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Modular Pre-Amplifier Design
(Wireless World, July 1969)

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

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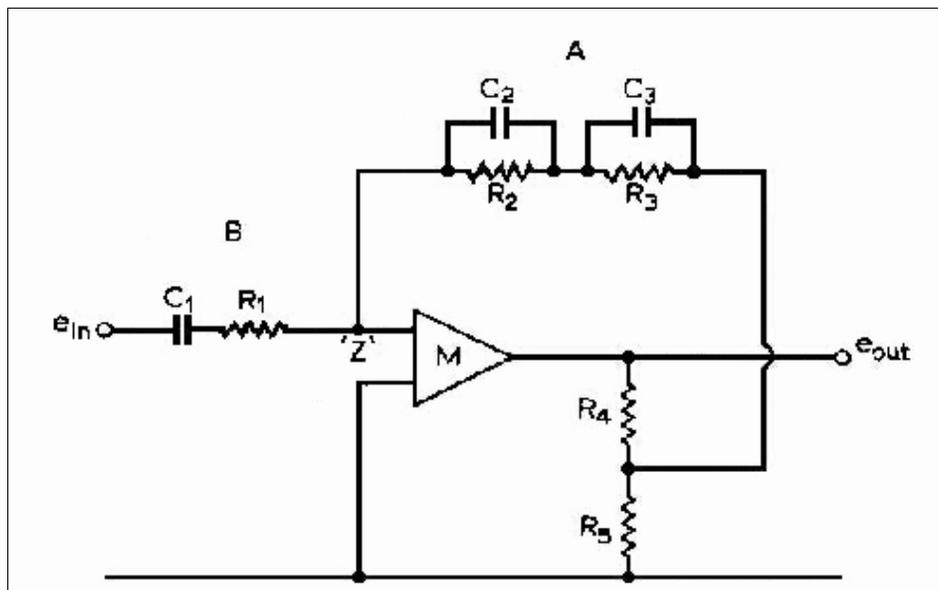
Fig. 2. Phase-inverting amplifier stage used to obtain R.I.A.A. replay characteristic.

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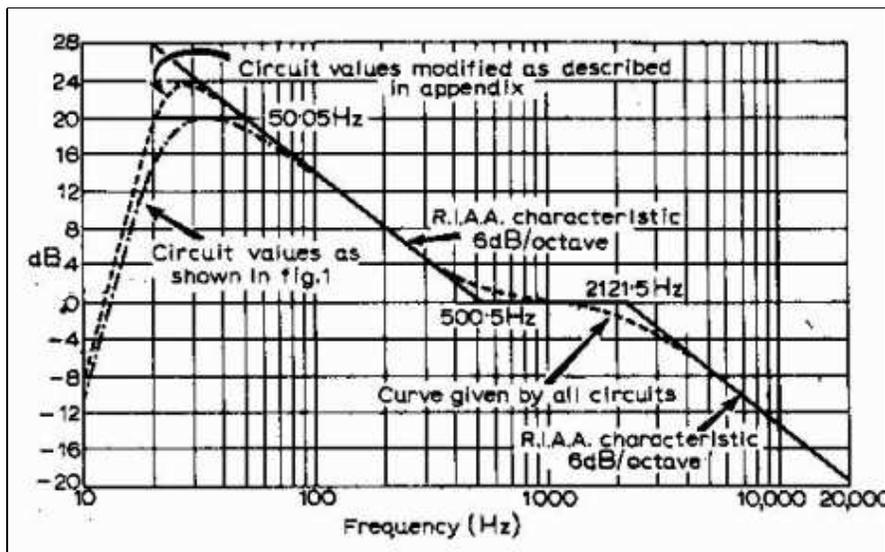
Fig. 3. Required R.I.A.A. frequency-response curve and circuits approximation to this.

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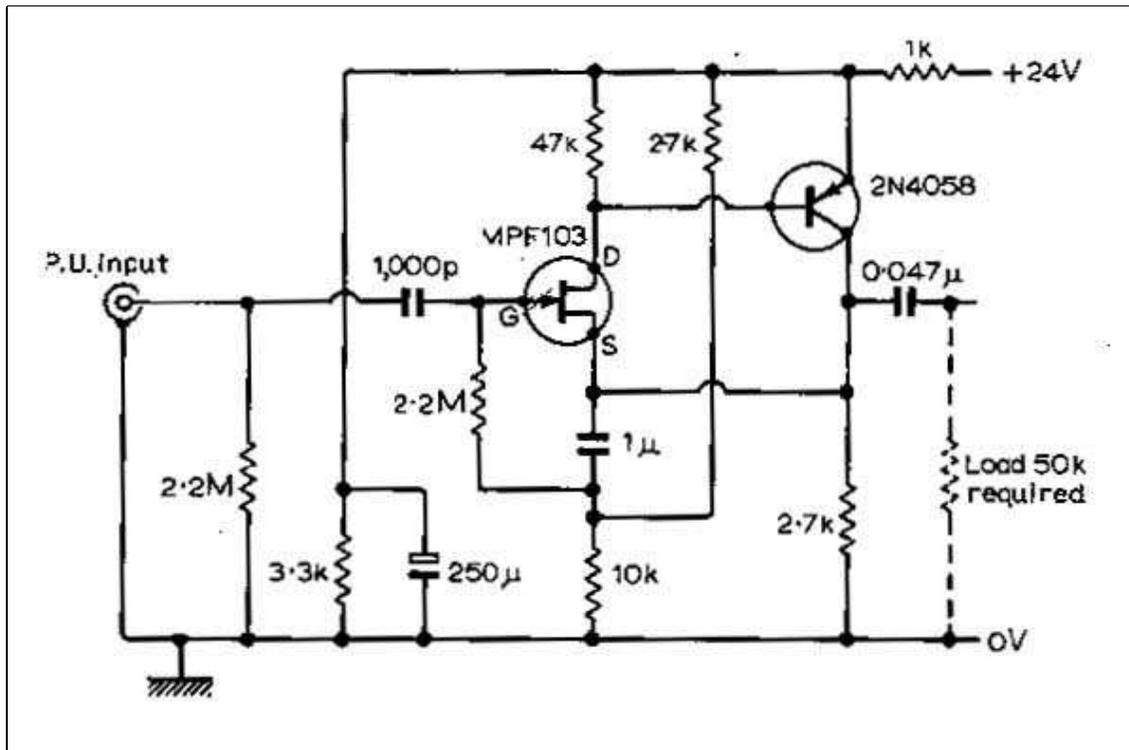
Fig. 4. Impedance conversion stage for use with ceramic cartridge. This may be
directly substituted for the magnetic cartridge stage at the front of Fig. 1.

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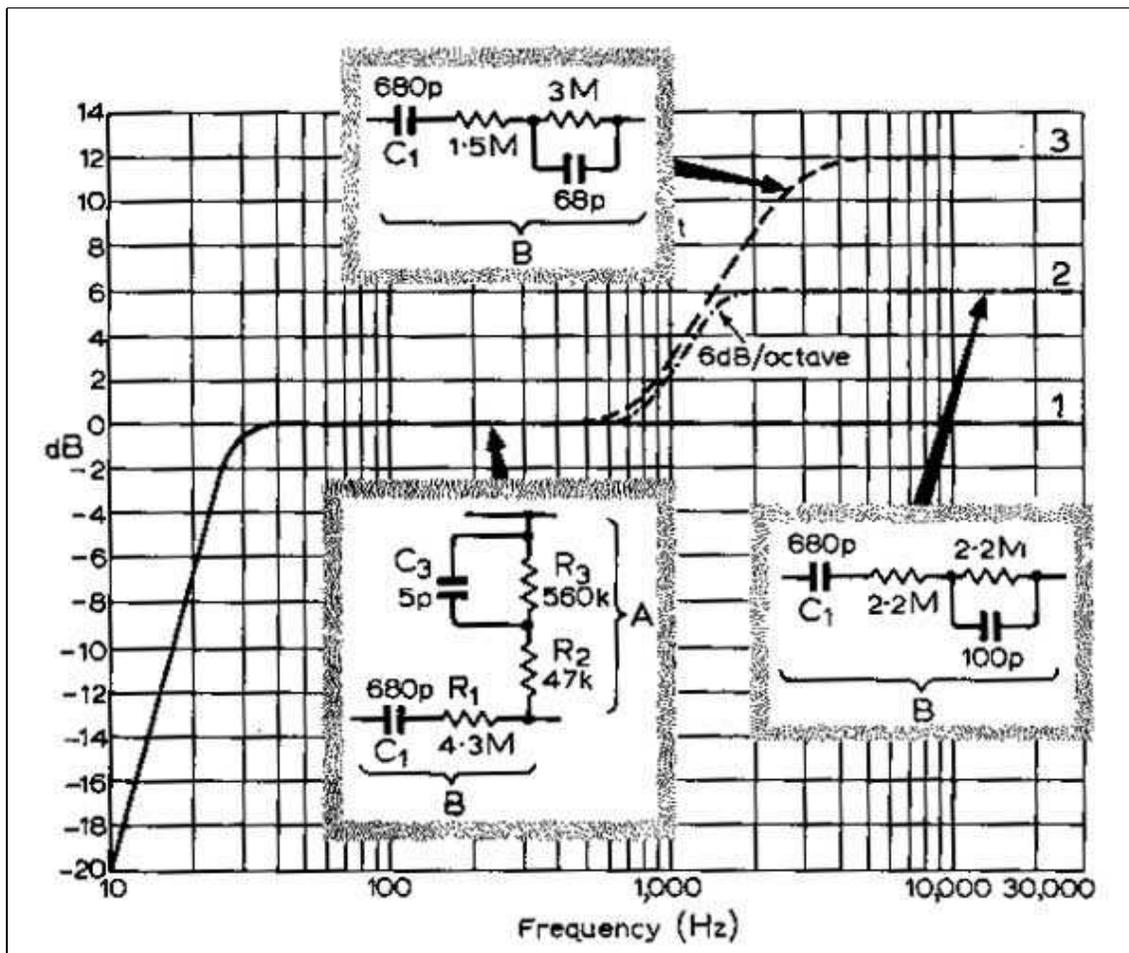
Fig. 5. Changes in equalization networks A and B of the magnetic cartridge
input stage allowing direct use of ceramic cartridge. Components for network
"A" are the same for the three curves shown.

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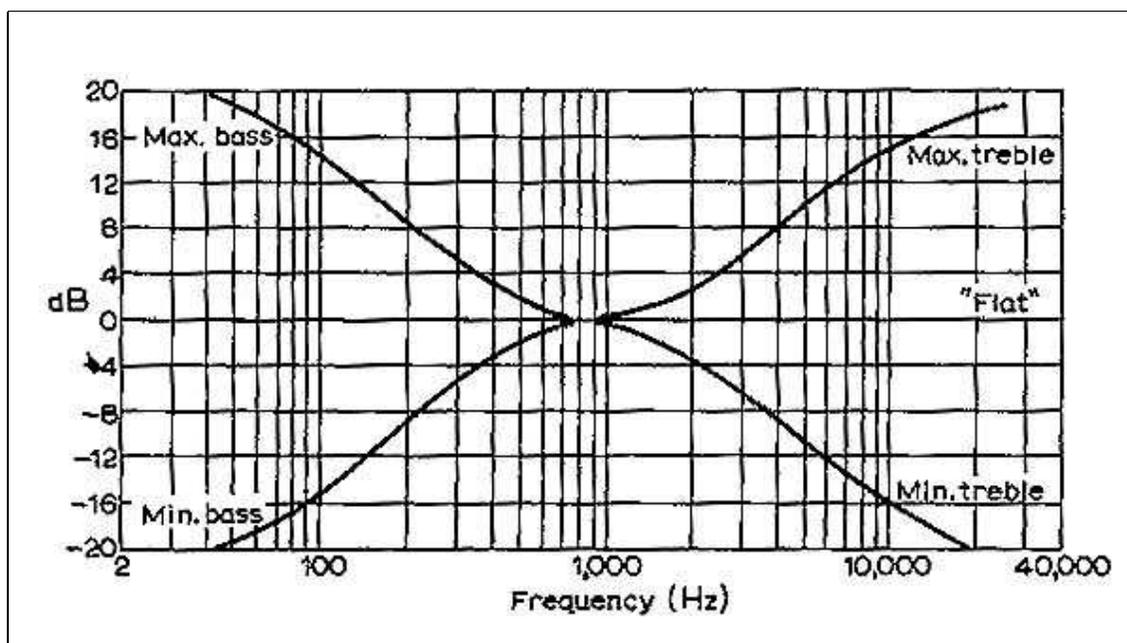
Fig. 6. Gain/frequency characteristics of tone control stage.

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.
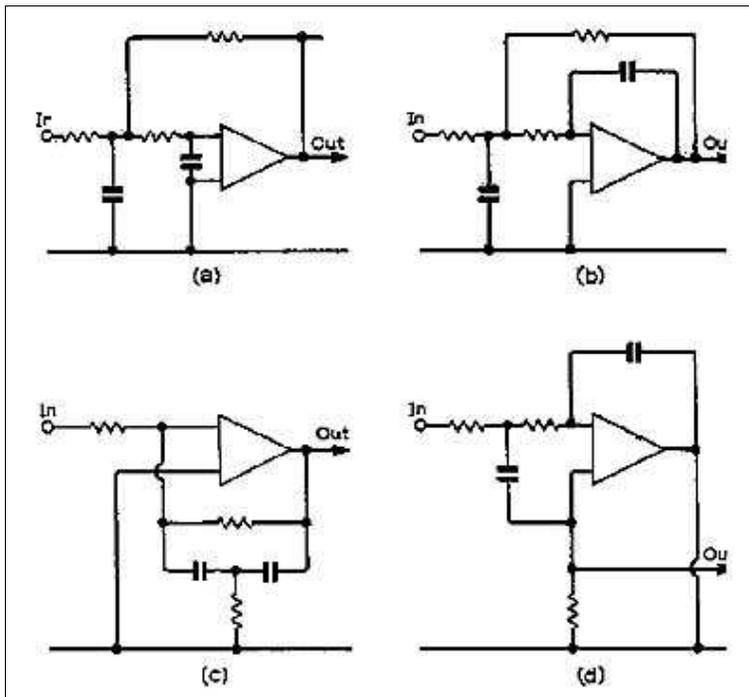
Fig. 7. Circuit arrangements for active low-pass filter design.
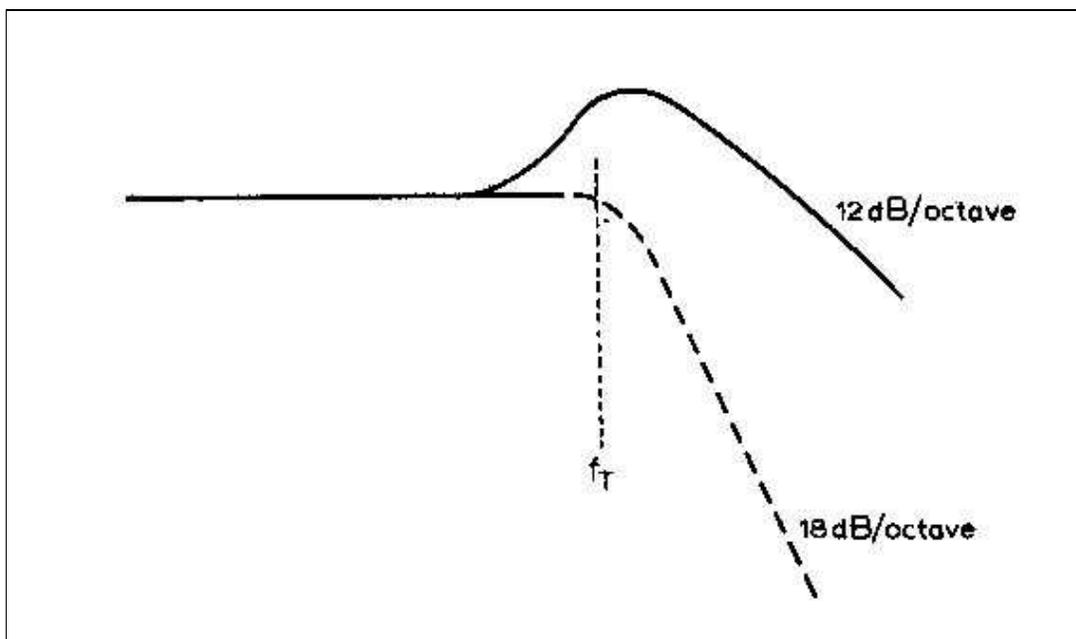
Fig. 8. Frequency response of the active filter circuits is 12dB/octave. Preceding
the filter with RC network gives response shown in broken line.

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.
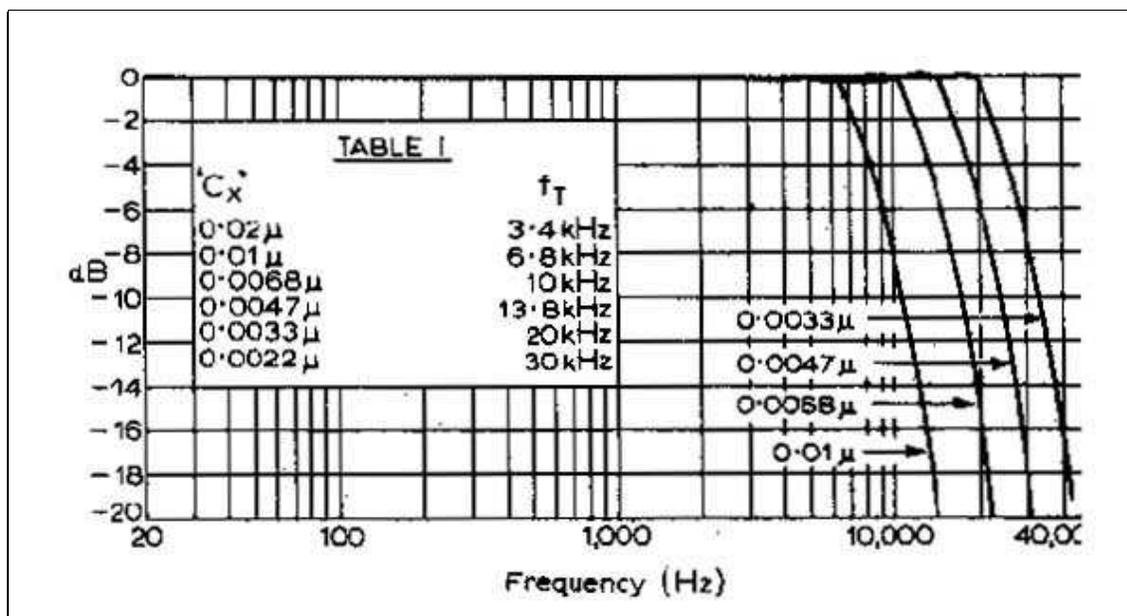

Fig. 9. Graph and table of turn-over frequencies for different value of 'Cx'.

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.
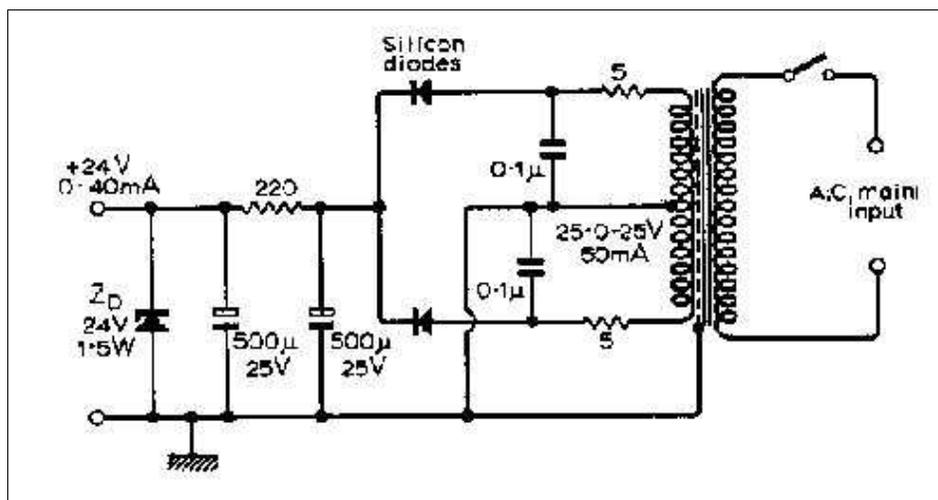

Fig. 10. Suitable power supply for any combination of stages.

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%.
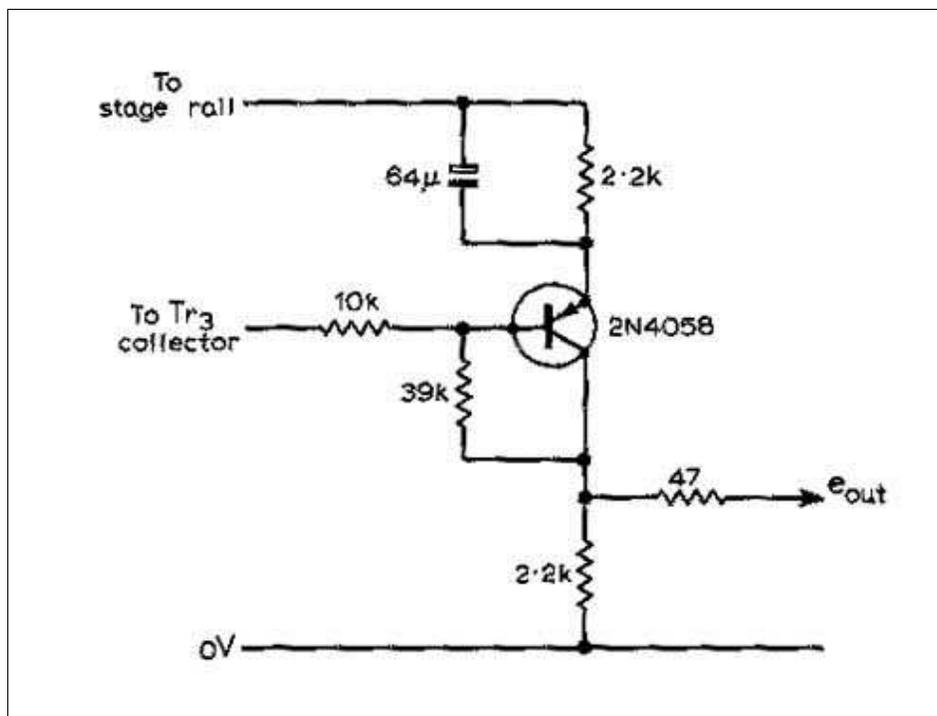

Fig. 11. Floating emitter collector-follower circuit referred to in Appendix II.

References

1. Langford-Smith, F., "Radio Designers Handbook", Vol.4 ch.72.
2. Baxandall, P. J., "Negative-Feedback Tone Control", Wireless World, October 1952
3. Baxandall, P. J., "Gramophone and Microphone Pre-amplifier", Wireless World, January 1955
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 avoidable 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.
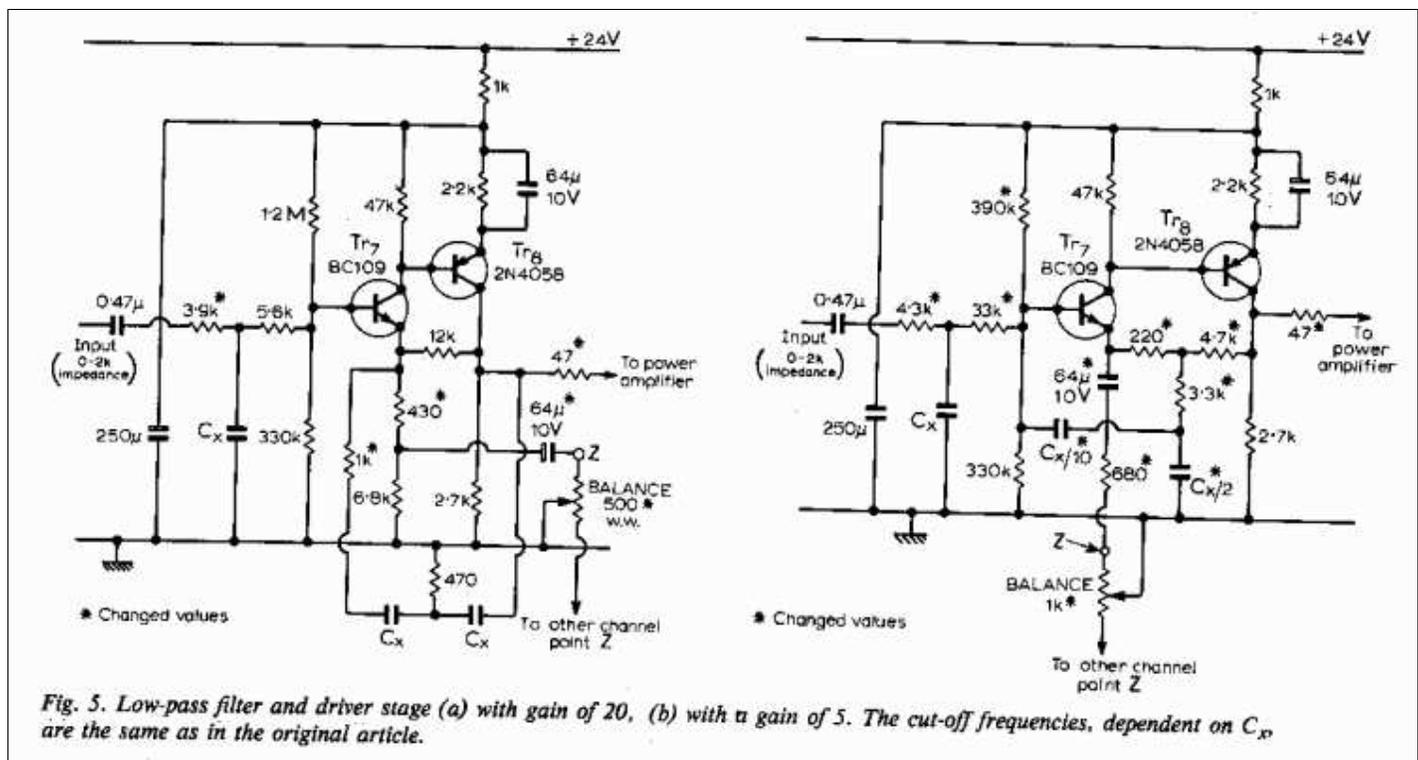
To meet this need more conveniently, two further versions of the driver amplifier, incorporating steep-cut low-pass filter characteristics which are identical to that of the original circuit, and having gains of 20 and 5, are shown in Figs. 5(a) and 5(b). An alternative, three-transistor arrangement 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.
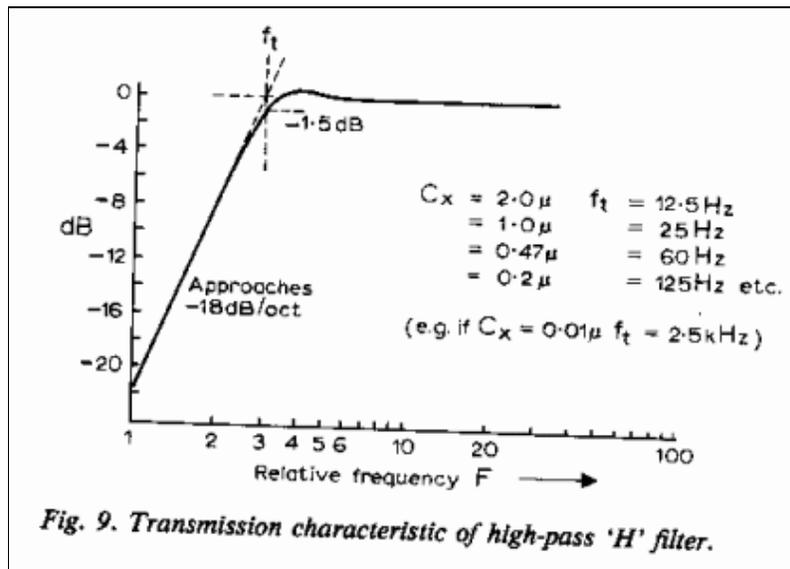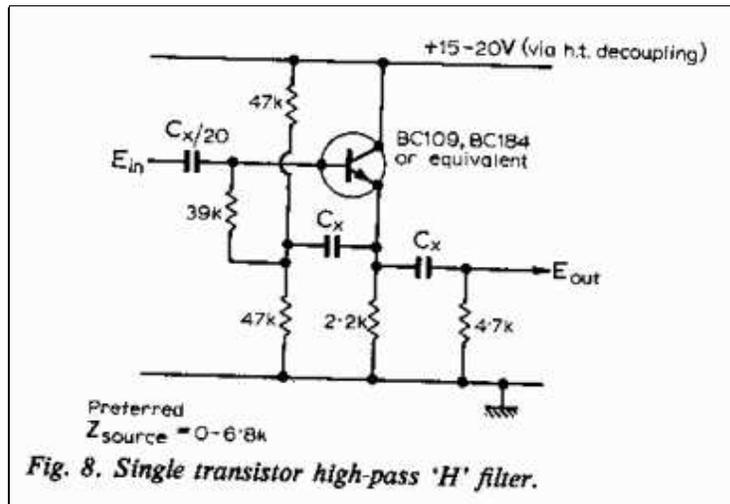
If single transistor 'H' filters are to be used at output signal levels exceeding 100mV a Darlington

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.
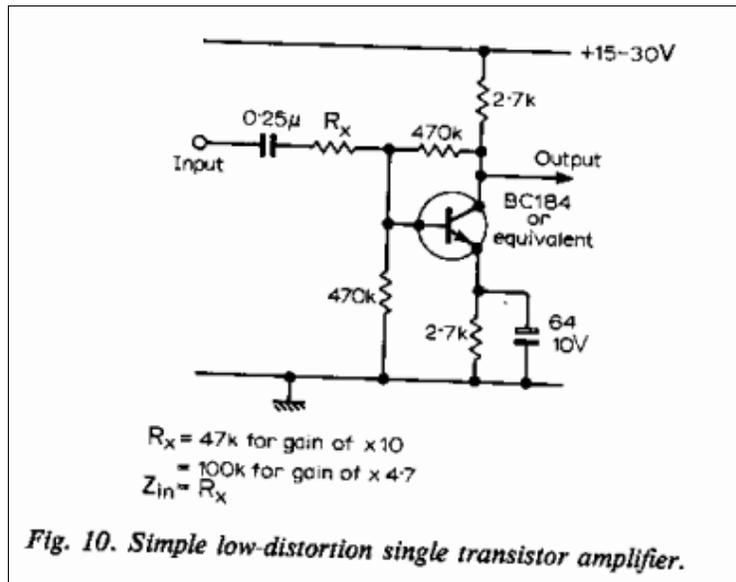
Magnetic pickup equalisation circuit. Some requests have been received for component values for the use of this circuit for tape-replay characteristic equalization. The author remains of the opinion that this type of provision is best left to the manufacturers of the tape recorder, in that the actual 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 I00mV, 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.

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