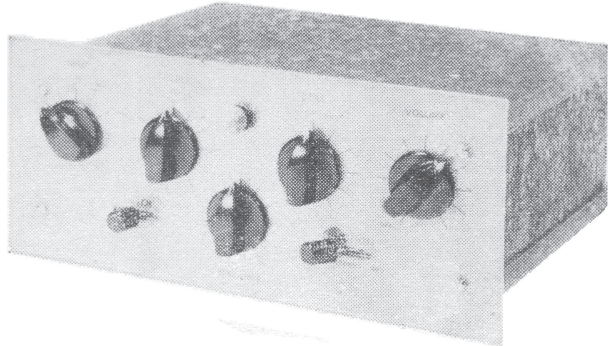
Stereo/Mono Integrated Version, Mk II



TRANSISTOR HIGH-QUALITY AUDIO AMPLIFIER

By J. DINSDALE, * M.A.

SINCE the publication three years ago of articles describing a transistorised high-quality amplifier[1] and pre-amplifier[2], prolonged listening tests and series of measurements have been conducted both on this design and on several commercial systems which are becoming available. In addition, I have received a wealth of constructive comment (and criticism) from colleagues and readers of *Wireless World,* many of whom bear out my own feelings. The appearance, moreover, of several articles on transistor amplifiers, notably by P. Tharma, T. D. Towers and the series by O. Greiter, show that the trend to transistor units has become firmly established. In short, I believe that the time has arrived to examine in greater detail some aspects of the original design and to describe some modifications which may be made to improve the performance.

General Specification

It was decided to maintain the basic specification of the original amplifier and follow normal commercial figures as far as possible. Thus the input sensitivities are 4 mV for magnetic pickups and 400 mV for crystal pickups. The "auxiliary" input may be used for radio tuner (80 mV) microphone (5 mV) or tape head (3 mV).

Treble and bass controls are provided, together with a Switched low-pass filter and an improved infinitely variable balance control. The use of low-noise transistors maintains a satisfactory background level even at full volume. The input level to the power amplifier is 100 mV for 10 watts output, and the frequency response of the 'complete system is flat within 2 dB from 35 c/s to 20 kc/s, with a total harmonic distortion of under 0.2 % at 10 watts.

Input Impedance

It was stated in the originall article (and has been reiterated more recently) that the most efficient way of designing pre-amplifiers for low sensitivity magnetic pickups and tape replay heads is to work into a low-impedance load and utilise the inductance of the transducer in series with this low impedance to achieve the h.f. de-emphasis in the replay characteristics, as shown in Fig. 1. The advantages, claimed for this mode of operation are twofold.

(i) The transistor is a current-operated device and therefore a low-impedance source will raise the sensitivity of the input stage and hence that of the pre-amplifier.

(ii) The noise generated in the input transistor may be minimised by connecting the base to ground via a low-impedance (e.g., the pickup). Unfortunately there are certain side effects which detract from the apparent

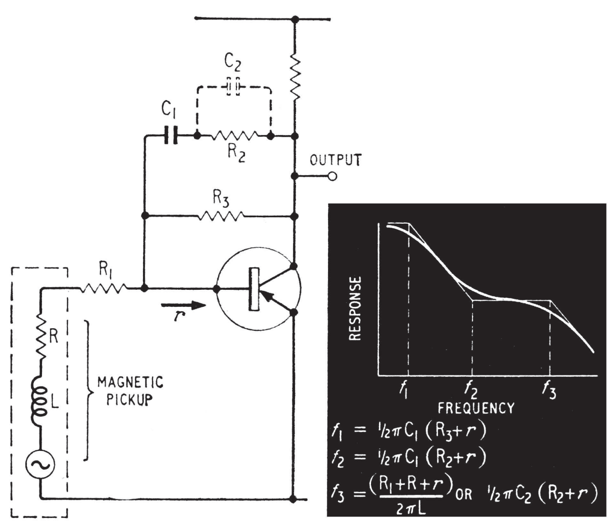


Fig. 1. Method used in the original (1961) design for equalisation of magnetic pickups. C_2 is used only when R_1 is large enough to remove the effect of L from the audio spectrum.

* Elliott Brothers (London) Ltd.

advantages of this system and, indeed, when a stereo pickup is used, can become most undesirable.

For example, one of the main problems in the design of stereophonic pickup heads is to reduce crosstalk between channels to a satisfactory level (better than 20dB) over the full frequency range, but especially above 2kc/s, where most of the directional information is found. While this figure may be approached in the modern high-quality instrument, it will only apply when the currents flowing in the component coils are negliglible, for two reasons.

(i) The magnetic field set up by the current will be sufficient to cause a reduction in channel separation due to the proximity of the coils in the confined space available inside the head causing direct transformer coupling between channels.

(ii) A more serious effect occurs with the sum-and-difference type of pickup (e.g. the Decca" ffss" and E.M.I. EPUl00) where a common coil carries the lateral signal before being combined with each phase of the vertical signal to form the complete output. Both left- and right-hand currents flow in the common lateral coil and owing to the impedance of this coil (2 to 12kΩ depending on the frequency) considerable crosstalk will result. Indeed the Decca "ffss" Mk. I head and Decca Stereo test record SXL 2057 produced 2 kc/s signals in each channel of the original pre-amplifier of only 6 dB difference, although only one channel contained recorded information. An analysis is given in the Appendix.

A further practical difficulty concerns the use of pickups with widely differing inductances (as may happen for example in demonstrations and tests). It is inconvenient to change the input resistor continually and, moreover, there comes a time when the inductance of a particular pickup is so low that even without any series resistor the input impedance of the transistor is itself too high to

TABLE I—COMPONENT VALUES FOR FIG. 2

R1	100kΩ	2%, high stability	R17	8.2kΩ	2%, high stability
R2	68kΩ	2%, high stability	R18	56kΩ	2%, high stability
R3	100kΩ	2%, high stability	R19	10kΩ	
R4	82kΩ	2%, high stability	R20	8.2kΩ	2%, high stability
R5	10kΩ	2%, high stability	R21	15kΩ	2%, high stability
R6	12kΩ		R22	3.3kΩ	
R7	12kΩ		R23	3.3kΩ	
R8	22Ω		R24	2.7kΩ	
R9	5.6kΩ		R25	2.7kΩ	
R10	1.2kΩ		R26	2.2kΩ	
R11	10kΩ		R27	6.8kΩ	
R12	4.7kΩ		R28	33kΩ	
R13	470Ω		R29	2.2kΩ	
R14	1.8kΩ		R30	2.2kΩ	
R15	180kΩ	2%, high stability	R31	1kΩ	
R16	15kΩ	2%, high stability	R32	10kΩ	If used with 40-volt supply.

All resistors ¼watt, and 5% unless otherwise stated.

C1	200μF	6V wkg. electrolytic	C14	0.12μF	5%
C2	10μF	6V wkg. electrolytic	C15	0.01μF	5%
C3	50μF	6V wkg. electrolytic	C16	0.0068μF	5%
C4	50μF	6V wkg. electrolytic	C17	0.1μF	5%
C5	10μF	6V wkg. electrolytic	C18	25μF	15V.W. electrolytic
C6	100μF	6V wkg. electrolytic	C19	0.0033μF	5%
C7	0.0047μF	5%	C20	0.01μF	5%
C8	0.015μF	5%	C21	100μF	6V.W. electrolytic
C9	0.0068μF	5%	C22	0.001μF	5%
C10	0.047μF	5%	C23	0.0047μF	5%
C11	0.0047μF	5%	C24	0.01μF	5%
C12	10μF	15V.W. electrolytic	C25	10μF	6V.W. electrolytic
C13	50μF	20V.W. electrolytic	C26	50μF	20V.W. electrolytic

In the prototype the small electrolytics used were Wima "Printilyt", or Plessey miniature (not subminiature), larger (1000μF+), Plessey "CE", and other capacitors Mullard Polyester 125V.

RV1	Bass	50kΩ+ 50kΩ linear	Morganite Type AG
RV2	Treble	25kΩ+ 25kΩ linear	Morganite Type AG
RV3	Balance	10kΩlog.+10kΩ, antilog.	Morganite Type AG
RV4	Volume	10kΩ+10kΩ, log	Morganite Type AG

S1	Input	4-pole 5-way (2-bank)	Model DH, NSF
S2	Filter	2-pole 3-way	Model TG, NSF
S3	Function	3-pole 3-way	Model TG, NSF

Vt 1,2,3. OC44,OC75, ACI07, GET880, etc. (h_{fe}>60 at I_c=1mA.)

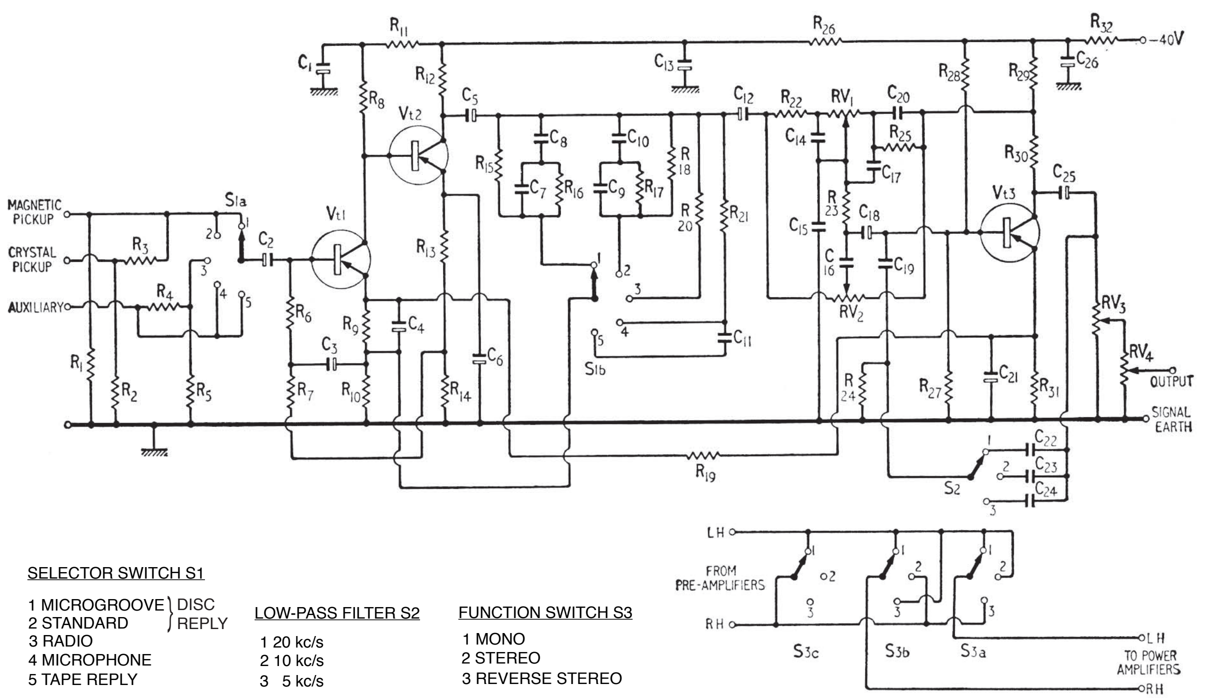


Fig. 2. Circuit of revised pre-amplifier with high input impedance.

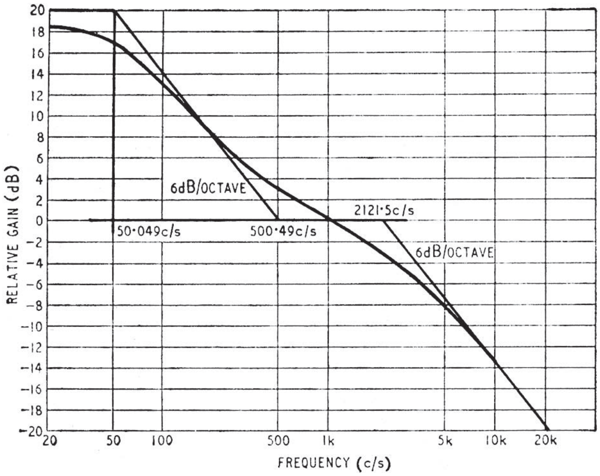


Fig. 3. R.I.A.A. microgroove choracteristic.

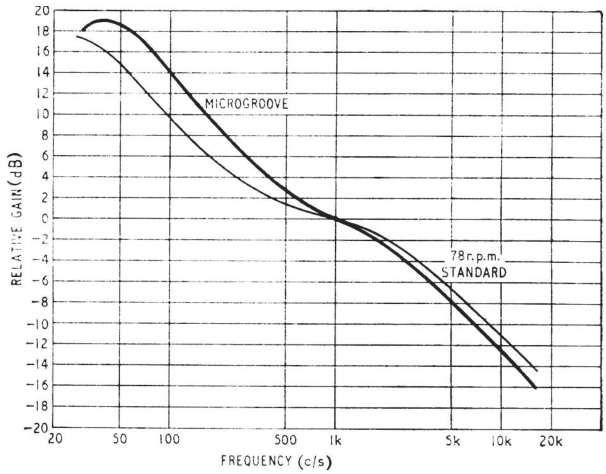


fig. 5. Disc replay response curves.

allow for accurate equalization. This also raises another point: that one should not be forced to rely on the inductance of a particular manufacturer's device being maintained to close limits over a prolonged period of production. All the above stresses the need for an input impedance high enough to keep the accompanying L/R roll-off well outside the audio band, provided that the noise figure does not rise unduly. Equalisation for the recording characteristic may then be achieved by conventional feedback networks.

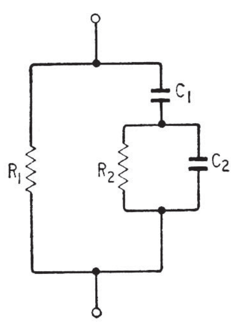
Revised Pre-amplifier

The new pre-amplifier circuit Fig. 2 follows the basic form of the original, but with a completely new input stage. Two transistors are used in this stage with heavy a.c. feedback via C3 and R6. This has the effect of increasing the input impedance at the base of Vt1 to about 500kΩ, while at the same time keeping noise to a minimum since the base of Vtl can still be loaded to ground via the impedance of the transducer. (This, of course, would not happen if the problem of raising the input impedance had been solved merely by adding a suitable series input resistor and increasing the sensitivity.) Since the actual impedance is frequency-dependent and is in any case far higher than is normally required, a padding resistance has been incorporated in each input network to stabilise the value over the audio band.

Equalisation is again performed by feedback, but taken this time from the collector of Vt2 to emitter of Vt1, since the first stage is no longer a virtual-earth amplifier with an input impedance of effectively zero, and any feedback from the collector of Vt2 to the base of Vtl would reduce the high input impedance.

This first stage has a voltage gain of approximately 10 in its most sensitive mode (microphone and magnetic pickup inputs) and is followed by the output stage consisting of a single transistor Vt3. This stage is very similar to the original article, and incorporates negative feedback tone controls and low-pass filters. Overall negative feedback from the emitter of Vt3 via R_{19}, provides additional stability at very low frequencies.

Fig. 4. feedback network for R.I.A.A. disc equalisation.



$33\frac{1}{3}$ AND 45r.p.m.	78r.p.m.
C_1R_1 = 2,940	C_1R_1 = 2,780
C_2R_2 = 81·2	C_2R_2 = 57·3
$\frac{R_1}{R_2}$ = 12·4	$\frac{R_1}{R_2}$ = 7·08

(C IN MICROFARADS, R IN OHMS)

Magnetic Pickup Input

Since all pickups operating on the elctromagnetic principle have a velocity characteristic, i.e. the output is proportional to the velocity of the stylus, the output will be identical to the recording characteristic and equalisation is necessary. It was decided to provide equalisation to the RIAA specification for both microgroove and standard recordings, since this has been the International Standard since 1955. The microgroove characteristic is detailed in Fig. 3, and the calculated turnover frequencies occur at 50.049c/s, 500.49c/s and 2121.5c/s. The characteristic may be achieved by three separate networks with a buffer stage betwen each to prevent interaction, but if, as in most applications, a single network is to be used, the parameters are as shown in Fig. 4. A convenient way of determining the component values is to choose R, to provide the required sensitivity and then calculate the remaining values. Thus in the present design if R,= 15kΩ, for microgroove,

R_1=186 kΩ (omitted since its effect is negligible)†
C_1 = 16,300 pF, nearest value 15,000pF
C_2 =5,400 pF, nearest value 4,700 pF
Similarly if R, =8·2 kΩ for 78 r.p.m. standard discs,
R_1=58kΩ (omitted)†
C_1 =48,000 pF, nearest value 47,000pF
C_2 = 7,000 pF, nearest value 6,800 pF

In the prototype, the above values were found to give

† These resistors will, in fact, reduce the response by about 1dB at 50c/s, and provision for fitting them if desired has been made on the printed circuit. The curves in Fig. 5 were measured without these resistors.

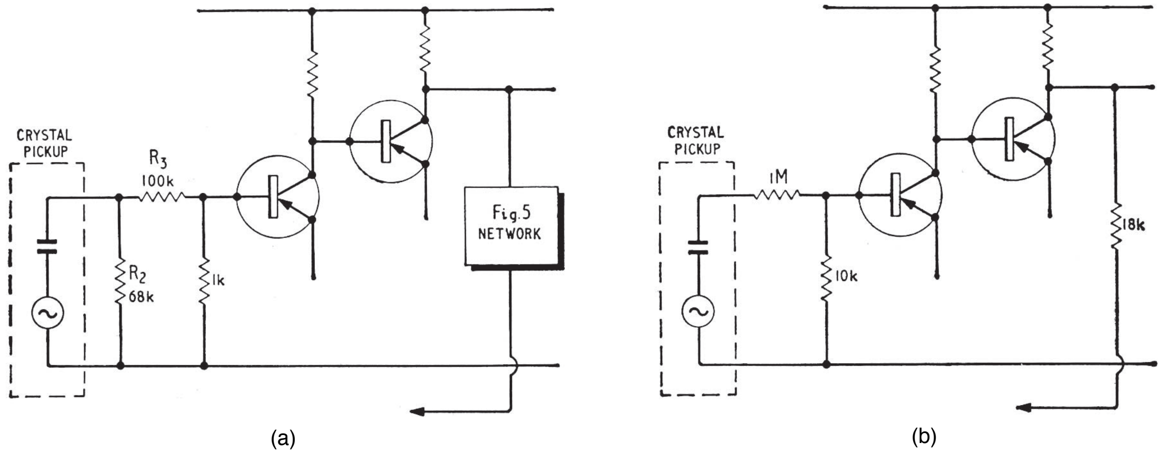


Fig.6. Two methods of equalisation for crystal pickups. (a) Velocity loading. (b) Normal loading.

responses within 1 dB of the standard curves as shown in Fig. 5.

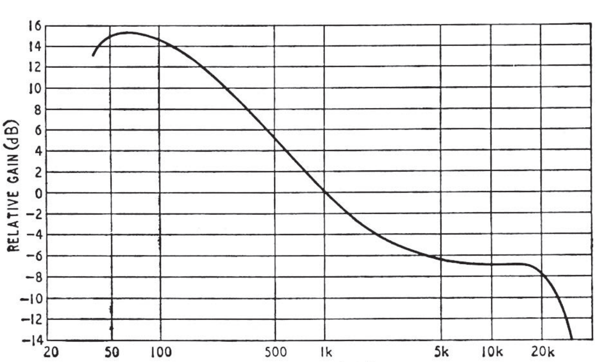
The frequency of the roll-off produced by the inductance of the pickup is given by $f = \frac{R_{in}}{2 \pi L}$ where R_{in} is the input impedance of the pre-amplifier (100 kΩ in this case). Thus for an inductance of 600 mH, a high figure, $f = \frac{10^5}{2\pi \times 0.6} = 25\text{kc/s}$ which is sufficently high to avoid interference with the characteristic.

Crystal Pickup Input

Piezoelectric (crystal) transducers are capacitive sources, and produce an output proportional to the amplitude of the signal. Thus provided negligible current flows no equalisation is necessary; an input impedance of 1 MΩ to 2 MΩ will achieve this, and an additional position- "Crystal pickup" –may be added to S1, the input consisting of a 1MΩ and 10 kΩ potentiometer (to allow for the high output level of crystal pickUps and also to load the base of Vt1 to ground via a low impedance to preserve the signal-to-noise ratio), and a with a single 18 kΩ resistor in the feedback network. This will give a sensitivity of 400 mV, which is typical of modern high-quality instruments. However, in the prototype it was decided to utilise the existing equalisation networks by "velocity loading" the crystal pickup, i.e. loading the pickup until the output approximated to an electromagnetic characteristic and then equalising as before. A load of approximately 68 kΩ will achieve this with the majority of pickups, and the resistor R_2 on the crystal pickup input provides this impedance.

It is difficult to maintain the necessary loading requirements for both magnetic and crystal pickups in this mode of operation, and the solution adopted here is to load the magnetic pickup input to ground via a low resistance when a crystal pickup is to be used. This may be achieved either by the use of jack plugs and sockets or (in this instance) by inserting a spare coaxial plug containing the extra resistor. It must be emphasised that this is by no means an accurate" impedance matching" component –it is used solely to allow for the high output of these pickups and to ensure a very low noise figure. A 1kΩ resistor will result in a sensitivity of 400 mV as before. The two arrangements are shown in Fig. 6. If a higher output device is used additional attenuation may be provided by reducing the additional resistor (and, incidentally, improving still further the signal-to-noise ratio). Thus if the resistor is made 470 ohms, the sensitivity is reduced to 1 volt.

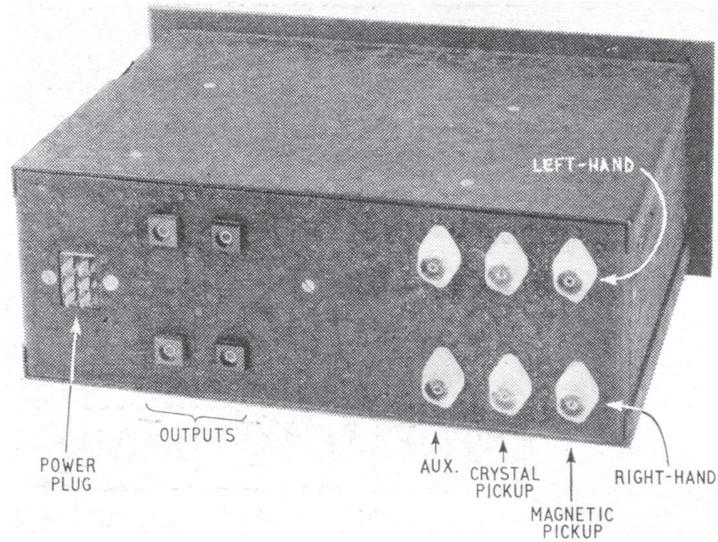
Fig. 7. Tape replay characteristic with equalisation for 7½in/sec.



Auxiliary Inputs

In the "Radio" and "Microphone" positions a flat response is achieved by simple resistive feedback via R_{20} and R_{21}, sensitivities being 80 mV and 5 mV respectively. The" Tape" input provides equalisation to the C.C.I.R. standard for a tape speed of 7½ inches per second. The characteristic within 1 dB of the standard from 70 c/s to 15 kc/s is shown in Fig. 7 and is independent of the replay head inductance. If preferred, equalisation may be carried out at the tape deck, and the selector switch set to the "Radio" position for a flat characteristic. The sensitivity as drawn is 3 mV.

The signal-to-noise ratio is typically −70 dB (wideband measurement), but this may be increased to −85 dB by selecting transistors. Harmonic distortion is of the order of 0.01 %, for signals up to 20 dB overload, when it rises sharply.



Rear view showing arrangement of input and output sockets.

The pre-amplifier will operate off the main amplifier 40-volt line with a suitable series dropping resistor R_{32}. Otherwise a 12-volt source is adequate. The current consumption is 2·7mA

Tone Controls

The tone control and filter circuits are very similar to the original design, being performed by negative feedback around the final stage. The tone control consists of a Baxandall network at the input to Vt3. For correct operation, the two capacitors C_{14} and C_{17} should be close-tolerance components. The use of switched tone controls is of doubtful value in a mono equipment except to obtain an accurate "level" position at the centre of the control. However, in stereo systems the more accurate ganging between channels obtainable with switched controls may be an added advantage. Ganged potentiometers are now available, however, accurate to within 1 dB at little extra charge. The tone control characteristics are shown in Fig. 8.

Three fixed low-pass filters were provided giving cut-off frequencies of 20 kc/s ("flat") 10 kc/s and 5 kc/s As in the original design, the slope of attenuation varies with the setting of the corresponding tone control, since both filter and tone control are achieved by feedback round the same transistor. The maximum boost position of the tone control gives the greatest slope of the corresponding filter. This ensures maximum discrimination against frequencies outside the audio band, when they would otherwise prove most objectionable. The filter characteristic is shown in Fig 9.

Balance

The original form of balance control (Fig. 10) was abandoned since its operation (by varying the collector load of the output transistor) resulted in noise at the output, and gave insufficient variation in channel gains to compensate (for example) for a different loudspeaker on each channel. In addition the tone controls were affected unequally by altering the fraction of the output fed back, since this fraction is determined by the ratio $R_A : R_A + R_B$, and for identical performance from both channels this ratio must be maintained.

The new design uses a conventional log/antilog ganged potentiometer at the output for minimum attenuation at the balance point and infinite variation beween channels.

It would be possible to use a dual ganged linear potentiometer, but as will be seen in Fig. 11, this results in an attentuation of 50% at the central (balance) point,

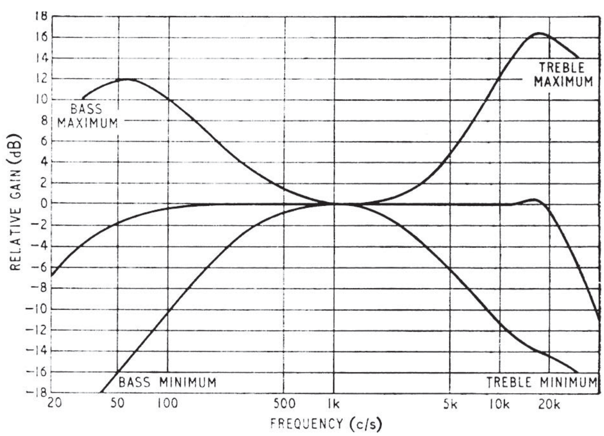


Fig. 8. Pre-amplifier tone control characteristics (low-pass filter at 20kc/s).

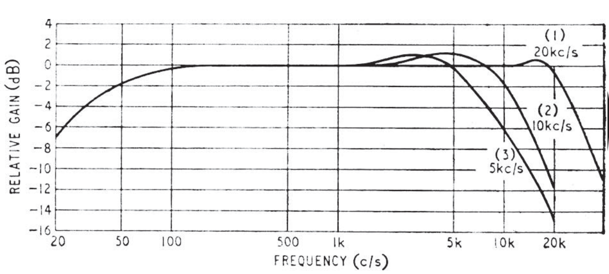


Fig. 9. Low-pass filter response (radio input).

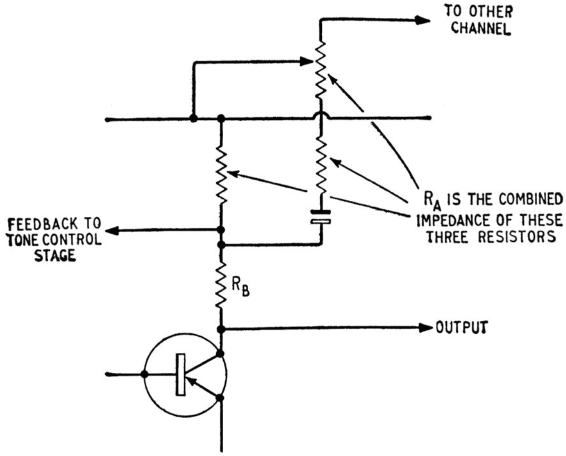


Fig. 10. Original balance control.

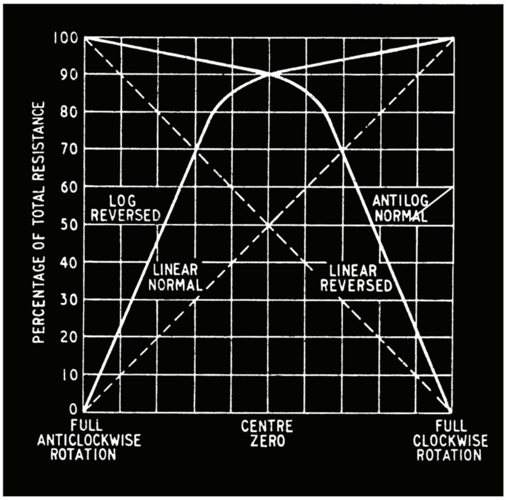


Fig. 11. Dual-gang log-antilog balance control gives less attenuation at centre point.

while the use of a log/antilog combination results in an attenuation of only 10%.

Since the volume control further loads the wiper of the balance control to earth via 10kΩ, the above attenuation figures become approximately 60% and 15% respectively.

Power Amplifier

Turning now to the power amplifier circuit, Fig. 12, several points arise that may be modified quite readily to give improved performance.

(i) When switching on this amplifier there is a loud "plop" from the loudspeaker owing to the output capacitor C_7 charging to approximately half the supply voltage. While this may be aggravating (acting as it does like the "Surprise" in Haydn's Symphony) it does at least signify (in the absence of hum and noise) that the equipment is "on." However, on a more serious note it may spell ruin to a loudspeaker system costing four times the price of the amplifier. A simple, if bulky, remedy is to use two capacitors in series across the supply, thus providing an artificial a.c. centre tap.

(ii) The bootstrap capacitor C_6, while linearising the l.f. response may also induce distortion by pulling Vt3 into the non-linear (bottoming) region of its characteristic on large negative-going signals. It has therefore proved worthwhile to connect a 1 kΩ resistor in series to minimise this effect. The capacitor still improves the l.f. response, though to a slightly less extent.

In both the original articles the need for correct earthing was emphasised since a 1-amp pulse of current in 1 milliohm of wire will produce a p.d. of 1 millivolt, one quarter of the input sensitivity on magnetic pickups. The current pulses in the earth line are asymmetrical, causing severe even-harmonic distortion.

Unfortunately, the stereo equipment will inevitably destroy the utility of all the above by causing an obvious but unavoidable earth loop. Fig. 13 shows two similar amplifiers operating from the same power supply (with individual decoupling). The signal earth points at the inputs will both differ by several hundred millivolts from true earth depending on the signal in each channel. Connection of a commercial stereophonic pickup or microphone with a common earth line now causes each

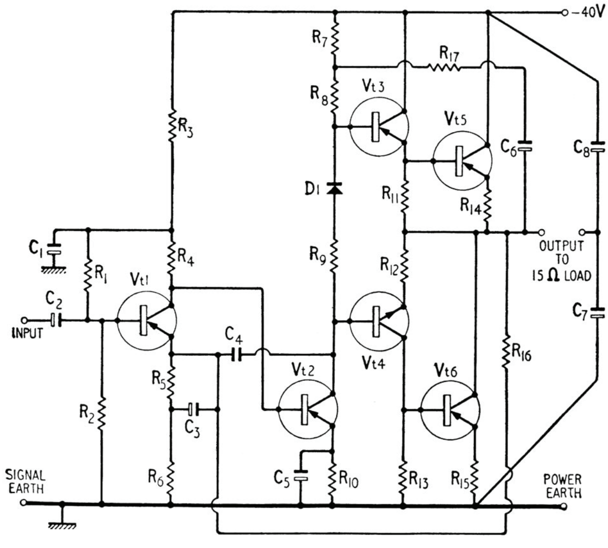


Fig. 12. Circuit of original power amplifier.

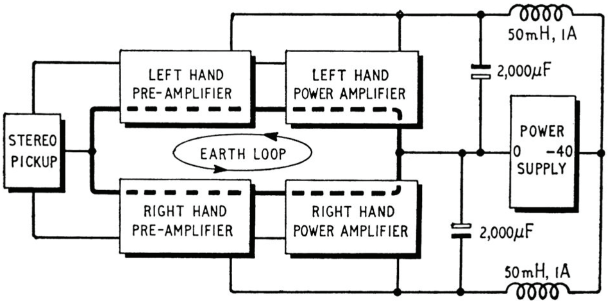


Fig. 13. Earth loop in stereo system.

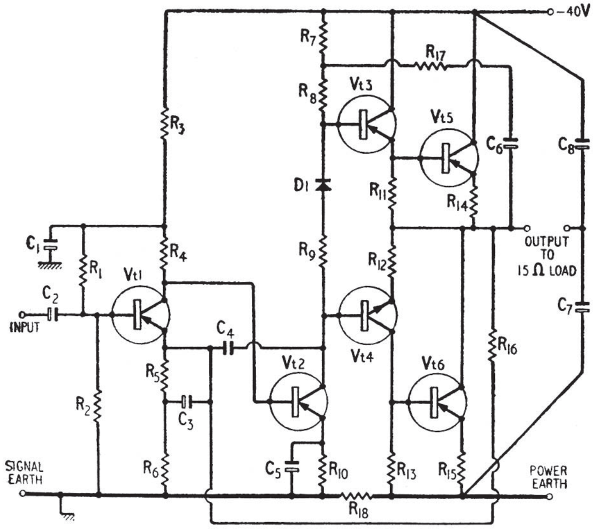


Fig. 14. Circuit of revised power amplifier.

TABLE II—COMPONENT VALUES FOR FIG. 14

R1	330kΩ	Note A	R10	560Ω	
R2	56kΩ		R11	150Ω	
R3	68kΩ		R12	10Ω	
R4	22kΩ		R13	150Ω	
R5	220Ω		R14	1Ω	1 watt, wirewound
R6	33Ω	2% high stability	R15	1Ω	1watt, wirewound
R7	1kΩ		R16	3.9kΩ	2% high stability
R8	4.7kΩ		R17	1kΩ	
R9	22Ω	Note B	R18	10Ω	10%
All resistors ¼ watt and 5% unless otherwise stated.					
C1	50μF	25V.wkg.electrolytic	C5	200μF	6V. wkg. electrolytic
C2	10μF	15V.wkg.electrolytic	C6	25μF	50V. wkg. electrolytic
C3	200μF	6V.wkg. electrolytic	C7	1,000μF	50V.wkg. electrolytic
C4	1,000pF		C8	1,000μF	50V. wkg. electrolytic
D1	OA5				

Note A. Adjust for collector of Vt6 to sit at half supply volts ± 1 volt.
Note B. Adjust for output stage quiescent current (measured as difference in supply current when R_9 is shorted out) to be 15mA ± 5mA.

channel to become distorted by the even harmonic products of both channels, causing unpleasant distortion particularly when one channel has a transient such as a cymbal clash.

To overcome this effect it is necessary to add some small impedance (but large compared with the few milliohms of lead resistance involved) into the earth loop in such a way as not to accentuate the even-harmonic distortion. Such a place is within the main feedback loop of the power amplifier where no large voltage amplifications take place. Fig. 14 shows the final power amplifier circuit. Although this, causes a slight increase in overall distortion on mono signals the improvement on stereo signals is impressive. (A far more costly but admittedly more elegant method would be to use two independent power packs).

The original design described two versions:
Version 1-40 volt supply-10 watts in 15 ohm load
Version 2-24 volt supply-10 watts in 3 ohm load
or 3½ watts in 15 ohm load

In view of the large currents flowing in the version 2 design, and the difficulty in avoiding earthing problems, version 1 is to be preferred although the lower voltage design will give very satisfactory results at lower power levels. All the component values and performance figures mentioned here refer to version 1. The circuit diagram of the 40 V power supply is given in Fig. 15.

Certain minor alterations have been made to the component values in the light of further knowledge of component tolerances, and to improve the bass response. The quiescent current in the output stage (measured as the variation in supply current which occurs when Vt3 and Vt4 bases are shorted together) should be between 10 and 20mA, R_9 being altered if necessary to obtain this value. Similarly the output sitting-point may be set up by altering R_1. It is important to mount the thermal stabilising diode D_1 on the same heat sink and in close proximity to the output transistors.

Acknowledgements

The author gratefully acknowledges the work of R. Tobey who designed the basic circuit of the improved

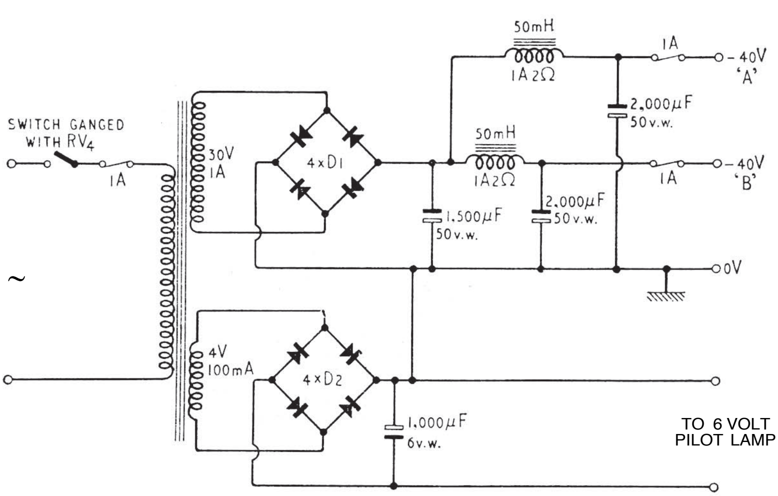


Fig. 15. Circuit of 40-V power supply.

pre-amplifier and offered valuable comments and Criticism during the development. He is also indebted to P. Lepine for carrying out the experimental work in optimizing and plotting the response curves.

[To be concluded. Full dimensioned drawings of metalwork, printed circuit and wiring diagrams will appear in the next issue together with hints on construction.]

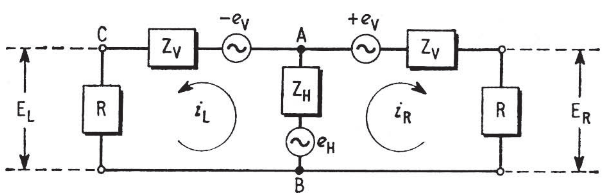
REFERENCES

[1] R Tobey and J. Dinsdale: "Transistor Audio Power Amplifier." *Wireless World,* November 1961.
[2] R Tobey and J. Dinsdale: "Transistor High Fidelity Pre-Amplifier." *Wireless World,* December 1961.

APPENDIX

$$i_L = \frac{(e_H - e_V)}{|Z_H + Z_V + R|}$$ and similarly i_R.

$$E_L = (e_H - e_V) - i_L|Z_V + Z_H| + i_R|Z_H|$$ and similarly E_R.



R = INPUT IMPEDANCE OF AMPLIFIER E = APPLIED SIGNAL FROM PICKUP

The channel separation is given by

$$\frac{\dfrac{(e_H - e_V)}{|Z_H + Z_V + R|}\cdot Z_H}{(e_H - e_V) - \left(\dfrac{(e_H - e_V)}{|Z_H + Z_V + R|}\right)(|Z_V + Z_H|)} = \frac{\dfrac{|Z_H|}{|Z_H + Z_V + R|}}{1 - \dfrac{|Z_H + Z_V|}{|Z_H + Z_V + R|}} = \frac{|Z_H|}{|Z_H + Z_V + R| - |Z_V + Z_H|}$$

Where R is large, the effects of Z_V and Z_H are negligible. However where R is of the same order as $|Z_H + Z_V|$, about 2 to 12K, the channel separation will be seriously degraded; and since Z_H and Z_V are frequency-dependent, this effect will worsen with increasing frequency.

In addition, the lateral vertical sensitivity will depart from the ideal ratio of 1 (an effect pointed out by Mr. D. G. Jaquess). The effective lateral sensitivity measured between points A and B will be $(e_H - 2i_L|Z_H|)$ because both currents flow in Z_H, while the effective vertical sensitivity measured between points A and C will be $(e_V - i_L|Z_V|)$ Thus the ratio $\left(\dfrac{e_H - 2i_L|Z_H|}{e_V - i_L|Z_V|}\right)$ will be dependent on both frequency and current and hence on loading.

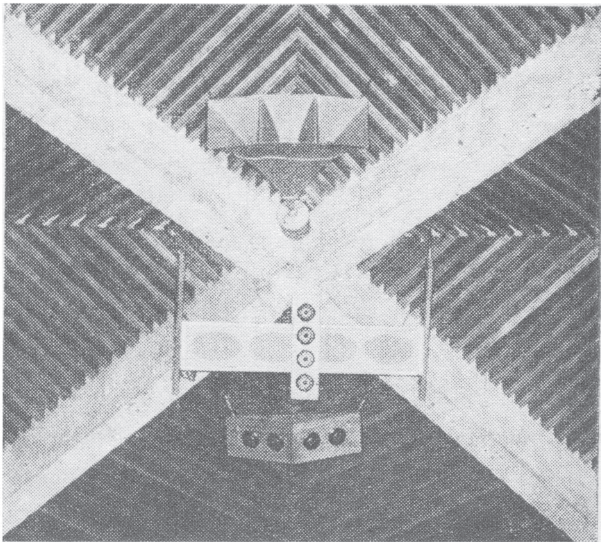
Measurements taken on the Decca "ffss" Mk. II pickup tend to support this theory, but there are other factors involved, such as the mutual coupling mentioned earlier.

CATHEDRAL SOUND

MANY of the beautiful architectural features of the new Coventry Cathedral-the high absorbent roof canopy, the tapestry at one end of the nave and the highly reflecting engraved glass curtain windowat the other-have presented unique problems for the acoustic consultants who have been called in to provide electroacoustic reinforcement, so that all members of the large congregations can hear well. Sound projected longitudinally down the aisle, even when suitably delayed to compensate for time lags in propagation, has failed to remove confusion of sound because it is overlayed by the strong return from the end window.

A new system, designed by F. H. Brittain and his colleagues at the Hirst Research Centre of the General Electric Company Ltd., relies on a single group of loudspeakers high in the roof and has satisfactorily solved the problem. The sound is directed principally toward the back of the nave where the requisite delay (about 60 millisec) is obtained simply by the increased height of the source. If the sound level is correctly adjusted one is not conscious that the sound is coming downwards, but the steep angle ensures that reflections from the floor of the aisle and from the back window are returned upwards instead of longitudinally over the congregation, and are soon lost in the absorbent roof canopy. The directional properties of the loudspeaker array are such that in the front rows there is negligible reinforcement (delayed) to interefere with direct hearing of the preacher's voice.

There are 27 microphone positions, controlled from a console with mixing facilities at the back of the navc. Audio power is supplied by two 50 watt amplifiers, but for normal speech only about 7 watts is, in fact, required.



The loudspeaker group consists of two short column arrays, a medium-frequency 3-cell horn and two pairs of h.f. units to improve articulation in the front corners of the nave.