Simple Class A Amplifier

A 10-W design giving subjectively better results than class B transistor amplifiers

by J. L. Linsley Hood, M.I.E.E.

During the past few years a number of excellent designs have been published for domestic audio amplifiers. However, some of these designs are now rendered obsolescent by changes in the availability of components, and others are intended to provide levels of power output which are in excess of the requirements of a normal living room. Also, most designs have tended to be rather complex.

In the circumstances it seemed worth while to consider just how simple a design could be made which would give adequate output power together with a standard of performance which was beyond reproach, and this study has resulted in the present design.

Output power and distortion

In view of the enormous popularity of the Mullard "5-10" valve amplifier, it appeared that a 10-watt output would be adequate for normal use; indeed when two such amplifiers are used as a stereo pair, the total sound output at full power can be quite astonishing using reasonably sensitive speakers.

The original harmonic distortion standards for audio amplifiers were laid down by D. T. N. Williamson in a series of articles published in Wireless World in 1947 and 1949; and the standard, proposed by him, for less than 0.1% total harmonic distortion at full rated power output, has been generally accepted as the target figure for high-quality audio power amplifiers. Since the main problem in the design of valve audio amplifiers lies in the difficulty in obtaining adequate performance from the output transformer, and since modern transistor circuit techniques allow the design of power amplifiers without output transformers, it seemed feasible to aim at a somewhat higher standard, 0.05% total harmonic distortion at full output power over the range 30Hz-20kHz. This also implies that the output power will be constant over this frequency range.

Circuit design

The first amplifier circuit of which the author is aware, in which a transformerless transistor design was used to give a standard of performance approaching that of the "Williamson" amplifier, was that published in Wireless World in 1961 by Tobey and Dinsdale. This employed a class B output stage, with series connected transistors in quasi-complementary symmetry. Subsequent high-quality transistor power amplifiers have largely tended to follow the design principles outlined in this article.

The major advantage of amplifiers of this type is that the normal static power dissipation is very low, and the overall power-conversion efficiency is high. Unfortunately there are also some inherent disadvantages due to the intrinsic dissimilarity in the response of the two halves of the push-pull pair (if complementary transistors are used in unsymmetrical circuit arrangement) together with some cross-over distortion due to the Ic/Vb characteristics. Much has been done, particularly by Bailey[1], to minimise the latter.

An additional characteristic of the class B output stage is that the current demand of the output transistors increases with the output signal, and this may reduce the output voltage and worsen the smoothing of the power supply, unless this is well designed. Also, because of the increase in current with output power, it is possible for a transient overload to drive the output transistors into a condition of thermal runaway, particularly with reactive loads, unless suitable protective circuitry is employed. These requirements have combined to increase the complexity of the circuit arrangement, and a well designed low-distortion class B power amplifier is no longer a simple or inexpensive thing to construct.

An alternative approach to the design of a transistor power amplifier combining good performance with simple construction is to use the output transistors in a class A configuration. This avoids the problems of asymmetry in quasi-complementary circuitry, thermal runaway on transient overload, cross-over distortion and signal-dependent variations in power supply current demand. It is, however, less efficient than a class B circuit, and the output transistors must be mounted on large heat sinks.

The basic class A construction consists of a single transistor with a suitable collector load. The use of a resistor, as in Fig. 1(a), would be a practical solution, but the best power-conversion efficiency would be about 12%. An l.f. choke, as shown in Fig. 1(b), would give much better efficiency, but a properly designed component would be bulky and expensive, and remove many of the advantages of a transformerless design. The use of a second, similar, transistor as a collector load, as shown in Fig. 1(c), would be more convenient in terms of size and cost, and would allow the load to be driven effectively in push-pull if the inputs of the two transistors were of suitable magnitude and opposite in phase. This requirement can be achieved if the driver transistor is connected as shown in Fig. 2.

This method of connection also meets one of the most important requirements of a low distortion amplifier - that the basic linearity of the amplifier should be good, even in the absence of feedback. Several factors contribute to this. There is the tendency of the Ic/Vb non-linearity of the characteristics of the output transistors to cancel, because during the part of the cycle in which one transistor is approaching cut-off the other is turned full on. There is a measure of internal feedback around the loop Tr1, Tr2, Tr3 because of the effect which the base impedance characteristics of Tr1 have on the output current of Tr3. Also, the driver transistor Tr3, which has to deliver a large voltage swing, is operated under conditions which favour low harmonic distortion - low output load impedance, high input impedance. A practical power amplifier circuit using this type of output stage is shown in Fig. 3.

The open loop gain of the circuit is approximately 600 with typical transistors. The closed loop gain is determined, at frequencies high enough for the impedance of C3 to be small in comparison to R4, by the ratio (R3 + R4)/R4. With the values indicated in Fig. 3, this is 13. This gives a feedback factor of some 34dB, and an output impedance of about 160 milliohms.

Since the circuit has unity gain at d.c., because of the inclusion of C3 in the feedback loop, the output voltage, Ve, is held at the same potential as the base of Tr4 plus the base emitter potential of Tr4 and the small potential drop along R3 due to the emitter current of this transistor. Since the output transistor Tr1 will turn on as much current as is necessary to pull Ve down to this value, the resistor R2, which together with R1 controls the collector current of Tr2, can be used to set the static current of the amplifier output stages. It will also be apparent that Ve can be set to any desired value by small adjustments to R5 or R6. The optimum performance will be obtained when this is equal to half the supply voltage. (Half a volt or so either way will make only a small difference to the maximum output power obtainable, and to the other characteristics of the amplifier, so there is no need for great precision in setting this.)

Silicon planar transistors are used throughout, and this gives good thermal stability and a low noise level. Also, since there is no requirement for complementary symmetry, all the power stages can use n-p-n transistors which offer, in silicon, the best performance and lowest cost. The overall performance at an output level of 10 watts, or at any lower level, more than meets the standards laid down by Williamson. The power output and gain/frequency graphs are shown in Figs. 4 - 6, and the relationship between output power and total harmonic distortion is shown in Fig. 7. Since the amplifier is a straight-forward class A circuit, the distortion decreases linearly with output voltage. (This would not necessarily be the case in a class B system if any significant amount of cross-over distortion was present.) The analysis of distortion components at levels of the order of 0.05% is difficult, but it appears that the residual distortion below the level at which clipping begins is predominantly second harmonic.

Stability, power output and load impedance

Silicon planar n-p-n transistors have, in general, excellent high frequency characteristics, and these contribute to the very good stability of the amplifier with reactive loads. The author has not yet found a combination of L and C which makes the system unstable, although the system will readily become oscillatory with an inductive load if R3 is shunted by a small condenser to cause roll-off at high frequencies.

The circuit shown in Fig. 3 may be used, with very little modification to the component values, to drive load impedances in the range 3 - 15 ohms. However, the chosen output power is represented by a different current/voltage relationship in each case, and the current through the output transistors and the output-voltage swing will therefore also be different. The peak-voltage swing and mean output current can be calculated quite simply from the well-known relationships $W=I^2.R$ and $V=I.R$, where the symbols have their customary significance. (It should be remembered, however, that the calculation of output power is based on r.m.s. values of current and voltage, and that these must be multiplied by 1.414 to obtain the peak values, and that the voltage swing measured is the peak-to-peak voltage, which is twice the peak value.)

When these calculations have been made, the peak-to-peak voltage swing for 10 watts power into a 15-Ohm load is found to be 34.8 volts. Since the two output transistors bottom at about 0.6 volts each, the power supply must provide a minimum of 36 volts in order to allow this output. For loads of 8 and 3 ohms, the minimum h.t. line voltage must be 27V and 17 volts respectively. The necessary minimum currents are 0.9, 1.2 and 2.0 amps. Suggested component values for operation with these load impedances are shown in Table 1. C3 and C1 together influence the voltage and power roll-off at low audio frequencies. These can be increased in value if a better low-frequency performance is desired than that shown in Figs. 4 - 6.

Since the supply voltages and output currents involved lead to dissipations in the order of 17 watts in each output transistor, and since it is undesirable (for component longevity) to permit high operating temperatures, adequate heat sink area must be provided for each transistor. A pair of separately mounted 5in by 4in finned heatsinks is suggested. This is, unfortunately, the penalty which must be paid for class A operation. For supplies above 30V Tr1 and Tr2 should be MJ481s and Tr3 a 2N1613.

If the output impedance of the pre-amplifier is more than a few thousand ohms, the input stage of the amplifier should be modified to include a simple f.e.t. source follower circuit, as shown in Fig. 8. This increases the harmonic distortion to about 0.12%, and is therefore (theoretically) a less attractive solution than a better pre-amplifier. A high frequency roll-off can then be obtained, if necessary, by connecting a small capacitor between the gate of the f.e.t. and the negative (earthy) line.

Z_L	V	I	R1	R2	C1	C2	V_{IN} (rms)
3Ω	17V	2.0A	47Ω	180Ω	500µF 25V	5000µF 25V	0.41V
8Ω	27V	1.2A	100Ω	560ΩΩ	250µF 40V	2500µF 50V	0.66V
15Ω	36V	0.9A	150Ω	1.2kΩ	250µF 40V	2500µF 50V	0.90V

Table 1. Summary of component combinations for different load impedances.

Suitable transistors

Some experiments were made to determine the extent to which the circuit performance was influenced by the type and current gain of the transistors used. As expected the best performance was obtained when high-gain transistors were used, and when the output stage used a matched pair. No adequate substitute is known for the 2N697 / 2N1613 type used in the driver stage, but examples of this transistor type from three different manufacturers were used with apparently identical results. Similarly, the use of alternative types of input transistor produced no apparent performance change, and the Texas Instruments 2N4058 is fully interchangeable with the Motorola 2N3906 used in the prototype.

The most noteworthy performance changes were found in the current gain characteristics of the output transistor pair, and for the lowest possible distortion with any pair, the voltage at the point from which the loudspeaker is fed should be adjusted so that it is within 0.25 volt of half the supply line potential. The other results are summarized in Table 2.

The transistors used in these experiments were Motorola MJ480 / 481, with the exception of (6), in which Texas 2S034 devices were tried. The main conclusion which can be drawn from this is that the type of transistor used may not be very important, but that if there are differences in the current gains of the output transistors, it is necessary that the device with the higher gain shall be used in the position of Tr1.

When distortion components were found prior to the onset of waveform clipping, these were almost wholly due to the presence of second harmonics.

Test No.	Current Gain Tr1	Current Gain Tr2	Distortion (at 9 watts)
1	135	135	0.06%
2	40	120	0.4%
3	120	40	0.12% (pair 2 reversed)
4	120	100	0.09%
5	100	120	0.18% (pair 5 reversed)
6	50	40	0.1%

Table 2. Relation of distortion to gain-matching in the output stage.

Constructional notes

Amplifier. The components necessary for a 10 + 10 watt stereo amplifier pair can be conveniently be assembled on a standard “Lektrokit” 4in x 4.75in s.r.b.p. pin board, as shown in the photographs, with the four power transistors mounted on external heat sinks. Except where noted the values of components do not appear to be particularly critical, and 10% tolerance resistors can certainly be used without ill effect. The lowest noise levels will however be obtained with good quality components, and with carbon-film, or metal-oxide, resistors.

Power Supply. A suggested form of power supply unit is shown in Fig. 9(a). Since the current demand of the amplifier is substantially constant, a series transistor smoothing circuit can be used in which the power supply output voltage may be adjusted by choice of the base current input provided by the emitter follower Tr2 and the potentiometer VR1. With the values of the reservoir capacitor shown in Table 3, the ripple level will be less than 10mV at the rated output current, provided that the current gain of the series transistors is greater than 40. For output currents up to 2.5 amps, the series transistors indicated will be adequate, provided that they are mounted on heat sinks appropriate to their loading.

However, at the current levels necessary for operation of the 3-ohm version of the amplifier as a stereo pair, a single MJ480 will no longer be adequate, and either a more suitable series transistor must be used, such as the Mullard BDY20, with for example a 2N1711 as Tr2, or with a parallel connected arrangement as shown in Fig. 9(b).

Amp Z_L	I_{OUT}	V_{OUT}	C1	Tr1/2	MR1	T1
15Ω	1A	37V	1000μF 50V	MJ480 / 2N697	5BO5	40V 1A
2 x 15Ω	2A	37V	5000μF 50V	MJ480 / 2N697	5BO5	40V 2A
8Ω	1.25A	27V	2000μF 40V	MJ480 / 2N697	5BO5	30V 1.25A
2 x 8Ω	2.5A	27V	5000μF 40V	MJ480 / 2N697	5BO5	30V 2.5A
3Ω	1.9A	18V	5000μF 30V	MJ480 / 2N697	5BO5	20V 2A
2 x 3Ω	3.8A	18V	10,000μF 30V	MJ480 / 2x2N697	7BO5T	20V 4A

Table 3. Power-supply components

The total resistance in the rectifier "primary" circuit, including the transformer secondary winding, must not be less than 0.25Ω. When the power supply, with or without an amplifier, is to be used with an r.f. amplifier-tuner unit, it may be necessary to add a 0.25uF (160V.w.) capacitor across the secondary winding of T1 to prevent transient radiation. The rectifier diodes specified are International Rectifier potted bridge types.

Transistor protection circuit

The current which flows in the output transistor chain (Tr1, Tr2) is determined by the potential across Tr2, the values of R1 and R2, and the current gain and collector-base leakage current of Tr2. Since both these transistor characteristics are temperature dependant the output series current will increase somewhat with the temperature of Tr2. If the amplifier is to be operated under conditions of high ambient temperature, or if for some reason it is not practicable to provide an adequate area of heat-sink for the output transistors, it will be desirable to provide some alternative means for the control of the output transistor circuit current. This can be done by means of the circuit shown in Fig. 10. In this, some proportion of the d.c. bias current to Tr1 is shunted to the negative line through Tr7, when the total current flowing causes the potential applied to the base of Tr6 to exceed the turn-on value (about 0.5 volt). This allows very precise control of the series current without affecting the output power or distortion characteristics. The simpler arrangement whereby the current control potential for Tr7 is obtained from a series resistor in the emitter circuit of Tr1 leads, unfortunately, to a worsening of the distortion characteristics to about 0.15% at 8 watts, rising to about 0.3% at the onset of overload.

Performance under listening conditions

It would be convenient if the performance of an audio amplifier (or loudspeaker or any other similar piece of audio equipment) could be completely specified by frequency response and harmonic distortion characteristics. Unfortunately, it is not possible to simulate under laboratory conditions the complex loads or intricate waveform structures presented to the amplifier when a loudspeaker system is employed to reproduce the everyday sounds of speech and music; so that although the square wave and low-distortion sine wave oscillators, the oscilloscope, and the harmonic distortion analyser are valuable tools in the design of audio circuits, the ultimate test of the final design must be the critical judgement of the listener under the most carefully chosen conditions his facilities and environment allow.

The possession of a good standard of reference is a great help in comparative trials of this nature, and the author has been fortunate in the possession, for many years, of a carefully and expensively built "Williamson" amplifier, the performance of which has proved, in listening trials, to equal or exceed, by greater or lesser margins, that of any other audio amplifier with which the author has been able to make comparisons.

However, in the past, when these tests were made for personal curiosity, and some few minutes could elapse in the transfer of input and output leads from one amplifier to the other, the comparative performance of some designs has been so close that the conclusion drawn was that there was really very little to choose between them. Some of the recent transistor power amplifier circuits gave a performance which seemed fully equal to that of the "Williamson", at least so far as one could remember during the interval between one trial and the next. It was, however, appreciated that this did not really offer the best conditions for a proper appraisal of the more subtle differences in the performance of already good designs, so a changeover switch was arranged to transfer inputs and outputs between any chosen pair of amplifiers, and a total of six amplifier units was assembled, including the "Williamson", and another popular valve unit, three class B transistor designs, including one of commercial origin, and the class A circuit described above. The frequency response, and total harmonic distortion characteristics, of the four transistor amplifiers was tested in the laboratory prior to this trial, and all were found to have a flat frequency response through the usable audio spectrum, coupled with low harmonic distortion content (the worst-case figure was 0.15%).

In view of these prior tests, it was not expected that there would be any significant difference in the audible performance of any of the transistor designs, or between them and the valve amplifiers. It was therefore surprising to discover, in the event, that there were discernable differences between the valve and the three class B transistor units. In fact, the two valve designs and the class A transistor circuit, and the three class B designs formed two tonally distinct groups, with closely similar characteristics within each group.

The "Williamson" and the present class A design were both better than the other valve amplifier, and so close in performance that it was almost impossible to tell which of the two was in use without looking at the switch position. In the upper reaches of the treble spectrum the transistor amplifier has perhaps a slight advantage.

The performance differences between the class A and the class B groups were, however, much more prominent. Not only did the class A systems have a complete freedom from the slight "edginess" found on some high string notes with all the class B units, but they appeared also to give a fuller, "rounder", quality, the attractiveness of which to the author much outweighs the incidental inconvenience of the need for more substantial power supply equipment and more massive heat sinks.

Some thought, in discussions with interested friends, has been given to the implications of this unlooked-for discovery, and a tentative theory has been evolved which is offered for what it is worth. It is postulated that these tonal differences arise because the normal moving-coil loudspeaker, in its associated housing, can present a very complex reactive load at frequencies associated with structural resonances, and that this might provoke transient overshoot when used with a class B amplifier, when a point of inflection in the applied waveform chanced to coincide with the point of transistor crossover, at which point, because of the abrupt change in the input parameters of the output transistors the loop stability margins and output damping will be less good. In these circumstances, the desired function of the power-amplifier output circuit in damping out the cone-response irregularities of the speaker may be performed worse at the very places in the loudspeaker frequency-response curve where the damping is most needed.

It should be emphasized that the differences observed in these experiments are small, and unlikely to be noticed except in direct side-by-side comparison. The perfectionist may, however, prefer class A to class B in transistor circuitry if he can get adequate power for his needs that way.

Listener fatigue

In the experience of the author, the performance of most well-designed audio power amplifiers is really very good, and the differences between one design and another are likely to be small in comparison with the differences between alternative loudspeaker systems, for example, and of the transistor designs so far encountered, not one could be considered as unpleasing to the ear. However, with the growing use of solid-state power amplifiers, puzzling tales of "listener fatigue" have been heard among the *cognoscenti*, as something which all but the most expensive transistor amplifiers will cause the listener, in contradistinction with good valve-operated amplifiers. This seemed to be worth investigation, to discover whether there was any foundation for this allegation.

In practice it was found that an amplifier with an impeccable performance on paper could be quite worrying to listen to under certain conditions. This appears to arise and be particularly associated with transistor power amplifiers because most of these are easily able to deliver large amounts of power at supersonic frequencies, which the speakers in a high quality system will endeavour to present to the listener. In this context it should be remembered that in an amplifier which has a flat power response from 30Hz to 180kHz, 90% of this power spectrum will be supersonic.

This unwanted output can arise in two ways. It can be because of wide spectrum "white noise" from a preamplifier with a significant amount of hiss – this can happen if a valve preamplifier is mismatched into the few thousand ohms input impedance of a transistor power amplifier, and will also cause the system performance to be unnaturally lacking in bass. Trouble of this type can also arise if transient instability or high frequency "ringing" occurs, for example when a reactive load is used with a class B amplifier having poor cross-over point stability.

Reference

1. Bailey, A.R., "High-performance Transistor Amplifier", Wireless World, November 1966; "30-Watt High Fidelity Amplifier", May 1968 and "Output Transistor Protection in A.F. Amplifiers", June 1968.

Figures

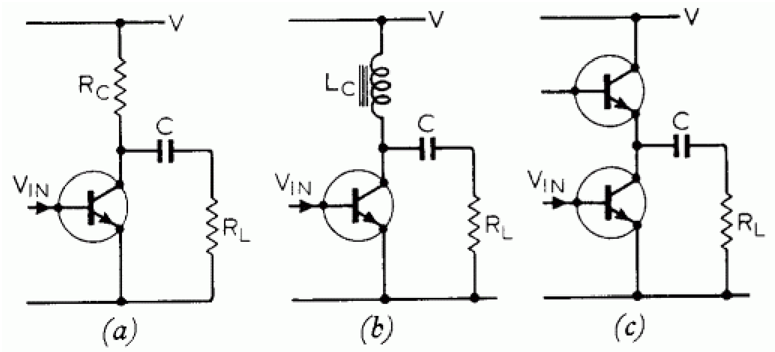

Fig. 1. Basic class A circuits using (a) load resistor Rc giving power conversion efficiency of about 12%, (b) l.f. choke giving better effieciency but being bulky and expensive, and (c) a second transistor as collector load.

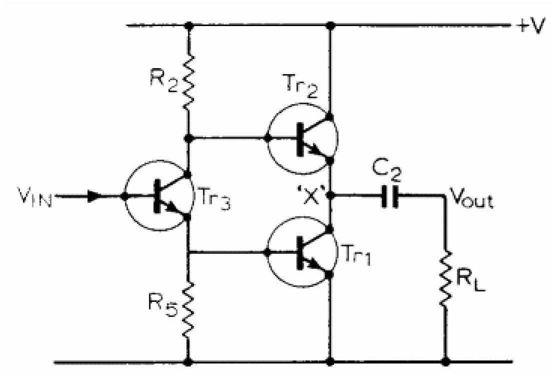

Fig. 2. Arrangement for push-pull drive of class A stage.

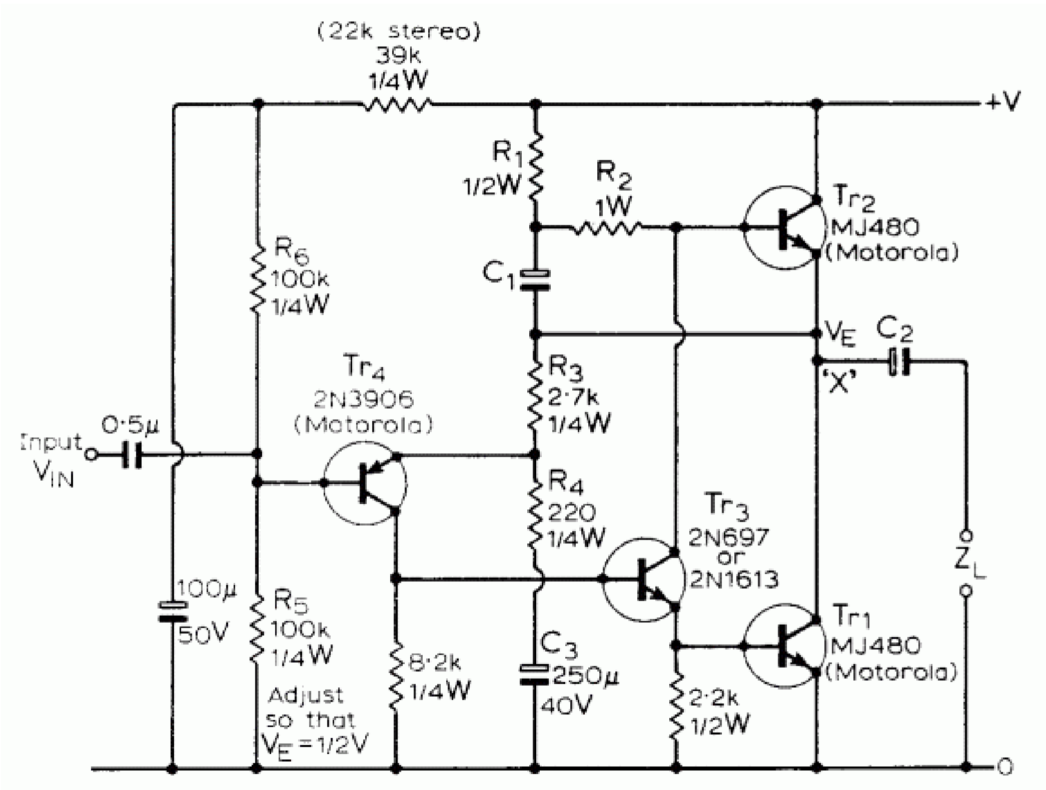

Fig. 3. Practical power amplifier circuit.

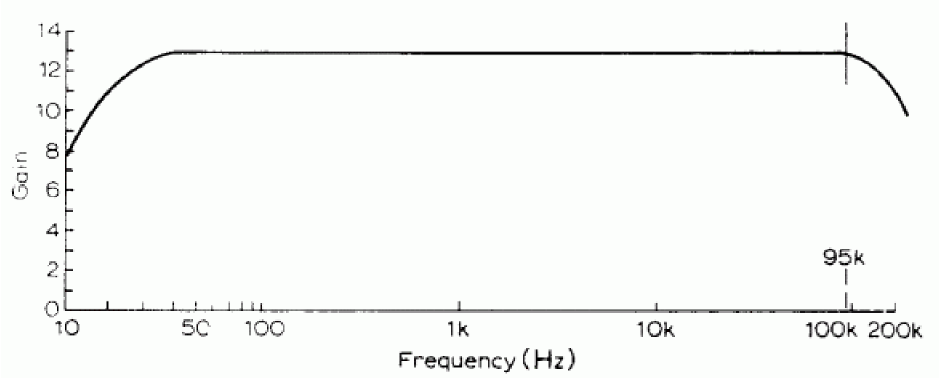

Fig. 4. Gain/frequency response curve of amplifier.

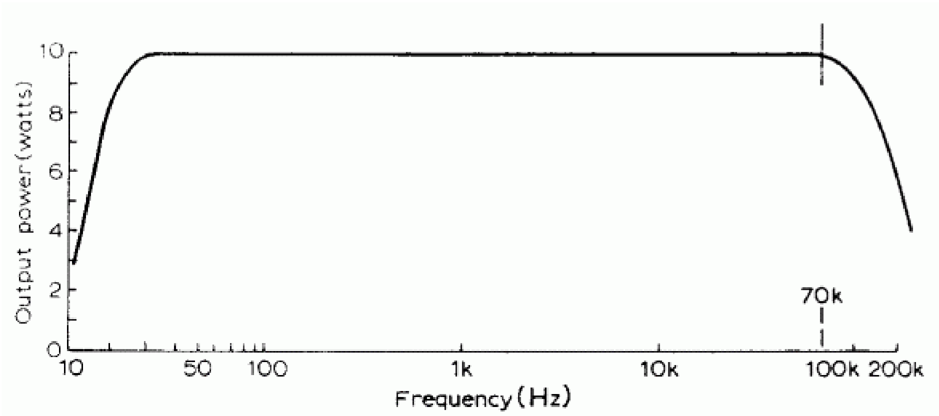

Fig. 5. Output power/frequency response curve of amplifier.

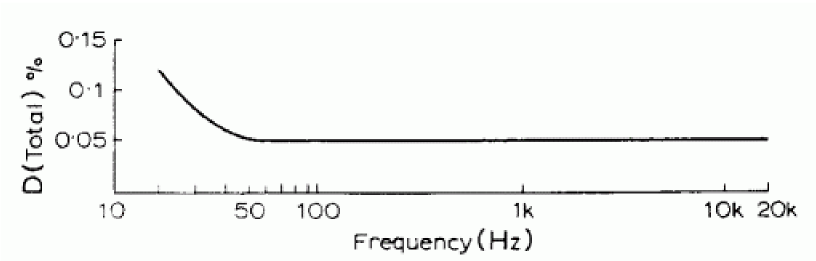

Fig. 6. Distortion/frequency curve at 9W.

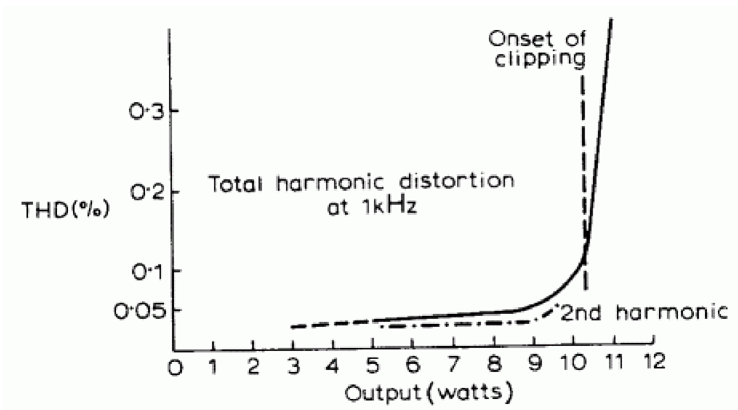

Fig. 7. Distortion/output power curve.

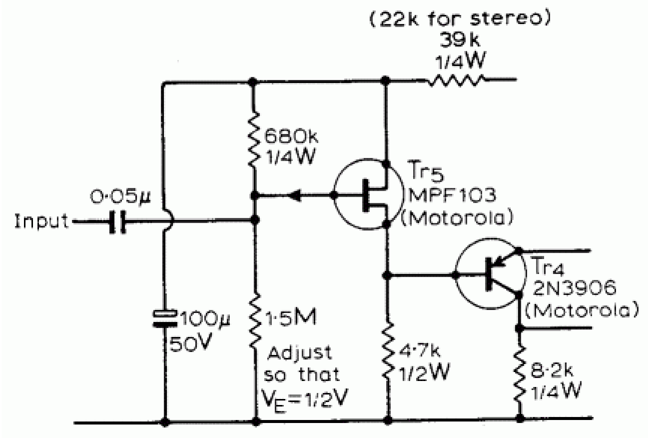

Fig. 8. Modified input circuit for high input impedance.

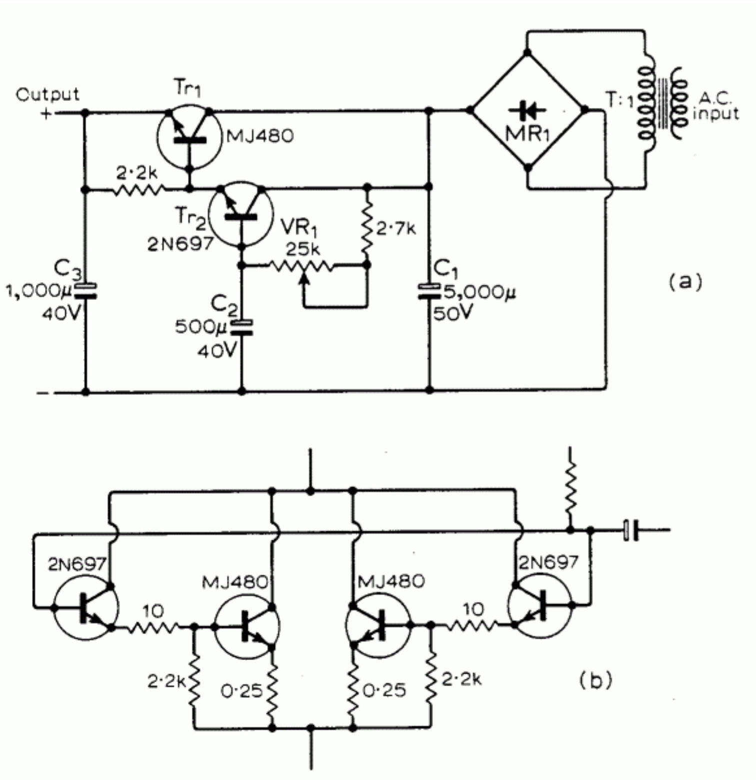

Fig. 9. (a) Power supply unit, and (b) parallel connected transistors for high currents.

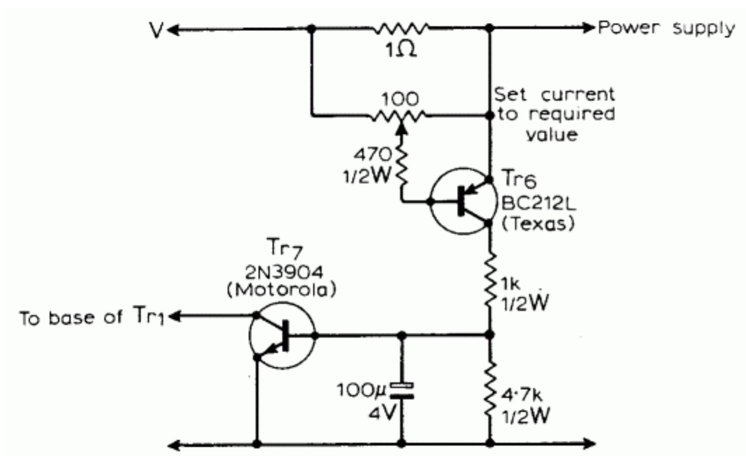

Fig. 10. Amplifier current regulation circuit.

Simple Class A Amplifier

A 10-W design giving subjectively better results than class B transistor amplifiers

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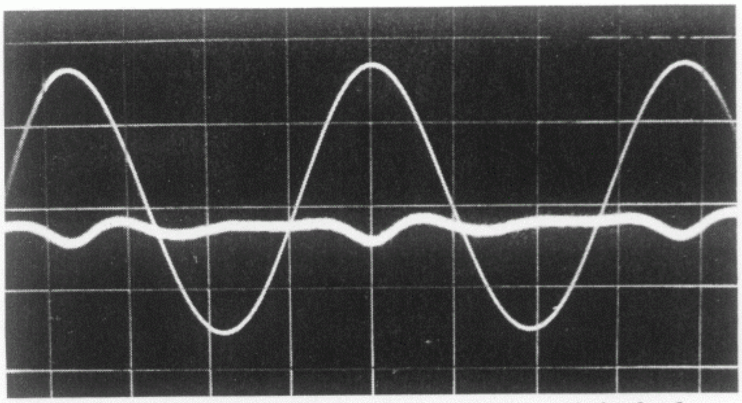
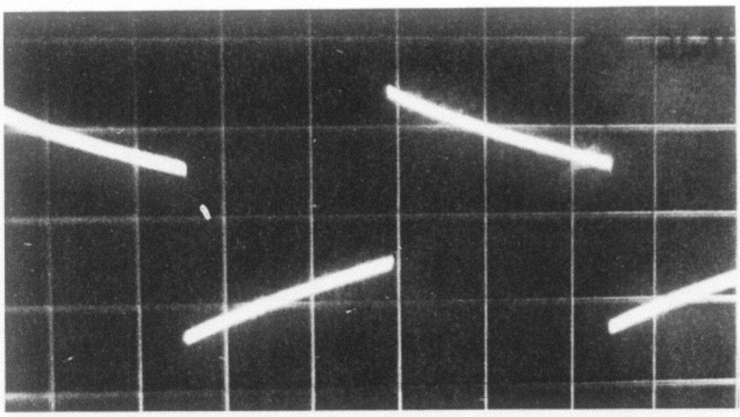
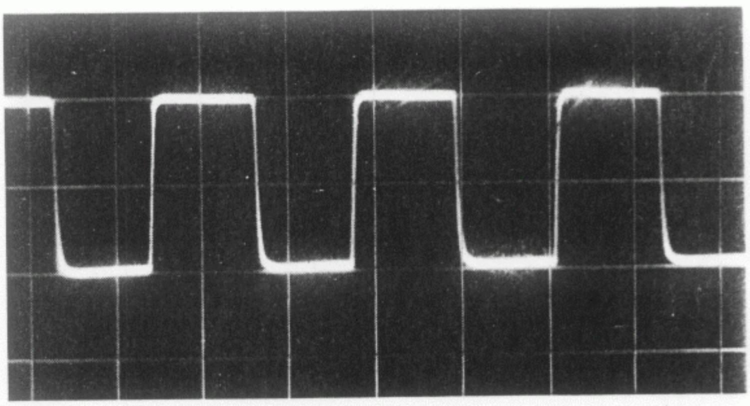
Oscilloscope Traces



Sine wave performance at 1kHz. 9 watts, 15 ohm resistive load. Fundamental on scale of 10V/cm. Distortion components on scale of 50mV/cm with r.m.s. value of 0.05%.

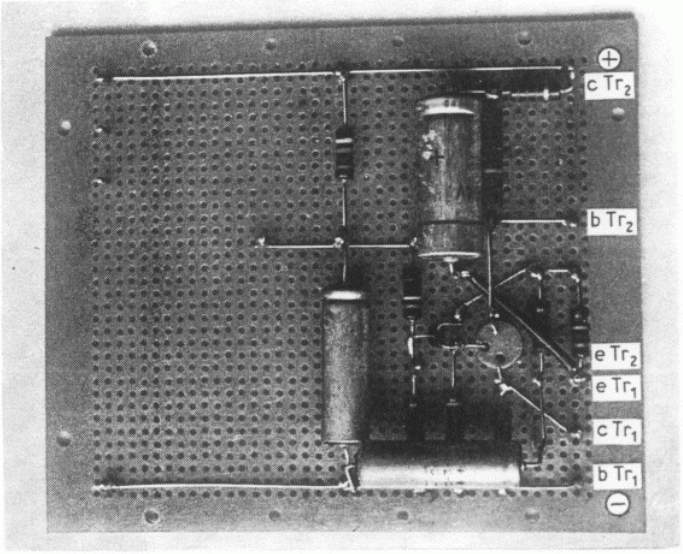


Square wave response at 50Hz.

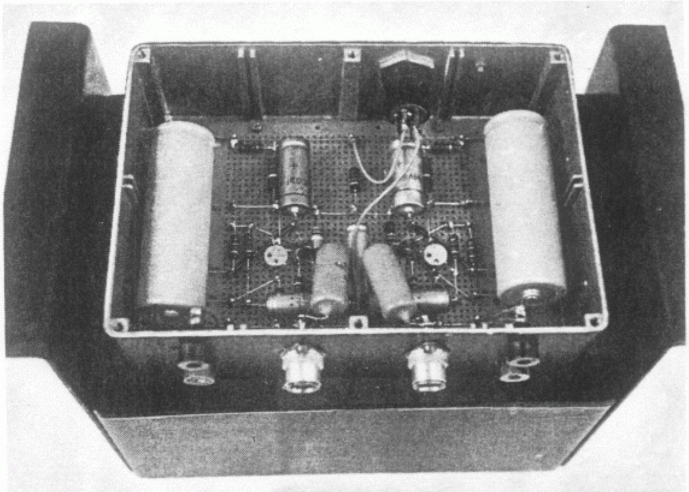


Square wave response. Scale 10V/cm. Frequency 50kHz. 15 ohm resistive load.

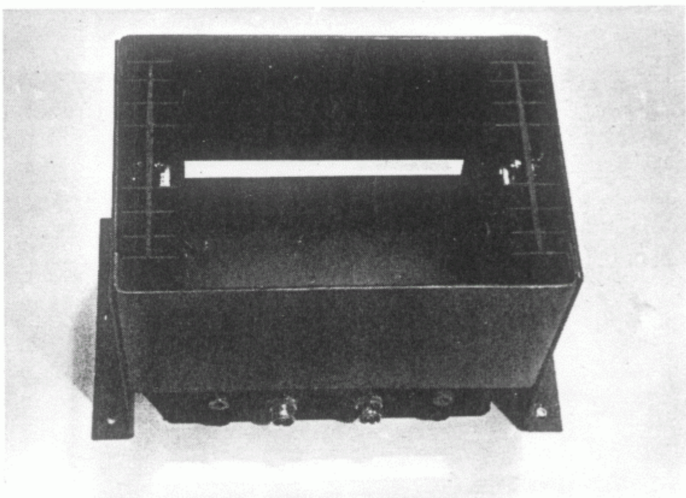
Photographs



Layout of single channel of 10 + 10 watt amplifier on standard 4in × 4¾in 'Lektrokit' s.r.b.p. pin board.



Underside of completed amplifier, with base cover removed, showing external box-form heat sink.



Looking down on the completed amplifier.

Letters to the Editor

Linsley Hood class A amplifier

Recent measurements on this amplifier have indicated that the gain and power bandwidths of this design, using the component layout shown on page 152 of your April 1969 issue, are wider than indicated by the Figs. 4 and 5 of the article. The apparent fall-off in gain beyond about 100kHz was, in fact, due to shortcomings in the measuring apparatus, and measurements made with better equipment suggest that the –3dB points for voltage gain are above 1.5MHz although the power output falls beyond 200kHz.

Since the output is in phase with the input, it is necessary to take care that the output leads and the output capacitor are not close to the input. (A 2-inch separation will be adequate for normal lead lengths.) However, an additional point must also be noted. If a capacitive load is connected with short leads between the output and the earth line near the input connection, the potential developed along the earth line, due to its inductance, can inject an in-phase signal, and thereby cause instability, in the MHz region. To avoid this possibility, it is recommended that the earthy lead to the loudspeaker terminal be returned to the earth line at the same point as the emitter of Tr1. The inclusion of a small r.f. choke (25 turns of 26-28 s.w.g. wire wound round the outside of a 10-ohm 1-watt resistor is ideal) between the output (point 'X') and C2 will also prevent this possibility of trouble.

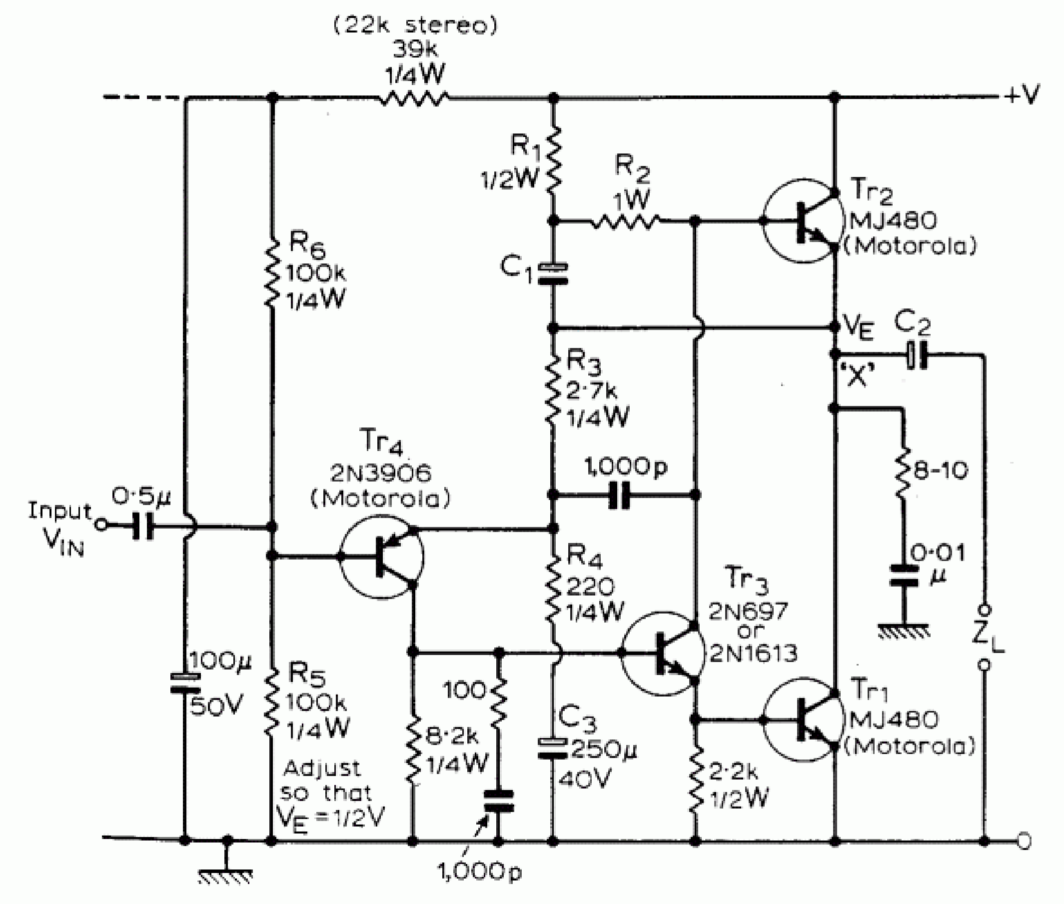
In practice, with the components and layout suggested, the inductance of the normal 12 to 18 inches (or more) of loudspeaker connecting lead prevents instability with capacitive loads, so this should be only of academic interest.

As an alternative, it is possible to reduce the r.f. response of the amplifier to give a smooth 6dB roll-off beyond 50kHz – which removes much of the need for care in the layout of components, without detriment to the harmonic distortion in the audible range, and without any audible alteration to the performance – by connecting a 1,000pF capacitor between the collector of Tr3 and the emitter of Tr4; a 1,000pF in series with 100 ohms between the collector of Tr4 and earth; and a 0.01µF in series with 8 to 10 ohms between the output ('X') and earth. (It should be noted that either all of these components should be added or none at all, they are not alternatives.) If the r.f. response is reduced in this manner, the use of a series r.f. choke would be unnecessary.

A series of measurements has also been made, using the amplifier design exactly as described in the article (without r.f. chokes or other modifications), to determine the voltage waveform produced, actually across the loudspeaker, with a square wave input to the amplifier. It was found, in practice, with several different loudspeaker systems, that the output waveform was virtually identical to that obtained with an equivalent resistive load – photographs of which were reproduced in the April issue. It was, in fact, the discovery that a good square wave was reproduced up to the 1MHz limit of the generator in use which prompted a reassessment of the r.f. response of the amplifier. The absence of any overshoot or significant ringing also provides confirmation of the stability of the amplifier under practical conditions.

A correspondent has reported that this design has been up-rated successfully to 15 watts into a 15-ohm load, to give a direct power equivalent to the Williamson amplifier, using 2N3055 output transistors with a 43-volt supply (1.1 amp per channel), and rather larger heatsinks. There would seem no good reason why this could not also be done using MJ481s.

J. L. LINSLEY HOOD



Mr Linsley Hood's amended circuit of his class A amplifier originally described in the April 1969 issue.

Simple Class A Amplifier

A postscript to the design published last year

by J. L. Linsley Hood

The author has had the benefit of an extensive and frequently helpful correspondence with readers following the publication of the circuit design. Attention has been drawn to some obscurities in the original article and to certain possible improvements in the design. Details are given below.

Power supply

Although much interest was aroused among constructors by the good technical (and audible) performance given by the amplifier, it is clear that the principal feature in the eyes of many users was the relative simplicity of the circuit. This being so it must have seemed a pity that the power-supply unit was somewhat complex. However, the purpose of this power-supply design was to avoid possible degradation of the amplifier performance by h.t. ripple. The sawtooth ripple voltage across the reservoir capacitor in a class-A amplifier drawing some 2-3A will be many times greater than in a class-B system, particularly at the audibly important low-signal levels.

However, a number of measurements have been made since the publication of the original article on an amplifier of this type, operated from a simple supply unit of the type shown in Fig. 1. There is little difference in the performance above 100Hz either in total harmonic distortion or in intermodulation distortion, although the shape of the output power/distortion curve at the onset of overload is modified, as would be expected, by the ripple on the h.t. line. Below 100Hz the distortion curve rises more steeply to about 0.2% at 20Hz.

A thermister is necessary, in this case, to slow down the rate of rise of the h.t. voltage. This will get hot in use.

It now appears that the mains transformers used in the development of the prototype of this amplifier were not as efficient in respect of apparent secondary circuit resistance or secondary leakage reactance as some of those which have been supplied for this purpose since the publication of the article. The reservoir voltage found with the 15Ω system may be above that given by the author. At switch-on this can cause a transient overloading of the transistors specified for the series regulator circuit in the original article. In view of this, it is suggested that these should be an MJ481 or 2N3055, used in conjunction with a 2N1613 or, better still, a 2N699. These amendments are shown in Fig. 2.

Adjustment of amplifier output current and centre-line potential

The author had supposed, somewhat naively, that most constructors of the circuit would have somewhere in their workshops a collection of odd-value resistors needed for trimming circuit parameters, and it was mentioned in the original article that the desired quiescent levels could be set by adjustment to R2 and R5 or R6. This sort of comment is unhelpful if one is writing away for a kit of parts. In view of this it is suggested that R2 should be replaced by a resistor in series with a potentiometer, as shown in Fig.1. The necessary value of resistors R5 and R6 to give an entirely adequate accuracy in the mid-point voltage setting can be predicted, and the suggested amended values are shown.

Some obscurity arose, inadvertently, in the original diagram concerning the reason for the different values of input decoupling resistor quoted for mono and stereo use. This was because it was intended that the one decoupling circuit should serve both channels. Where an unsmoothed h.t. supply is used it is recommended that the decoupling capacitor should be increased in value to 250µF.

Stability of output current setting

Some criticism has been voiced because there is no specific control over the output current value in the simplest form of this circuit, other than that due to the stability of the current gain of Tr2, whose performance determines this parameter. In order to meet this point (in anticipation) a circuit was described in the original article which allowed precise control over the operating ‘quiescent’ current without detriment to the performance of the amplifier.

However, measurements made on an amplifier without this addition have shown no significant change in operating current in somewhat over two years use, and there is also little measurable difference in current from a minute or so after switch-on to the end of a six-hour period of continuous use. In practice therefore, in temperate climates at least, the simplest form of the circuit is adequate in this respect. If any user cares to experiment with an alternative and somewhat more elegant form of quiescent-current control another regulation circuit is shown in Fig. 3. The transistor used as Tr5 requires to be somewhat more massive than that used for Tr3 since the mean collector current is twice that of Tr3 and the maximum voltage and current occur simultaneously. The 2N2905A is just about adequate with a good heatsink, but a larger power device such as the 2N4919 is preferable.

Alternative transistor types

The amplifier has been built successfully with a wide variety of transistors, including fully complementary versions to operate from an existing negative h.t. line, and in one case two identical amplifiers have been made for use with the inputs in paraphase, in order to double the available output voltage swing. One constructor has, indeed, made a stereo 30W system using two such pairs of amplifiers plus input phase splitter, as shown in Fig. 4.

However, one transistor change which is recommended is the use of a 2N1711 as Tr3. This has a high voltage capability equal to that of the 2N1613, and a current gain which is double that of either the 2N1613 or the 2N697. The use of the 2N1711 instead of the former types suggested for Tr3 increases the feedback factor and approximately halves the typical distortion factor of the system (0.025% at 9W or 0.05% at full power) without detriment in other respects.

Also, a 2N1711 as Tr3 allows the use of 2N3055 devices as Tr1 and Tr2, with a final performance which is equal to that of the original specification below 100kHz. (The typical current gain of the 2N3055s is only half that normally found with the MJ480/1 output transistors, and their use was not originally recommended for this reason.)

Gain/frequency and power/frequency characteristics

These are, in fact, better than the curves published in April 1969. As mentioned in a letter to the editor published in October 1969, the h.f. fall-off shown was mainly due to an error in the measurement instrument. Although the performance at h.f. depends to some extent on the layout employed, the small signal voltage gain, with the component arrangement shown, is flat (within 1dB) to beyond 2MHz. This may be a snag in some cases because even a small feedback capacitance between output and input (as may happen, for example, if the output heatsinks are not earthed) may cause the amplifier to oscillate. A suitable circuit change to reduce the amplifier h.f. response to more normal levels was described in the letter above. This is not an essential modification – the author’s own units are still exactly as described in April 1969.

The output power response of the unmodified amplifier is flat within 1dB to 200kHz.

The l.f. response shown in the original gain/frequency and power/frequency graphs was that determined for an earlier prototype of the amplifier. During the development of the circuit the values of some of the capacitors were increased to improve the l.f. performance, and by an oversight the graphs accompanying the article were not amended. In fact the gain and power graphs can be shown as ‘flat’ from 10Hz-200kHz. In this respect, and that of transient response, the class-A design is probably better than any circuit so far published. The i.m. distortion, at 10W output, (70Hz and 7kHZ, 4:1) is less than 0.1%.

Miscellaneous

Surprise – and even alarm – has been caused to some constructors by the fact that the output transistors get hot. However, with adequate heatsinks, which should be black painted, the dissipations in the transistors are only a small fraction of the maker's permitted level, and provided that some care is taken in the layout to make sure that sensitive components, such as electrolytic capacitors, remain cool, no reduction in the working life of such a system, in comparison with an equivalent class-B unit for example, is to be expected.

Some difficulty has apparently been encountered by some constructors because the power supply regulation system is inoperative when the supply is operated without a load. If an equivalent resistive dummy load is connected for bench-testing, all should be found to be well.

Finally, it is prudent to wire a small resistor of about 2kΩ across the loudspeaker terminals to make sure that the output capacitor charges even with the speaker disconnected. Charging of the capacitor by an accidental short-circuit could cause damage. This addition is shown in Fig. 1. No damage is caused by operating the amplifier on an o / c output.

Figures

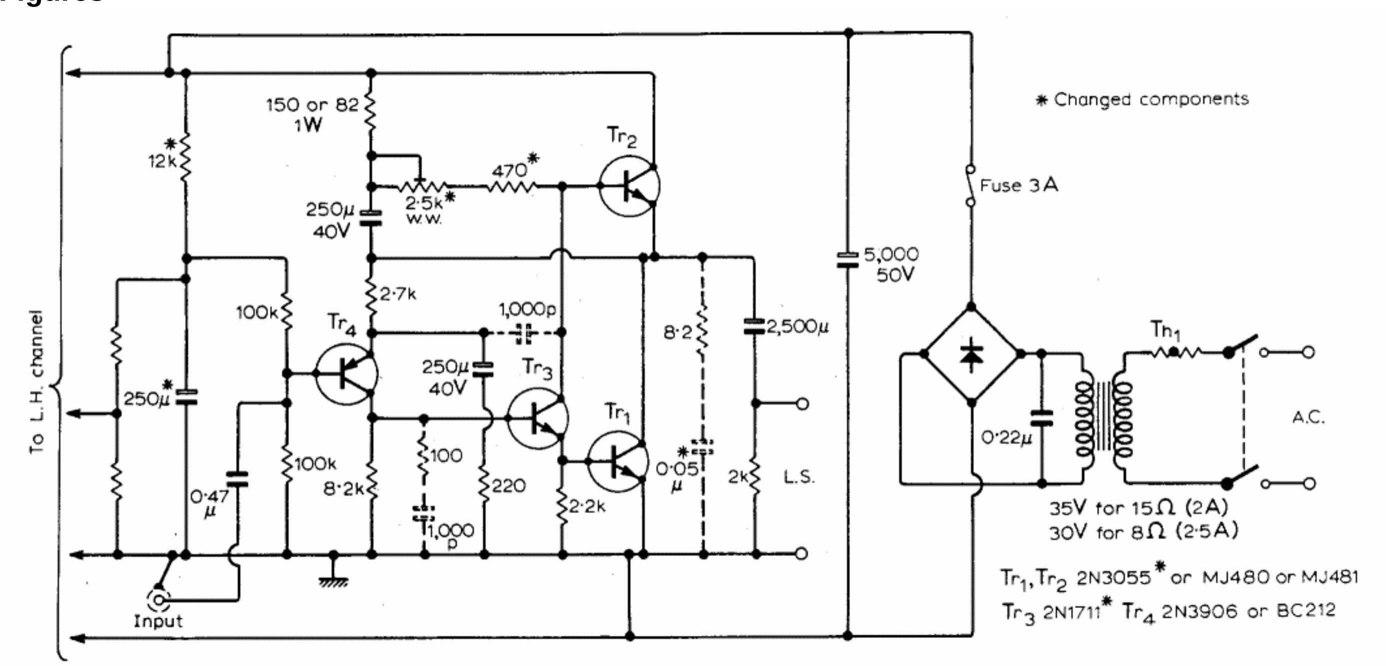


Fig. 1. Suggested amended circuit for 8 or 15Ω use employing a simplified power supply. The dotted components reduce the h.f. response and should be used with capacitive loads.

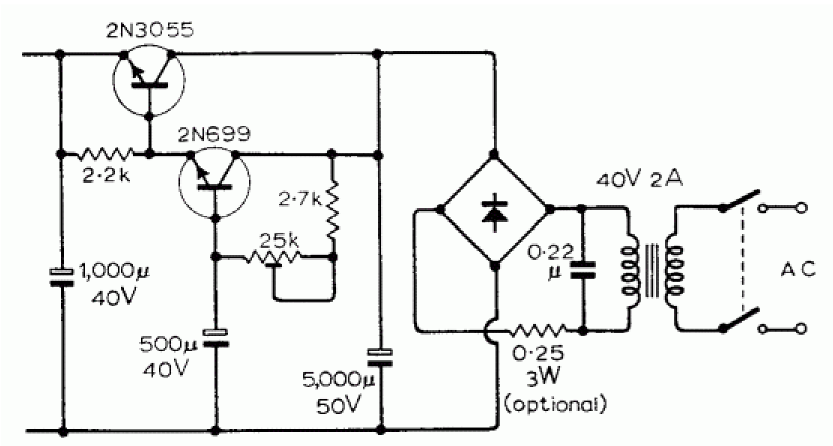


Fig. 2. Amended circuit of power supply for 15Ω systems.

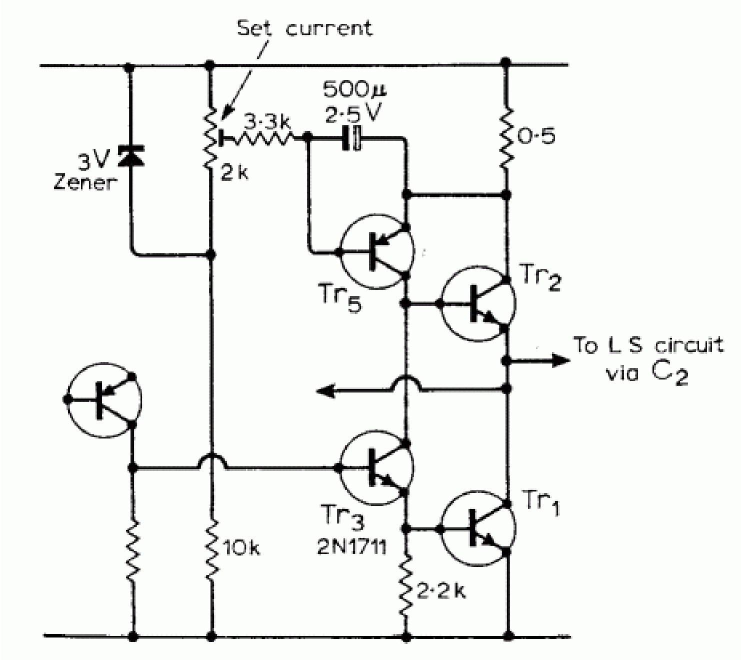


Fig. 3. Alternative method of quiescent-current control. R1, R2 and C1 in the original have been deleted. Tr5 is 2N4919 on heatsink or alternative type.

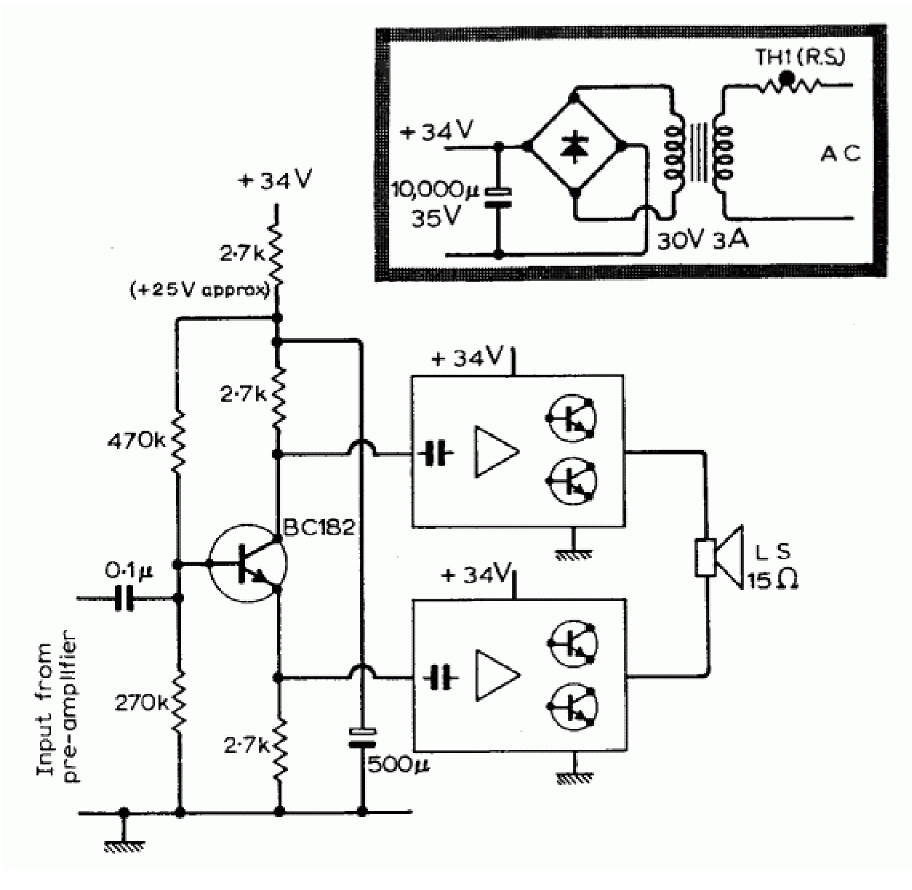


Fig. 4. Scheme for 30W class-A system. The two identical class-A amplifiers, each designed to give 15W into 8Ω (34V at 1.5A), are driven by a phase-splitter. The output capacitors have been removed.

Class-A Power

After two and a half decades, John Linsley-Hood's Class-A power amp is still rated among the best. Here, John explains how to bring the design up to date, adding enhancements such as dc-coupled output.

The current debate, among the more reactionary of the hi-fi devotees, about the relative merits of thermionic valve operated audio amplifiers makes intriguing reading, if only because, in a sense, this is 'where I came in'. I will explain.

I have had an interest in the reproduction of music, principally from gramophone records, for a very long time. I made my first, two-valve, battery-operated, audio amplifier as a twelve year old school boy, some time before the outbreak of the 1939-1945 war.

This gave way – in the interests of economy, – to a series of mains powered amplifiers, which were usually combined with a radio receiver. Electricity from the mains was free, to me at least, whereas high-tension batteries had to be bought from my pocket money.

My early work culminated, in 1951, with the assembly of a luxurious kit for the highly esteemed high-fidelity Williamson 15W amplifier design. Although, by this time, I had my first proper job – in the electronics labs of the Sellafield nuclear research establishment in Cumberland – and cash was a bit more plentiful, I still wouldn't have built that particular, rather expensive version of the hardware if I hadn't heard through the lab grapevine that one of the research chemists had bought himself a Williamson kit, but, on receiving the parcel, lacked the courage to assemble its contents. Rumour had it that he was open to offers, and I was happy when he accepted mine.

This was an excellent amplifier, and was better, in my judgement, by a greater or lesser extent, than any of its predecessors of my own design, or, indeed, any of the other valve amplifiers, belonging to my friends, with which I had had a chance to compare it. It gave me great pleasure until early 1968, when I replaced it with a solid-state equivalent.

What I replaced it by, and the circumstances of this replacement, were described in an article in Wireless World in April 1969, entitled 'A simple class A amplifier'. This was a long time ago. In the light of the current debate, it seems possible that both my listening trials at the time, and an up-dated version of my original class A design, may be of interest to you. By up-dated, I mean using more modern components and delivering a bit more power output.

The Williamson Amplifier

In the inter-war years, with the improvement in audio quality of both gramophone records and radio broadcasts, considerable attention was paid to improving the quality of ac mains-powered audio amplifiers. A number of interesting designs were offered. These were mainly based on the use of push-pull output stage layouts. Relative to straight single ended circuits, push-pull stages would give greater output power for a given distortion level.

At that time, there were audiophiles who decried the use of push-pull output stage layouts. They claimed that the best audio quality was only obtainable from the much less efficient single ended arrangements, i.e. those in which the output valve had a simple resistor, choke or output transformer load. Interestingly, this is a claim which was examined and dismissed by Williamson at the time, but which has recently been resurrected.

Using negative feedback

Almost all valve operated audio power amplifiers require an output transformer to match the relatively high output impedance of the valve output stage to the low impedance load presented by the loudspeaker.

In general, the transformer is the most difficult and expensive part of the system to design and construct. This is because of the following conflicting demands:

• For a low leakage reactance – combining both leakage inductance and inter-winding capacitance – from the primary to the secondary windings, to avoid loss or impairment of high frequency signal components.

• For a low level of leakage inductance from one half of the primary to the other, to reduce the discontinuities due to push-pull operation, and the odd-order harmonic distortion resulting from these.

• For a high primary inductance, to give a good low-frequency response.

• For a low winding resistance, to avoid power losses.

• For a good quality grade of core laminations to ensure a low level of core-induced distortion, due to magnetic hysteresis and similar effects.

Intrinsic signal distortion of a valve amplifier stage could range from 0.5 to 10%, depending on its circuit form and operating characteristics. It had been appreciated for some time that such intrinsic distortion could be reduced significantly by applying local negative feedback. Various amplifier designs incorporating local negative feedback had been proposed. However, this still left the output transformer – however well made – as a major source of transfer and frequency response non-linearities.

At this point, D. T. N. Williamson, who was working at the time as a development engineer for the valve section of the GEC Research Laboratories, described a high-quality audio amplifier design, using the recently developed GEC ‘kinkless tetrode’ output valve, namely the KT66. In this design, a single overall negative feedback loop embraced both the whole of the amplifier and the loudspeaker output transformer.

With the exception of the output valves, which were triode connected KT66s, Williamson’s design employed triode amplifier valves exclusively, because these had a lower intrinsic distortion figure. He also made use of extensive local negative feedback, provided by un-bypassed cathode-bias resistors. This had the additional benefit of eliminating the electrolytic bypass capacitors – a philosophy which is in accord with much of contemporary thinking.

Williamson also used non-polar rather than electrolytic high-tension reservoir and smoothing capacitors, in the interests of more consistent ac behaviour. Electrolytic capacitors were much worse at that time.

If overall negative feedback was to be applied without causing either high or low-frequency instability, careful design was essential – both in the amplifier stages and in the output transformer. These problems had frustrated earlier attempts to do this – but Williamson demonstrated that it could be done.

The performance given by his design, if his detailed specifications were carried out to the letter, was superb. The performance criteria of better than 0.1% thd, at 15W output, from 20Hz to 20kHz, and a gain bandwidth from 10Hz to 100KHz +/- 1dB, are at least as good as those offered by many of today’s better commercial designs.

The series of articles written by Williamson, in Wireless World over the period 1947 – 1949 described the power amplifier and its ancillary units. This series had enormous impact on audio design thinking, and if I may quote the WW editor of the time, in his introduction to a reprint of all these articles.

“Introduced in 1947 as merely one of a series of amplifier designs, the ‘Williamson’ has for several years been widely accepted as the standard of design and performance wherever amplifiers and sound reproduction are discussed. Descriptions of it have been published in all the principal countries of the world, and so there are reasonable grounds for assuming that its widespread reputation is based solely on its qualities”.

All in all, the Williamson was a hard act to follow.

Alternative hardware

The world had not stood still since 1951. My equipment had remained monophonic, while the rest of the audio world was changing over to stereo.

My main interest was in music, not in circuitry, so I thought it would be prudent to ask my ears what they thought of the alternatives, before I started to replace my hardware.

To this end, I built or borrowed six well thought-of audio amplifiers, my own Williamson, a Quad 2, two dissimilar but recently published class AB transistor amplifiers, a commercial 30W solid-state unit, and a simple Class-A unit of my own design.

I included the Class-A design out of curiosity. If it turned out to be any good, it would be cheap and easy to build. It was not expected to offer any special merit in performance.

In the event, as I reported at the time, (WW April 1969, p.152), the six amplifiers divided quite clearly into two separate tonal groups. The three class AB transistor amplifiers formed one group, while the two valve amplifiers and the simple class A amplifier formed the other.

To be fair, the differences between any of these were not very great – but they were audible. Once they were noticed, they tended to become more apparent on protracted periods of listening. Certainly, for me – and I was doing these tests for my own benefit – in these comparative trials, the two best were the Williamson and the class A. They were virtually indistinguishable. Of these two, the Williamson was vastly more massive and costly to construct.

The only remaining question was, if I replaced the 15W Williamson with the 10W Class-A design, would the output be adequate? Connecting an oscilloscope across the loudspeaker terminals showed that I seldom needed more that 2-3W from the power amplifier – even under noisy conditions.

I suppose that the final proof of my satisfaction with the class A transistor amplifier was that, a year or so later, I gave my old Williamson to a friend.

Valves versus transistors

Not all of the considerations of valves versus transistors relate solely to performance. It is worth bearing in mind that products involving obsolete technology will be disproportionately expensive, difficult to obtain and possibly of inferior quality.

Valves can also vary in operating characteristics from sample to sample – especially where two valves of the same type are obtained from different sources. Characteristics that can vary are mutual conductance, gain, operating grid bias, anode current impedance, and even usable anode voltage.

By comparison, the performance characteristics of, say, a range of 2N3055 epitaxial base output transistors are almost identical, whether made in the Philippines or in Toulouse.

Again, all valves deteriorate in use, exhibiting a gradual loss of cathode emission over a typical 3000 hour service life. If a valve is persistently over-driven, the heating of the anode may cause the metal to out-gas. This impairs the vacuum essential to proper operation, and shortens the valve’s life.

A further consideration is that valves are high voltage devices, which can be dangerous. And the need for high working voltages can lead to more rapid failure of other components in the circuit – especially capacitors.

The class A design

My original design is shown in Fig. 1. This is still a valid design, except that the MJ480/481 output transistors are now obsolete. However, they can be replaced by the more robust 2N3055. In this case, the epitaxial-base version of this device should be chosen rather than the hometaxial, since the f_T of the output transistors should be 4MHz or higher.
As I commented, at the time, the design gave a somewhat lower distortion if the h_{FE} of Tr1 was greater than that of Tr2. This caused the output circuit to act as an amplifier with an active collector load rather than an output emitter follower with an active emitter load.

A simple modification which takes advantage of this effect is the use of a Darlington transistor such as an MJ3001 for Tr1. At 1kHZ, this reduces the distortion level at just below the onset of clipping from about 0.1% down to nearer 0.01%. As before, the residual distortion is almost exclusively second harmonic. Also, as before, it fades away into the general noise background of the measurement system as the output power is reduced.

While this transistor substitution seems to be a good thing, it was not a modification whose effect I was able to check, in listening trials, against the Williamson. As a result, for the sake of historical fidelity, I would still recommend the use of epitaxial-base 3055s as Tr1 and Tr2.

I have checked all the other changes which I have proposed with the exception of the power increase.

Improving performance

With regard to the original 10W design, as published, I feel the following improvements will be beneficial:

- Provide a more elegant means of controlling output transistor operating current by including a variable resistor in the base of Tr2.

- Arrange the circuit so that it would operate between symmetrical power supply lines, allowing the amplifier to be directly coupled to the loudspeaker.

- Increase output power from 10 to 15 watts per channel

- Up-grade the smoothed but not regulated power supply arrangement.

In my postscript to this design, which WW published in December 1970, I suggested both alternative transistor types and an improved method of adjustment and control of the output transistor current flow, Fig. 2.

Although, in theory, this layout should give a superior performance, when I changed my prototype amplifier to this arrangement, I found little change in measured thd and I couldn't hear any difference in sound quality.

Although directly coupling the amplifier to the loudspeaker will not have much effect on thd, it is still beneficial since it eliminates the output coupling capacitor. The most obvious way of doing this is to rearrange the input layout, around Tr4, so that it becomes the input half of a 'long-tailed' pair.

I am reluctant to do this because this would alter the overall gain/phase characteristics of the amplifier. It would also require additional high-frequency stabilisation circuitry, with all its incipient problems of transient intermodulation or slew-rate limiting.

Fortunately, the need to remove the dc offset at the output can be achieved without altering the good phase margins of the design, by simply injecting an appropriate amount of current into the base circuit of Tr4.

Output power and dissipation

In essence, all that is required to increase the power output from the amplifier is to increase the rail voltages and the standing current through the output devices. Restrictions are that power consumption must remain within the confines of what the mains transformer and rectifier can deliver. Also, the heat-sinks must be able to dissipate the extra heat and the output transistors must be adequately rated.

For a 15W (sinusoidal) output into an 8Ω load, an $11V_{RMS}$ drive voltage is required. This, in turn means a $31V_{P-P}$ voltage developed across the load, and an output current into the load of $2A_{P}$. Since the circuit is a single-ended configuration, in which the collector current will not increase on demand, this means that the output transistor operating current must be at least 2A to allow this.

With the circuit shown, using the improved current control layout – which is rather less efficient than the boot-strapped load for Tr3 which I originally proposed – the rail voltage needed is +/- 22V.

This will lead to a dissipation, in each output transistor, of 44W. Prudence suggests that a heatsink having a rating of no more than 0.6°C/W, should be used for each output pair.

Most 2N3055s have a Vce of 60V, a maximum collector current of 15A, and a maximum dissipation, on a suitable heatsink, of 115W. However, RCA's 3055, and its complementary MJ2955, are rated at 150W.

Working conditions for the output transistors are entirely within the devices safe operating area, so no specific overload protection circuitry is needed. Even so, the inclusion of a 3A fuse in the loudspeaker output line would seem prudent.

DC offset cancellation

Figure 3 shows the full circuit for one channel of the 15W Class-A audio amplifier. I have inserted a 15V three-terminal regulator ic into the positive rail to prevent any unwanted signal or hum intrusion into the emitter of Tr4.

It is easy to set the dc offset to within +/-50mV. The offset does not change greatly with time, although this assumes that Tr5 is not allowed to warm up too much. This is because the base-emitter potential of this transistor controls the operating current, which in turn, affects the output dc offset.

Small-signal bandwidth

In the original circuit the small-signal bandwidth was 10Hz–250kHz, +/-3dB, which was needlessly wide. Because of this, I have added an input high-frequency roll-off network, R3/C2, to the input circuit to limit the top end response to some 50kHz. This assumes an input source impedance of 10kΩ or less.

As it stands, the low-frequency –3dB point is about 7Hz. It can be lowered even further, if necessary, by making C1 larger – say to 1µF.

Supplying power

As was shown in the 1970 postscript, it is possible to operate this amplifier from a simple rectifier/reservoir capacitor layout. Fig. 4 is an example. The only penalty is a small 100Hz background hum, probably about 3mV in amplitude. However, I feel that, if you are seeking the best, a proper regulated power supply is preferable, Fig. 5.

The circuit shown for the current booster pass transistors, Tr1/Tr2, is one suggested by National Semiconductor. It takes advantage of the internal current limiting circuitry of the 7815/7915 devices to limit the short-circuit current of these ICs to 1.2A. By choosing the correct ratios of R5:R7 and R8:R10, the short-circuit current drawn from Tr1 and Tr2 will also be limited.

For a satisfactory ripple free dc supply of +/-22V, the on-load voltage supplied to the regulator circuit should be +/-27V.

Performance

I prefer measurements made with appropriate instruments to judgements based on listening tests.

Measured distortion is less than 0.1% near the onset of clipping. It fades away into the background noise level of the measuring system as output power level is reduced.
For me, the fact that the distortion given by this circuit is almost pure second harmonic is more persuasive of its performance than any 'golden eared' judgement of tonal purity.

If you then add the observation that the circuit remains stable on a square-wave drive into typical reactive loads, I am not surprised that its performance was capable of equalling the Williamson on listening tests. No significant overshoot is observed on the square-wave, and stability is achieved without the need for internal high-frequency compensation arrangements.

So, as a final thought, if any of you want to find out how a top quality valve amplifier like the Williamson sounds, you can find out at a tenth of the cost of building one by making up this Class-A design. It has the additional advantage of incorporating readily available and modern components.

Figures

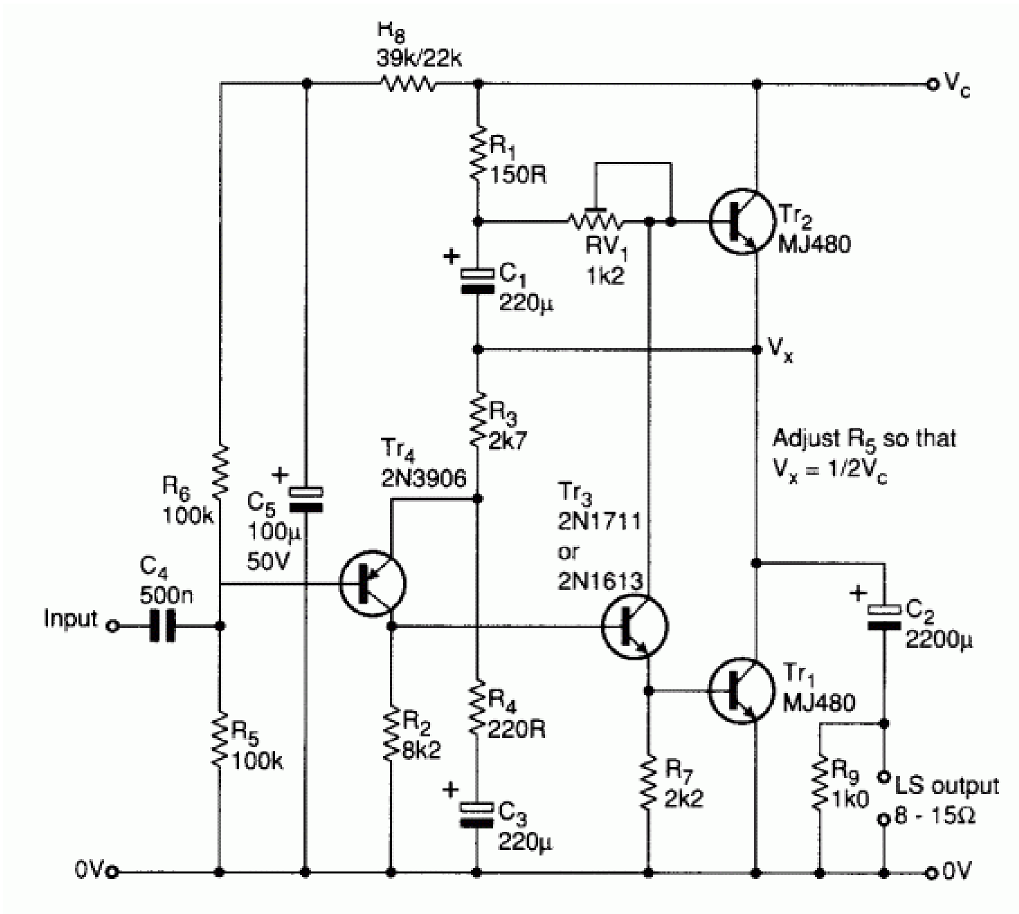


Fig. 1. Original 10W Class-A design is still valid, but the power devices are now obsolete.

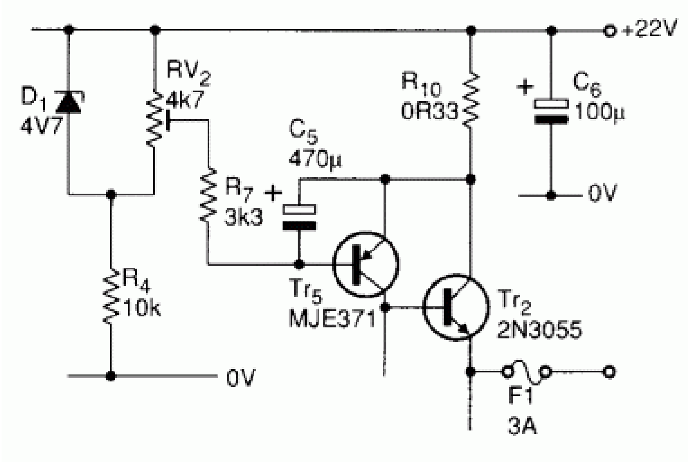


Fig. 2. Improved method of adjusting quiescent current, suggested as a postscript to the original design.

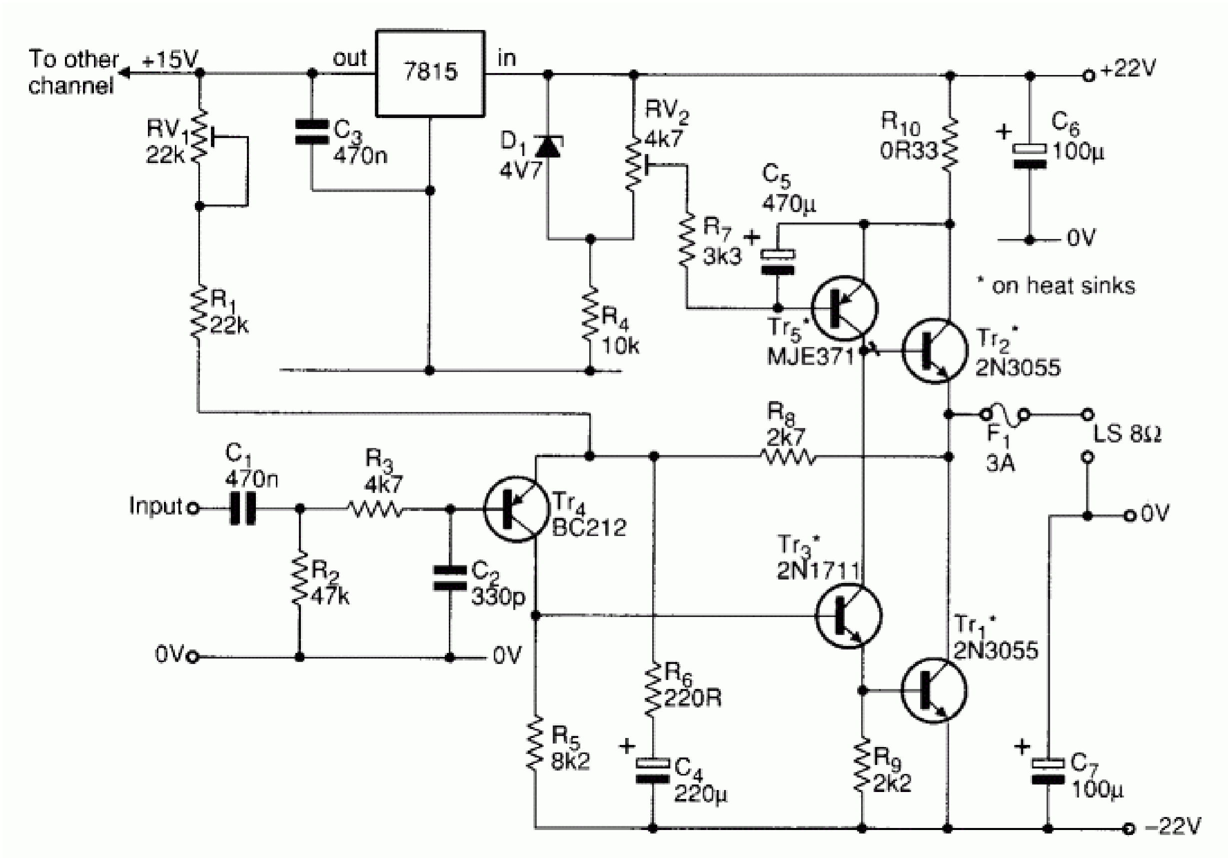


Fig. 3. One channel of the enhanced 15W Class-A design incorporating – amongst other things – direct loudspeaker coupling.

Note: There is an error in this diagram. The negative end of C4 should be connected to the 0V (earth) point and not the –22V supply rail as shown. Failure to do this will result in excessive hum due to supply rail ripple being injected into the negative feedback path (Tr4 emitter).

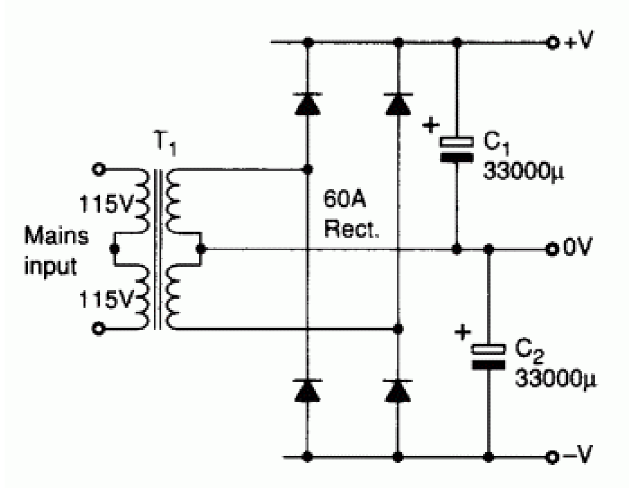


Fig. 4. Simple but adequate dual-rail supply using a single bridge.

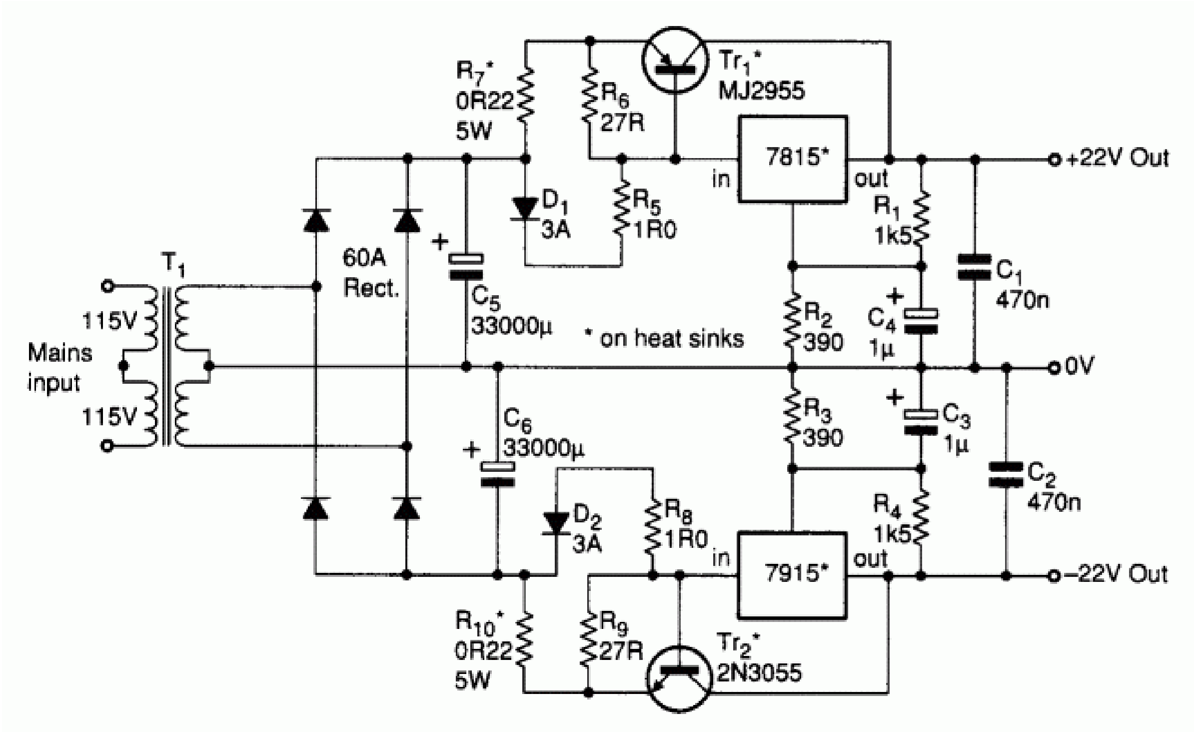


Fig. 5. Regulated power supply for the Class-A amplifier uses boosters around the three-terminal regulators. These take advantage of the regulators' current-limiting feature.

Extract from:

JLH - a lifetime in electronics

……….. …However, these were early days, and I was quite content with my Williamson mono set up - the snag lying in the 'mono' bit.

Turntable trauma
When my old friend, the Seascale carpenter, made the cabinets for our two radiograms, he fitted each with a Decca turntable and its XMS sapphire-stylus pick-up. This seemed quite a good choice in 1952, before the advent of stereophonic LPs, even though the playing weight of the spring counterbalanced head was 40g or thereabouts.

Sadly though, there was no way that the Decca system would play a 'stereo' disc without jumping up and down during the loud bits.

Replacing the turntable and pick-up would have presented little difficulty, though any new system would probably not have fitted as snugly as the original joiners installation.

The real problem was what to do about the single Williamson power amplifier. Quite apart from the cost and difficulty of buying and installing another Williamson, with its separate 450V power supply, there simply wasn't room in the cabinet to accommodate it.

So my thoughts turned to constructing two equivalent transistor power amplifiers. This solution would certainly save space, but it had to sound as good too.

I treated the requirement as if it were for a typical industrial control system. The emphasis would be on simplicity - on the premise that 'what you don't put in won't go wrong'. I designed and made up the experimental four transistor power amplifier circuit shown in Fig. 8.

Why was it n-p-n throughout?
At that time, 1965, silicon p-n-p transistors were not very good, so the design used only n-p-n power devices. The opposite was the case with germanium devices, where it was the n-p-n ones which were relatively poor in performance.

This design operated in class A, which removed any problems which might arise with class 'AB' output bias level settings.

On being tested with a 35V supply line, this amplifier worked very well. It had an output power of a little over 10W, a THD figure rather better than 0.1%, and a bandwidth of 10Hz to 100kHz,+/-0.5dB. This was very encouraging - especially when I compared its sound quality against the Williamson, and concluded that it was at least as good.

On seeing these results, I built a tidied-up stereo version as a Christmas present to myself in 1967. Some time later, I replaced the output transistors with Motorola 'epitaxial base' 2N3055s.

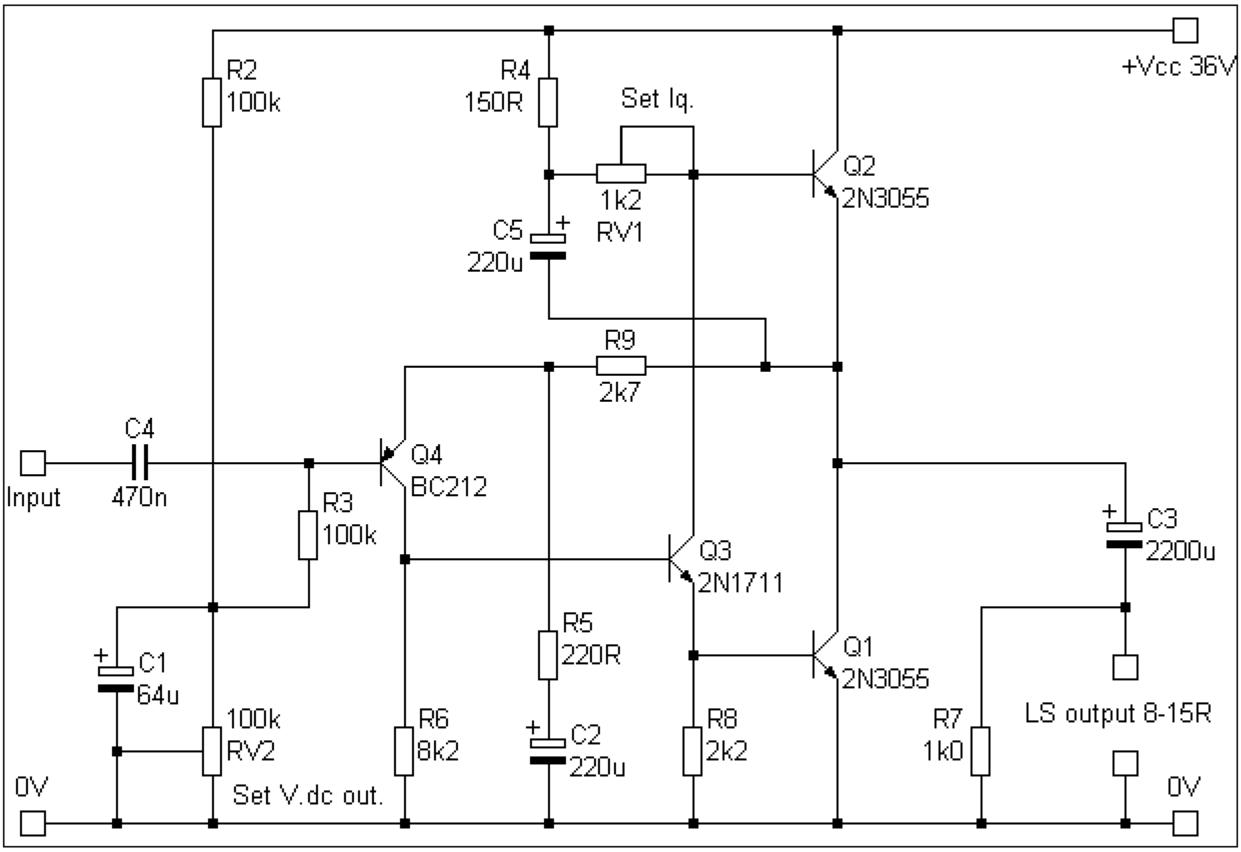
With an increase in the supply voltage to 45V, the new transistors allowed an output power of more than 15W. This was equivalent to the Williamson, though I am unconvinced that I ever needed or used more than two watts.

Unlike the Lin design, my 10W Class-A circuit did not use a push-pull pair of output devices to provide the required low output impedance. Instead it used a 'Darlington pair' connected amplifier stage, comprising Q1 and Q3, driving Q2 as an active load. Transistor Q4 gave increased loop gain for the AC and DC feedback loops.

My first audio article
It had not occurred to me to seek to publish my design. But two of my friends urged me to do so.

John Greenbank, an assistant editor on Wireless World at the time, greeted my contribution with enthusiasm. Unknown to me. the topic of Class-A versus Class-AB was one of current hi-fl debate. As a result of this publication, I suddenly found myself to be a hi-fi guru ……………

Fig. 8 The 10W class A design.



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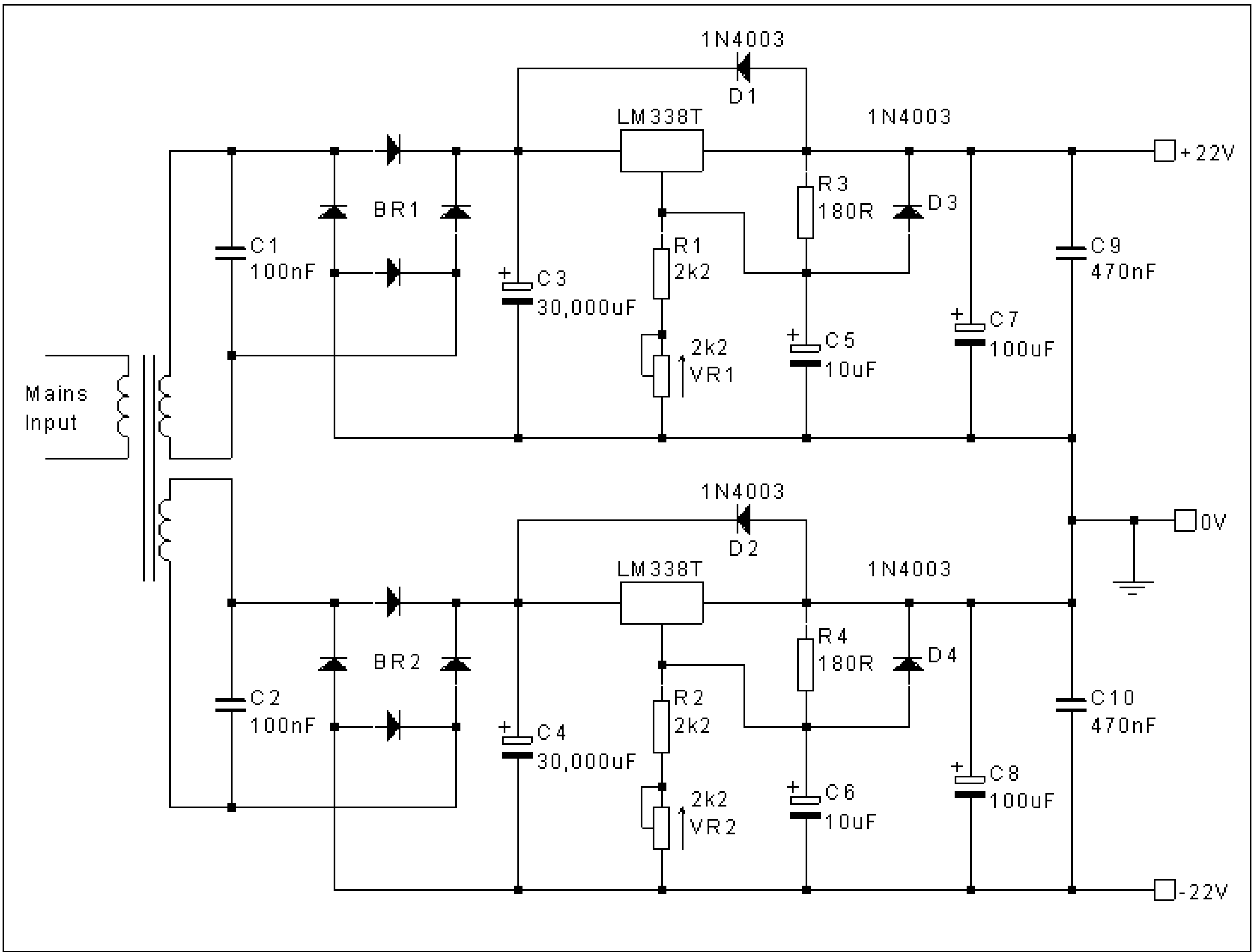
Updated Power Supply

A revised (but unpublished), regulated power supply for the 1996 JLH Class-A amplifier.



Notes

The LM338T was specified by the originator of this diagram. Due to its poor junction to case thermal resistance (4degC/W) and its low maximum junction temperature (125degC), the size of the heatsink and the maximum volt drop across the device are extremely critical for satisfactory working. The volt drop should be limited to between 2.5V and 6V if at all possible (based on a heatsink of 1degC/W for each device and an ambient of 25degC). The lower limit is set by the drop out voltage of the regulator and the higher voltage will be determined by the transformer secondary voltage, the transformer regulation, the diode losses and the fluctuations in the mains supply voltage.

The LM338K in a TO3 case (though much more expensive, at least here in the UK) has a greatly improved junction to case thermal resistance (1degC/W) and so will be more tolerant of heatsink size and will cater for a wider variation in volt drop across the device. I would therefore recommend that this version of the LM338 be used.

Bridge rectifiers BR1 and BR2 should have a minimum rating of 100V 25A (200V 35A preferred).

C3 & C4 are best made from 3 x 10,000uF capacitors in parallel.

C3 & C4 should be 50V minimum.

C5 & C6 should be 25V minimum.

C7 & C8 should be 35V minimum.

VR1 & VR2 are adjusted, under load, to give +/-22V supply rails.

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HISTORY: Page created 01/05/2001
16/05/2001 Diagram redrawn and components renumbered
14/08/2001 LM338 notes revised
01/05/2004 Reference to separate regulators for each channel
fed from a common rectifier/capacitor

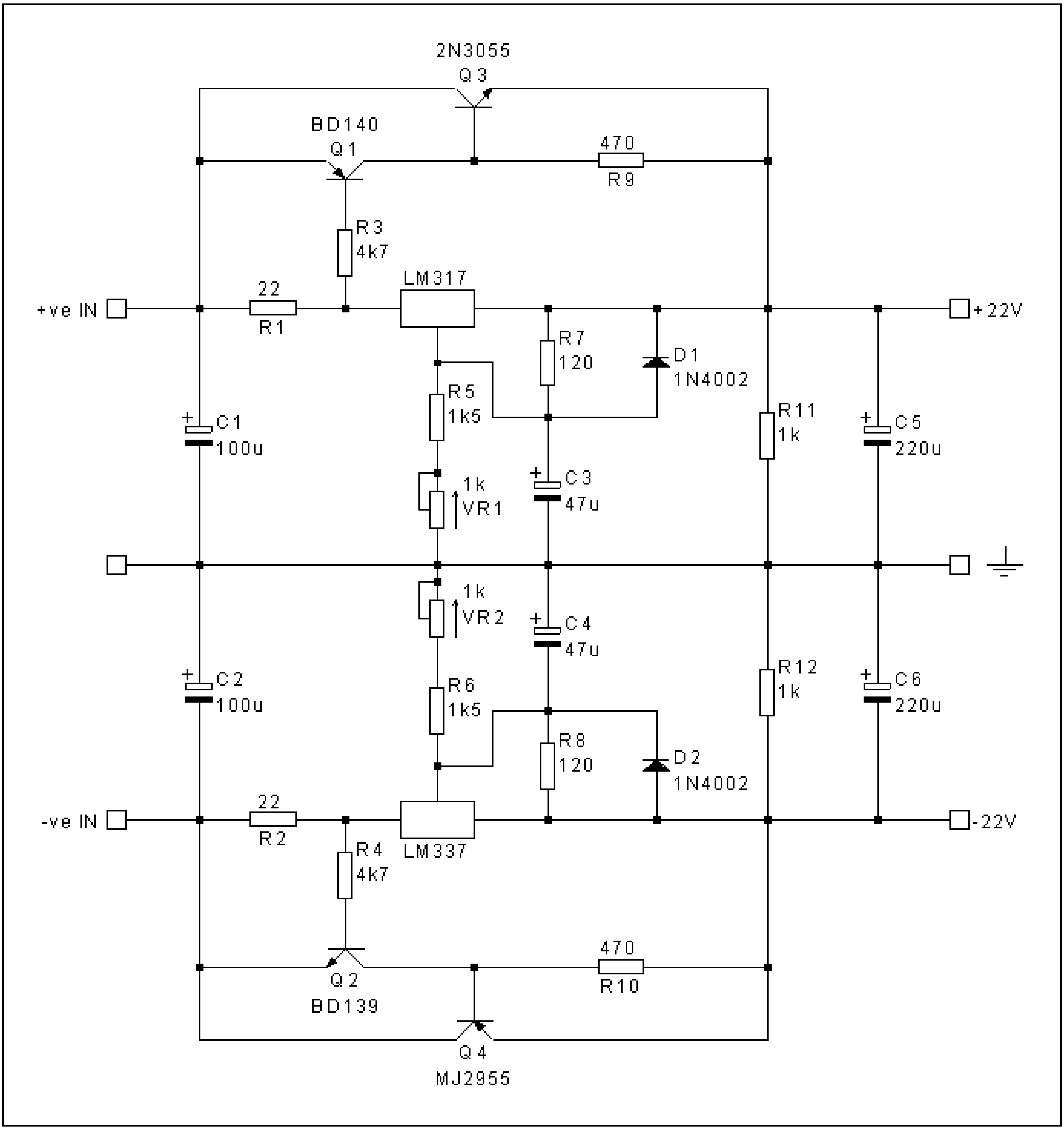
Current Boosted LM317/LM337 Regulator

The LM338K is a relatively expensive device (at least here in the UK) and may not be easy to find in certain locations. The current boosted 7815/7915 regulator circuit in the 1996 article is an alternative, but this requires power resistors in the supply line. It also has reduced regulation due to the voltage lifting arrangements necessary to provide a 22V output from a 15V regulator.

The following current boosted LM317/LM337 circuit is essentially that included on several manufacturers' datasheets. The capacitor values have been changed from those on the datasheets to reflect the use of electrolytic capacitors as opposed to tantalum devices (I'm not happy with the reliability of tantalum capacitors even though they do have some desirable characteristics) and the transistors types have been altered to ones that are more readily available. The capacitor values are not critical and may be halved or doubled to suit available components, though a minimum voltage rating of 35V should be observed.

The circuit shown is suitable for supplying a single amplifier (2A quiescent current). If two amplifiers are to be supplied from a single regulator, I recommend that the pass transistors (Q3, Q4) be duplicated using a parallel arrangement. In either case, a heatsink of between 2 and 3degC/W will be required for each transistor. The exact size of heatsink required should be determined whilst taking into account individual circumstances.

VR1 & VR2 should be adjusted, under load, to give +/-22V supply rails.



The Capacitance Multiplier

The power supply for the original 1969 JLH design included a form of capacitance multiplier to reduce the amount of voltage ripple on the supply rail. The capacitance multiplier circuit has been developed further, by Rod Elliott of Elliot Sound Products, and the results published as Project 15 at the ESP Audio Pages. The modified circuit is suitable for both the original 1969 JLH amplifier (using only the positive half of the circuit) and the 1996 update. The design considerations for the capacitance multiplier, its benefits, and a comparison with voltage regulators are included in the Project article. I do not propose to repeat the information here, but have included a copy of the final circuit schematic for information (Figure 1.).

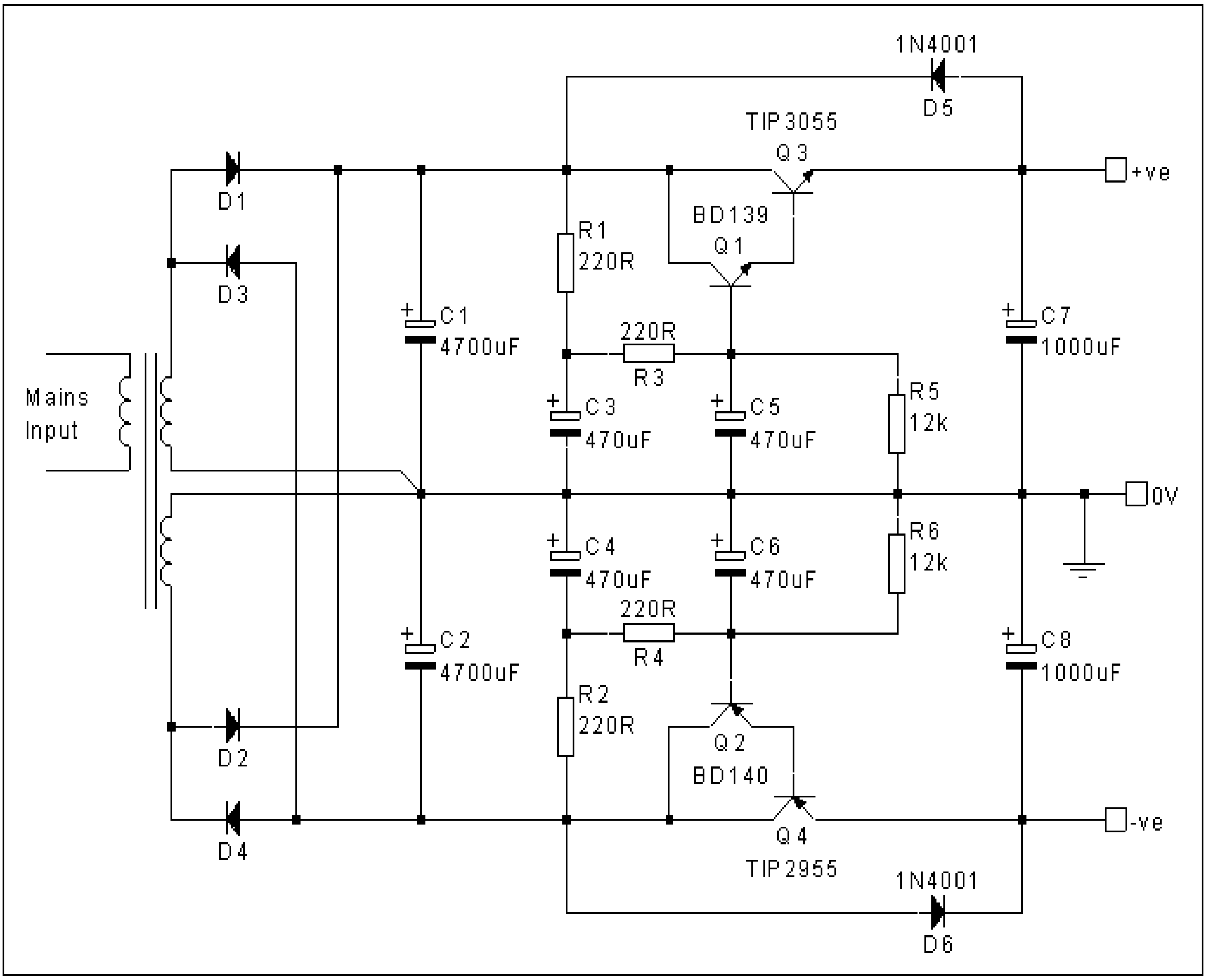


Figure 1. The final circuit (courtesy of Rod Elliott and The ESP Audio Pages)

If desired, this circuit can be modified slightly to use the same type of pass transistor in both the positive and negative supply rails. This will allow the same transistors to be used as those in the output stage of the amplifier, which could result in some cost saving by buying a quantity of the same type of device. A greater advantage results from the opportunity to match the gain of the amplifier output transistors. Matching the gain gives the minimum harmonic distortion for this amplifier and, after selecting suitable ‘pairs’ for the amplifiers, the ‘rejected’ transistors can be used in the capacitance multiplier(s). The modified circuit is shown in Figure 2. VR1 and VR2 are adjusted to give equal (but opposite) voltages on the supply rails and to give the required volt drop across the pass transistors of approximately 3V.

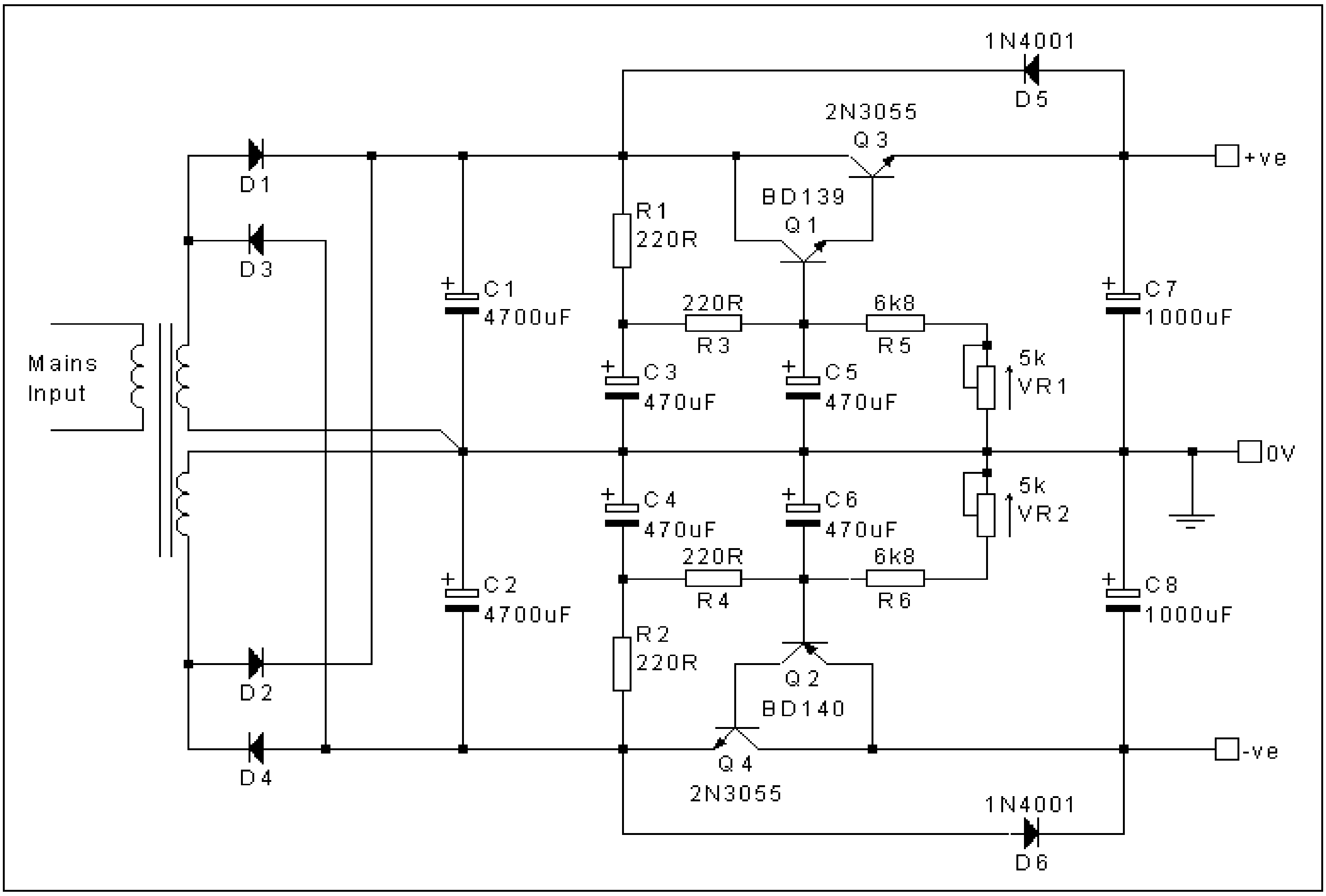


Figure 2. The revised circuit which allows both pass transistors to be of the same type.

The required size for the main smoothing capacitors (C1 & C2) depends upon the load current. I suggest that the minimum capacitance should be as shown the following table. The preferred value for the capacitors is also given (this is approximately 1.5 times the minimum).

Quiescent Current (A)	Peak Load Current (A)	Minimum Capacitance (uF)	Preferred Capacitance (uF)
1	1.5	4,700	6,800
2	3	6,800	10,000
3	4.5	8,200	12,000
4	6	10,000	15,000

Please note that these figures have been revised since the original publication of this page and reflect more recent (and accurate) simulations – I have changed my simulator program and models to ones that give more realistic results.

A simple capacitance multiplier circuit is shown in Figure 3. This gives the easiest possible physical construction (i.e. the fewest components). The performance of this simple circuit in terms of ripple voltage reduction is not as good as the previous circuits, but it is still more than adequate to reduce any hum to inaudible levels (unless you have extremely sensitive speakers).

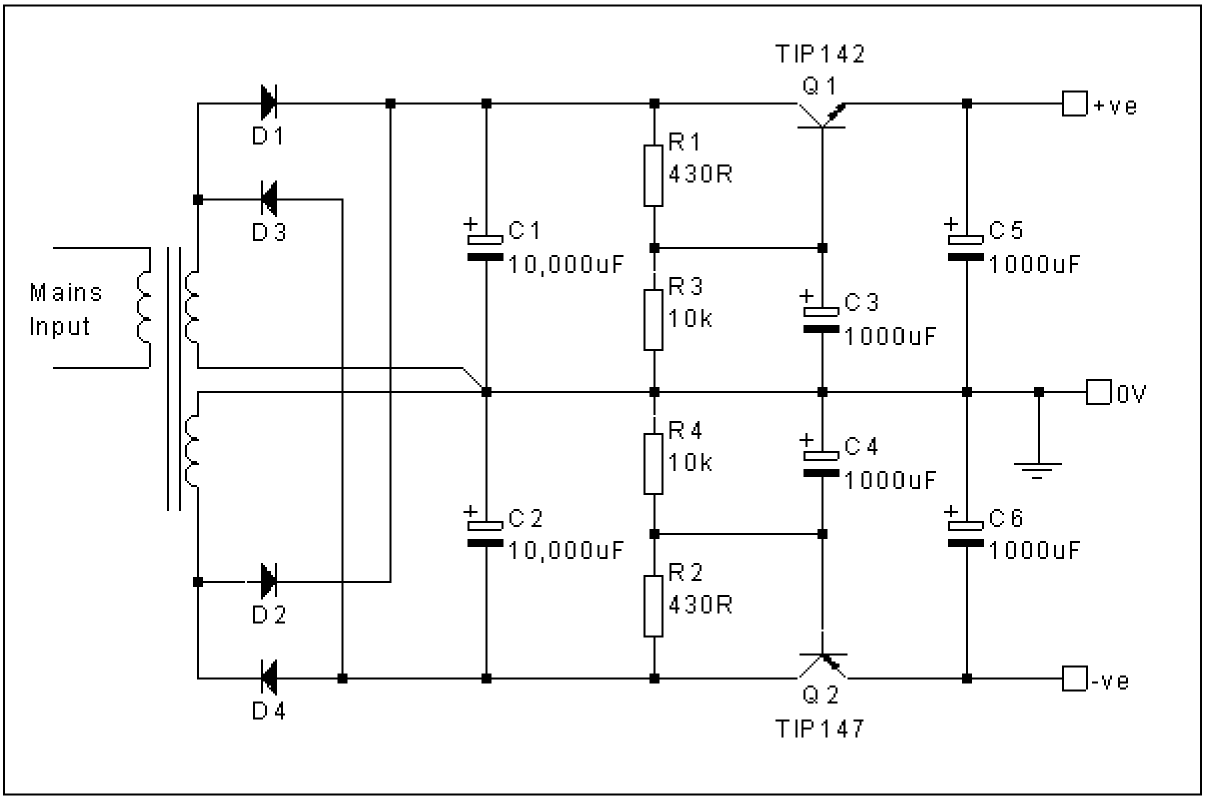


Figure 3. A simple capacitance multiplier

Again, the required size for the main smoothing capacitors (C1 & C2) depends upon the load current. It is generally greater than that for the previous circuits. The following table gives the minimum and preferred values.

Quiescent Current (A)	Peak Load Current (A)	Minimum Capacitance (uF)	Preferred Capacitance (uF)
1	1.5	4,700	6,800
2	3	6,800	10,000
3	4.5	10,000	15,000
4	6	15,000	22,000

In the two previous tables, the 1A and 2A figures are relevant to a single 1969 or 1996 version of the JLH amplifier. The 3A and 4A figures are appropriate for a pair of each amplifier version operating from a single power supply.

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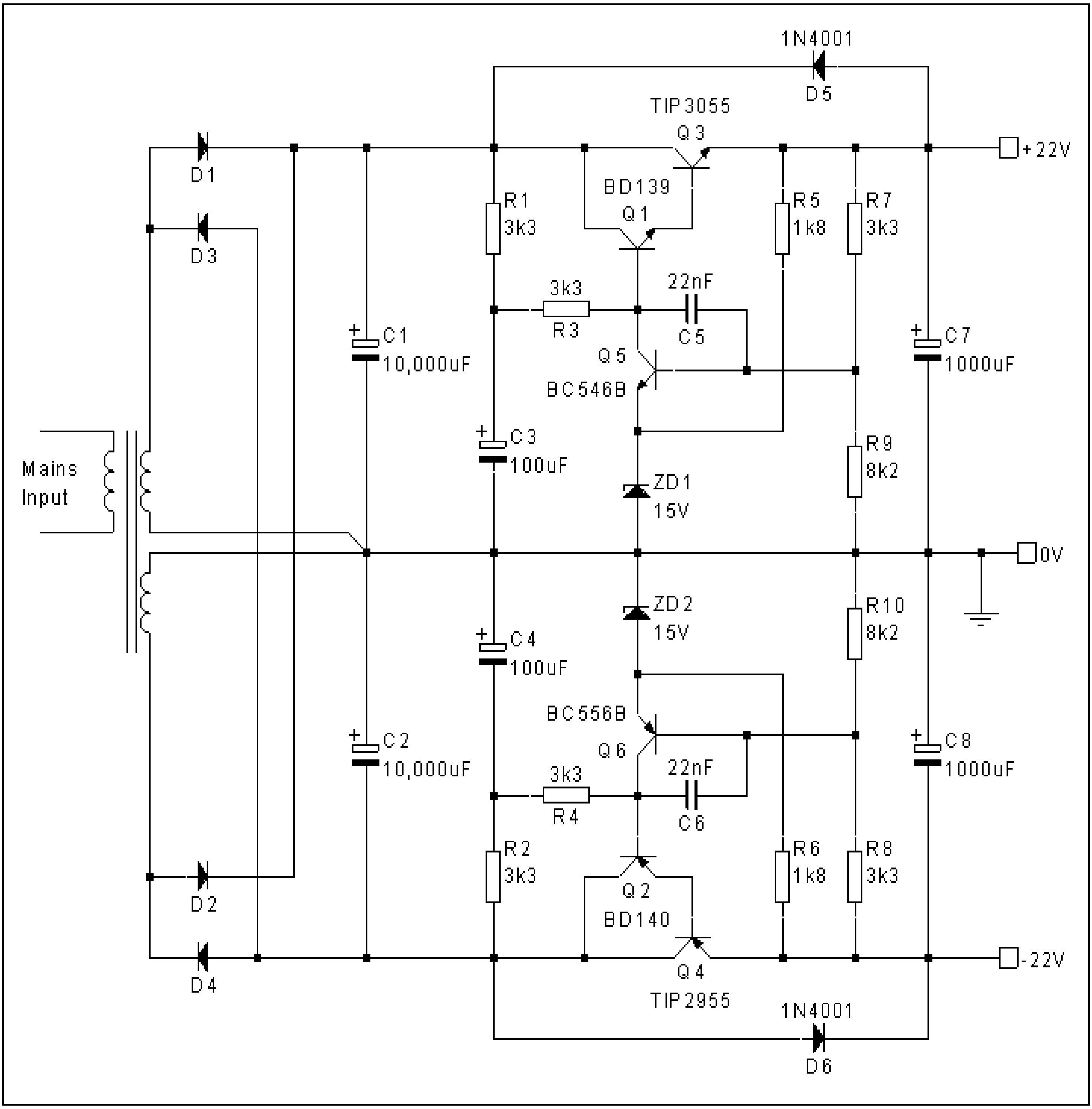
HISTORY: Page created 01/05/2001
16/05/2001 Diagrams amended to correct polarity of D6. Simple capacitance multiplier circuit added
17/05/2001 Recommended capacitor sizes updated. Minor text changes

A Simple Voltage Regulator

The following diagram is provided for anyone who would like to include a voltage regulator but who does not want to use one of the ic versions (as in the updated, regulated supply for the 1996 design). This is a very basic regulator circuit, without any additional features such as foldback current limiting, and does not have the same performance as an ic regulator.



Notes

The TIP2955/TIP3055 transistors should be satisfactory for a power supply feeding a single amplifier. If two amplifiers are to be fed from a single power supply, these transistors should be changed to higher power devices such as the MJ2955/2N3055. Adequate heat-sinking must be provided for whichever transistors are used.

R9 and R10 can be replaced with a 10k preset potentiometer (or a 5k potentiometer in series with a 5k6 fixed resistor) to provide adjustment of the output voltage.

As with the capacitance multiplier circuit, Q2 and Q4 can be changed to a complimentary feedback pair arrangement, if required, to allow the use 2N3055s as the pass device in both halves of the supply (see the capacitance multiplier page for details). If this done, R9 and R10 must be made variable to allow the supply rails to be set to equal (but opposite) voltages.

Zener diodes of a different voltage rating can be used for ZD1 and ZD2, but the value of R9 and R10 will need to be adjusted to maintain the +/-22V output.

For different output voltages or a different Zener diode voltage, the output voltage can be calculated from the following equations:

$+V_{OUT} = ((R7 + R9) / R9) * (V_Z + 0.6)$

$–V_{OUT} = ((R8 + R10) / R10) * (V_Z + 0.6)$

where $+V_{OUT}$ and $–V_{OUT}$ are the required supply rail voltages and V_Z is the Zener voltage.

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HISTORY: Page created 13/05/2001
16/05/2001 Diagram amended to correct polarity of D6
Resistor numbering corrected in voltage equation
17/05/2001 Minor text changes. Second voltage equation added

JLH Class-A Update

I had originally intended that this page would be a step-by-step record of the modifications carried out during the past year by one constructor – Tim Andrew. However, recent ill health has meant that I have been unable to spend much time sitting at my pc so, rather than incur yet more delay in publishing the results, I have decided to write a short summary instead. I am very pleased that Tim has taken the time to supplement this with his own comments. At the end of the page is a brief update on the higher power 'JLH for ESL' circuit.

Tim is a professional musician (a classical concert pianist) and so I trust his subjective judgement when it comes to assessing the accuracy and realism of sound reproduction. Before Tim first contacted me, he had built a kit version of the 1996 design, which he had subsequently upgraded with higher quality components. Though Tim was happy with the results, he was keen to see if further improvements could be made to the sound quality and I was pleased to be able to suggest various circuit modifications, the majority of which subsequently proved to be very worthwhile. Each of the modifications was carried out separately so that the results could be evaluated on an individual basis.

Rather than show schematics for each stage, I will start off with the penultimate circuit and include some appropriate comments.

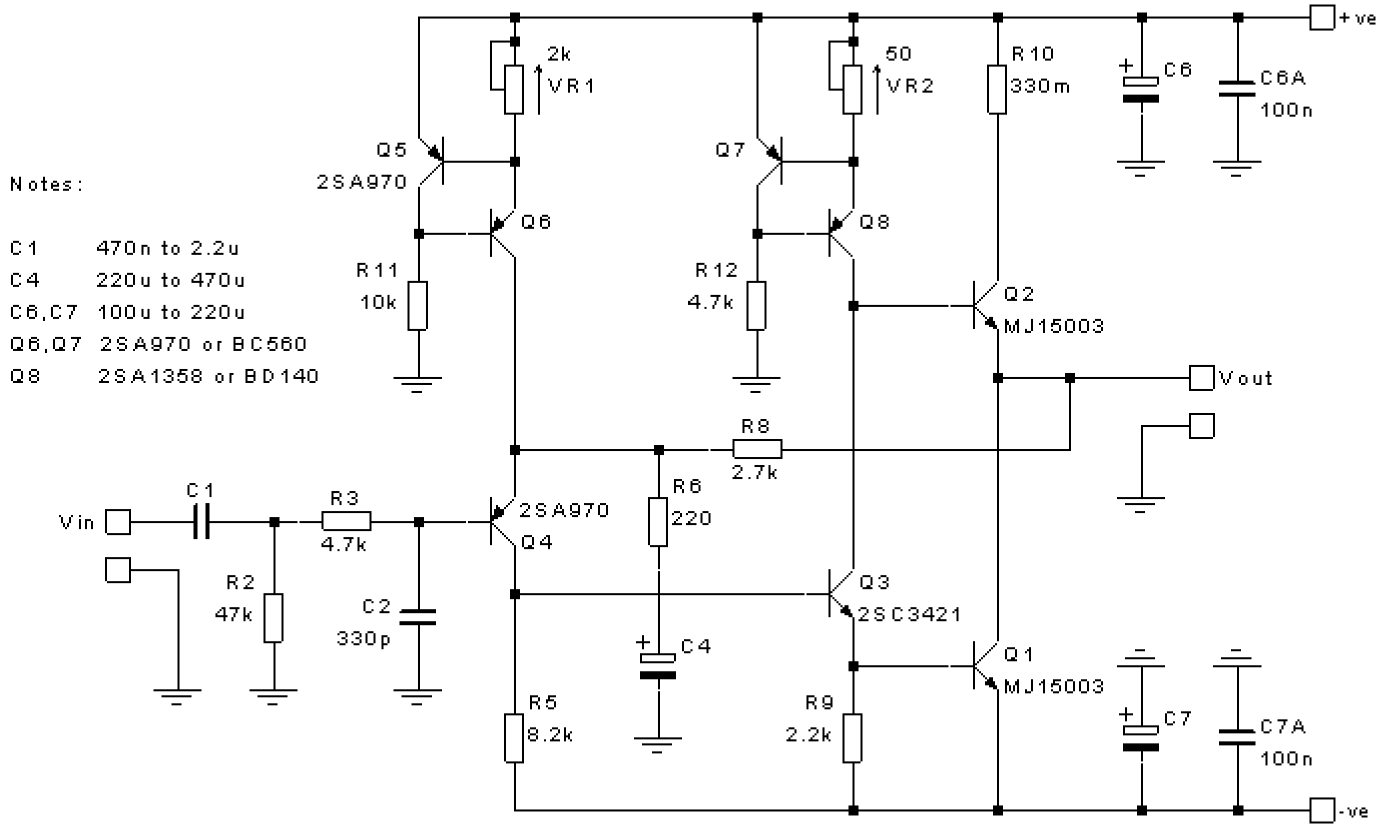


Fig 1 – The Penultimate Circuit

Transistor substitutions

One of the first modifications was to try alternative output transistors. The MJL3281A gave an audible indication of oscillation and was quickly rejected. The MJ21194 sounded significantly better than the 2N3055 but, in Tim's layout, introduced a low-level hum. The MJ15003 gave a similar improvement to the MJ21194, but without the hum, and so was retained for future use. At a later stage, the BC212 and 2N1711 (Q4 and Q3) were replaced with the 2SA970 and 2SC3421.

Output dc offset control

The standard dc offset control circuitry (7815 and associated components) was replaced with a two transistor constant current source (Q5/Q6). I had various reasons for suggesting this change. Firstly, three terminal regulators are not renown for their quietness and so it did not seem like a good idea to inject the noisy output from one directly into the feedback loop. Also, I had received reports that certain 7815s oscillated due to the low current conditions under which they were being operated.

However, one of the main benefits of the ccs is that the output dc offset variation as the amp warms up is greatly reduced. This is because the temperature coefficient of the ccs acts in the opposite direction to that of the input transistor (Q4) and negates the effect of temperature changes in Q4 (assuming that the temperature of Q5 follows that of Q4). This cancellation of temperature coefficient effects can be put to further good use as will be seen later.

Quiescent current control

I first suggested that Tim try the 1969 bootstrap Iq control circuit, partly because the simulated distortion figures were half those for the 1996 version but mainly because I wanted to know how the two methods of Iq control compared in the same amplifier. I had received reports that the 1969 circuit (modified to dual supply rails) sounded better than the 1996 version, but I could not be sure that there were no other variables involved. As it turned out, the bootstrap circuit was a retrograde step and Tim immediately reverted to the original 1996 arrangement.

I still had some nagging doubts about the 1996 Iq control circuit and so I suggested introducing another constant current source (Q7/Q8). As with the bootstrap circuit, the simulated distortion figures were still half those for the 1996 version but with the added advantage that the distortion did not increase at low frequencies due to a reduction in capacitor effectiveness. A further advantage was an increase in amplifier efficiency (or maximum output). The maximum output voltage swing with the ccs is greater than that for the standard 1996 circuit and the maximum output current increases from around 1.35 to about 1.5 times the quiescent current.

When carrying out this modification, Tim reused the existing MJE371 for Q8. R10 has been retained to provide an easy means of measuring the quiescent current. To my relief, Tim found the second ccs to be worthwhile improvement.

Power supply

Whilst making the other alterations, Tim also took the opportunity to upgrade his power supply, initially by fitting larger bridge rectifiers and snubber capacitors and then by replacing the LM338s with ‘follower’ type discrete regulators, in line with my desire to remove unnecessary feedback loops from the overall circuit. The ‘follower’ regulators, basically a capacitance multiplier circuit with a fixed voltage reference (derived from a resistor fed by a ccs), gave a small improvement. A much greater improvement was obtained when separate regulators were provided for each amplifier, whilst retaining a common transformer, rectifier bridges and reservoir capacitors.

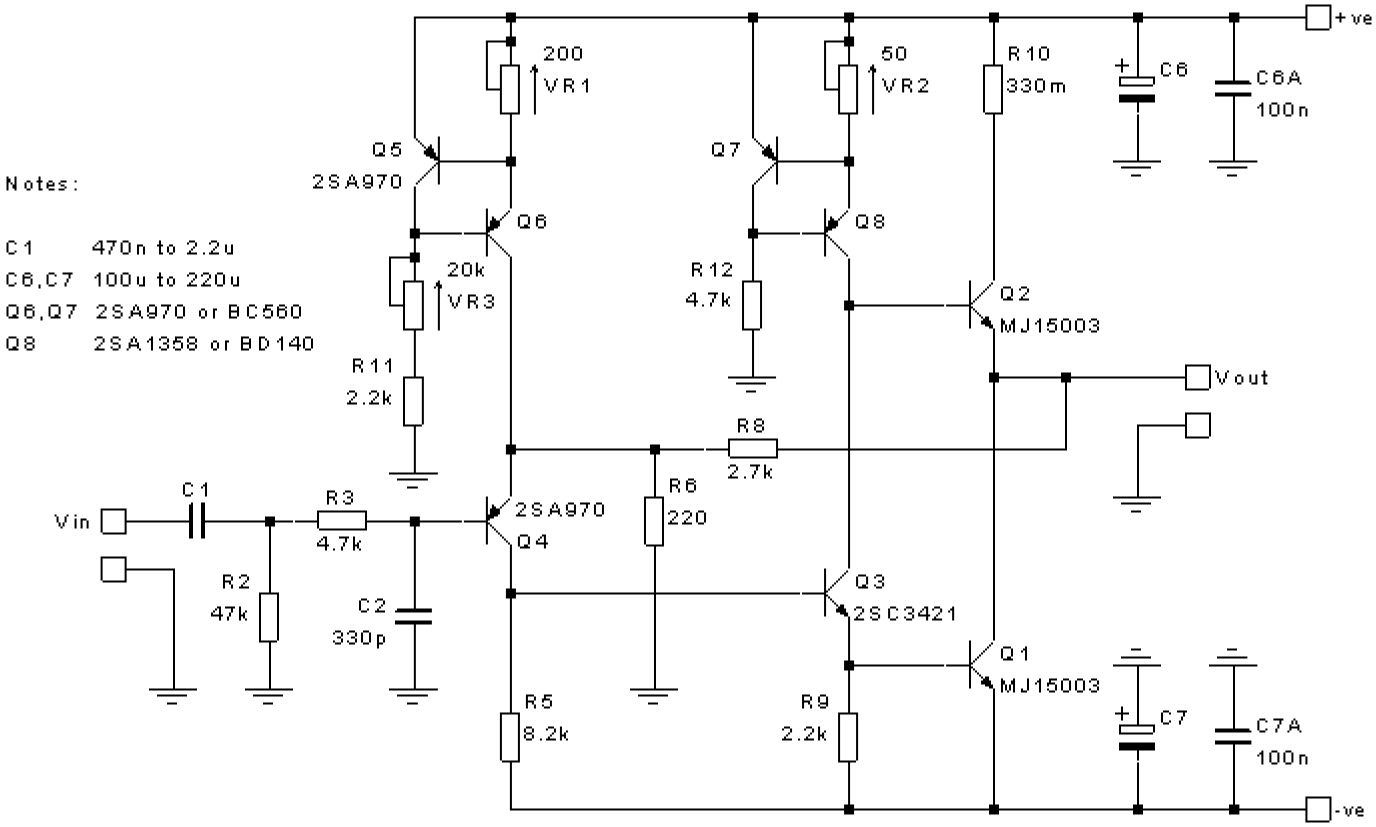


Fig 2 – The Final Circuit

Removal of the feedback capacitor

I had received emails from a couple of constructors reporting on the beneficial effects of removing the feedback capacitor (C4). I passed these comments on to Tim and he decided to try this modification for himself.

This modification should be treated with caution. I would not recommend trying it unless the dc offset ccs (Q5/Q6) modification has been done first because otherwise the output dc offset variation during the warm-up period is likely to be in the order of several hundred millivolts. In Tim’s case, with the dc offset ccs fitted, the output dc offset variation with the feedback capacitor removed was only slightly higher than that which he had previously with the standard 1996 circuit.

I believed that the offset variation could be reduced further by utilising the temperature coefficient of the Q5/Q6 ccs. I therefore suggested that R11 be made adjustable so that the temperature rise of Q5 could be varied. In this way, the output dc offset variation due to temperature changes in all stages of the amplifier could be compensated for, though this requires a lengthy, iterative process. With the amp at its normal operating temperature, the offset is adjusted to near zero using VR1. The offset when the amp is cold is then measured. VR3 is adjusted slightly, the amp is allowed to warm up and the offset is re-zeroed using VR1. The offset is then rechecked when the amp is cold and the process repeated until the minimum offset variation has been obtained. Tim has been able to achieve an output dc offset variation between switch-on and normal operating temperature of less than 50mV.

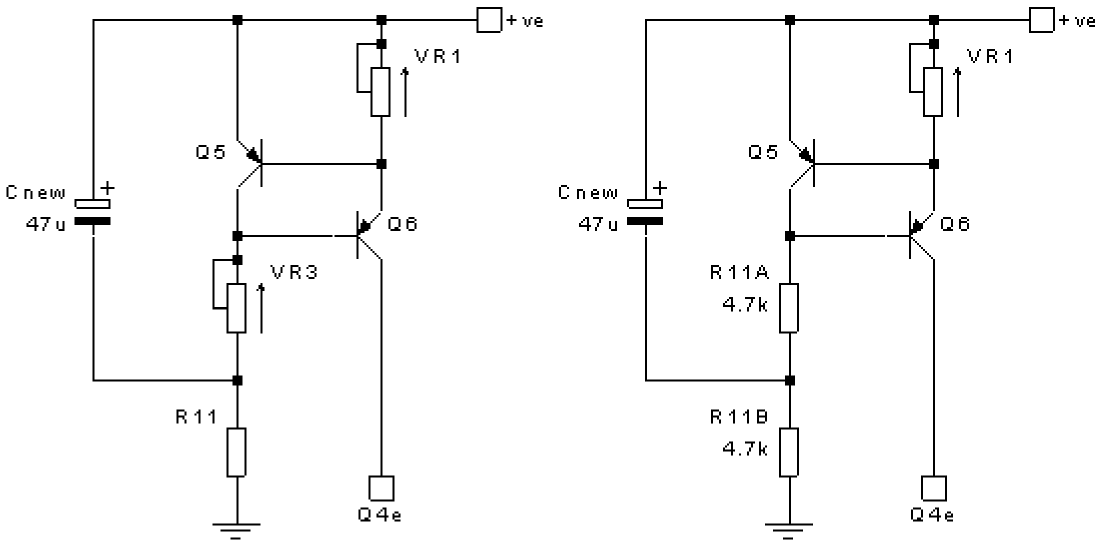
15/03/2003 Addendum

It has been brought to my attention (thanks Mietek and Rudy) that removing the feedback capacitor increases the hum level at the amplifier output, which is particularly noticeable with high sensitivity speakers and if a simple rectifier/capacitor power supply is used. I had not anticipated this, but some quick simulations soon indicated that removal of the feedback capacitor reduces the PSRR of the amp by a factor of about 3, causing any supply rail ripple to become more audible.

Fortunately, the cure for this problem is relatively simple. The PSRR of the input stage ccs can be improved by the addition of a single capacitor, connected between the junction of VR3/R11 (Fig 2) and the +ve supply rail. Doug Self's 'Audio Power Amplifier Design Handbook' indicates that this modification will improve the PSRR of the ccs by about 10dB. A capacitor value of 47uF will suffice, but higher values (within reason) can be used.

The higher power ('JLH for ESL') circuit can be similarly modified by splitting R11 (Fig 3) into two 4k7 resistors in series and connecting the capacitor from the mid-point of these resistors to the +ve supply rail.

This modification can also be carried out even if the feedback capacitor is not removed, and will give an improvement in PSRR with the corresponding reduction in hum.



Revised input stage ccs

17/08/2003 Addendum

Several constructors have found that adding the 47uF capacitor to the input stage ccs after having removed the dc blocking capacitor from the feedback network has caused the ccs to become unstable. This has manifest itself by relatively large output dc offset variations when taking voltage readings around the input circuit or when a hand is moved near to the ccs components.

In Tim's case, a successful solution to this problem has been to replace Q5 and Q6 with 'slower' transistors. The MPSA56 appears to work well in the ccs. Alternatively, the 47uF capacitor could be removed and the PSRR of the ccs improved by omitting VR3 and replacing R11 with a 1mA constant current diode (or an FET wired as a ccs to give a similar current).

Adding base resistors (100R to 1k) to Q5 and Q6 and/or a 1k resistor between Q6c and Q4e should also help to improve stability.

Tim's comments on the modifications (Updated 17/08/2003)

A few years ago I built the 1996 version JLH Class-A amplifier. Constructors of this amplifier have commented about its smooth sound, with many favourable comments and comparisons against valve designs and a few not so favourable comments with regard to its limited power output. In its standard 1996 form, which I built from a kit using cheap components, my first impressions of its sound were of smoothness coupled with a relaxed liquid musical flow which I found far preferable to anything else which I had previously heard. In the context of my system with speaker efficiency somewhere around 87dB/W and with volume set correctly such as is appropriate for the perspective as recorded, or in other words "at a realistic level", its limited power output has never been a problem. The amplifier and its power supply have since been subject to extensive component substitutions and substantial circuit modifications.

As this section is about my impressions of the modifications that have been made to the circuit, a brief word on what I consider to be an "improvement" might be in order. I want to hear, with ease, the ambient signature of the recording venue, with a distinct impression of the space between its walls. Also, I want to notice, for example, the sound of the felt hammer of a piano hit the string, followed not only by the sound of the string vibrating but also the more subtle reflected and attenuated sounds of the hammer and its mechanism as these reverberate between the walls of the recording venue. This is sometimes more noticeable in larger venues where the reflected sound arrives later, albeit weaker. Those delicate piano harmonics must be reproduced with the greatest accuracy, enabling subtle shadings of timbre to be noticed, again with ease. As a pianist, I want to hear the "pitch" of the note as it decays through to its quietest moment as acutely as possible, but I want no hint of hardness or roughness. With orchestral strings for example, where there are many instruments playing together, I don't want to hear one homogeneous group, and I want transparency, not brightness.

Professionally, I have a very close affinity with the piano. A difficult instrument to reproduce, it is perhaps more revealing of faults in the reproduction chain than can be the case with other instruments although the human voice is also very useful, for obvious reasons. It is my view that any modification that produces a more realistic rendition of the complex sound of this instrument, and the very subtle structure of its over-tones, will also represent an improvement in the accuracy of the amplifier overall. This has been the case during all my listening trials. It is worth mentioning that any modification which leads to an apparent decrease, for example in the level of the treble, will not necessarily be deemed to be an improvement, even if the new treble level is a welcome one, unless it is accompanied by an improvement elsewhere, improved detail or portrayal of nuance for example. From this, you will gather that I am not in the habit of 'voicing' the system, adjusting one thing to correct for another, but that I prefer to address the transparency of the system as a whole, with the aim of neutrality. Only then will I look at altering the balance, perhaps with a slight adjustment to the treble. It is through this approach (transparency first, followed by tonal balance) that I am now able to enjoy the vast majority of recordings in my collection, previously I had found many of these to be deficient in one way or another. Almost without exception, each modification has improved "difficult" recordings, whilst further improving others, often revealing a warmth and atmosphere, the previous lack of which had been wrongly attributed to the recording.

Though considerable time has been expended on both the amplifier and its power supply, I find it sobering to say the least that improvements made to power supply, specifically to the method of its delivery into various parts of the amplifier circuit have been so rewarding. The following is a list of the modifications that, with considerable help from Geoff, I have been able to carry out on the 1996 version of the JLH. Also included are my opinions of the results of these. Each substitution has been carried out individually, this has enabled subsequent and hopefully accurate (but not always positive!) evaluation. !

The Amplifier

Input capacitor.

The cheap polycarbonate(?) 1uF input capacitor was replaced with a 470nF Mcap "Audiophile" polypropylene type. This led to an improvement in both bass firmness and in detail, treble sounded less bright. Later, I replaced the Mcap 470nF with Audio Note paper-in-oil 470nF. This sounds very different, smooth, warm and open with much more textural detail and firmness in the bass. There is some loss of focus when compared with the better plastic types and the positioning of instruments within the stage is not as precise as it could be, however none of the plastic types I have tried has approached the naturalness and openness of the paper-in-oil, particularly in the treble, and any shortcomings are easily forgiven in light of considerable improvements elsewhere. This simple modification has since proved to be one of the most effective. I have also tried a polystyrene type (333nF) which sounds more detailed and focussed than anything else tried previously, though there is a tendency to sound a little "squeaky" on occasions (placing a small paper-in-oil capacitor across it improves this considerably), nevertheless I prefer this to most polypropylene types, many of which sound hard and slightly blurred to me.

Resistors.

All standard grade metal film resistors in both critical and semi-critical parts of the circuit were replaced with tantalum film types.
Improved smoothness and texture, with a more fluid sound. A slight "mumbling" quality has been removed.

Output transistors.

The 2N3055s were replaced with MJ21194. In comparison with these the 2N3055s sound grey and rather diffused with less sense of authority, less detail and a more prominent treble quality. In contrast, the MJ21194s have a noticeably firmer sound with more ambience in the treble and greater detail. More natural generally. Reluctantly, they were removed from the circuit due to a faint hum which was not present with the 2N3055s.
Wanting to try something else, and now with the strong impression that the 2N3055s were less than ideal, I tried some MJ15003s.
This time, a substantial improvement over the 2N3055s. The MJ15003's bass is both tauter and more authoritative, with cleaner treble and greater textural detail.

DC offset control.

Replace 7815 with constant current source.
Result...Cleaner, smoother and weightier, with what can only be described as an organic flow. It was obviously all there before, but I suppose it was masked somewhat by the noise of the regulator. The volume can be increased further without sounding "loud". A substantial improvement in all respects.

Iq control circuit.

The Iq control circuit was replaced with a bootstrap circuit (using an Elna "Silmic"). Less clarity was the result, with less tonal variety and focus, sounding more shut-in. The bootstrap simply doesn't sound as detailed. I assume this is due to the presence of the bootstrap capacitor connected to the signal path. Perhaps a Black Gate might improve things, but I suspect not enough to equal the MJE371 circuit which is more transparent, open, dynamic and uncoloured, the female voice sounds less "female" with the bootstrap circuit. It strengthens my theory that those who prefer the earlier version of the JLH do so because of the absence of the 7815 in the earlier circuit. I would go further and say that due to the absence of both a bootstrap capacitor, and an output capacitor, and with the ccs in place of the 7815, they might well prefer the 1996 version, all other things being equal. My original Iq control circuit was very quickly re-instated!

It was not long until the original Iq control circuit was removed again, this time replaced with a constant current source and with better results this time. The initial reaction is to think that the treble detail and "air" have been diminished with a reduction of transparency. On prolonged listening things are rather different. There is actually more detail coming across, coupled with a growing sense of "rightness". Sounds are presented in a more natural light, gone is the spotlight effect with its admittedly pleasant but artificial treble detail. String harmonics are more balanced and proportioned with a sense that they now belong to the fundamental, part of the whole. The gaps between rapid piano notes are often missed by amplifiers, the JLH reproduces these well and they are even clearer now than before. Familiar recordings of woodwind and brass instruments sound remarkably smooth and natural. Differences in scale between smaller chamber music recordings and larger scale works are now more clearly conveyed. It is interesting to compare the sound of the Iq ccs circuit with that of the bootstrap which shared many of the attributes of the ccs but had a lumpy and coloured, slightly congested characteristic which I found unpleasant. Returning to the standard 1996 Iq circuit the next day was quite a relief, this time I have no plans return. I would miss the qualities that the Iq ccs circuit has brought to the amplifier. Final thought........Recommended for those who want to sit down for an evening of good music and a fine wine.

Feedback capacitor.

The 470uF Oscon (previously a very similar sounding 220uF Silmic) feedback capacitor was replaced with link (needing a small change in value to the DC offset ccs preset). The result of this change was a more open and natural treble with an increased sense of fluidity, depth and ease. Hot/cold offset variation are much greater without the feedback capacitor, in my circuit a variation of 150mV was observed (with the feedback capacitor it was around 65mV), this was reduced by controlling the current through the ccs in an effort to adjust the temperature compensation, but on a recent re-build of the circuit this arrangement proved ineffective and was subsequently removed.

Driver transistor (2N1711).

This was replaced with a 2SC3421. As with the other transistor substitutions I have made in the JLH, the actual pitch of a note is more easily heard with the 2SC3421s. The same characteristics introduced by the Iq ccs circuit are still there but each single note now conveys more "meaning", more clearly defined in time. Timing, of course, is a musician’s greatest asset! The Iq ccs circuit introduced a smoother, rounder sound with a somewhat darker hue, the extra transparency and openness brought about by the 2SC3421s has lifted that slight darkness away whilst apparently retaining the smoothness and naturalness of the Iq ccs.

Input transistor.

The BC212 was replaced with 2SA970 with similar improvements to those noticed with the 2SC3421.

The Power supply.

Rectifier diodes.

Having tried snubber capacitors across the original "standard" diodes with no noticeable improvement, the originals (and snubbers) were replaced with schottky types. This seemed to be beneficial with more smoothness and an improved "woody" quality with woodwind.

Regulators.

The LM338K regulator circuit was replaced with a capacitance multiplier. The bass now conveys more authority and the amplifier sounds a little warmer, also with more detail.

Dual regulators.

The single capacitance multiplier was replaced with a new (adapted) dual version allowing separate regulation for each channel. This warrants a detailed write-up so I shall list my observations in the order in which I noticed them and in descending order of their magnitude.

It is only now that I have heard the new dual power supply, that I can identify the sonic effects of the single supply. For the first, and most important observation, I shall use a single piano note as an illustration. With the single supply, when the note is struck there is an initial transient 'bump' as the hammer hits the string, followed by the decay, which starts after the initial 'bump' has subsided. With the dual supply, this initial transient is less 'loud' (better controlled?) and it carries more weight and meaning, this is followed by the decay which not only conveys better pitch, leading to more emotion and tunefulness, but the decay starts sooner, its first moments not masked by the apparently exaggerated impact of the hammer blow introduced by the single supply. Also, due to the increased definition, the note seems to decay more slowly, incidentally this is one of the more significant differences between a small grand piano, and a large 'concert' grand where, due to the increased string length of the larger instrument, its sustaining power is much greater. A single note can therefore be followed more easily from start to finish. The tonal signature and real colour of all instruments are now better conveyed.

There is also a significant improvement in the quality of the treble where there is greater transparency. For most of the time, it is less obvious than before, and smoother, but little details previously almost un-noticed are conveyed more clearly and with improved texture. This treble improvement was unexpected and is a constant pleasure!

The third improvement I have noticed is an improvement in the positioning of individual instruments. The perceived stage width is not obviously any wider than before, although I couldn't fault it before, on a good recording the stage width was almost limitless, on a bad recording it had definite limits. This hasn't changed, what has improved is the positioning of instruments within the limitations of the stage width imposed by the recording, with instruments on the edge of the stage more clearly conveyed in space with a better "floating" feel to the acoustic coupled with a more acute sense of the venue.

Filter capacitors.
Having previously bypassed the standard grade electrolytics with Elna "Silmic" 100uF with little, if any improvement, this time the original capacitors (30,000uF per rail) were replaced entirely with "Silmics" (18,000uF per rail). A superb improvement in definition. The scale of which came as quite a surprise.

Conclusion.
I consider the JLH in its present form, to be a very special amplifier. Its ability to portray the acute sense of emotion and excitement contained in a fine performance, through its accuracy and with such grace, coupled with its ability to scale music's dynamic heights so convincingly, is rare. My most sincere thanks to Geoff who, through spending so much time helping others like me, has so far not had time to carry out these modifications for himself *.

* *Unfortunately not the only reason - Geoff*

Higher power circuit

The ‘JLH for ESL’ circuit, which can be used with conventional speakers as well as electrostatics, already has a ccs for dc offset adjustment but it would benefit from the other modifications outlined above. In particular, the use of a ccs for quiescent current adjustment obviates the need for a high power preset, which can sometimes be hard to find.

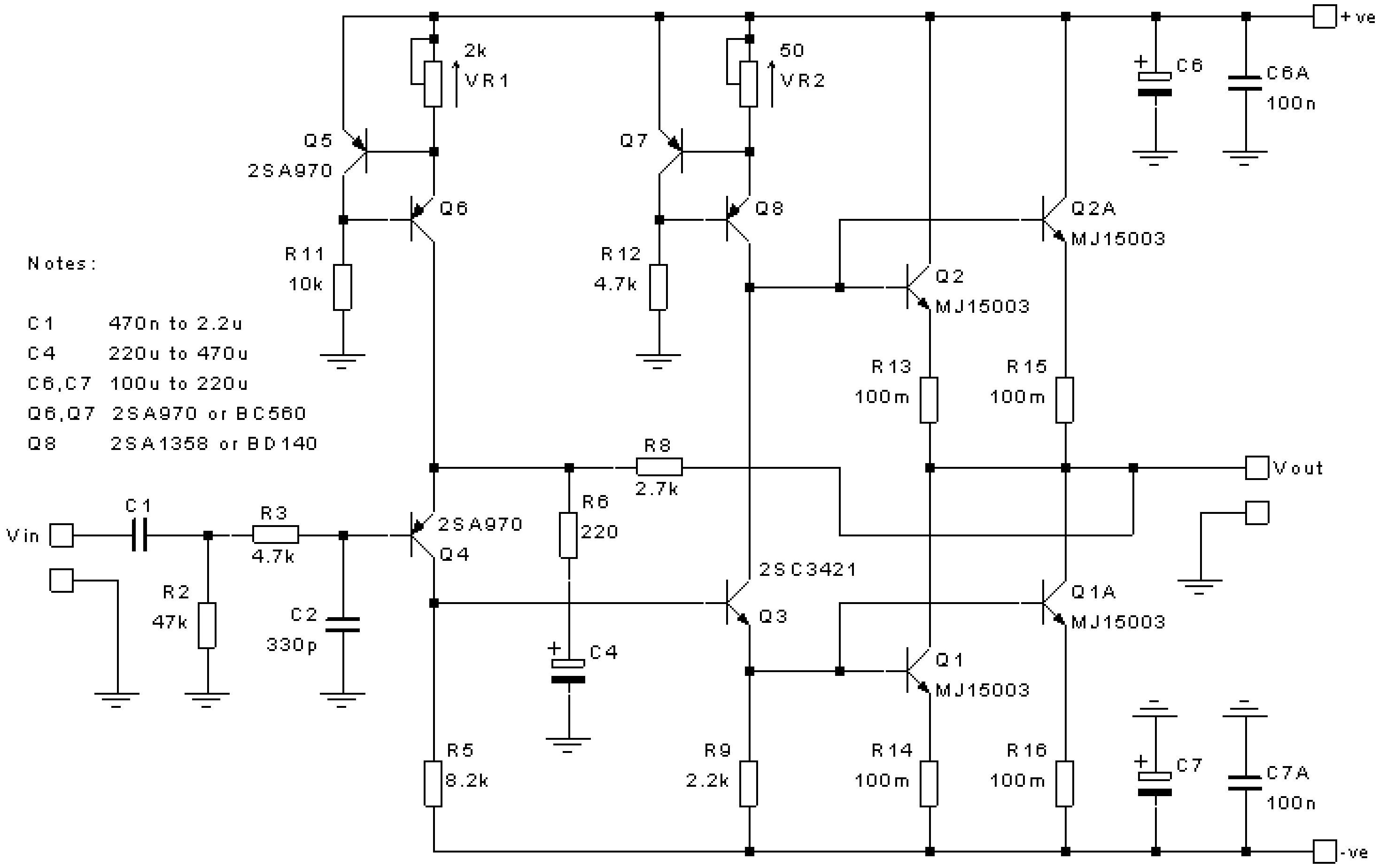


Fig 3 – The Higher Power Circuit

When used with conventional speakers, this circuit can deliver over 40W provided the supply rail voltage and quiescent current are selected to suit a specific load impedance. The supply rail voltage needs to be a couple of volts higher than the peak output voltage swing and the total quiescent current should be about 0.7 times the maximum output current. The power dissipated in each output transistor (supply rail voltage times half the quiescent current) should be limited to about 40 to 45W, assuming decent sized heatsinks are used (0.6 to 0.8degC/W per transistor).

The peak load voltage and current can be calculated from required power and the speaker's impedance in the normal way using:

Vpk = sqrt(2*Pwr*Rload) and Ipk = sqrt(2*Pwr/Rload)

To allow for speaker impedance variations, I would suggest that current is calculated using ¾ of the speaker's nominal impedance and voltage using 1½ times the nominal value. Of course, you are free to make your own assumptions about speaker impedance variations and to calculate the required supply rail voltage and quiescent current accordingly. From feedback I have received, higher quiescent currents tend to sound better so you may wish to bias the compromise between voltage and current accordingly (whilst keeping the power dissipation in the output transistors at a safe level).

The following table indicates the maximum power output into 8, 6 and 4ohm loads for some standard transformer secondary voltages, assuming a resistive load and without any allowance for the impedance variations mentioned above. The supply rail voltages assume a regulated supply, with the consequential volt drop, and the quiescent current has been calculated from either the maximum current into 4ohm or, in the case of the 25 and 30Vrms secondary, the transistor power dissipation limit.

Secondary Voltage (Vrms)	Supply Rail Voltage (V)	Quiescent Current (A)	Power 8ohm (W)	Power 6ohm (W)	Power 4ohm (W)
18	18	2.8	16	21	32
22	23	3.7	28	37	56
25	28	3.2	42	56	42
30	33	2.7	60	45	30

A JLH Class-A for the Quad ESL57

Credits: Original design – John Linsley Hood
Circuit modifications – Geoff Moss
Layout, pcbs and construction – Nick Gibbs

This version of the JLH Class-A amplifier is the result of a series of emails and design discussions which culminated in the subsequent construction of two high current amplifiers specifically optimised to drive Quad ESL57 electrostatic speakers.

Some months ago, I received an email from Nick Gibbs regarding his 1969 JLH and a possible upgrade to a 1996 version with a higher quiescent current. Nick's 1969 JLH was about 16 years old and had been in almost daily use. It had a 27V supply rail and a quiescent current of 1.2A and Nick was using it to drive his Quad ESL57s, since neither his Quad 405 nor JLH MOSFET amps would do so without tripping the protection circuits. The little 10W JLH worked well with the ESL57s, albeit with some occasional clipping on louder passages, and Nick felt that a higher current version would best meet his needs.

The ESL57 is a difficult load to drive in that it is capacitive and its impedance drops to a low of around 2ohm at 15kHz. A high current delivery is therefore required but, to offset this, the maximum voltage that should be applied to the ESL57 is only 33Vp-p. It seemed that a JLH Class-A with a reduced supply rail voltage and a higher quiescent current would be ideal.

After we exchanged a number of (sometimes lengthy) emails, the final design evolved. It was a cross between the 1969 and 1996 versions (hopefully with the better parts of each ☺) operating off +/-20V supply rails and with a quiescent current of between 3.5A and 4A. The circuit is shown in Fig.1, but it should be noted that Nick used MJ802 output transistors in place of the 2N3055s since he already had these devices available.

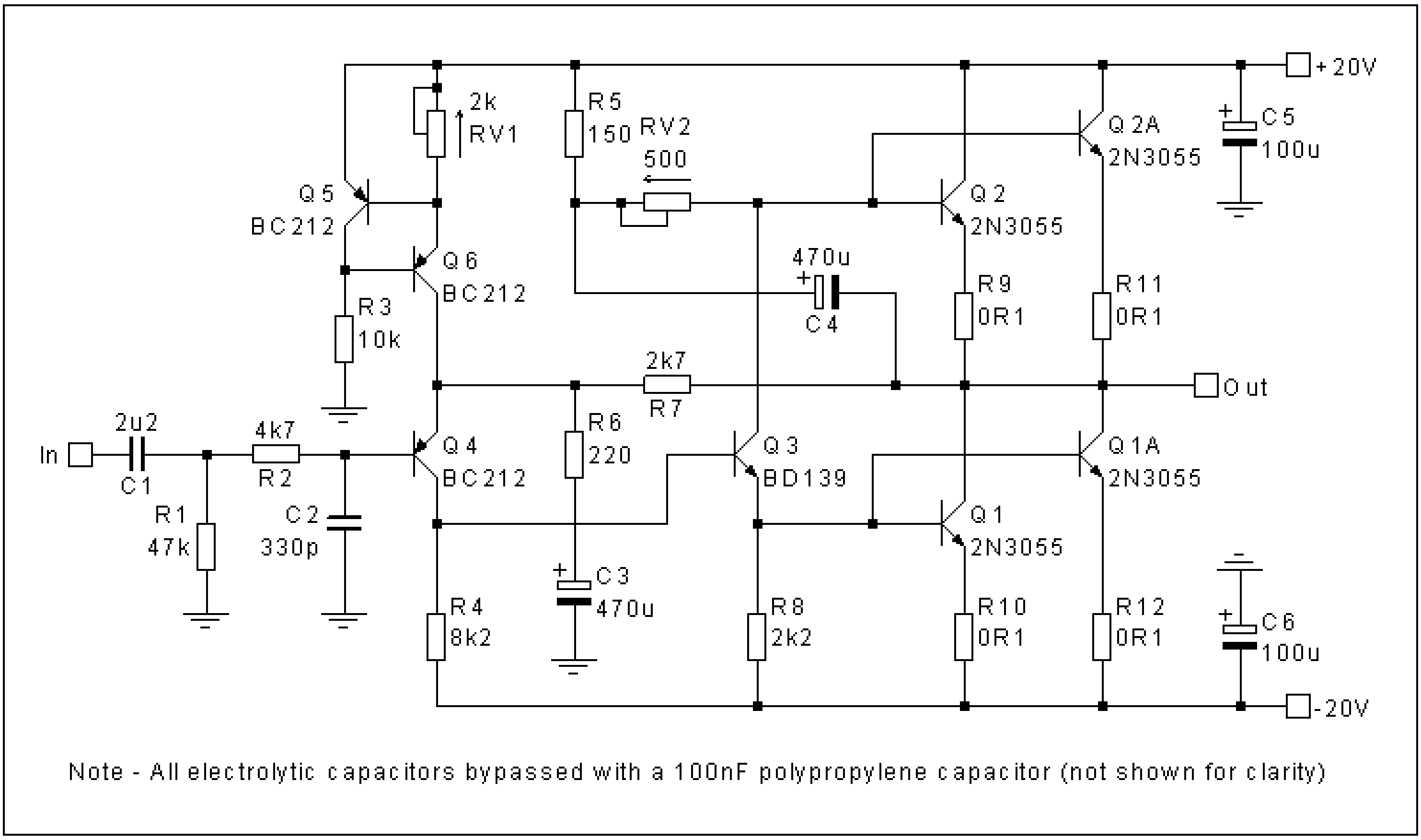


Fig. 1 - The Final Circuit.

As can be seen, the circuit is a mixture of the two original JLH versions, with modifications to enable an increase in quiescent current. Parallel pairs of output transistors have been used to keep the dissipation in each device at an acceptable level. The 0R1 emitter resistors are included to ensure equal current sharing between each device. The quiescent current control is the standard 1969 bootstrap method whereby C4 maintains a constant voltage across RV2 and thus a constant dc current into the bases of the output transistors.

The input stage of the 1996 version has been utilised, but for dc offset control the 7815 has been replaced with a constant current source to avoid the instability problems that have been encountered when the 7815 is operated at a low current. Several capacitor values have been increased to modify the low frequency -3dB point and to reduce low frequency distortion. High quality components have been used throughout.

Addendum - 4 February 2002

Note, care must be taken to ensure that R5 and RV2 are adequately rated. The current through these components is slightly greater than the sum of the output transistor base currents. The output transistor base current is the output transistor quiescent collector current (Ic) divided by the current gain (Hfe) of the device. The current through R5 and RV2 is therefore approximately equal to 4 x Ic / Hfe and this should be calculated for the chosen output transistor quiescent current and output transistor type. It is recommended that output transistors with a gain of 100 or more at the working collector current are used in this design to reduce the power rating requirements for R5 and RV2.

Whilst it should not be difficult to obtain fixed resistors with the required power rating, the preset potentiometer could be more of a problem since the more common ones are only rated at 0.5W or 1W, though higher rated devices are available. It must be remembered that the power rating of a preset, when connected as a rheostat, is proportional to the length of track in use. The required power rating must therefore be calculated from the current flowing through the preset and the full preset resistance value.

With high gain (>100) output transistors and a quiescent current of 3A, a 1W device should be adequate for R5 and 2W for RV2, provided RV2 is no greater than 500ohm. If a larger value of RV2 is found to be necessary, it will be best to use a 2W fixed resistor in series with RV2 to avoid the need for a higher power rated preset. (Note, the original value shown in the Fig. 1 for RV2 was 2kohm. This value has been changed due to the power rating considerations).

If RV2 needs to be set to below about 300ohm due a particular combination of quiescent current and transistor gain, I suggest that R5 be reduced to between 50 and 100ohm to avoid the need for increasing the size of the bootstrap capacitor C4.

The power supply (one for each channel) is shown in Fig. 2. This is basically the standard LM338K circuit, included elsewhere on this site, with some capacitor variations/additions.

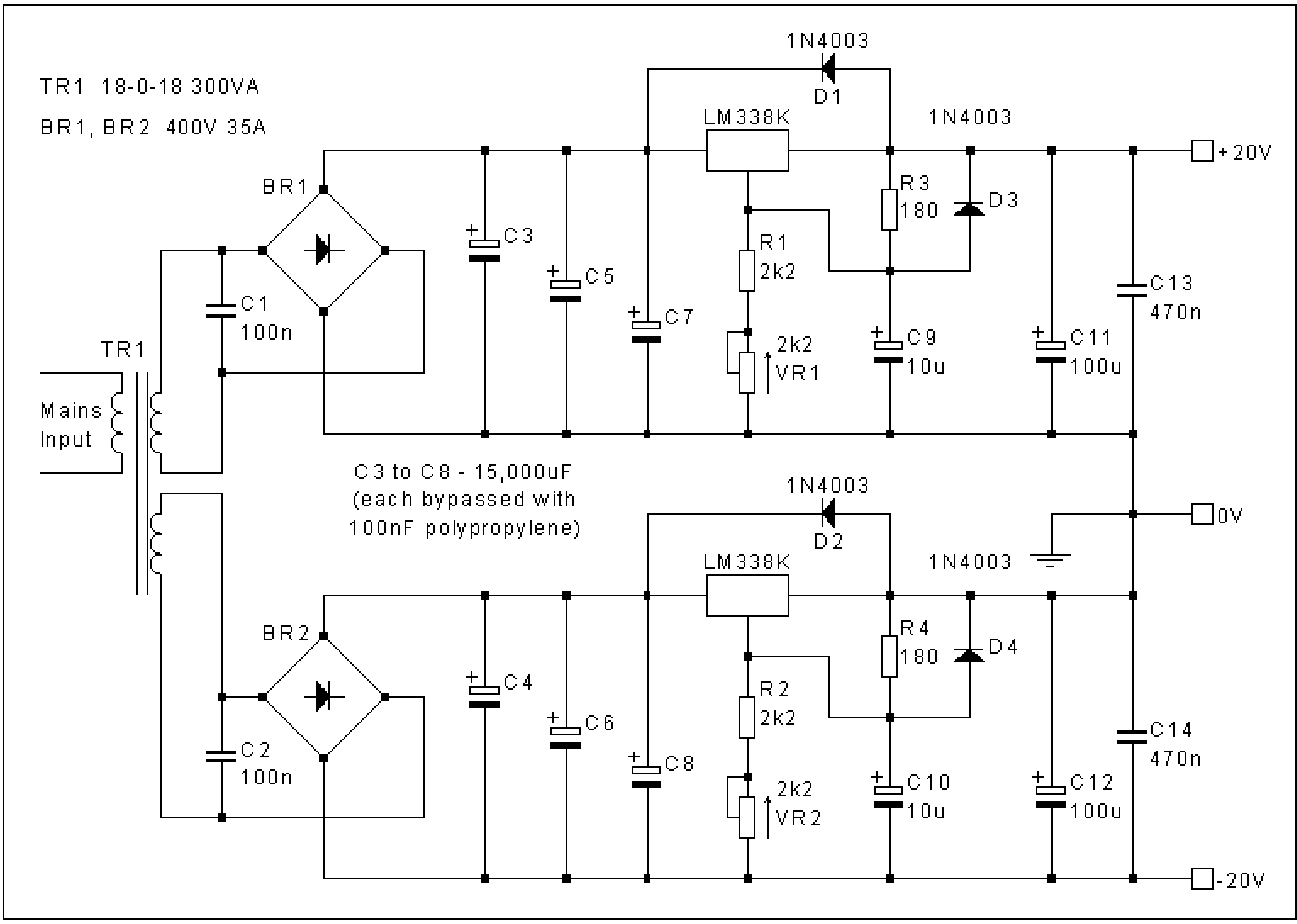


Fig. 2 - The Power Supply

Nick initially adjusted the quiescent current to 3.5A but found after listening tests that increasing this to 4A gave a noticeable improvement. Even at an Iq of 4A the amps run cool due to the substantial heatsinking (0.5degC/W for each output device and each LM338K). Variations in quiescent current and dc offset with temperature are minimal, with an Iq of 3.8A and a dc offset of less than 35mV at switch-on.

As for the sound quality, Nick’s initial comments are summarised below:

> “I have just spent two hours listening …….. I cannot believe the improvement over my old JLH. I have ended up with 20V rails and an Iq of 4A. You may well understand that I am feeling a little emotional at the moment so I will attempt to quantify the sound in a point form:-

Female vocal - incredible, makes the hairs on the back of my neck stand up.

Instruments and singers now appear as solid 3D objects, they have constant depth, if that makes sense?

No blurring of image or loss of depth on loud moments.

Acoustic guitar - real!

Bass - although the ESLs are 3dB down at 55Hz everything is so well defined, I would say at this point that my old JLH was brilliant here too.

I can hear more hiss from the source material, although I cannot as yet fault the top end reproduction, the ESLs are very revealing.

I think more than anything else it's the fact that the amps NEVER appear to get confused (?) (increasing the Iq from 3.5A to 4A prevented slight confusion/clipping (recovery) with heavy mid to upper band periods, at the levels I listen at). Constant image solidity, depth and remarkable detail at all times are what these amps are about, bloody brilliant!

I have broken into my Cambletown Whisky as celebration ………"

And a few days later:

"The amplifiers just get better the more you listen, real instruments and human voice are truly superb and very involving, plus of course the sensational 3D solid imaging. I wish you could hear them.

I am slowly working my way through my CD collection with the new amps, and it just gets better!"

And Nick's most recent comments:

"The combination of Marantz CD17 MkII + new amps + ESL57 is the first system that I have EVER heard that can do justice to the sound of a piano. I have been listening to a 'Deutsche Grammophon' recording of Liszt's Hungarian Rhapsody, a little hissy, but for the first time, the attack (?, I don't know how to describe this), the first instances of a piano note and all that goes with it to convince you that you are listening to a piano, is there. I have friends who play the piano and so I often listen to the 'real thing'. I consider this ability of the new amps very important. I thought the ESLs would give me this with pretty much any amp, but it has taken the new JLH amps to actually do it. Additionally, the insight into Bizet's Carmen, again on 'Deutsche Grammophon', is exceptional.

Increasing the size of the electrolytics from 220uF as in my original JLH to 470uF has very noticeably extended the bass response.

The original JLH is a magnificent amplifier, but with the modifications it has become outstanding."

Circuit Boards

Nick has kindly supplied me with a copy of his pcb layout for both the amplifier board and the regulator board in case they are of interest to other constructors. These are reproduced below at full size. It should be noted that the amplifier board is laid out for Caddock MP930 series power resistors, on heatsinks, for R5, R9, R10, R11 and R12 and also that Nick has used two resistors in series (100ohm and 50ohm) for R5 as these were more readily available (and cheaper). Component overlay diagrams have also been included after the pcb diagrams.

The actual board sizes are: Amplifier board – 8.55" x 5.25" (217mm x 133mm)
Regulator board – 3.3" x 2.9" (84mm x 74mm)

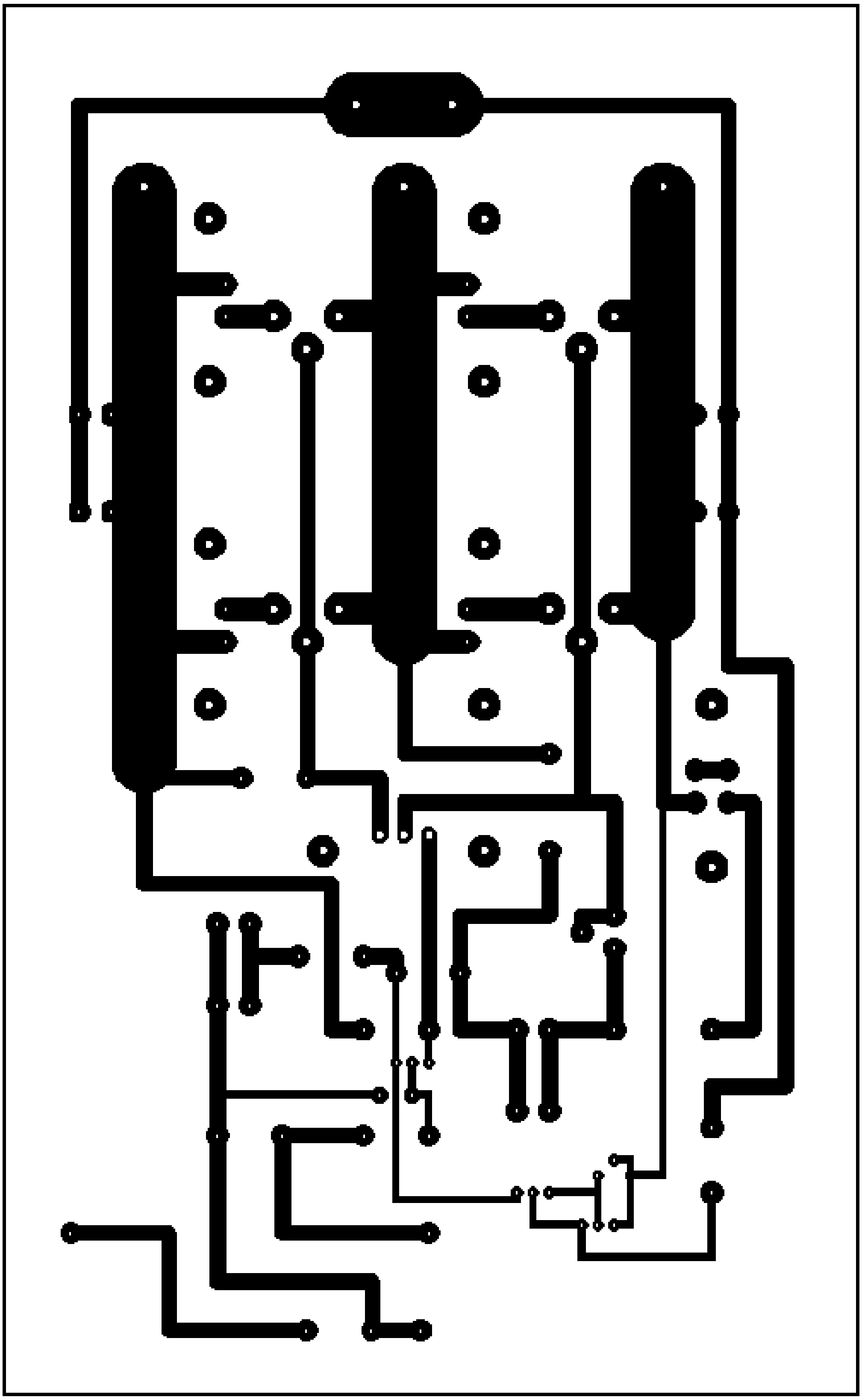


Fig. 3 – Amplifier pcb (viewed from copper side)

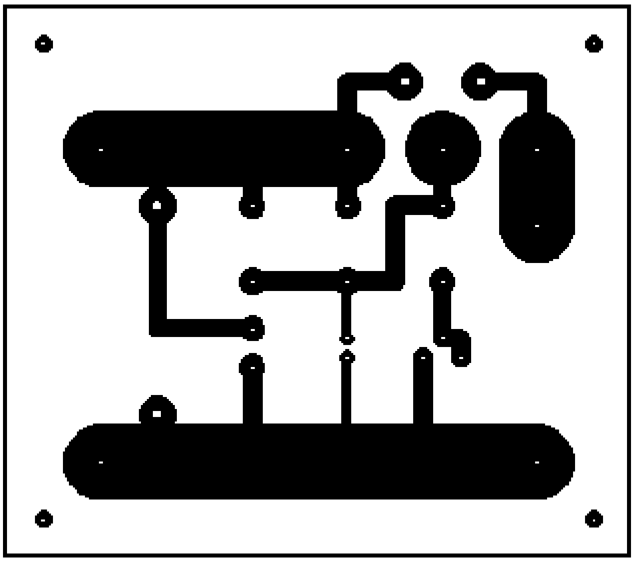


Fig. 4 – Regulator pcb (viewed from copper side)

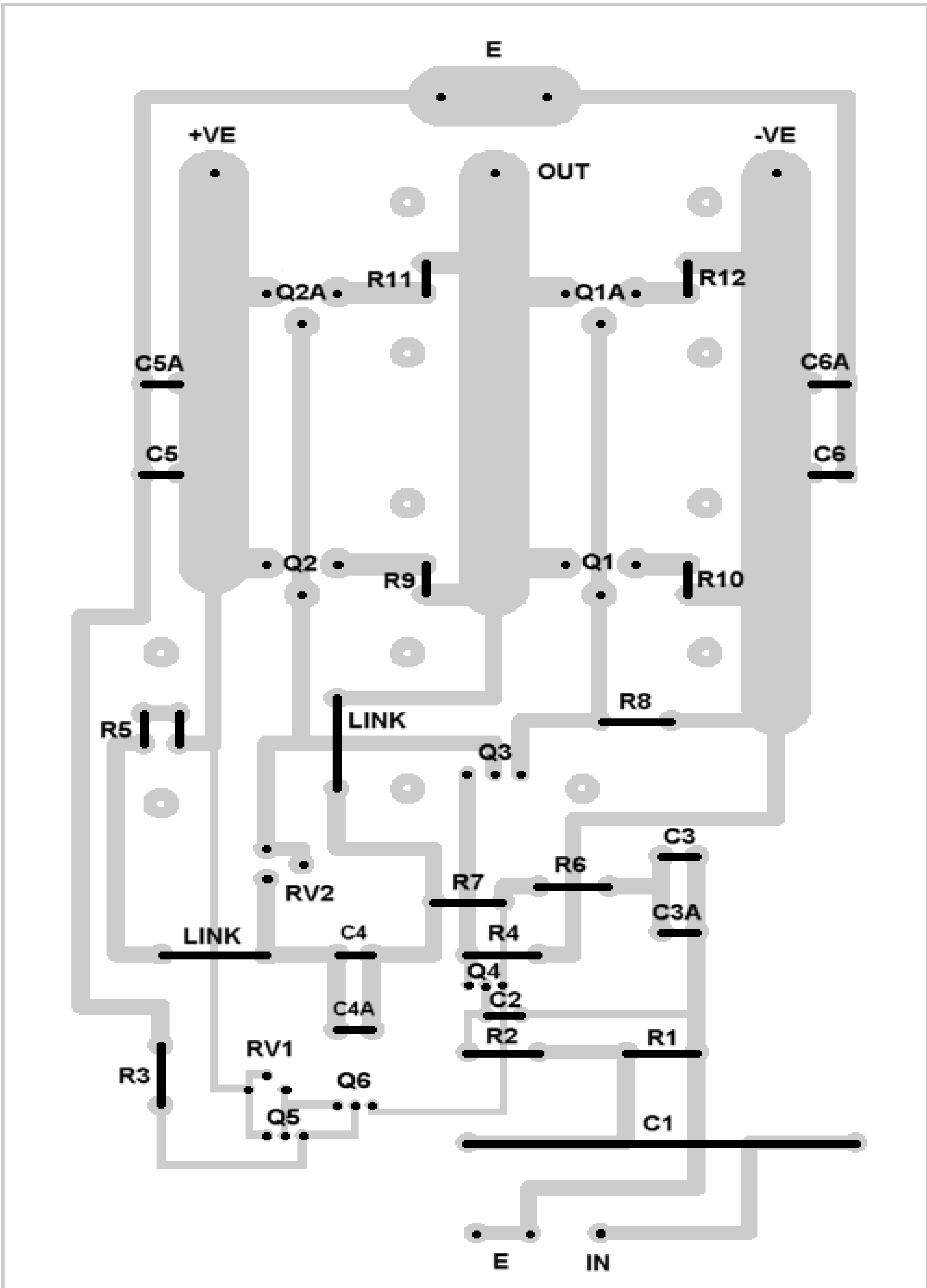


Fig. 5 – Amplifier pcb overlay (viewed from component side)

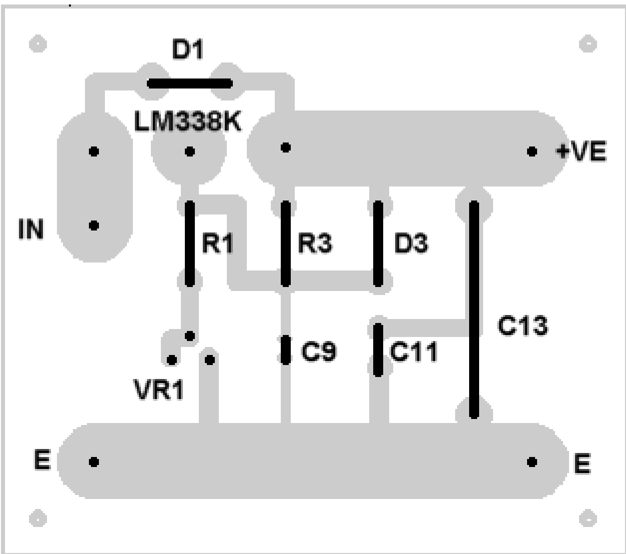


Fig. 6 Regulator pcb overlay (viewed from component side)

Finally, for those of you interested in seeing the results of Nick's labours:

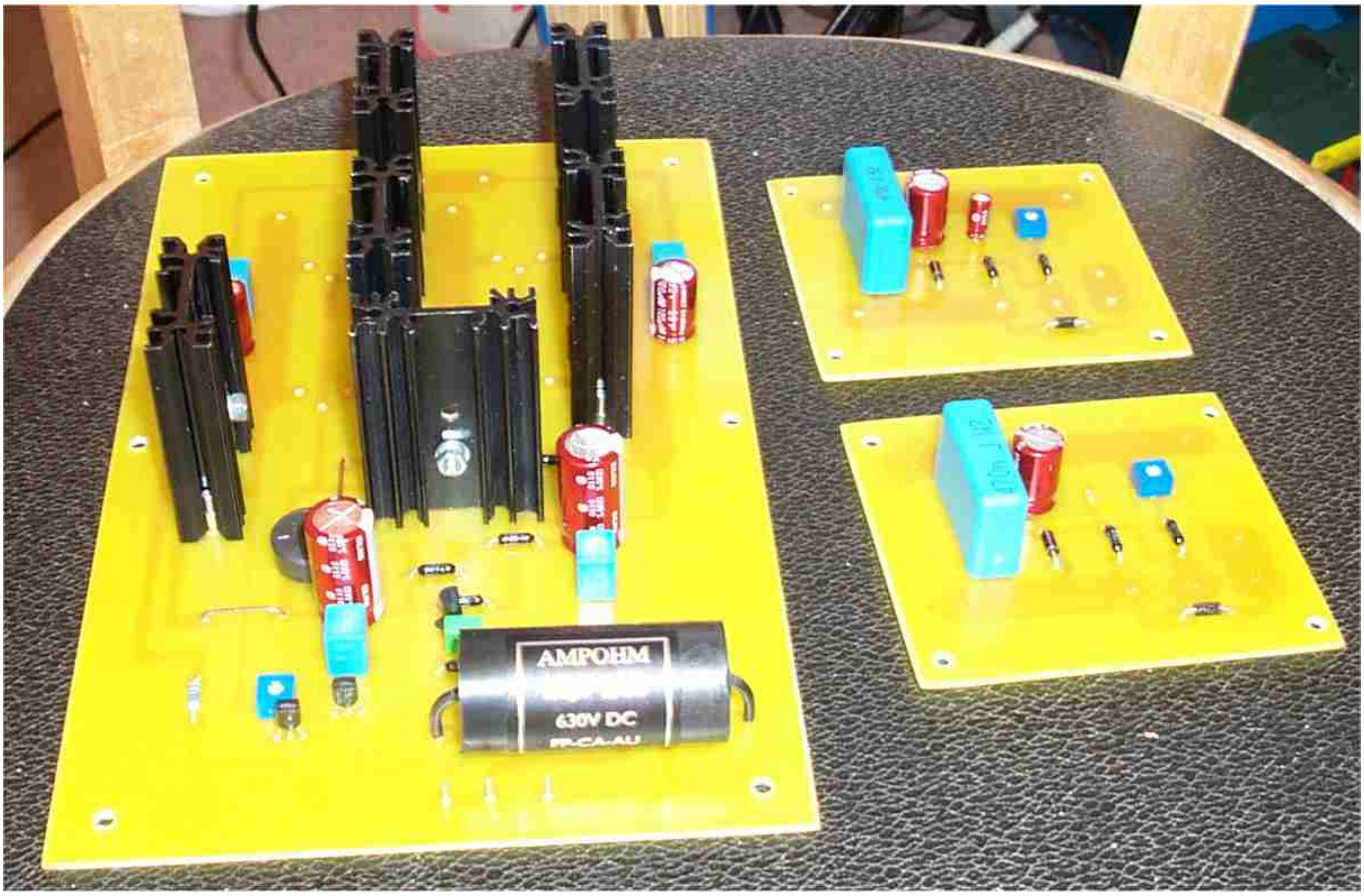


Photo. 1 – The pcbs

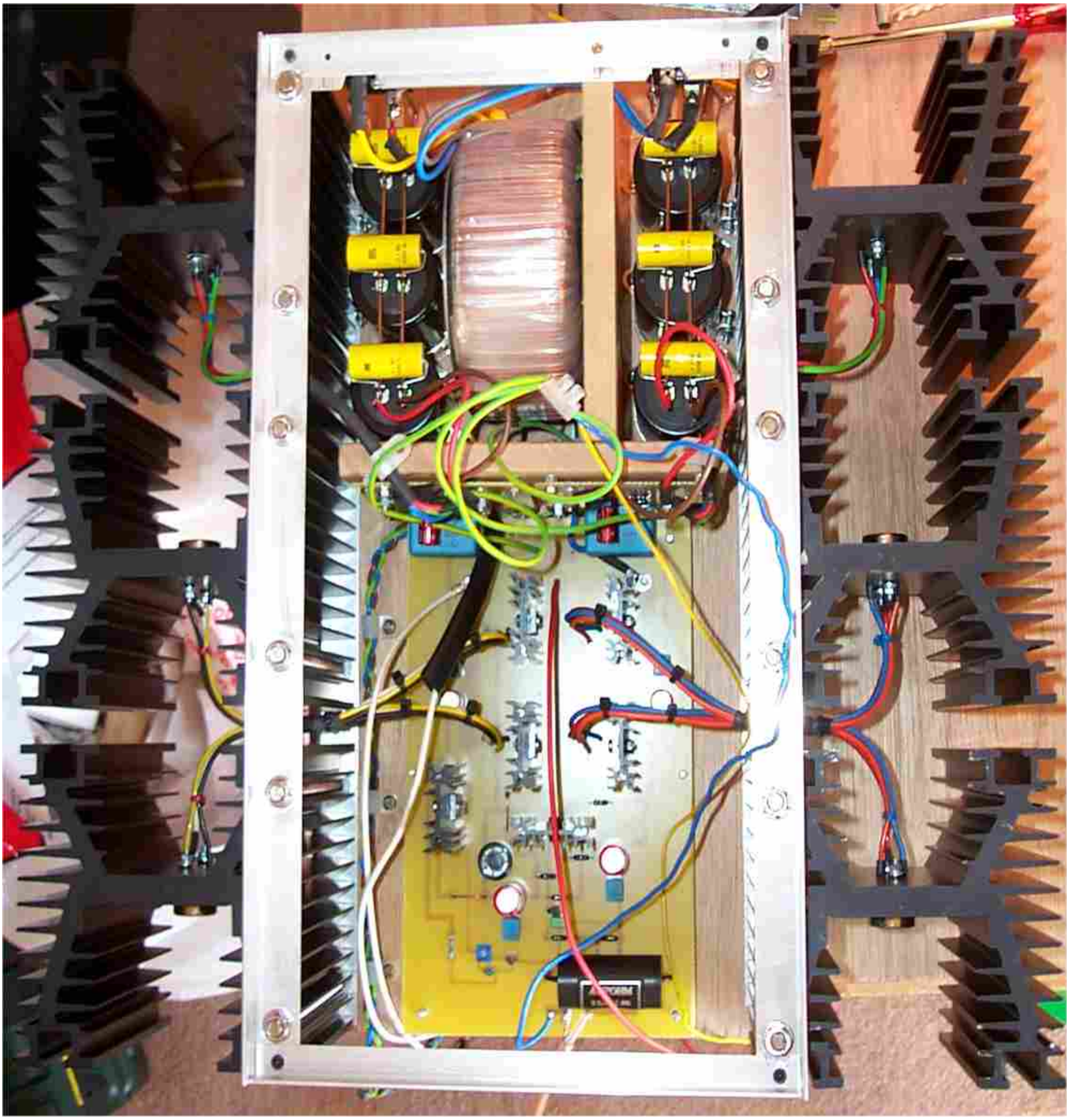


Photo. 2 – Nearing completion, a plan view

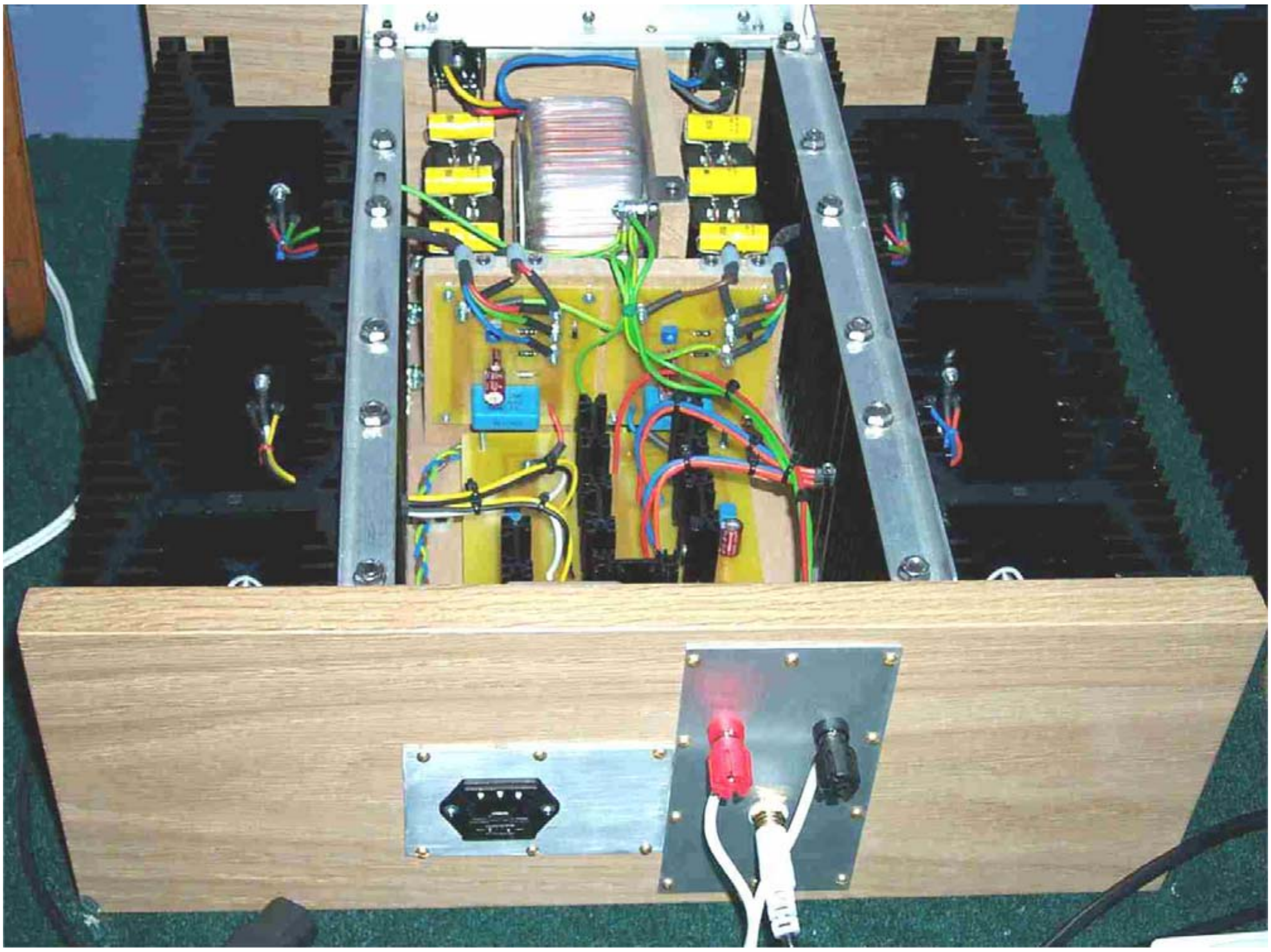


Photo. 3 – The finished amplifier with top cover removed

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Design Notes

Design notes for the JLH Class-A amplifier.

The 1996 version using the specified transistors with +/-22V supply rails and a quiescent current of 2A has an approximate rms power output, into a resistive load, of 10W into 16ohm, 20W into 8ohm, 15W into 4ohm and 10W into 2ohm.

For minimum distortion, Tr1 and Tr2 should be a matched pair. If this is not possible, the device with the higher gain should be used in the Tr1 position.

Low gain output devices such as the 2N3055 should only be used with a high gain driver transistor, for example the 2N1711 or 2N3019 (or a suitable alternative - perhaps a specially selected BD139).

The output transistors should ideally have an f_T of 4MHz or more, though many amplifiers have been successfully used with 3MHz devices (or even lower). For information, the 2N3055 datasheet from ON-Semi quotes an f_T of 2.5MHz and that from ST, 3MHz (though other 2N3055 manufacturers quote a figure of 0.8MHz). TIP power devices are usually 3MHz.

Because of the high dissipation in the output devices (about 45W each for the 1996 version), I suggest that any proposed output transistor should have a thermal resistance (junction to case) of less than 1°C/W. Alternatively, a parallel pair arrangement (with 0R1 emitter resistors) can be used. For guidance on heatsink sizing and transistor mounting see the ‘Heatsinks’ article at the ESP Audio Pages.

To minimise quiescent current and dc offset drift due to temperature rise, resistor R10 (0R33) should be a 7W or 10W type or 3 x 1R0 3W in parallel. The resistor(s) should be stood-off the pcb to ensure adequate ventilation. For the same reason, Tr5 should have an adequate heatsink. From experience, I would suggest a minimum of 10 °C/W, though around 6°C/W would probably be better. If a BD140 is used instead of the MJE371, a larger heatsink will be required since its thermal resistance (junction to case) is much higher than that for the MJE371. When laying out the pcb, try to keep R10 and Tr5 away from the output transistor heatsinks.

I must stress that the circuit diagram (Figure 3.) in the original article for the 1996 version contains an error. The negative end of the feedback capacitor (C4) is shown connected to the –ve supply rail. This will result in excessive hum due to the supply rail ripple voltage being injected into the feedback path (Tr4 emitter). To prevent this problem, the negative end of C4 should be connected to the 0V (earth) point.

The value of the input capacitor (C4 or C1) can be usefully increased to lower the low frequency – 3dB point and improve the bass response of the amplifier. I suggest a value of between 1uF and 2.2uF. A polypropylene capacitor is preferred in this position.

The value of the blocking capacitor in the feedback circuit (C3 or C4) can be usefully increased to reduce the low frequency distortion of the amplifier. Values between 470uF and 1000uF would be suitable. Rudy van Stratum has tried values up to 1000uF and has found that 470uF sounded best in his modified (dual rail) 1969 version.

In theory, and in simulation, increasing the value of the bootstrap capacitor (C1) in the 1969 version to between 470uF and 1000uF reduces the frequency at which low frequency distortion starts to increase due to the non-linearity of the current source that controls the output stage quiescent current. Rudy has also tried values up to 1000uF in this position. Contrary to expectations, 1000uF caused a ‘thickening’ in the bass and a loss of ‘air’ and ‘finesse’ in the treble. He has now found that 470uF gives the best results.

I previously suggested by-passing all electrolytic capacitors with a 100nF polypropylene capacitor (in parallel with the electrolytic). This may not have an audible effect, but it ensures a low esr at high frequencies. Rudy has reported that, when he has tried paralleling capacitors in the past, the sound quality has deteriorated in comparison to a single capacitor. His exact comments were:

“About 10 years ago this was standard practice for me, everywhere I used a small film cap to better the high frequency behaviour. But I believe now that a good design does not need such things, the best amplifiers I heard use no such things. And on more than one occasion this bypassing produced sharp edges to the sound. Somehow it seems that you can hear two different capacitors. Compare it to the difference between a good broadband (full-range) speaker vis-a-vis a two-way system: you always hear two units. Some of the natural integration is gone.”

As other articles I’ve read appear to come to the opposite conclusion, I’ll keep an open mind on this issue.

I have received two reports recently regarding a problem with oscillation of the 7815 voltage regulator in the 1996 design. There are several cures for this problem. One would be to replace the 7815 (and C3, RV1 and R1) with an adjustable constant current source (a decoupled resistance, an FET, an LED/transistor or a two transistor circuit). The current source will need an adjustable output of between 0.4 and 0.5mA. The second solution is to improve the stability of the 7815 by increasing the output capacitor (C3) to between 22uF and 100uF. In addition, it could be worthwhile adding a resistor from the output of the 7815 to earth to ensure a minimum output current. A value between 3k and 4k7 should be suitable. None of the 7815 data sheets that I have been able to find has specified a minimum output current, but adjustable regulators such as the LM317 call for a minimum current of around 3.5mA. From one constructor’s experience, the 78L15 seems to be more prone to oscillation than the standard 7815 so, even though the current draw is less than 0.5mA (or 1mA if one regulator is used to feed both channels as in the original diagram), I suggest using the latter.

When using the original (1969) bootstrap arrangement for quiescent current control, care must be taken to ensure that R1/R2 (1969 article Fig. 3) or R1/RV1 (1996 article Fig. 1) are adequately rated. The current through these components is slightly greater than the sum of the output transistor base currents. The output transistor base current is the output stage quiescent current (Iq) divided by the current gain (Hfe) of the output devices. The current through R1/R2 or R1/RV1 is therefore approximately equal to 2 x Iq / Hfe.

If the current/resistor values of Table 1 (1969 article) and output transistors with a current gain of 100 or more are used, the resistor power ratings shown in Fig 3 (1969 article) are adequate. If low gain (circa 50) output devices are fitted, the resistor power ratings should be increased to about double those shown in Fig 3. For other resistor values or quiescent currents, the required power rating of R1/R2 or R1/RV1 should be calculated.

Whilst it should not be difficult to obtain fixed resistors with the required power rating, a preset potentiometer could be more of a problem since the more common ones are only rated at 0.5W or 1W, though higher rated devices are available. It must be remembered that the power rating of a preset, when used as a rheostat, is proportional to the length of track in use. It is therefore necessary to determine the power rating from the current flowing through the preset and its total resistance value. It may be necessary to use a fixed resistor in series with a lower value preset to form RV1 in order to keep within the power limits of the preset.

For those who prefer the greater simplicity of the 1969 version, but wish to avoid the output capacitor (C2), the circuit can be modified to operate off dual supply rails. Figures 1 and 2 illustrate two methods of achieving this. It must be stressed that Option 1 has yet to be verified in practice (so far as I am aware), but Option 2 has been successfully implemented by at least one constructor.

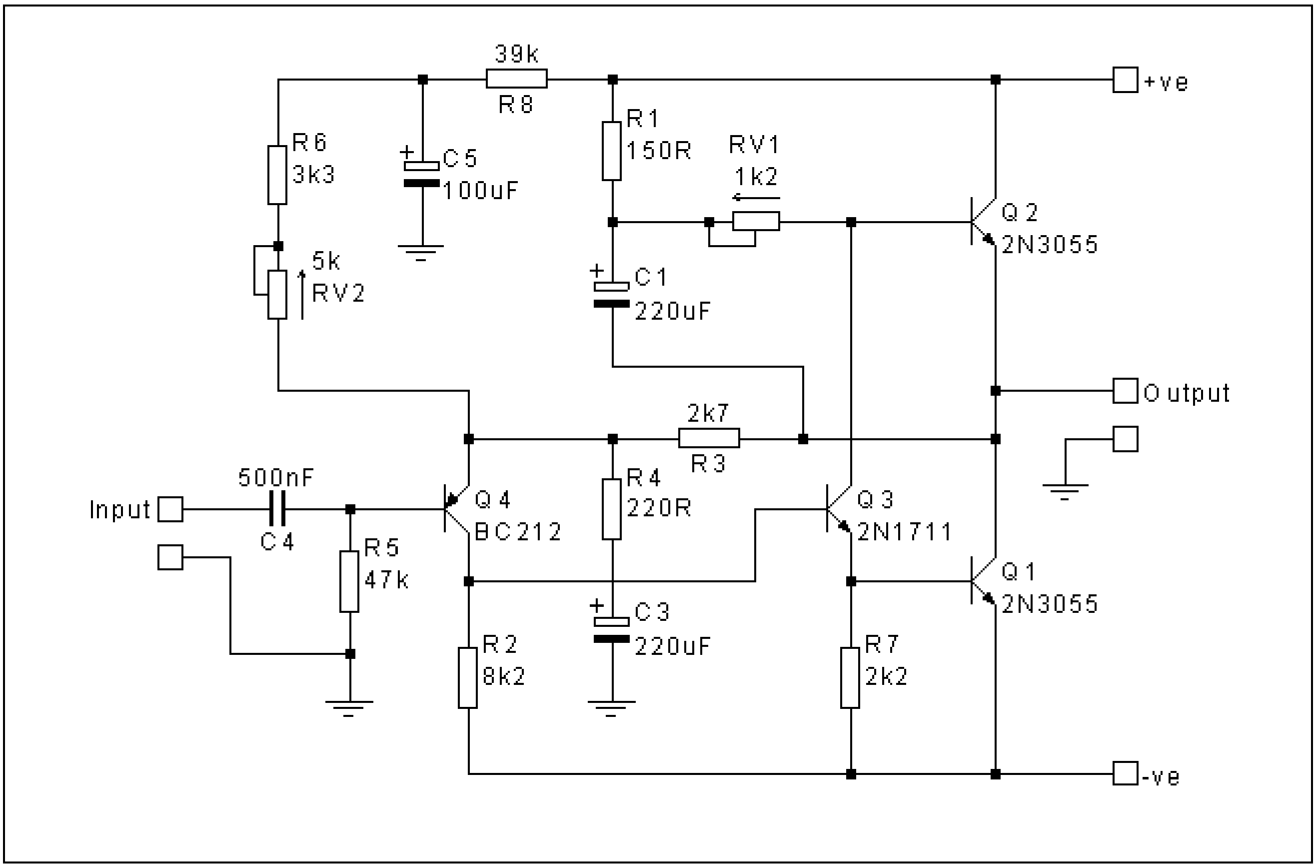

Figure 1. 1969 design with dual supply rails (Option 1)

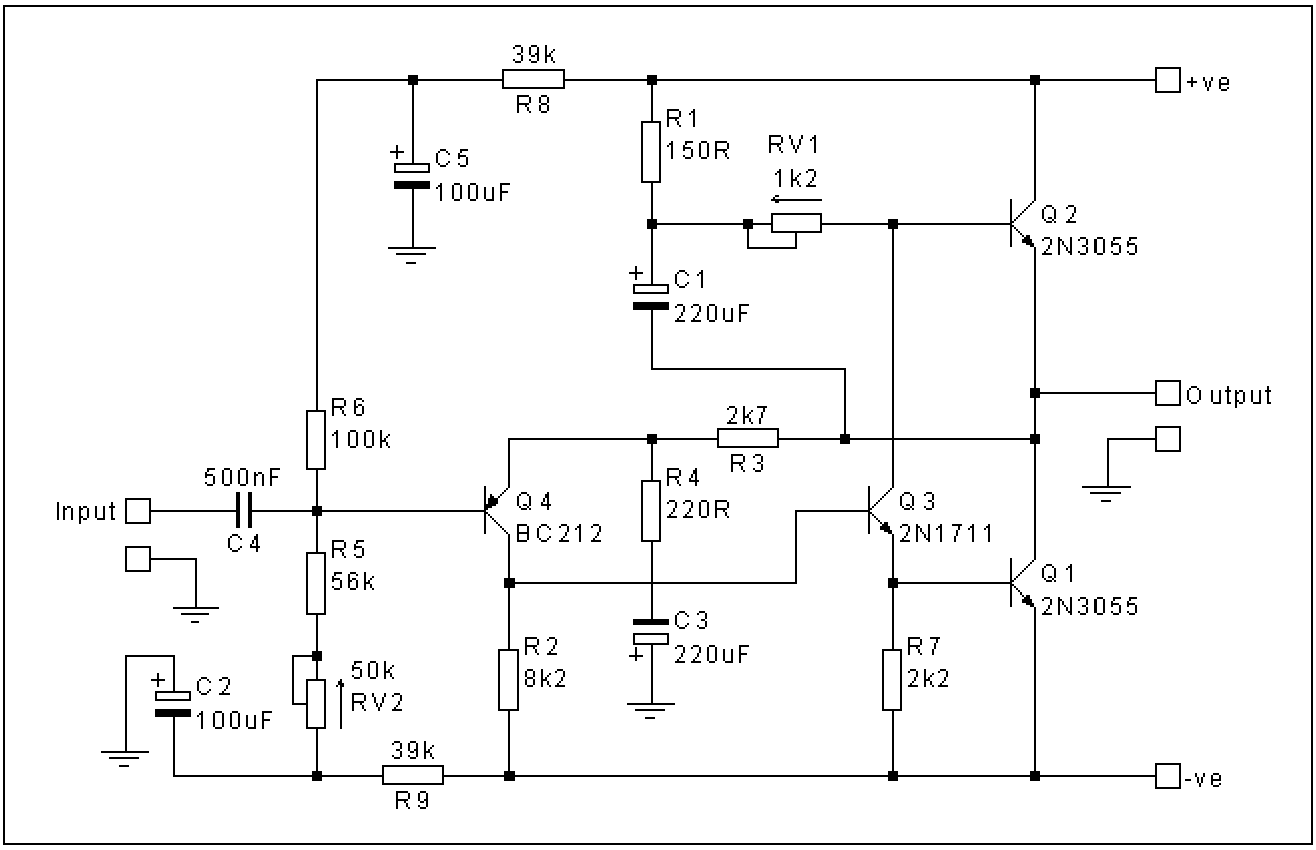

Figure 2. 1969 design with dual supply rails (Option 2)

For more design information relevant to this type of amplifier, see Project 36 at the ESP Audio Pages. The amplifier in this project is very similar to the JLH 1969 version (the main difference being the addition of a transistor to the quiescent current control circuitry) and Rod Elliott gives a good explanation of how he determined that this is the optimum topology for a simple solid-state Class-A amplifier.

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HISTORY: Page created 01/05/2001
10/05/2001 Added link to Quiescent Current and DC Offset page
16/05/2001 Diagrams redrawn
05/06/2001 Polarity of C3 in Figure 2 corrected
05/08/2001 Capacitor notes revised and 7815 oscillation notes added
31/01/2002 1969 bootstrap resistor power rating notes added
27/11/2002 dc offset servo paragraphs removed

Transistor Substitutes

The semiconductors used in the original 1969 circuit are, naturally, no longer available and even some of those shown in the 1996 update article can be difficult to source in some localities. The following list of substitutes has been prepared to assist those who are having difficulties in finding the specified devices. I have also included details of the working voltage, current and power dissipation for each transistor, when used in the 1996 circuit, so that other alternative devices may be considered.

Device	Original Device 1969	Original Device 1996	Substitutes
Tr1 / Tr2	MJ480 / MJ481	2N3055	2N3055 / 2 x TIP3055

Tr3	2N697 / 2N1613	2N1711	2N3019 / BD139
Tr4	2N3906	BC212	BC559 / BC560
Tr5	None	MJE371	BD140

Table 1. Commonly available or preferred transistor substitutes.

Notes to Table 1:
The 2N3055 should be epitaxial-base type with high f_T (preferably 4 MHz)
The 2 x TIP3055 are a parallel pair with 0R1 emitter resistors
The BD139 should preferably be selected for high gain to minimise distortion. If possible, use the BD139-16 (the manufacturer's higher gain device)

The use of more modern 'audio' power transistors with a high current gain-bandwidth product (f_T), such as the 2SC5200, 2SC3281 and MJL3281A, is not recommended at present. The >30MHz f_T of these devices causes the open-loop gain to remain above unity when the phase shift through the amp reaches 180°. This results in instability and oscillation, which requires additional compensation such as a dominant pole capacitor. In a simple circuit such as this, the provision of a compensation capacitor can significantly increase distortion levels unless other circuit changes are made (which perhaps defeats the object of this simple design). However, I will be investigating various possible options for solving the instability problems since I would really like to try the highly linear MJL3281A device.

I have received feedback from one constructor, Tim Andrew, who has been trying alternative output transistors in his JLH 1996 version. The MJL3281A gave clearly audible oscillation. The MJ21194 gave a noticeable improvement in sound quality, but introduced a low frequency hum, the cause of which has yet to be determined. The MJ15003 gave a significant improvement in sound quality, similar to the MJ21194 but without the side effects. Tim's opinion is that, when compared to the 2N3055, the bass is tauter and faster and the top end less 'splashy'. In a subsequent email about the MJ15003, Tim went on to say:

> "It's no good, I just had to email you again to say how good these transistors are. Recordings that were previously hard and bright are now sumptuous with crystal clarity, while recordings that were dull are now alive with a new sense of vibrancy. They seem to go particularly well with the tantalum film resistors that I have just fitted. I know you plan at some point to change to full range speakers, but I would seriously recommend that instead, you try these transistors with paper-in-oil caps, preferably copper foil on the input. Audio Note are introducing large value 50 volt P-in-Os for speaker crossovers. If you try these too, I would say you would be very happy indeed. It seems people just don't realise what they are missing with these P-in-Os. They have a total lack of hardness that has to be heard to be believed. My tweeters are metal domes, people say they don't like them because they sound metallic, but here they have a smoothness and clarity that is difficult to describe. Anyway, thanks again for the suggestion of these transistors, they are a big step up from the 2N3055s and I wouldn't go back now."

Following Tim's successful trial of the MJ15003, another constructor, Jason Wou, tried the substitution and sent me the following feedback:

> "I just replaced 2N3055 with MJ15003. It was direct swap. I didn't really have to adjust anything, it was
> basically a one-to-one swap. I had lots of MJ15003 to build a Leach Amp.
>
> My impression is the MJ15003 is DEFINITLY better!! I was getting goosebumps. ;) Sounded so real. Smoother highs and midrange (I won't comment on bass since I use a subwoofer). It improved the already superb sounding amplifier even more! I guess I won't be using those transistors for the Leach Amp any more!
>
> The MJ15003 is more expensive than the 2N3055, but not by much. Maybe a dollar or two more. From now on if there's any amp project with 2N3055 in it, I will be using the MJ15003!

How exciting. My amp is singing at this very moment, it sounds just so much better."

The MJ802 has also been proven to work in place of the venerable 2N3055, see 'A JLH Class-A for the Quad ESL57'

If alternative power transistors are required, they should be selected to meet the requirements of Table 2 and should have an f_T of around 4MHz. Devices with a low junction-case thermal resistance are preferred.

I have not yet found a commonly available alternative for the 2N1711 (Tr3), other than the (selected) BD139. The 2N1711 and 2N3019 are preferred (if one or the other can be found) over the BD139, due to their higher gain.

Other substitutes for Tr4 include, amongst others, the BC212L, BC556, BC557 and 2SA872. Low noise devices such as the BC559, BC560 and 2SA872 are preferred.

Note, the substitutes given above do not necessarily have the same case style or lead-out arrangement as the original devices. Manufacturer's data sheets should be consulted to determine the relevant differences.

The following table can be used to assist in the selection other suitable transistors. The table shows the peak values (derived from simulation) of voltage, current and power in each transistor for a 1996 design with +/- 22V supply rails and a quiescent current of 2A. The simulations were run using 4, 8 and 16 ohm resistive loads and full-load figures were checked with source voltages set to give the maximum (non-clipping) output and with source frequencies of both 50Hz and 1kHz. The maximum figures obtained in the simulations are included in Table 2. Note, the maximum figures for a 1969 design will be lower as the power output is less if the original article is adhered to. When selecting alternative devices, an allowance must be made to provide a factor of safety. I suggest as a minimum that the voltage and current be multiplied by a factor of 1.5 and the power by a factor of 2.

Device	Voltage (Vce)	Current (Ic)	Average Power	Maximum Power
Tr1	40V	3.1A	45W	49W
Tr2	40V	2.7A	43W	56W
Tr3	40V	47mA	475mW	575mW
Tr4	23V	0.41mA	6mW	8mW
Tr5	39V	50mA	985mW	2W

Table 2. Maximum voltage, current and power for transistors in a 1996 design.

Before I get any queries, please note that the maximum power, under load, does not coincide with the maximum voltage or the maximum current, therefore the power figures cannot be derived from the multiplication of columns 2 and 3.

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HISTORY: Page created 01/05/2001
22/05/2001 2N3019 added
27/05/2001 Reference to BD139-16 added
09/09/2001 Caution regarding high ft output transistors added
07/11/2001 Notes re MJ15003 and MJ802 added

D C Voltages

The following tables of dc voltages are provided to assist in initial testing and any fault-finding that may be required. They have been prepared from simulations of the 1969 and 1996 versions. The 1969 version was simulated with a supply rail voltage of 27V and a quiescent current of 1.2A and the 1996 version with +/-22V supply rails and a quiescent current of 2A.

The last three columns in the tables have been included to allow calculation of nodal dc voltages at other supply rail voltages. Vs is the supply rail voltage (the value of a single rail for dual-rail supplies), Vbe is the base-emitter potential for a transistor (typically 0.7V) and Iq is the quiescent current.

1969 Version

Device	Emitter	Base	Collector	Emitter	Base	Collector
Tr1	0V	0.7V	13.5V	0	Vbe	Vs / 2
Tr2	13.5V	14.2V	27.0V	Vs / 2	(Vs / 2) + Vbe	Vs
Tr3	0.7V	1.4V	14.3V	Vbe	2.Vbe	(Vs / 2) + Vbe
Tr4	12.9V	12.3V	1.4V	(Vs / 2) – Vbe	(Vs / 2) – 2.Vbe	2.Vbe

1996 Version

Device	Emitter	Base	Collector	Emitter	Base	Collector
Tr1	–22V	–21.3V	0V	–Vs	–Vs + Vbe	0
Tr2	0V	0.7V	21.3V	0	Vbe	Vs – (Iq / 3)
Tr3	–21.3V	–20.5V	0.7V	–Vs + Vbe	–Vs + 2.Vbe	Vbe
Tr4	0.7V	0.1V	–20.5V	Vbe	0.1	–Vs + 2.Vbe
Tr5	21.3V	20.7V	0.7V	Vs – (Iq / 3)	Vs – (Iq / 3) – Vbe	Vbe

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HISTORY: Page created 07/05/2001

D C Voltages

The following tables of dc voltages are provided to assist in initial testing and any fault-finding that may be required. They have been prepared from simulations of the 1969 and 1996 versions. The 1969 version was simulated with a supply rail voltage of 27V and a quiescent current of 1.2A and the 1996 version with +/-22V supply rails and a quiescent current of 2A.

The last three columns in the tables have been included to allow calculation of nodal dc voltages at other supply rail voltages. Vs is the supply rail voltage (the value of a single rail for dual-rail supplies), Vbe is the base-emitter potential for a transistor (typically 0.7V) and Iq is the quiescent current.

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Device	Emitter	Base	Collector	Emitter	Base	Collector
Tr1	0V	0.7V	13.5V	0	Vbe	Vs / 2
Tr2	13.5V	14.2V	27.0V	Vs / 2	(Vs / 2) + Vbe	Vs
Tr3	0.7V	1.4V	14.3V	Vbe	2.Vbe	(Vs / 2) + Vbe
Tr4	12.9V	12.3V	1.4V	(Vs / 2) – Vbe	(Vs / 2) – 2.Vbe	2.Vbe

1996 Version

Device	Emitter	Base	Collector	Emitter	Base	Collector
Tr1	–22V	–21.3V	0V	–Vs	–Vs + Vbe	0
Tr2	0V	0.7V	21.3V	0	Vbe	Vs – (Iq / 3)
Tr3	–21.3V	–20.5V	0.7V	–Vs + Vbe	–Vs + 2.Vbe	Vbe
Tr4	0.7V	0.1V	–20.5V	Vbe	0.1	–Vs + 2.Vbe
Tr5	21.3V	20.7V	0.7V	Vs – (Iq / 3)	Vs – (Iq / 3) – Vbe	Vbe

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Earthing

Correct earthing is essential to minimise the possibility of noise being injected into signal lines and to reduce the likelihood hum being created by ground-loops. The diagram below gives my suggested earthing arrangements.

For safety, the mains/chassis earth must be connected to the amplifier. This connection must not be made at the transformer centre-tap or the reservoir ground (point ‘A’ in the diagram) since the voltage at this point will be affected by the high capacitor charging pulses. Connection at this point will cause severe ground-loop hum when the amplifier input is connected to source equipment that has its own mains earth connection.

Connecting the mains earth to the star point (point ‘C’ in the diagram) is better, but this can still cause audible hum due to the resistance of the connection between the input and the star point, since this connection will still carry any ground-loop currents. The best arrangement is to connect the mains earth to the chassis and then to the input socket, as shown in the diagram. This will minimise the possibility of hum due to ground-loops.

Supply rail decoupling capacitors and other non-signal carrying parts of the circuit should have a separate earth return path to the reservoir ground point so as to avoid injecting noise into the signal earth. Similarly, the earth returns from the components in capacitance multiplier or voltage regulator (if used) should have separate paths back to the reservoir ground.

Every effort should be made to keep the input earth and the feedback earth at the same potential since any difference between the two will appear at the output of the amplifier.

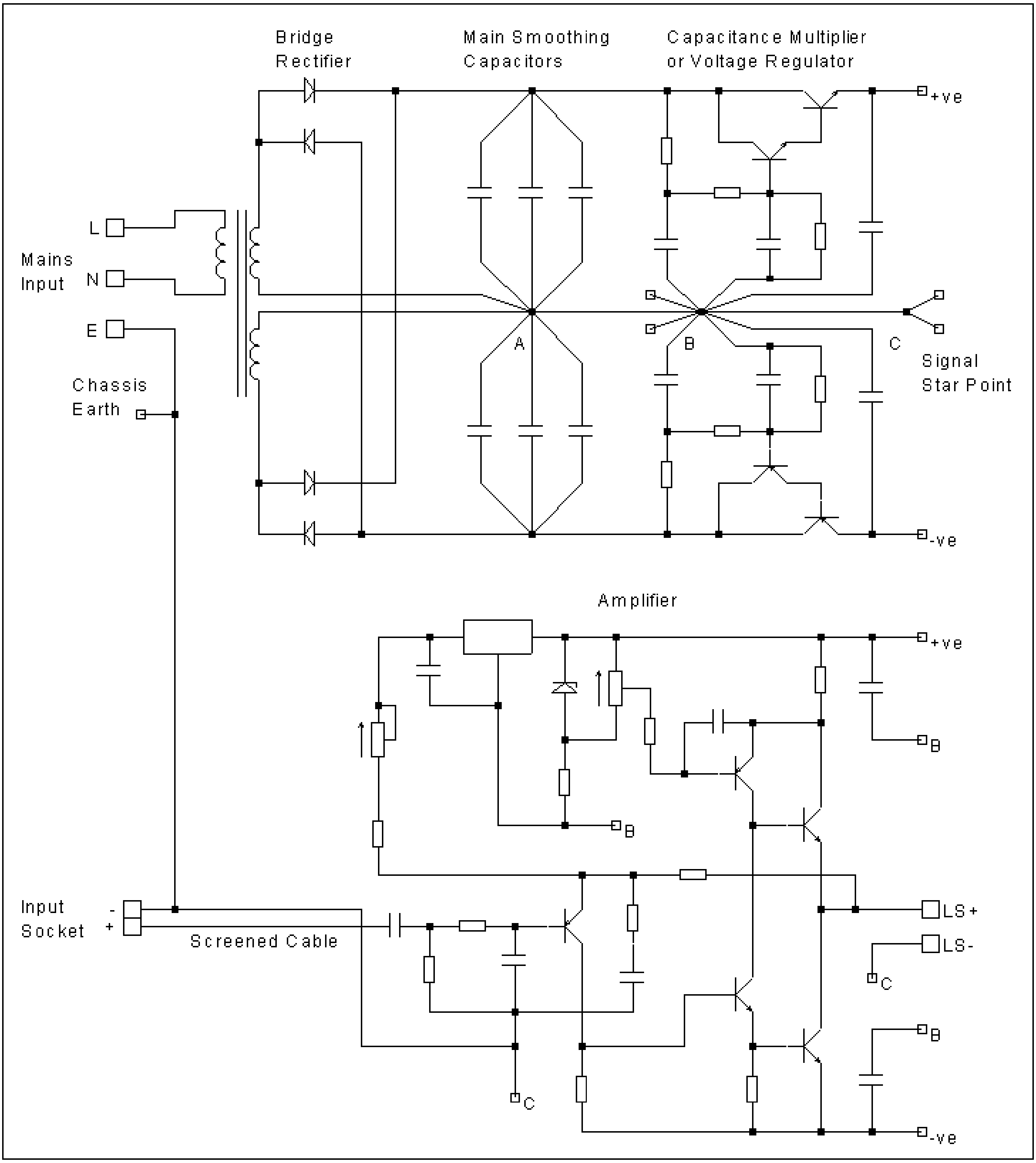
Ideally, points ‘A’ and ‘B’ in the diagram should be the same physical location (for example a large earth-bar). However, this is not always a practical arrangement if a capacitance multiplier or voltage regulator is used and so two, separated points have been shown. If a simple rectifier/capacitor power supply is used, the supply rail decoupling capacitors etc. should be returned to point ‘A’.

The signal star point should be joined to the reservoir ground through a short, thick connection. Under no circumstances should the reservoir ground be used as the signal star point, due to the high capacitor charging pulses present in this part of the circuit.

Note also that the output from the rectifiers should be connected directly to the smoothing capacitors and then the dc output taken from the same point to the capacitance multiplier, voltage regulator or amplifier. Under no circumstances should the capacitors be ‘teed-off’ as this will put sharp pulses on the supply rail and will cause an increase in hum.

Please also note that mains switching, ac fuses, dc fuses and output fuses have not been shown in the diagram. This is not to suggest that these essential safety requirements are not necessary. The diagram is solely intended to show the preferred earthing layout, not the full circuit.

Though I have shown the circuit of the 1996 version with a dual-rail power supply, the same principles apply for the 1969 version with a single supply rail.



For further guidance on earthing, and pcb layout in general, I recommend Doug Self's 'Audio Power Amplifier Design Handbook' (2nd edition). Chapter 13 gives plenty of useful information on these topics (and the rest of the book is worth reading as well). Additional information can also be found in the 'Earthing' article at the ESP Audio Pages.

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HISTORY: Page created 07/05/2001

Constructors’ Comments – Sound Quality

This page contains comments I have received from other constructors regarding the sound quality of the JLH Class-A amplifier. Though the quotes are extracts from emails, I hope that they can still be read in context. I have deliberately excluded my own comments since I am more than a little biased (I must be, otherwise I wouldn’t have spent the time needed to set up this site ☺). If you would like your views on the sound quality of the JLH (good or bad) adding to this page, please email me.

From Rudy van Stratum, Holland – Modified JLH 1969 version (dual supply rails, no output capacitor)

Now the sound is to my taste, very good indeed, comparable with several good tube amplifiers I have at my disposal.

Sounds better, more open, more airy, tauter bass, etc than my old and trusted C-coupled version.

I've listened extensively to the differences between the Hiraga and the modified JLH 69. Of course this need not be a definitive judgement, so .....

For a start: these two amps are very very good in transistor terms, indeed they are belonging to a remarkable class of all-time classics.

Differences between the two are very subtle. It's certainly not so that one of the two 'blows away' the other. If there are differences I should say that the JLH seems to flow somewhat more (vague terms, but alas). The general character of the sound is very similar (warm sounding, full bodied, airy).

Rudy has now tried different values (up to 1000uF) for the bootstrap (C1) and feedback (C3) capacitors and has sent the following additional comments:

I have settled on values of 470uF for both C's (bootstrap and feedback), they seem to work fine and are marginally better than both 220uF and 1000uF………. In comparison with my Hiraga the JLH now sounds very open and quick and with a fine texture in the highs. Very good indeed. But the Hiraga sounds 'fuller' and has more weight somewhere around 100-500 Hz I guess. My old JLH also had that full bodied 'tube' sound (but not the air and texture of the symmetrical version).

From Mike Jonasson, New Zealand – JLH 1996 version

I've compared my 1996 JLH Class A to some very costly valve amps - a single ended 300B project and a reworked Classic design from the 50's and I've listened critically to many commercial examples.

These have midrange charm which makes them attractive but valve devotees seem oblivious to shortcomings elsewhere - fairly obvious ones I find completely unacceptable. I have resolved that this probably has a lot to do with music choice - they listen to a lot of female cabaret stuff which is fairly light in musical texture, not too much outside the midrange and not a lot going on at the same time - avoiding intermodulation and bass problems that test equipment out on rock and classic.

It's not my cuppa tea and efficient speakers are mandatory for a lot of them.

The JLH Class A is also better than any SS amp I've built / heard. These include some Mosfet and Bipolar Class B designs with worldwide DIY popularity, as well as commercial products. The JLH simply has more finesse.

I believe the 1996 version betters the 1969 original as there is no capacitor to degrade the output signal.

From Nick Gibbs, England – JLH 1969 Version

I only expected the amp to be a stop gap until something more suitable came along. Well the amp has now been in use for 16 years (used on an almost daily basis) without a single fault or modification. During this time I have built and used the JLLH MOSFET design published in ETI, however, I always returned to the Class A amp after a few days.

Now I am using a pair of Quad ESL57 electrostatic speakers which present around 2 Ohms at 15KHz and about 30 Ohms at 80Hz, my Class A is running a 27V rail and 1.2A standing current, so I get a bit of clipping now and then. I am looking to build a version of the 1996 design with a substantially higher standing current to satisfy the Quad's.

When I first got the Quad's I used a Quad 405 amp to drive them, however, this didn't prove very successful as apparently it can only deliver a few watts into 2 Ohms. Clipping occurred at low levels and was particularly awful. Next I tried the JLLH MOSFET expecting better results. This amp was audibly superior, however, it is capable of very limited current supply into low impedance loads, after a quick clip the PSU shuts down. I have done very limited listening with the MOSFET amp as the PSU shuts down very easily. Then I tried the little Class A not really expecting any surprises. Well, (and I don't read ANY hifi mags) the stereo image and ambience of a well recorded performance were unbelievable, clipping appears very gentle ?, it is surprising how much material falls within the bounds of 10W even on the Quads. This amp combined with the Quad's really is fantastic, even friends who consider my interest a little strange said "Wow". However, the combination appears utterly ruthless in its reproduction, bad recordings are bad.

From Jason Hubbard, England – JLH 1969 version

I have built the amp (I used it for about 6 months but gave up on it - I had inadequate heatsinking and the fan I needed to run in order to keep it all cool bugged me too much). Sounded great compared to anything I'd used before but I yearned for more power to drive woefully inefficient speakers in a large room.

From Asen Tutekov, Bulgaria – JLH 1969 (3 ohm) version

The sound is good. I was a bit disappointed at the very beginning - maybe because I expected a miracle to happen. That was because I hadn't listened to a SE power amp before that moment. After several hours of listening I found out that the amp is very detailed, doesn't tire out the ears and controls the bass better than my Quad 405-2 ..... In short - I'm content with it.

From Jason Wou, Australia – JLH1996 version

….. the amp sounds fantastic …..

For some CD's (like GRP's Rippington) the amp sounds just amazing. I can nod nod nod throughout the
CD. But for some other CD's I can hear distortion-like sound which I couldn't hear from my other amps.
Most MP3's sound terrible with this amp. It looks as if this amp + B&W Solid speakers seems to be somewhat "selective" to the type of music and brands, or it's just way too revealing. Any opinion?

(Yes, several people have commented that this amp shows up poor recordings and my findings are the same. I have a box of about 40 CDs that I can no longer listen to but that seemed alright when using other, well-reviewed, Class-AB amps – Geoff)

From Ian Mackenzie, Australia – JLH 1996 version

My feelings echo those other builders in the comments page. The amp sounds very liquid, subtle but very detailed and coherent on good recordings compared to any conventional A/AB amp.

There also appears to be a very even perspective and uniformity of the tonal balance in both timbre and dynamics. In short this amp is excellent and a boon for such a simple diy project.

From David Smith, England – JLH 1969 version

I have used Rod Elliot's pre-amp designed for the DoZ amplifier to feed the JLH amps and I am truly very pleased with the results; the sound is very smooth and easy on the ears, particularly noticeable is the absence of unpleasant sibilance with broadcast female voices. For the first time I can see why the amplifier is so highly rated.

From Tim Andrew, UK – JLH 1996 version

My version of the 1996 JLH design uses paper-in-oil capacitors on the input, Elna Silmic electrolytics elsewhere, with Vishay bulk foil and Tantalum film resistors in all signal carrying parts of the circuit. At the suggestion of Geoff Moss, I have also replaced the 2N3055s with MJ15003s. These modifications have been carried out individually so as to enable me to evaluate each in turn. Each one has produced a very noticeable improvement and, in particular, the capacitor and the MJ15003 transistor changes must be singled out as making a larger improvement than I had expected. The amplifier now **sounds** far more powerful than many 200 watt amplifiers that I have heard and owned but has a warmth, purity, delicacy and speed that has eluded them all.

From Chris Ma, Canada – JLH 1996 version

The JLH compares to the Rotel multi-channel (power section only) as follows:- The brightness/harsh high is gone. The vocal is a lot fuller. The four string bass has more emotion to the notes. The kick drum is easier to distinguish from the electric bass guitar. It has more depth in the sense of stage but narrower than the Rotel. The focus or image position is better with the JLH. It reviews the fine detail much, much more with ease. Certain tracks in some CDs I would not like to listen to before with the Rotel because they sounded really bad but now I can enjoy them with the JLH. The background noise is really quiet. I can enjoy heavy rock music again with the JLH because it can handle a lot of things going on musically without tiring me out with just noise. For such a simple design and inexpensive final product it is a very good amp. Now the JLH makes me really miss the Pink Triangle turntable.

Chris's full email giving some background to his comments can be found here.

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HISTORY: Page created 24/06/2001
08/07/2001 Ian Mckenzie's comments added
24/07/2001 David Smith's comments added
05/08/2001 Additional comments from Rudy van Stratum
18/08/2001 Tim Andrew's comments added

14/09/2001 Chris Ma's comments added
05/12/2001 Tim Andrew's comments updated

A Dutch journey in JLH-land

A tribute to the JLH 10-15 Watt Class-A amplifier

By: Rudy van Stratum, Dutch audio-hobbyist

Why this article?

Simple: Geoff Moss asked me to write this article. And if that's not enough, I want to convince you that building a JLH is a wise thing to do. During the last few months, Geoff and I seemed to have a continuous hotline of e-mails about the progress I made in building a (second) JLH. Because I think we made some steps forward and gained some new insights, we decided we'd better share these thoughts with you as potential new constructors.

How I met the JLH

The first time I came across the JLH amplifiers was in 1997. A friend of mine sent me the 1996 article that appeared in Electronics World. I was immediately intrigued by the man and his ideas. To be honest I have not experienced much enthusiasm for any transistor amplifier since I discovered good tube amplification. The only transistor amplifier I have used every now and then during the last 10 years or so was the famous Hiraga 20 Watt amplifier (I have built several of these amplifiers, in a number of variants and settings). In my humble opinion, this is one of the best simple transistor amplifiers available to the diy public, up to this very moment. Before that I enjoyed many fine hours with the now obsolete Musical Fidelity A1 budget amplifier. Nevertheless, the story of the JLH inspired me to take another leap at transistors. I'm not a great fan of complexity and regulation, so I chose the original 1969 design, as shown in figure 1 of the 1996 article. This was even simpler than the Hiraga. Component selection did not seem very critical. Most of the components needed I had somewhere lying around the house. It took me several days and the first prototype (not yet in a neat enclosure) was playing before me. Finally, I built the prototype in true Hiraga-style. Here are some features:

- • A monstrous power supply, 3 big 40 000uF/75V Sprague capacitors per channel.
- • 35 Ampere bridge rectifiers, a transformer of around 200 VA per channel, double mono construction.
- • Passive power supply that gave me around 44 Volts of DC voltage. I set the thing at an idle current of 1.5 Ampere, so in effect I already had a 15 Watter from the start.
- • I used Motorola transistors everywhere, 2N3906, 2N1711 and 2N3055's.
- • Beyschlag 1 Watt resistors everywhere (I was very pleased with these resistors in my earlier designs, comparable with Holco's which are far more expensive).
- • Philips capacitors everywhere else, except at the output, where I in the end used a Roederstein 4700 uF/63V type.
- • An input C4 of 1.3 uF, 100 Volt, of American make (I once got 25 of these capacitors from a friend of mine), polyester type, very musical.

First results with the JLH-69

My first JLH worked right from the start, no hum whatsoever, no hiss, no problems, just music. Before I forget: I introduced one 'new' thing into the design. I put a 0.33 ohm/5 Watt resistor in the line from the collector of Tr2 to the Vc. I did this to have an easy way of measuring the idle current through the power resistors. Later, when everything worked properly, I could easily bypass this resistor with a few inches of thick wire. Actually, I never bothered to bypass this resistor, so I always listened with this power resistor in place. I fine-tuned the amplifier by trying at least a dozen output capacitors. Every capacitor sounded different and the one that really made the amplifier sing was the Roederstein (other capacitors made the amplifier sound more 'mundane', more clinical, more sterile). I changed the transformer a few times for other types, and here also differences in sound quality could clearly be heard.

From the beginning it was clear to me that this was a very special amplifier indeed. In some respects it gave me more pleasure than did the Hiraga. This amplifier did not sound like a transistor amplifier at all, it sounded so round and full and gave an immense depth into the music. Amazing. The amplifier gave a somewhat coloured overall view I guess, like many fine-sounding tube-amplifiers of the 1960s. I missed some speed and texture in the high frequencies. The low frequencies came out too dark and not so 'quick' and up-tempo. Think of the loudness controls of yesteryear.

Good but not perfect

My JLH did not sound perfect. The Hiraga was a more neutral and more transparent performer. I started to wonder how the amplifier would sound without the output capacitor, because this was the most obvious candidate for the few shortcomings of the JLH. I wondered how the 1996 version would sound to my ears. In the article the author did not take a lot of trouble to persuade the builder to go for the newer version. Actually, he did not hear any significant differences between the '69 and '96 versions, good is good. I acquired the newer version of the JLH on loan from a friend. As I expected, it did not sound as good as my old version. My friend had built the amplifier from a Hart Electronics kit. Maybe (now this is what I think) there was still one remaining error in the design, as pointed out by Geoff on this site (the incorrect connection of the feedback capacitor to the –ve supply rail). Indeed a fairly loud hum was clearly audible.

So in 1999 I started to think about building the old JLH anew, with a symmetrical power supply. Why change something more than is strictly necessary? If the old design could sound that good, there clearly is not much wrong with the design itself. Therefore, in building my second JLH, I insisted on using exactly the same components where possible. Refer to figure 2 in the 'Design Notes' article for the circuit used. When I first switched on this prototype (now it really was a prototype, I did not know if and how it would work) there was a terrible hum. More seriously, after switch-on I got more than 10 Volts DC for a few moments on my speaker. I've got myself a big DC offset problem here. After consulting several friends at the time, no one could see a solution, so away went the thing into a box.

When Geoff Moss entered the game

Then a few months back I discovered the DIYaudio site and I posted my problem of 2 years back on the forum. Geoff is a regular visitor to the forum and within 24 hours there was a solution to my problem: insert one resistor and the hum should be gone. Geoff gave me some extra advice, remove the 0.33 ohm resistor and change the polarity of the feedback capacitor.

Ok, now for the first time I had a properly working amplifier, new-old style. The DC offset swing was minimized within reasonable margins (it stays within 0.5 Volts during one second, you can hear the woofer give a very soft thud sound). First conclusion: clarity and focus and speed are much better than the old JLH. Very good.

The 1969 dual-supply JLH: an evaluation of its sound

After a few weeks (and comparing it many times with the Hiraga), I came to the conclusion that the sound missed some of the old warmth and body. Just a logical consequence or could this be 'solved'? Now a painstaking inspection of all components and several experiments followed.

As mentioned before, the Geoff’s advice was also to cut out the 0.33 ohm power resistor. This sounded logical, it had no use after all. It took me several weeks to find out that the sound of the amplifier changed just because of this removal of the resistor. We both have no clue why this is so. Geoff made several computer simulations and concluded that it did not seem to matter whether the resistor was in or out (there was virtually no difference in the simulated distortion figures, the square wave performance or the bandwidth). Okay, then I prefer having it in, and my advice to you builders is to try it for yourself. Without this resistor, the newer design sounded more clinical than my old version. The transparency was all right though. With the resistor in place the sound becomes more relaxed, everything is smoother than before. Now, when I place the original Roederstein capacitors between the amp and the speakers (just for an experiment) I have my original sound back (and I checked it by making comparisons between the two amplifiers). Taking away or bypassing the Roederstein (do this only with the newer version!!) shows clear gains in transparency and detail richness. No doubt about that. The output capacitor introduces some colouration of the sound, it is a very nice and pleasant kind of distortion that is introduced here though, not bad at all.

Furthermore there was a slight gain in quality to be had by bringing the bootstrap and feedback capacitors (referring to figure 2 on the Design Notes page, C2 and C3) to a higher value of 470 mF/63V (this is a departure from my old JLH).

Now 4 years after first meeting the JLH I have a version that really shines. I think it flows more organically than the Hiraga, but the Hiraga still has the edge in a number of other areas. I dare not say which amplifier is the best in the long term. Both amplifiers stay way ahead of any transistor competition that I know of.

How does the JLH compare to a good tube amplifier?

As a postscript, let me say something about the statement that the JLH equals a good tube amplifier (and, as a bonus, for a tenth of the price of a good tube amplifier). Well, this certainly is not so. I have made many tube amplifiers in my life that, without any doubt on my side, bettered both the JLH and the Hiraga in almost all respects (of course these are subjective statements, it can not be demonstrated by hard figures). Also, and meant as a warning, it should be clear that I have also heard a good number of tube amplifiers that could not match the quality of either the JLH or the Hiraga. Recently I built a very simple EL84 push-pull amplifier (see www.hifi.nl) using old transformers from a Bocama/Lafayette LA-224B amplifier of the 1960s. I used no overall feedback, set the EL84's in triode-mode and used a paraphase phase-splitter. This baby tube amp didn’t cost me as much as the JLH (in terms of components). And sorry folks: everyone in my place prefers the tube amp. You really have to bring in a normal commercial transistor amplifier to be able to hear the special qualities of the JLH again.

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HISTORY: Page created 19/08/2001

15-20W Class AB Audio Amplifier

A design with class-A performance but reduced thermal dissipation

by J. L. Linsley Hood
(Wireless World, June/July 1970)

I have included this article because, in a way, it is a follow-up to the original JLH Class-A design. The topology is very similar to the Class-A circuit with the exception that the output stage operates in push-pull, therefore the amplifier can continue to deliver the necessary power (within the limits of the power supply and output transistors) when the load requires currents above the Class-A bias level. This circuit may be of interest to those with a limited size of heatsink or who are looking for a simple Class-AB design that has proven subjective qualities.

There are plenty of Class-AB designs on the Web that abound with differential input stages, constant current sources, current mirrors, cascoding, Darlington/compound pair output stages etc. etc., but I have seen very few, if any, that offer a simple, current feedback circuit of proven ability.

The article gives an insight into JLH's thought process when designing amplifiers and further confirms that his designs are based on both subjective listening and objective measurement (contrary to some of the suggestions that I have seen).

The amplifier can be biased into full Class-A operation and, as with the 1969 Class-A amplifier, this circuit could be modified to operate from dual supply rails.

Class Distinction in Audio Amplifiers

A discussion of design problems and how to overcome them

by J. L. Linsley Hood *(1)*
(Wireless World, June 1970)

Since the publication of "Simple Class A Amplifier" the author has received numerous letters asking whether it would be feasible to increase the power output to 15W, or even 20W, to provide a greater reserve for use with inefficient loudspeaker systems.

Whilst it would be possible, the problems associated with increased heat dissipation and the provision of suitable power supplies makes this unattractive. In view of the low average power required for normal listening, the question inevitably arose whether it would be practicable to design an output stage which would operate in class A with an inherently low level of high order distortion up to a watt or two, but progress further into class B operation if and when higher powers were momentarily demanded.

There are, unfortunately, a number of snags with the class B operation of transistor output stages, to which the answers are not fully known.

It was pointed out some years ago, by Bailey (2) and others, that the use of quasi complementary symmetry in such output stages led to an increase in high-order harmonic distortion, associated with the non-linearities in the crossover characteristics at low volume levels, and although the level of total harmonic distortion at maximum power output could be quite low, the distortion content at typical listening levels could be many times greater than this, and would also be of an audibly objectionable type.

A number of schemes have been proposed to overcome this problem, including the use of full complementary symmetry (2 3 4), and various methods of ensuring that there are an equivalent number of forward biased junctions in each limb have been described (5 6), including the ingenious semi-complementary triples arrangement used in the "Quad" amplifier (7).

However, in the author's experience, some class B transistor amplifiers - including those employing full symmetry, which is presumed to eliminate the major fundamental snags of this type of operation - having an impeccable performance on paper, did not have the tonal quality which had been expected. Since harmonic distortion at both high and low power levels had been found to be well below the level at which audible effects might reasonably be expected in some of the designs tested, it seemed more probable that the audible ill-effects were due either to transient instabilities associated with loudspeaker loads -perhaps related to changes in the reactance of the base-emitter junction at the current cut-off point - or to high-frequency crossover type distortion arising from hole-storage effects. Hole-storage depends on the presence of holes produced when current flows in a semiconductor - even though the current is due to majority carriers (electron flow). The greater the current the greater the number of holes and the worse the problems of hole storage.

Hole-storage phenomena

The expected result of hole storage in the base region of a transistor, following the attempted termination of a high emitter collector current, is that the transistor remains in a conducting state after the forward base bias has been removed. This has the effect, amongst other things, that the normal crossover discontinuity shown in Fig. 1(a) becomes displaced from the mid-point of the transfer waveform as the frequency is increased, as shown in Fig. 1(b).

These waveforms were generated in a simple complementary pair emitter-follower circuit, without additional negative feedback, driving a resistive load. (In order to assist its display the crossover effect was deliberately exaggerated by the use of an inadequate quiescent current.) Provided that the peak currents flowing through the transistors are small, this effect is innocuous. However, if the peak currents are increased, by reducing the load resistance, the crossover waveform rapidly deteriorates as shown in Fig. 1(c), and increasing the forward bias to give a more suitable quiescent current has little effect in removing this prominent notch, until the forward bias is almost equivalent to that of class A operation.

It is known from experience that these effects can be minimized by the use of transistors with good high-frequency characteristics and low-impedance base-emitter return paths. A low-impedance driver stage will also be effective provided that it does not become cut off (as in the case of the Darlington pair) when the input signal reverses polarity.

The effect of reducing the driver circuit impedance from 2000ohm to 100ohm is shown in Fig. 1(d).

The lack of effective symmetry between the upper n-p-n device and the lower p-n-p is also shown in Fig. 1(c). This effective asymmetry is reduced if the source impedance is reduced.

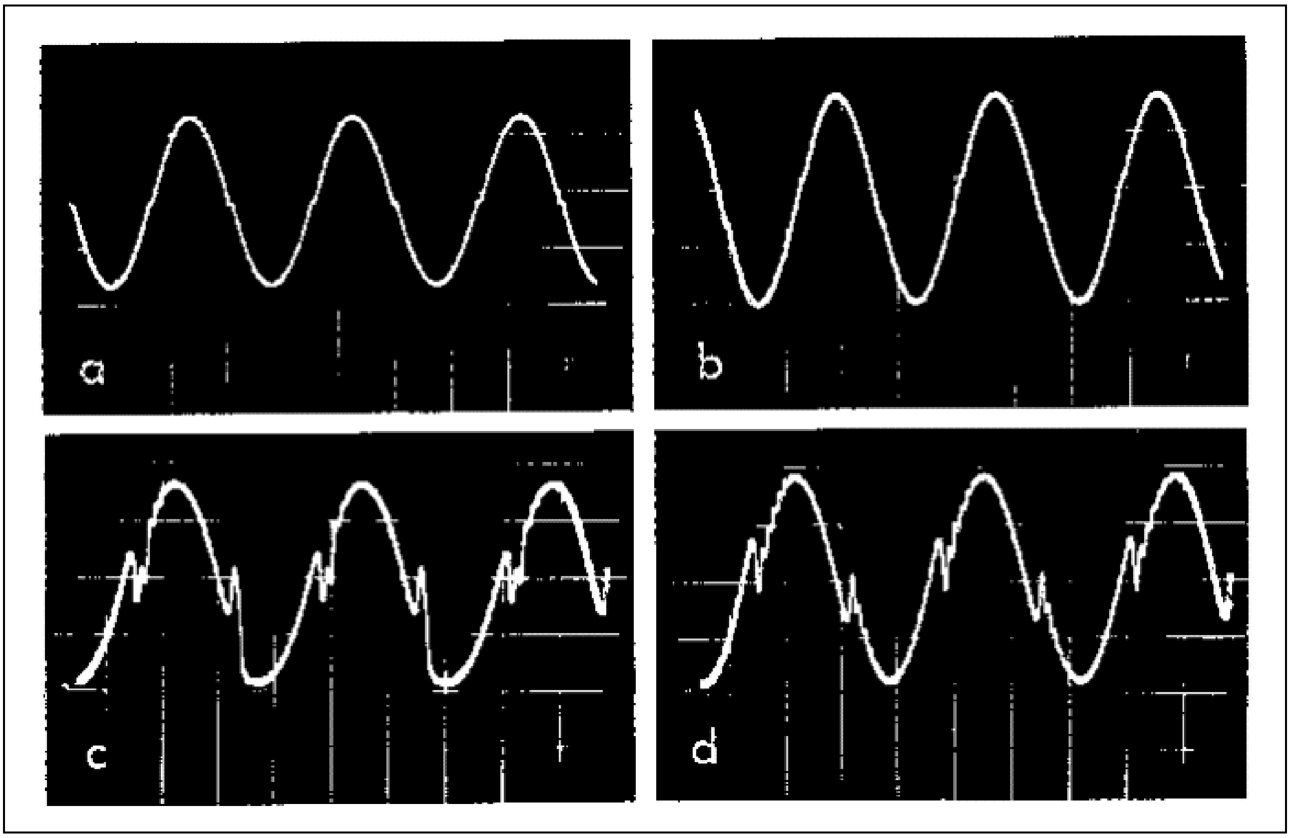

Fig. 1. Crossover distortion in a class B stage employing transistors with an f_T of about 2MHz. (a) Low frequency sine wave at 10mA. (b) High frequency sine wave showing the effect of hole storage on the crossover discontinuity under light load conditions. (c) Influence of hole storage and n-p-n/p-n-p asymmetry under high current conditions at 200kHz. (d) Improvement of conditions in (c) by reducing source impedance.

It was noted that this effect did not become apparent, even under high emitter current conditions, until the operating frequency approached 0.05 f_T. At 0.1 f_T, the problem was severe and this argues that the occurrence of high transient currents - which may arise with certain loudspeaker systems - and high driver stage output impedances, is most undesirable unless the highest frequency components of the waveform are low in relation to the transition frequency of the output transistors. With the availability of power transistors having transition frequencies of the order of 4MHz (such as the MJ480/490 series) it is unlikely that hole-storage phenomena will be troublesome at the rates-of-change of signal voltage likely to be encountered in audio amplifier practice so long as the driver stage does not leave the output transistor base open-circuited on cut-off. However, the use of a driver output, or base circuit, impedance not in excess of a few hundred ohms appears prudent. With earlier designs using germanium diffused junction power output transistors, which usually have very poor h.f. performance, this problem could be important, and Dinsdale has referred to a "subjective audible improvement" resulting from the replacement of low transition frequency output transistors with types having better h.f. characteristics.

Transient instabilities on loudspeaker loads

Phase-angle measurements made with a variable frequency sine wave input, from a high impedance source, reveal that even a simple single-unit loudspeaker can present quite complex characteristics. The reactance - which is normally inductive - changes rapidly, and sometimes even becomes capacitive, at frequencies in proximity to cone and structure resonances.

In general, the characteristics of most of the common designs of transistor power amplifiers are such that instability problems do not arise with inductive loads, and the inclusion of a small choke, of a few microhenries inductance, in the speaker output lead is a well known technique for avoiding instabilities under adverse load conditions. However, capacitive loads can frequently impair the stability margins of the feedback loop, and it is in this respect that

the reactive characteristics of the loudspeaker load are most significant Since it was suspected that the region of the output waveform where this might arise most readily was that at which the output transistors were being driven from the conducting to the cut-off state, an input waveform which provided a transient of controllable steepness (by varying the input amplitude), but arrested at the mid-point, was provided by the circuit of Fig. 2.

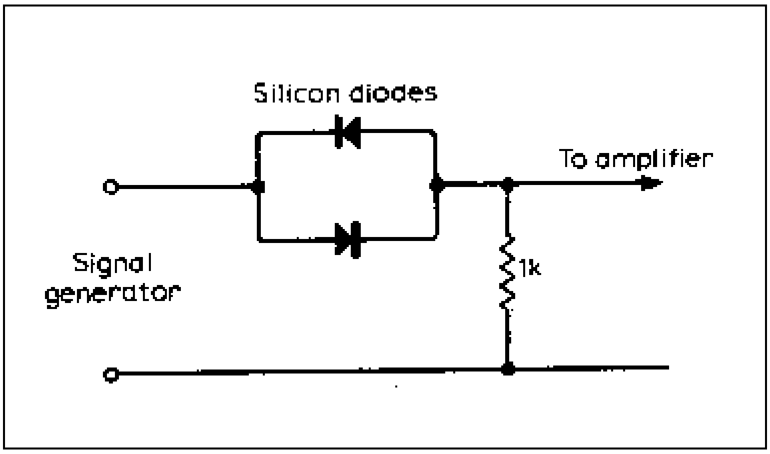

Fig. 2. Circuit for generating the test waveform shown in Fig.3.

The waveform generated by this device is shown in Fig. 3 and the result of introducing such a waveform into an amplifier of poor stability margins, coupled to a resistive load shunted by an appropriate value of capacitance is shown in Fig. 4(a). (The broadening of the oscilloscope trace in the horizontal regions at the mid-point of the waveform was due to inadequately recorded h.s. oscillation.)

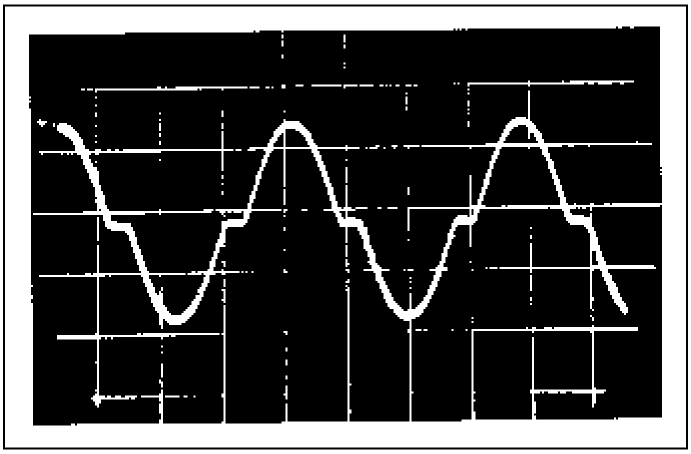

Fig. 3. Test waveform for providing arrested transient input.

The output waveform obtainable from a design with better stability margins and improved bandwidth is shown in Fig. 4(b). In both cases the magnitude of the input signal was adjusted so that clipping occurred on both negative- and positive-going peaks.

Since the h.f. instability shown in Fig. 4(a) - which did not occur in the absence of a large input signal, and which required a particular range of shunt capacitance to provoke it at all - also occurred on parts of the waveform preceding the arrested transient, it was concluded that the change in reactance of the base-emitter junction at cut-off or switch-on, was not a major cause of the transient induced instability observed in this particular design.

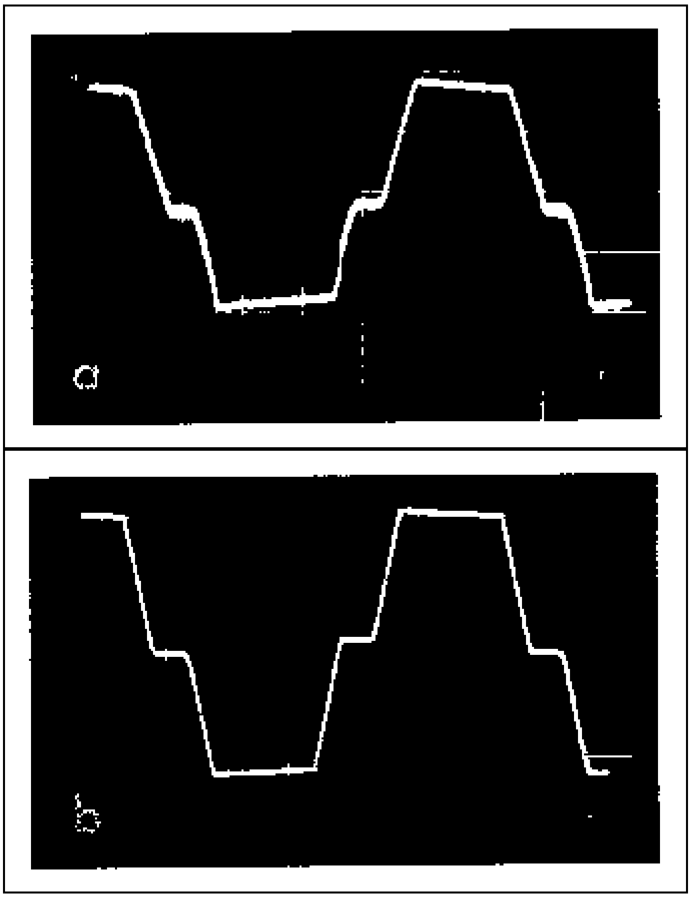

Fig. 4. Amplifier performance using 10kHz test waveform. (a) Response of amplifier showing inadequate stability with reactive load. (b) Response of improved amplifier with reactive load.

Square-wave performance and tonal quality

In view of the fact that a loudspeaker system can present a reactive load, of a type which is found in certain circumstances to cause signal induced instability, and since this instability could be provoked by a square-wave input into an amplifier with a suitable reactive load, a series of tests and comparative listening trials was conducted to determine whether there was any audible relationship between the two. In the event, it was found, beyond doubt, that an amplifier system which did not show any sign of instability over the range of load shunt capacitances up to, say, 0.33uF had a better tonal quality on even a simple loudspeaker system than one in which some shunt capacitor value could cause h.f. oscillation. Moreover, in a more complex loudspeaker system, with a crossover network and high-frequency capacitively coupled "tweeter", it was possible to hear the difference between systems which would, in the lab., with some RC load combination, give a square-wave response such as that of Fig. 5(a) and those which had a response like that shown in Fig. 5(b). No positive distinction could be drawn in listening trials between a system giving a waveform such as Fig. 5(b) and one in which a square-wave input could produce a single overshoot "spike".

Since the frequency of the "ring" waveform in Fig. 5(a) is well beyond the upper limits of the audible spectrum, it is clear that it is not this of itself which produces the undesired sound quality, but rather that this type of behaviour is symptomatic of a different and more objectionable effect when the amplifier is used with a loudspeaker load.

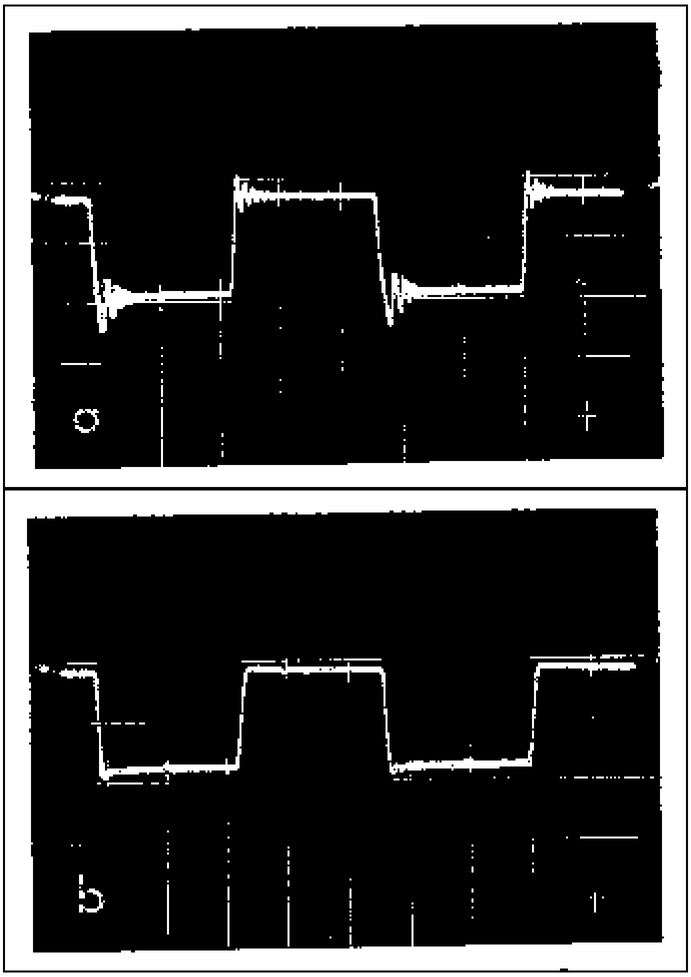

Fig. 5. Amplifier response driving a reactive load (15ohm, 0.47uF) with a 10kHz square wave. (a) The ringing gives evidence of instability. (b) No transient ring indicates better stability.

The conclusions which have been drawn from this series of experiments are these: (1) that it is desirable to employ output power transistors in which the transition frequency is at least ten times higher than the highest signal frequency component which is passed to the amplifier from preceding stages; (2) that it is preferable to drive the output transistors from a source which has a low impedance over the whole signal voltage swing, or at least to provide a reasonably low-resistance base-emitter current path; and (3) that the phase/frequency characteristics of the feedback loop should be such that a square-wave output devoid of overshoots is obtained when the amplifier is bench tested with a wide range of shunt capacitance values in an RC dummy load. This latter requirement probably implies either a fairly limited number of stages within the feedback loop or a relatively restricted h.f. bandwidth.

When these requirements had been met, and when the harmonic distortion levels over the range 40mW up to the maximum rated power output were of a suitably low level, there was no audible difference, in the most careful listening trials, between several different designs. However, it is difficult in class B systems to obtain the desired low level of harmonic distortion at low signal levels without the use of substantial amounts of negative feedback, and this leads to a worsening of the amplifier response to signals containing transients.

The use of a class AB system, if the problems in maintaining the correct forward bias level can be solved satisfactorily, should facilitate the attainment of these desired standards, particularly if the h.f. negative-feedback loop can be made fairly simple.

Next month full details will be given of a 15-20W class AB amplifier with the following characteristics:-

Power output: 15W into 15ohm, or 18W into 8ohm (20W with modified output circuit component values.)
Bandwidth: 10Hz-100kHz +/- 0.5dB at 2V output; 20Hz-50kHz +/- 1.0dB at maximum power output.
Output impedance: 0.03ohm (at 1kHz).
Total harmonic distortion: 0.02% at 15W/15ohm or 18W/8ohm; less than 0.02% at all power levels below maximum output.
Intermodulatlon distortion: Less than 0.1% at 10W (12.3V r.m.s. into 15ohm) and 70Hz, and at 1V r.m.s. at 10kHz.
Square-wave transfer distortion: Less than 0.2% at 10kHz.

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15-20W Class AB Audio Amplifier

A design with class-A performance but reduced thermal dissipation

by J. L. Linsley Hood
(Wireless World, July 1970)

Many class B designs can be operated in class A at low power levels if the quiescent current is increased. However, this often worsens the distortion characteristics of the output stage, particularly at intermediate (and audibly important) power levels, by displacing the crossover point to a region where the transfer slope is much steeper, and the crossover discontinuity therefore much more prominent. This effect is considerably accentuated by the fact that almost all modern transformerless power amplifier systems use either Darlington pair or augmented (p-n-p/n-p-n) emitter follower output pair configurations, and these have a very high mutual conductance.

The use of a complementary pair of emitter followers, driven from a voltage source having an output impedance which is very much lower than the normal input impedance of the output devices, appeared from this line of thought to offer the best way of minimizing the several problems mentioned above.

In practice, the necessary low impedance base-emitter paths can be arranged quite simply by driving the output transistors from a suitably tapped emitter load resistor in a conventional emitter-follower circuit, provided that the current flow in this load circuit is adequate to deliver the necessary output drive.

Moreover, this type of circuit arrangement will also operate, in class A, as a straightforward cascaded emitter follower, as can be seen from the circuit arrangements shown in Fig. 1. In (a), the transistors Tr1 and Tr2 act as a conventional Darlington pair, with a resistive emitter load to which the output load ZL is coupled through C1. In (b), essentially the same circuit is employed, but using a complementary type of transistor as the second stage emitter follower.

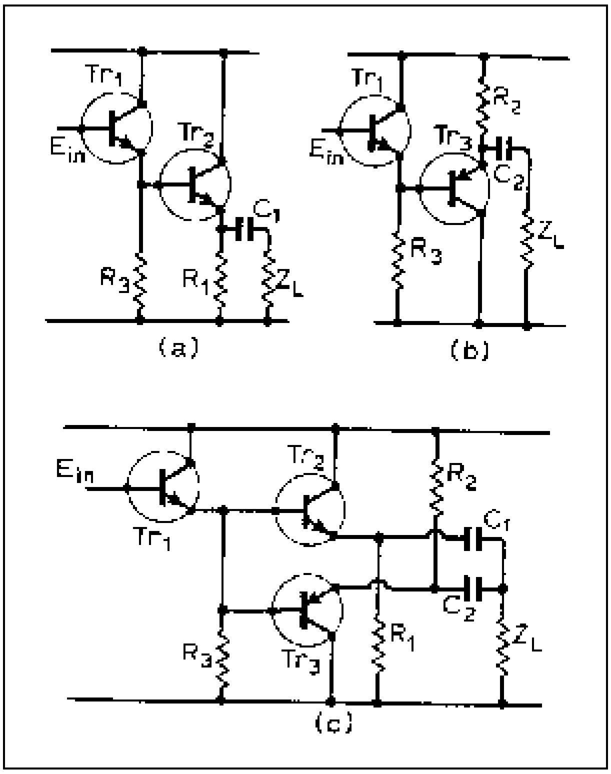


Fig. 1. Emitter follower Configurations for class A operation

It is then possible to arrange the circuit as shown in (c), so that both of these configurations are employed simultaneously. Resistors of double the ohmic value can then be employed as R1 and R2, with half the emitter current in each transistor, to give an identical matching impedance to the output load. In practice, this circuit arrangement can be simplified into the form shown in Fig. 2, and the resistors R1 and R2 deleted since the load current for each transistor can flow through the other. This also improves the efficiency since the transistors have a very high dynamic impedance and form good emitter loads for each other. The two small value resistors Rx and Ry are included to assist in stabilizing the output transistor working points.

The actual value of the quiescent current in the output stage can be set by adjustment to VR1. To avoid asymmetry, at low audio frequencies, the bypass capacitor should have as high a value as convenient.

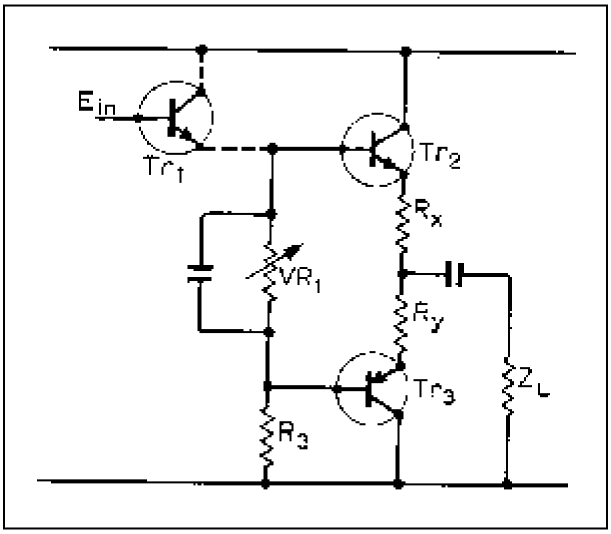
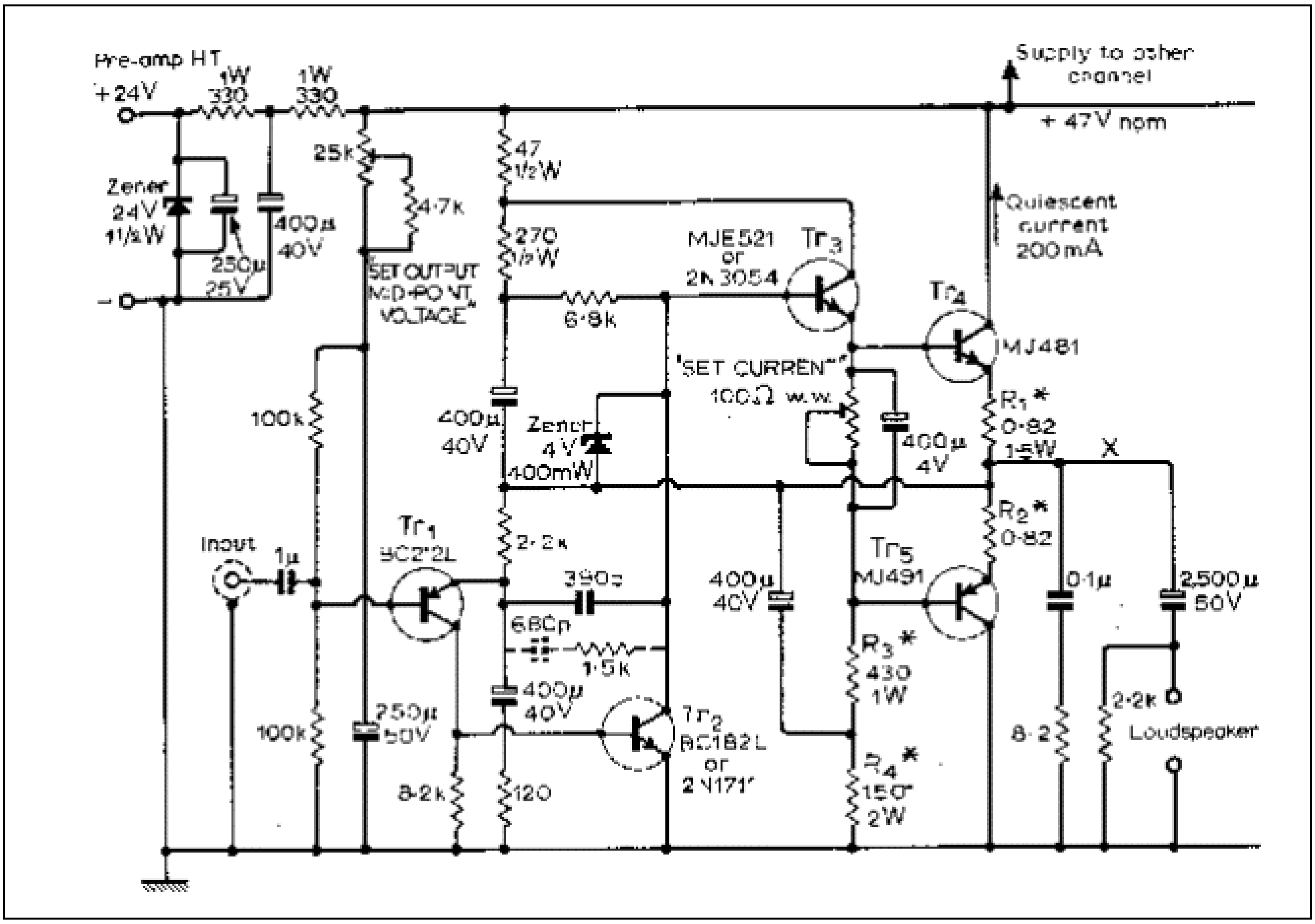


Fig. 2. Simplification of Fig. 1(c).

This arrangement of the output transistors was of particular interest to the author, since the first three stages of such an amplifier could be substantially the same as those used in the previously described class A design, of which the performance was known. In fact, the system could be constructed on the basis of the class A design, with the quiescent current reduced to a much lower level, and a pair of suitably biased back-to-back emitter followers interposed between the output and the loudspeaker load. However, this would not have made the most of such a system. In particular, it will be noted that if the potential at the emitter (or base) of Tr1 in Fig. 2 is held constant, the current through the resistor chain R3, VR1 will be constant for any particular value of VR1 and therefore the turn-on potential applied between the bases of Tr2 and Tr3 will also remain constant (or virtually so). This allows the standing current of the output transistors to be defined precisely, since the d.c. output potential can be controlled by the use of unity gain d.c. negative feedback, and this effectively controls the emitter potential of Tr3
.

Also, since the last voltage amplifier stage is not required to deliver significant power, it can be optimised for voltage gain, with an increase in the available negative feedback. A practical amplifier circuit of this type is shown in Fig. 3.



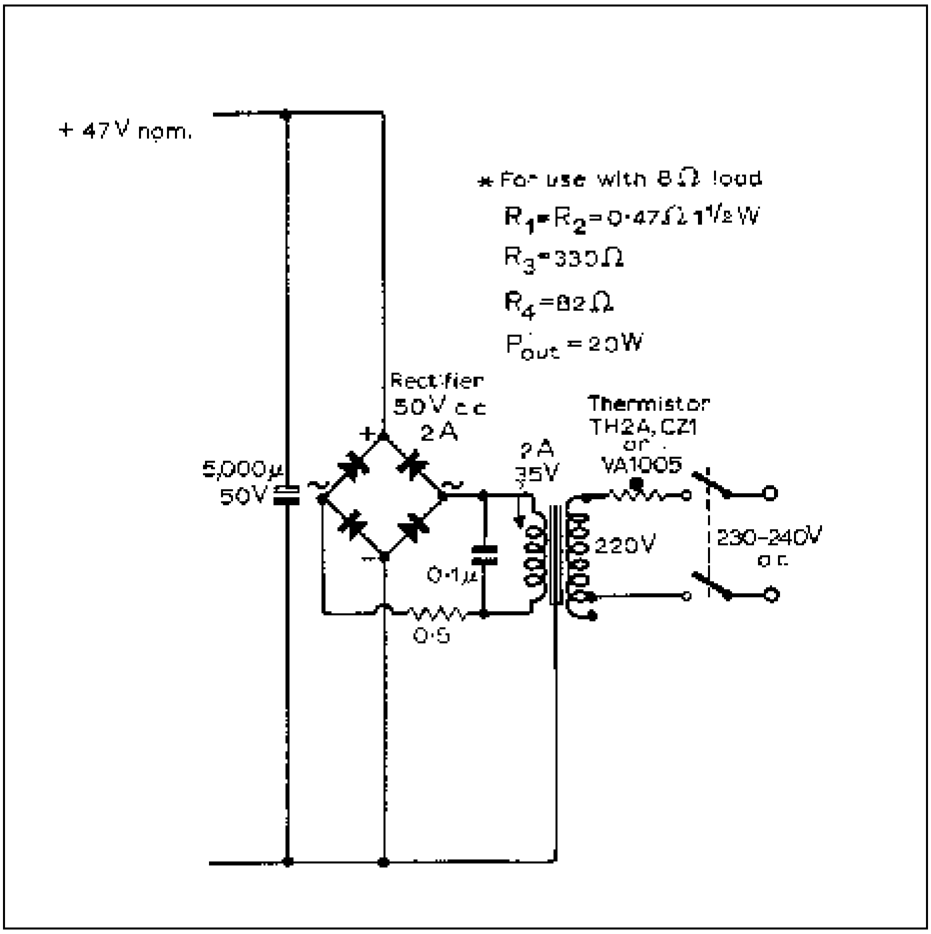


Fig. 3. Power amplifier circuit. The dotted components (680pF, 1.5kohm) can be added if electrostatic speakers are used

The first two transistor voltage amplifier stages of this follow conventional design practice, with the collector load resistor of Tr2 boot-strapped to obtain large voltage swing at the base of Tr3 with as little second harmonic distortion as practicable. The collector of Tr3 is also partially boot-strapped in order to reduce the peak voltage swing, and improve the symmetry of the output waveform prior to the application of the loop negative feedback. (Without overall n.f.b. the distortion at full output power is a little less than 4%, almost entirely second harmonic. This is similar to the performance of a good triode valve output stage prior to the application of n.f.b.) The lower end of R3 is also fed with the output signal to improve the output voltage swing obtainable from Tr5.

The 390pF capacitor between the emitter of Tr1 and the collector of Tr2, and the 8.2ohm resistor in series with the 0.1uF capacitor across the output, provide the necessary phase-angle correction and define the high-frequency gain of the feedback loop. With the values shown there is a 6 dB/octave roll off beyond 100kHz, and the system is completely stable under all load conditions. However, with the use of a large value capacitive load there will be some overshoot on a rapid transient. The author believes that it is desirable for tonal purity, for such overshoots to be eliminated, and it is recommended, therefore, that the 390pF capacitor be shunted with a 680pF 1.5kohm combination where it is intended to drive electrostatic speaker systems. However, on normal loads this merely reduces the h.f. roll-off point, and the power output available in the 30-50kHz region, and can well be omitted.

The 100ohm wire-wound potentiometer between the bases of Tr4 and Tr5 is used to set the quiescent current level to about 200mA. The chosen current level determines the power level at which the system changes from class A to class B operation. With the suggested level of 200mA, this transfer will occur at approximately 1.2 W with a 15ohm speaker (640mW for 8ohm).

If the standing current through the output stage is increased, progressively larger output power levels can be obtained within the class A region, up to the level at which the amplifier acts as a pure class A system. The only observed penalty for this exercise is that the power supply demand and the thermal dissipation in the output transistors are both proportionately increased. However, if the output transistors are of dissimilar origin or are otherwise badly paired the operation of the circuit in class A will ensure that the distortion levels and other performance standards are attained in spite of this.

Performance characteristics

The specifications given below were obtained using the power supply system shown in Fig. 3. The amplifier was specifically designed to work from a poorly smoothed h.t. line, the values and positions of the h.t. decoupling and 'bootstrap' capacitors being chosen to avoid the intrusion of ripple into the signal circuits. The only significant difference observed in using a good quality stabilized and smoothed power supply is a small improvement in the already extremely good hum and noise levels.

Power output. 15W into 15 ohm, or 18W into 8ohm (20W with modified output circuit components values).
Bandwidth. 10Hz-l00kHz +/- 0.5dB at 2V output. 20Hz-50kHz +/- 0.5dB at maximum power output.
Output impedance. 0.03ohm (at 1kHz).
Total harmonic distortion. 0.02% at 15W/15ohm or 18W/8ohm; less than 0.02% at all power levels less than maximum output.
Intermodulation distortion. Less than 0.1%. l0W (12.3Vr.m.s.) l5 ohm, 70Hz. 1V r.m.s. 7kHz (or 10kHz).
Square-wave transfer distortion. Less than 0.2W at 10kHz.
Rise time. 3us.
Input impedance. 20kohm (approx.).

Gain. 18x.
Hum level. (Simple power supply) -70dB w.r.t. 1W.
Noise level. (Simple power supply) -80dB w.r.t. 1W. (These figures are, respectively, better than –80dB, and -85dB with the regulated power supply).
Feedback factor. 46dB (typical).
Input voltage for max. output. 850mV r.m.s.
Load stability. Unconditional.

For the perfectionist, a suitable design for a regulated d.c. power supply, with re-entrant short-circuit and overload protection is shown in Fig. 10. This gives approximately 10dB improvement in the hum and (r.m.s.-weighted) very low frequency noise.

The gain/frequency, and power output/frequency graphs are shown in Figs. 4 and 5, and the relationship between output power and distortion, and signal frequency and distortion are shown in Figs. 6 and 7. The square wave performance into a 15ohm resistive load, with any value of shunt capacitance up to 0.1uF, at 1kHz, 10kHz, and 50kHz are shown in Fig. 8. The sine wave output at 1kHz, and 15W with a 15ohm resistive load (42.5Vp-p) and the associated harmonic distortion (representing 0.02%) is shown in Fig. 9.

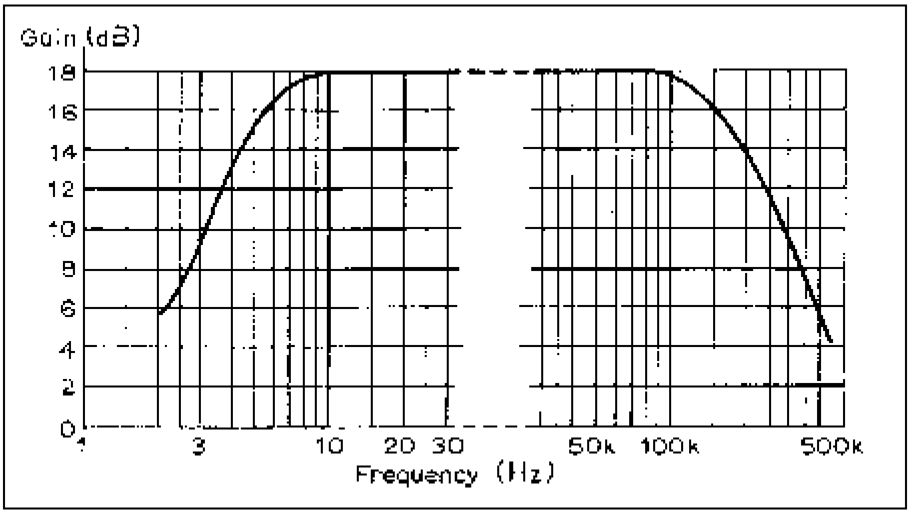


Fig. 4. Gain/frequency characteristics.

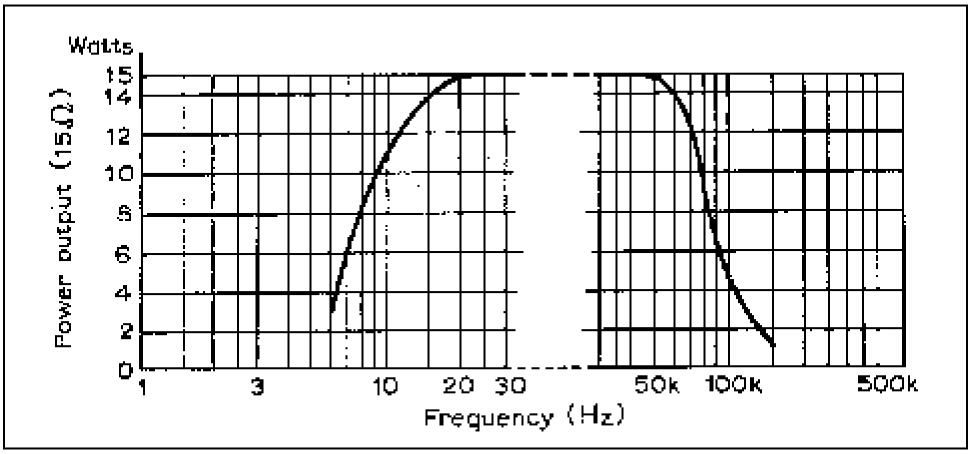


Fig. 5. Power output/frequency characteristics.

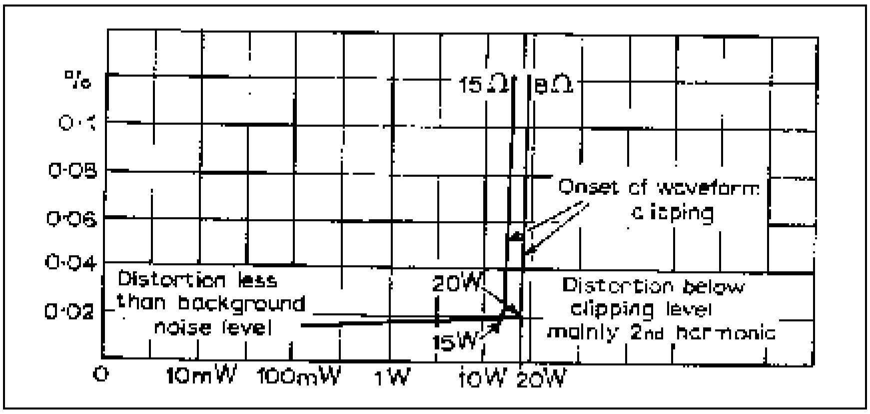


Fig. 6. Power output/distortion characteristics. The 8ohm load characteristic was measured using the modified output-stage components.

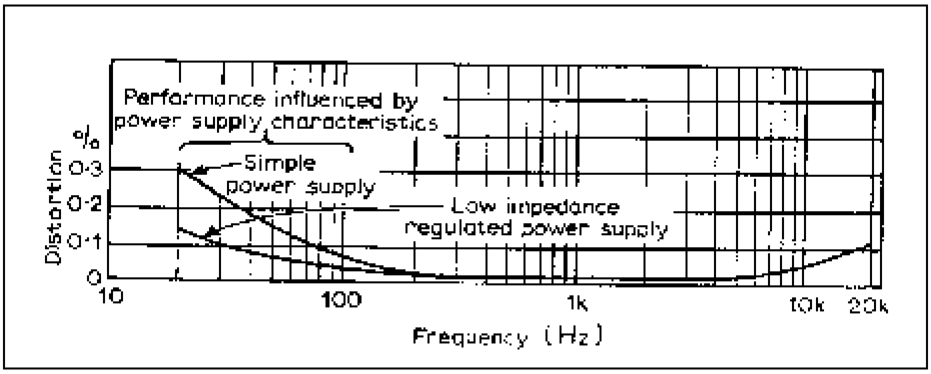


Fig. 7. Influence of signal frequency on distortion (1W into 15ohm).

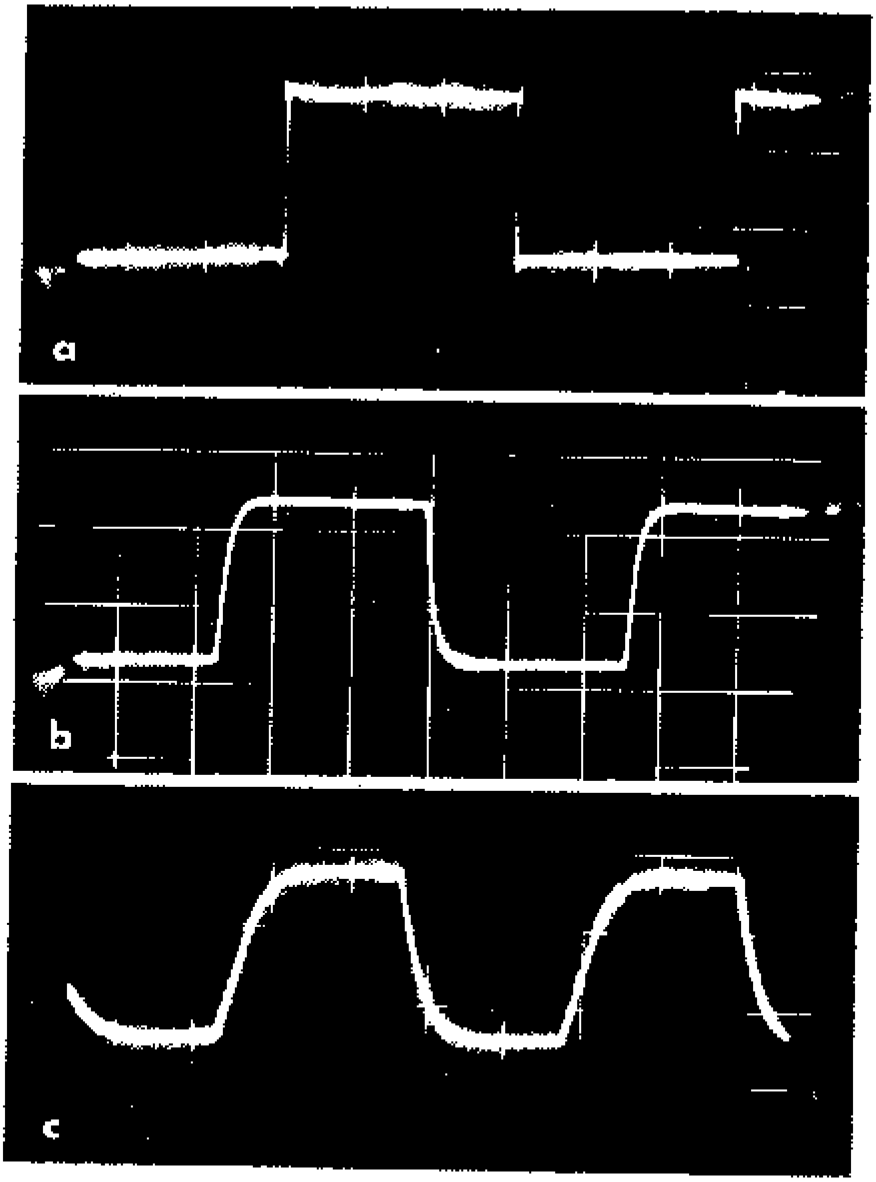


Fig. 8. Square-wave performance into 15Ω in parallel with 0–0.1µF. (Scale 2V/cm) (a) 1kHz, (b) 10kHz, (c) 50kHz.

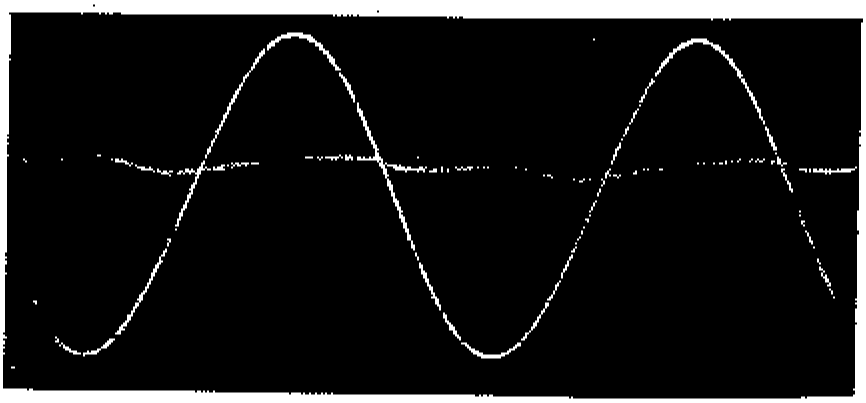


Fig. 9. 14-W 1-kHz sinewave into 15-Ω resistive load. Distortion 0.018% on scale 35mV/cm. Fundamental on scale 10V/cm.

Listening trials

As described last month, a number of experiments were done during the development of this circuit to try to relate audible effects to the phenomena observable and measurable in the laboratory, and a transfer distortion analyser (British patent application No.7925/1970) was made to judge the performance with non-sinusoidal waveforms. (A point was reached in the earlier stages of the design where the author's ear was no longer able to detect the subsequent improvements.)

The transient response of the 10-watt class A design (as originally published(1), without the modifications(2), suggested in October 1969 to reduce the h.f. bandwidth) is superior to that of the present circuit in the range 50kHz-2Mhz under load conditions of fairly low capacitive reactance. Under more adverse load conditions the present design will be (technically) better. However, the most careful comparative listening trials, with several of the author's long-suffering friends, have failed to uncover any audible difference between these two designs, both of which will almost certainly surpass in performance the best available valve-operated, transformer-coupled units.

Constructional points

The layout used in one of the prototypes of this design is shown in Fig. 11, using a 0.15in. matrix copper strip board. The layout should not be particularly critical provided that normal precautions are observed, such as keeping the output and input circuits reasonably well separated, and making sure that the power supply leads, and the loudspeaker return lead, connect to the board at a point close to that to which the collector leads of the output transistors are soldered.

Since the circuit has unity gain at d.c. the occurrence of a switch-on 'plop' in the loudspeaker can be avoided by the use of a suitably long time-constant in the decoupling circuit which provides the base bias for Tr1. The voltage at 'X' (Fig. 3) will then follow the base potential of Tr1 as it slowly rises following switch on. It is undesirable to have the full h.t. voltage applied during this period, and this is avoided by the incorporation of a thermistor (Radiospares TH2A or equivalent) in the mains transformer primary circuit. Since this will cause a drop of some 10-15V, this should be allowed for in the tapping point on the mains transformer. Also, since the thermistor becomes quite hot under operating conditions (this is necessary) it is important to mount it in such a way that this does not damage associated components or wiring.

The dissipation of the output transistors is normally about 8W, and the output pair can both be mounted on a single 3.5in. x 4in. black anodised, ribbed heat sink. The heat sink should be earthed - very simply by omitting the mica washer on the MJ491.

The driver transistor dissipation is of the order of 2W in some circumstances, and this is somewhat in excess of the power which can be handled safely by the normal TO-5 cased device, such as the 2N1613, unless very careful heat sinking arrangements are employed. The use of such devices as the 2N3054 or the Motorola MJE521, mounted on a small piece of black-painted aluminium sheet, say 1in. x 1.5in., gives a very large safety margin in this stage. The performance of the Motorola MJE52 1 is slightly to be preferred, and was used in all the prototypes. This stage, however, is not a very critical one, and these transistor type variations are unlikely to make a significant difference to the system's overall performance.

The Texas BC212L and 182L are the preferred transistor types for Tr1 and Tr2, although the 2N1613 was also used in some development models as Tr2 with identical results. The Motorola 2N3906 and 3904 could also be used in the Tr1, Tr2 positions with almost equivalent performance, but this has not been tried. The use of 0.5W carbon film 5% resistors is suggested except in the points where higher wattages are required. R1 and R2 should be of small diameter or low inductance. The various electrolytic capacitors can be of higher value or voltage working without ill effect.

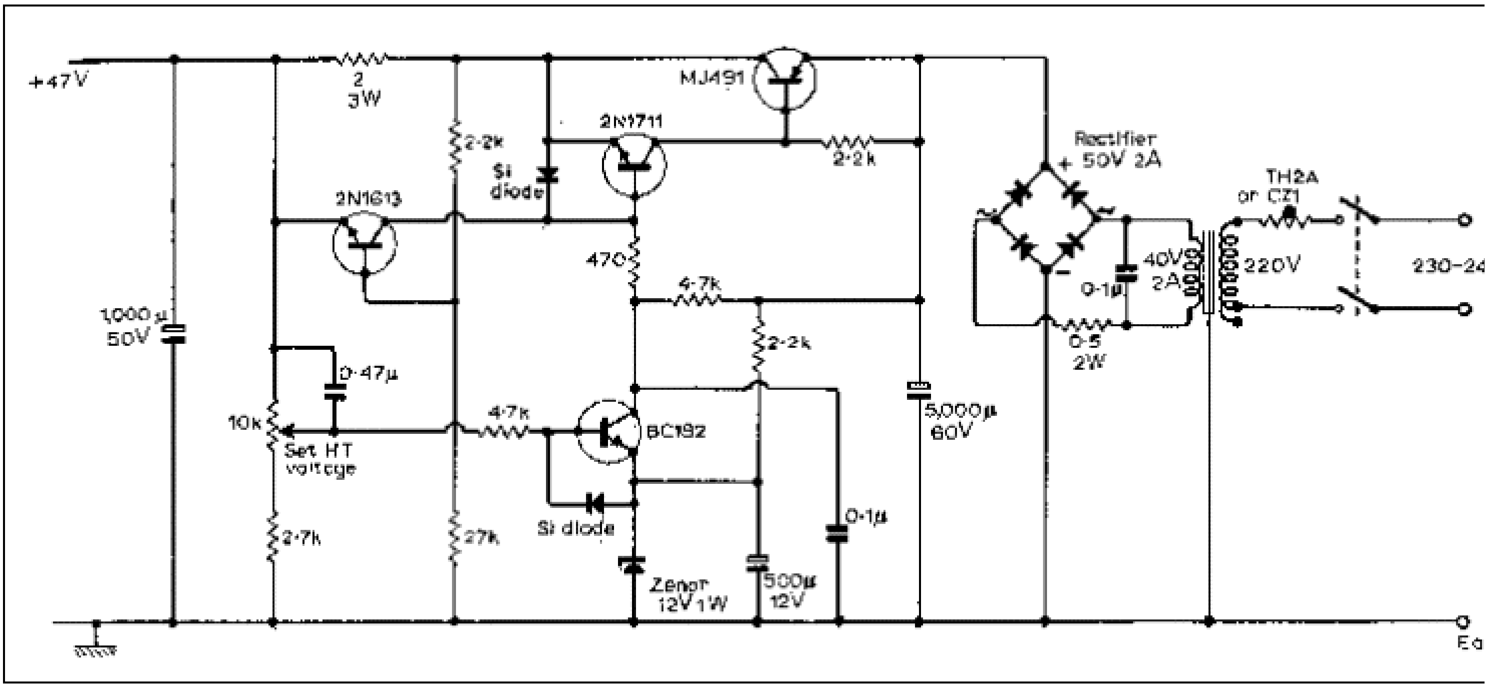


Fig. 10. Stabilized power supply with re-entrant short-circuit protection (12-49V).

Appendix 1

Calculation of power output levels obtainable with given quiescent current in class A operation.

The maximum output power which can be obtained from a power output stage such as that in Fig. 3, in class A, is entirely determined by the quiescent current and the load impedance provided that adequate h.t. voltage is available. At frequencies which are low enough for the 'wattless" components of the load current to be ignored, the maximum current excursion which can be caused to flow through the load without taking one or other of the output transistors beyond cut-off is equal to twice the quiescent current through the output stage. Since this is the 'peak' current through the load, if the waveform is sinusoidal, the r.m.s. equivalent current will be 2 x Iq / √2, and at low frequencies, the power developed in the load will be 2 x Iq^2 x R_L.

For example, if the stage is required to operate in class A up to one watt, with a 15 ohm load, the peak current swing through the load must be 1 = 2 x Iq^2 x 15, or Iq = l83mA. Similarly, for an 8ohm load, Iq = 250mA.

With the standing current suggested (200mA), 1.2 watts or 640mW will be given for 15ohm and 8ohm loads respectively. This should be adequate for most normal listening. For full class A operation up to 15W, quiescent currents of 710mA and 970mA respectively will be required.

Appendix 2

Output transistor protection

The use of class B output circuit configuration (and class AB comes within this category at the power levels concerned) in transistor power amplifiers of this general type leads to the possibility that very high instantaneous currents can flow, which will lead, regrettably, to the equally instantaneous destruction of the transistors involved, if the amplifier is operated at maximum drive into an effective short circuit, and this could be a load with a very high capacitive reactance, in some cases.

The classic system for output transistor protection, using two input bypass transistors, is that due to Bailey(3), and this is also applicable to the output circuit of this design. However, because of the d.c. asymmetry between the potential at the base of Tr3 and the output point 'X', a much simpler arrangement can be used, consisting solely of a good quality (low leakage) zener diode between these two points, with the positive zener end connected to the base of Tr3. Any 4 - 4.7V zener will do provided that the leakage current at 3V reverse, and 0.4V forward, is less than 10uA. The ITT400mW series ZF4.7 is quite suitable. Again, for 20W output into 8ohm, the resistors R1 and R2 must be reduced to 0.47ohm.

REFERENCES

1. J. L. Linsley Hood, "Simple Class-A Amplifier", Wireless World, April 1969.
2. "Letters to the Editor", Wireless World, October 1969.
3. A. R. Bailey, "Output Transistor Protection in A.F. Amplifiers", Wireless World, June 1968.

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HISTORY: Page created 20/07/2001

15-20W Class AB Audio Amplifier

Letters to the Editor of Wireless World

Class AB amplifier (August 1970)

Mr. Linsley Hood is quite correct when he states that the operation of transistor output stages in class AB can cause increased distortion, because of the change in the slope of the transfer characteristic around the crossover point. However, I fear that he is wrong in supposing that a low source impedance overcomes the problem.

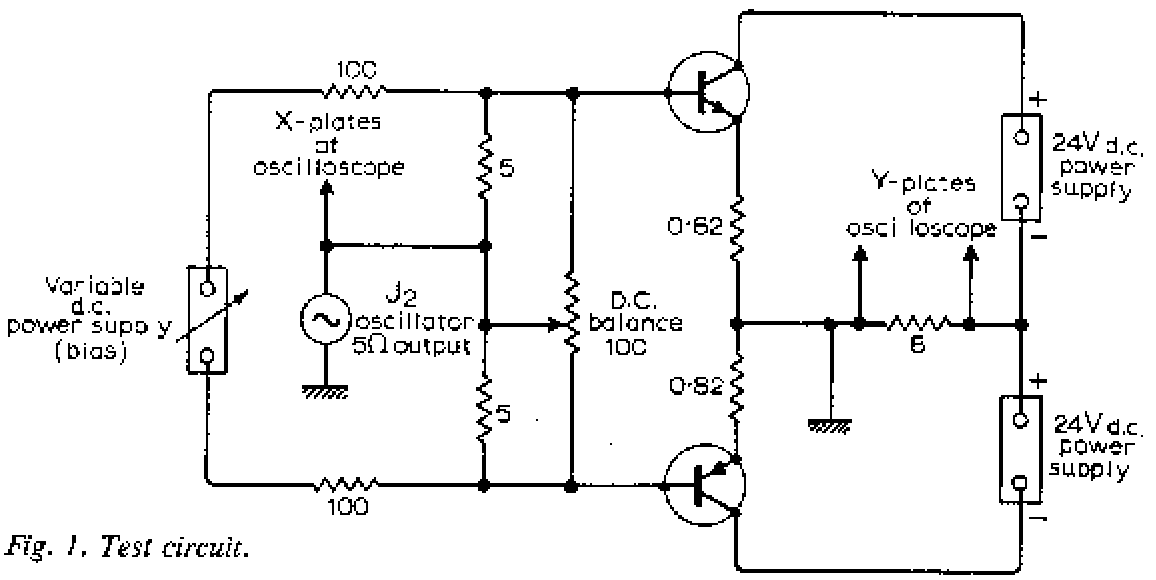


Fig. 1. Test circuit.

Fig. 1 shows a test circuit which I constructed to measure the transfer characteristic of the output stage under various bias conditions and the results are shown in Fig. 2 for 200mA, 20mA and 0mA. Note the prominent change in slope at 200mA bias. In the test circuit the transistors are operated in the common emitter mode to enable the changes in the slope of the transfer characteristic to be seen more easily, but this does not alter the validity of the results since the effect of putting the load into the emitter circuit is only to provide local negative feedback. Under the same conditions a push-pull emitter follower using an output stage with the transfer characteristic of Fig. 2(b) will produce less distortion than a similar output stage with the transfer characteristic of Fig. 2(c).

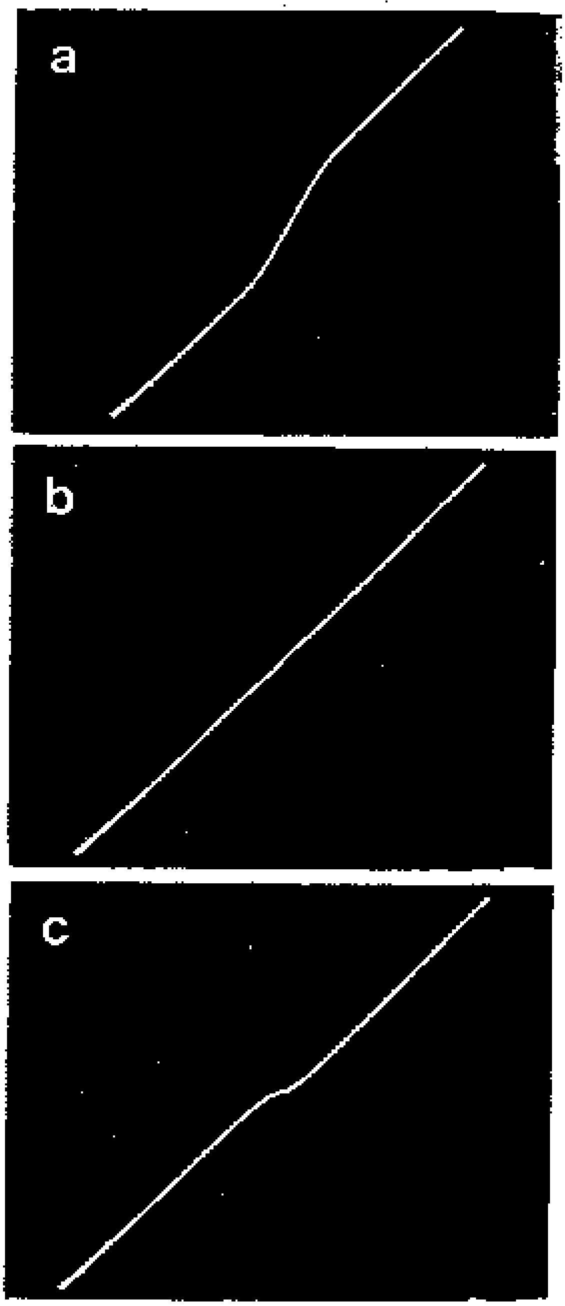


Fig. 2. Transfer characteristics at 200mA bias (a), 20mA (b), and 0mA (c). Vertical scale 500mA/division, horizontal scale 500 mV/division.

To check this I constructed Mr. Linsley Hood's amplifier and measured the distortion at 200mA and 20mA bias current with a Marconi TF2330 wave analyser and TF2100/1M1 low distortion oscillator. The results are shown in Fig. 3 and show clearly the improvement in distortion at intermediate output levels produced by the lower bias current. However, in spite of the excellent results obtained I would not advise constructors of this amplifier to use a bias current as low as 20mA as it tends to be rather unstable. A bias of 50mA would be about the optimum and at this level there would still be a "hump" in the distortion curve but it would be smaller than at 200mA bias and removed to a lower power level. I would also consider the use of a temperature compensating diode or transistor in the bias network strongly advisable, to minimize thermal variations.

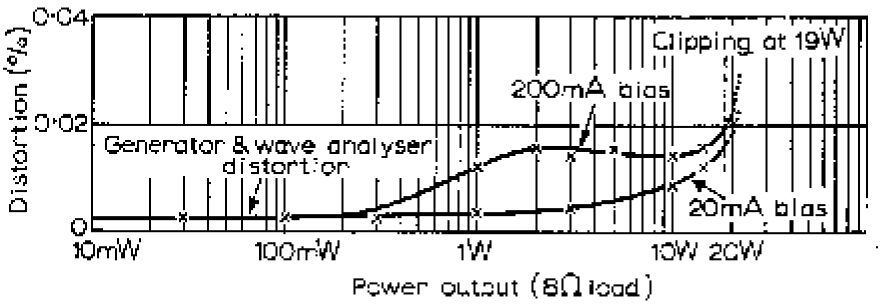


Fig. 3. Distortion Versus output power for bias currents of 200 and 20mA (load 8Ω, frequency 1kHz).

Mr. Linsley Hood is also incorrect when he states that the emitter follower driver Tr3 presents the output transistors with a low source impedance. This would be true if it were not for the bootstrap capacitor which raises the effective value of the 6.8kohm load resistor in Tr2 collector to around 5Okohm. Thus the source impedance seen by the output transistors is about 1kohm, i.e. about twice their input impedance with an 8ohm emitter load.

A further point concerns the current gain of the output transistors. The specified gain spread for the MJ481/MJ491 devices used is 30-200 at 1A. As only 40mA is available from the driver stage the peak collector current with minimum gain devices is only l.2A. This corresponds to an output power of about 8 watts into 15ohm and 5 watts into 8ohm. To achieve the output power claimed by the author the output transistors need to have a minimum current gain of around 80 at 1A. Perhaps the author could suggest alternative component values for those unfortunate enough to get low-gain transistors.

One last point. The author obviously attaches great importance to "square wave transfer distortion" but he has not yet told us how he defines it. It is well known that any network, whether it be active or passive, that does not have a linear phase/ frequency characteristic will produce transient distortion of a square wave. Does the author consider that, for example, an L-C filter with a sharp cut-off at 50kHz would produce audible distortion? The ringing produced by such a filter would be very similar to that produced by an audio amplifier with a load of 15ohm and 2uF.

D. S. GIBBS,
Bury, Lancs.

The author replies:

Mr. Gibbs' letter raises a number of interesting points, with some of which I concur. However, I regret that he has misunderstood the argument in some cases.

To take his points separately.

1. Optimum quiescent current: The fact that there is an optimum value of quiescent current in a class B output stage for minimum harmonic distortion is well known and is not in dispute. This optimum current depends, among other things, on the current gain of the output transistors (or the product of the current gains if a Darlington pair or a similar output stage configuration is used) and, to a first approximation, the higher the effective current gain of the individual halves of the output stage the lower the optimum value of quiescent current. From

the figures Mr. Gibbs quotes it would seem that the transistors he chose for this experiment had a high value of current gain.

However, this is not the point. I believe that the bulk of normal listening is done with output power levels which are of the order of only 50-250mW, only the very occasional transients demanding power levels in the 1-2 watt region. I also believe that it is advantageous for the amplifier to operate in true class A bias conditions for normal listening power levels, in that this avoids most of the ill-effects which can arise in class B, for example due to mismatched output transistor characteristics. These ill-effects produce the bulk of the high order harmonic and intermodulation distortions which appear to be objectionable to the ear.

Therefore, the question is simply which output stage configuration will operate best overall, with a forward bias of say, 200mA (this being chosen to allow class A operation up to 600mW-1.2 watts with 8-15 ohm loads). The simple complementary emitter follower combination appears to be the best one for this purpose.

The measurement of very low order harmonic distortion levels is difficult, and is influenced by such things as h.t. supply impedances, lead connections, etc. and I am grateful therefore to find that Mr. Gibbs' measurements confirm my own findings that such a design, with such an output stage and forward bias does not give rise to harmonic distortion levels in excess of 0.02%. My own subsequent measurements with a harmonic analyser show that the distortion produced in the 'hump' region is mainly 3rd harmonic, whereas the higher magnitude of distortion produced by a more conventional complementary Darlington pair biased to 200mA, in a similar circuit, also contains more of these audibly objectionable higher order harmonics (see my Fig. A). Whether one has 0.015% or 0.005% t.h.d. is probably only of academic interest to the user.

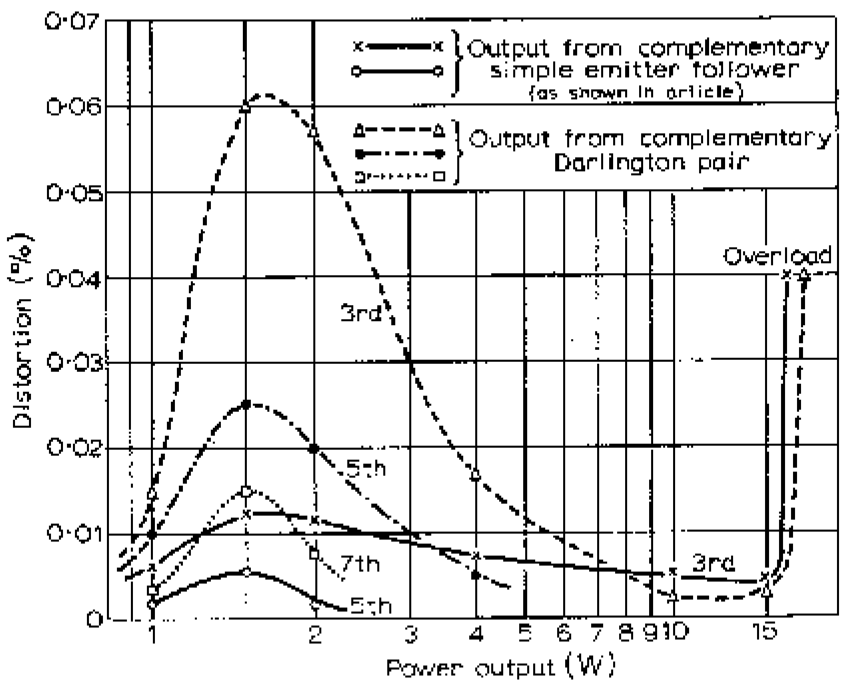


Fig. A. Measurements of Class AB amplifier with 200mA quiescent current and 15 restive load. Second harmonic distortion below 0.01% was similar in both circuits.

2. Base-emitter impedance: For good high-frequency and transient performance it is desirable, I believe, that the impedance between base and emitter of the output transistors should be low. In the case of the class AB amplifier circuit, this condition is met by the 100ohm potentiometer, 400uF combination connected between the bases of the two output stage transistors, since when one of these is cut-off the other is conducting and provides the necessary base-to-emitter return path. The use of a relatively high driver impedance is actually advantageous in minimizing harmonic distortion due to the transistor base impedance non-linearity.

3. Output power: The question of the range of current gains to be found with the M481-491 series transistors has been raised before in different contexts in these columns. My own experience with quite a large number of these is that the lowest current gain encountered, at 1A, is of the order of 75, and most, in fact, lie in the 100-150 bracket. However, this is not really an important limitation under dynamic conditions, because the effect of the bootstrap connection to the emitter load of Tr3 allows adequate drive current even with low-gain transistors.

4. Audible effects of transient overshoots on reactive loads: My experimental findings are that there is an occasional audible difference between an amplifier whose stability under reactive load conditions is such that no overshoots are produced with a transient input and one which 'rings'. I do not think that this has anything to do with the nature of the h.f. response curve although it is evident that a 'ring' can be produced by a steep-cut low-pass filter. In the case of an audio amplifier driving a loudspeaker load, my own hypothesis is that some loudspeaker systems, under some dynamic conditions, can provide a negative reactive impedance, and this, however transitory, can exaggerate incipient reactive load instabilities present in the amplifier, and introduce spurious (and audible) waveform distortions.

I will take this opportunity of adding a personal note. In the original draft of my article, I walked into a philosophical booby-trap on the output power calculations, through overlooking the fact that current can flow both ways through the load. On subsequent consideration I became aware of this error, and the calculations shown in the Appendix 1 are correct. That part of the article relating to this - the last half of the third paragraph on page 322 - is however, in error. The values 1.2W and 640mW should be substituted for the 300 and 160mW figures shown and the remaining 35 words of that paragraph deleted. I apologize to readers for this contradiction appearing in the text.

J. LINSLEY HOOD.

Class AB - some questions (September 1970)

Following the two articles on a class AB amplifier design by Mr. Linsley Hood and also the correspondence in the August issue, we would like to raise several points concerning the specification.

Total harmonic distortion is specified as less than 0.02% at all power levels below maximum output, but this is presumably (see Figs. 6 and 7) only at 1kHz though not specified as such. What are the distortion levels at 100Hz and 10kHz at full output, for example?

When quoting a noise level for the amplifier, the noise bandwidth of the measurement was unspecified thus rendering the result as meaningless as quoting a frequency response without limits (e.g. +/- 3dB).

A value for "square-wave transfer distortion" is given as 0.2% at 10kHz but the power level is not specified. As "square-wave transfer distortion" is a non-standard quantitative measurement, for the result to be meaningful, an explanation is required as pointed out by Mr. Gibbs in his letter in the August issue. Also results for other amplifiers, for example a good class B amplifier, would be useful for comparison.

MARTIN SMITH and H.P. WALKER,
Southampton, Hants.

Notwithstanding the perfection of Mr. Linsley Hood's latest amplifier in practice, I would differ with him over some of the points he raises in the July issue.

A Darlington pair has a lower mutual conductance than the output transistor on its own. The converse can only be true of the complementary pair configuration. His first paragraph attributes a higher value to both pairs.

The overall linearity of the output stage of his Fig. 2, when driven from a genuinely low source impedance, does depend on the quiescent current contrary to his expectations. A high drive impedance is the answer, with a low inter-base impedance. This does not impair the cut-off performance as the conducting transistor presents a low base-emitter impedance to the one being cut off.

The output stage of Fig. 3 operates between the common emitter and the common collector modes. The true emitter follower of Fig. 2 has an inherent distortion of about 100 times less than Fig. 3, provided that the source impedance is low enough and the quiescent current is appropriate. Infinite values of bootstrap capacitance are necessary to secure pure common emitter operation; this circuit is predominantly common emitter above 30Hz. His calculation of class A output power assumes that the output transistors have a constant mutual conductance. Due to the bend in this characteristic at low collector currents they do not cut off as soon as expected. The class A output of either version is nearly 2 amps pk-pk. Using a standing current of 100 mA and no emitter resistors, a class A output of over 5 amps pk-pk is available. (The traditional definition of class A does not preclude current ratios between the two halves of 10^8.)

A high class A power is not, ipso facto, a particular virtue. The correct quiescent current is related to the linearity of the output stage under dynamic conditions, and this ought to be significantly lower than that required by full class A operation, in a good class AB design.

The mutual conductance of MJ 481/491 with 0.82ohm emitter resistors is 1mho at high currents; this falls to 0.5mho at a collector current of around 20mA. If Tr3, 4, 5 have high current gains, so that the drive impedance really is low, this is the optimum quiescent current with a bandwidth of a few kHz. Higher quiescent currents worsen the performance. A current of 200mA is undoubtedly right for bandwidths greater than this, but no compromise would be necessary if the drive impedance was high enough for all combinations of transistors.

Poor matching of the output transistors is extremely unlikely to cause any noticeable deterioration of the performance, except to a distortion meter; low gains may even be advantageous in certain cases. Full class A operation is unnecessary in both these circumstances.

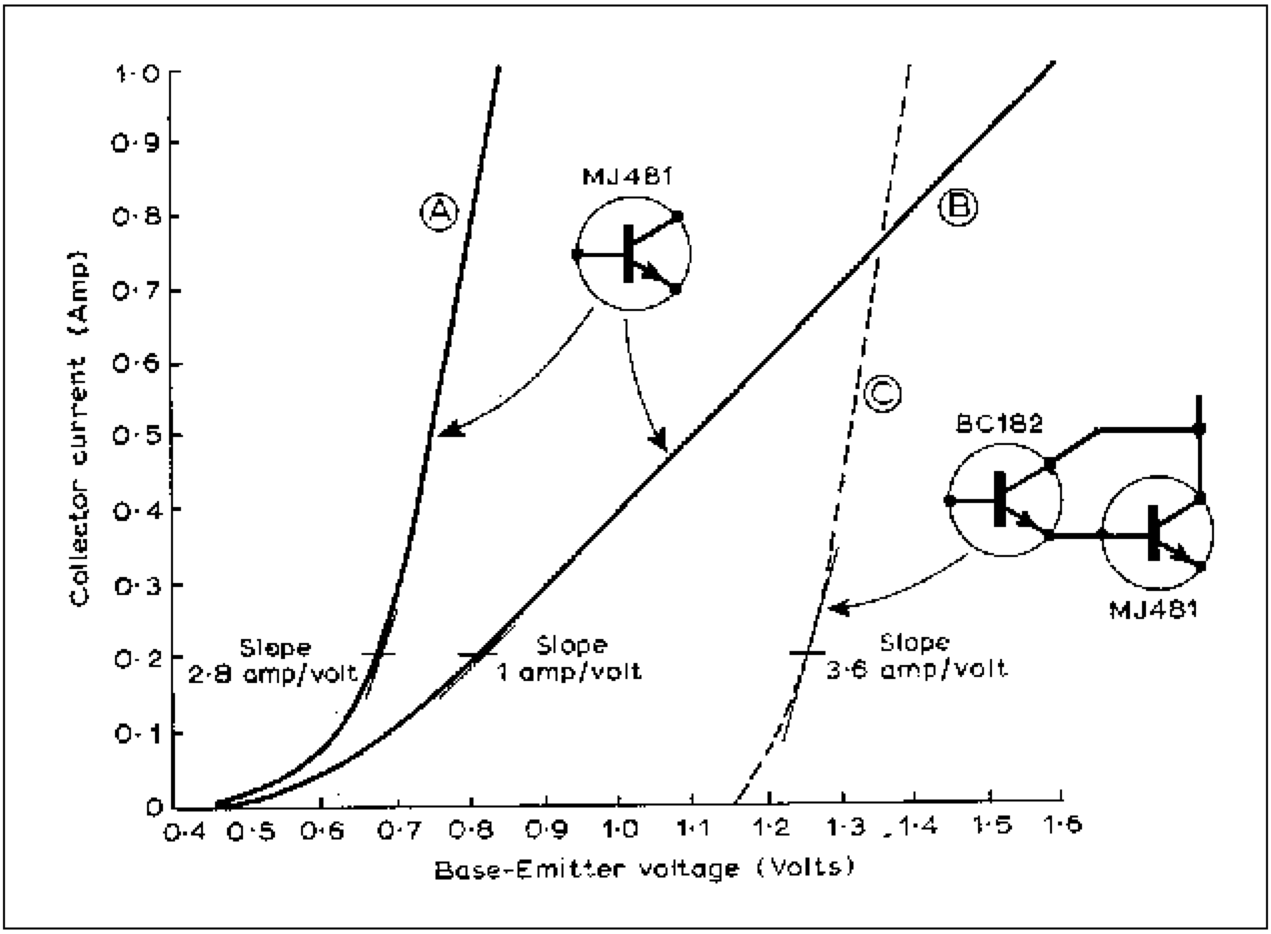
My final point concerns the avoidance of temperature-compensation in the biasing of the output stage. The penalty for this is very poor thermal stability in the 8ohm version.

D. L. D. MITCHELL,
University of Bradford.

Class AB amplifiers (October 1970)

I am grateful to Mr. Mitchell for his letter in the September issue concerning my class AB amplifier, but there are some points which he makes which, I feel, should not pass without challenge.

In particular he states that a Darlington pair output stage has a lower mutual conductance than the output transistor on its own. While, in theory, this could follow from the fact that the second transistor imposes an impedance in the emitter circuit of the first, this situation does not arise under any but near zero source impedance systems, as I have illustrated in the transfer characteristic graphs on the next page.



Curve A is the transfer characteristic of a simple (MJ481) transistor with a source (base input circuit) resistance of 10 ohms. Curve B shows the performance of the arrangement but with a source resistance of 100 ohms. Curve C is that of the same output transistor, but with an input Darlington configuration using a BC182 input transistor. There is no measurable difference in performance, in this configuration, with source resistances of 10, 100 or 1,000 ohms.

In the event, the slope of the Darlington pair, at 200mA, which was my chosen quiescent current, is 3.6 amps/volt as compared with 2.8 amps/volt for the simple output transistor.

The presence of as little as 100 ohms input circuit resistance reduces this to 1 A/V, which confirms the point I made in my article, which was concerned, implicitly, with the circumstances which would exist in a practical design.

The second point on which I differ from Mr. Mitchell concerns the conditions of operation of a class A stage. I believe this classification should be restricted to systems in which each component of the output stage operates in its linear region over the whole of its effective output swing. The mere fact that one or other of the output transistors is not completely cut off is not enough to satisfy this requirement.

Although I had not mentioned this point specifically in the article, the use of the amplifier in true class A does bring about a reduction in the distortion typically to below some 0.01%, at power levels below 15 watts, over the frequency range 100Hz-5kHz, and the distortion content then decreases linearly with reduction in output signal magnitude.

My decision, in the design of the amplifier, to employ a variable resistor, as a source of bias, between the bases of the output transistors, rather than a more complex temperature compensation network was based partly on the convenience of adjustment of such a biasing system, as compared with, say, a string of diodes (two forward biased silicon diodes will, in fact, give almost the correct quiescent current, and this arrangement was used in some of the prototypes in use by friends) and partly on its lesser proneness to catastrophic failure than transistor "amplified diode" systems.

My curve B indicates the relative insensitivity of the single transistor output stage to variations in forward bias (and the choice of 200mA quiescent current very much reduces thermal effects, even with an 8 ohm load!) as well as the excellent transfer linearity of such a system which contributes to the lower harmonic distortion figures obtainable with such an output stage in comparison with the more normal push-pull configurations.

Both Mr. Mitchell and Mr. Gibbs (letters, Aug.1970) have taken me to task for my observation in the article that "the use of a complementary pair of emitter followers 'driven from a low source impedance' appeared to offer the best way of minimizing the several problems" described in the introduction.

The article in question was in fact written as one, rather lengthy, article which was divided in two for convenience of publication, and this division, coupled with some editorial deletions, resulted in the observation above being given an unexpected degree of prominence. Since I was, at this stage, reviewing the thought processes which had led to the choice of this output stage configuration, it would have been better if I had continued "and this type of stage was therefore chosen as the starting point for this design".

In the event, both the preliminary calculations and the initial experiments indicated that it was neither practicable nor desirable, from the point of view of linearity of operation, that the output stage should have a low source impedance and the solution suggested by Mr. Mitchell in his letter, that of a relatively high driver impedance with a low inter-base impedance, was the configuration which had been adopted in the final design.

In reply to the letter from Messrs Smith and Walker in the September issue I would point out that the total harmonic distortion was quoted at 1000 Hz, because this is the recommendation of the B.S. and DIN specifications. The t.h.d. figures, at full output, at 100Hz and 10kHz, are typically 0.04% and 0.06% respectively. At low frequencies the harmonic distortion is mainly influenced by the impedances of the power supply bypass capacitor and the decoupling and 'bootstrap' capacitors, and an improvement can be made, if necessary, by increasing the value of these.

At high frequencies, the distortion content is mainly determined by the deliberate and necessary reduction in the open-loop gain, and feedback factor, required to maintain good reactive load stability, although the circuit layout and stray capacitances have some effect.

I apologise for the omission of the bandwidth limits for the noise figure measurements. These were effectively those imposed by the amplifier gain/frequency characteristics, as would be measured by a very wide bandwidth millivoltmeter. The use of a more restricted bandwidth, say 20Hz-20kHz, would allow an apparent improvement in the specified noise figure. (It is, in fact, quite inaudible.) However, on looking through back numbers of Wireless World I find that other authors have been equally remiss in omitting measurement bandwidths when quoting noise levels. This point will, perhaps, be noted in the future.

I regret that the measurement parameter "square wave transfer distortion" was not accompanied by some further explanation. In practise, transfer distortion is measured by comparing electrically the waveforms at the input and output of the system under test, and then expressing the error arising in the transfer as a percentage of the input waveform, as measured on an r.m.s. calibrated voltmeter such as that used for conventional t.h.d. measurements. Any convenient waveform may be used for this purpose.

Typical values for transfer distortion with conventional audio amplifier designs using a 10kHz square wave and a resistive load range from 0.2% to 10%. Square-wave transfer errors as high as 30% are fairly common under reactive load conditions, and this, in conjunction with the relatively high distortion levels sometimes found at low volume levels, may account for much of the so-called 'transistor sound'. Unlike harmonic distortion, transfer distortion with reactive loads may worsen as the amount of negative feedback is increased.

J. L. LINSLEY HOOD,
Taunton, Somerset.

Class AB amplifiers again (December 1970)

Mr. Linsley Hood's reply in the October issue to my letter (August) does indeed clear up the difficulties I experienced in following his article and his reply to Mr. Gibbs (August issue), but I feel bound to justify my objections more fully. I understand the mutual conductance of a transistor or a pair of transistors to be dIc/dVbe. Vbe is measured between the input base and output emitter, under precisely those near zero source impedance conditions to which he refers. With values of less than an ohm the shape of the basic mutual characteristic of the MJ481 is preserved. The curve obtained with 100 ohm source resistance looks much more like the current gain characteristic, except at low collector currents. If the effect of the 10 ohm resistor is removed from Mr. Linsley Hood's curve A, the slope does become steeper than that of curve C. Consider an MJ481 with and without a 0.2 ohm emitter resistor and with and without a 40361 driver in the Darlington pair configuration, with zero source impedance (Fig. 1), with modifications where appropriate. It is easier to work in terms of mutual resistances than conductances, and representative values of these are shown in Table 1 (R is infinite here).

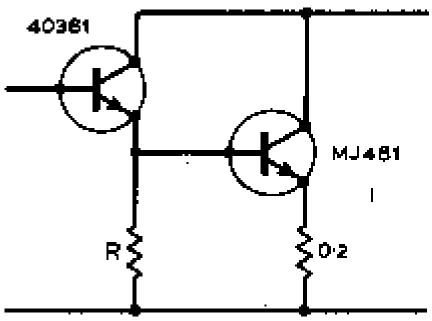


Fig. 1.

The mutual resistance of combinations of these three, including the MJ481, is the sum of these resistances seen at the output emitter. The MJ481 is assumed to have a current gain of 100; this does not prejudice the argument as the characteristic of the 40361 is nearly exponential, so that the slope is approximately inversely proportional to Ic. The results for the four cases are shown in Table 2. The optimum quiescent current for a voltage driven stage is normally the collector current at which the resistance slope is twice its high current value.

Table 1

$\frac{dV_{be}}{dI_c}$ of MJ481	I_c of MJ481	I_c of 40361	$\frac{dV_{be}}{dI_c}$ of 40361 / 100
Ω	mA	mA	Ω
0.16	1.000	10	0.04
0.25	200	2	0.15
0.32	100	1	0.30
0.50		0.5	0.50

Table 2

Combination of components	Slope at 1 A output current	Optimum quiescent current	Slope at this current
	Ω	mA	Ω
MJ481	0.16	100	0.32
MJ481 +0.2Ω	0.36	50	0.70
MJ481 + 40361	0.20	200	0.40
MJ481 + 40361 + 0.2Ω	0.40	100	0.82

It can seen that the addition of an emitter resistor reduces the optimum quiescent current and of a driver increases it although either addition reduces the overall mutual conductance at all currents. The effect of finite values of R is to reduce the change introduced by the driver.

The p-n-p/n-p-n configuration is more complicated (c.f. Mr. Baxandall's letter in the September 1969 issue), but in general it has a higher mutual conductance (Fig. 2, r=0) than the simple output transistor. With common values of r the combination is linear down to much lower collector currents in the output transistor, giving a lower half-slope current. With a high source impedance the optimum quiescent current for a complementary or quasi-complementary output stage is not so readily defined. It may well be Mr. Linsley Hood's experiences in these circumstances which leads him to the conclusion (August issue) that the optimum quiescent current varies inversely with the absolute magnitude of the current gain in half of the output stage.

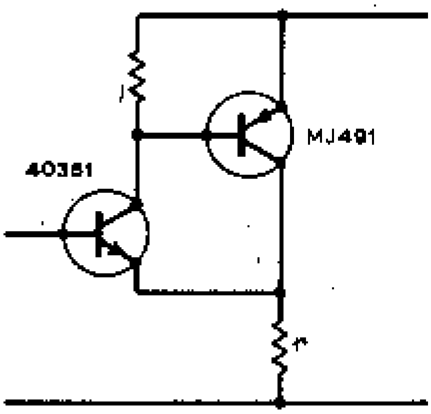


Fig. 2

The “circumstances which would exist in a practical design" are precisely those put there by the designer; source impedances of under 1 ohm are perfectly feasible. It begs the question to insert resistors in the base lead before even measuring the basic properties of the transistors. The mutual characteristic so obtained is only relevant to a complete amplifier which has these impedances in series with each half of the output stage - resistors R1 & R2 in Figs. 3 & 4 - excepting pure class B using transistors which cut off perfectly and do so with zero base-emitter voltage. If R1 & R2 are zero, and R3 is finite, the overall transfer characteristic of the complete output stage is best not looked at in terms of the mutual conductance measured when one transistor is omitted.

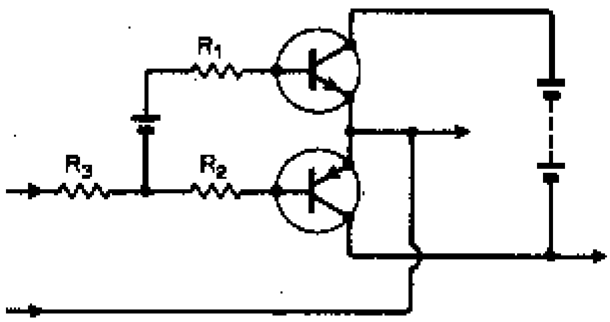


Fig. 3

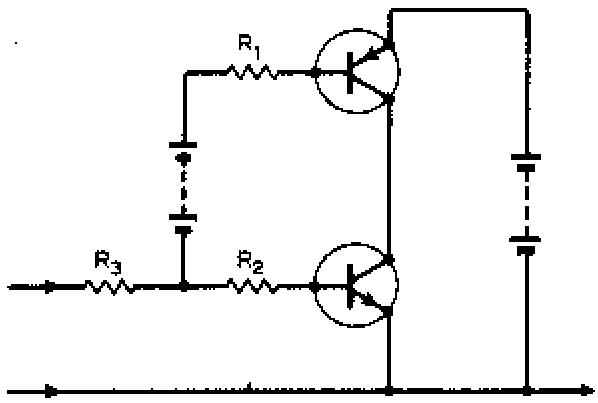


Fig. 4

I apologize for making objections in terms of the article, since it does not convey the sense that the author intended, but I based my arguments on the design itself. The source impedance to the output stage is genuinely low. The minimum current gain of an MJE521 at 50 mA collector current is about 80, giving a drive impedance of 70 ohm at the most (derived from the 6.8 kohm resistor). The input impedance of the output stage varies between 50 and 100 ohms in the 15 ohm version with output transistors of current gain 100. It is the inappropriate ratio between these two quantities which is responsible for the effects to which I referred.

It would be convenient if the bootstrap capacitor could supply the extra current required to drive low gain MJ491s which need a base current in excess of the standing current in the driver stage. This could only occur if the bootstrap capacitor temporarily sustained a greater voltage than it does under static conditions. This situation arises during a short negative transient (MJ491 on) a short time after a long positive excursion (MJ481 on). Short and long are referred to the time constant of the bootstrap capacitor and R4 in Fig. 3 of the article. Quite how common these conditions are in music (with whatever d.c. components there might have been removed well before bootstrap capacitor has its say) I can't imagine.

The other points I should like to make are best left to a future date - we both appear to be drawing on material which should see the light of day in articles rather than in letters.

DUNCAN MITCHELL,
Postgraduate School of Electrical and Electronic Engineering,
University of Bradford.

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HISTORY: Page created 20/07/2001
12/01/2002 December 1970 letter added

The JLH Class-A Output Stage
Single-ended or push-pull?

I have been asked on a number of occasions if the output stage of the JLH Class-A is single-ended or push-pull. My usual reply has been “do you want the short or the long answer” and then, without waiting for a response to what was in any case a rhetorical question, I have proceeded to send the long one. This brief article may save me from having to write it again.

Let's start off by defining what we mean by single-ended and push-pull when applied to the output stage of an audio amplifier.

A single-ended stage has a single output device (or combination of devices eg a Darlington or Sziklai pair) which handles all of the signal. The stage must operate in Class-A and the peak output current is equal to the quiescent current.

A push-pull output stage can operate in either Class-A or Class-B (where each transistor switches off for half the cycle). I will only consider the Class-A scenario. Two output devices (or combination of devices) are required and these are usually complementary types when BJTs or MOSFETs are being used. The circuit is arranged such that as one half of the output stage turns on more and delivers or accepts more current, the other half is turned off by a similar amount and its current is reduced. In this way, the current in the load is twice the current change in one half of the output stage and the peak output current, whilst maintaining Class-A operation, is equal to twice the quiescent current (not quite, but as near as dammit).

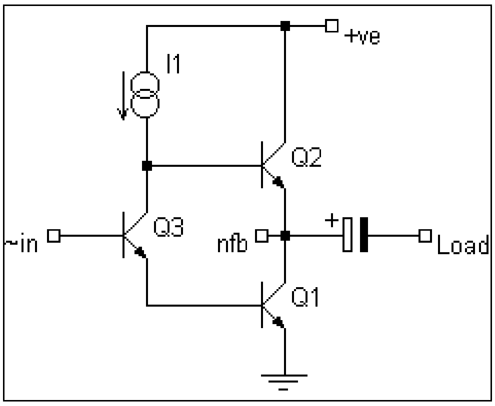
The JLH output stage can be analyzed in various ways. Doug Self describes the circuit as push-pull, with Q2 as an emitter follower and Q1 acting as a voltage controlled current source. James Sugden offers a similar opinion ".... the output stage is more akin to the White cathode follower. The upper transistor is the 'driven' element acting as an emitter follower with the second device as the emitter load but driven again in such a manner to vary itself in a beneficial way."

On the other hand, JLH himself opined in the 1996 article that, if matched devices were not available, it was better to use the transistor with the higher gain as Q1 so that the output circuit worked as an amplifier with an active collector load rather than an output emitter follower with an active emitter load. He later stated in an article in Electronics World (May 2000) that "..... my 10W Class-A circuit did not use a push-pull pair of output devices to provide the required low output impedance. Instead it used a 'Darlington pair' connected amplifier stage, comprising Q1 and Q3, driving Q2 as an active load."

The statements in the last two paragraphs all seem to indicate a single 'driven' device (thus meeting the definition of single-ended) with an active emitter or collector load providing some push-pull action. This is similar in many respects to the Pass Aleph circuit which is single-ended with a modulated current source for the output MOSFET load. The main difference between the two is that the Aleph modulates the current source by way of the output current whereas the JLH does it via the input signal.

Though JLH has stated, as indicated above, that the Class-A circuit does not use a push-pull pair of output devices, this is at variance with his comment in the original 1969 article that ”The use of a second, similar, transistor as a collector load ……………. would allow the load to be driven effectively in push-pull if the inputs of the two transistors were of suitable magnitude and opposite in phase. This requirement can be achieved if the driver transistor is connected as shown in Fig. 2.”

However, now for an alternative view based on information published by Wireless World back in 1973. The output stage of the JLH is basically that shown in the following diagram. A constant current source feeds the junction of Q2b/Q3c. Assuming Q1 and Q2 are perfectly matched, with no signal present half the ccs output feeds in to the base of Q2 and the other half passes through Q3 into the base of Q1.



The ccs current, coupled with the gain of Q1/Q2, determines the quiescent current of the output stage. When a signal is present at the base of Q3, Q3 will either pass more or less current depending on the polarity of the signal. If it turns on more, the current into the base of Q1 will increase. Because of the fixed ccs output, the current into the base of Q2 must decrease by a similar amount. The converse is true if the signal causes Q3 to pass less current. We therefore have an accurate current phase splitter and true push-pull working of Q1 and Q2 is achieved.

If transistors were perfect devices, the push-pull action would enable the peak output current to equal twice the quiescent current. However they are not and Ic does not vary linearly with Ib, particularly when Vce is changing as well. This means that for a particular change in base current, the change in Ic of the transistor that has a reduced base current will be greater than the increase in Ic for the other transistor. Simulation has shown that the peak output current for the JLH will be between 1.3 and 1.55 times the quiescent current depending on the particular output transistors being used.

So, what happens if the transistors are not perfectly matched? With no signal present, the base currents to Q1 and Q2 will not be equal (the ccs current will be split by the inverse ratio of the two gains). As I proved in one of my posts at the diyAudio forum, this limits the possible base current change for one of the output devices. At low signal levels this does not make much difference, the push-pull action is maintained, but as the signal level increases push-pull action ceases earlier than it would do with matched transistors. In fact it can be shown that, with a significant difference in gain between the two output transistors, the peak output current (in one direction) can equal (or even be less than) the quiescent current. If the peak (symmetrical) output current is equal to the quiescent current do we still have push-pull working?

In reality, I doubt that any of these analyses or explanations is strictly accurate due to the complex interaction between Q1, Q2 and Q3 and the internal feedback around the loop formed by these three devices. However, suffice to say that this design produces very good sound quality so perhaps it is unnecessary to ponder too long on the intricacies and inner workings of the circuit.

Well, I don't seem to have answered the original question "is the JLH single-ended?". Perhaps I had better give the short answer instead –

Yes .... or no.

Modular Pre-Amplifier Design

(Wireless World, July 1969)

Optimally designed stages that may be used separately or in several different combinations

by J. L. Linsley Hood, M.I.E.E.

The type of distortion introduced by a class A transistor amplifier operating at a low signal level will be predominantly second harmonic and inoffensive to the ear. Although harmonic distortion is a convenient thing to measure, and makes a reasonable yardstick for comparative purposes, at low levels its presence is less important than that of the intermodulation effects it causes. When a complex signal is transmitted through a non-linear element, intermodulation products between the separate components of the signal are formed, and these are readily apparent in the final audible result as a “blurring”, and the loss of separate identity, of the individual components which make up the whole. A measure of this is the ease (or difficulty) in distinguishing the words of a choral performance in the presence of an orchestral background, or in identifying the presence and nature of individual instruments in a large orchestra.

Measurements by a number of workers (1) have indicated that the magnitude of intermodulation products can be much greater than that of the total harmonic distortion level, and the non-linearities which are likely to be of the most importance in this respect are those at the low- and high-frequency ends of the audible range.

At the moment, the performance of audio amplifiers is much superior in this respect to that of f.m. transmissions, tape recordings, disc replay systems, or loudspeakers. However, advances in manufacturing techniques of gramophone records, pickup cartridges and loudspeakers have allowed a continuing improvement in the performance of these in harmonic and i.m. distortion, and it is clear that any amplifier design offered at this time should have a very high standard of performance if it is to remain of continuing value over the next decade.

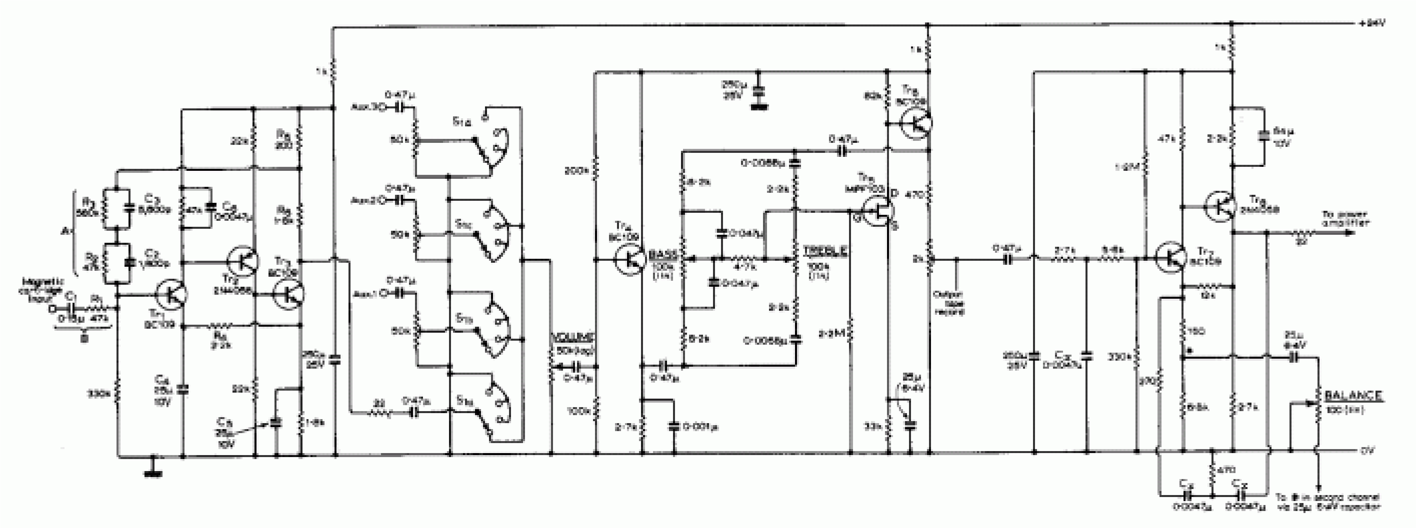


Fig. 1. A likely combination of stages.
(Click on figure for a higher resolution image)

The author has designed a range of high-quality pre-amplifier stages. Each stage performs its required operation with negligible noise and distortion. When joined together, as for example in Fig. 1, the total harmonic distortion level is below 0.1% over the frequency range

20Hz-20kHz, at any tone control setting, and for up to 2V r.m.s. output. Each stage is capable of operating on its own and has an output impedance low enough for screened cable inter-connections to be made without high frequency loss.

Magnetic pickup equalization circuit

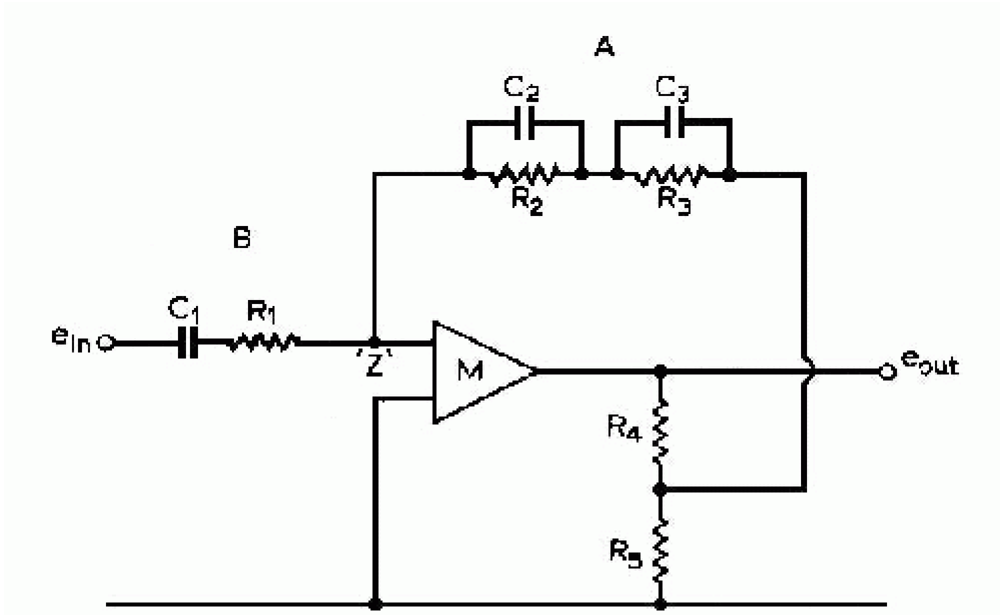


Fig. 2. Phase-inverting amplifier stage used to obtain R.I.A.A. replay characteristic.

The required R.I.A.A. replay characteristics can be approximated by several different circuit arrangements. The most straight-forward from the point of view of performance calculation is that shown in Fig. 2, employing a simple phase-inverting amplifier stage. If the gain of amplifier M is high enough, point Z becomes a virtual earth (see Appendix I), and the input impedance of circuit equivalent to that of the input network B. The load resistance required by the pickup cartridge, usually 47-50kohm, is provided by a suitable choice of R1. With resistor R2 equal to R1, stage gain is given by R4 + R5/R5 at the mid-point frequency (usually 1kHz) if the impedance of C2 is large, and that of C3 small in relation to R2. Since the voltage output to be expected from most good quality magnetic pickup cartridges is in the range 4-10mV for a 5cm/sec recorded velocity, a gain of 10 is adequate for this stage. The required replay frequency-response curve shown in Fig. 3 can be obtained by a suitable choice of C2 and C3. Since the two networks A and B determine the frequency response of this circuit, it is apparent that substitution of these can be made to provide a wide range of different performance characteristics without alteration to the circuit of amplifier unit M.

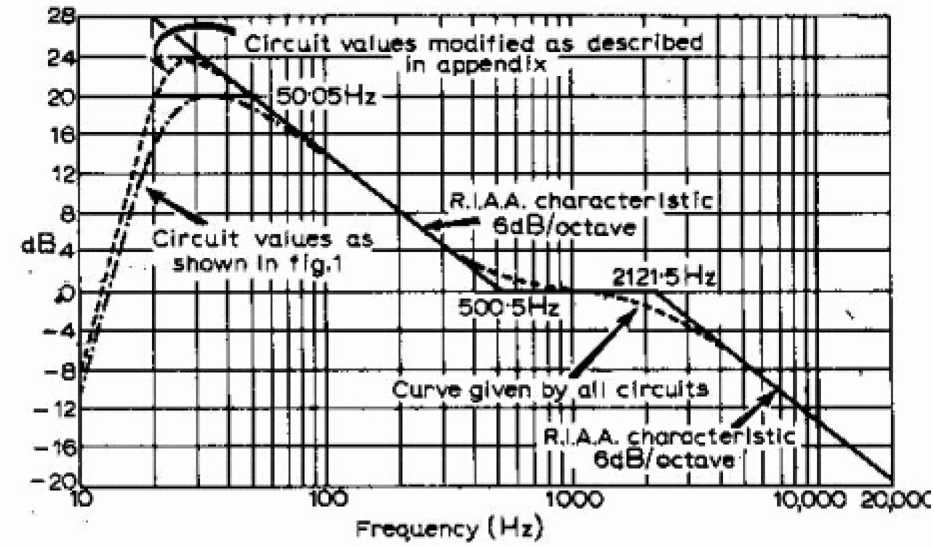


Fig. 3. Required R.I.A.A. frequency-response curve and circuits approximation to this.

The final circuit can be seen at the front of Fig. 1. Because phase inversion between input and output is required, and because the necessary gain is higher that can be obtained from any single transistor arrangement, a triplet circuit has been used. Tr1 and Tr3 are high-gain, low-noise voltage-amplifying stages, and Tr2 is a phase and voltage transformation stage allowing the input transistor to be used in its most linear region. The output transistor has a low collector load resistance, to reduce distortion to the lowest possible level.

D.C. working-point stability is ensured by D.C. negative feedback through R3 and R2 to the base of Tr1, and through R4 to the emitter circuit of the same transistor. The circuit R4, C4, and C5 also provides the feedback path necessary, in conjunction with the input capacitor C1, to provide an 18dB/octave steep-cut rumble filter, with a turn-over frequency of 25Hz (see Appendix II), and an ultimate attenuation of more than 40dB at 8Hz.

Capacitor C6 provides phase correction, and is essential for a clean square-wave response, and freedom from transient ringing, when used with a capacitive load.

The response of this circuit is particularly good, and it can deliver up to 1 volt output with distortion less than 0.02% from 100Hz to 10kHz.

Stages for ceramic cartridge equalization

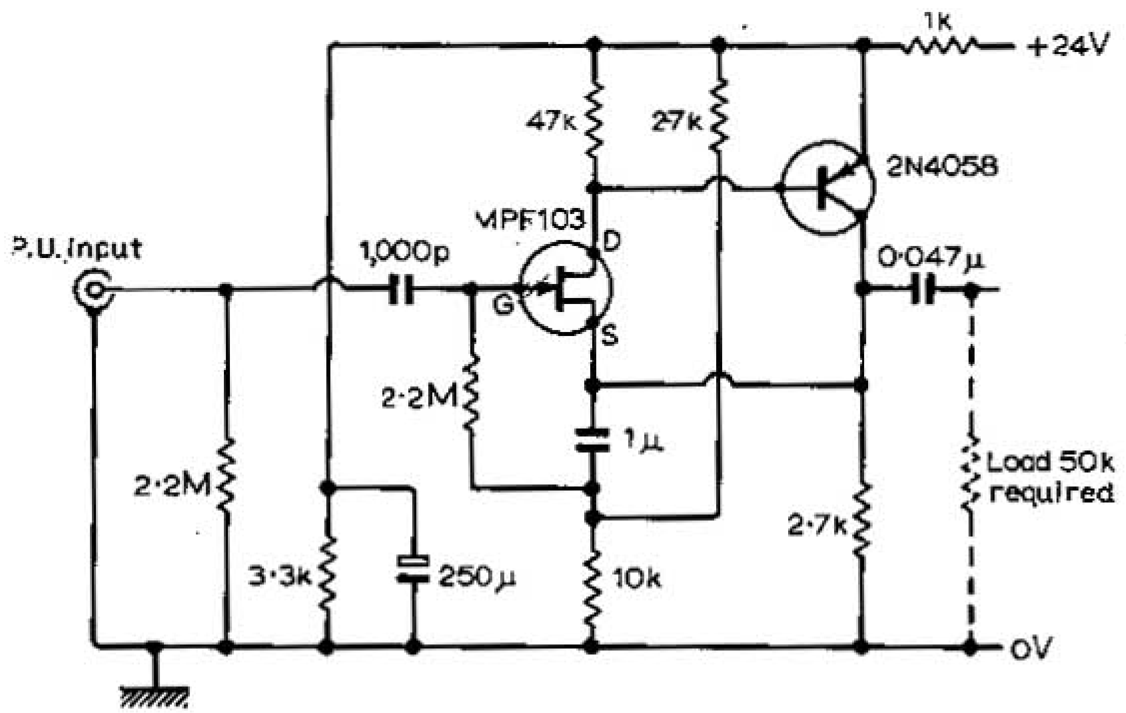


Fig. 4. Impedance conversion stage for use with ceramic cartridge. This may be directly substituted for the magnetic cartridge stage at the front of Fig. 1.

Fig. 4 is an impedance conversion stage contributing less than 0.05% distortion at 1kHz and having a flat response from 35Hz to greater than 200kHz, with 18dB/octave roll-off below 35Hz. This simple stage may be directly substituted for the magnetic cartridge stage of Fig.1.

Alternatively, should it be required that the pre-amplifier be able to cope with inputs from both magnetic ceramic cartridges, then switchable equalization networks for A and B can be provided. These are shown in Fig. 5. When used with a ceramic cartridge the output is from 50 to 200mV. To preserve the required shape of the rumble filter characteristic it is necessary to alter the values of C4 and C5 from 25uF to 12.5uF. The pre-amp response is then as shown in Fig. 5, curve 1.

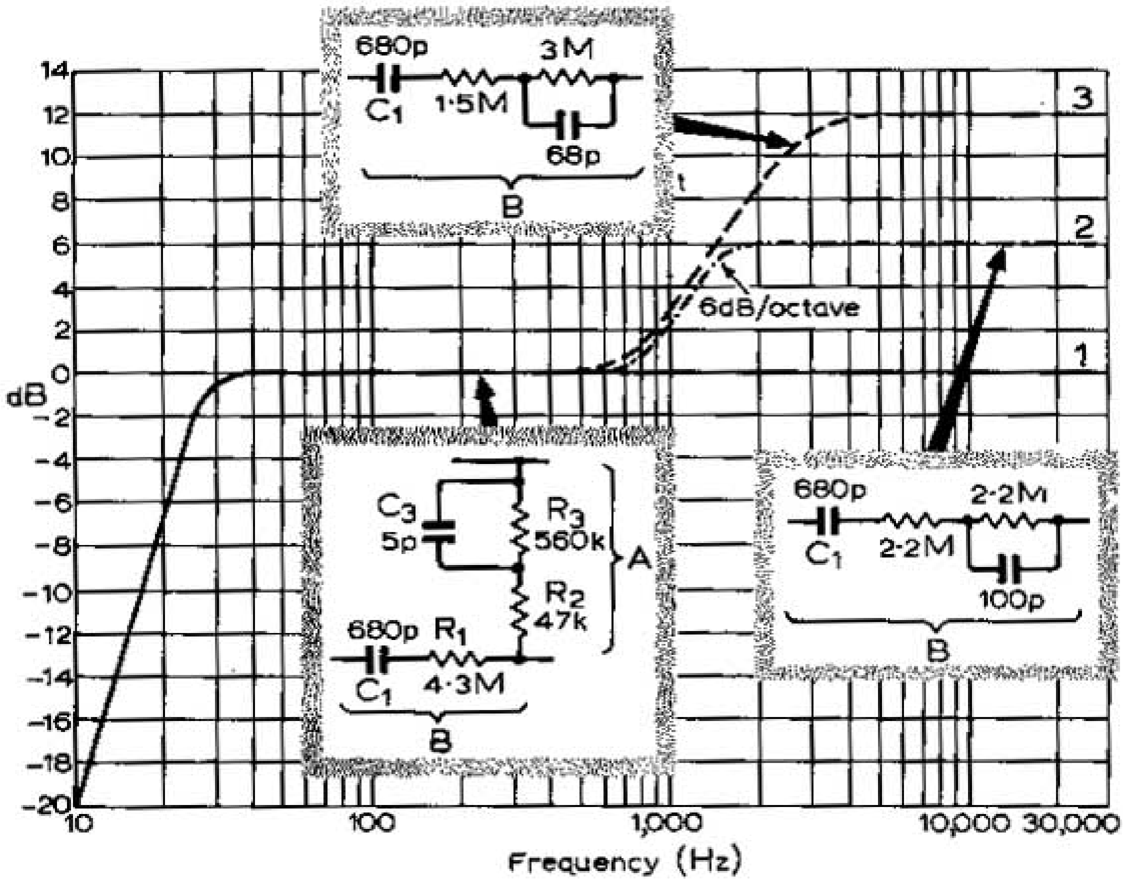


Fig. 5. Changes in equalization networks A and B of the magnetic cartridge input stage allowing direct use of ceramic cartridge. Components for network A are the same for the three curves show.

The performance of many ceramic pickup/amplifier combinations is disappointing in comparison with that obtainable from a good magnetic cartridge with a similar amplifier. This is sometimes due to the mismatching between cartridge and amplifier, or through inadequate input impedance provision (in the modification shown in Fig. 5 this is 4.4Mohm), or due to the failure of the piezoelectric element within the cartridge to provide the required equalization for the 12dB fall in voltage output anticipated when a recording having R.I.A.A. velocity characteristics is replayed on a displacement sensitive device. In the latter case, a very considerable improvement in the relative performance of the ceramic cartridge may be obtained by shunting part of the input resistor in the input network B by a small capacitor. Curves 2 and 3 in Fig. 5 show partial and complete correction respectively.

Tone-control stage

The tone-control stage is of conventional type, and uses a negative feedback system derived from the design due to Baxandall (2). However, it differs from normal practice in that a junction field-effect transistor is used as the active element. Field-effect transistors have both lower noise levels and better linearity than bipolar transistors, and in this type of circuit the high input impedance results in negligible loading of the tone-control network. The stage gain needed in this circuit requires a high value drain load resistor, and the f.e.t. must therefore be followed by an emitter-follower to provide the low output impedance desired for easy interconnection of the separate units.

If the feedback tone-control network is to perform satisfactorily, both the input and output impedances seen by the network at its ends must be low in relation to the network input impedance when the sliders of the potentiometers are at the position nearest to the point being measured. Some form of impedance conversion circuit is therefore also needed between the volume control and the tone-control circuit. An emitter follower is also used at this point. The 0.001uF capacitor in the emitter circuit of Tr4 is to avoid the possibility of high frequency parasitic oscillation occurring if long screened leads are used to connect the base of Tr4 to the volume control.

The input to this section is taken through a switch from the gramophone pre-amplifier section, and other inputs provided with preset gain-equalization potentiometers. The switch is arranged to earth the inputs not in use, to minimize breakthrough between programme channels.

The gain/frequency characteristics of the stage are shown in Fig. 6.

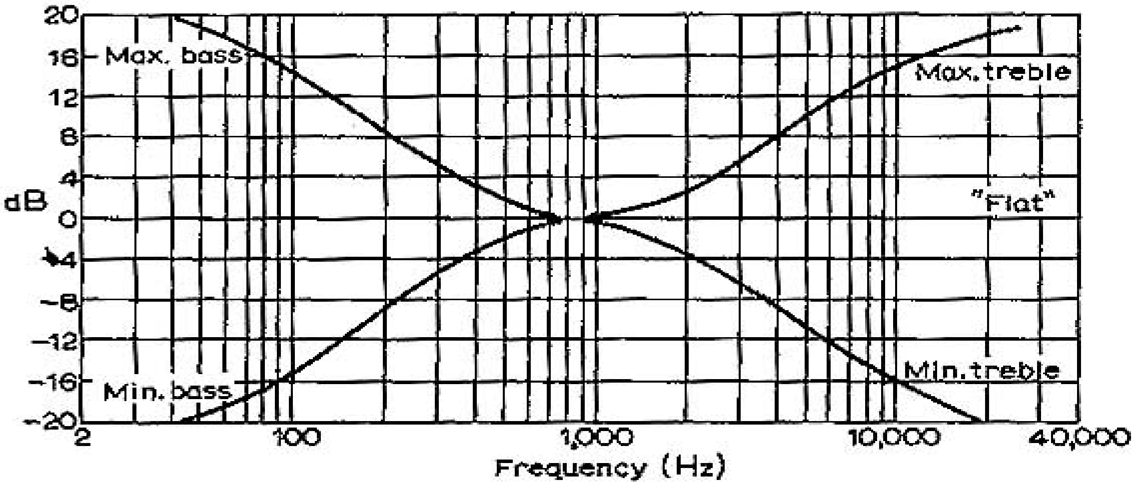


Fig. 6. Gain/frequency characteristics of tone control stage.

Low-pass filter circuit

The voltage amplifying stage preceding the main amplifier should include a steep-cut low-pass filter that can be set to remove unwanted high frequencies. This can be done either by a suitable LCR filter arrangement, or by an active filter giving an equivalent performance without the use of inductors. The circuit arrangements available for low-pass active filters are shown in Fig. 7. (b) is the well known circuit arrangement first employed in an audio amplifier design by P. J. Baxandall (3), and (d) is the unity gain rearrangement of this circuit introduced by Sallen and Key (4). The frequency response of all these circuit arrangements is similar, *mutatis mutandis*, to that shown in Fig. 8, and the circuit should be preceded or followed by a simple RC filter if the type of response shown in the dotted line is required.

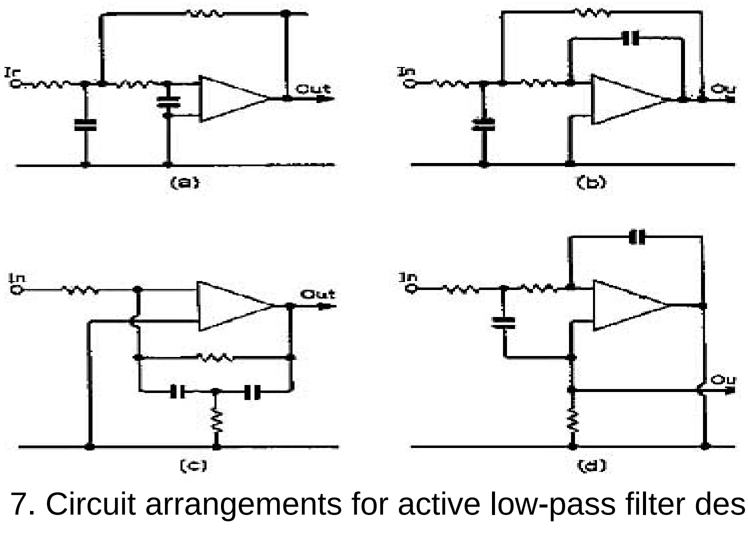


Fig. 7. Circuit arrangements for active low-pass filter design.

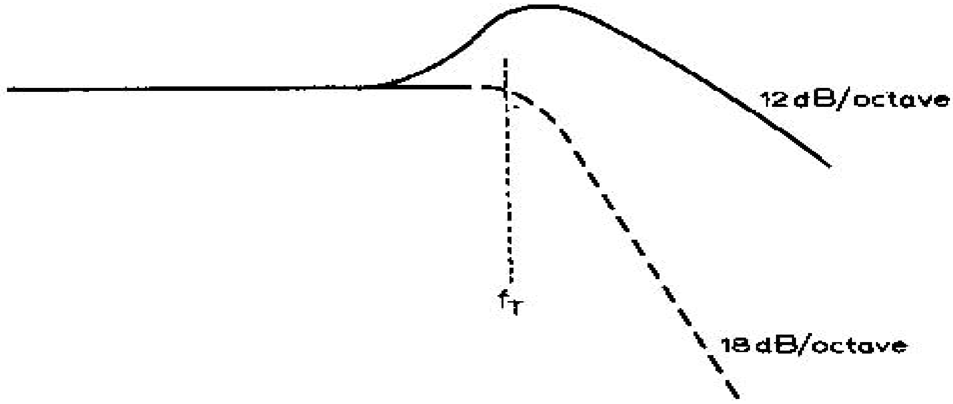


Fig. 8. Frequency response of the active filter circuits is 12dB/octave. Preceding the filter with RC network gives response shown in broken line.

For a given overall stage gain, type (b) gives much better distortion factor near the region of cut-off than (a), and (c) is marginally better than (b) when used with non-linear amplifier elements. The particular advantage of (c) however, is that it can be used conveniently with a very low-distortion two-transistor circuit.

The final stage, with the filter circuitry, is shown in Fig. 1. As a matter of practical convenience, the component values of this circuit have been chosen so that the required low-pass response is obtained when all of the capacitors 'Cx' are of equal value to each other. The frequency response obtained with a given value of 'Cx' can be found in Fig. 9. The user can interpolate between these to obtain turn-over frequencies at any points to suit his own requirements. If a ganged selector switch is employed to give a range of turn-over frequencies, the switch arms (moving contacts) should be connected to the junction of the resistors in the RC filter and to the 470ohm resistor in the main filter network. In Fig. 1 the 0.0047uF capacitor for 'Cx' results in response being 3dB down at 18kHz. With good quality programme sources this is a recommended capacitor value.

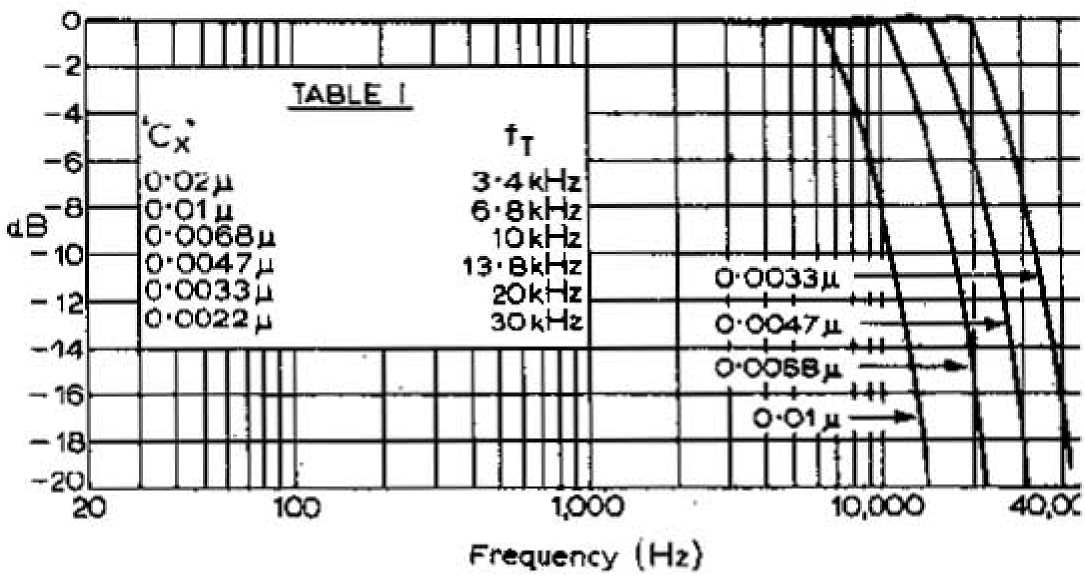


Fig. 9. Graph and table of turn-over frequencies for different value of 'Cx'.

With capacitors of zero value, the response of the circuit is flat to about 100kHz. The user should however arrange for the response to fall off above 25kHz. (It is unlikely that the listener will find anything to gain from the parts of the sonic spectrum beyond this point.)

The optimum performance of this particular type of circuit arrangement is obtained when the overall gain is about 50 with feedback. A 20-40mV input is therefore adequate for this stage for the output voltages required.

The distortion level of this circuit is less than 0.03% at 2 volts r.m.s. output or less, at any frequency within the pass band. The output impedance is less than 150 ohms over the range from 20Hz to the cut-off frequency selected.

It is convenient, for several reasons, to operate at the 60-100mV level through the tone-control stages. At this output voltage level the distortion introduced by a RC coupled f.e.t. stage is less than 0.1% even without feedback, so that the maximum ‘lift’ settings of either ‘bass’ or ‘treble’ controls cannot give rise to unacceptable levels of distortion. It is also large enough for the noise and inevitable 50Hz pickup to be unobtrusive. Some attenuation is therefore desirable between the tone control unit and the steep-cut filter circuit. This is obtained by the preset 2kohm potentiometer in the tone control circuit, which provides a convenient means for setting the overall gain of the amplifier system, and also as a coarse ‘balance control’ in a stereo system. Fine balance between channels is obtained by adjusting the 100ohm balance potentiometer in the output stage. This alters the stage gain over the ratio 6:10.

Constructional notes

The constructional technique used by the author in building the prototype of this amplifier is similar to that used in the 10-watt class-A design described in Wireless World in April 1969, with the separate units laid out in mirror image form, as a stereo pair on a single 4in X 4¾in s.r.b.p. pin board, Two units of each type can be accommodated on each board, laid out more or less in the form of the circuit diagram(or its mirror image).

In general, reasonable care should be taken to separate input from output leads, and where the boards are to be mounted as a group within the same box, it would be wise to interpose a sheet metal screen between them.

The units are separately coupled by 250uF capacitors from a common 24-volt line, derived from a zener diode stabilized RC filter power supply. This supply is separate from the main amplifier, and a 30mA output is ample. Details of a suitable power supply are given in Fig. 10. The expected working voltage on each of the unit sub-rails is about 15volts.

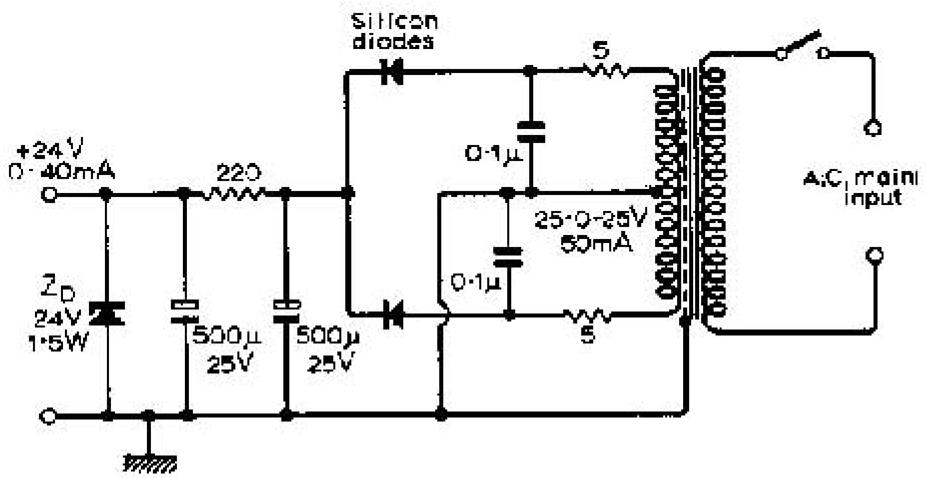


Fig. 10. Suitable power supply for any combination of stages.

Apart from the input transistor in the gramophone pre-amp unit (Tr1) for which the BC109 is to be preferred, there is no particular reason why any modern silicon planar types should not give an indistinguishable performance. For example, the n-p-n types could be 2N3904, BC107/8/9, 2N3707, or BC184Ls. Similarly, the p-n-p types could be 2N4058, 2N3906, or BC214Ls.

Although, in many cases, the use of 1/4 watt resistors is sufficient, it would probably be found simpler to use 1/2 watt units throughout. 5% tolerance carbon film resistors are to be preferred.

The author has mounted the gramophone pickup equalization circuit in a separate small diecast box, immediately under the gramophone turntable unit, so that the leads from the gramophone are taken at a low impedance from the output of this unit. This has been very effective in reducing the hum picked up on the output leads to an imperceptible level.

Appendix I

The use of 'virtual earth' (null seeking) amplifier circuit arrangements is superficially ill-advised with input elements such as pickup cartridges, because it appears that as the operating frequency is increased, the input half of the balancing limbs will also change, with a resultant change in the gain of the circuit. In particular a magnetic pickup cartridge may have an inductance of some 300-800mH and the impedance of this will exceed that of the input circuit in the range 12-20kHz. This should clearly reduce the gain of the system by reducing the ratio of A to B.

However, on reflection, it can be seen that the amplifier operates as a null generating device, sensitive only to the current flowing in the input circuit to the 'virtual earth'. As the operating frequency increases, so the current flow through R1 will decrease, but so it would in any case, regardless of the amplifier, were the element simply connected across network B as the load recommended by the cartridge manufacturers (at these frequencies the impedance of C1 can be ignored), and the voltage across R1 measured by a perfect voltage amplifier. The decrease in current input into a given resistive loads from a source having a series inductance is simply an unfortunate fact of life, from which one cannot escape, whatever one's technique of measurement, and high impedance voltage amplifiers connected across the load, or low impedance current amplifiers connected in series with it, are alike in this respect, except that with transistors, the latter are a bit easier to contrive. The same argument is also applicable, in the appropriate context, to high impedance capacitative elements such as piezo-electric pickup cartridges. Once again, the voltage amplifier and current amplifier see the same phenomena in identical form. The necessary, and inevitable, corrections can be accomplished by simply by the tone control settings.

Appendix II

Although the R.I.A.A. replay characteristics suggest an approximately flat velocity response from 20-50Hz, this would effectively imply recording bass lift in this region and the author suspects that this is not done and a constant modulation characteristic being used instead. The author has therefore, for his own use, modified the values of the feedback elements as follows: R5 – 470 ohms; R6 – 1.5kohms; C1 – 0.47uF; C3 – 6800pF; and C6 – 6800pF. These changes maintain the velocity response flat down to 25Hz, with rapid attenuation below this frequency. Unfortunately the mid point gain of the circuit is reduced to 5, and some additional amplification is therefore needed if it is desired to avoid working with the tone control circuit at the 20mV level. The simple floating emitter collector-follower circuit of Fig. 11 is therefore interposed, without coupling capacitors, between the output series resistor and the collector of Tr3. The distortion contributed by this is less than 0.05%.

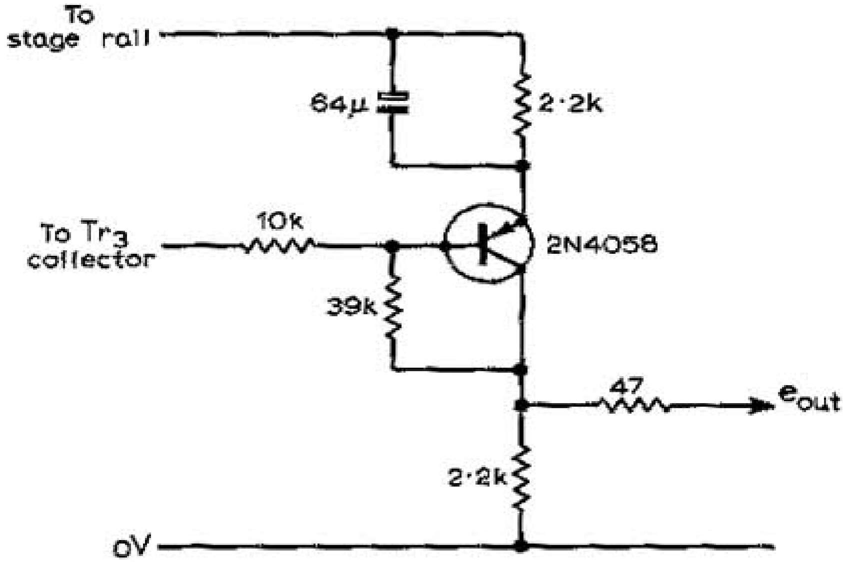


Fig. 11. Floating emitter collector-follower circuit referred to in Appendix II.

References

1. 1. Langford-Smith, F., “Radio Designers Handbook”, Vol.4 ch.72.
2. 2. Baxandall, P. J., “Negative-Feedback Tone Control”, Wireless World, October 1952
3. 3. Baxandall, P. J., “Gramophone and Microphone Pre-amplifier”, Wireless World, January 1955
4. 4. Sallen, R.P. and Key, E.L., I.R.E. Trans. Circuit Theory, March 1955, p. 74-85

Postscript (December 1970)

Modular pre-amplifier

The intention in the original article was not to offer a complete pre-amplifier design, but rather to describe a series of versatile ‘building blocks' from which the potential user could assemble a 'custom built’ pre-amplifier to suit his own needs or preferences. To increase the scope of this some additional circuit modules are described below.

Steep cut low-pass filter. It is certainly prudent to include a low-pass filter somewhere fairly close to the input of the main amplifier whenever a wide-bandwidth main amplifier is to be used with a good-quality loudspeaker system. Doing so will prevent unwanted high-frequency components, arising from component noise, record surface noise, and similar causes, from impairing the long-term listening comfort of the user, and from producing avoid-able intermodulation effects due to non-linearities in the loudspeakers.

The combination of such a steep-cut low-pass filter with a low-distortion, low-output impedance driver stage, with a gain of 50 and an output capability of some 2V r.m.s. at 0.02% t.h.d., appeared to provide the most versatile system for use with a wide variety of power amplifiers.

However, many power amplifiers require an input voltage of only 0.25 - 0.8V r.m.s., and there are snags in respect of hum and component noise if the stages following the volume control are operated at levels below some 50mV. The preferred level to achieve an optimum balance of noise and distortion components is probably in the 100 – 200mV region. In these circumstances a driver-stage gain of 50 is excessive, and much of the available gain must be removed by an input attenuator, and if a potentiometer is used for this it can introduce noise.

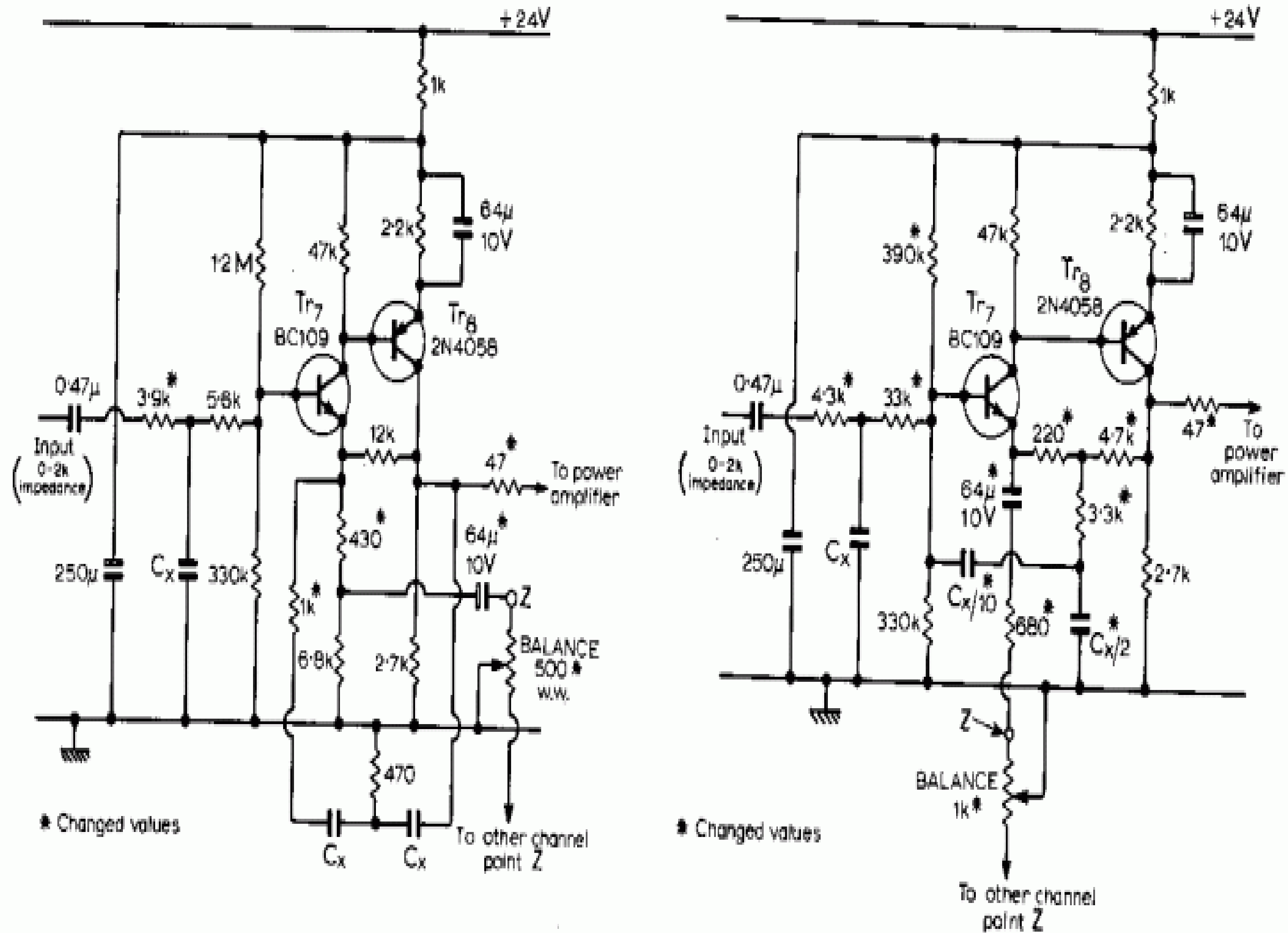


Fig. 5. Low-pass filter and driver stage (a) with gain of 20, (b) with a gain of 5. The cut-off frequencies, dependent on C_x, are the same as in the original article.

To meet this need more conveniently, two further versions of the driver amplifier, incorporating steep-cut low-pass filter characteristics which are identical to that of the original circuit, and having gains of 20 and 5, are shown in Figs. 5(a) and 5(b). An alternative, three-transistor arrangement whose cut-off slope is variable over the range –6 to –18dB/octave, at any chosen (switchable) frequency, is shown in Fig. 6. This consists of a single transistor version of the ‘H’ filter used in the two previous pre-amplifier designs (the nomenclature derives from the shape of the component layout in the ‘op-amp’ form), followed by a very low distortion two-transistor amplifier whose gain can be chosen as required, over the range 5 to 100, by adjustment of Ra and Rb. If a unity-gain stage is all that is required (actually the gain is about 0.9) the output ran be taken from the point marked 'A' on the diagram, and Tr7 and Tr8 omitted.

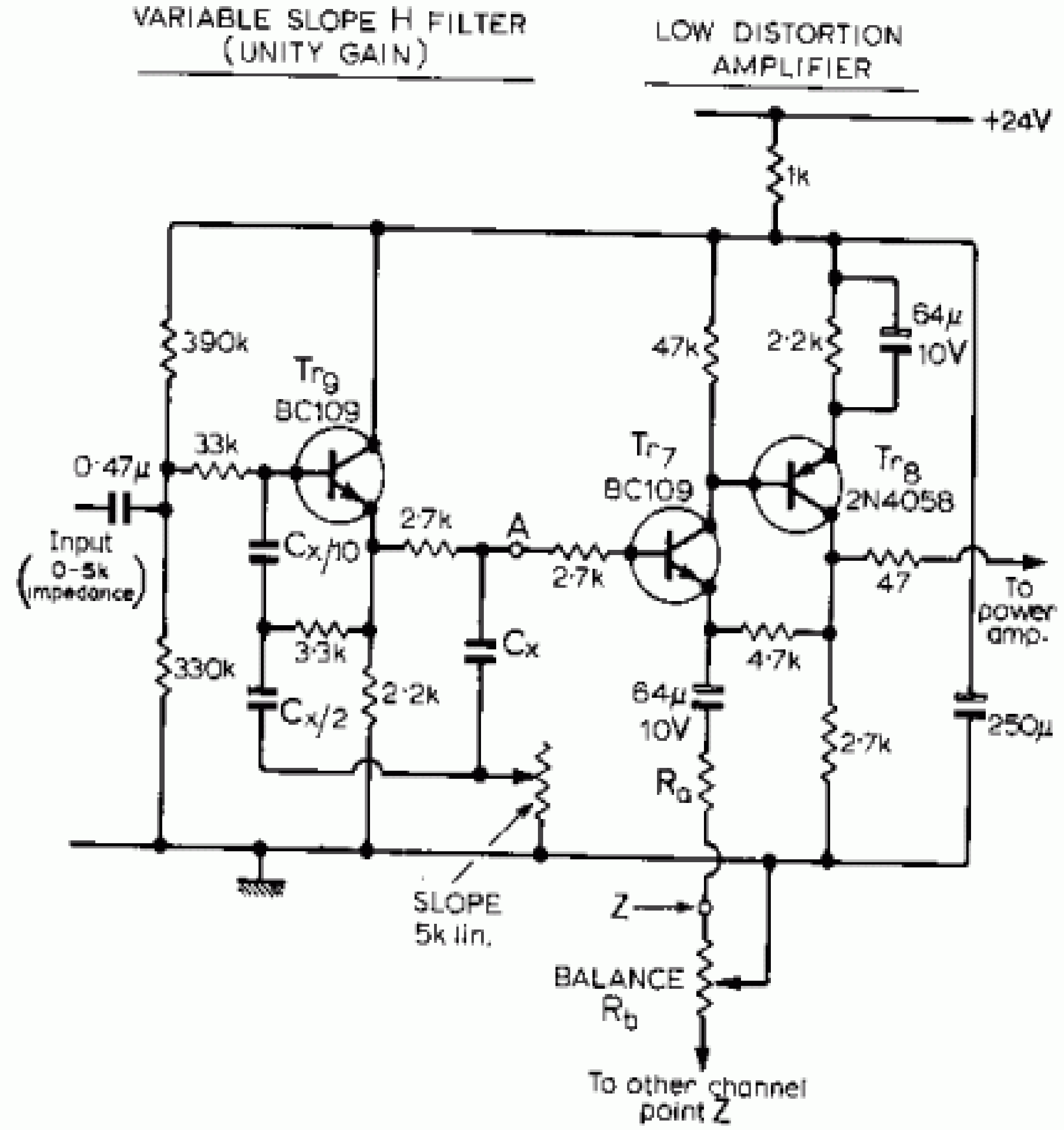


Fig. 6. Pre-amplifier driver stage incorporating a switched low-pass filter with slope variable from 6-18dB/octave. The gain of the filter is variable through the range 5-100 by choice of R_a and R_b.

$$Gain = 1 + \frac{4.7}{R_a + R_b/2}$$

Output can be taken from point A if only unity gain is required.

The response curve of the filter circuit, at any chosen turnover frequency is shown in Fig. 7. The slope is smoothly variable by adjustment to the 5kohm pot. If the slope pot. Is open circuit the response is flat to 20kHz and beyond, but in this case the load impedance should not be less than 50kohm.

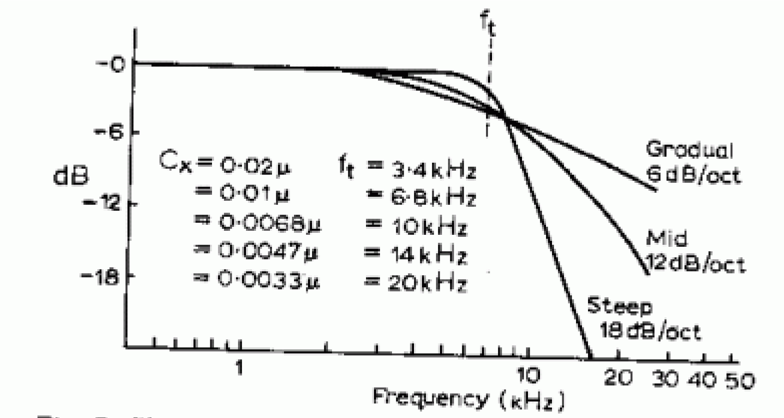


Fig. 7. Slope characteristics and turn-over frequencies of variable-slope 'H' filter.

For completeness, an equivalent single-transistor high-pass filter, having a cut-off slope approaching 18dB/octave, and suitable for use as a 'rumble' filter or a pre-amplifier woofer/tweeter cross-over filter, is shown in Fig. 8. The frequency response characteristics of this filter are shown in Fig. 9. Both of these filter circuits should be driven from a source having a fairly low impedance – not higher than 6kohm.

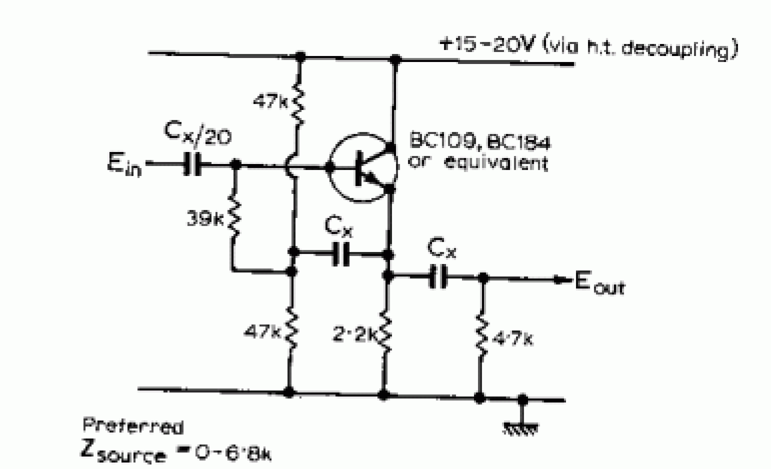


Fig. 8. Single transistor high-pass 'H' filter.

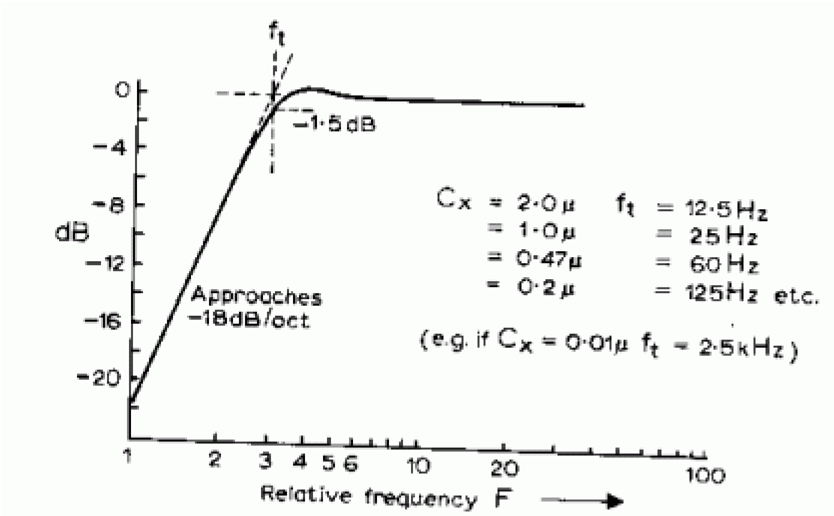


Fig. 9. Transmission characteristic of high-pass 'H' filter.

If single transistor 'H' filters are to be used at output signal levels exceeding 100mV a Darlington transistor, e.g. Motorola MPSA14, is to be preferred.

The apparent noise level, referred to the input, of the two transistor driver amplifiers, using reasonably low noise transistors and an input impedance of the order provided in the normal circuit, is about 4 – 6uV. The output noise voltage in the original circuit was 0.2 – 0.3mV, which should be inoffensive. With a lower gain driver stage this noise will be reduced even further.

The use of a variable negative feedback type of balance control in these circuits is deliberate, in that it permits a low output impedance to be obtained from the driver stage. Measurements made with a wide range of published transistor-operated power amplifiers have shown that substantially lower distortion levels are often given by using a low-impedance drive circuit, and that there is frequently an advantage also in terms of hum, noise, and transient response.

Tone-control circuit. This stage has a worst case (bass and treble controls set to maximum 'lift') distortion level which is typically less than 0.1% at 1V r.m.s. output. It is perfectly capable of driving a normal high-quality power amplifier without the interposition of other pre-amplifier stages. The required signal amplification could then be provided prior to the volume control. This is tending to be the normal practice in commercial 'hi-fi' amplifiers, in that it gives the highly-sought-after zero noise-level at minimum volume control settings, and makes for economies in the use of components.

Noise in the tone-control stage due to the f.e.t. has caused occasional troubles. This should not occur with the f.e.t. now recommended for this part of the circuit (the Amelco 2N4302), which appears to have a consistently low noise level. The necessary bias adjustments were described in a letter to the editor published in April 1970.

The input impedance level suggested for the tone-control stage was 50kohm, because it was thought that most of the other systems likely to be used with this unit would be transistor operated; and this would be a suitable level for this purpose, while avoiding some of the hum pick-up problems likely to be encountered at higher impedance levels. However, if this impedance is too low, and if a high gain (beta greater than 400) transistor is selected for Tr4 – in fact most BC109s will do – the base bias resistors can be increased to 1Mohm and 560kohm (instead of 200kohm and 100kohm) enabling the volume control and auxiliary control potentiometers to be increased to 25kohm.

If an even higher input impedance is required, the f.e.t. impedance conversion shown in Fig. 4 in the original pre-amp article can be substituted in its entirety for Tr4. To preserve the function of the rumble filter in this circuit, with the 0.47uF capacitor desired to feed the tone-control network, a 4.7kohm resistor should be connected from the output side of this capacitor to the earth line. A low noise f.e.t. is of course preferable.

If additional amplification is required on any signal source prior to the tone-control stage (if this is working at the 100mV level) a simple single-transistor feedback amplifier such as that shown in Fig. 10, can be used with confidence, in that its performance is stable, its noise is low, it is almost impossible to damage by an input overload, and its distortion is well below 0.1% at output voltages up to 0.25V r.m.s., and with gains up to 10.

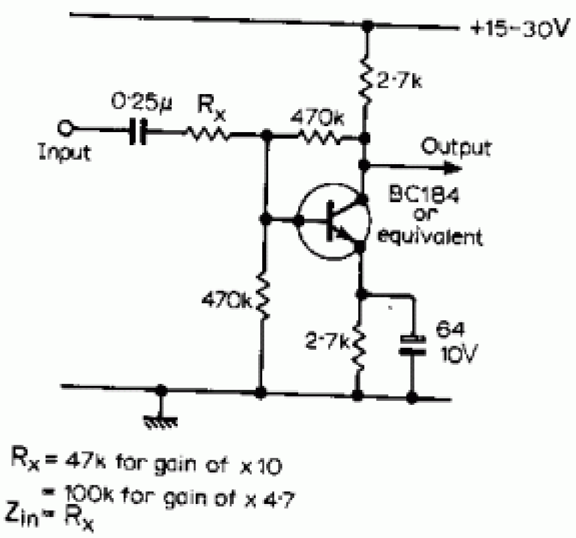


Fig. 10. Simple low-distortion single transistor amplifier.

Magnetic pickup equalisation circuit. Some requests have been received for component values for the use of this circuit for tape-replay characteristic equalization. The author remains of the opinion that this type of provision is best left to the manufacturers of the tape recorder, in that the actual head characteristics can influence the replay frequency/voltage characteristics.

However, a fairly close approximation to the replay curve theoretically required for 7.5 i.p.s. is given if C2 and R2 in the original equalization network A are altered to I00pF and 27kohm.

The noise level of this circuit is almost entirely determined by the performance of Tr1 The BC184C and 2N5089 transistor types may be of interest in this position.

The maximum output which can be obtained from this circuit at 0.02% t.h.d., is 2V r.m.s. If the normal input to the tone control circuit, or other following stages, is l00mV, this gives a 26dB overload capability. The gain of the equalization circuit can be increased by a factor of 3, (i.e. to 30 at 1kHz) without upsetting the rumble filter characteristics if R5 is reduced to 68ohm and C4 increased to I00uF.

Miscellaneous. An omission from the original article was the suggestion that high value resistors (2 – 5Mohm) should be connected across the switch contacts, from slider to each Cx. This removes 'plops' on switching ranges.

A number of correspondents have queried the need for a separate h.t. power supply for the pre-amp. (The reservoir capacitors for the unit shown should have read 35V working, not 25V). It is always possible to run the pre-amp via a suitable voltage-dropper circuit from the main amplifier power supply and if a zener diode is included in this line, this scheme may be satisfactory. However, measurements on channel separation and harmonic and i.m. distortion, with identical amplifier systems invariably show some advantage, particularly at the low-frequency end of the audible spectrum, in the use of a separate power supply for the pre -amp (even when the electrolytic bypass capacitors are still new) and this arrangement is still recommended by the author as well worth the small additional cost.

One point which has not been published, to the best of the author's knowledge, concerns the particular advantage conferred by the feedback pair amplifier using complementary transistors, such as that used in the low-pass filter circuit, in comparison with the more usual n-p-n/n-p-n pair, where the bias for the first transistor is derived from the h.t. line. In the case of the n-p-n/p-n-p pair, any h.t. line feedback, due to inadequate h.t. line bypass, will be negative rather than positive, and this can assist in obtaining good t.h.d. figures down to low signal frequencies.

My thanks go to Malcolm Jenkins for providing the copy of the article used to make this web page and to Lynn Miller for converting it into web format.

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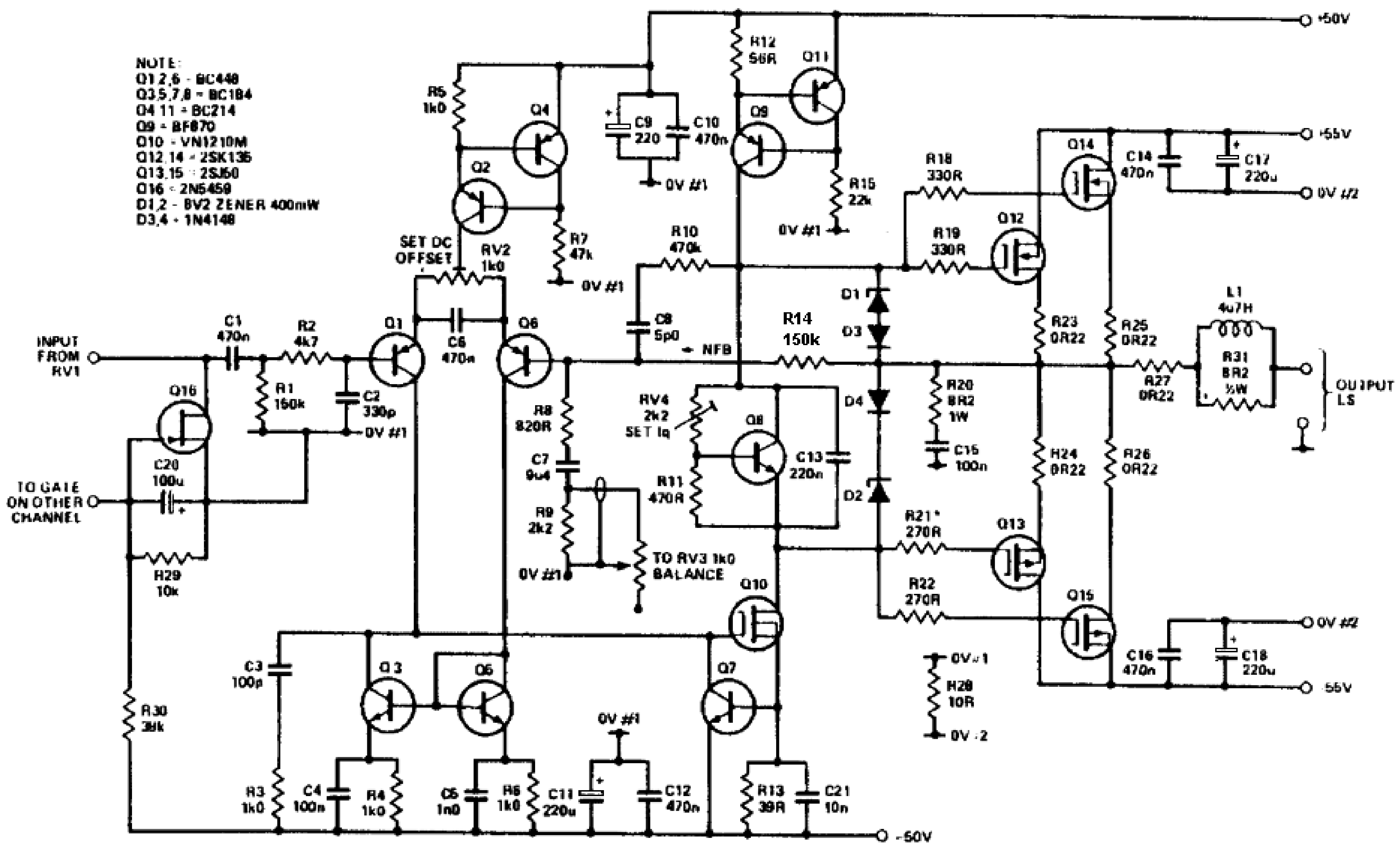
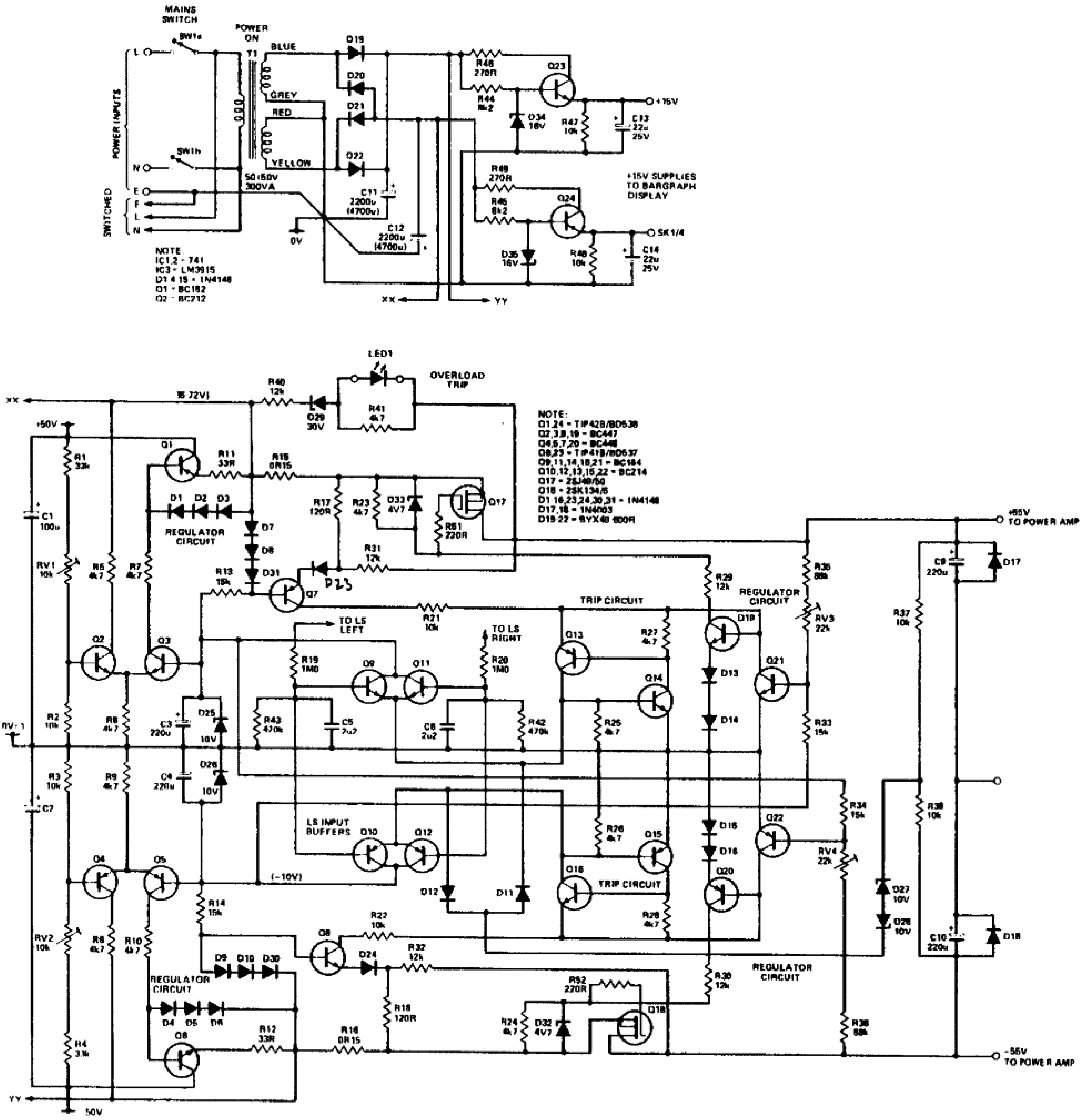
HISTORY: Page created 06/01/2002
13/01/2002 Hi-res Fig. 1 and December 1970 postscript added

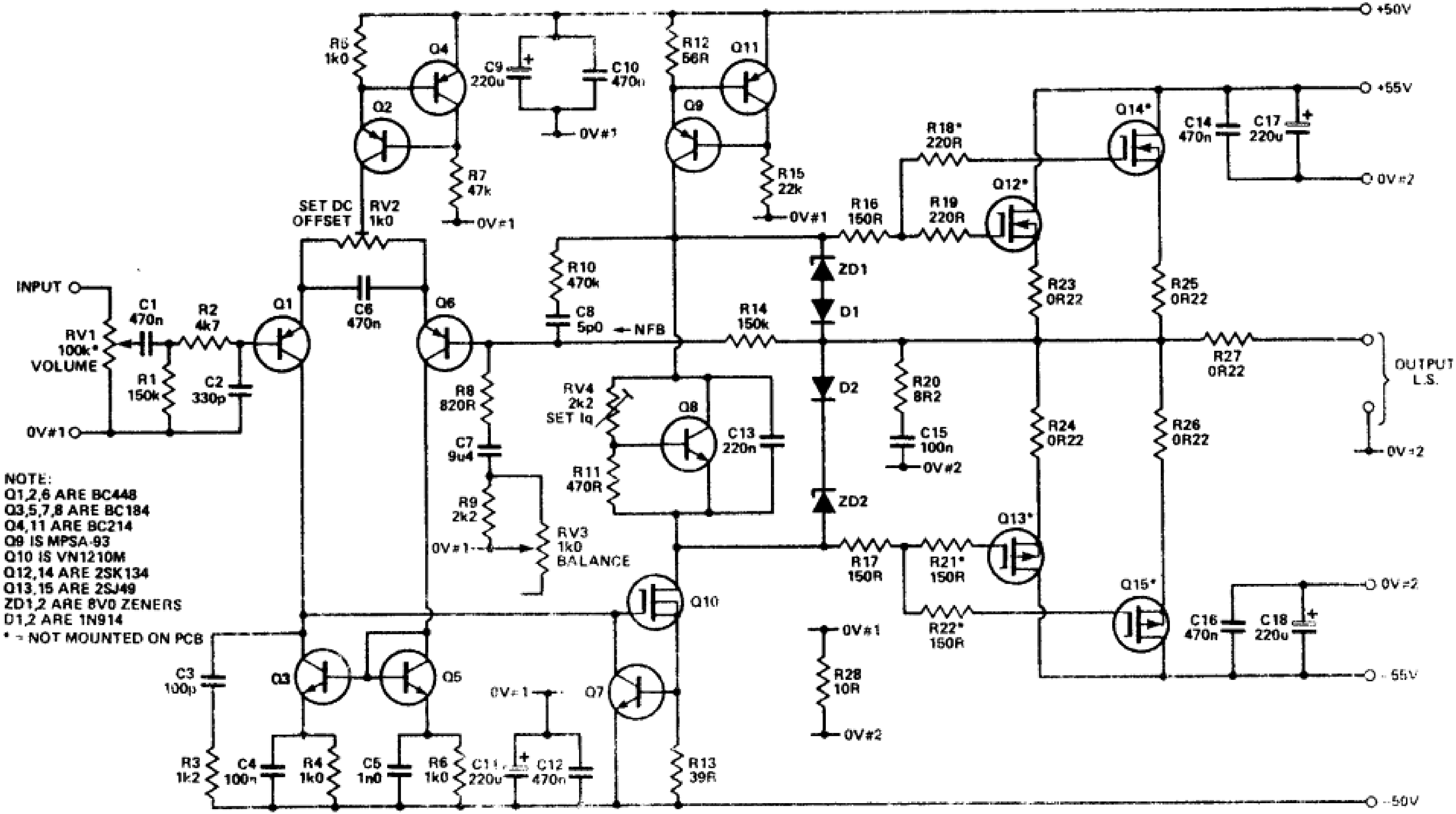
Audio Design 80W MOSFET Audio Amplifier - ETI May 1989



Power Supply for 'Audio Design 80W MOSFET Audio Amplifier' - ETI May 1989

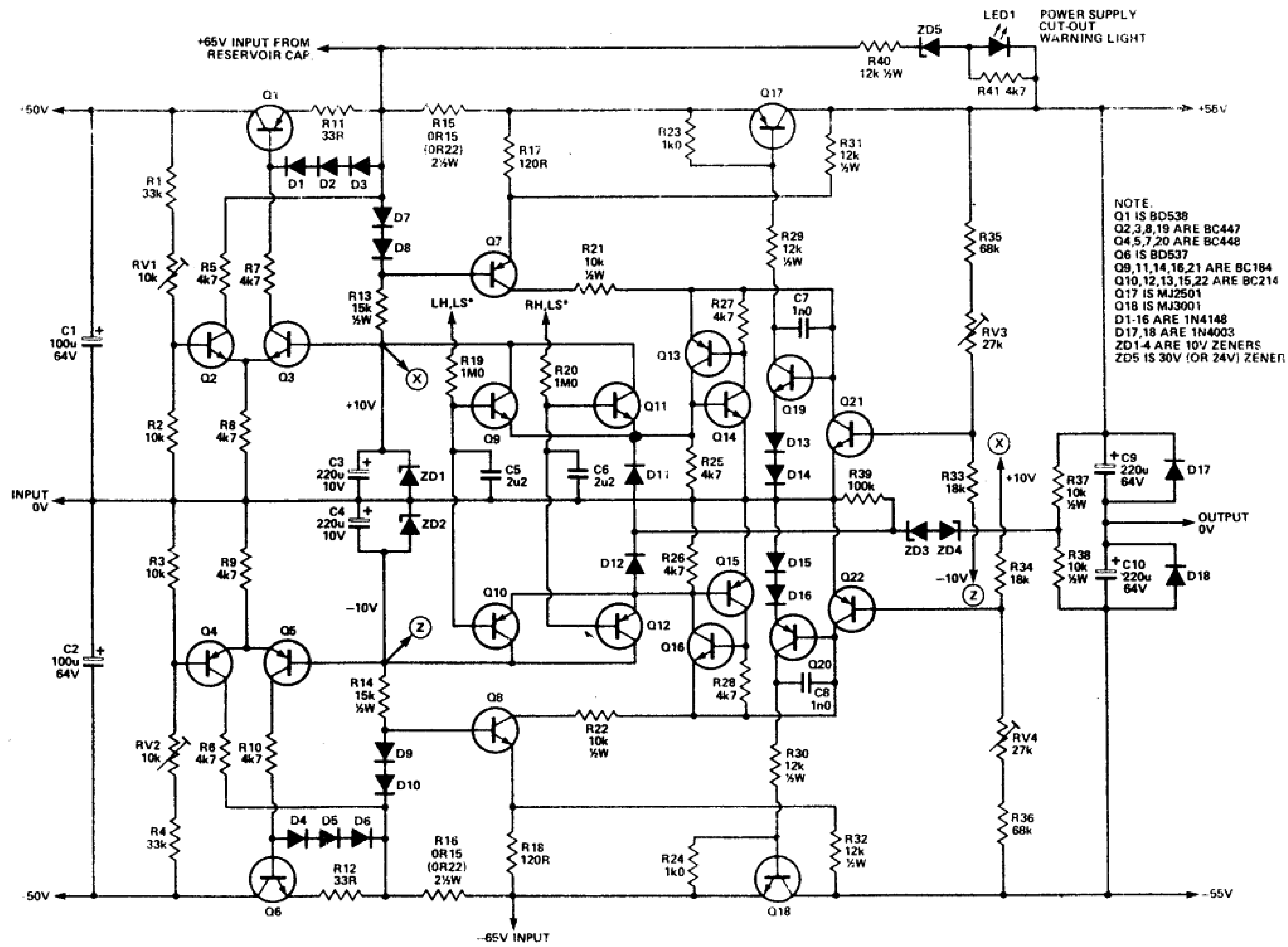


Audio Design Amplifier - ETI July 1984

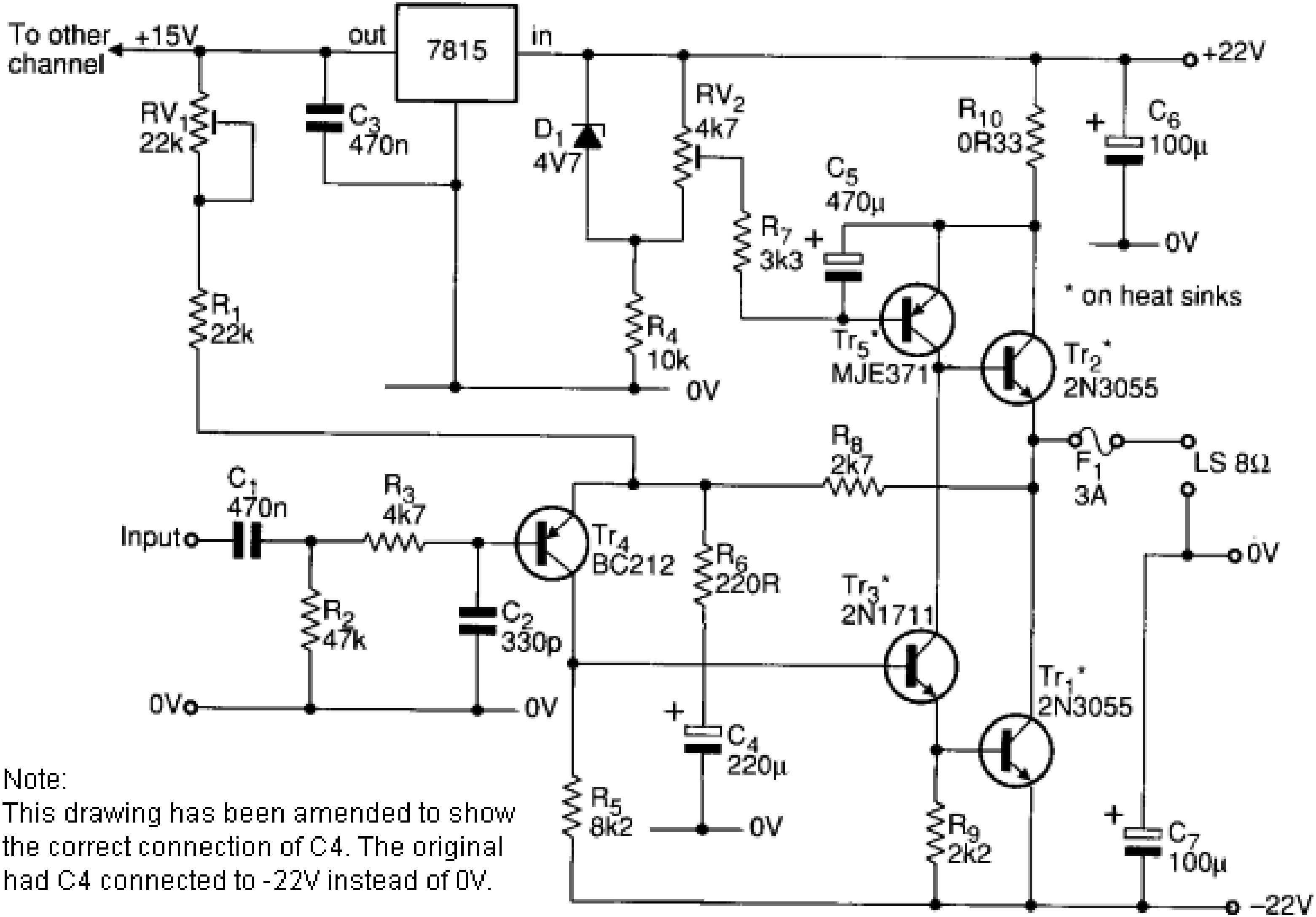


NOTE:
Q1,2,6 ARE BC448
Q3,5,7,8 ARE BC184
Q4,11 ARE BC214
Q9 IS MPSA-93
Q10 IS VN1210M
Q12,14 ARE 2SK134
Q13,15 ARE 2SJ49
ZD1,2 ARE 8V0 ZENERS
D1,2 ARE 1N914
* = NOT MOUNTED ON PCB

Power Supply for ' Audio Design Amplifier' - ETI August 1984

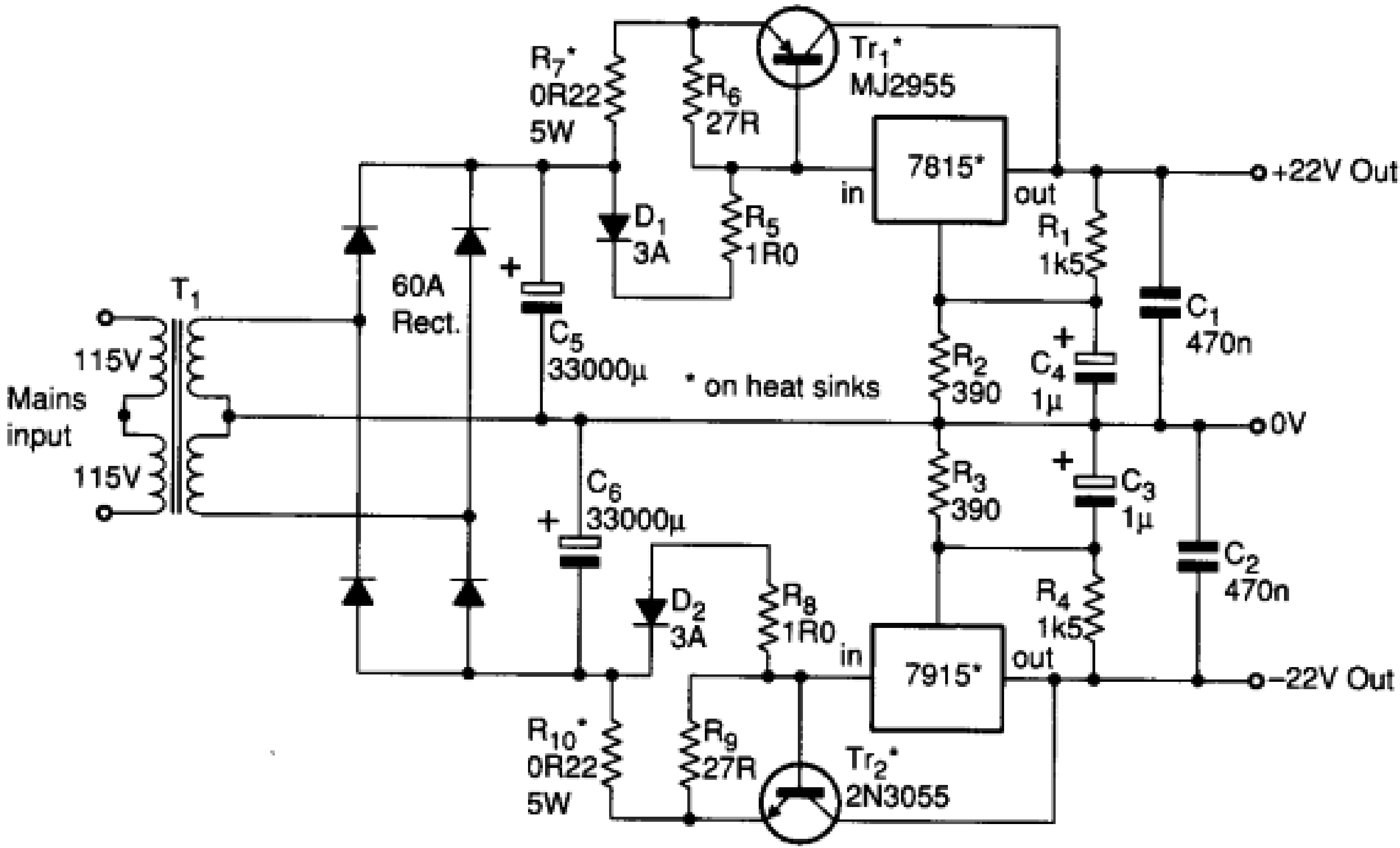


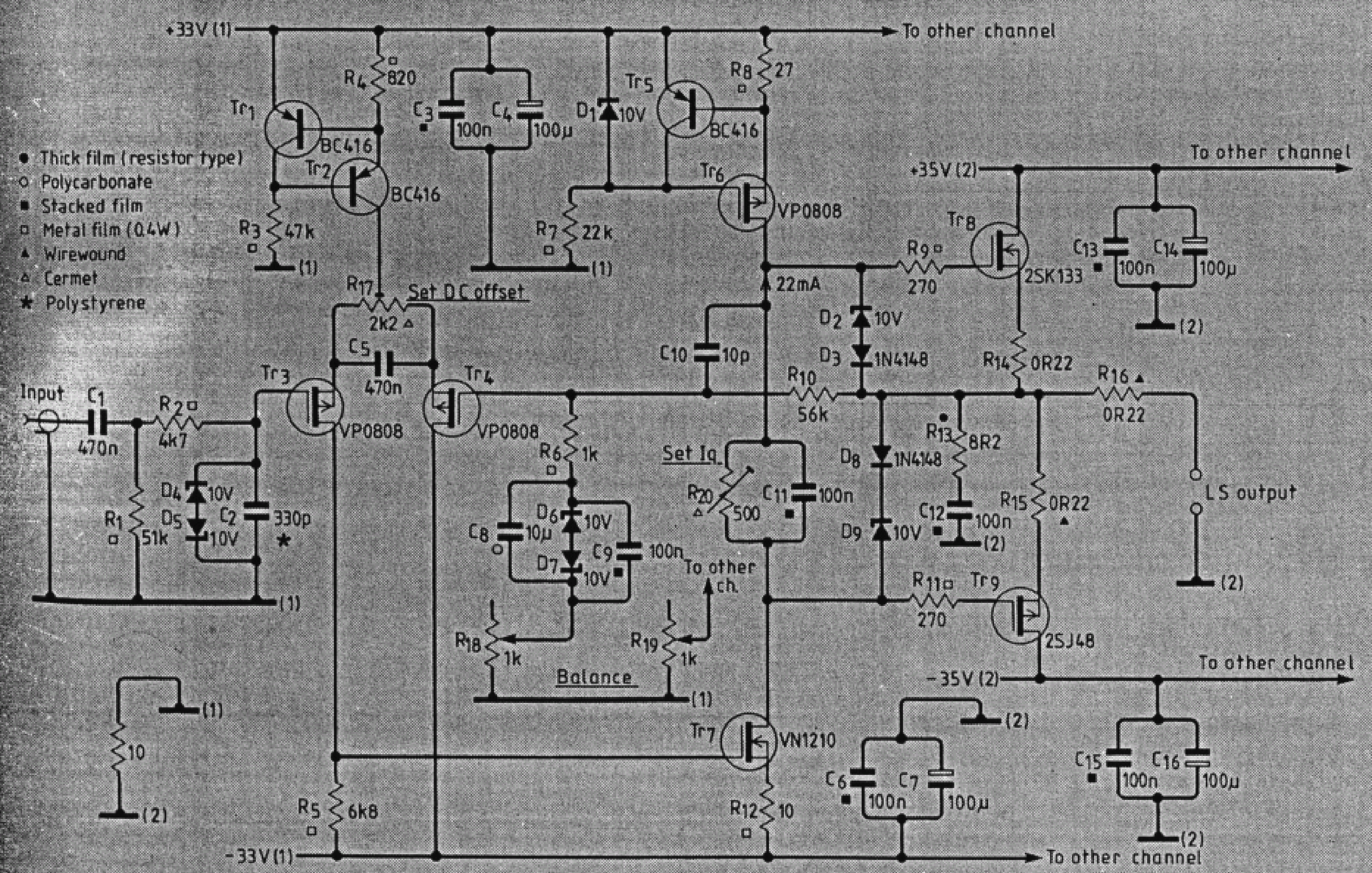
Class-A Power - Electronics World September 1996



Note:
This drawing has been amended to show the correct connection of C4. The original had C4 connected to -22V instead of 0V.

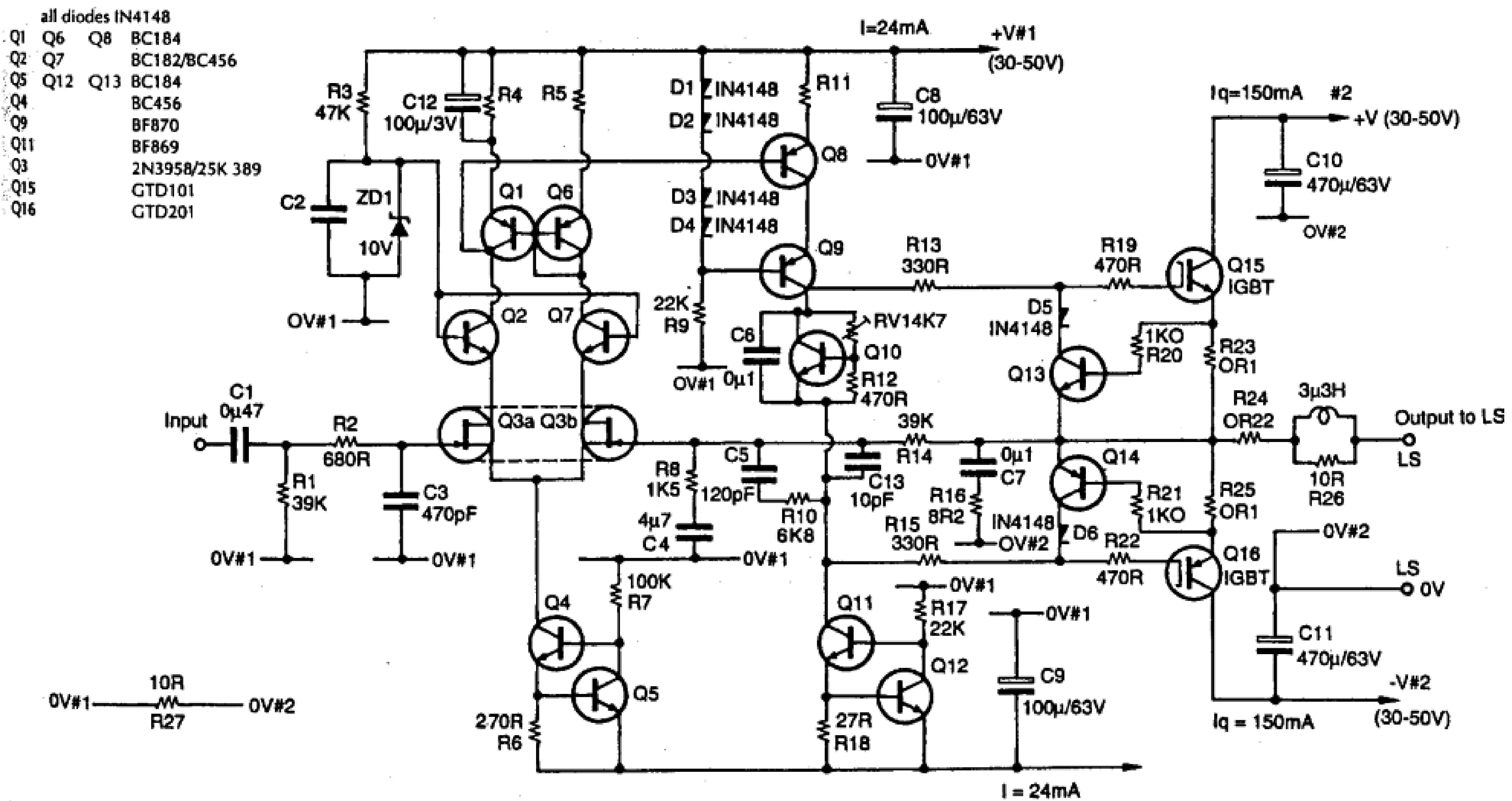
Power Suply for 'Class-A Power' - EW September 1996



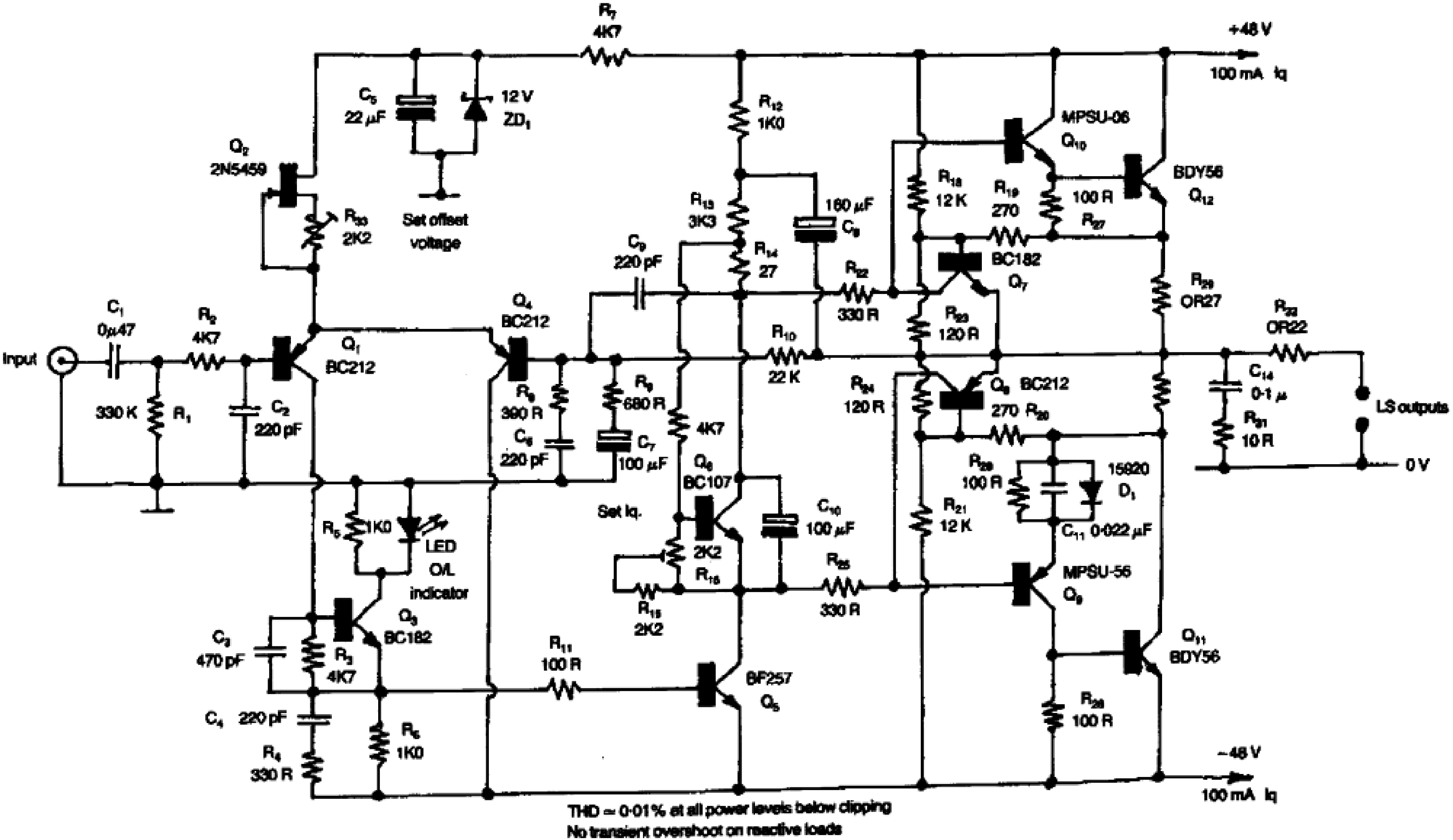


Class A/AB MOSFET Power Amplifier - Electronics & Wireless World March1989

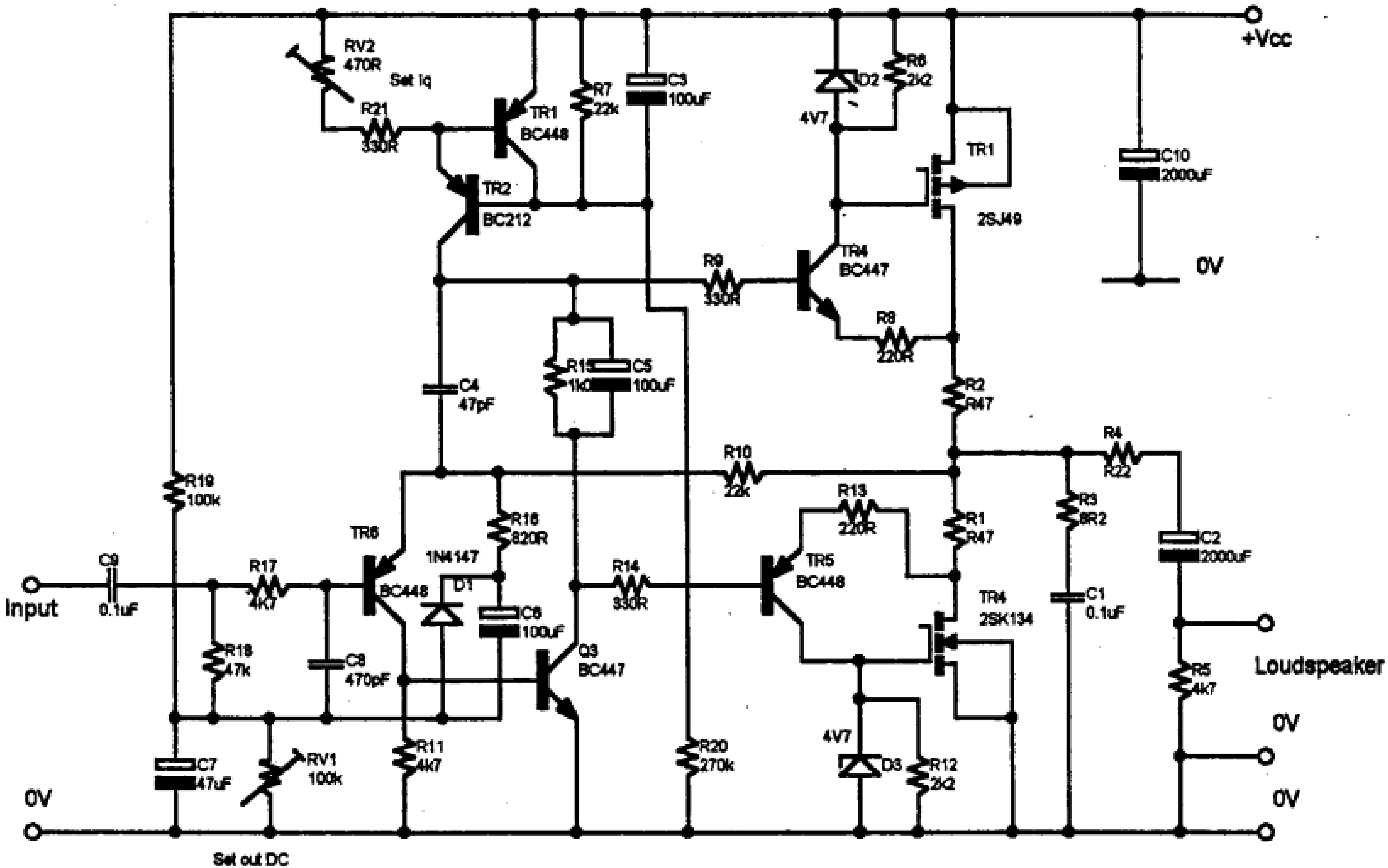
IGBT Audio Amplifier - Electronics World & Wireless World May 1992



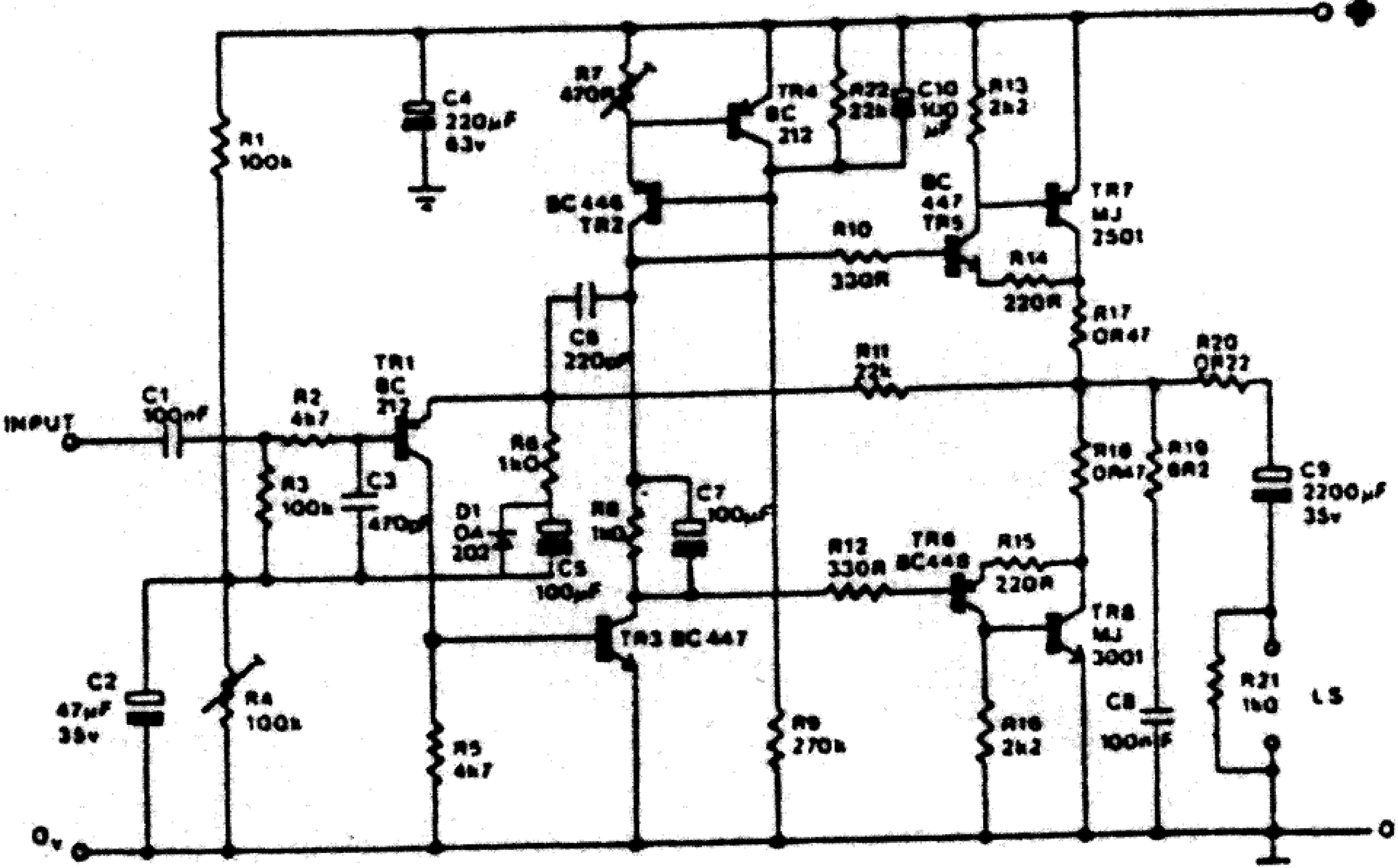
A Direct-Coupled High Quality Stereo Amplifier - Hi-Fi News November 1972

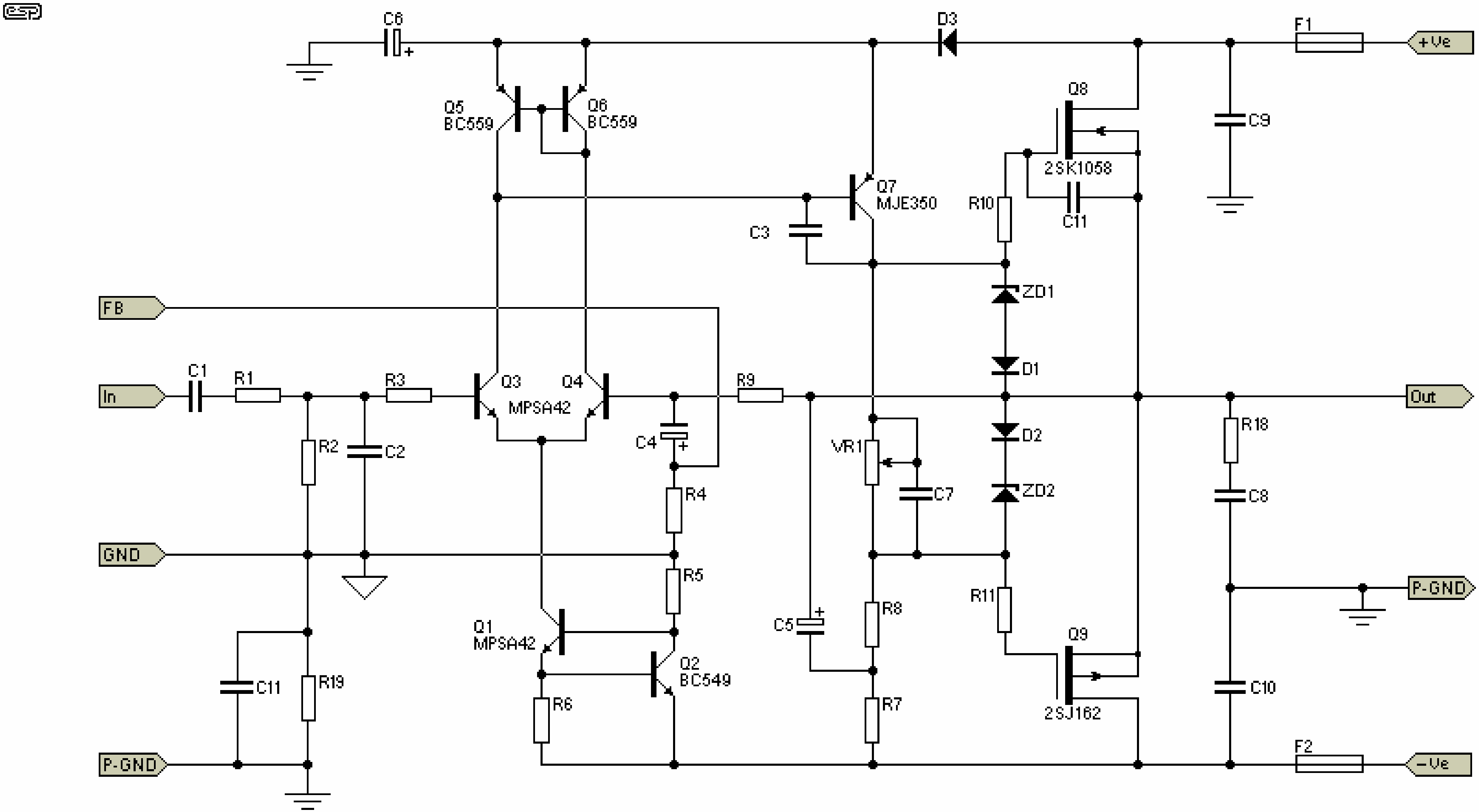


An Introduction to Power MOSFETS - HFN December 1980
(An update to the 'Simple 30 Watt Integrated Amplifier')



A Simple 30 Watt Integrated Amplifier - Hi-Fi News January 1980





ESP
C6
D3
F1
+Ve
Q5
BC559
Q6
BC559
Q8
C9
2SK1058
Q7
MJE350
R10
C11
C3
ZD1
FB
D1
C1
R1
R3
Q3
Q4
MPSA42
R9
In
Out
C4
VR1
D2
R18
R2
C2
R4
C7
ZD2
C8
GND
R5
R11
P-GND
C5
R8
Q1
MPSA42
Q9
Q2
BC549
C11
R19
R6
R7
C10
2SJ162
P-GND
F2
-Ve

Simple Class A Amplifier - Wireless World April 1969

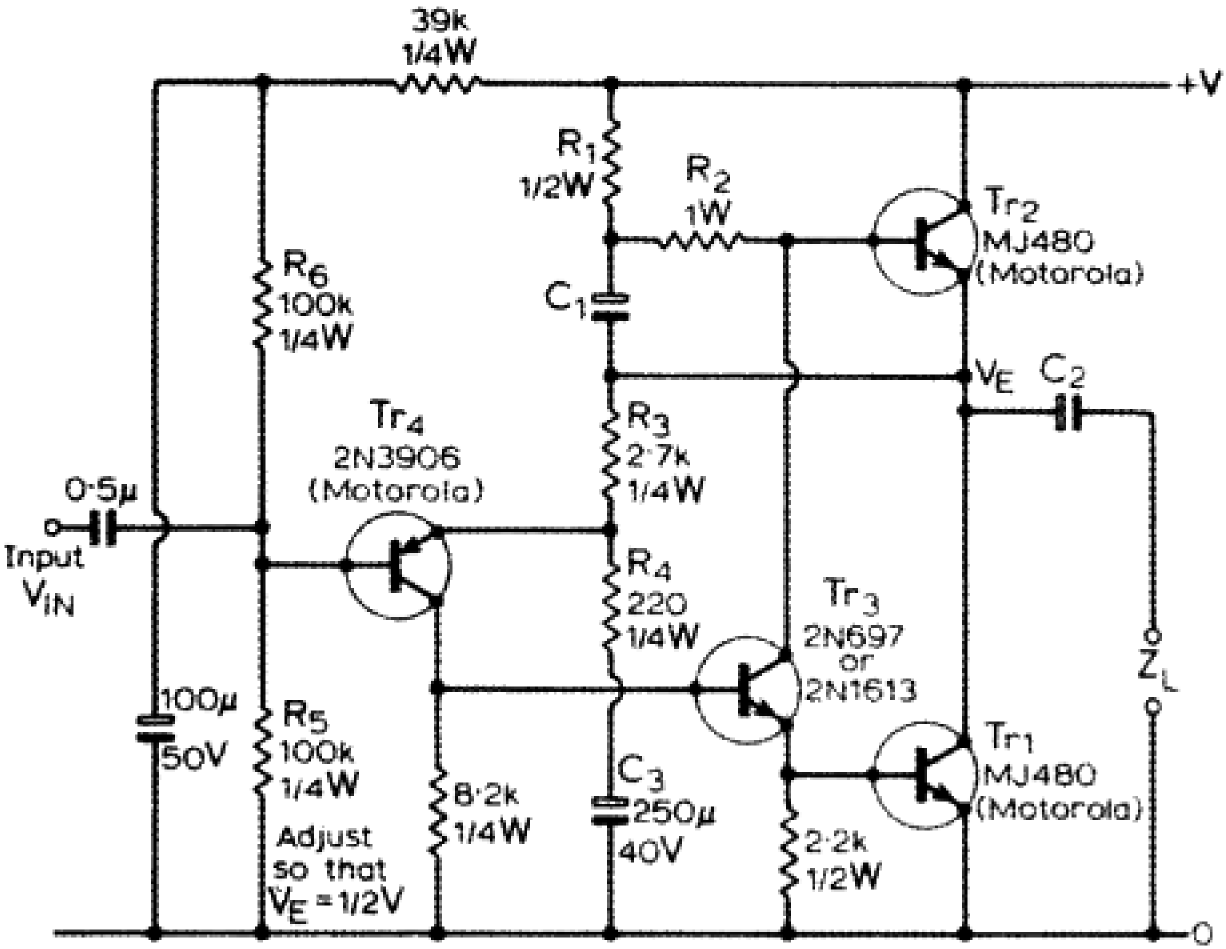


Fig. 3. Practical power amplifier circuit.

Table 1: Summary of component combinations for different load impedancices

Z_L	V	I	R_1	R_2	C_1	C_2	V_{IN} (r.m.s)
3Ω	17V	2A	47Ω	180Ω	500µ25V	5000µ25V	0.14V
8Ω	27V	1.2A	100Ω	560Ω	250µ40V	2500µ50V	0.66V
15Ω	36V	0.9A	150Ω	1.2kΩ	250µ40V	2500µ50V	0.9V

Power Supply for 'Simple Class a Amplifier - WW April 1969

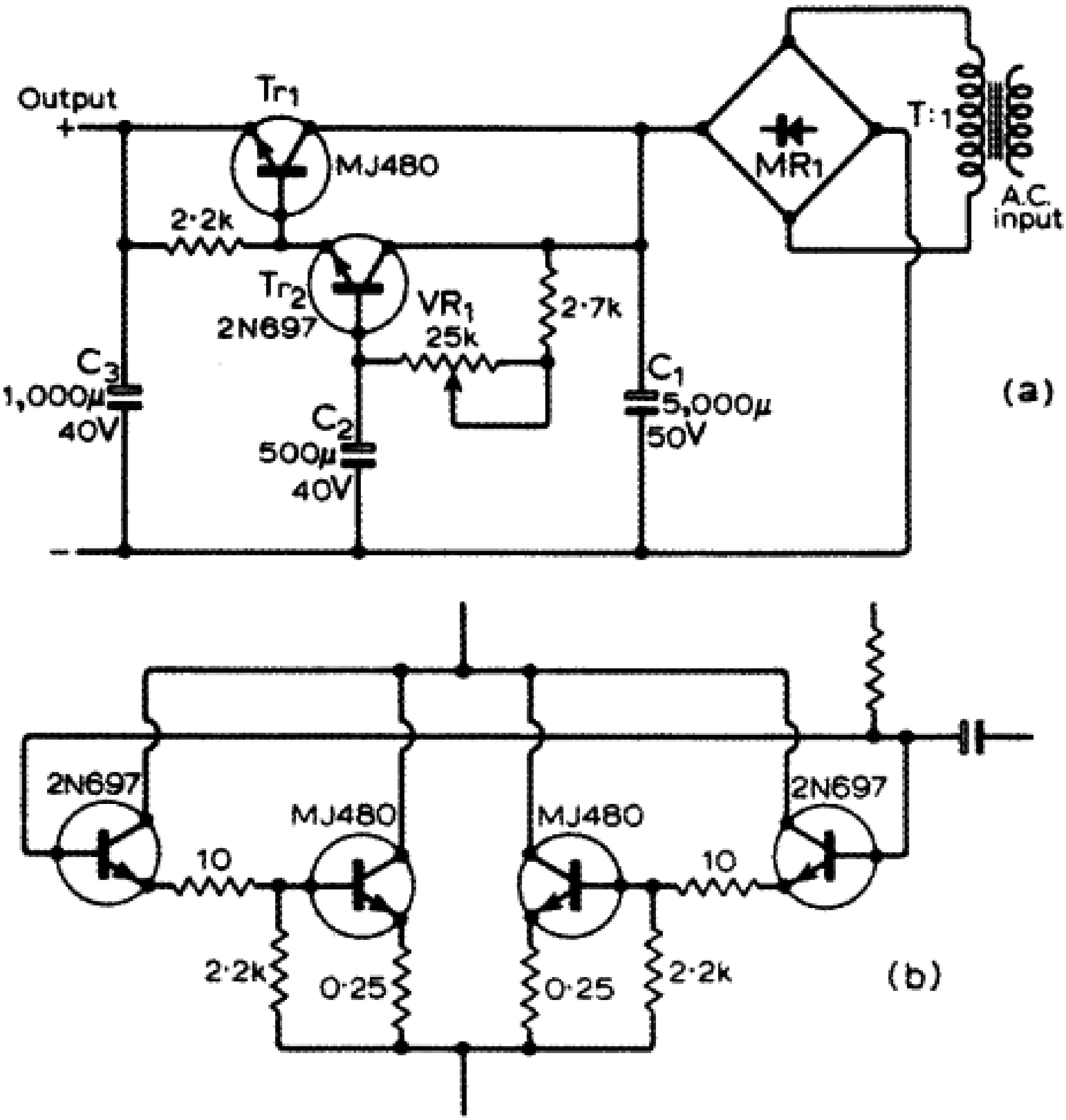


Fig. 9. (a) Power supply unit, and (b) parallel connected transistors for high currents.

15-20W Class AB Audio Amplifier - Wireless World July 1970

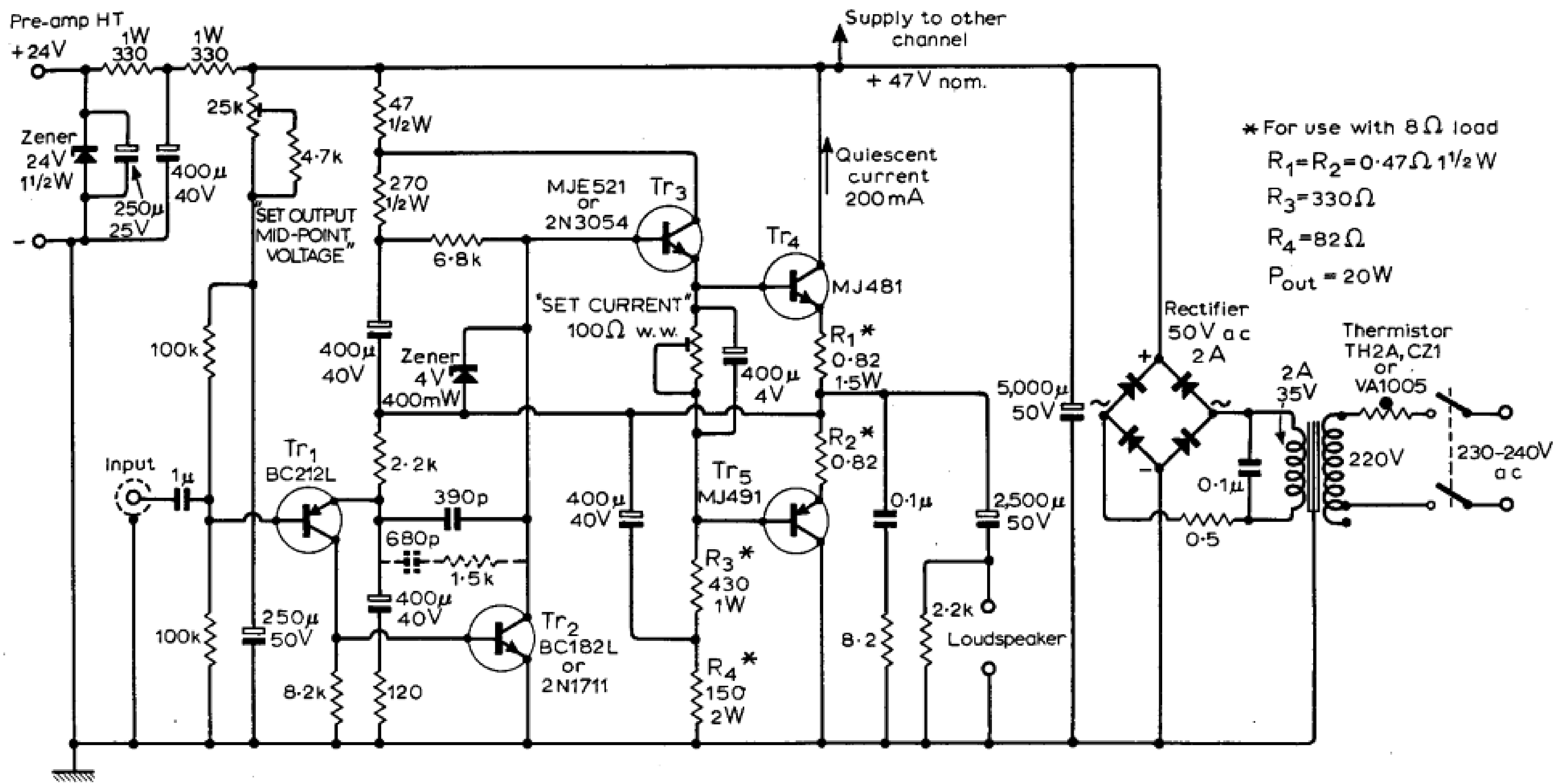


Fig. 3. Power amplifier circuit. The dotted components (680pF, 1.5kΩ) can be added if electrostatic speakers are used.

15-20W Class AB Audio Amplifier - Wireless World July 1970
Alternative stabilized power supply

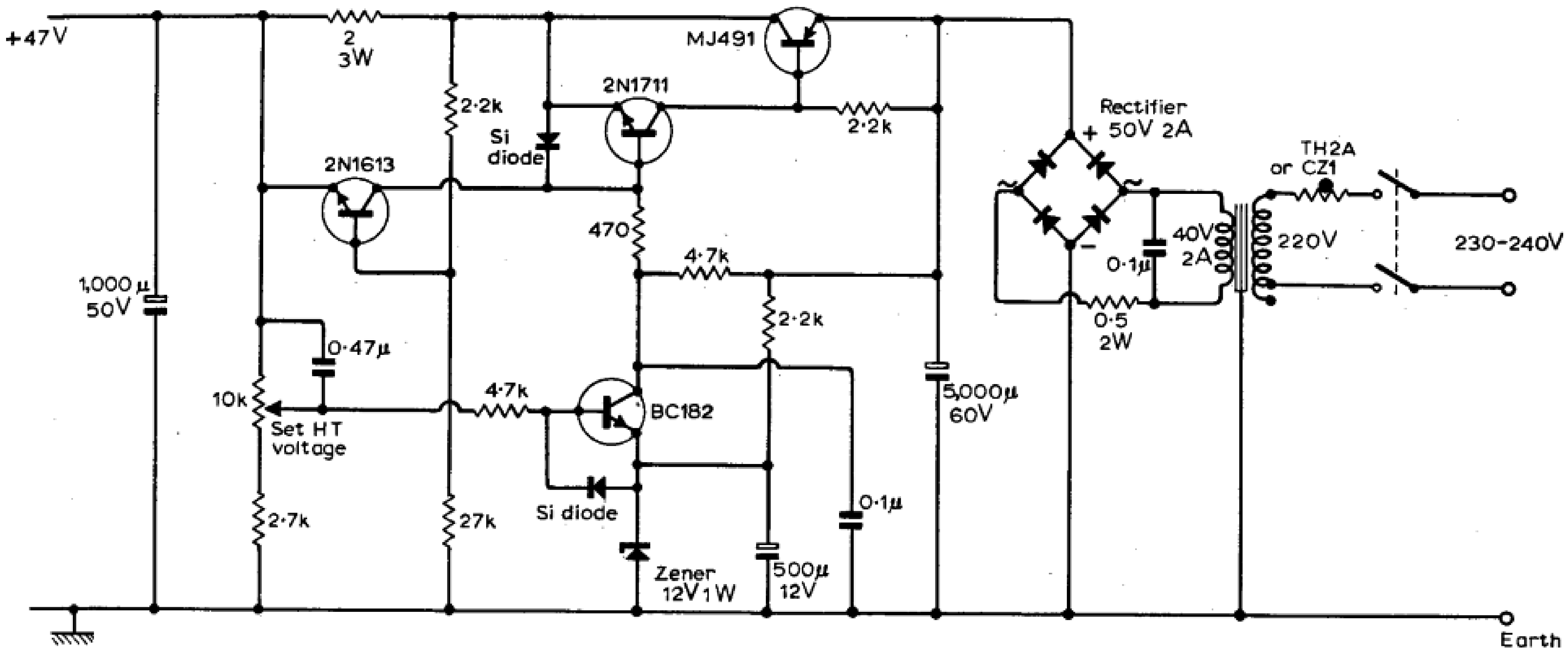


Fig. 10. Stabilized power supply with re-entrant short-circuit protection (12-49V).

80-100W MOSFET Audio Amplifier - Wireless World August 1982