitachi output devices are not protected by internal Zener diodes, it is unnecessary to exclude the possibility of reverse gate biassing, provided that this is within the ±14V gate-source breakdown voltage limits. This gate breakdown protection can therefore be provided by a pair of back-to-back 8V zeners, while the gate-source capacitance and the 680 ohm gate 'stopper' resistor will exclude the possibility of very rapid extraneous noise pulses which could escape Zener limiting due to lead inductance or turn-on time delays. Ideally, $R_{17,19}$ and the Zeners should be mounted close to the power mosfet pins.

Feedback loop, and loop stability

Although the use of a two-stage voltage amplifier will not automatically gurarantee, under all load conditions, that the internal phase shift will not approach 180° until the open-loop gain is negligible, the necessary conditions for an adequate phase margin, at unity gain, are very much easier to contrive in circuits in which only two successive gain stages are employed – provided that the additional phase shift of any other element in the feedback path is small enough to be neglected.

Unfortunately, in the case of the conventional junction-transistor Darlington or compound (p-n-p/n-p-n.) emitter follower this additional phase shift is significant, even at a few hundred kilohertz where the loop gain is still high, so this loop gain must be artificially reduced at higher frequencies to preserve closed-loop stability. Two basic methods exist for this, of which the first, and simpler, is simply to connect an external capacitor across the whole of the gain stage so that this acts as an integration network with a gain decreasing linearly by 20dB/decade from some l.f. break point. This has the advantage of allowing a wide phase margin of stability, and predictable performance characteristics. The second method is to tailor the h.f. performance so that it is maintained at as high a level as possible up to the point at which the loop phase shift approaches 180°, and then to reduce the gain rapidly, and in a manner chosen not to exceed the 180° stability threshold, until it is less than unity.

This method is commonly employed in



Prototype amplifier. Loudspeaker protection circuit is at rear right.

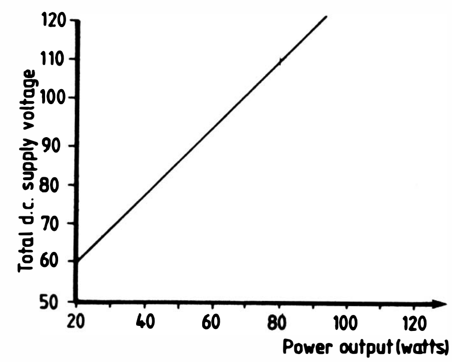


Fig. 15. *Amplifier output power as function of supply voltage.*

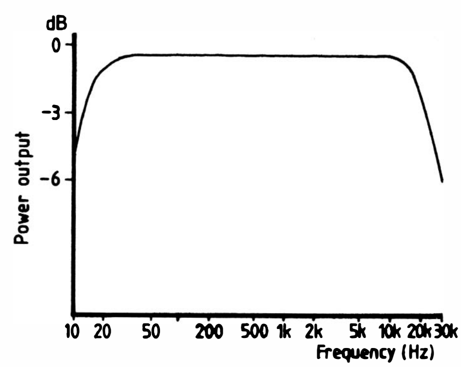


Fig. 16. *Power bandwidth of amplifier.*

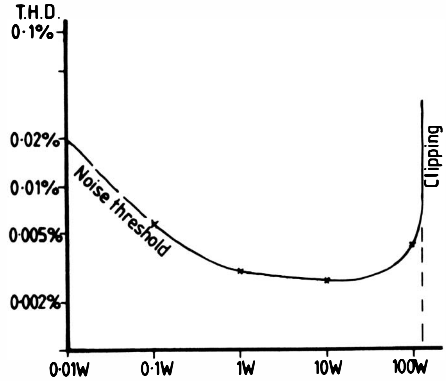


Fig. 17. *Harmonic distortion as a function of output power (1kHz, 8Ω load).*

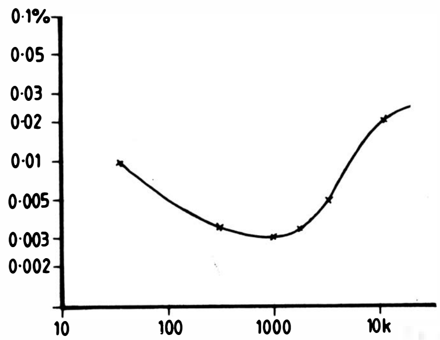


Fig. 18. *Harmonic distortion as a function of output frequency (80W, 8Ω load).*

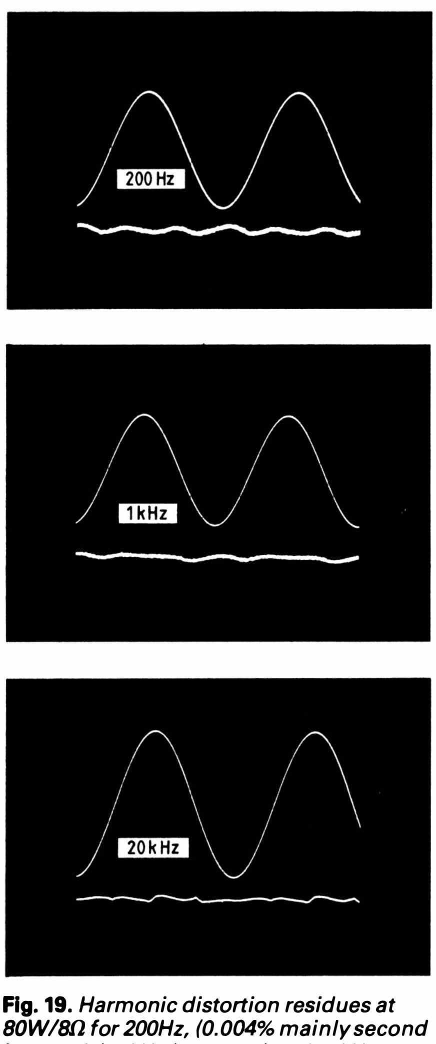


Fig. 19. *Harmonic distortion residues at 80W/8Ω for 200Hz, (0.004% mainly second harmonic), 1kHz (0.0025%) and 20kHz (0.025%).*

commercial transistor amplifier designs, often by the simple artifice of a capacitor between collector and base of the second stage amplifier transistor, because it allows better h.f. t.h.d. figures – and consequently better reviews in the 'Hi-Fi' journals. It does, however, carry with it the penalty that the phase margin of the amplifier is less good, with a consequently inferior transient response – manifest in respect of a less good 'settling time'[6] – and a less predictable performance with differing loudspeaker load characteristics. In addition, the intermal slew-rate limiting imposed by the second-stage collector-base capacitance (which is the mechanism by which the h.f. gain is reduced) leads to the predictable problem that signals accompanying large transient inputs will be blotted out during the period in which the amplifier is slew-rate limited. This is the phenomenon called 'Transient Intermodulation Distortion' by Otala[7]. This problem does not exist with the first method of h.f. compensation. A very good analysis of this problem was given by Jung[8] (with a small addendum by myself[9]).

The biggest advantage, in this respect, conferred by power mosfet output devices, is that the inherent phase-shift of the output emitter-follower impedance conversion stage is sufficiently small that it may be neglected up the megahertz region. This means that, with care, feedback audio amplifiers having high orders of negative feedback (open-loop gain) can be designed without the need for any external control of h.f. gain, and which will exhibit the desirable characteristics given by systems in which the gain decreases with frequency at 20dB/decade, and the loop phase shift does not significantly exceed 90°.

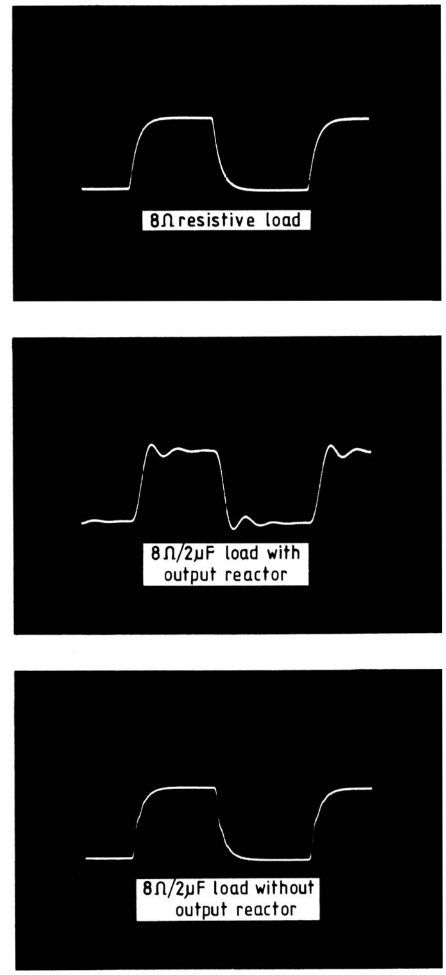


Fig. 20. *Response of amplifier to 10Vp-p 100kHz square wave on resistive and reactive loads, with and without output inductor.*

Influence of negative feedback

The use of negative feedback is, unfortunately, not as well understood, even among electronics engineers, as one might sometimes wish, and this misunderstanding has spilled over into the more emotive, and less logical, realm of the 'Hi-Fi' fraternity, where the ill effects attendant upon the improper use of this technique have encouraged the attempt to design amplifiers believed by their authors to employ no negative feedback whatever – a case of discarding the baby along with the bath water, if there ever was one.

The necessary conditions which must be satisfied if the potential benefits are to be gained have been examined both by Baxandall[10,11,12], in his series on audio amplifier design in this journal, and also, from a different angle, by *Wireless World*'s own Cathode Ray[13]. The message from all these contributions, if I may presume to precis, is that the amplifier in question must be made as linear as possible before negative feedback is applied; that the gain – at the frequency under consideration – must be enough, or the customary simplification of the mathematics will be inappropriate; and that a small amount of n.f.b., by injecting into the input an additional distorted signal, will worsen the harmonic distortion which would have been present without it.

Translated into design requirements, this implies that a high stage gain, coupled with good linearity and the lowest practicable phase shift, is the necessary design objective – most easily attained if not more than two gain stages are employed. The inclusion of a positive feedback path within the overall n.f.b. loop as a means of increasing the loop gain brings with it some supplementary requirements. These are that the phase shift within the positive feedback loop must be very small over the range of interest, since the p.f.b. will worsen it, and that the linearity of this part of the circuit must be much better than that of the remaining circuit outside the p.f.b. loop, or the benefits will be negated. Looked at in this light, the use of a bootstrapped driver load in an audio amplifier is not well advised, since the loop containing the 'bootstrap' will include the output devices whose linearity it is desired to improve.

In the particular case of the feedback loop built around Tr_2, Tr_5, R_8 and R_7, the linearity of this is very good because it is only driving a high-value resistive load, and the dominant phase shifts are those due to C_6 at the l.f. end, and the circuit stray capacitances in Tr_5 collector circuit at the h.f. end of the pass band. This gives a phase-linear bandwidth which is greater than that of the overall n.f.b. gain loop, and therefore satisfies the conditions for improving the overall amplifier performance.

Because of the capacitive nature of the load presented to Tr_6 by the gate-source capacitance of the power mosfets, the h.f. loop gain of the amplifier falls below unity at about 30MHz, which is sufficient to give an adequate margin of stability, while still allowing some 60dB of negative feedback at 30kHz, the chosen upper operating frequency limit. No additional h.f. roll-off components are required.

Stability with capacitative loads

A minor problem associated with power mosfets, discussed by Hitachi in their design note[14] is that the very high-frequency −3dB point of the mosfet used as a source follower (typically 30-40MHz for the Hitachi devices) allows the inductance of the internal gate-contact lead – some 70nH – to produce a negative resistance condition, with consequent parasitic oscillation, under conditions of small capacitative load (0.01μF-0.22μF). Oscillation, under these conditions, but due to other causes, is not uncommon in audio amplifiers, and can be the cause of amplifier failure when used with the so-called low-impedance loudspeaker cables, even when the amplifier is completely stable under the 8ohm/2μF load combination frequently chosen by reviewers. Needless to say, this possibility of parasitic oscillation should be avoided and this is most easily done in this type of design by the inclusion of a small inductor of some 5μH inductance, (20 turns of 24s.w.g. enamelled wire, wound round the case of a 10ohm, 1watt carbon-rod resistor) in the output lead to the loudspeaker load.

This output inductance has two practical effects, apart from the avoidance of parasitic oscillation. The first of these is to reduce the total harmonic distortion of the circuit, as measured at the output at high audio frequencies, simply because it acts as an output low-pass filter. The second effect, due to the same cause is a 'ripple' on the square-wave/reactive-load test waveform, which is an inevitable effect of any steep-cut, low-pass filter. Without this output inductor, the 8ohm/2μF test waveform is smoothly rounded and free of any overshoots.

Output stage protection

Because of the freedom of power mosfets from secondary breakdown, and because they have an inherent positive temperature coefficient of resistance, output stage protection can be much simpler than is the case with normal junction transistors, and a simple fuse in the output circuit is quite adequate. This has a practical advantage over many of the electronic protection methods normally employed, in that it avoids hard clipping under dynamic conditions when the amplifier is required to drive fast h.f. transients into loudspeakers having a low h.f. impedance.

Overall performance and sound quality

The power bandwidth, the t.h.d. as a function of output power, and the t.h.d. as a function of signal frequency are shown in Figs. 16-18, and the distortion waveforms and 10kHz reactive load waveform, with and without the output inductor, are shown in the oscilloscope photographs of Figs. 19 and 20.

Inevitably, the question must be asked whether, in the event, the sound quality given by a well designed power mosfet amplifier is better than, or indeed noticeably different from, that given by an equally well designed power amplifier using junction transistors. The designer is not a good person from whom to seek an answer to this question, if only because his awareness of the inevitable design compromises in the circuit, and of the imperfections which remain as a result of the impossibility of achieving all design objectives simultaneously, colour his expectations in respect of its perceived performance. However, having said this, I believe that power mosfet output devices, in appropriately designed circuitry, can offer an improved performance in the upper-middle and top end of the audible spectrum, which is apparent as an improved clarity and transparency in tonal quality, particularly at low output levels, in comparison with equivalent junction transistor designs.

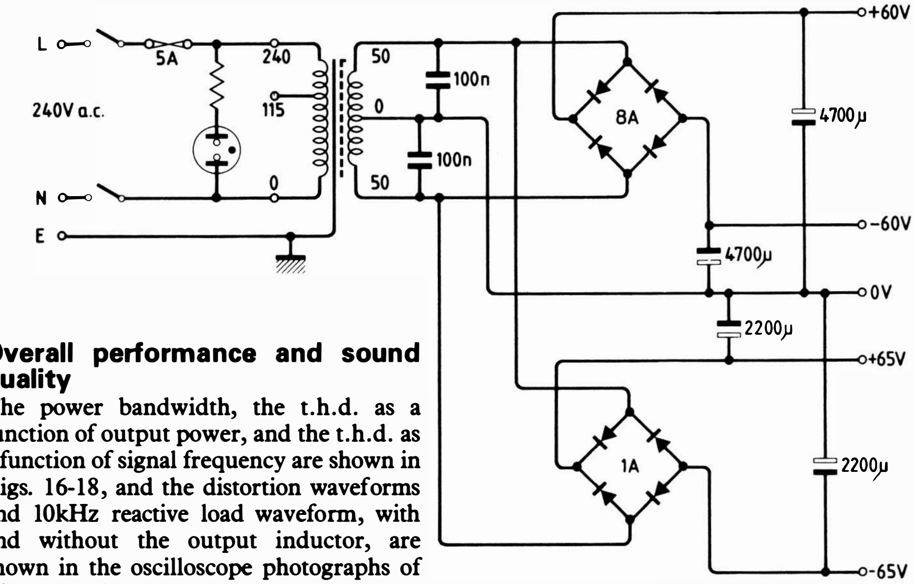


Fig. 22. *Power supply used in prototype.*

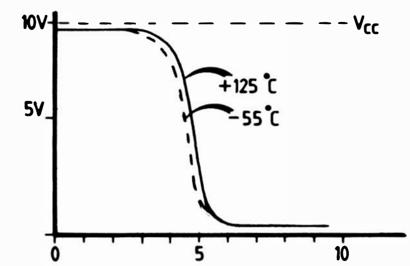


Fig. 23. *Typical transfer characteristic of c.m.o.s. gate.*

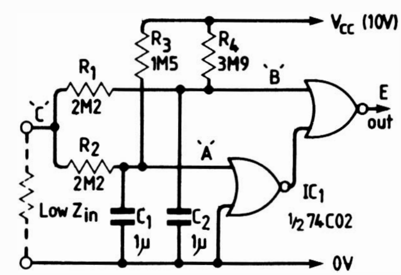


Fig. 24. *Input d.c. level monitor using c.m.o.s. Nor.*

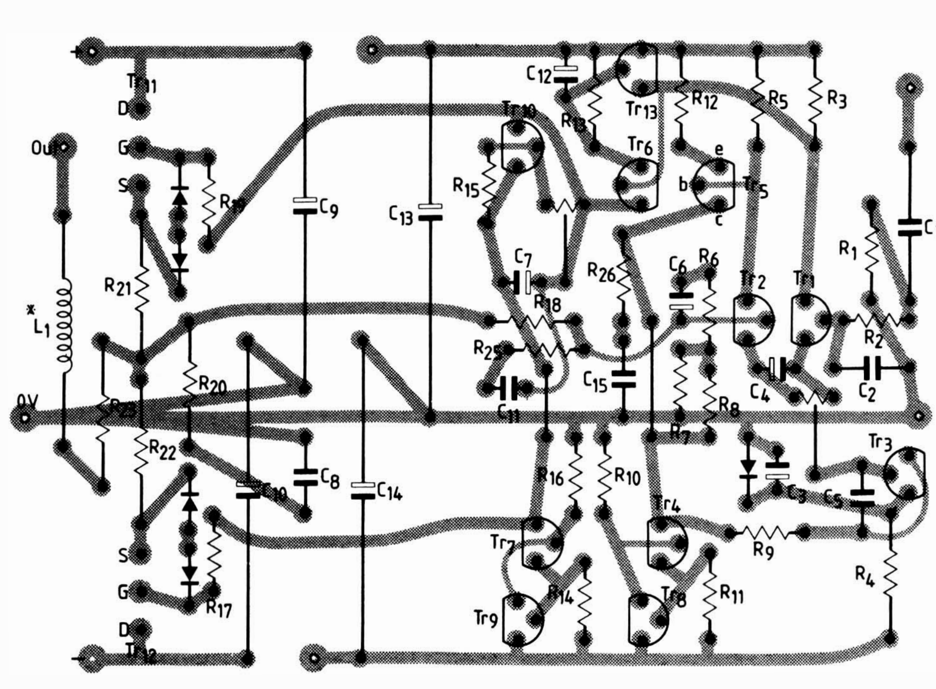


Fig. 21. *Printed-board for power amplifier.*

* L_1 is wound on R_{24}

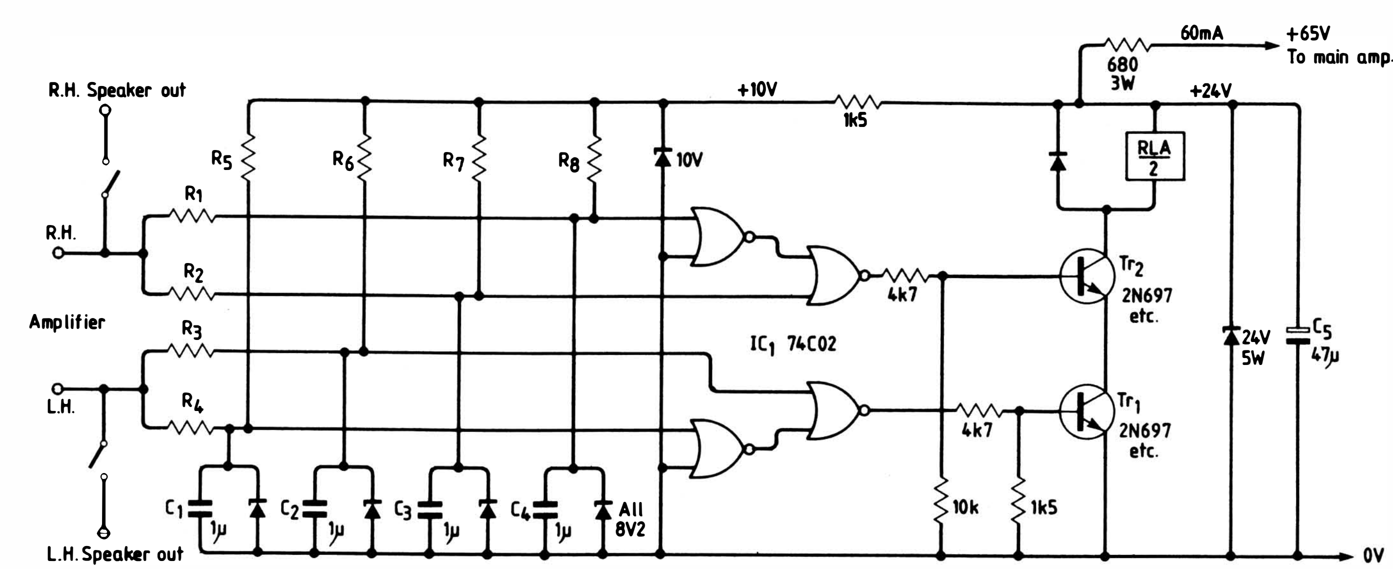


Fig. 25. *Complete two-channel loudspeaker protection circuit with switch-on delay.*

Power supply

A suitable power supply circuit is shown in Fig. 22. As mentioned above, the output power of the amplifier depends almost entirely on the supply line voltages, and the original design was based on a conventional 'E' and 'I' cored transformer with a nominal 50-0-50 secondary winding, which gave a quiescent output d.c. voltage, after rectification, of ±62V. This was subsequently replaced by a 250VA 50-0-50V toroidal cored unit, in the interests of a lower residual 50Hz field, and this gave a d.c. output of ±65 volts, and increased the power output, at 1kHz across an 8ohm, water-cooled, resistive load, from 83watts/channel to some 105watts/-channel. It was thought prudent to uprate the reservoir capacitors to 80V types, but no other changes are necessary.

Loudspeaker protection circuit

Although the use of direct coupling between loudspeaker and amplifier output, together with the use of split positive and negative h.t. rails, undoubtedly helps in the economical design of high-powered audio amplifiers by limiting the necessary voltage rating of the reservoir capacitors, it does carry with it the implicit hazard that, in the event of a component failure within the power amplifier, the whole output of one or other of the supply lines may be switched into the output circuit, with expensive consequences.

The most elegant way of avoiding this hazard is to employ a small supplementary circuit to monitor the average d.c. potential of the amplifier output terminals, and to disconnect the loudspeakers in the event that an averaged d.c. offset of more than a volt or so is detected. Experiments over a periold of time have shown that the loudspeaker can be connected through a pair of gold-plated relay contacts without audible or measurable signal degradation. Silver-plated contacts are excellent when new and clean, but tend to become partially rectifying if sulphided by exposure to urban atmospheres, and should therefore be avoided if possible.

An inevitable problem in the use of an 'average d.c. potential' monitoring circuit is the necessity for some compromise between speed of response, in disconnexion following a fault condition, and the need not to diagnose a large but legitimate v.l.f. signal – especially if assymmetrical – as such a fault. My own choice is an integrating time-constant of about 2 seconds. This ignores all the normal l.f. signal components, at least at the largest signal levels I have so far used, but allows a switch-off in better than 80 milliseconds in the event of a large direct voltage being applied to the input. This should be adequate to avoid thermal damage to the loudspeaker.

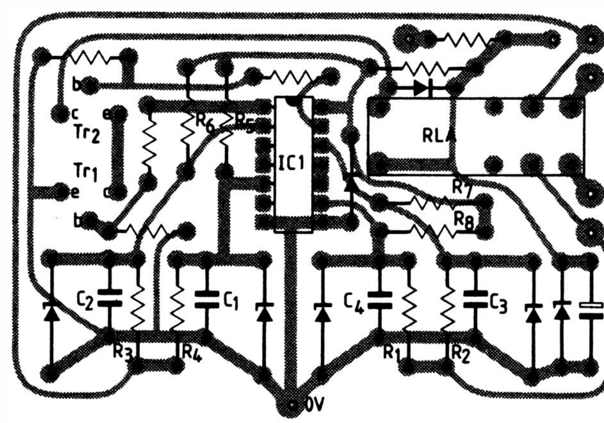


Fig. 26: *Layout of printed board for circuit of Fig. 25.*

In order to accommodate a fairly long integrating time-constant with the use of non-polarized capacitors, a high-input-impedance offset-detection logic circuit is essential. C.m.o.s. logic elements of the 74C or CD4*** series are well suited to this task, especially since the switching potentials are well defined in relation to the supply voltage line employed. Typical gate transfer characteristics are shown in Fig. 23. Because of this, if the gates are biassed by an input resistor chain, as shown in Fig. 24, so that one sits below and one sits above this threshold level, a pair of Nor gates will effectively act as an input-threshold d.c. monitor circuit, in which the output will only be high so long as input A is high and input B is low. With the resistor values quoted, this condition will be met while input C is within ±2V d.c., for a 10V supply line. The circuit also will provide a switch-on delay of a few seconds while C_1 charges up through R_3 to a potential above the $\frac{1}{2}V_{cc}$ level.

The complete, two-channel, loudspeaker protection circuit based on this arrangement needs only one Quad 2-input Nor gate, and a pair of switching transistors. The final circuit is shown in Fig. 25. It is 'fail-safe' in the sense that the relay contacts are normally open, and can only operate if the h.t. supply is present and both transistors are energized. The relay used is an RS Components p.c.b.-mounting, 24V unit, with 5A, 250V a.c.-rated gold-plated contacts, of d.p.d.t. operation. H.t. supply for this is best obtained from the output stage +65 volt line.

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Editor's note: *We understand that a kit of components for the amplifier is to be made available by Hert Electronic Kits, Ltd, Oswestry, Shropshire.*

A preamplifier design to match the mosfet power amplifier will be described later in the year. □