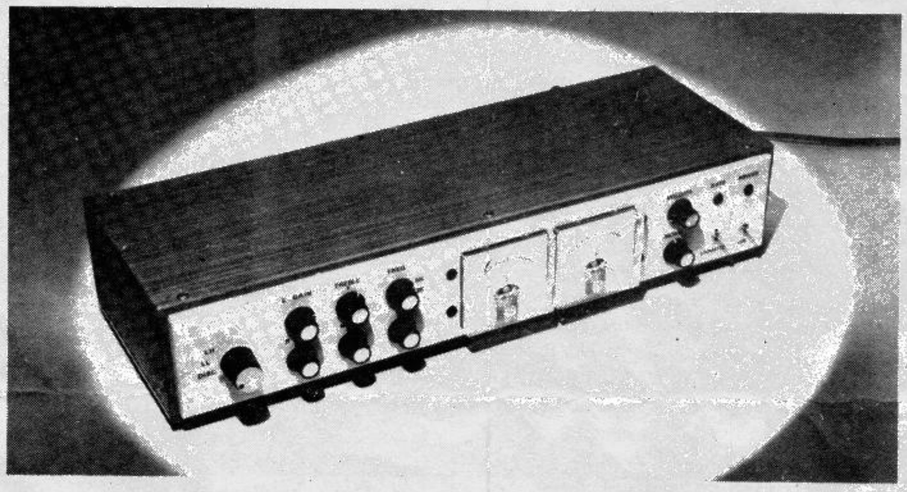
Advanced preamplifier design

A no-compromise circuit with noise gating

by D. Self, B.A., *Electrosonic Ltd*

This preamplifier design offers a distortion figure of below 0.002%, an overload margin of around 47dB, and a signal-to-noise ratio of about 71dB for the disc amplifier. A novel noise gate mutes the output when no signal is presented to the disc input and conversely, by using the subsonic information present on record pressings, eliminates the problem of muting low level signals.



This article describes a stereo pre-amplifier that equals or exceeds the performance of many of those available. The circuit incorporates a novel method of muting the signal path, when the disc input is quiescent, by using a noise gate that never mutes a wanted low-level signal.

Many of the important performance factors, such as signal-to-noise ratio, overload margin, and accuracy of the RIAA equalization, are essentially defined by the design of the disc input circuitry. This therefore merits close attention. The best attainable s/n ratio for a magnetic cartridge feeding a bipolar transistor stage with series feedback is about 71dB with respect to a 2mV r.m.s. input at 1kHz, after RIAA equalization. This has been clearly demonstrated by Walker[1]. The equivalent amplifier stage with shunt feedback gives an inferior noise performance over most of the audio band due to the rise in cartridge source impedance with frequency. This limits the maximum s/n ratio after equalization to about 58dB. These facts represent a limit to what the most advanced disc input stage can achieve.

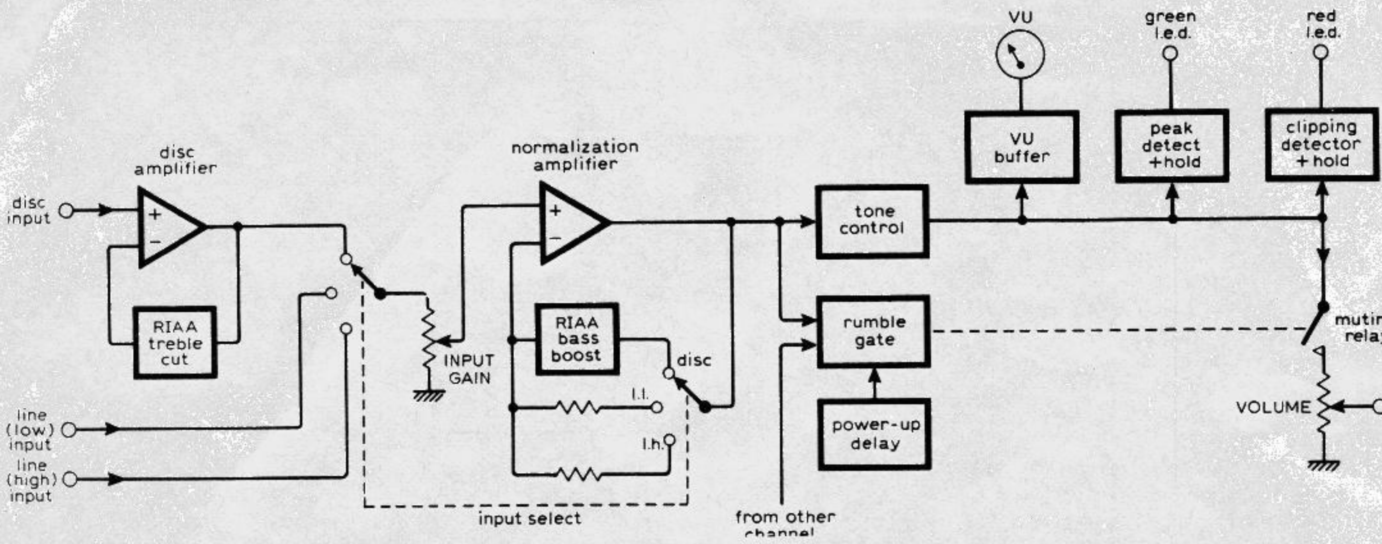
Overload margin appears to be receiving little attention. The maximum velocities recorded on disc seem to be steadily increasing and this, coupled with improved cartridges, means that very high peak voltages are reaching disc inputs. Several writers have shown that short-term voltages of around 60 to 80mV r.m.s. are possible from modern discs and cartridges, and higher values are to be expected. This implies that to cater for signal maxima, a minimum overload margin of 32dB with respect to 2mV r.m.s. at 1kHz is essential. Obviously a safety factor on top of this is desirable. However, most pre-amplifiers at the top end of the market provide around 35-40dB only. There are certain honourable exceptions such as the Technics SU9600 control amplifier which achieves an overload margin of 54dB, mainly by the use of a staggeringly high supply of 136V in the disc input amplifier. The Cambridge P50/110 series offers a margin in excess of 60dB by the artifice of providing unity-gain buffering, for correct cartridge loading, but no amplification before the main gain control. This allows the use of an 18V supply rail, but does limit the maximum s/n ratio.

The overload margin of a pre-amplifier is determined by the supply voltage which sets the maximum voltage swing available, and by the amount of amplification that can be backed-off to prevent overload of subsequent stages. Most pre-amplifiers use a relatively high-gain disc input amplifier that raises the signal from cartridge level to the nominal operating level in one jump. Low supply voltages are normally used which reduce static dissipation and allow the use of inexpensive semiconductors. The gain control is usually placed late in the signal path to ensure low-noise output at low volume settings. Given these constraints, the overload performance is bound to be mediocre, and in medium-priced equipment the margin rarely exceeds 30dB. If these constraints are rejected, the overload margin of the system can be improved.

Two separate gain controls remove the most difficult compromise, which is the placement of the volume control. This approach is exemplified in the Radford ZD22 and the Cambridge P60 circuitry. One gain control is placed early in the signal path, preceded by a modest amount of gain. Cartridges of high output can be accommodated by the use of this first control. The second is placed late in the pre-amplifier and is used as a conventional volume control, see Fig. 1.

The other performance criterion which is largely defined by the disc input circuitry is frequency response, as defined by the accuracy of the RIAA equalization. Assuming that the relevant amplifying stage has sufficient open-loop gain to cope with the bass boost required, the accuracy of the equalization depends entirely on the time constants within the feedback loop. Careful design, and the use of close-tolerance components can assure an accurate response to within ±0.2dB from 30Hz to 20kHz.

Pre-amplifier distortion seems to have received little attention compared with that generated by power amplifiers, perhaps because the former has



traditionally been much lower. However, power amplifiers, with such low t.h.d. that the residual harmonics can no longer be extracted from the noise at normal listening levels, are now commonplace, particularly with the advent of techniques such as current dumping. This desirable state of affairs unfortunately does not extend to pre-amps, which in general produce detectable distortion at nominal operating levels, usually between 0.02% and 0.2%. In this design the t.h.d. at 1kHz is less than 0.002% even at 25dB above the nominal operating level of 0dBm. A Sound Technology 1700A distortion measurement system was used during development.

At this point it is convenient to consider the noise gate principle. When the pre-amplifier is being used for disc reproduction the output from each channel is continuously sampled to determine if a signal is present; if nothing is detected within a specified time interval, dependent on the previous signal levels received, the pre-amplifier is muted by the opening of a reed relay in series with the output signal path. This allows only power amplifier noise to reach the loudspeakers and considerably reduces the perceived noise generated by a quiescent sound system. Noise in the quiescent state is particularly noticeable when headphones are in use. The reed relay is also used to prevent switch-on transients from reaching an external power amplifier. So far this circuit appears to be a fairly conventional noise gate. The crucial difference is that signals from disc that have not been subjected to rumble filtering are always accompanied by very low frequency signals generated by record ripples and small-scale warps. Even disc pressings of the highest quality produce this subsonic information, at a surprisingly high level, partly due to the RIAA bass boosting. The l.f. component is often less than 20dB below the total programme level but this is quite sufficient to keep the pre-amplifier unmuted for the duration of a l.p. side.

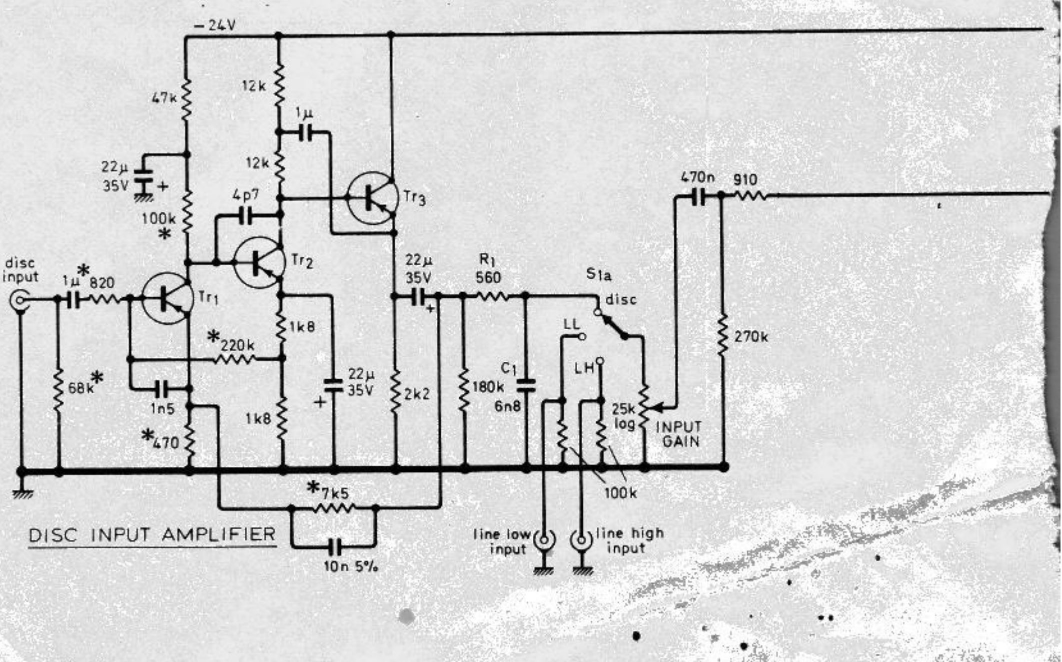
Fig. 1. Block diagram of the complete circuit. Two gain controls are used in the signal path to allow a substantial increase in overload margin.

The pre-amplifier is unmuted as soon as the stylus touches the disc, and muted about a second after it has been raised from the run-out groove. This delay can be made short because the relative quiet at the start of the run-out groove is sensed and stored. The rumble performance of the record deck is largely irrelevant because virtually all of the subsonic information is generated by disc irregularities.

Audio circuitry

A detailed block diagram of the pre-amplifier is shown in Fig. 1, and Fig. 2 shows the main signal path. The disc input amplifier uses a configuration made popular by Walker, but the collector load of the second transistor is bootstrapped. This increases the open-loop gain and hence improves the closed-loop distortion performance by a factor of about three to produce less than 0.002% at an output of 6.5V r.m.s. (1kHz). This stage gives a s/n ratio (ref 2mV) of about 70dB and a gain of 15 at 1kHz. This is sufficient to ensure that the noise performance is not degraded by subsequent stages of amplification. The maximum output of this stage before clipping is about 6.5V r.m.s. and the nominal output is 30mV r.m.s. Because this is the only stage before the input gain control, these two figures set the overload margin at 47dB. To ensure that this overload margin is maintained at high frequencies, the treble-cut RIAA time-constant is incorporated in the feedback loop. This leads to slightly insufficient cut at frequencies above 10kHz because the gain of the stage cannot fall below unity, and hence fails to maintain the required 6dB/octave fall at the top of the audio spectrum. This is exactly compensated for outside the feedback loop by the low-pass filter R_1 C_1, which also helps to reject high frequencies above the audio band.

Fig. 2. Circuit diagram of the signal path. Constant-current sources are biased from a l.e.d./resistor chain for improved thermal stability.



For convenience I have referred to the next stage of the circuit as the normalization amplifier because signals leaving this should be at the nominal operating level of 0dBm by manipulation of the input gain controls. Separate controls are provided for each channel to allow stereo balance. A later ganged control is used for volume setting and causes no operational inconvenience. In the disc replay mode, the normalization amplifier provides the RIAA bass boost, by the feedback components $R_{2,3}$ and C_2. Two line inputs are also provided; line low requiring 30mV and line high 100mV to give 0dBm from the normalization stage with the input gain control fully advanced. When these inputs are selected, the feedback networks are altered to adjust the gain and give a flat frequency response. Ultrasonic filters are incorporated to ensure stability and aid r.f. rejection. Capacitor C_3 in the feedback arm reduces the gain to unity at d.c. for good d.c. stability. If a fault causes the amplifier output to saturate positively the capacitor is protected by a diode which has no effect on the distortion performance.

The circuitry of the normalization amplifier is complicated because its performance is required to be extremely high. The harmonic distortion is far below 0.002% at the maximum output of

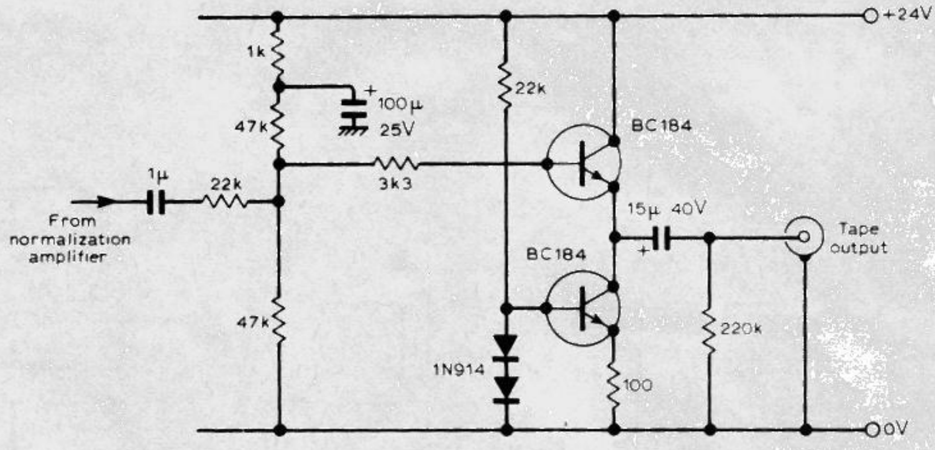
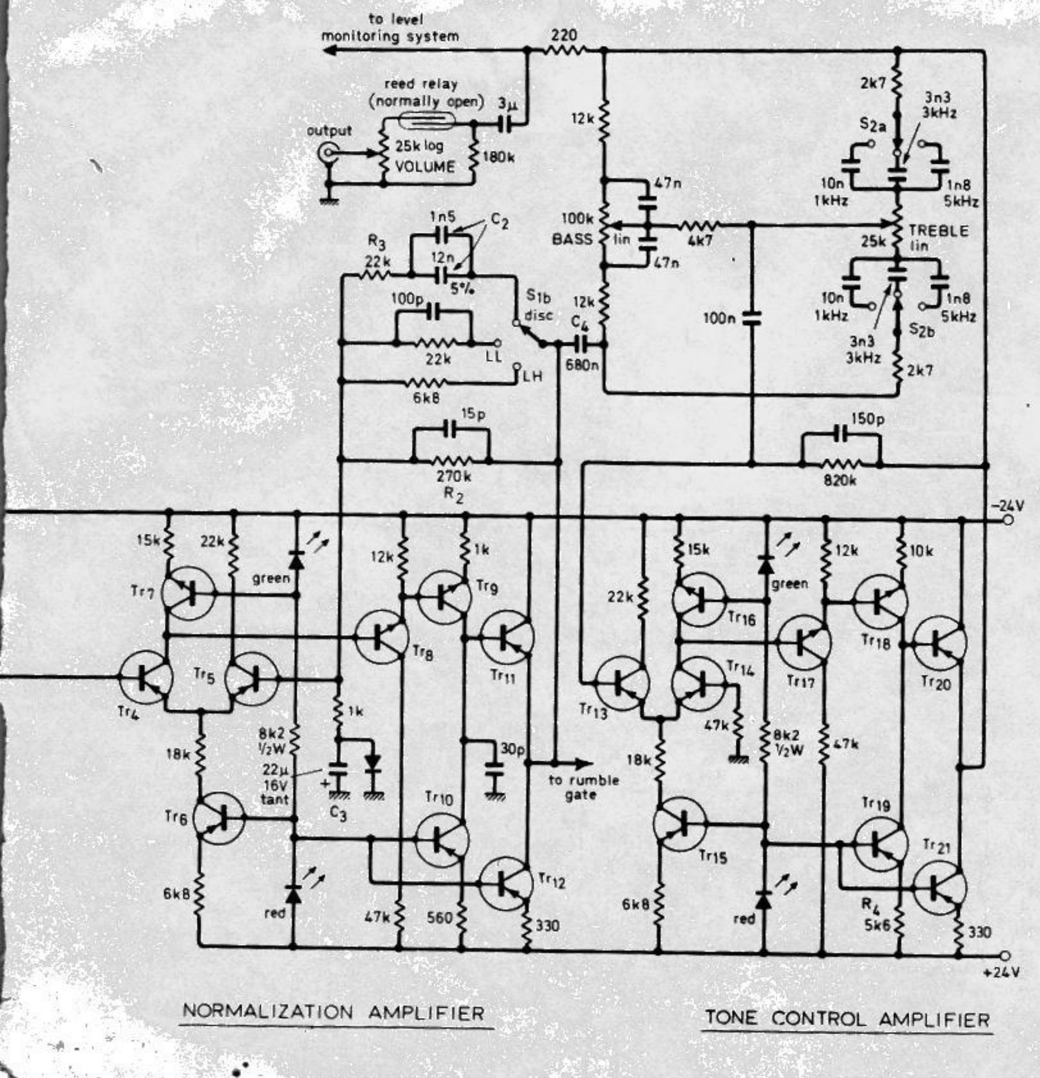


Fig. 3. Tape output circuit. The smallest allowable load impedance for an undistorted output is about 2.2kΩ. Line inputs of the pre amplifier are suitable for playback purposes.

14.5V r.m.s. which is 25dB above nominal operating level. This large amount of preamplifier headroom allows gross preamplifier overload before clipping. The input stage of the amplifier is a differential pair with a constant-current source for good common-mode rejection. The operating currents are optimized for good noise performance, and the output is buffered by an emitter-follower. The main voltage amplifier, Tr_9 has a constant-current collector load so that high voltage gain at low distortion can be obtained. This performance is only possible if the stage has very little loading so it is buffered by the active-load emitter-follower. The various current sources are biased by a l.e.d.-resistor chain because the forward voltage drop of an l.e.d. has a negative temperature coefficient that approximates closely to that of a silicon transistor V_{be} drop. Hence, this method provides exceptionally stable d.c. conditions over a very wide temperature range.

After the normalization stage the signal is applied to a tone-control circuit based on the Baxandall network. The main limitation of the Baxandall system is that the turnover frequency of the treble control is fixed. In contrast, the bass control has a turnover frequency that decreases as the control nears the flat position. This allows a small amount of boost at the low end of the audio spectrum to correct for transducer shortcomings. The equivalent adjustment at the high end of the treble spectrum is not possible because boost occurs fairly uniformly above the turnover frequency for treble control settings close to flat. In this circuit the treble turnover frequency has been given three switched values which have proved useful in practice. Switch 2 selects the capacitors that determine the turnover point. The maximum boost/cut curves are arranged to shelve gently, in line with current commercial practice, rather than to continue rising or falling outside the audio range. In addition, the coupling capacitor C_4 has a significant impedance at 10Hz so that the maximum bass boost curve not only shelves but begins to fall. Full boost gives +15dB at 30Hz but only +8dB at 10Hz. The tone control system has a maximum effect of ±14dB at 50Hz and ±12.5dB at 10kHz.

The tone-control amplifier uses the same low distortion configuration as the normalization stage, but it is used in a virtual-earth mode. The main difference is that the open-loop gain has been traded for open-loop linearity by increasing the emitter resistor of the



main voltage amplifier from 1kΩ to 10kΩ thus increasing local feedback. Resistor R_4 has been increased to 5.6k to maintain appropriate d.c. conditions. This modification makes it much easier to compensate for stability in the unity-gain condition that occurs when treble-cut is applied.

Level detection circuitry

From the tone-control section the signal is fed to the final volume control via the muting reed-relay. Note that this arrangement allows the volume control to load the input of the external power amplifier even when the relay contacts are open, thus minimising noise. The signal level leaving the tone-control stage is comprehensively monitored by the circuitry shown in Fig.4. Each channel is provided with two peak-detection systems, one lights a green l.e.d. for a pre-determined period if the signal level exceeds 1V peak, and the other lights a red l.e.d. if the tone-control stage is on the verge of clipping. Each channel is also provided with a VU meter driver circuit. Transistor Tr_{22} forms a simple amplifying stage which also acts as a buffer. Voltage feedback is used to ensure a low-impedance drive for the meter circuitry. The first peak detector is formed by IC_1 and its associated components. When the voltage at pin 2 goes negative of its quiescent level by one volt, the timer is triggered and the l.e.d. turns on for a defined time. The relatively heavy l.e.d. current is drawn from an unstabilized supply to avoid inducing transients into any of the stabilized supplies.

The clipping detector continuously monitors the difference in voltage between the tone-control amplifier output and both supply rails. If the instantaneous voltage approaches either rail, this information is held in a peak-storage system. Normally Tr_{24} and Tr_{25} conduct continuously but if the junction of D_1 and R_5 approaches the +24V rail then Tr_{24} and hence Tr_{25} turn off. This allows C_5 to charge and turn on Tr_{26}, and Tr_{27} and hence the l.e.d. until the charge on C_5 has been drained off through emitter-follower Tr_{26}. If the measured voltage nears the —24V rail, then D_1 conducts to pull up the junction of R_6 and R_7, which once again turns off Tr_{25}. In this way both positive and negative approaches to clipping are indicated. This comprehensive level indication does of course add significantly to the task of building and testing the preamplifier. If desired, any or all of the three sections may be omitted.

Noise gate

The final section controls the muting reed-relay. At switch-on, the +12V rail rises rapidly until stablized by the zener diode. Pin 2 on IC_4 is, however, briefly held low by C_6, and the 555 is therefore immediately triggered to send pin 3 high. This saturates Tr_{28} which prevents Tr_{29} from turning on. At the end of the time delay, pin 3 goes low and relay driver Tr_{29} is no longer disabled.

The noise gate uses two amplifiers with gains of about 100. These sample both channels at the output of the normalization stage and the inputs are clamped with diodes so that the normalization amplifiers may use their full voltage swing capability without damaging the 741s. Due to their high gain, under normal signal conditions the op-amp outputs move continuously between positive and negative saturation which keeps the storage capacitor C_7 fully charged. In the silent passages between l.p. tracks the l.f. signal is not normally of sufficient amplitude to cause saturation but will usually produce at least +3 to +4 volts across C_7 which gives a large margin of safety against unwanted muting. To facilitate this the response of the amplifiers is deliberately extended below the audio band. When the stylus leaves the record surface and the l.f. signals cease, C_7 slowly discharges until the non-inverting input of comparator IC_3 falls below the voltage set on the inverting input. At this point the 741 switches and its output goes low to cut off the base drive to Tr_{29}, and switch off the relay. When the stylus is replaced on a record, the process takes place in reverse, the main difference being that C_7 charges at once due to the low forward impedance of D_2. To prevent the relay sporadically operating when the preamplifier is handling signals presented through the line inputs, an extra wafer on the source-select switch is arranged to override the rumble-sensing circuit, and provide permanent unmute. This is achieved by pulling the inverting input of comparator IC_3 negative of the +15V rail by the 10kΩ resistor so that even when C_7 is fully discharged, IC_3 will not switch. In addition, S_3 provides a manual override for testing and comparison purposes.

The power supply is shown in Fig.6. Regulators are used to provide stabi-

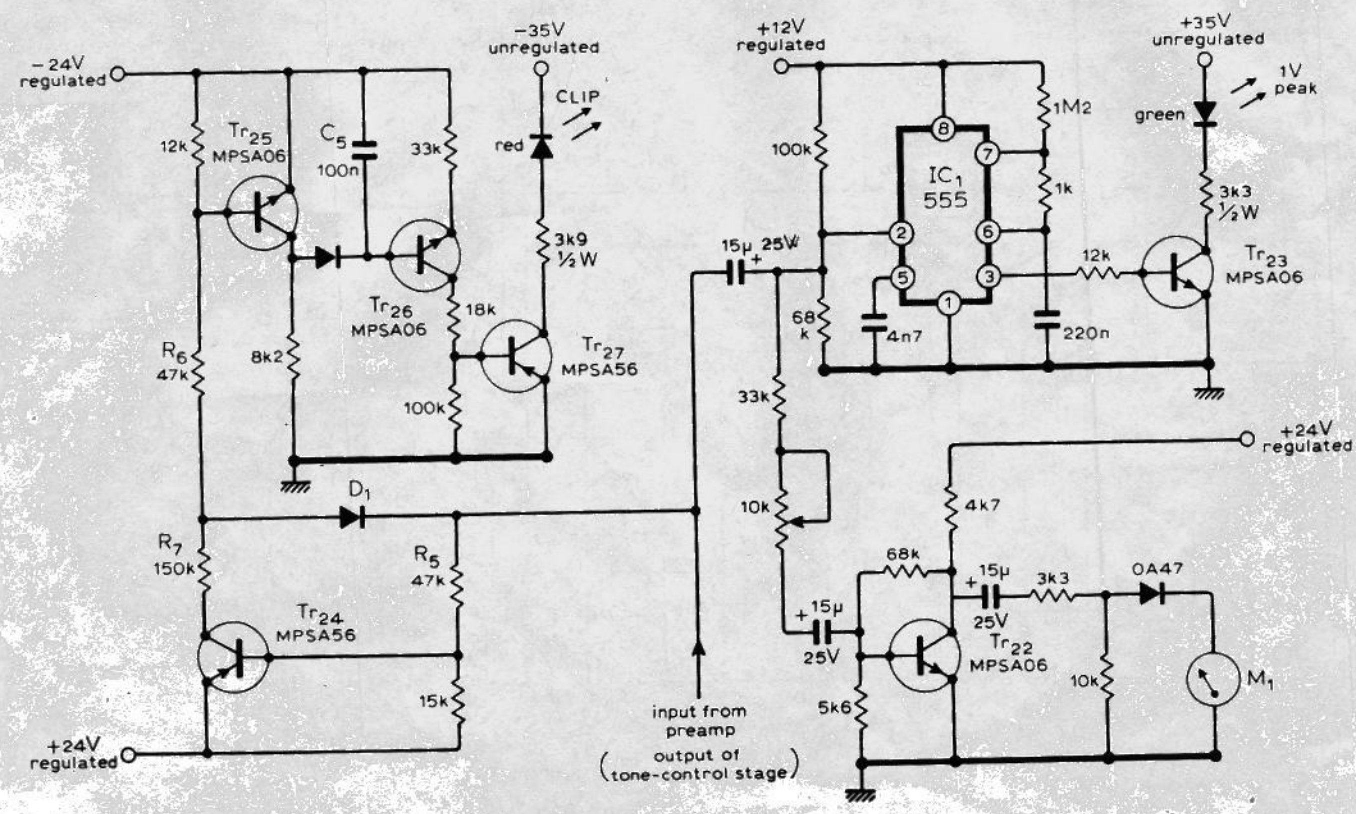


Fig. 4. Level monitoring circuitry. Although three separate circuits are shown, these may be omitted as required.

Component notes

All unmarked diodes are 1N914 or equivalent.
Red bias l.e.ds are TIL209 or equivalent.
Green bias l.e.ds are TIL211 or equivalent.
Resistors marked with an asterisk should be metal oxide types
Tr_1 to Tr_6 and Tr_{13} to Tr_{15} are BCY71
$Tr_{7, 8, 9, 16, 17, 18, 22, 23, 25, 26, 28}$ are MPS A06.
$Tr_{10, 11, 12, 19, 20, 21, 24, 27}$ are MPS A56
Tr_9 is BFX85 or equivalent.
The muting reed relay should be a 2 pole make type with an 18V coil. If a different coil voltage is used, the value of the dropper resistor should be adjusted.
The VU meter should have a 1mA movement.
If an internal diode and series resistor are fitted, the external components should be omitted.
Switch 1 (source select) is a 5 pole 3 way
Switch 2 (treble frequency) is a 4 pole 3 way

bility of spurious e.m.fs arising between stages. The only problem likely to be encountered is the formation of an earth loop when the preamplifier is connected to a power amplifier. Therefore, it may be satisfactory in a permanent installation to have the preamplifier circuitry connected to mains earth only through the signal lead to the power amplifier. The preamplifier case must of course be connected to the mains earth for safety reasons. It is preferable to define the potential of the preamplifier even if the power amplifier is disconnected. In the prototype the 0V rail was connected to the mains earth via a 22Ω resistor which stops the formation of an earth loop and prevents the signal circuitry from taking up a potential above earth due to leakage currents etc.

Testing is relatively straightforward, providing the preamplifier is constructed and checked stage by stage. Dynamic parameters such as t.h.d. are not accurately measurable without expensive test gear, but it has been found in the course of experimentation that if the d.c. conditions are correct then the various signal stages almost always show the desired a.c. performance. The non-signal circuitry should be relatively simple to fault-find. No problems should be encountered with the noise gate section which has proved to be very reliable throughout a protracted period of testing. The only preamplifier adjustment is for the VU meter calibration. This should be set to 1V r.m.s. = 0VU, which is completely non-standard but very useful in terms of the dynamic range of the signal path. For normal operation the input gain controls should be set so that the meter indications do not exceed 0VU, to preserve a safety margin in the later stages. This completes the preamplifier design.

References

1. Walker, H.P. "Low-noise Audio Amplifiers", *Wireless World*, May 1972.

2. King, Gordon J. *"The Audio Handbook"*, Newnes-Butterworth, 1975.

3. Heidenstrom, P.N. "Amplifier Overload", *Hi-Fi News*, December 1974.

Printed circuit boards

Wireless World has arranged a supply of glass fibre boards for the advanced pre amplifier. Two boards are available, one accommodates two audio channels, the other accommodates the stereo level detection and noise gating circuitry. The set of boards is priced at £7.00 inclusive. Available from M. R. Sagin at 11 Villiers Road, London N.W.2.

Literature Received

"Service planning and propagation" is No. 7 in a series of Technical Reviews from the IBA, which deals with the initial planning stage of new transmitters for sound and television broadcasting. Seven articles, concerned with service areas, field strength surveys, television microwave links and the business of negotiations with land interests are contained in the 64-page book, which is obtainable from Engineering Information Service, IBA, Crawley Court, Winchester, Hants SO21 2QA WW401

A cross-reference guide between CV, BS9000 and Nato semiconductors is available from Swift Hardman, containing currently-available types. Copies are obtainable from Semiconductor Sales Department, Swift Hardman Ltd, PO Box 23, Baillie Street, Rochdale, Lancs.................. WW402

A new GDS catalogue is now ready, containing lists and application information on a wide range of semiconductors, including Fetrons and integrated circuits. Components, tools and accessories are also covered and all prices are included. GDS Sales Ltd, Michaelmas House, Salt Hill, Bath Road, Slough, Berks SL1 3UZ...... WW403

We have received another weighty tome in the Eurolec series, this time No. 50, seventh edition of the guide to UK electronic components. In 507 pages, the book lists, component manufacturers, importers and distributors, with company information and contracts, in both alphabetical order and by location. People in the industry are listed and there is a section in which associate companies are linked with their principals. A buyer's guide completes the book, which is available from Eurolec, Little Waltham, Chelmsford CM3 3NU, at £15.80 by post.

Modular telemetry outstations in the Serck Flex-C-Mos range are described in a four-page leaflet, from Serck Controls, Queensway, Leamington Spa, Warks...... WW404

KGM have sent us a folder illustrating the principle behind the M-range edge-lit indicators. Characters of any form in sizes from 4mm to 150mm can be displayed. KGM Electronics Ltd, Clock Tower Road, Isleworth, Middlesex TW7 6DU.... WW405

"All in one package" is the title of an application booklet, produced by Marconi Instruments, on their TF2008 a.m./f.m. swept signal generator, which covers the frequency range 10kHz to 510MHz. Copies can be obtained from Publicity Department, Marconi Instruments Ltd, Longacres, St. Albans, Herts. WW406

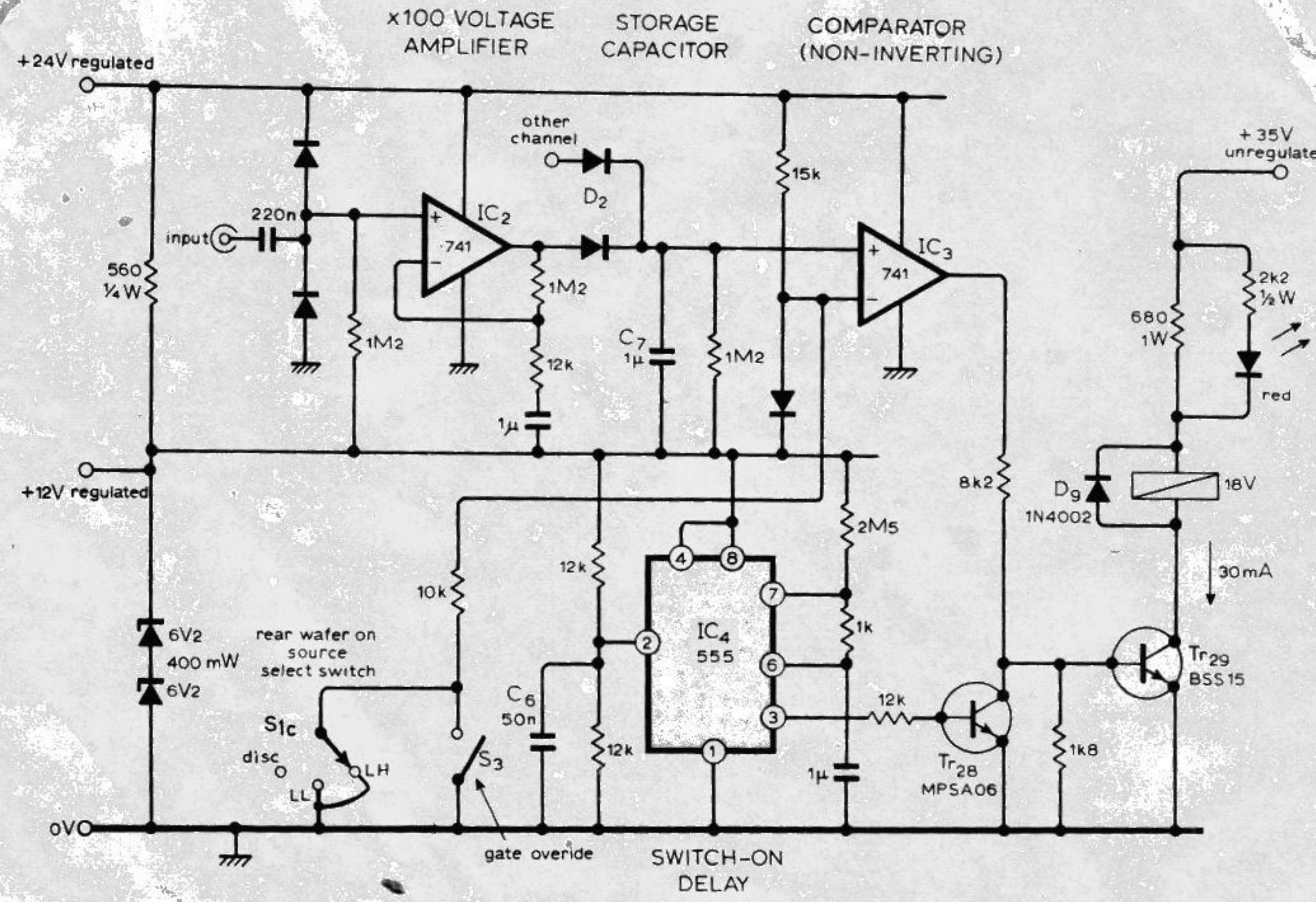
A series of leaflets, produced by Omron, provides full electrical, mechanical and application information on the E3N range of photoelectric switches. The leaflets can be had from IMO Precision Controls Ltd, 349 Edgware Road, London W2 1BS.... WW407

Counter/timers with maximum input frequencies of 32MHz to 1.2GHz are described in a brochure from RCS Electronics. Specifications of six instruments are included. Copies from 6 Wolsey Road, Ashford, Middx TW15 2RB...................... WW408

A list of UK-produced bibliographic data bases (abstracts, lists, reviews, information services, etc.) has been compiled by the British Library. The inventory is the basis of the UK contribution to the European information network and covers an enormous range of scientific and technical subjects. A subject index is provided. The "Inventory of bibliographic data bases produced in the UK" is available at £2.50 from Publications Ltd, British Library Lending Division, Boston Spa, Wetherby, West Yorkshire LS23 7BQ.

Ferranti have sent us leaflets on six new integrated circuits, an i.c. price list and an application report on the ZN417 analogue processor, which comprises three operational amplifiers and a voltage regulator in one package. The other i.c.s are the ZN1084E oscillator-divider timer, the ZN423T 1.26V reference source, the ZN10404 up/down counter, a gated linear amplifier, the ZN424, and an 8-bit a.-d./d.-a. converter type ZN425E. Ferranti Ltd, Electronic Components Division, Gem Mill, Chadderton, Oldham OL9 8NP.................... WW409

Technical Bulletin 7a-123 (UK) from Alpha Metals is entitled "Alpha M.E.G. Resin-based Solder Creams". The range of materials available is described, with methods of application and the removal of residues. The bulletin is obtainable from Alpha Metals Ltd, 457 Kingston Road, Ewell, Surrey .. WW410

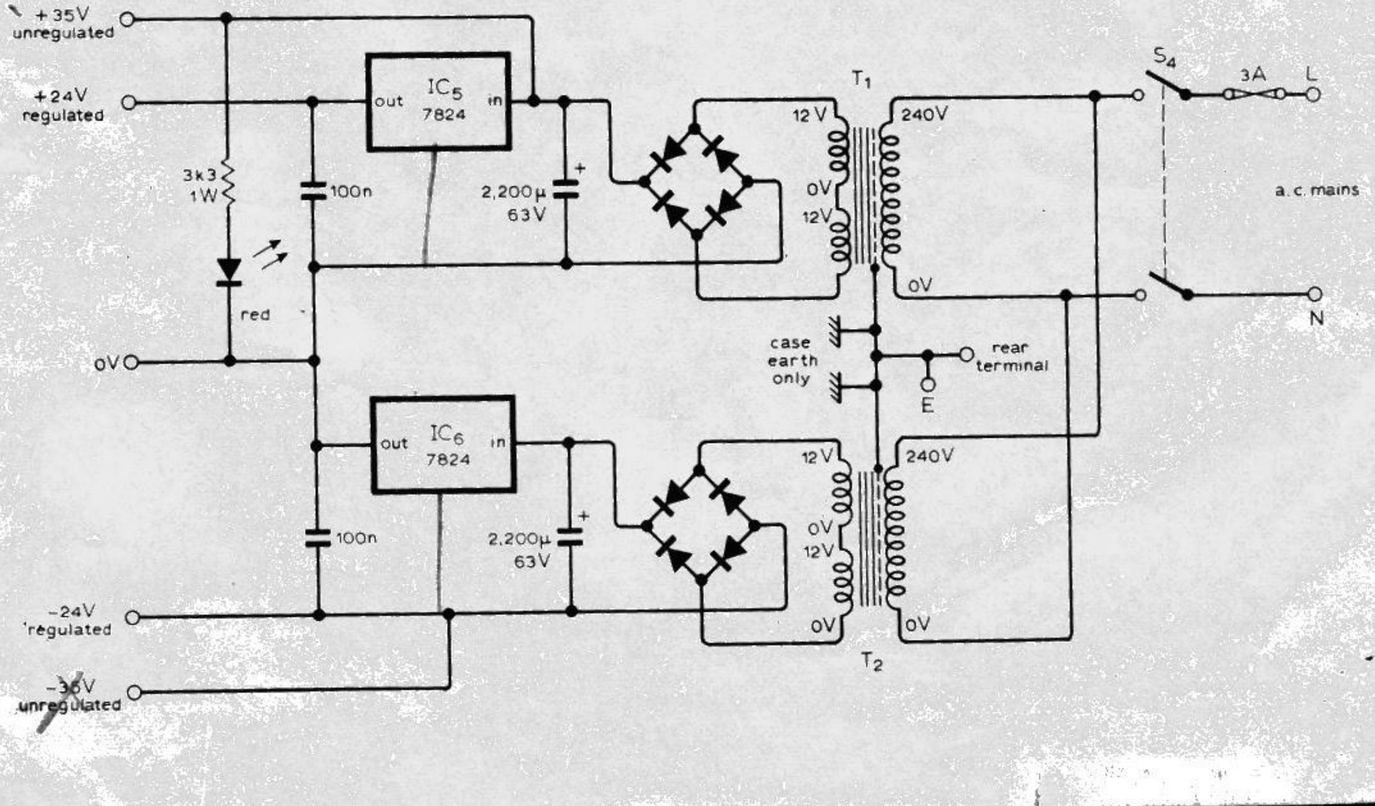


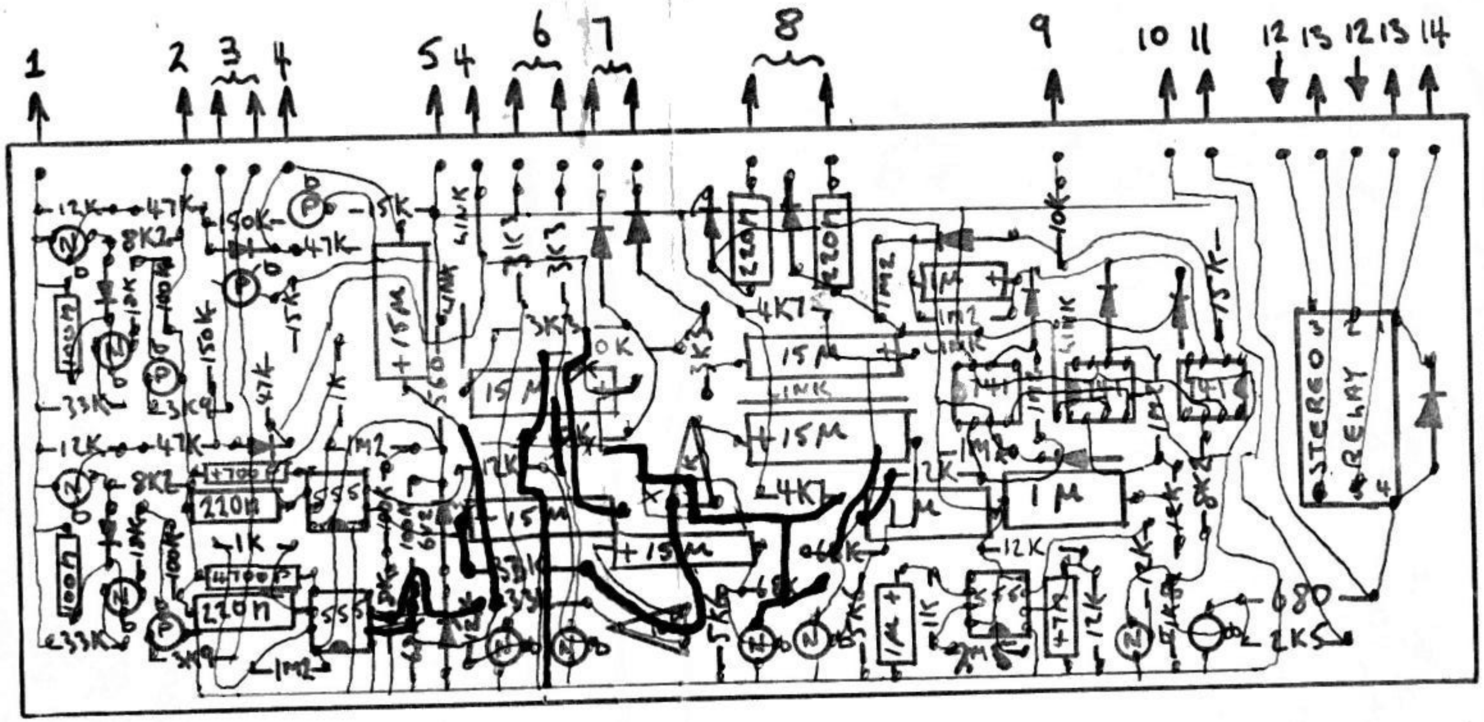
lized ±24V rails. The unregulated supply rests at about ±35V. The signal circuitry has been designed to withstand ±35V appearing on the supply rails, so that even in the unlikely event of both regulators failing, no further destruction will arise. Each regulator i.c. requires about 7cm² of heat sink area.

Physical layout of the preamplifier is no more critical than that of any other piece of audio equipment. In general it is wise to use a layout that places the disc input amplifier as close as possible to its input socket, and as far as possible from the mains transformer. Screened cable should be used between the disc input stage and its input socket, and between the final volume control and the output socket. The earthing requirements are straightforward and the circuit common 0V rail is led from the input sockets through the signal path to the output volume control, and finally to the 0V terminal of the power supply. This arrangement minimises the possi-

Fig. 5. Noise gate and delay switch on circuitry. The noise gate is provided with an override switch for use with line input signals. The delay switch-on overrides all of the circuitry. Amplifier IC_2 is repeated for a stereo system.

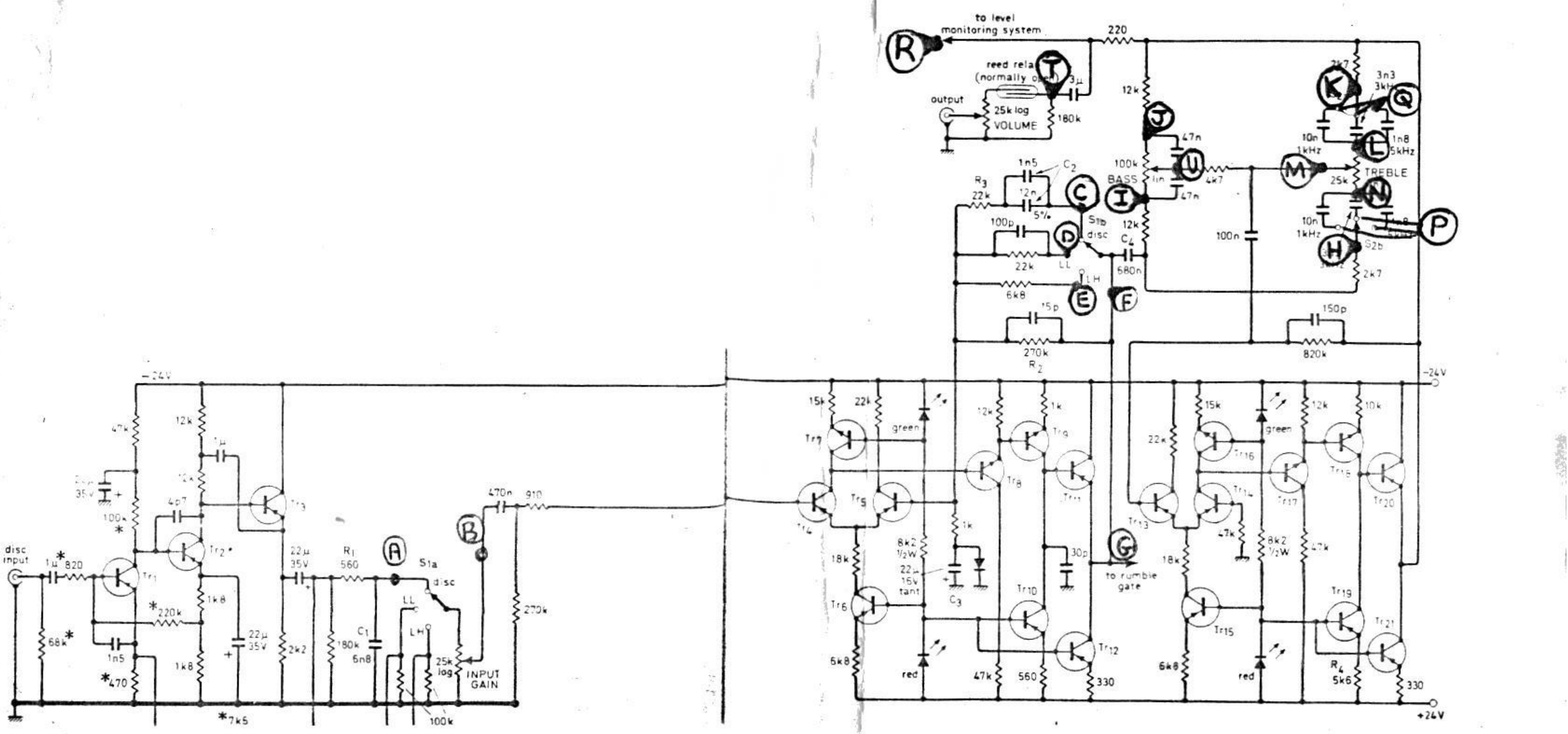
Fig. 6. Power supply. Two regulator i.cs are used which should be mounted on heat sinks.





COMPONENT SIDE

BOARD CONNEXIONS



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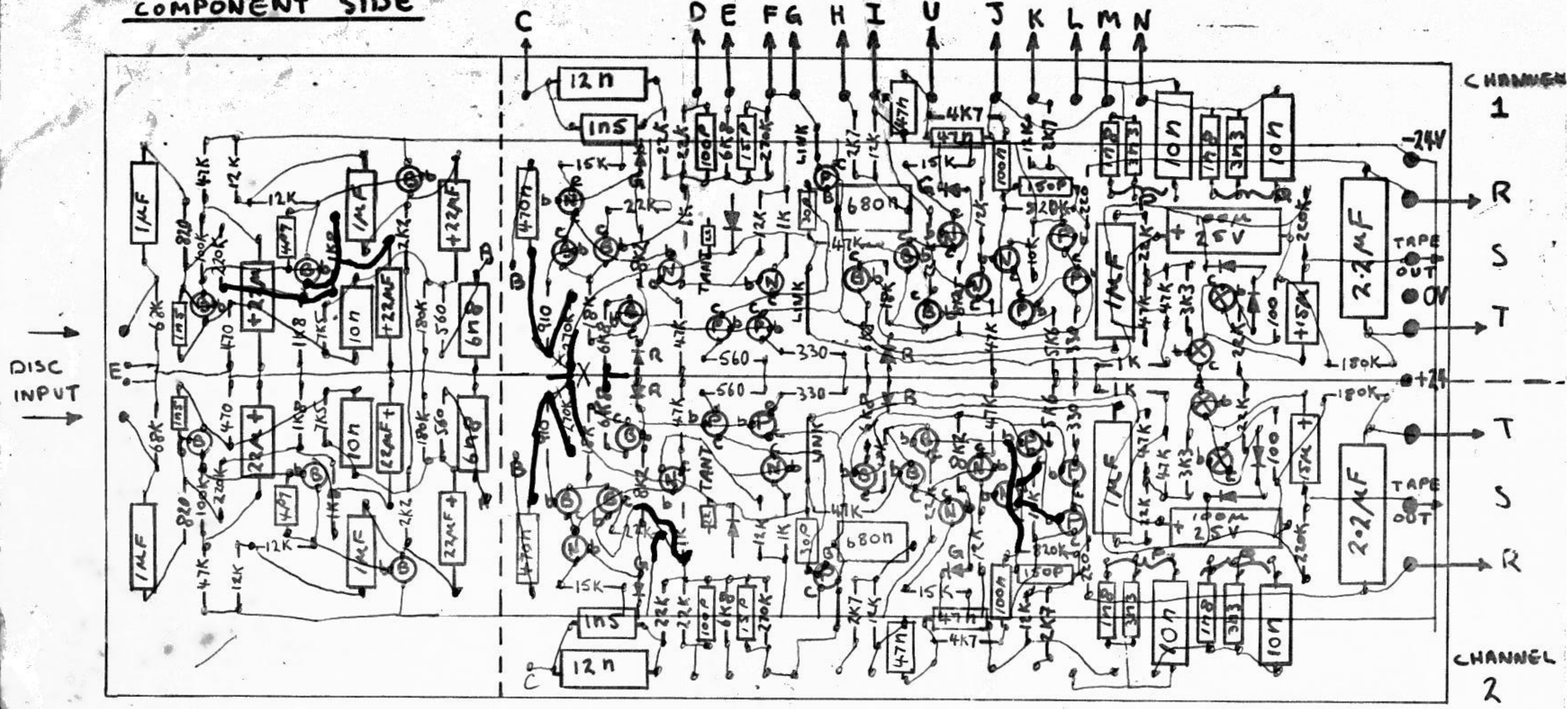


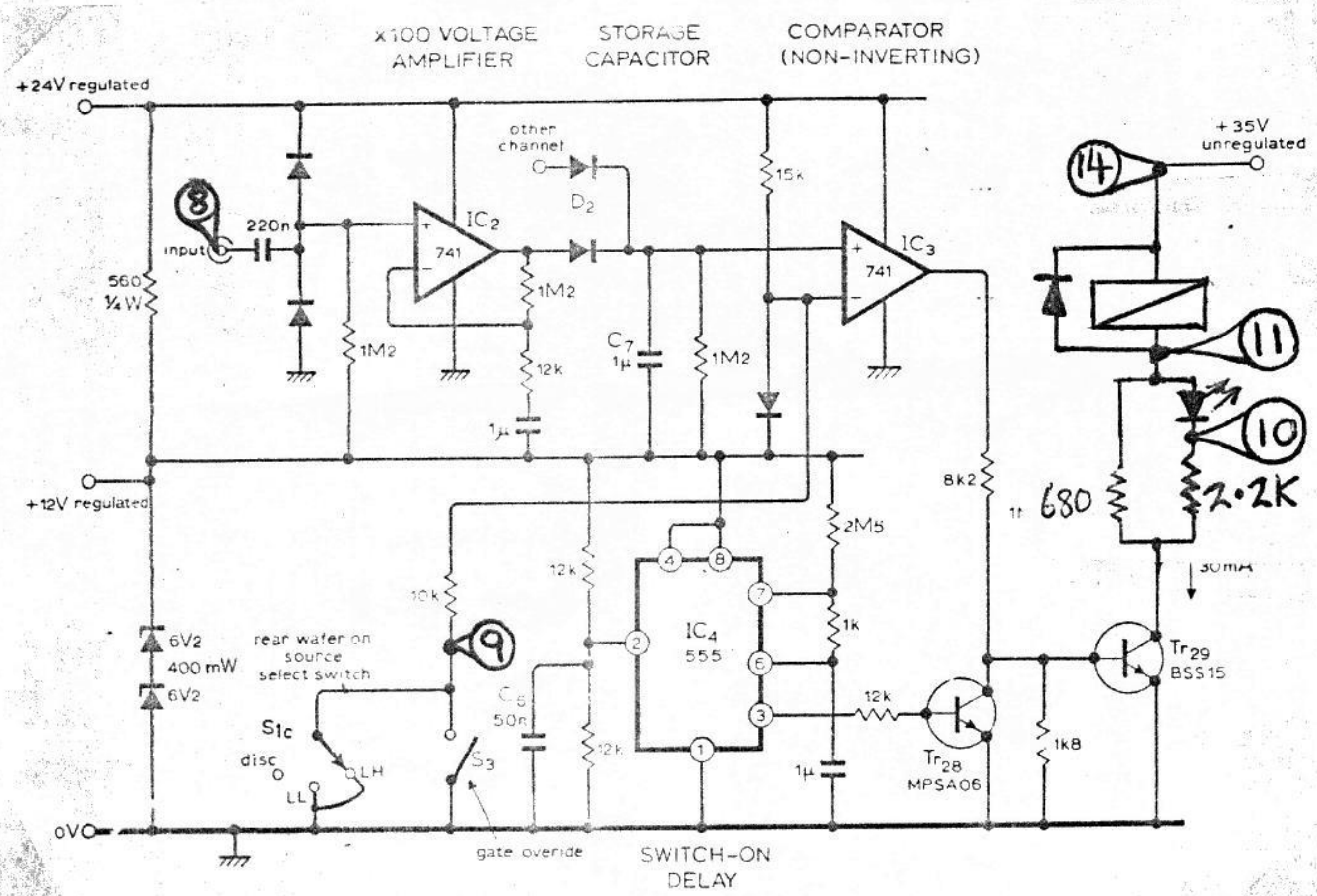
DISC AMP
MAY BE SEPARATED
BY CUTTING ALONG
DOTTED LINE

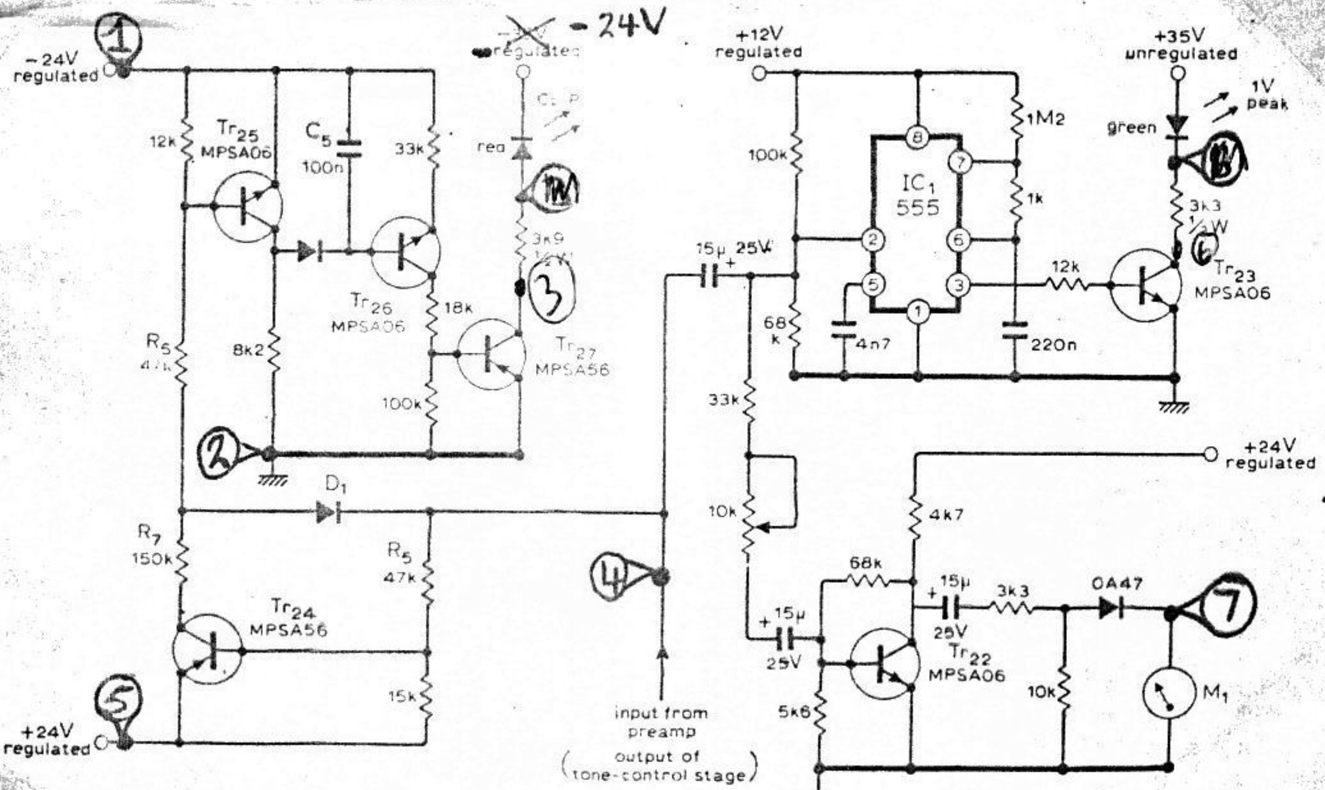
B = BCY 71
N = MPS A06
P = MPS A56
X = BC 184

ALL TRANSISTORS ARE IN THE STANDARD FORM (BOTTOM VIEW)

TAPE OUTPUT
CIRCUIT

BOARD CONNEXIONS





in large measure to the wider consultation now entered into. The Mobile Radio Users' Association pressed for wider consultation when the first Warden report was produced (long before *Wireless World* became involved in the subject) and we were naturally pleased to see the same flag being flown in these pages.

It was surprising, therefore, to read in the April editorial that "... discreet trusties referred to in December ... made, at first, no effort to press for a programme that might dilute their own bargaining strength." Your January article "Who is warden over the Wardens?" referred to myself as joint secretary of the Home Office Mobile Radio Committee representing p.m.r. users through the Mobile Radio Users' Association. May I please take some of your space to explain to readers how the MRUA contributed to considerable widening of consultation, and thus enlighten those of your readers with the unlikely image of myself or MRUA Chairman J. W. Tayler (also representing users at the MRC) as "discreet trusties"!

Following the submission of the Warden report to the Mobile Radio Committee in 1975, when intense and vigorous discussion took place, it was recognised, as Mr Carlton of the EEA mentioned in his letter in your April issue, as the first study of private mobile radio in depth, and likely to be of considerable importance in shaping policy. The MRUA felt, however, that the Home Office approach at WARC ought to be influenced by wider investigation and therefore decided to carry out an independent user survey of private mobile radio. Accordingly in December 1975 *every private mobile radio user in the United Kingdom* was sent a survey questionnaire together with a covering letter outlining the main conclusions of the Warden report. The results of the survey were published in the MRUA magazine *Talk Through* and appeared as an MRC paper, via which we hope the conclusions drawn may contribute to UK policy at WARC. I would submit that the circularisation, not only of all our members, but of *all* p.m.r. users hardly indicates a lack of effort on the part of the MRUA to widen discussion.

Alan Ford,
Secretary, The Mobile Radio Users' Association,
London SW1.

ADVANCED PREAMPLIFIER DESIGN

From his comments on my letter in the March issue on his preamplifier, I am afraid Mr Self did not understand the point of my letter.

The point was that, with the circuits I had tested, the circuit with part passive equalisation did sound better – though it needed music as complex as the opening of Mahler's 8th symphony to show initially that the sound was indeed better rather than just different.

To answer some of the points in Mr Self's reply. An amplifier with a low slew rate can be represented by an amplifier with infinite slew rate followed by a suitable RC filter. If this is capable of distortion, then alternative circuits with reactive components elsewhere within the feedback loop are likely to give distortion. Remember that the rules of negative feedback do not necessarily apply if the feedback is not exactly 180°.

I cannot agree with Mr Self that both amplitude/frequency and phase/frequency responses are identical for similar passive and active equalisation circuits. To a first approximation they may be equal, but the ear is capable of detecting very small differences. Such differences would appear to be attributed to second order effects such as:

(a) A finite closed and open loop gain of the circuit. The gain of a feedback circuit is not

$$G = \left(\frac{R_1 + R_2}{R_2} \right)$$

but

$$\frac{1}{1 + G/A}$$

where A is the open loop gain of the circuit and R_1, R_2 are feedback dividing resistors.
(b) The feedback input has a finite impedance. When the feedback is fed to the emitter of the first transistor this impedance is negative.
(c) The open loop bandwidth of the stage.

Attempts at mathematical analysis would appear to reveal second-order differences attributed to these three factors, but even deciding what form the analysis will take is complicated, let alone doing the calculations.

Obviously the overload margin on passive preamplifiers is much less than feedback equalisation circuits and waveform clipping has been heard on certain records with a high treble content. But it still sounds better and clipping can be avoided by a small increase in feedback. If Mr Self would like to offer his preamp to a qualified hi-fi reviewer for comparison against one of my passive preamplifiers, it would be interesting to see which sounds better when used with equipment of suitable (the highest) quality.

Graham Nalty,
Borrowash,
Derby.

CURRENT DUMPING AMPLIFIER

I was very interested to read the letter in your April issue by Divan and Ghate commenting on the "current dumping" amplifier described in your December 1975 issue. At first it seems incredible that one can entirely cancel out the distortions produced by a pair of output transistors, but having worked through the mathematics of it, I am now convinced. Indeed it will work even if the transfer function of the output pair is complex as well as non-linear, provided of course that the system is stable and the amplifier "A" is perfect and can produce adequate drive to compensate for the imperfections in the output pair.

The best explanation of "current dumping" is that feedback from the output pair to the amplifier is applied in the normal way, but can never completely cancel the distortion, so the error signal generated in the amplifier is fed forward and applied to the load, exactly cancelling any small remaining errors.

I would like to bring to your attention two errors in the equations:
(2) $Z_f || Z_3 || Z_{in}$ should read $Z_f || Z_3 || Z_{in} || Z_2$
(4) $Z_{in} || Z_2 || Z_3 || Z_4$ should read $Z_{in} || Z_2 || Z_3 || Z_f$

D. T. Ovens,
Havant,
Hants.

ADVANCED PRE-AMPLIFIER DESIGN

If that was an "Advanced preamplifier design" in your November 1976 issue then I can only hope that when it is fully developed it will look different from the circuit published.

First a few fundamentals:

1. Magnetic cartridges give output voltages dependent on the velocity of the needle; keeping the recorded amplitude fairly constant with frequency, the record makers therefore force the output of the cartridge to rise at +6dB/octave.

2. Normal cartridges today, because of development in magnetic materials (stronger, smaller magnets), give outputs of much more than 2mV, around 10mV at 1kHz for 5cm/s velocity.

3. If the disc is cut with an overhead of +20dB (peaks of 50cm/s) and the frequency is 20kHz not 1kHz, giving another rise in output of +20dB, then you can see the signal at 20kHz can be 1V.

Reality is not as bad as this since the spectral density of music is not constant with frequency and falls off at high frequencies. However, outputs from cartridges do rise to 200mV peaks and do have fast slew rates.

Mr Self's talk of overload margins is a little confused when he compares amplifier performance. If the normal operating (0dB) level of an amplifier is 10mV input then to cope with 1V inputs there must be no limiting of distortion anywhere before a gain control for signals of +40dB above normal. This is best known as an overhead of 40dB and *is required at 20kHz* relative to 1kHz.

Now RIAA amplifiers have some peculiar problems coping with the high transient signals from magnetic cartridges just because the output does rise with frequency: this rise causes a high spectral density of high frequency signals and high slew rates. The prime requirements in the input stages are therefore wide bandwidth (to give fast slew rate) and low transient distortion when handling the excess high frequency spectral density.

Mr Self's preamplifier does little for either of these: the open loop bandwidth is not clearly defined. If the second stage is guessed at 100 then the stage has a —3dB point of 3kHz. The bandwidth of the amplifier is further limited by the input capacitor (1n5) and by the output loading network R_1/C_1 on the output: in fact what can the amplifier drive into C_1 at 20kHz to give a respectable overhead margin?

More problems!

The input impedance will fall rapidly to high frequencies because the output signal is fed back via 10nF to the emitter of Tr_1 then by 1.5nF to the input itself. Therefore the magnet won't be given a chance to generate the correct h.f. signals, for to do so it must have a resistive load right up to 20kHz.

More problems!

The first two transistors are connected in a classic phase shift oscillator configuration. I have often had this configuration burst into l.f. oscillations when fed from a low impedance (which a cartridge has at l.f.). The reason is simple: there are two phase shift networks, first the r_e of Tr_2 and the decoupling capacitor 22μF ($\phi = 90°$ below 10Hz), the second the resistor 220k and the input capacitor 1μF ($\phi = 90°$ at 0.1Hz). Thus towards l.f. even if the circuit has insufficient loop gain to oscillate (it fortunately has by a factor of about four) it will have a characteristic l.f. peak of a few dB.

All amplifier designs of this type have some sort of l.f. peak; it could be suppressed by increasing the 22μF to 2200μF, thus reducing the feedback by 100 times, or best of all don't use this configuration.

Actually the component values don't seem to have been chosen consistently: the input capacitor of 1μF has a f–3dB point of about 1Hz but the decoupling capacitor has a f–3dB point about 100Hz which is rather the wrong way round to achieve a proper control of the l.f. response.

Only one more eyebrow to raise on the input amplifier! I quote, "insufficient cut at frequencies above 10kHz" (to give the correct RIAA which should be 6dB/oct. fall from 2.1kHz to >50kHz). I shudder to think what is happening to this amplifier's phase response with all the "tricky dicky" empirical networks hung on it. This really is the last straw....

Shall I go on to the l.f. amplifier? O.K., I will. But first some comments on the system.

I don't agree with the gain control where it is, the amount of gain following it is over 65dB at maximum bass boost. No matter how good the noise performance of Tr_4/Tr_5 some l.f. hum and noise will be present at the output *all* the time. By all means vary the input preset gain to allow for high output cartridges but the system volume control must be later on in the chain, or does Mr Self have another control on his power amplifier? The l.f. boost amplifier is a nightmare: why not use any one of the perfectly good op-amps available (L148T1, TBA231)? Why use a design with an obviously wide bandwidth and enormously high gain to do a job that a lower bandwidth, lower gain (more stable) amplifier can do? There isn't, you see, the problem in this stage of lots of h.f. spectral density and fast slew rates — this has all been removed by the input amplifier!

The design here has the following major problems:

1. The open loop gain depends on the transistor h_{FE}s (very variable).

2. The open loop compensation is not calculated to ensure good transient response and/or stability. Is it calculated?

3. The response of the network 270k + 22k + (1n5//12nF) does not give anything like the correct l.f. response for RIAA. This should start to fall at 50Hz, all 20dB at 6dB/oct. to 500Hz then go flat to >20kHz. Mr Self's circuit, if he wants to know, starts to fall at 37.4Hz and falls at 6dB/oct. for 24dB.

Finally, the tone control is the usual "Baxandall" horror, for two reasons. The first, the lift and cut of ±15dB is too large, giving audible phase shift problems, and anyway whose power amplifier can handle more than 10dB? The other reason is that the bass lift and cut varies both amplitude and frequency at once. On top of which there is the absurdity of providing selected treble roll frequencies alongside completely unknown and variable bass-roll frequencies!

O.K. I am willing to accept the challenge, if *Wireless World* is. [Yes — Editor.] I will describe my alternative version of preamplifier, with details of each design decision and performance objective.

Until then, Mr Self...?

A. J. Watts,
SGS-ATES (United Kingdom) Ltd,
Aylesbury,
Bucks.

Mr Self replies:
To deal with Mr Watts' main points in the order that he makes them:

He is correct in stating that the outputs from cartridges have high frequency peaks and large slew rates, and that this represents a potential problem in the design of RIAA-equalized disc input stages. However, if the treble-cut portion of the RIAA curve is incorporated in the first stage, in the form of frequency-dependent negative feedback, the falling high-frequency gain means that the signal the stage puts out is substantially "tamed" and so enormous slew rates are simply not required; the open-loop bandwidth of the published disc input stage is quite adequate.

He is wrong in stating that the closed loop bandwidth is limited by the 1n5 input capacitor; this component, in conjunction with the associated 820Ω resistor, forms an r.f. attenuation network to prevent breakthrough of radio signals, and has no effect within the audio band. This is because the input stage is in a series feedback configuration, and hence almost the same signal voltage appears on the emitter of the first transistor as at the base, due to the high open-loop again; hence at audio frequencies the capacitance is "bootstrapped" and has no effect.

Similarly Mr Watts is incorrect in saying that the input impedance of this stage will fall significantly at high audio frequencies. A.c. feedback is returned to the emitter of the first transistor, and not the base; this series feedback raises the input impedance of the stage, in accordance with the elementary laws of feedback, so that it has a negligible effect on the impedance seen by the cartridge, which is completely defined by the parallel combination of the 68k and 220k resistors. This gives a constant impedance across the audio band.

The first two transistors are not connected in a classic phase shift oscillator configuration; this requires three RC networks, not two. Hence the circuit cannot oscillate at low frequencies, though it is possible for diminishing phase margins at low frequencies to cause an l.f. hump, if the d.c. feedback time constants are poorly chosen. This is why the input and decoupling time constants are markedly different. I would prefer not to comment on Mr Watts' phase-shifts and frequencies as of course a single pole cannot ever give a 90° lag; it can only approach it asymptotically.

If a low gain input stage is used to allow a very high overload margin, then there will always be a problem in persuading the stage to give less than unity gain at the highest extremes of the RIAA curve. The extra treble cut network (560Ω and 6n8) does not alter the overall phase response, as its extra phase lag is compensated for by the falling phase lag of the input stage due to the h.f. gain levelling out at unity. Since we are dealing with a minimum-phase system (in the sense of having no all-pass filters), then the amplitude/frequency response completely defines the phase/frequency response. In other words, if the RIAA curve is correct, then the phase response will be indistinguishable from that of a more conventional circuit using only one treble-cut time constant.

And now to the next stage...

Mr Watts appears to have overlooked the system volume control at the end of the preamplifier chain; one can hardly have a volume control later in the proceedings than this. Since this control is used for day-to-day volume manipulation, and hence is rarely fully up, the residual hum and noise is attenuated with the signal, as Mr Watts suggests, and the desirable "zero noise at zero volume setting" condition is in fact attained.

If this stage is a nightmare to Mr Watts then I venture to suggest he will find trying to extract the same performance from a TBA231 even more of a bad dream. Integrated circuit operational amplifiers were not chosen as they give an inferior noise performance, due to the processes involved in integrating the input stages, and in general only accept lower supply voltages, hence giving less overload margin. As for the "major problems": 1. The open-loop gain certainly does depend on the transistor current gains. However, since this is the case for every amplifier ever built, I am unrepentant. To return to the laws of feedback, one of the prime motivations of negative feedback is to render closed-loop gain predictable by making the effect of open-loop gain changes negligible.

2. If Mr Watts can calculate the phase and gain stability margins of this stage, then I shall be interested to see his results. I find a flat assertion unconvincing and I imagine others will too.

3. If Mr Watts rechecks his calculations, or better still, measures the actual circuit instead of theorising, he will find that the combined response of the first two stages is very close indeed to the RIAA curve.

As for the tone control stage, I suggest it is probably impossible to design a tone control without phase shift.

As explained in the text, the variable turn-over frequency over the bass control is advantageous rather than otherwise. I fail to see how this makes the provision of switched treble turn-over frequencies "absurd."

In conclusion, I can only say that I would like to thank Mr Watts for the friendly and constructive nature of his comments. I can hardly wait to see his own preamplifier design.

CITIZENS'-BAND IN THE UK?

I note with regret that R. C. S. Withers' organization (UK Citizens' Band Campaign), is advocating the use of 27MHz for a citizens' band service in the United Kingdom ("Letters" December 1976).

Such a service is essentially short range and therefore the selected frequency range should not be one usable for long distance communication when the maximum usable frequency is high.

A u.h.f. band remote from broadcast television and amateur frequencies would be a first choice. Alternatively a v.h.f. band could be used but there would appear to be many demands for the use of v.h.f. for other services.

There exists a Citizens' Band Association which is promoting the establishment of a v.h.f./u.h.f. citizens' band service in the United Kingdom. They have published proposals for a service, including a technical specification.

H. Turner,
Derby.

ADVANCED PREAMPLIFIER DESIGN

The letter from Mr Watts in your February issue and the answer from Mr Self is notable for two factors — the abrasive language of the former's criticism and the surprisingly temperate reply from the latter. Frankly, I, too, could find much to fault in the design but, of course, there are ways of expressing it, aren't there?

My main criticism of Mr Self's design is that it is over-engineered, conceived by a hi-fi enthusiast who apparently has not been too involved in the costing process when putting together the elements of a circuit. The principle of Occam's Razor is also the essence of good design technique. He has also overlooked the simple facts of life — that despite the extremes to which one may go in designing equipment of this type, the aberrations that are inherent in all programme sources available to the domestic user are likely to be far greater than those introduced by even the most modestly designed reproducing equipment.

But Mr Watts is guilty of worse errors, in dealing with pure theory, opinion, and dressing it up as fact. Let me take one example — and since he seems to invite challenges, here's another from me. If he is able to produce for me a *high grade pickup cartridge* capable of the sort of amplitude linearity input when correctly loaded that he insists should be observed in the equalised input stage and will deliver consistently peaks in excess of 200mV, then there is £5 ready in my hand for any charity he cares to name.

Reg Williamson,
Norwich.

ADVANCED PRE-AMPLIFIER DESIGN

In reply to Mr Williamson (Letters, April), I think there are mainly two points to be made. One, that any pre-amplifier should have adequate signal handling capacity in excess of the performance of any pickup cartridge both dynamically and in pure consideration of the amplitude of signals. Second, that as far as I am concerned the two pickup cartridges which are capable of giving peaks in excess of 200mV are the Ortofon SL15 with appropriate transformer and the Decca London cartridge.

The reference to signal peaks of 80 cm/s observed on gramophone records came from the book "Hi-Fi Systems" by G. King where there is a graph illustrating the velocities measured on gramophone records at various frequencies.

I nominate my favourite charity as the Musicians Union!

A. J. Watts,
SGS-ATES (United Kingdom) Ltd,
Aylesbury,
Bucks.

Advanced preamplifier additions

Rumble and scratch filter, virtual earth mixer, meter suppression circuit

by D. Self, B.A. *Electrosonic Ltd*

The original design did not include a scratch or rumble filter because it was felt that an attempt to make no compromise in the range of facilities provided, as well as in performance, could lead to a design that was over-complex. Furthermore, use of the treble control in the 5kHz setting was thought to give some of the advantages of a variable-slope scratch filter. Nevertheless, the ultimate slope is only 6dB/octave, and therefore the high-frequency rejection is less than that obtainable with the usual second-order low-pass filter. It should also be noted that the low-frequency response of the original preamplifier was not extended to d.c., but rolled-off in a controlled manner to be 3dB down at 7Hz. However, a good deal of interest has been shown in further filtering facilities, and therefore the design shown in Fig.1

This article describes a number of additional facilities that may be easily added to the preamplifier design published in the November 1976 issue of *Wireless World*. These additions are in the form of independent circuits, one or all of which may be added to a completed preamplifier with a minimum of disturbance to the existing circuitry. Even if all the circuits are incorporated, the extra demands on the stabilized power supplies should cause no problems.

was evolved. This uses a switched-frequency second-order low-pass filter, and a fixed-frequency third-order high-pass filter that discriminates against the subsonic disturbances generated by record warps etc. The last mentioned is not really a rumble filter because it does not attenuate within the audio band. The frequency response is —1dB at 20Hz and —17dB at 10Hz, hence subsonic signals are greatly attenuated without perceptible loss of lower bass frequencies. The filter has a Butterworth characteristic. The low-pass filter incorporates switch-selected cut-off frequencies at 10kHz and 7.5kHz as measured at the —3dB point. These frequencies were chosen after listening to records suffering from varying degrees of damage and wear, and are believed to be a good compromise. The 10kHz filter has a relatively subtle effect which gives a smoother upper frequency response with records that are only slightly past their best. The 7.5kHz setting is effective with more severe cases, while discs in very poor condition may be improved by the use of an even lower cut-off frequency using the treble control. Restricting the low-pass filter to two switched frequencies allows the use of a 3-way 4-pole rotary switch. With the switch in the out position the low-pass filter still operates but the —3dB point is at 40kHz. This prevents

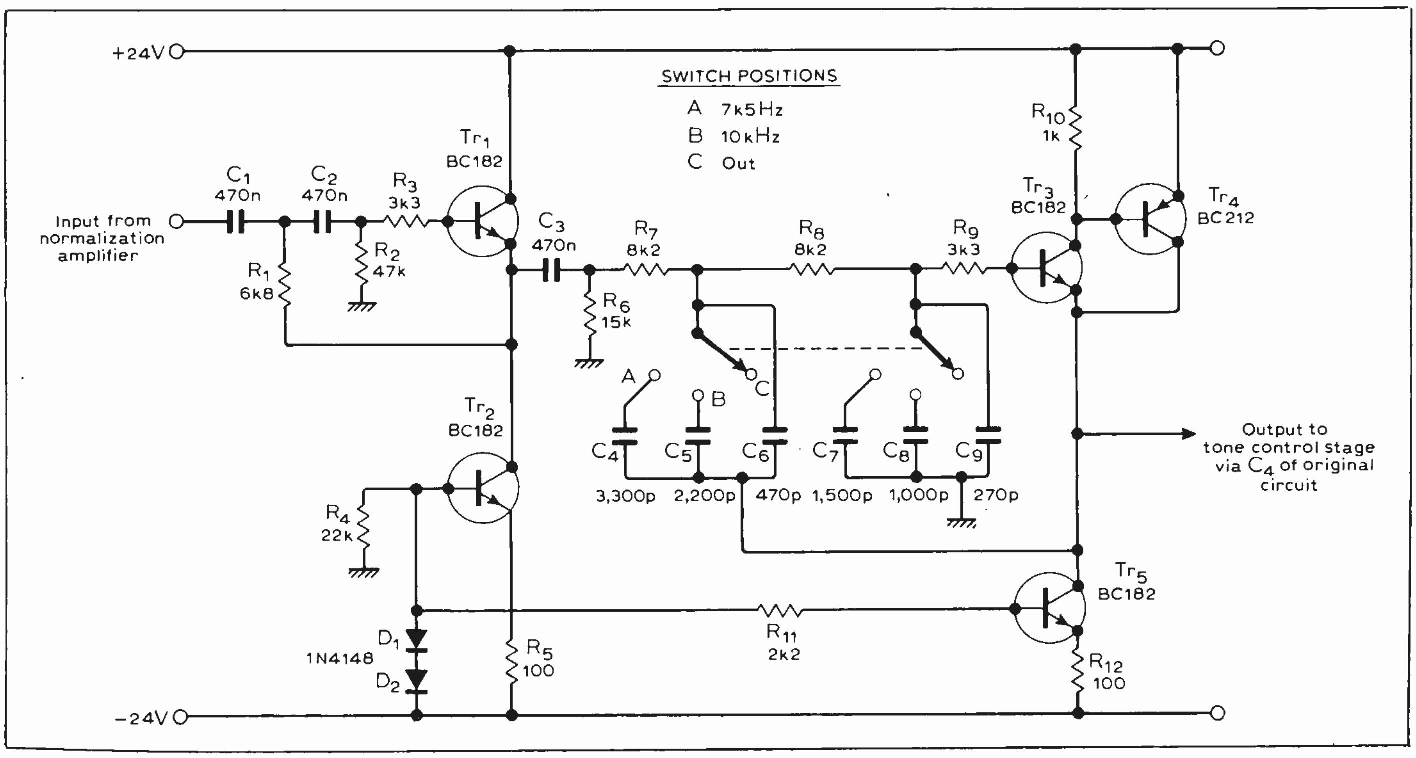


Fig. 1. Rumble and scratch filter. This circuit has a switched frequency low pass stage and a fixed frequency third-order high pass stage.

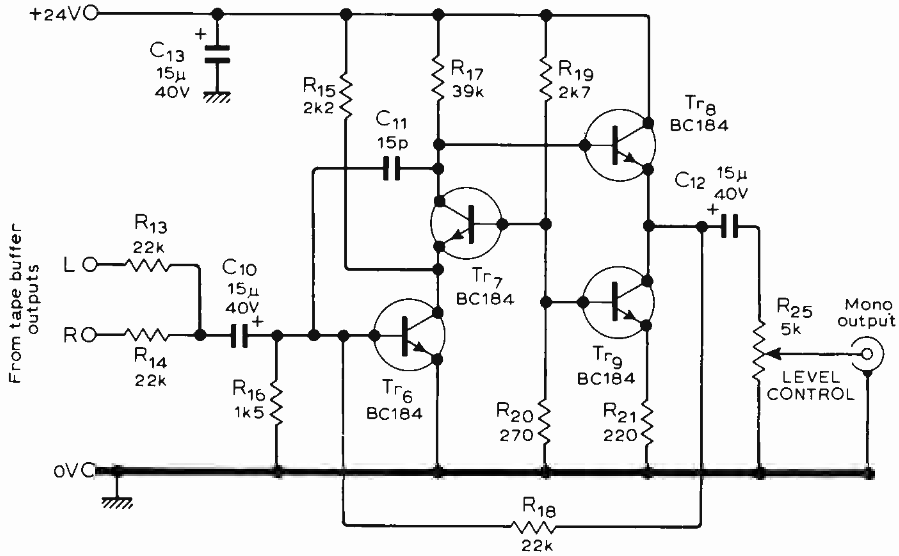


Fig. 2. Virtual earth mixer. This circuit allows simultaneous stereo and mono signals from the original preamplifier.

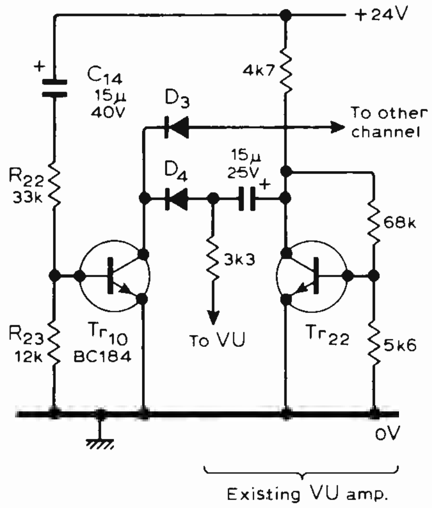


Fig. 3. Meter suppression circuit.

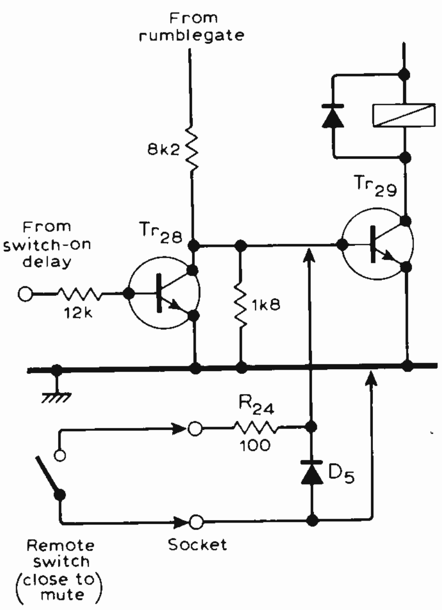


Fig. 4. Remote muting arrangement.

out-of-band frequencies from being fed to the power amplifier, which could cause t.i.d.

Both high and low-pass sections use the well-known Sallen & Key configuration, with unity-gain buffering provided by emitter-followers with current-source loading. As described in the original article, current-source emitter-followers generate very low levels of distortion and have better load-driving characteristics than conventional emitter-followers. Both current sources are biased from diodes D_1, D_2. Resistor R_{11} prevents Tr_5 from affecting Tr_2 in the event of a fault. The prevention of interaction makes fault diagnosis much easier. One point of interest is that the final unity-gain buffer is in fact a compound emitter-follower incorporating a complementary pair. This is desirable because the second unity-gain buffer, unlike the first, is driven from a substantial resistive source impedance, and under these conditions a simple emitter follower would generate a relatively high level of t.h.d. The circuit shown produces a t.h.d. figure of about 0.008% for 12V r.m.s. at 1kHz.

It is recommended that the filter is connected between the normalization amplifier and the tone-control stage so that the filters come immediately before C_4 on the original diagram. Note that the feed to the rumble gate must be taken off before the filters because the detection of subsonic frequencies is fundamental to the correct operation of the gating circuitry. The signal to the tape output emitter-followers may be taken before or after the filter system. A stereo filter draws an additional 30mA from the two stabilized supply rails.

The original preamplifier design had no provision for mono/stereo switching, as it was felt that it was unlikely to be used with a mono power amplifier/speaker system. It should also be appreciated that mode selection circuitry that does not compromise headroom or distortion performance in either mono or stereo would add a significant amount of circuitry to the preamplifier. The prototype preamplifier has now been in regular use for over a year, and in this time the only real need for a mono output has been for recording stereo material on a single-track tape machine. It was therefore decided to design a virtual-earth mixer that would give a mono tape output while the main preamplifier outputs remained in stereo. This simultaneous mono approach has the advantage that it can be easily added to an existing preamplifier without the need of a mode switch on the front panel. The prototype mono mixer incorporates an internal preset for controlling the output level. This can be brought out to the front panel or omitted as required.

The circuit of the mixer is shown in Fig.2. Mixing resistors R_{13} and R_{14} are fed from the existing tape buffer outputs, see Fig.3. of the original article. Shunt feedback is applied through R_{18} to give unity voltage gain and a virtual earth point at the junction of C_{10} and R_{16}. The design is derived from a configuration described by Butler[1] and offers a very low t.h.d. figure without the use of the usual circuit artifices such as bootstrapping or current-source collector loads. At an output level of 5V r.m.s. the t.h.d. is below 0.005% from 1kHz to 20kHz. Transistors Tr_6 and Tr_7 produce all of the voltage gain and are arranged as a cascode. Resistor R_{15} bypasses Tr_7 and allows Tr_6 to operate at a much higher current level than the collector load R_{17}. This arrangement appears to be crucial for good distortion performance. Transistors Tr_8 and Tr_9 operate as the now-familiar current-source emitter-follower so that low-impedance loads down to 3kΩ may be driven without loss of headroom due to premature clipping on the negative half-cycle. Note that current-source Tr_9 is biased from the same potential divider as Tr_7. The mixer circuit draws an additional 26mA from the +24V rail.

When the prototype preamplifier is switched on, the VU meter needles smartly strike their end-stops and then fall back. This behaviour is due to the initial charging current of the 15µF capacitor, coupling the collector of Tr_{22} to the meter, see Fig.3, passing through the meter movement itself as the supply rail rises. The degree of overload does not appear to be excessive as no degradation of meter accuracy has occurred during the past year. Nonetheless, the sound of meter needles against end-stops is not pleasing and the circuit shown in Fig.3 has been provided to prevent this effect. At the moment of switch-on, the +24V rail rises rapidly and current flows through C_{14}, R_{22} and R_{23}, to turn on Tr_{10}. This transistor shorts the switch-on surge to ground via D_4, and reduces the overswing to an inoffensive twitch. After a few hundred milliseconds C_{14} becomes fully charged and insufficient current flows through R_{22}, R_{23} to keep Tr_{10} conducting. The transistor then

Continued on page 86

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turns off and has no further effect on the VU circuitry. Diodes D_3 and D_4 isolate the two VU circuits from each other and also prevent Tr_{10} from being driven into conduction on negative half-cycles.

A further point about the VU system, which should have been emphasized in the original article, is that the 10kΩ resistor and germanium diode associated with the meter itself are only relevant if inexpensive milliammeter movements are fitted. If professional VU meters are used, which contain internal bridge rectifiers, the above components are unnecessary.

Because the amplifier is already fitted with a relay that switches off the main outputs, it is simple to add a socket for remote muting. Fig.4 shows part of the original rumble-gate system, together with the new components required. All that is involved is the closure of a switch between the base and emitter of Tr_{29} so that the transistor turns off and causes the relay contacts to open. Note that, like the switch-on delay, this facility overrides all other control functions. The resistor and diode are included to protect the circuit if a wrong connection to the remote-muting socket is made. As the control lead only handles a small direct voltage the audio signal cannot be degraded.

Reference

Butler, F. Transistor wide-band cascade amplifiers, *Wireless World*, March 1965, pp.124-128.

Printed circuit boards

A p.c.b. which accommodates a stereo rumble and scratch filter, virtual earth mixer, and meter surge suppression circuit, will be available for £3.50 from M. R. Sagin at 23 Keyes Road, London N.W.2.