REVISED EDITION

PHILIPS

HI-FI AMPLIFIER CIRCUITS

with

EF 86

ECC 83

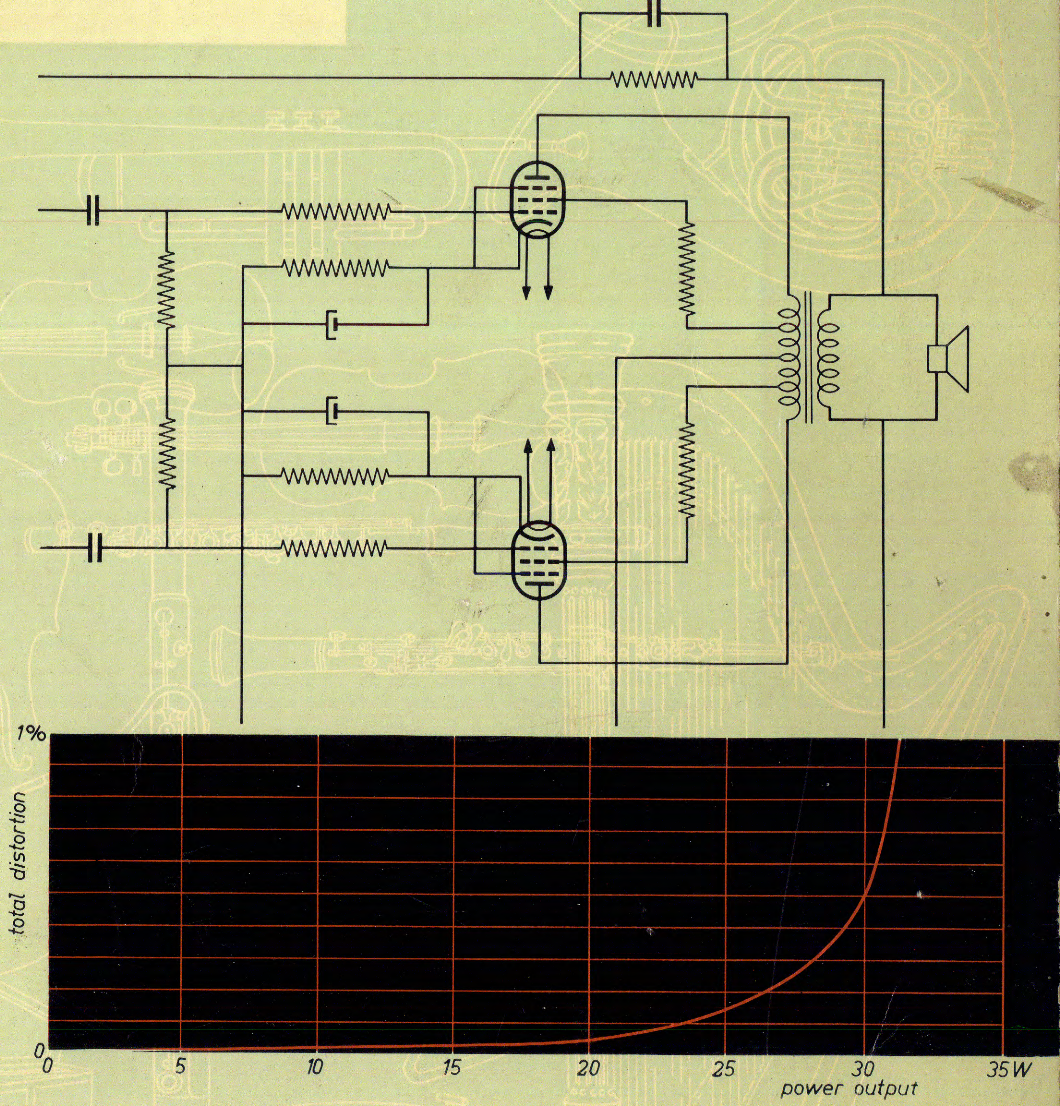
EL 34

EL 84

EL 86

EZ 81

GZ 34



PHILIPS ELECTRON TUBE DIVISION

PREFACE

The general interest in high-quality reproduction of recorded music, either on disc or tape, need not be discussed. Everyone connected with the production and sales of HiFi equipment knows that its market is growing rapidly, and this tendency will undoubtedly continue. With these perspectives in mind, the present Bulletin has been compiled. It contains not only general information on the design of HiFi amplifiers, but also a selected collection of power amplifier and pre-amplifier circuits. These circuits may be considered as representative for the present-day state of the technique as far as low- and medium-priced HiFi amplifiers are concerned.

We trust that the information given will be valuable for all setmakers who wish to compete on this interesting market. Additional advice will gladly be given on request.

CONTENTS

Chapter I

General Notes on High-Quality Amplifier Design

In general, a Hi-Fi installation consists of a turntable, a pick-up, a pre-amplifier/equalizer, a power amplifier and one or more loudspeakers. The outfit may be completed with an AM/FM tuner and a tape recorder. All these components contribute to the quality of reproduction (or to the distortion). At their best, however, they cannot do more than reproduce what is available as a signal source, viz. a recording on a disc or tape, or a broadcast transmission. The source will certainly contain some distortion. As far as linear distortion is concerned, complete correction may be achieved by the equalizer, or obtained by means of the tone controls. Against all other kinds of distortion in the signal source there is no remedy. Record producers and engineers responsible for the quality of broadcasts do their utmost to minimise distortion. As a rule they are very successful, which has resulted in an increased interest by a large public for reproduction equipment that does full justice to the fine quality of the "sources".

The amplifiers used in the chain from source to ear are of great influence on the quality. Morever, they constitute the only links of which the important characteristics such as gain and frequency response can be controlled. Nothing, whatsoever, can be done to modify the characteristics of pick-ups, tone heads and loudspeakers. The controls of the amplifiers, however, open the possibility of adapting the complete Hi-Fi installation not only to the recording characteristics of discs or tape, but also to the acoustic properties of a room. Last but not least, a listener may adjust the response characteristic in accordance with his personal taste or hearing faculties.

Since amplifiers are the only flexible parts in Hi-Fi equipment, and therefore the most interesting, some general remarks will be made first about their design. Subsequently a number of circuits is discussed of complete power amplifiers and pre-amplifiers with built-in equalizers that at present may be considered representative as far as low-and medium-priced equipment is concerned. AC/DC circuits are deliberately left out.

According to the safety regulations in most countries, the maximum current that can be measured between a terminal of an equipment and earth must not exceed 0.7 mA peak value. This imposes restrictions on the circuit impedances in such a measure that, for instance on 220 V a.c. mains, the earthing capacitors must be limited to about 7000 pF. Moreover, the output transformer secondary and hence the loudspeaker cannot be included in the overall feedback circuit. These restrictions render the design of ac/dc Hi-Fi amplifiers impossible, although a reasonable quality may be reached with a low-gain amplifier combined with a high-output pick-up coupled to the amplifier by means of an input transformer.

HIGH-FIDELITY AMPLIFIER REQUIREMENTS

The basic requirements set to a good amplifier are summarized below:

(1) low harmonic distortion (max. 0.5 %):
(2) low intermodulation distortion (max. 2 %);
(3) low beat note distortion (max.0.8 %);
(4) linear frequency response up to at least one octave above the audible range;
(5) little phase shift in this range;
(6) low hum and noise levels;
(7) ample power reserve to allow peak passages being reproduced without overload;
(8) low output resistance to provide electrical damping of the loudspeaker.

Although these requirements are, of course, of influence on the complete amplifier design, they are of pre-eminent importance on the conception of the output stage.

A good power-handling capacity is essential for realistic reproduction of orchestral music. It is generally considered that in normal rooms a peak output power of 10 W is required, but conditions in large rooms and small halls may call for a power output of at least 15 W.

There are two basic forms of output stage from which an effective output of 10 to 15 W, with low distortion, can be delivered to the loudspeaker system, viz.:

(1) the class AB push-pull pentode stage;
(2) the class A or class AB push-pull triode stage.

Both circuits have their typical merits, and the choice between them is mainly a question of economy and performance.

PENTODE OUTPUT STAGE

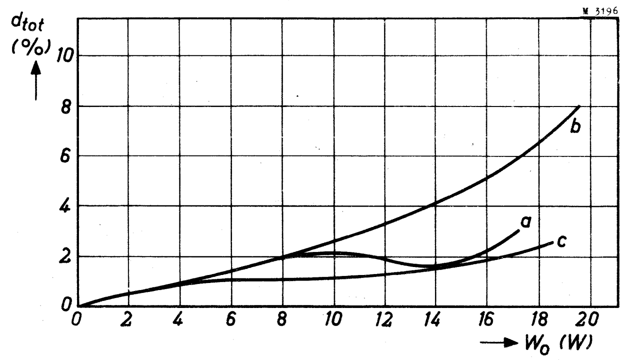
The power pentodes EL 34 and EL 84, which were introduced some years ago, have anode dissipations of 25 W and 12 W respectively. They allow the design of push-pull output stages in class AB giving well over 25 or 12 W effective output (assuming output transformers of about 80 % efficiency, which value is typical of present practice).

The overall efficiency of such stages is fairly high, being of the order of 40 to 50 %. Harmonic distortion is, however, of the order of 3 to 4 % at full output, and consequently a high degree of negative feedback is necessary to reduce distortion to a level of, say, below 0.5 % at the rated output.

The operating conditions for class AB normally recommended and published by the tube manufacturers are based on measurements with continuous sine-wave drive. The cathode resistor is so chosen that under zero signal conditions the tubes are operated in class A, whilst at full drive the working point is shifted to class B setting. The anode-to-anode load resistance is chosen for optimum performance in class B setting at full drive. The shifting of the working point is due to the influence of the increased anode and screen-grid currents on the cathode bias. For a typical output stage with two EL 84 pentodes on a supply voltage of 310 V, the increase in cathode current, and hence in control-grid bias, is about 40 % with a sinusoidal input voltage.

When, however, such a power stage is used for the reproduction of speech and music, operating conditions are rather different. The mean amplitude signal is now very small compared with the peak value which occurs from time to time, and the mean variations in cathode current are therefore also very small. Due to the relatively long time constant of the cathode resistor and its bypass-capacitor, the shifting of the working point, even under peak signal conditions, is small enough for the stage to be considered as working with a virtually fixed bias. If the normal class AB stage (cathode biased) is measured under the corresponding fixed bias conditions with a sine-wave input, it is found at high output levels that distortion is greater than when cathode bias is used. These two conditions are illustrated for the EL 84 output pentode by curves *a* and *b* in Fig. 1. The quiescent bias is the same in both cases, curve *a* showing normal published operation with cathode bias, curve *b* operation with fixed bias. These results indicate that, in practice, a cathode-biased class AB stage designed on a sinusoidal drive basis will produce increased distortion when peak passages of speech or music are being reproduced.

Fig.1. Comparison of distortion curves under steady-state sinusoidal input conditions for a pair of EL 84 tubes in class AB push-pull (*a*) with normal cathode bias, (*b*) with fixed bias under the same conditions, (*c*) with load reduced for optimum fixed-bias operation.



One method of improving performance is to adjust the quiescent operating conditions in the output stage so that they are nearly optimum for fixed bias working, although cathode bias is still used. This entails a smaller standing current and lower anode-to-anode load resistance. These changes result in larger variations in the instantaneous anode and screen-grid currents when the stage is driven, but the effect of these is at least partially compensated, since the time constant in the cathode circuit has also been increased. The excursion of the working point is still kept very small under driven conditions.

It is found that good short-term regulation of the power supply is ensured by the use of large-value electrolytic capacitors for anode and screen-grid supplies. Peak currents corresponding to near overload conditions are effectively supplied by the capacitors with a reduction in line voltage of well under 0.5 %, and the instantaneous power-handling capacity of the stage is not impaired.

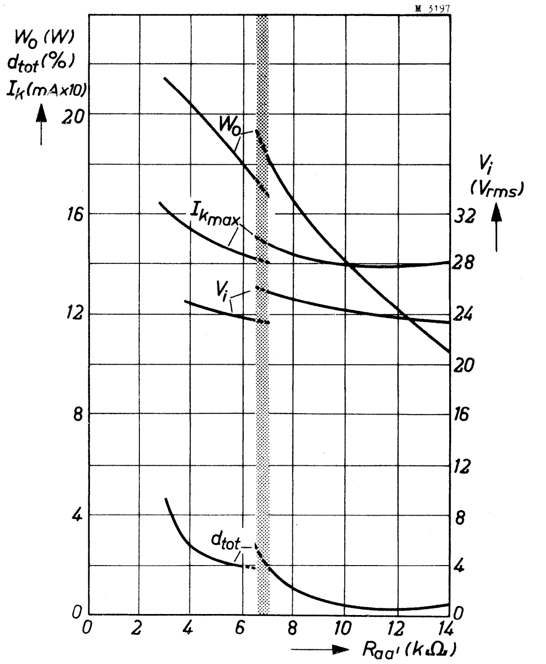
Such a design, combined with a high degree of negative feedback (26 dB), which includes the output transformer, is an alternative operating condition in the output stage of the 10 W amplifier circuit described on page 29, and has proved very satisfactory. Another feature of the use of these operating conditions is that the 12 W output pentodes each run at a mean anode dissipation of only 7.5 W. The corresponding fixed bias conditions in this case are illustrated in curve c of Fig. 1.

It should be noted that this form of operation is suitable only for use in speech or music reproduction, and cannot be fully driven with a sine-wave input without excessive distortion. For this reason it is difficult to measure directly the distortion levels which occur under practical conditions.

A second method of improving performance, described later, is to use distributed load conditions in the output stage. Depending on the precise loading used, the variation in anode and screen-grid currents can be reduced to such a level that almost identical performance is obtained under cathode and fixed bias conditions.

TRIODE OUTPUT STAGE

A low level of inherent distortion can be obtained in a push-pull triode stage operating under virtually class A conditions. It is found that with 25 W power pentodes or tetrodes, strapped as triodes, a power output of 12 to 15 W can be obtained at harmonic distortion levels below 1 % using a supply voltage of about 425 V.



Maximum power output and the corresponding distortion vary appreciably with the value of the load impedance, and Fig. 2 illustrates the typical performance of the EL 34 power pentode, triode-connected and operating slightly below its rated anode dissipation of 25 W.

Fig. 2. Performance of two triode-connected EL 34 tubes in push-pull.

For anode-to-anode load impedances below 7 kΩ either one common or separate cathode resistors (bypassed) can be used; above 7 kΩ improved operation is obtained with an unbypassed common cathode resistor. Operating conditions approach class A as the load impedance is raised, and optimum performance for high-quality output stages is obtained with a load impedance of about 10 kΩ. An output of 14 W is then delivered by the tubes with the total harmonic distortion well below 0.5 %.

This type of output stage (Williamson c.s.) has found great favour for a number of years in high-quality amplifiers giving about 12 W effective output. Because of the low inherent distortion, less negative feedback can be used to give acceptable linearity as compared with that required in pentode or tetrode output stages giving similar power output. Furthermore, in 3- or 4-stage amplifier designs, with most of the feedback applied over the whole amplifier (including the output transformer), it will be possible to attain increased stability for a given distortion level.

ULTRA-LINEAR OUTPUT STAGE

Although the triode push-pull output stage has great merits with a view to distortion, the low efficiency and limited power output are felt as serious disadvantages. This is the reason that increasing interest is being shown in various forms of distributed loading on the output stage, which circuits gained popularity under the name of *ultra-linear output stages*. These involve the application of negative feedback in the output stage itself. The screen grids of the power tubes (see Fig.3) are fed from suitable tappings on the primary of the output transformer, and the stage can be considered as one in which the negative feedback is applied in a non-linear manner via the screen grids. The characteristics of the ultra-linear stage are intermediate between those for pentode and triode operation, approaching the latter as the percentage of the primary turns, common to the anode and screen-grid circuits, increases.

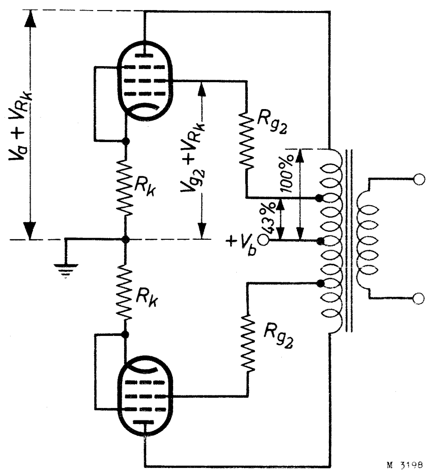


Fig.3. Circuit diagram of the ultra-linear push-pull output stage.

With the ultra-linear circuit under optimum conditions, about 65 % of the power handling capacity of the corresponding pentode stage can be realized with considerably lower distortion, whilst with an output level of the corresponding triode stage a similar order of distortion is obtained. At the same time the output impedance is reduced to a level comparable with that of the conventional push-pull triode stage.

High-quality amplifiers with power outputs of over 20 W may be designed with two pentodes of the 25 W class in an ultra-linear circuit, so that the power handling capacity is considerably greater than with triode operation. The performance of 12 W pentodes can be considerably improved with the ultra-linear circuit, although the power handling capacity is somewhat reduced. Effective power outputs of 10 to 12 W can still be obtained.

Table 1

Tube type	Mode of operation	Operating conditions V_a (V)	V_{g2} (V)	R_k (Ω)	R_{a-a} (kΩ)	R_{g2} (Ω)	Total distortion in % at 10 W	14 W	20 W	30 W
2 x EL 34	triode connection	400	+)	470 each tube	10	+)	0.5	0.7		
	ultra-linear, 43 % common winding	400	400	470 each tube	6.6	1000 each tube	0.6	0.7	0.8	1
	pentode connection	375	375	130 common	3.4	470 common	1.5	1.9	2.5	3.8

Tube type	Mode of operation	V_a (V)	V_{g2} (V)	R_k (Ω)	R_{a-a} (kΩ)	R_{g2} (Ω)	5 W	10 W	15 W
2 x EL 84	triode connection	300	+)	150 common	10	+)	1		
	ultra-linear, 20 % common winding	300	300	270 each tube	6.6		0.8	1.0	1.5
	ultra-linear, 43 % common winding	300	300	270 each tube	8.0		0.7	0.9	
	pentode connection	300	300	270 each tube	8.0		1.5	2.0	2.0

+) Screen grid strapped to anode

Table 1 gives a comparison of triode, pentode and ultra-linear operation of the EL 34 and EL 84 power pentodes. For the EL 34 in an ultra-linear circuit comparison with triode operation is the most interesting, because ultra-linear operation using a tapped primary transformer enables the power handling capacity to be more than double that possible with triode operation, whilst at the same time distortion is held to a very low level.

Although experiments showed that with a common winding ratio of 0.2 (i.e. with 20 % of the winding common to anode and screen-grid circuits), the distortion level is comparable with triode conditions, it has been found that appreciable improvement is obtained at higher outputs if the common winding ratio is further increased. The best compromise in overall performance has been obtained with the percentage of common primary winding increased up to 40 - 45 %. Although the power handling capacity is reduced, at least 35 W output can be obtained with distortion at the onset of grid current at about 2.5 %.

Fig.4 shows the performance typical of the EL 34 when used with an output transformer having primary windings tapped at 43 % of the turns. The values of the power output quoted are those delivered to the load in the secondary circuit.

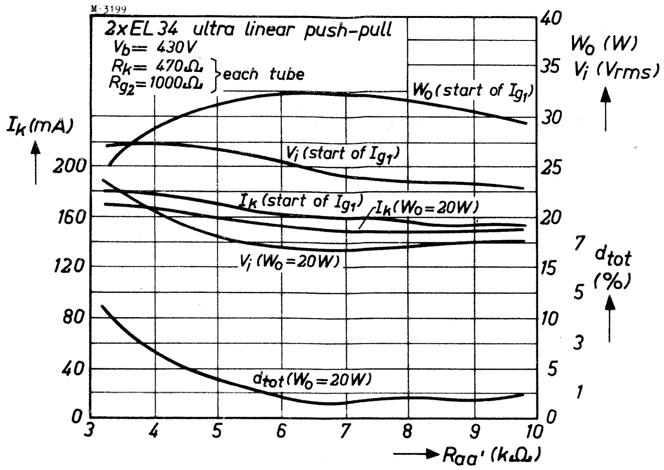


Fig.4. Performance of two EL 34 pentodes under ultra-linear push-pull conditions with screen-grid tappings at 43 % of the primary turns.

With the EL 84, comparison of ultra-linear with normal pentode operation is more significant. Considerable reduction in distortion is again obtained under ultra-linear conditions. Approximately 15 W is delivered by the tubes with a common winding ratio of 0.2.

From the figures in Table 1 little advantage would appear to be gained by further approaching triode conditions. There are, however, at least two advantages in using a tapping at about 40 % of the primary turns, particularly with the EL 34 where a high output power is still available. In the first place almost identical performance is obtained under cathode-bias and fixed-bias conditions, since with the closer approaching to class A triode operation, variations in anode and screen-grid currents are reduced when the stage is driven. Secondly, as with normal triode operation, power output and distortion are less dependent on the value of the load impedance. With the tapping at about 40 % of the turns, little change in performance is observed by a change in anode-to-anode load impedance of 6 to 9 kΩ.

NEGATIVE FEEDBACK

The distortion as mentioned in Table 1 is not considered low enough in real high-quality amplifiers. Distortion is further reduced by the application of negative feedback, usually from output to input of the complete amplifier, consisting of a pre-stage, a driver/ phase-inverter stage and the output stage (single-loop feedback). The feedback loop includes also the output transformer, and as this is a highly frequency-dependent component, the amount of feedback that can be applied depends largely upon the quality of this component.

Instability will occur in a feedback amplifier when the loop gain, being the product of gain without feedback and the attenuation of the feedback network, exceeds unity at frequencies for which the total phase shift round the loop becomes either 0 or 360° and so renders the feedback voltage in phase with the input. As the conditions for negative feedback imply a phase shift of 180°, it will be clear that instability occurs as the additional phase shift in the amplifier and feedback network approaches 180°.

Phase shift is usually difficult to measure. It is normal practice to utilize for design purposes the relationship between phase shift and attenuation. A simple high-pass or low-pass *RC* filter has an ultimate phase shift of 90°, and the attenuation approaches 6 dB/octave asymptotically. From this it follows that an ultimate phase shift of 180° corresponds to a final rate of attenuation of 12 dB/octave. Adequate margins of stability are preserved when the attenuation rates do not exceed 10 dB/octave in that part of the loop gain characteristic where it varies from about 10 dB through unity gain (0 dB) to -10 dB.

From the above it follows that the amplifier characteristics should be controlled far in excess of the audible frequency range. This control becomes increasingly difficult as the degree of feedback increases, and since the present practice is to apply from 25 to 29 dB negative feedback, it will be clear that the phase shift introduced by the output transformer sets the limit to the amplifier design.

It is hardly possible to provide a constant and high level of feedback over the whole audible frequency range in a three- or four-stage amplifier where the feedback loop includes the complete circuit and the output transformer. Adequate stability is very difficult to obtain. It is, therefore, more usual to find that the effective feedback decreases towards the highest and lowest audible frequencies.

Adequate feedback, however, must be available in the region of the fundamental resonance of the loudspeaker system, to provide the low output impedance needed for efficient electrical damping. It should also be available up to the highest frequency of which harmonics lie within the audible range, a frequency which can be taken to be about 10 kc/s.

OUTPUT TRANSFORMER

The output transformer is the most critical part of any amplifier; an incorrectly designed transformer can be the cause of distortion that may be generally looked for in another part of the amplifier. The various kinds of distortion that may originate from the transformer can be described as follows:

(1) *Frequency distortion*, caused either by too low a primary inductance, by too high a stray inductance, or by effects of resonance.

(2) *Phase distortion*, caused by phase shift when the feedback voltage is taken from the transformer secondary. As a rule this distortion takes the form of parasitic oscillations at high frequencies, caused by the phase shift due to the high stray inductance and the self-capacitance.

(3) *Intermodulation and harmonic distortion in the output stage* caused by overload at the lower frequencies when the primary inductance is too low. Firstly this leads to a reduction in effective load impedance, and secondly it causes reactive load at the lower frequencies. The elliptical load characteristic then approaches the form of a circle, which does not fit very well into the anode-current versus anode-voltage characteristics of the output tubes.

(4) *Intermodulation and harmonic distortion* as a result of the non-linear relation between the flux and the magnetic field strength in the transformer core. This distortion is always present, but can be reduced to a very low value when the flux density B_{max} is kept below a certain limit (approximately 7000 gauss with normal transformer sheet).

(5) *Harmonic distortion* introduced by too high a resistance of the transformer windings, which is also unfavourable with respect to the efficieny.

From this it follows that a good output transformer must meet a number of requirements which may be specified as follows:

(1) the primary inductance should be high,

(2) the stray inductance and self-capacitance of the windings should be low,

(3) the magnetic flux density B_{max} should not be excessive,

(4) the efficiency should be high (low losses in windings and core),

(5) matching should be correct.

It will be clear that a transformer design must be a clever compromise between these often conflicting requirements. High inductance implies a large number of turns, which is conflicting both with the resistance and with the self-capacitance. A low value of B_{max} means ample core cross-section, which also leads to large dimensions, with the same results as the requirement of high primary inductance.

The stray inductance and the capacitances, however, can be kept within reasonable limits when special winding methods are used. With such a method the stray resonant frequency can be shifted to a higher frequency range. The requirements set to the frequency response of the complete amplifier impose a minimum value on this resonant frequency. When sound reproduction free of resonance up to 20 kc/s is required, the frequency at which an attenuation of 3 dB is allowed is at least 60 kc/s. When parasitic oscillations caused by phase shift in the feedback circuit must be suppressed with one or more *RC*-filters, and the frequency response up to 60 kc/s may not be influenced, then the stray resonant frequency must even be at least 200 kc/s. (It should be remembered that the output transformer secondary forms part of the feedback circuit.)

The stray inductance can be kept low by taking the feedback voltage from a separate winding which can be coupled very closely to the primary, even with a relatively simple winding method. Experiments in a 20 W amplifier with 2 x EL 34 revealed, however, that although the stability was good with every kind of load, the frequency response curve was 7 dB down at 20 kc/s because the loudspeaker winding was not included in the feedback circuit.

Better results were obtained with a transformer in which the primary was divided into a number of parallel windings, between which the secondary windings (also connected in parallel) were applied, but the coupling between the primary windings could not be made sufficiently close, which resulted in a relatively high stray-inductance.

Improvement can be achieved by using very complicated winding methods, but such a transformer is difficult to produce. By reversing the winding directions, the capacitances can be balanced, but the slightest upsetting of the symmetry or inaccuracy during the winding process makes the improvement illusive.

The best results are obtained with a series winding method for the primary. The coil former contains two identical sections, either carrying one half primary, subdivided into e.g. five windings with the secondary windings in between. So there are ten primary and eight secondary windings, the former being connected in series and the latter either in parallel or in parallel-groups connected in series with other groups, as the matching requirements may be. This leads to a versatile transformer design which may be employed for several transformer ratios. This construction offers the further advantage that it is easy to select and to make the tapping of the screen-grid connection for ultra-linear operation.

It will thus be clear that the output transformer is a very critical and expensive component.

SINGLE-ENDED PUSH-PULL OUTPUT STAGE

It is not surprising that several attempts have been made for matching loudspeakers directly to a conventional amplifier without making use of an output transformer.

It proved to be possible to manufacture loudspeakers with a centre-tapped voice coil and having an impedance of 2 x 2000 Ω, so that they could be connected directly to a conventional push-pull output stage. This circuit, however, was not very successful. The efficiency was low due to the low inductance-to-resistance ratio of the voice coil, whilst the coupling between the two sections of the coil was too weak.

This may be explained by investigating the conventional push-pull output stage as shown in Fig.5*a* and its equivalent circuit (Fig.5*b*). It is seen that the equivalent circuit contains two separate generators the outputs of which are coupled only magnetically by the output transformer. Imperfections in this coupling introduce distortion. The coupling between the two windings of the high-impedance loudspeaker mentioned above is inferior to that of a transformer, so that the distortion is increased.

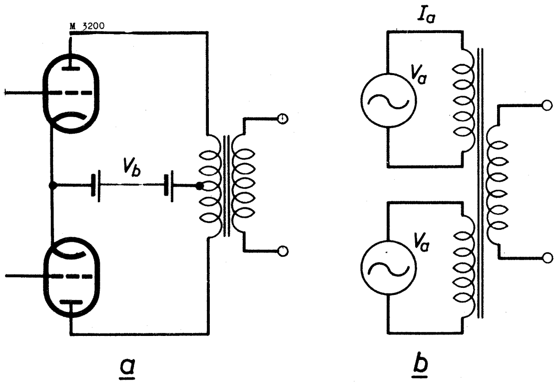
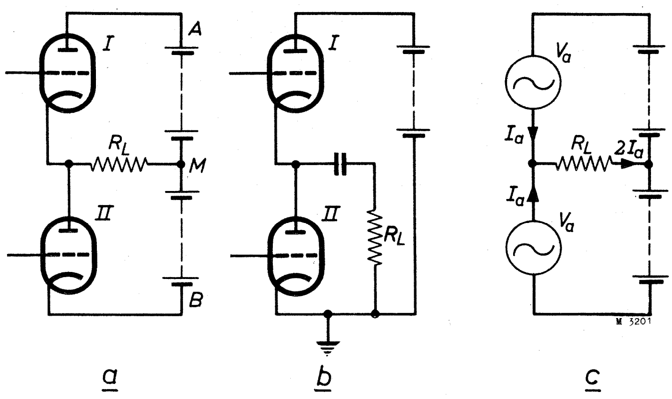


Fig. 5. Basic circuit (*a*) and equivalent circuit (*b*) of a conventional push-pull stage.

Fig.5*a* reveals that the output tubes, as far as the d.c. supply is concerned, are connected in parallel but in series to the (a.c.) load. If the supply and the load are interchanged, the circuit of Fig.6*a* is obtained. To obtain the same output power, the supply

Fig. 6. Basic circuit of a single-ended push-pull output stage, (*a*) with the load resistance connected to the centre tapping *M* of the supply source, (*b*) with the load resistance connected to earth. The equivalent circuit is shown in (*c*).



voltage should now be twice that of the normal push-pull circuit; the anode current, however, is halved, so that the input power is kept constant. The load resistance, which is connected to a centre tap on the supply source, is one quarter of the anode load resistance (R_{aa}) of the conventional push-pull circuit (Fig.5*b*), in which

$$R_{aa} = 2V_a/I_a,$$

whereas in the single-ended push-pull circuit according to Fig.6c:

$$R_L = V_a/2I_a = \tfrac{1}{4}\,R_{aa}.$$

The circuit of Fig.6*a* has the disadvantage that the load (the voice coil) is at a high potential with respect to earth. The circuit of Fig.6*b* shows that it can be separated from the d.c. source and connected to earth potential by means of a series capacitor. This circuit has the additional advantage that the supply source need not be centre-tapped.

The optimum load resistance of the single-ended push-pull circuit can be reduced by using tubes which are specially designed for the purpose, such as the EL 86, PL 84 and UL 84. These tubes have a low d.c. resistance so that a high anode current flows at a relatively low anode voltage. Moreover, their internal (a.c.) resistance and hence their optimum load resistance are considerably lower than with conventional output pentodes.

The principle of the single-ended push-pull output stage can be further modified, leading either to extremely high quality amplifiers, or to very economical output stages, depending on whether high fidelity or saving in the layout is considered of greater importance. The latter is of particular interest for the design of radio receivers. Circuits have been developed in which the output stage operates simultaneously as a phase splitter. These circuits can be realized with a relatively small number of components. The quality of reproduction and the output power are quite favourable compared with the normal class A output stage with matching transformer. A discussion of these circuits, however, is beyond the scope of this documentation in which only designs with the highest quality of sound reproduction are described.

TYPICAL PROBLEMS INHERENT IN SINGLE-ENDED PUSH-PULL CIRCUITS

The single-ended push-pull circuit of Fig.6 operates satisfactorily when triodes or triode-connected-pentodes are used. When pentodes with their much higher efficiency are to be used in the output stage, the screen-grid supply offers some problems that will be discussed in short.

The screen-grid supply (see Fig.7) should be so arranged that the screen grids are at the direct potentials of the corresponding anodes, but should carry no signal voltages with respect to the corresponding cathodes. The simplest way of achieving this for tube *II* consists in feeding its screen grid from the H.T. line via a dropping resistor R_1, bypassed to the cathode via C_3. The dropping resistor should be of such a value that the screen-grid voltage is about ½ V_b under no-signal conditions. However, the screen-grid current increases with the drive, so that the screen-grid voltage drops correspondingly, resulting in a shift of the tube setting and increased distortion.

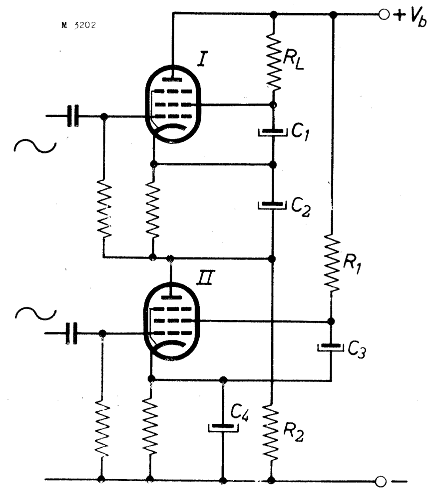


Fig. 7. Single - ended push-pull output stage with pentodes. The screen grid of tube *II* is fed via R_1, that of tube *I* via the load resistance R_L. The anode current of tube *II* exceeds that of tube *I*, since the tubes are connected in series. Therefore R_2 is shunted across tube *II* to restore the balance.

It is even more difficult to obtain the screen-grid voltage for tube *I*, since the voltage drop between the anode and the screen grid should be negligible. A screen-grid dropping resistor forms a shunt across the output as far as the signal voltage is concerned, so that a resistor of a low value would consume a considerable part of the available output power. The simplest solution is to feed the screen grid of tube *I* via the load resistance R_L itself, hence by connecting the high-impedance loudspeaker between the anode and the screen grid of tube *I*. The relatively low direct current in the screen-grid circuit will not affect the operation of the customary 800 Ω loudspeaker, and full output power is delivered to the voice coil without loss. It may be objectionable, however, that the voice coil is only capacitively earthed, and at a high positive potential. To prevent flash-overs in the loudspeaker between the voice coil and the metal parts, the latter should be connected to the H.T. instead of being earthed, but this may be conflicting with safety regulations.

A compromise is to use a dropping resistor of e.g. 6.8 kΩ in the screen-grid circuit and to connect the loudspeaker to earth, feeding it via the usual electrolytic capacitor. This circuit gives satisfactory operation for tube *I*, but the loss in output power is about 12 %. Another solution is the use of a choke instead of a resistor. Its dimensions can be much smaller than those of a normal output transformer, and its air gap may be small in view of the low direct current flowing through the choke. An inductance of about 5 H is sufficient at a load resistance of 800 Ω and a lower frequency limit of 30 c/s.

The screen-grid voltage of tube *II* may be stabilised by using a voltage stabilizing tube or a VDR resistor. It is also possible to take a voltage of ½ V_b from the H.T. supply unit, especially when it is equipped with four semi-conductor rectifiers in a bridge circuit as indicated in Fig.8. The required voltage is present at the centre tapping on the supply transformer secondary. The smoothing choke should be included in the common negative lead, so that the screen-grid current of tube *II* is also smoothed. The voltage drop across the choke gives rise to some asymmetry in the supply voltages, as a result of which the output is slightly reduced. It may therefore be preferable to split the choke winding into two identical halves, which are included in the positive and negative leads.

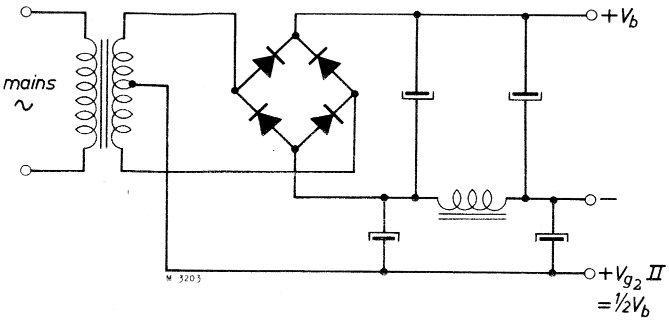


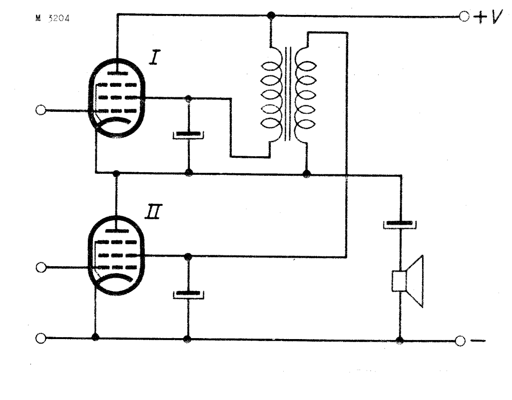
Fig.8. Circuit of an H.T. supply with metal rectifier. The screen-grid voltage for tube *II* is taken from a centre tapping in the output transformer.

Since the two output tubes are connected in series as far as the d.c. is concerned, the anode current of tube *II* will exceed that of tube *I* by the value of the screen-grid current of the latter. The setting of the tubes will therefore differ in spite of the fact that their cathode resistors are of the same value. To meet the conditions imposed by push-pull operation, it is necessary to bypass tube *II* by a resistor (R_2 in Fig.7) which compensates the influences of the screen-grid current from tube *I*, so that the cathode currents and the settings of the two tubes can be made identical. It may be true that this resistor is also shunted across the output, but since its value is relatively high, it consumes but little power.

The most efficient solution of the problems around the screen-grid supply has been given in the simplified circuit of Fig.9. A double choke is used for the screen-grid supply, so that each screen grid is almost at the d.c. potential of its corresponding anode, and the cathode currents of both tubes are equal.

The current flows through the windings in opposite directions, so that the magnetic fields cancel out. The core can be stacked without air gap, so that a high inductance may be acquired at small

Fig. 9. Screen-grid supply for a single-ended push-pull output stage with a double choke. The screen grids are at the same direct potential as their respective anodes. The core magnetization of the choke is neutralised by passing the d.c. of the screen grids in opposite directions through the coils. Thus a high inductance can be obtained in a small choke.



dimensions. The H.T. supply requires no centre tapping; smoothing is obtained with a double electrolytic capacitor and a resistor.

The circuit described is used in the extremely high-quality amplifier described on page 45. Its unprecedented quality of reproduction implies the use of high-impedance loudspeakers.

PHASE SPLITTER

The push-pull output stage must be driven by a phase splitter which delivers signals of opposite phase and adequate amplitude to the grids of the output tubes. These signals should be well balanced and of low distortion content. If a considerable gain can be attained in such a stage, so much the better. A high gain offers the advantage that the number of stages can be kept small, which results in little phase shift, so that stability can be maintained even with heavy negative feedback.

Some designers prefer a high gain at the cost of higher inherent distortion, and a higher negative feedback factor, so that the distortion is reduced to a reasonable value, others prefer a minimum of distortion in the phase splitter, be it at low gain.

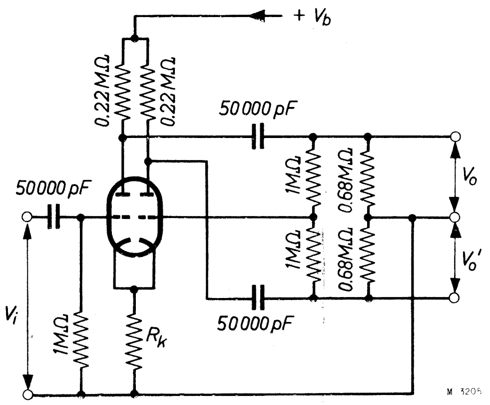


Fig.10. Phase splitter circuit with a fairly high gain and output voltage, but with rather large distortion.

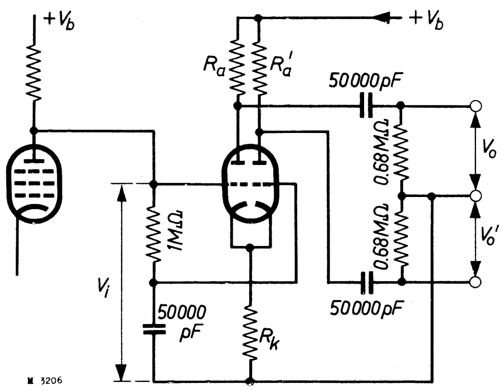


Fig.11. Cathode-coupled phase splitter. The gain is about half that of Fig.10, but the distortion is low.

A typical circuit employed by designers of the first mentioned group is given in Fig.10, whereas the circuit of Fig.11 is the "cathode-coupled phase splitter", which has very low distortion, be it at a considerable reduction in gain. The maximum obtainable output signal is also lower than in the circuit of Fig.10. Both circuits may be used with the high-mu double triode ECC 83; the operating conditions are given in Table 2.

Table 2

Phase splitter	Fig.10		Fig.11		
Anode supply voltage	250	350	250	350	V
Cathode resistor	1200	820	68 000	82 000	Ω
Anode resistors	0.22	0.22	0.1	0.15	MΩ
Output voltage	35	45	20	35	V_{rms}
Gain	58	62	25	27	
Total distortion	5.5	3.5	1.8	1.8	%

The distortion with triodes decreases linearly with the output voltage. At comparable values of 20 V_{rms} at 250 V supply and of 35 V_{rms} at 350 V, the distortion of the circuit of Fig.10 is 3.2 % respectively 2.7 %, against 1.8 % in both cases when the cathode-coupled circuit is used. The first circuit has the advantage of a higher output voltage, the second that of low distortion and (because the anode of the pre-stage tube can be coupled directly to the grid of the phase splitter) of little phase shift. Since the gain of the ECC 83 is still reasonable in the cathode-coupled circuit, it has become very popular with designers of Hi-Fi equipment.

The drive required for EL 34 type pentodes for full output is about 2 x 25 V_{rms}, whilst EL 84 type pentodes require 2 x 10 V_{rms}. In both cases the requirements are similar for triode, pentode or ultra-linear connection. This implies that, when the circuit of Fig.10 is used in an amplifier without feedback, the input voltage of the ECC 83 should be of the order of 0.4 V_{rms} for a stage preceding 2 x EL 34 and with 350 V anode supply; with the cathode-coupled phase splitter roughly 0.9 V_{rms} is required. For 2 x EL 84 these figures are 0.17 and 0.4 V_{rms} respectively at 250 V anode supply.

With 26 dB feedback, the input voltage requirements are increased by a factor of 20. This implies that the phase splitter should be preceded by a pre-stage; this is usually equipped with an EF 86 pre-amplifier pentode *). If this tube is adjusted for maximum gain, which is general practice, the total sensitivity of the complete amplifier is too high for use with a sensitive crystal pick-up. In the 10 W high-fidelity amplifier described on page 25, advantage is taken of the excessive gain by incorporating a very effective tone control circuit. On the other hand, the sensitivity is insufficient when a magnetic pick-up or microphones are used. Modern amplifiers are, therefore, mostly used with a separate pre-amplifier which contains all the controls and equalizers.

With these ideas in mind a new type of phase splitter has been developed experimentally, in which a combination of positive and negative feedback leads to an extremely high gain combined with adequate stability and reasonably low distortion. A gain of about 800 can easily be obtained with this circuit, but the attenuation at the higher frequencies in the audible range is considerable. Therefore the gain of the practical circuit given in Fig.12 has been reduced to about 220.

The anode of the left-hand section of the ECC 83 is fed from the cathode of the right-hand section. The cold end of the cathode resistor of the latter is connected to the cathode of the first, so

*) Some designers are inclined to use the ECC 83 as a pre-stage tube at high sensitivities. The ECC 83, however, is only warranted to be free of microphony when it is used with a maximum sensitivity of 50 mV for 5 W output and a loudspeaker with an efficiency of 5 % placed in the immediate vicinity of the tube. It is known that the ECC 83 has occasionally been used at a sensitivity of even 1 mV, and although some specimens gave satisfactory results, this should not be expected from the majority of the tubes. In mass-production, especially of tape recorders and equipment with built-in loudspeaker, this practice cannot be advised, and the EF 86, which is a noise-free amplifier pentode, designed for this purpose, is the recommended type.

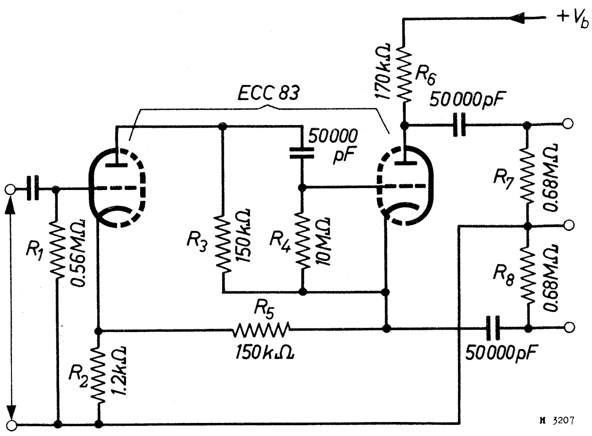


Fig.12. High-gain phase splitter with positive and negative feedback. A gain of 220 can be obtained with good stability and frequency response. At a higher gain the attenuation of the high frequencies becomes excessive.

that the positive feedback of the right-hand section exceeds the negative feedback of the left-hand section. The positive feedback via R_2 might easily lead to instability, if the lower end of R_3 were not connected to the cathode of the right-hand tube section. Since this cathode voltage is in phase with the anode voltage of the left-hand section, additional negative feedback is applied.

This phase splitter, followed by a conventional push-pull output stage and with 26 to 30 dB negative feedback across the complete amplifier, renders the use of a pre-stage superfluous, provided a separate pre-amplifier containing the various controls and equalizers is used.

PRE-AMPLIFIER AND PHASE SPLITTER FOR SINGLE-ENDED PUSH-PULL

One of the main advantages of the single-ended push-pull output stage is the ease with which feedback can be applied, because the phase shift has been reduced to a minimum by the omission of the output transformer. This offers special facilities as to the design of the pre-stage and the phase splitter, in which the positive feedback may be adjusted on the verge of oscillating, whilst the negative feedback across the complete amplifier provides the required stability.

Fig.13 shows the principle of operation of this circuit. The left-hand section of the double triode operates as a pre-amplifier. Its signal voltage is applied both to the control grid of the right-hand section of the ECC 83 and to that of the output tube *II*. The first operates as a phase inverter, and its signal voltage is fed to the control grid of output tube *I*.

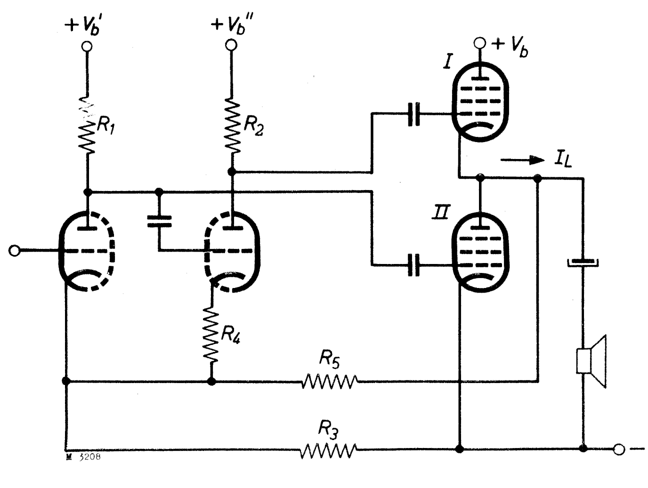


Fig. 13. Simplified circuit diagram of a positive feedback phase-splitter in combination with a single-ended push-pull output stage. Stability is obtained by negative feedback from the output to the input via R_5.

It should be recognized that the driving voltages for the output tubes are entirely different. Output tube *II* is driven in the normal way. Output tube *I*, however, must be considered as a cathode follower with tube *II* acting as a varying cathode resistor, its alternating cathode potential being equal to the output voltage. This implies that the driving voltage should be equal to the sum of the output voltage and the driving voltage that would be required to obtain an output current equal to that of tube *II*, if the cathode were at zero potential, hence a driving voltage of about 105 V_{rms} for full output. A driving voltage of that value, however, cannot be delivered by an ECC 83 under normal operating conditions, so that special measures must be taken.

The output voltage requirements imposed on the phase inverter can be satisfied in a fairly simple way by feeding its anode from a d.c. source on which the output voltage of the amplifier is superimposed. Such a voltage is available at the screen grid of output tube *I* at which almost the full direct supply voltage and simultaneously the full output voltage are present. Because output tube *I* is a cathode follower, its input and output voltages are in phase, so that, when the anode supply ($+V_b''$) is taken from the screen grid mentioned, the phase inverter need only deliver the normal driving voltage for an output tube plus the voltage loss across the resistor R_2. Such an output voltage can easily be obtained from one section of an ECC 83.

The anode supply of the left-hand section of the ECC 83 ($+V_b'$) may be taken from the H.T. supply via a normal *RC*-decoupling filter.

The gain of the pre-stage and the phase inverter can be increased appreciably by applying positive feedback. This is achieved by means of the resistor R_3 in the common cathode circuit of the triode sections, whilst the required additional negative feedback is applied from the output to the input via R_5.

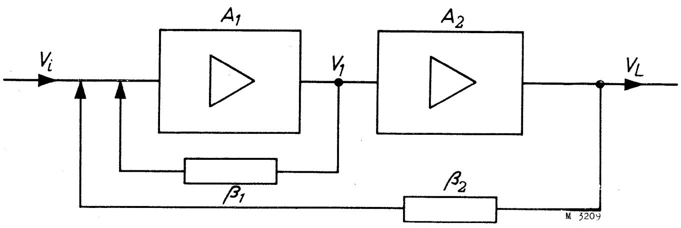


Fig.14. Block diagram of two amplifiers, A_1 and A_2, connected in cascade. The output of A_2 is fed back to the input of A_1 via β_2.

The effect of the combined positive and negative feedback may be explained with reference to the block diagram shown in Fig.14. A_1 and A_2 represent two amplifiers connected in cascade, for example a pre-stage and an output stage. The voltage gain of these stages will also be denoted by A_1 and A_2. The part $\beta_1 V_1$ of the output signal V_1 supplied by A_1 is fed back to the input of A_1, and so is the part $\beta_2 V_L$ of the output voltage V_L supplied by A_2. Denoting the input signal by V_i then:

$$V_1 = A_1\ (V_i + \beta_1 V_1 + \beta_2 V_L)$$

and

$$V_L = A_2 V_1.$$

The total voltage gain is therefore:

$$A = \frac{V_L}{V_i} = \frac{A_1A_2}{1 - A_1\beta_1 - A_1A_2\beta_2} = \frac{A_1A_2}{N},$$

where

$$N = 1 - A_1\beta_1 - A_1A_2\beta_2.$$

In an analogous manner it can be shown that if the distortion of the individual amplifiers is denoted by d_1 and d_2 respectively, the total distortion is:

$$d = \frac{1}{N} \cdot d_1 + \frac{1 - A_1\beta_1}{N} \cdot d_2 + \frac{1 - A_1\beta_1}{N} \cdot d_1d_2.$$

This expression reveals that something peculiar happens when $A_1\beta_1$ is made equal to unity, for in that case d is reduced to d_1/N, in other words the distortion of the amplifier A_2 no longer contributes to the total distortion.

The resulting distortion d_1/N is determined by the normally low value of d_1 of the first amplifier and by the quantity N which, in the case of $A_1\beta_1 = 1$, becomes equal to $-A_1A_2\beta_2$. Expressed in absolute values, the product $A_1A_2\beta_2$ may be much greater than unity, and the total distortion d will then become even much smaller than d_1. The absolute value of the total gain A now becomes equal to $1/\beta_2$, thus being independent of A_2.

The condition $A_1\beta_1 = 1$ can be satisfied for a wide frequency range, provided A_1 and β_1 are real in this region, that is if the amplifier A_1 and the feedback link β_1 contain no phase-shifting elements. It is quite possible to realize this very approximately in a pre-amplifier stage.

To satisfy the condition $A_1\beta_1 = 1$, the positive feedback of the amplifier A_1 must be such that it is on the verge of oscillating. This does not imply, however, that the combination $A_1\beta_1 - \beta A_2\beta_2$ is unstable. Provided the second amplifier and its feedback link contain no elements that introduce unfavourable phase shifts, the entire circuit can be kept stable by rendering the feedback β_2 negative. The absence of an output transformer with its unavoidable phase shifts offers the possibility of designing practical circuits based on this principle. In this way an amplifier can be obtained with remarkably low distortion and relatively high sensitivity, the loss of gain caused by the heavy negative feedback being partially made good by the positive feedback. Such a design is given on page 45.

POWER SUPPLY

The power supply of high-quality amplifiers offers no great problems. Although the first Hi-Fi amplifiers where provided with a separate supply unit equipped with ample smoothing filters containing a choke, it became apparent in the course of time that the supply transformer could be mounted on the same chassis as the main amplifier, especially when a separate pre-amplifier for low input signals was used. This arrangement requires, of course, a judicious circuit lay-out, but is nevertheless relatively easy to design.

The smoothing choke may be omitted when the anodes of the power tubes are fed from the reservoir capacitor and the screen grids from a smoothing capacitor via a series resistor. The phase splitter and pre-stage tubes are usually fed via adequate smoothing filters consisting of a resistor and an electrolytic capacitor of 8 to 25 μF. The hum level can thus easily be kept at -70 to -90 dB with respect to full output.

The rectifier tubes are preferably indirectly heated and their heating-up time should be somewhat longer than that of the power tubes, so that the voltage rating of the smoothing capacitors need not exceed the operating voltage appreciably.

For currents up to 150 mA and a transformer voltage up to 2 x 350 V, the EZ 81 is recommended; the GZ 34 is the preferred type for currents ranging from 150 to 250 mA, and a transformer voltage up to 2 x 450 V.

Care should be taken that the transformer resistance (R_t) is not below the minimum value quoted in the Limiting Values of the tube data. This transformer resistance may be determined as follows:

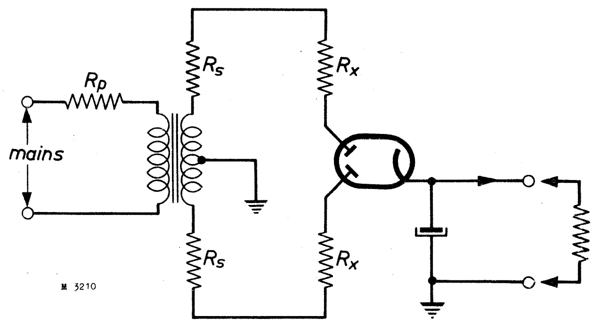


Fig. 15. Circuit of a two-phase rectifier circuit. The ohmic resistance of the primary winding of the mains transformer is denoted by R_p, that of each half of the secondary by R_s. R_x is the additional resistance to be connected in series with each anode of the rectifying tube, to obtain the lowest permissible transformer resistance.

The ohmic resistance (R_p) of the primary and the ohmic resistance of one of the secondaries (R_s) are measured. The transformer resistance is $R_t = R_s + n^2R_p$ in which n is the winding ratio (i.e. the voltage ratio) between one secondary and the primary (see Fig.15). If the value thus found is below R_{tmin} given in the tube ratings, a resistor with a value of $R_x = R_{tmin} - R_t$ must be connected in series which each rectifier anode. Each resistor carries half the direct current and half the ripple current, so that its power rating should be in accordance with a current of 1.2 times the total direct current delivered by the rectifier. Alternatively, the two resistors may be replaced by a single one of the same value, connected in series with the cathode of the rectifier tube, the required power rating then being 2.4 times the total direct current.

The reservoir capacitor in the H.T. unit should be able to pass the ripple current which can be estimated to be 1.4 times the total direct current delivered by the supply unit. In single-phase rectifiers (which are rarely used in Hi-Fi equipment), the ripple current is 2.4 to 2.7 times the direct current.

Chapter II

Power Amplifiers

SIMPLE 3 W GRAMOPHONE AMPLIFIER

This circuit contains a small number of tubes and components and has been designed for those who wish to make a very simple amplifier with a reasonably high quality. It can be used with all types of crystal pick-up, the sensitivity being sufficient to insert a correction network between the pick-up and the input terminals. The output is 3 W at a total harmonic distortion of 1 %.

SPECIFICATION

Tubes:	EF 86 pre-amplifier, EL 84 power tube, EZ 80 double anode rectifier.
Output power:	3 W, at 1.0 % total harmonic distortion.
Frequency response:	Flat within ± 1 dB (relative to the level at 1 kc/s) from 20 c/s to 40 000 c/s.
Tone control:	Maximum treble cut approximately 20 dB at 10 kc/s; maximum bass boost approximately 15 dB at 70 c/s.
Sensitivity:	100 mV for 3 W output.
Hum and noise levels:	-70 dB at full output.

CIRCUIT DESCRIPTION

The amplifier which is to be operated from the a.c. mains uses three tubes, viz. EF 86 as pre-amplifier, EL 84 as power tube and an EZ 80 rectifier. The circuit (Fig. 16) includes three controls: volume control by R_1, a square law potentiometer of 0.5 MΩ; treble cut is obtained with R_2, a 0.5 MΩ linear potentiometer, and the bass boost inserted in the feedback circuit and controlled by R_8, a square law 50 kΩ potentiometer.

The EF 86 is used under "starvation" conditions. The tube voltages and currents are very much smaller than under normal operating conditions owing to the high anode load resistance (1 MΩ), and the low screen-grid supply voltage which is taken from the cathode of the power tube. Grid current biasing is obtained by means of a high grid leak (10 MΩ). The gain is very high, viz. of the order of two or three times above the level of conventional operating conditions. The low operating voltages allow direct coupling of the pre-amplifier anode to the control grid of the output tube. To maintain correct grid biasing of the latter, the cathode resistor should have a much higher value than usual. A wire-wound type of 560 Ω with 5 % tolerance and 3 W power rating should be used.

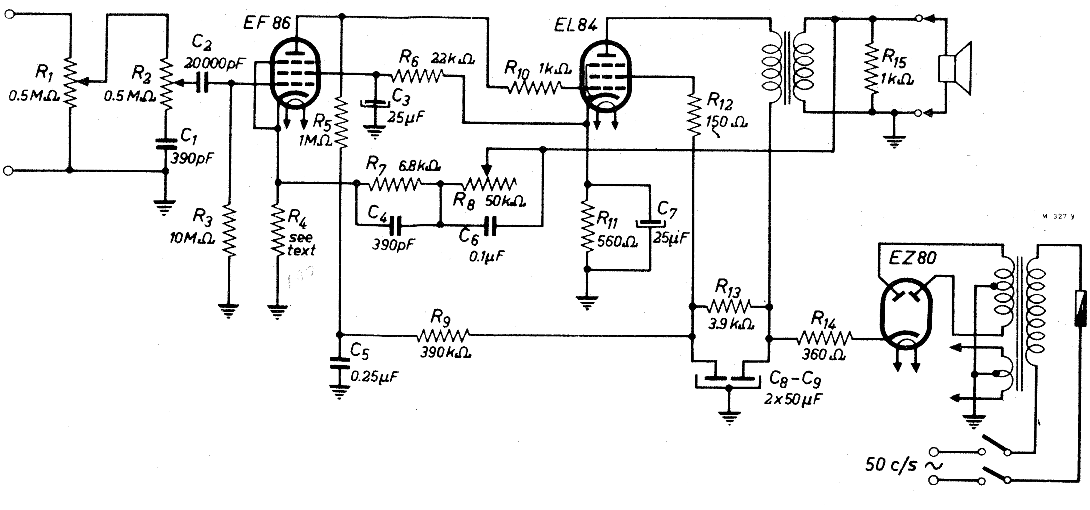


Fig.16. Circuit diagram of the 3 W amplifier.

R_1 = 0.5 MΩ, log.
R_2 = 0.5 MΩ, linear
R_3 = 10 MΩ ± 20 %, ¼ W
R_4 = see text, ¼ W
R_5 = 1 MΩ ± 10 %, high stab. ¼ W
R_6 = 22 kΩ ± 10 %, ¼ W
R_7 = 6.8 kΩ ± 10 %, ¼ W
R_8 = 50 kΩ, log.
R_9 = 0.39 MΩ ± 10 %, ¼ W
R_{10} = 1 kΩ ± 20 %, ¼ W
R_{11} = 560 Ω ± 5 %, 3 W, wire wound
R_{12} = 150 Ω ± 20 %, ¼ W
R_{13} = 3.9 kΩ ± 10 %, ½ W
R_{14} = see text, 2 W
R_{15} = ¼ kΩ ± 20 %, ¼ W
C_1 = 390 pF ± 10 %
C_2 = 22 000 pF
C_3 = 25 µF, 50 V wkg.
C_4 = 390 pF ± 10 %
C_5 = 0.25 µF, 350 V wkg.
C_6 = 0.1 µF ± 10 %, 150 V wkg.
C_7 = 25 µF, 50 V wkg.
C_8 = 2 x 50 µF, 350 V
C_9 = 2 x 50 µF, 350 V

The screen-grid voltage of the EF 86 is taken from the cathode of the output tube via a filter R_6C_3 of 22 kΩ and 25 µF respectively. This filter and the high cathode resistor of the output tube provide a heavy d.c. feedback so that the operating conditions are stabilized against mains voltage variations and the influence of variations in tube characteristics within the normal production spreads.

Because of the inherently high distortion associated with single-tube output stages, appreciable negative feedback around the output stage and transformer is necessary for producing an output of acceptable quality. Therefore a.c. negative feedback has been applied from the output transformer secondary to the unbypassed cathode resistor of the EF 86. The value of this resistor depends upon the output transformer and the loudspeaker used. Its value is 82 Ω for matching to 15 Ω voice coil impedance, 100 Ω for 7 Ω, 120 Ω for 5 Ω and 150 Ω when a 3.75 Ω loudspeaker is used with an appropriate output transformer. The primary of the transformer should match the speaker to 5 kΩ load impedance of the output tube. Fig. 17 shows the distortion as a function of the output power.

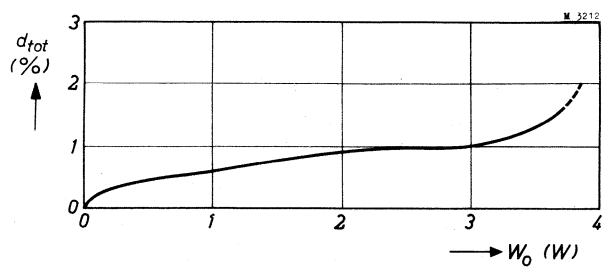


Fig.17. Harmonic distortion d_{tot} as a function of the output power W_o measured on the 3 W amplifier at 400 c/s.

The gain of the amplifier is not high enough to permit the use of complete bass and treble control. Therefore a bass boost control is inserted in the negative feedback circuit, and a treble cut is connected between the volume control and the control grid of the EF 86. With both controls at a minimum, the response curve is flat within ± 1 dB between 20 c/s and 40 kc/s, the 3 dB points being at 15 c/s and 50 kc/s. Maximum bass boost is 15 dB at 70 c/s and maximum treble cut is 20 dB at 10 kc/s, so that proportionate reproduction can be obtained of all types of record, and needle scratch of worn standard records can sufficiently be suppressed (see Fig. 18). A 1000 Ω resistor is connected across the secondary of the output transformer to maintain stability when the loudspeaker is disconnected or when the capacitive load is increased by use of a long loudspeaker cord.

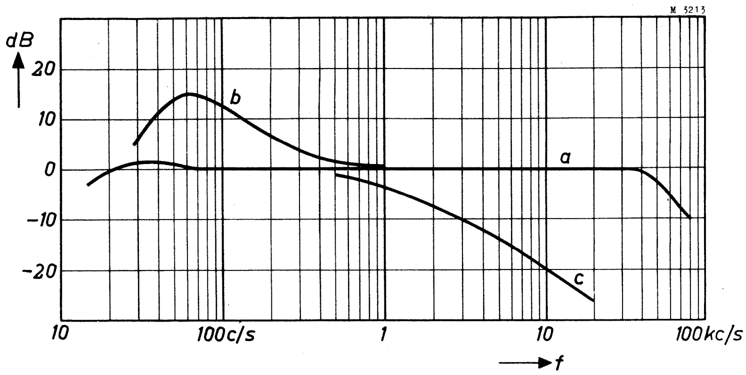


Fig. 18. Maximum bass boost (b) and maximum treble cut (c) of the 3 W amplifier. With both controls at minimum, the frequency response curve is virtually flat between 20 c/s and 30 kc/s (curve a).

The supply is conventional. The heater of the EZ 80 full-wave rectifier may be connected to the same transformer winding as the amplifier tubes, as indicated in the circuit diagram. A separate winding, however, is to be preferred. If one winding is used it should have a 2 A rating, if separate windings are used either should be for 1 A. The H.T. winding has been dimensioned for 2 x 300 V 60 mA, but a 2 x 275 V winding and R_{14} reduced to 100 Ω may also be used. The primary may be tapped for matching to 110 V, 127 V, 150 V, 220 V mains.

The anode of the output tube is fed from the storage capacitor, the screen grid from the first smoothing capacitor, the smoothing resistor being 3.9 kΩ, ½ W. The anode of the pre-amplifier tube is fed via a 390 kΩ dropping resistor, decoupled by a 0.25 μF paper capacitor for 350 V working voltage.

VOLTAGES AND CURRENTS MEASURED

(voltages measured against chassis)

Voltage across C_9	310 V
Voltage across C_8	290 V
EL 84	
Anode voltage	290 V
Anode current	43 mA
Screen-grid voltage	290 V
Screen-grid current	5 mA
Cathode voltage	27 V
EF 86	
Anode voltage	20 V
Anode current	195 μA
Screen-grid voltage	27 V

10 W HIGH-FIDELITY AMPLIFIER

WITH TWO EL84 PENTODES IN PUSH-PULL

The original circuit of this amplifier was published some years ago. It was reprinted and discussed in most of the leading radio and audio periodicals because it was the first design to make real Hi-Fi reproduction possible at a moderate price.

It will be clear that many amplifiers were built according to this attractive design, and thus much experience was gained, which was usually, but not in all cases, favourable. There were some complaints of instability, either at high frequencies or at low, and occasionally at both. Disregarding the cases of bad and careless mounting, the instability was always caused by the output transformer, which differed in some respect from the original design. Some transformers showed a higher primary inductance, others a higher stray inductance and a lower primary inductance than originally specified.

Successful attempts were made to reduce the influence of a relatively wide spread in the characteristics of the output transformer on the stability. The influence on the quality of reproduction is almost imperceptible, provided the amplifier is adequately mounted.

Two modified circuits, which have become quite popular and have been realized on a large scale without showing any instability, are discussed below.

OUTPUT TRANSFORMER

The data of the original output transformer are :

Transformer ratio		34.5 : 1
Primary inductance (50 c/s, 10 V)	without d.c.	40 H
	at 5 mA	28 H
Leakage inductance		22 mH
Resistance of primary		2 x 240 Ω
Resistance secondary		0.4 Ω
Efficiency at 1000 c/s		85 %

The details of this transformer are as follows:

Laminations (see Fig. 19)	: normal dynamo sheet 0.5 mm
Core	: shell type
Overall dimensions	: 84 x 70 mm
Width of core	: 28 mm
Stacking	: without air gap
Height of stacking	: 28 mm
Core cross-section	: 7.86 mm^2.

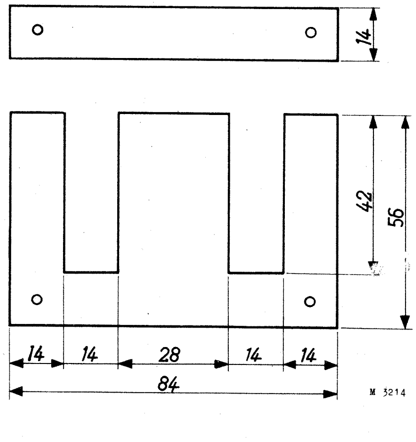


Fig.19. Dimensions in mm of the output transformer core laminations of the 10 W push-pull amplifier.

phase splitter. The power stage is slightly modified, which goes hand in hand with the omission of the choke in the power supply. It proved possible to feed the anodes of the power tubes directly from the reservoir capacitor, without introducing noticeable hum. The smoothing choke is replaced by a resistor of 1.5 kΩ. The screen grids of the power tubes are fed from the smoothing capacitor; no common series resistor is used in the screen-grid circuit, because in this circuit the screen-grid voltage would become too low and the output power would be reduced.

In this circuit use is made of the rectifier tube EZ 81, which was not yet available when the first version of the circuit was published. The d.c. output is sufficient to feed an additional AM/FM tuner from the power supply, provided the mains transformer is large enough.

SENSITIVITY

The sensitivity of the amplifier is about 400 mV for 10 W output, so that it may be used with crystal pick-ups, and with a broadcast AM/FM tuner. For reproduction of tape recordings an equalizer is required.

FREQUENCY RESPONSE AND TONE CONTROL

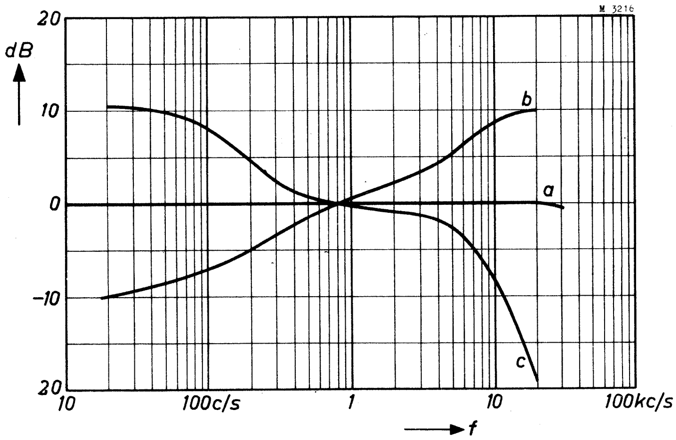


Fig. 21. Frequency response and tone-control characteristics of the 10 W amplifier according to the first modified circuit. (a) With both controls in the middle position, (b) with maximum treble lift and maximum bass cut, (c) with maximum bass lift and maximum treble cut.

Fig. 21 shows the frequency response curve with tone controls at maximum and at minimum. The amplifier has a virtually flat frequency response curve from about 10 c/s to at least 30 kc/s with both controls in middle position.

DISTORTION

The harmonic distortion of the amplifier is less

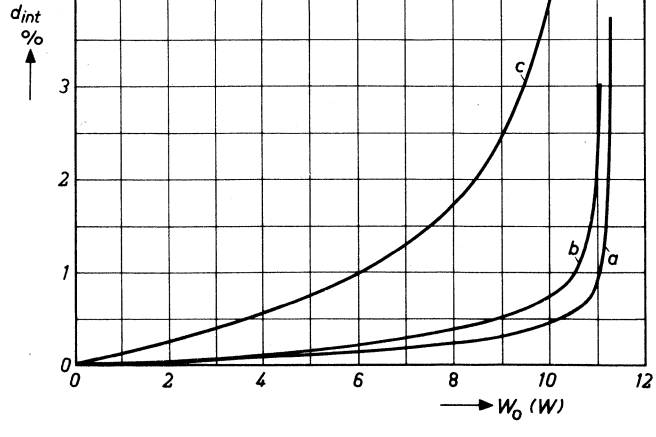


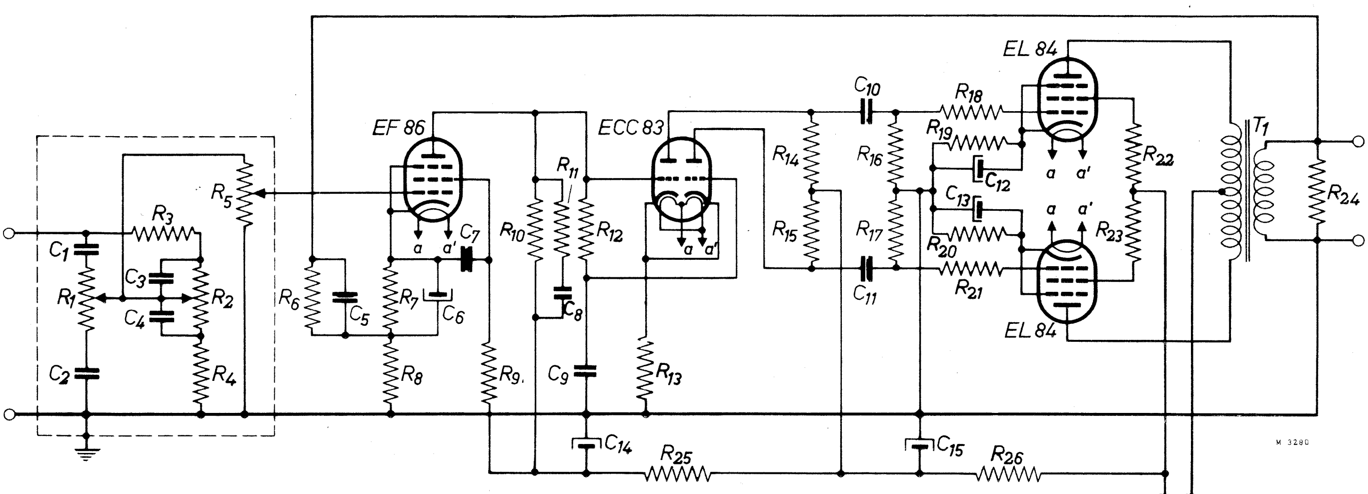
Fig. 22. Harmonic distortion (d_{tot}) measured (a) at 400 c/s and (b) at 40 c/s, and intermodulation distortion (c) measured at 40 c/s and 10 kc/s in a 4 : 1 amplitude ratio on the 10 W amplifier according to the first modified circuit.

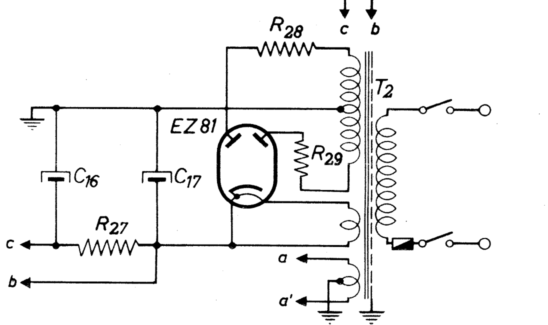
than 0.5 % at 400 c/s and 10 W output; at 40 c/s the distortion is about 0.8 %. Intermodulation distortion was measured with signals of 40 and 10 000 c/s in a 4 : 1 ratio. The equivalent output power appeared to be 8.2 W at 2 % intermodulation distortion. Both the harmonic distortion curves and the intermodulation curve as functions of the output power are plotted in Fig. 22.

SECOND MODIFIED CIRCUIT

CIRCUIT DESCRIPTION

Fig. 23 shows the circuit diagram which is recommended when the primary inductance of the output transformer is much larger than the value given in the original design. The circuit modifications





R_1 = 2 MΩ, log.
R_2 = 2 MΩ, log.
R_3 = 1.5 MΩ ± 10 %, ¼ W
R_4 = 150 kΩ ± 10 %, ¼ W
R_5 = 1 MΩ, log.
R_6 = high stab. resist. ± 5 %, ¼ W
3.75 Ω load: 15 kΩ
7 Ω load: 27 kΩ
15 Ω load: 33 kΩ
R_7 = 1.8 kΩ ± 10 %, ¼ W
R_8 = 100 Ω ± 5 %, high stab. ¼ W
R_9 = 470 kΩ ± 10 %, high stab. ¼ W
R_{10} = 100 kΩ ± 10 %, high stab. ¼ W
R_{11} = 10 kΩ ± 10 %, ¼ W
R_{12} = 1 MΩ ± 10 %, ¼ W
R_{13} = 68 kΩ ± 10 %, 1 W
R_{14} = 100 kΩ ± 10 %, ½ W
R_{15} = 100 kΩ ± 10 %, ½ W
R_{16} = 0.82 MΩ ± 10 %, ¼ W
R_{17} = 0.82 MΩ ± 10 %, ¼ W
R_{18} = 4.7 kΩ ± 10 %, ¼ W
R_{19} = 270 Ω ± 5 %, 2 W (norm.loading)
= 437 Ω ± 5 %, 2 W (low loading)
R_{20} = 270 Ω ± 5 %, 2 W (norm.loading)
= 437 Ω ± 5 %, 2 W (low loading)

Fig.23. Circuit diagram of the 10 W amplifier (second modified circuit).

R_{21} = 4.7 kΩ ± 20 %, ¼ W
R_{22} = 47 Ω ± 20 %, ¼ W
R_{23} = 47 Ω ± 20 %, ¼ W
R_{24} = 1 kΩ ± 20 %, ½ W
R_{25} = 100 kΩ ± 10 %, ¼ W
R_{26} = 33 kΩ ± 10 %, ½ W
R_{27} = 1.2 kΩ ± 10 %, 1 W
R_{28} = depending on R_{tr}
R_{29} = depending on R_{tr}
C_1 = 33 pF ± 10 %
C_2 = 680 pF ± 10 %
C_3 = 270 pF ± 10 %
C_4 = 3300 pF ± 10 %

C_5 = 270 pF ± 5 %; load 3.75 Ω
150 pF ± 5 %; load 7 Ω
120 pF ± 5 %; load 15 Ω
C_6 = 100 μF, 12.5 V wkg.
C_7 = 47 000 pF, 350 V d.c. wkg.
C_8 = 330 pF ± 10 %
C_9 = 0.1 μF, 350 V d.c. wkg.
C_{10} = 0.1 μF, 350 V d.c. wkg.
C_{11} = 0.1 μF, 350 V d.c. wkg.
C_{12} = 50 μF, 25 V wkg.
C_{13} = 50 μF, 25 V wkg.
C_{14}, C_{15} } 2 x 8 μF, 350 V
C_{16}, C_{17} } 2 x 50 μF, 350 V wkg.

resemble those previously discussed. Moreover, the gain of the first stage is reduced to obtain a loop gain not exceeding 26 dB at 1 kc/s, thus increasing the overall margin of stability.

The by-pass capacitors of the separate cathode resistors of the output stage are decreased to 50 μF to reduce the gain and improve the stability at the low-frequency end. The time constant of the screen-grid decoupling of the first stage is also decreased for the same purpose.

High-frequency stability, particularly with capacitive loads, is improved by modifying the time constants C_8-R_{11} and C_8-R_{10} in the phase-shifting network in the anode circuit of the first stage. Because the loop gain is reduced, the value of C_5 in the feedback loop must also be modified. Although the value given in the List of Components (time constant $C_5 \cdot R_6 = 4 \times 10^{-6}$ sec) generally proves satisfactory, it is possible that a slightly different value may be necessary for optimum high-frequency performance with any particular make of output transformer.

It should be possible to increase the feedback by 6 dB (by reducing the value of R_6) before instability occurs under normal operating conditions, which is a good test for checking the stability of the mounted amplifier.

Tests have been made with a number of output transformers having widely differing primary inductances, ranging from 40 H to > 100 H at 10 V, 50 c/s and with a leakage inductance not higher than 30 mH. It proved that stability is maintained even under open-circuit conditions and with capacitive loading up to at least 0.05 μF, which implies that no difficulties need be feared even when very long loudspeaker cords are used.

SENSITIVITY

The sensitivity of the modified design is 40 mV for 10 W output when the tone control circuits are omitted. The attenuation of these circuits is about 20 dB, so that the sensitivity with tone control is 400 mV for 10 W output. The sensitivity as a function of the output power is plotted in Fig. 24. The amplifier with tone

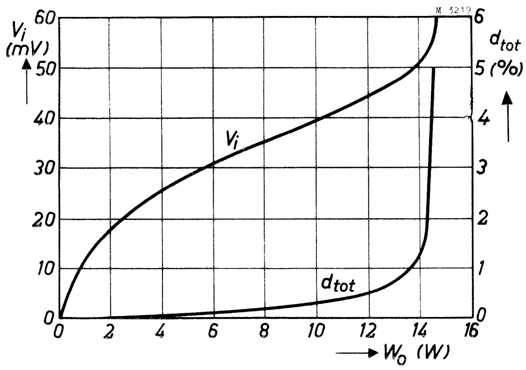


Fig. 24. Total distortion d_{tot} and input voltage V_i as functions of the output power W_o, measured at the 10 W amplifier according to the second modified circuit, at a frequency of 400 c/s.

control is therefore suitable for use with crystal pick-ups and may also be used with AM/FM broadcast tuners. For reproducing tape recordings an equalizer should be used. When magneto-dynamic pick-ups are used, an additional pre-amplifier is essential in which also the tone control circuits can be mounted. A choice can be made of one of the circuits described in Chapter III.

FREQUENCY RESPONSE AND PHASE SHIFT

Typical characteristics for the frequency response, the loop gain and the phase shift both in the amplifier and in the feedback loop are given in Fig. 25. These curves apply to an anode load (R_{a-a}) of 8 kΩ. The curves are also shown below 10 c/s to give a good idea of the stability.

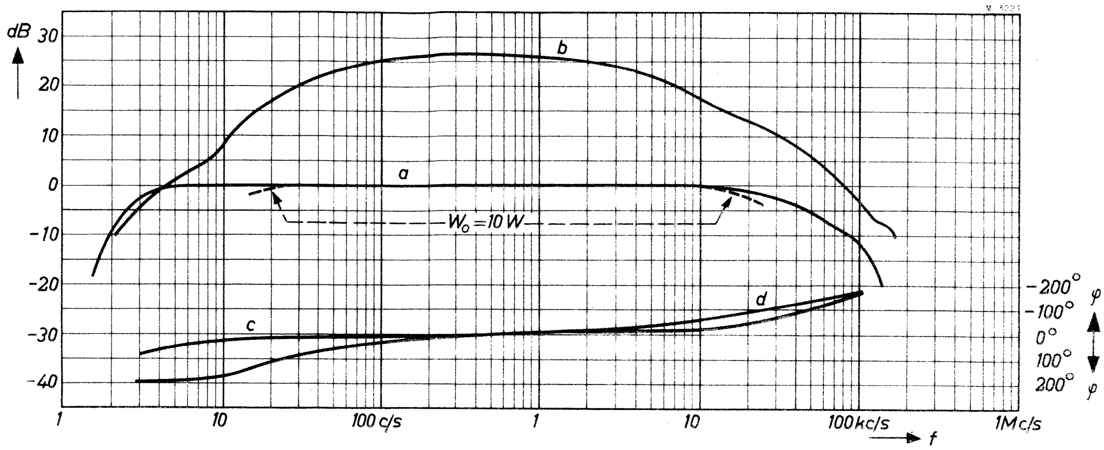


Fig. 25. Performance of the 10 W amplifier. The frequency response and power response (*a*), the loop gain (*b*), and the phase shift of the amplifier (*c*) and of feedback loop (*d*) plotted against the frequency.

DISTORTION

Fig. 24 shows the total harmonic distortion as a function of the output power, together with an input voltage curve. The intermodulation distortion is plotted in Fig. 26 for frequencies of 40 c/s and 10 kc/s and for frequencies of 70 c/s and 7 kc/s. In both cases the amplitude of the lower frequency is four times that of the higher frequency.

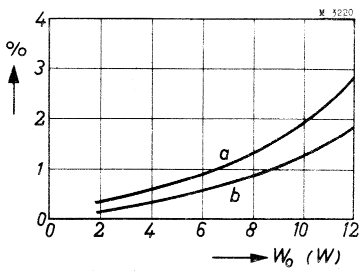


Fig. 26. Intermodulation as a function of the equivalent power output (W_o) measured with signals of 40 c/s and 10 kc/s (*a*), and of 70 c/s and 7 kc/s (*b*), both at a 4 : 1 ratio.

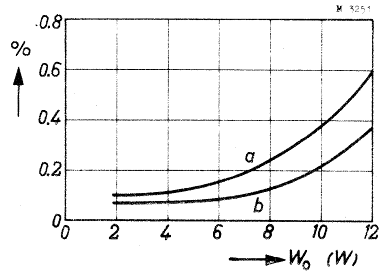


Fig. 27. Beat note distortion as a function of the equivalent power output (W_o), (*a*) measured with signals of equal amplitude of 14 and 15 kc/s, and (*b*) with signals of 9 and 10 kc/s.

The magnitude of the beat note distortion was measured with frequencies of 9 kc/s and 10 kc/s, the two input signals having the same amplitude, and also with 14 kc/s and 15 kc/s. The percentage of the 1 kc/s difference component in the output voltage is plotted in Fig. 27.

LOW LOADING

In Chapter I low-loading operating conditions were dealt with briefly. In normal loading conditions the output tubes are adjusted according to the operating conditions for class AB given in the tube data, viz. either separate 270 Ω cathode resistors are used or a common resistor of 130 Ω, the anode-to-anode load resistance being 8000 Ω, and the anode current 2 x 36 mA.

At the low-loading adjustment the anode-to-anode load is reduced to 6000 Ω, and the quiescent anode current is only 2 x 24 mA. Therefore cathode resistors of 437 Ω (390 Ω + 47 Ω) should be used in each cathode connection. With speech and music inputs the operation of the output stage approximates very closely to fixed bias conditions although it is connected for cathode bias (see Fig. 3).

The advantages may be summarized as follows. The maximum rated output of an amplifier is required only for a small part of the time, the *average* power being comparatively low. A high maximum power output has to be available to cater for the widely differing sound levels during the reproduction, especially of music. The low-loading adjustment provides reduced distortion at just these peak levels, although improvement may be hard to detect until the ear is accustomed to high-quality reproduction, because it affects such short intervals of time.

The H.T. consumption is smaller when the output stage is adjusted for low loading. The standing dissipation is reduced from 11 W to 7.5 W at each anode. The output tubes are thus being run well below their maximum anode dissipation of 12 W, and their expectation of life will correspondingly be increased. Also, with low loading there will be less ripple on the H.T. line; the mains transformer can be given a lower rating as a measure of economy.

Larger peak currents are produced in the output stage under low-loading conditions than with normal class AB operation. These peak currents are of short duration with speech or music and are supplied by the reservoir capacitor which has to be of high value (50 μF).

A disadvantage (which is in fact only theoretical) is that the amplifier may not be tested with a sine wave up to the full output power. Low-level sine-wave drive may be used to measure the frequency response provided the output does not exceed 1 to 1.5 W, above which point excessive distortion will occur. Normal square-wave testing can be undertaken, but the input should not exceed a level similar to that used for the low-level sine wave.

ULTRA-LINEAR OPERATION

The amplifier has also been tested with ultra-linear operation. The output transformer was tapped on 43 % of the number of turns counted from the centre tapping. The capacitor in the feedback loop (C_5) was reduced by about 20 %.

A comparison of performance is given below for (A) normal push-pull AB output stage, (B) ultra-linear push-pull output stage, and (C) low-loading output stage (speech and music only).

Type of circuit	A	B	C	
Rated power output	10	10	10 *)	W
Overload point	14	11	14 *)	W
Sensitivity across volume control	40	40	-	mW
Harmonic distortion (10 W) 400 c/s	0.3	0.1	-	%
Intermodulation distortion (10 W) 40 c/s - 10 kc/s, ratio 4 : 1	2	1	-	%
Beat note distortion (10 W)				
9 kc/s and 10 kc/s	0.25	0.25	-	%
14 kc/s and 15 kc/s	0.4	0.35	-	%
Loop gain at 1000 c/s	26	20.5	-	dB

*) Equivalent sine-wave output power.

20 W HIGH-FIDELITY AMPLIFIER

WITH 2 x 2 PENTODES EL84 IN PUSH-PULL

In many cases a higher output than 10 W is required, so that the amplifier may be equipped with two EL 34 power pentodes in push-pull. Another approach is to connect two EL 84 pentodes in parallel for each half of the push-pull circuit. Such a circuit has the disadvantage of more complicated assembly, but there are several advantages which may be of influence on the choice of circuit. They may be summarized as follows: (1) The output stage operates at a lower anode voltage so that the mains transformer and the smoothing capacitors are less expensive. (2) The driving voltage is the same as with the 10 W amplifier, so that only a relatively low grid drive is required for full output, which keeps the distortion low. (3) The anode-to-anode load resistance is substantially lower than with two EL 34 tubes so that it is easier to keep the leakage inductance of the output transformer low; a high factor of negative feedback can therefore be applied without occurrence of instability.

SUMMARY OF PERFORMANCE

Rated power output: 20 W
Peak power output: 35 W
Frequency response: flat within 1 dB from 10 c/s to 100 kc/s
Power response: constant within 0.5 dB from 50 c/s to 50 kc/s
Harmonic distortion: (400 c/s) 0.3 % at rated output
Sensitivity (at rated output): 0.5 V_{rms}

CIRCUIT DESCRIPTION

The circuit diagram is given in Fig. 28. The amplifier is equipped with the following tubes, viz. one ECC 83 double triode as phase inverter and pre-stage tube in a positive feedback circuit, four EL 84 power pentodes in a push-pull circuit and one GZ 34 double-anode rectifier tube.

The output tubes are connected in parallel in pairs so that the mutual conductance is doubled and the load impedance is halved; the power output is also doubled compared with the conventional 10 W push-pull circuit.

The output tubes are adjusted with fixed bias. It is true that if separate cathode resistors with bypass capacitors were used, the differences in the cathode currents of the individual power tubes within the normal production spread would be substantially equalized. The fixed-bias operation with pre-set voltage control (R_{12}) and balance adjustment (R_{13}), however, makes the circuit very interesting from the point of view of output stage setting. The versatility of the output stage is such that normal and low-loading class AB and even class B setting can be chosen at will by adjustment of the grid bias. The great number of tappings on the output transformer secondary enable matching of the load to the various optimum load conditions of the output stage.

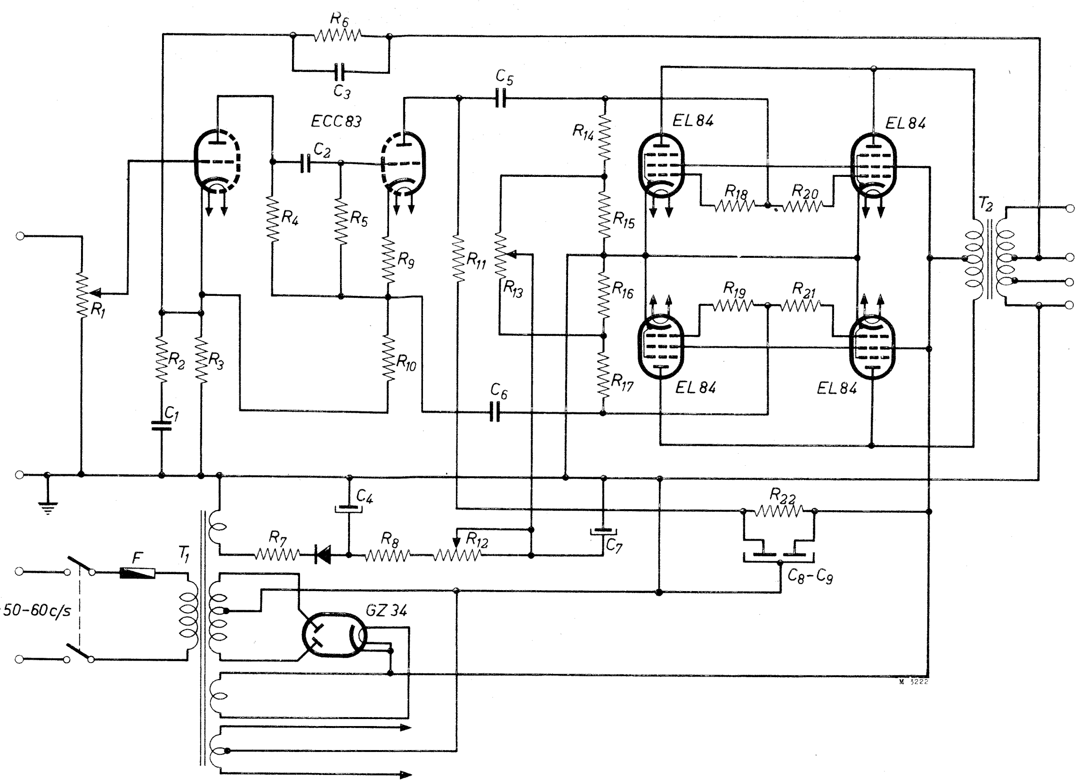


Fig.28. Circuit diagram of the 20 W amplifier with 2 x 2 EL 84 power pentodes.

R_1 = 0.5 MΩ, lin.
R_2 = 3.9 kΩ ± 5 %, high stab. ¼ W
R_3 = 1.2 kΩ ± 5 %, high stab. ½ W
R_4 = 150 kΩ ± 10 %, ½ W
R_5 = 1 MΩ ± 20 %, ¼ W
R_6 = 27 kΩ ± 5 %, high stab. ¼ W
R_7 = 2.7 kΩ ± 20 %, ¼ W
R_8 = 27 kΩ ± 20 %, ¼ W
R_9 = 3.3 kΩ ± 5 %, ½ W
R_{10} = 150 kΩ ± 5 %, high stab. ½ W
R_{11} = 160 kΩ ± 5 %, high stab. ½ W
R_{12} = 25 kΩ, lin.
R_{13} = 50 kΩ, lin.
R_{14} = 330 kΩ ± 10 %, ¼ W
R_{15} = 56 kΩ ± 10 %, ¼ W
R_{16} = 56 kΩ ± 10 %, ¼ W
R_{17} = 330 kΩ ± 10 %, ¼ W
R_{18} = 1 kΩ ± 20 %, ¼ W
R_{19} = 1 kΩ ± 20 %, ¼ W
R_{20} = 1 kΩ ± 20 %, ¼ W
R_{21} = 1 kΩ ± 20 %, ¼ W
R_{22} = 27 kΩ ± 20 %, ½ W

C_1 =2700 pF ± 5 %
C_2 = 0.1 μF ± 20 %, 350 V wkg.
C_3 = 56 pF ± 5 %
C_4 = 25 μF, 50 V wkg.
C_5 = 0.1 μF ± 10 %, 350 v wkg.
C_6 = 0.1 μF ± 10 %, 350 V wkg.
C_7 = 50 μF, 25 V wkg.
C_8 = 32 μF, 350 V wkg.
C_9 = 64 μF, 350 V wkg.

Low-loading operating conditions are of particular influence on the design of the power supply. When the amplifier is used for speech and music only (which is usually the case) a total anode current drain of 110 mA for the complete amplifier is required. The reservoir capacitor should have a high capacitance to cater for the instantaneous current supply during peak power passages. As a rule the supply voltage of 310 V should not drop more than 2 to 3 V during such a power peak.

With normal loading a power stage with four EL 84 tubes will take about 180 mA at 300 V, whereas with low loading the quiescent current drain will be about 110 mA, so that a less expensive mains transformer may be used in combination with an EZ 81 rectifier tube. The ripple current through the reservoir capacitor is also considerably lower in this case, so that a less expensive type can be chosen.

Low loading has, moreover, the advantage of less distortion with speech and music reproduction, and a lower hum component. The only disadvantage is that the distortion cannot easily be measured and expressed in comparative figures, but this is of little importance if the quality of reproduction is appraised by hearing.

The phase-splitter and pre-stage used in this circuit has been described in Chapter I. An ECC 83 double triode is used for this purpose in a circuit with positive and negative feedback. The gain is about 220, so that a high sensitivity is reached with few components and tubes.

A negative feedback of 27 dB is obtained by connecting the 15 Ω tapping on the output transformer secondary to the cathode resistor of the first section of the ECC 83 double triode via an *RC* network. At the high-frequency end, the loop gain is reduced by means of C_3 (56 pF) across the feedback resistor R_6 and by the high-pass filter R_2C_1 of 3.9 kΩ and 2700 pF respectively, across the cathode resistor of the first triode section. With a large number of amplifiers according to this design no low-frequency instability was experienced.

THE OUTPUT TRANSFORMER

The specification of the output transformer is as follows:

Primary matching impedance (R_{a-a})	3.6	kΩ
Secondary matching impedances	3 - 5 - 7 - 15 - 400	Ω (100 V)
Primary inductance (anode-to-anode)	90	H
Leakage inductance between halves of primary	3.5	mH
Leakage inductance between primary and secondary	10	mH
Leakage resonant frequency	100	kc/s
Primary resistance	2 x 95	Ω
Maximum primary d.c.	2 x 175	mA
Maximum permissible d.c. unbalance	7	mA
Efficiency at 1000 c/s	91	%

The transformer is wound on a high-permeability core, and the various windings are subdivided into multiple interdigital coils. It was thus possible to use it with matching impedances differing ± 30 % from the nominal values.

When a choice is made between normal and low loading, and when the secondary load is known, a less expensive transformer may be designed, according to the details given in Chapter I, without appreciable reduction in quality. Then, however, the matching impedance should be as close as possible to a primary impedance of 3000 Ω in the case of low loading, and of 4000 Ω in the case of normal loading.

POWER SUPPLY

The mains transformer should be designed for delivering at normal loading 2 x 300 V_{rms}, 225 mA; 25 V_{rms} 5 mA; 6.3 V_{rms}, 3.5 A and 5 V_{rms}, 2 A. An additional winding of 2 x 3.75 V_{rms}, 2 A is recommended for the supply of a pre-amplifier and/or AM/FM tuner. When the H.T. of the latter is also taken from the same supply unit, the transformer should be dimensioned for 260 mA. This implies that the ripple current through the reservoir capacitor will be of the order of 370 mA.

For low loading the H.T. winding of the mains transformer should be dimensioned for 160 mA and the heater supply of the EZ 81 rectifier tube should be 6.3 V_{rms}, 1 A. This tube can supply a rectified current of 150 mA and a peak current of 450 mA, so that adequate power supply is available for low-loading conditions.

A separate pre-amplifier may be fed from the same H.T. unit, but when an AM/FM tuner is used the latter should be fed from its own power supply.

The grid bias for the output stage is obtained from the 25 V winding on the supply transformer and rectified by means of a germanium diode type OA 85, which can withstand relatively high peak inverse voltages. The pre-set voltage control R_{12} is so adjusted that the anode current of each power tube is 36 mA when normal loading is applied, and at 24 mA for low-loading operating conditions. Balance is obtained by adjustment of R_{13}.

FREQUENCY RESPONSE AND DISTORTION

Fig.29 shows the frequency response curve and the response without feedback. The harmonic distortion measured at frequencies of 40 c/s and 1 kc/s is given in Fig.30. These curves apply, of course, to normal loading.

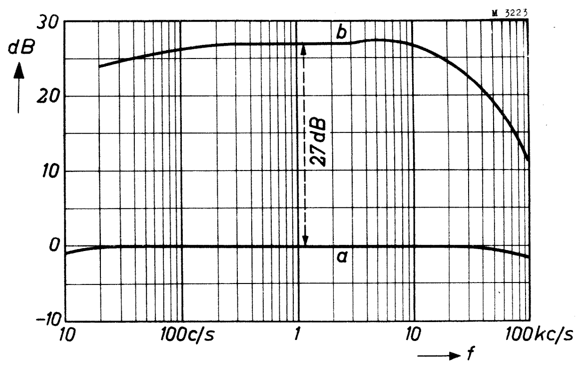


Fig.29. Frequency response (a) with and (b) without feedback of the 20 W amplifier with 2 x 2 EL 84 power pentodes in push-pull. The feedback is 27 dB at 1 kc/s.

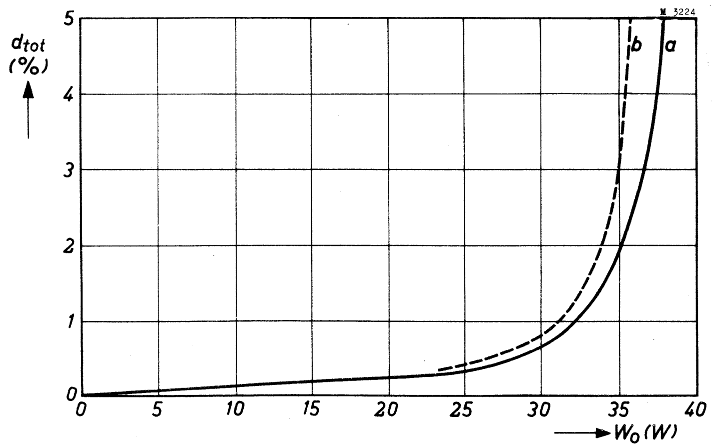


Fig. 30. Total harmonic distortion of the 20 W amplifier with 2 x 2 EL 84 power pentodes in push-pull, measured at 1 kc/s (a) and at 40 c/s (b).

20 W HIGH-FIDELITY AMPLIFIER

WITH 2 x EL34 IN AN ULTRA-LINEAR PUSH-PULL CIRCUIT

This amplifier uses two EL 34 power pentodes in an ultra-linear push-pull circuit. It is intended for use with a separate pre-amplifier containing all the controls, so that apart from the mains switch no controls are provided.

The amplifier gives sufficient power output for small halls and auditoriums. It can handle a peak output power of over 30 W without noticeable distortion.

SPECIFICATION

Tubes:	EF 86 pre-stage, ECC 83 driver/phase-splitter, 2 x EL 34 push-pull output stage, GZ 34 double anode rectifier.
Output power:	minimum 20 W from 30 c/s to 20 kc/s.
Power response:	flat within 0.5 dB referred to the 1 kc/s level at 20 W over a range from 30 c/s to 20 kc/s.
Frequency response:	at 1 W level flat within 1 dB referred to the 1 kc/s level, over a range from 2 c/s to 100 kc/s.
Harmonic distortion:	(400 c/s) $<$ 0.05 % at 20 W.
Intermodulation distortion:	(40 c/s and 10 kc/s; ratio 4 : 1) 0.7 % with peak corresponding to 20 W sine-wave power. 1 % with peak corresponding to 29 W sine-wave power.
Hum and noise:	-89 dB relative to 20 W with a 10 kΩ source resistance.
Sensitivity:	220 mV for 20 W output.
Phase shift:	10° maximum at 10 c/s; 20° max. at 20 kc/s.
Output impedance:	approximately 0.3 Ω (at 40 c/s, 1 kc/s and 20 kc/s) at 20 W output.

CIRCUIT DESCRIPTION

OUTPUT STAGE

Fig.31 shows the circuit diagram of the complete amplifier. The merits of this circuit with respect to power handling capacity and low distortion have already been discussed in detail in Chapter I.

The screen grids of the output stage are fed from tappings on the primary of the output transformer at about 40 % of the primary windings. The anode-to-anode load resistance is about 6.6 kΩ. The supply voltage is 440 V at the centre-tapping of the output transformer, and the combined screen-grid and anode dissipations of the output tubes are 28 W per tube.

With this particular screen-grid-to-anode load ratio it has been found that improved linearity is obtained at power levels above 15 W when resistors of the order of 1 kΩ are inserted in the

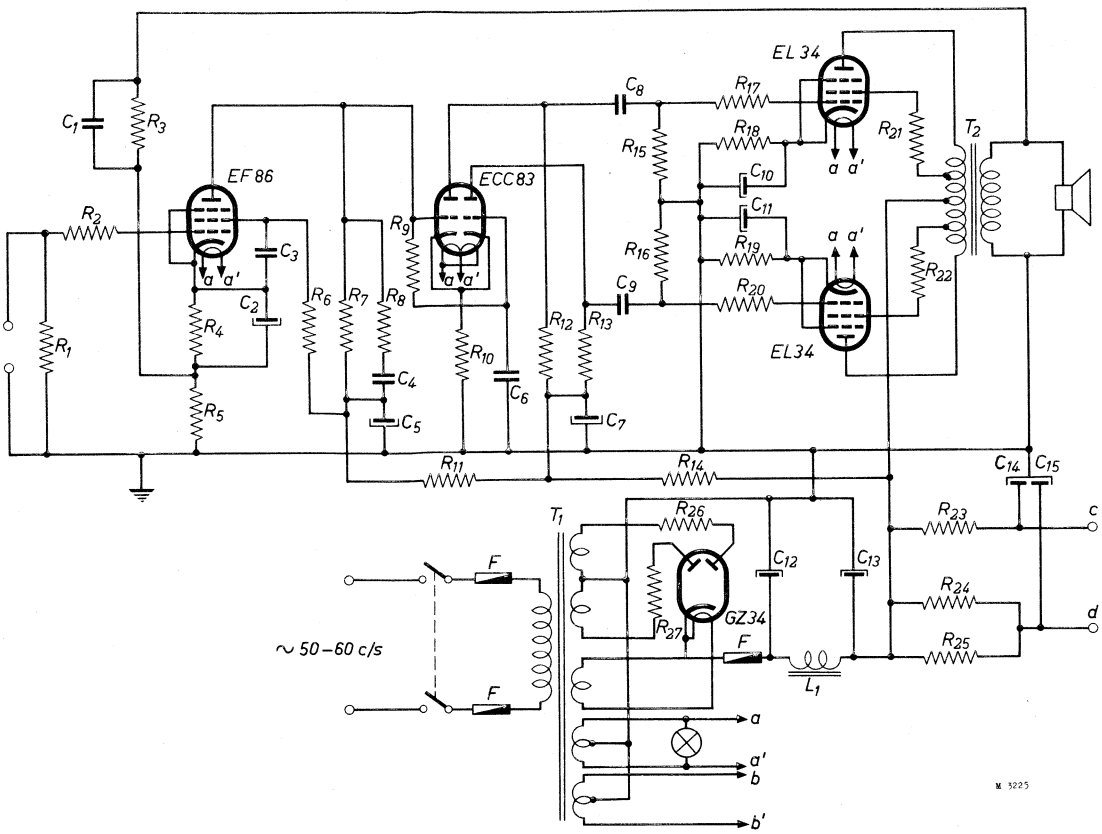


Fig.31. Circuit diagram of the 20 W amplifier with 2 x EL 34 power pentodes in an ultra-linear circuit.

R_1 = 1 MΩ ± 20 %, ¼ W
R_2 = 4.7 kΩ ± 20 %, ¼ W
R_3 = high stab. ± 5 %
12-16 Ω load: 8.2 kΩ
6-8 Ω load: 5.6 kΩ
R_4 = 2.2 kΩ ± 10 %, high stab.
R_5 = 100 Ω ± 5 %, high stab.
R_6 = 390 kΩ ± 10 %, high stab.
R_7 = 100 kΩ ± 10 %, high stab.
R_8 = 4.7 kΩ ± 10 %, ¼ W
R_9 = 1 MΩ ± 20 %, ¼ W
R_{10} = 82 kΩ ± 10 %, ½ W
R_{11} = 270 kΩ ± 10 %, ½ W
R_{12} = 180 kΩ ± 10 %, ½ W *)
R_{13} = 180 kΩ ± 10 %, ½ W *)
R_{14} = 15 kΩ ± 20 %, ½ W
R_{15} = 470 kΩ ± 10 %, ¼ W
R_{16} = 470 kΩ ± 10 %, ¼ W
R_{17} = 2.2 kΩ ± 20 %, ¼ W
R_{18} = 470 Ω ± 5 %, 3 W, wire wound
R_{19} = 470 Ω ± 5 %, 3 W, wire wound
R_{20} = 2.2 kΩ ± 20 %, ¼ W
R_{21} = 1 kΩ ± 10 %, ½ W
R_{22} = 1 kΩ ± 10 %, ½ W
R_{23} = 56 kΩ ± 10 %, 1 W
R_{24} = 12 kΩ ± 20 %, 6 W
R_{25} = 12 kΩ ± 20 %, 6 W
R_{26} = depending upon R_{tr}
R_{27} = depending upon R_{tr}
C_1 = ± 5 %
12-16 Ω matching: 220 pF
6-8 Ω matching: 330 pF
C_2 = 50 μF, 12 V wkg.
C_3 = 56 000 pF, 350 V wkg.
C_4 = 47 pF ± 10 %
C_5 = 8 μF, 450 V wkg.
C_6 = 0.22 μF, 350 V wkg.
C_7 = 8 μF, 450 V wkg.
C_8 = 0.47 μF, 350 V wkg.
C_9 = 0.47 μF, 350 V wkg.
C_{10} = 50 μF, 50 V wkg.
C_{11} = 50 μF, 50 V wkg.
C_{12} = 50 μF, 450 V wkg.
C_{13} = 50 μF, 450 V wkg.
C_{14}, C_{15} $\}$ 2 x 8 μF, 450 V wkg.

*) matched within 5 %, $R_{13} > R_{12}$

screen-grid connections. The slight reduction in the resulting peak power handling capacity is not significant in practice.

Separate cathode bias resistors are used to limit the out-of-balance direct current in the output transformer primary. The use of further d.c. balancing has not been considered necessary. It is likely, however, that some improvement in performance, particularly at low frequencies, would result from the use of d.c. balancing.

It is necessary in this type of circuit that the cathodes are bypassed to earth, even when a common cathode resistor is used. A low-frequency time constant in the cathode circuit cannot be eliminated when automatic bias is used.

DRIVER STAGE

An ECC 83 double triode fulfils the combined functions of driver and phase splitter. The cathode-coupled form has been chosen, which makes a high degree of push-pull balance at very low distortion possible. With the high supply voltage available the required drive voltage can be obtained at a distortion level of only 0.4 %. The resistors R_{12}, R_{13} of 180 kΩ must be matched within 5 % for optimum performance, R_{13} having the higher value.

Optimum balance is obtained when the effective anode loads differ by 3 %. It is necessary that the grid resistors R_{15}, R_{16} are of small mutual tolerance since they form part of the anode loads of the driver stage. High-frequency balance will be largely determined by wiring layout since equality of shunt capacitance is required. Low-frequency balance is controlled by the value of the time constant R_9C_6 in the grid circuits. The value of 0.25, which has been chosen, ensures adequate balance down to very low frequencies.

A disadvantage of the cathode-coupled phase splitter circuit is that the effective voltage gain is about one half of that obtained with one section used as a normal voltage amplifier. Due to the high mu of the ECC 83, however, the effective stage gain in the circuit is still about 25.

PRE-STAGE

A high-gain low-noise EF 86 pentode has been used in the pre-stage. The stage gain is about 120. High-stability cracked carbon resistors are used in the anode, screen-grid and cathode circuits, and give appreciable improvement in measured background noise level as compared with ordinary carbon resistors. The pre-stage is d.c. coupled to the phase splitter in order to bring the grids of the phase splitter at the required positive potential and simultaneously to minimize the low-frequency phase shift in the amplifier and to improve low-frequency stability when feedback is applied.

NEGATIVE FEEDBACK

The sensitivity of the amplifier without feedback is 6.5 mV for 20 W output, with feedback about 220 mV is required for the same output power, the designed overall loop gain being about 30 dB. The loop gain, the overall frequency response and the phase shift characteristics of the amplifier are shown in Fig.32.

Notwithstanding the high degree of feedback, an adequate margin of stability has been achieved. Complete stability is maintained under open-circuit conditions in the prototype amplifier. An increase in feedback of at least 10 dB, obtained by reducing the value of R_3, should be possible before signs of high-frequency instability occur. In this design oscillation with capacitive loads is the form of instability most likely to occur, but even with very long loudspeaker cords there is little risk of instability.

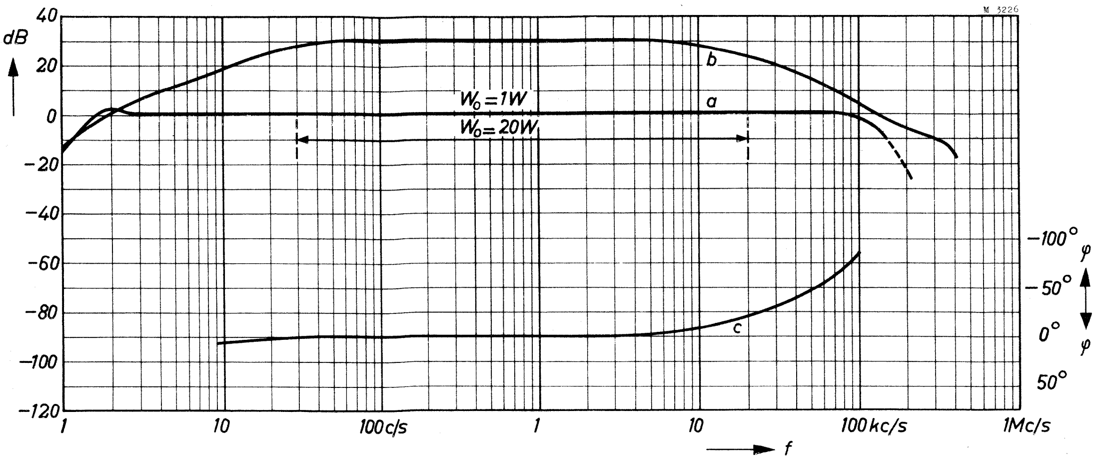


Fig.32.Performance of the 20 W ultra-linear amplifier with 2 x EL 34. The frequency response and power response (a), the loop gain (b) and the phase shift (c) are plotted against the frequency.

DISTORTION

The harmonic distortion of the experimental amplifier at 400 c/s, measured without feedback under resistive load conditions, is shown in Fig.33. The distortion towards the overload point is also shown with 30 dB feedback. At the 20 W level the distortion without feedback is well below 1 %, and with feedback it falls below 0.05 %. Harmonic distortion at 400 c/s reaches 0.1 % at about 27 W output.

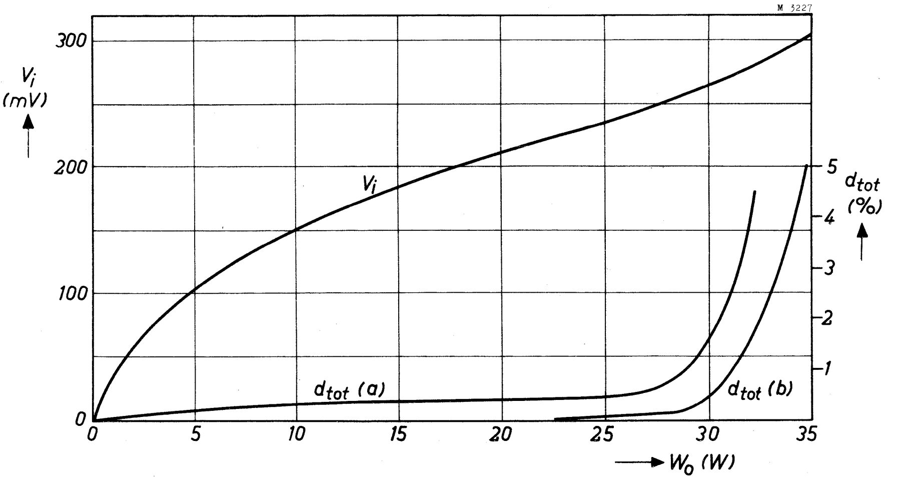


Fig.33. Total harmonic distortion measured without feedback (a) and with 30 dB feedback (b) at 400 c/s, and the input voltage (V_i) as functions of the output power.

The loop gain characteristics are such that at least 20 dB feedback is maintained from 15 c/s to 25 kc/s, and 26 dB down to 30 c/s.

Measurements of intermodulation distortion have been made, using frequencies of 40 c/s and 10 kc/s in a 4 : 1 amplitude ratio. With the combined peak amplitude of the mixed output at a level corresponding to the peak sine wave amplitude at 20 W output power, intermodulation products expressed in r.m.s. terms totalled 0.7 % of the 10 kc/s amplitude, and at 29 W about 1 %.

The output versus input voltage characteristic shown in Fig.34 reveals that excellent linearity is obtained up to 20 V across 15 Ω, corresponding to 27 W output power.

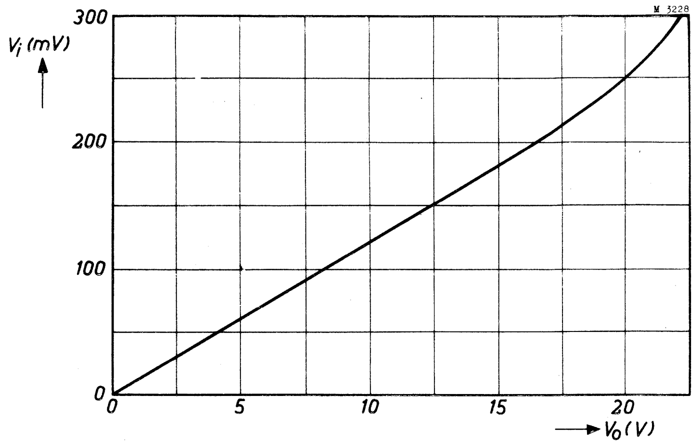


Fig. 34. Input voltage plotted against the output voltage with 15 Ω load on the secondary of the output transformer.

SENSITIVITY

The sensitivity of the amplifier is about 220 mV for 20 W output power and 300 mV at the overload point at mid frequencies. The background level in the experimental amplifier was 89 dB below 20 W, measured with a source resistance of 10 kΩ. This is equivalent to about 5.5 μV at the input terminals. It is possible to increase the overall sensitivity of the amplifier by 6 dB whilst still maintaining a low background level, high loop gain and a high margin of stability. However, considerations involved in the design of suitable pre-amplifier circuits, in particular the need for adequate signal-to-noise ratio, render a higher sensitivity of doubtful advantage.

POWER RESPONSE

The power handling capacity at the low-frequency end of the audible range is chiefly determined by the quality of the output transformer. It is desirable that associated pre-amplifier circuits should attenuate the very low frequencies which the amplifier is incapable of handling at the rated power output without excessive distortion. With the output transformer described in the following section, at least 20 W is available down to 20 c/s, and the frequency response at the 20 W level is linear from about 30 c/s to 20 kc/s.

OUTPUT TRANSFORMER

The output transformer has been made for matching impedances of 6 Ω to 8 Ω and for 12 Ω to 16 Ω. The primary inductance measured at 5 V, 50 c/s is 72 H; at 10 V, 50 c/s it is 120 H. The stray inductance is 8 mH with the secondary short-circuited, and 6 mH with one half of the primary short-circuited. The total primary resistance is 310 Ω, that of the secondary is 0.45 Ω at 12 to 16 Ω matching and 0.18 Ω at 6 to 8 Ω matching. The maximum flux density B_{max}, measured at 20 c/s and 500 V peak, is 5800 gauss.

The coils are wound on a former that is accurately divided in two sections, carrying each one half of the primary. Each of these is subdivided into five sections which are connected in series with the secondary windings in between. Hence there are ten primary and eight secondary coils, the latter being partially connected in parallel, partially in series.

The details of the output transformer are:

Core:	shell type, normal dynamo sheet.
Overall dimensions:	150 x 125 mm.
Width of core:	50 mm
Stacking:	without air gap
Height of stacking:	50 mm
Cross section:	25 cm^2

The dimensions of the laminations are given in Fig.35. The coil former has double flanges at the ends and one additional flange exactly in the middle. The windings are given below in succession as they are wound on the former.

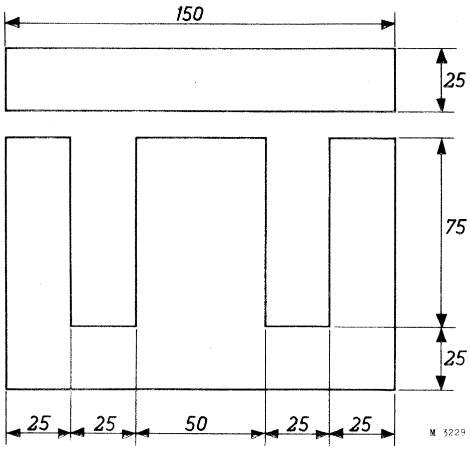


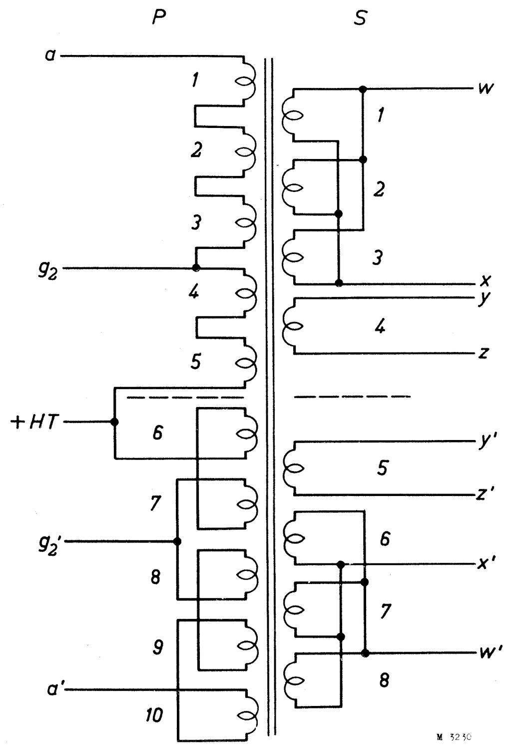
Fig.35. Dimensions in mm of the input transformer core laminations.

Winding	Number of turns	Diameter of wire	Width of winding	Number of layers
P_1. P_{10}	380	0.28 mm	32 mm	4
S_1. S_8	60	1.0 mm	33 mm	2
P_2. P_9	380	0.28 mm	32 mm	4
S_2. S_7	60	1.0 mm	33 mm	2
P_3. P_8	380	0.28 mm	32 mm	4
S_3. S_6	60	1.0 mm	33 mm	2
P_4. P_7	380	0.28 mm	32 mm	4
S_4. S_5	60	1.0 mm	33 mm	2
P_5. P_6	380	0.28 mm	32 mm	4

Enamelled copper wire is used throughout. In the experimental set-up the windings were interleaved with one layer of Fuller board of 0.1 mm and two layers of paper of 60 μ. The beginning of the windings should be against the outer flanges.

The following internal connections are to be made between the primary coils:

end of P_1 to beginning P_2	end of P_{10} to beginning P_9
end of P_2 to beginning P_3	end of P_9 to beginning P_8
end of P_3 to beginning P_4	end of P_8 to beginning P_7
end of P_4 to beginning P_5	end of P_7 to beginning P_6
end of P_5 to end of P_6	



The beginnings and ends of the secondary coils S_1, S_2, S_3 and of S_6, S_7, S_8 are interconnected, so that the coils of each group are connected in parallel (see Fig.36).

The external connections are:

Beginning of P_1 to anode EL 34 (I), beginning of P_{10} to anode EL 34 (II). Interconnection between P_3 and P_4 to screen grid EL 34 (I), interconnection between P_7 and P_8 to screen grid EL 34 (II). The interconnection between P_5 and P_6 to + H.T. supply.

Fig. 36. Connections of the primary and secondary windings.

The connections of the secondary can best be explained with the aid of Fig.36. For 5 to 8 Ω matching the beginnings of S_1, S_2 and S_3 are connected to the beginning of S_4, and the ends of these coils are also interconnected (*w* to *y* and *x* to *z*). This also applies to the other coils, viz. *w'* to *y'* and *x'* to *z'*. The combinations are connected in series, and the loudspeakers are connected to the terminals *w* and *w'*. For 12 to 16 Ω matching the coils S_4 and S_5 are connected in parallel (*y* to *y'* and *z* to *z'*). The three groups are connected in series, hence *y* and *y'* to *x*, and *z* and *z'* to *x'*, the loudspeakers being connected to *w* and *w'*.

POWER SUPPLY

The supply transformer should have the following secondary windings. The H.T. winding should be for 410 V, 180 mA (so that, in addition to the amplifier, an AM/FM broadcast tuner can be connected); a 5 V, 3 A winding for the heater supply of the GZ 34 double anode rectifier tube; a centre-tapped 6.3 V 4 A winding for the heater supply of the amplifier; and finally a centre-tapped 6.3 V, 3 A winding for the heater supply of the pre-amplifier and of a tuner for broadcast reception.

The values of the resistors R_{26} and R_{27} depend upon the d.c. resistance of the transformer windings. The transformer resistance with the GZ 34 should be at least 110 Ω.

The choke L_1 should have an inductance of 5 to 8 H at 180 mA, and a resistance of the order of 200 Ω.

The reservoir capacitor of the smoothing filter, C_{12}, should be suitable for a working voltage of 450 V, and be able to carry the heavy ripple current.

The H.T. supply of the pre-amplifier can be taken from the terminal *c*, a broadcast tuner may be connected to *d*. These supply voltages are decoupled by a double electrolytic capacitor, C_{14}-C_{15} of 2 x 8 μF 450 V.

10 W SINGLE-ENDED PUSH-PULL AMPLIFIER

OF EXTREMELY HIGH QUALITY WITH TWO TUBES EL86

Single-ended push-pull stages were originally designed for use in broadcast receivers, in combination with high-impedance loudspeakers. This type of circuit leads to a slight reduction in cost with an even higher output and quality of reproduction than is usually obtained in a normal class A stage. In these circuits *) the tone control is normally included in the feedback loop at the expense of the quality of reproduction. When, however, the circuit is slightly modified and used with a separate pre-amplifier containing the equalization and tone controls, an A.F. amplifier of extremely high quality can be obtained. The use of high-impedance loudspeakers obviously remains essential.

One difficulty encountered with the design of this circuit was, that the distortion could not be measured with the conventional audio oscillators and distortion meters. The distortion of the former is considerably higher than that of the amplifier, especially at a low output. With the customary distortion meters percentages lower than 0.1 % cannot be measured. Therefore tuned filters were connected between the audio oscillators and the input of the amplifier, and the distortion was measured with a wave analyzer.

SUMMARY OF PERFORMANCE

Power output:	10 W
Harmonic distortion:	at 10 W: < 0.3 %
	at 11 W: < 1 %
	at 2 W: < 0.02 %
Frequency response:	flat from 7 c/s to 40 kc/s
	3.8 dB down at 300 kc/s
Power response:	the 1 % line is flat from 30 c/s to 20 kc/s and could not be measured beyond this frequency owing to the limited frequency range of the distortion meter.
Sensitivity:	at 50 mW output power: 40 mW
	at full output: 0.6 V
Output impedance:	800 Ω
Internal resistance:	50 Ω

CIRCUIT DESCRIPTION

Fig. 37 shows the circuit diagram. Two pentodes EL 86 are used in the output stage; a high-mu double triode ECC 83 operates as a pre-stage amplifier and phase splitter in a positive feedback circuit.

One of the problems encountered in single-ended push-pull circuits with pentodes is the screen-grid supply, especially of the upper

*) See, e.g. Electr. Appl. *17*, p. 81, 1956/57 (No. 3)

pentode. This has been solved by feeding each screen grid from its own anode via one coil of a double choke. The magnetization of the core is neutralized as a result of the equal screen-grid currents passing in opposite directions through the windings. Thus an inductance of 60 H for each choke is easily obtained with a relatively small component.

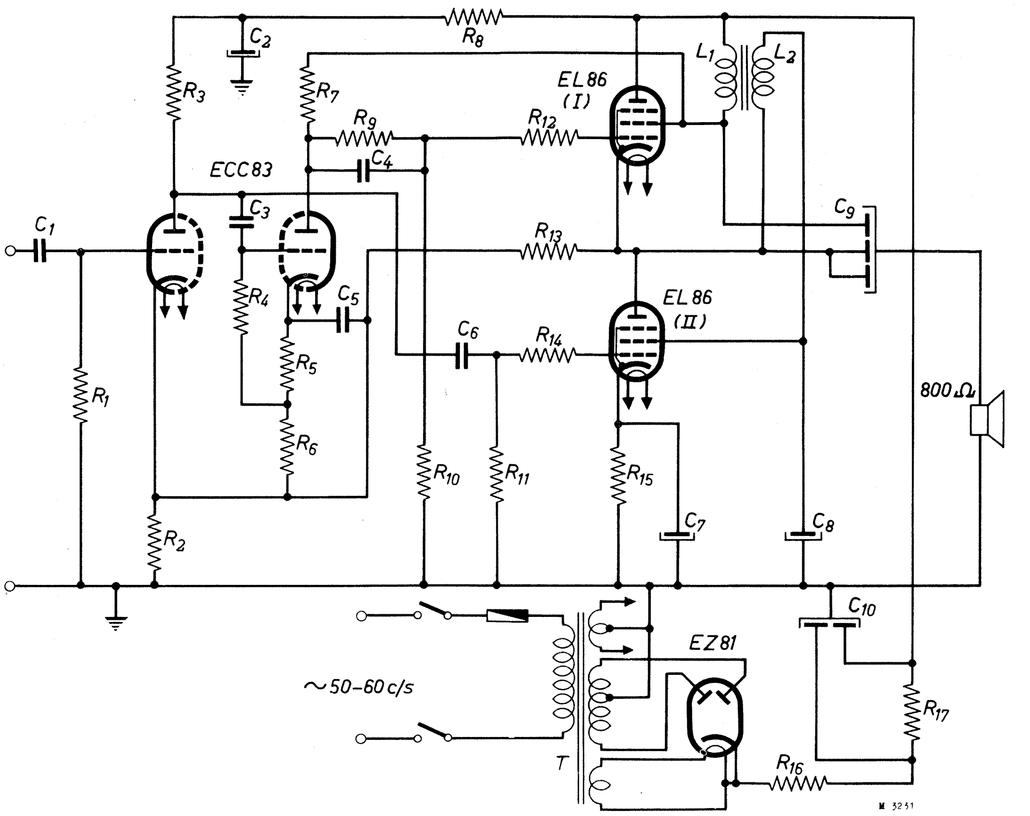


Fig.37.Circuit diagram of the single-ended push-pull amplifier with 2 x EL 86.

R_1	= 1 MΩ ± 20 %, ¼ W	R_{15}	= 150 Ω ± 5 %
R_2	= 680 Ω ± 5 %, high stab. ½ W	R_{16}	= See text
R_3	= 220 kΩ ± 10 %, high stab. ¼ W	R_{17}	= 150 Ω ± 20 %, 6 W, wire wound
R_4	= 1 MΩ ± 20 %, ¼ W		
R_5	= 5.1 kΩ ± 5 %, high stab. ¼ W	C_1	= 0.1 μF
R_6	= 33 kΩ ± 5 %, high stab. ¼ W	C_2	= 8 μF, 350 V wkg.
R_7	= 100 kΩ ± 5 %, high stab. ¼ W	C_3	= 0.1 μF, 350 V wkg.
R_8	= 100 kΩ ± 10 %, ¼ W	C_4	= 0.1 μF, 350 V wkg.
R_9	= 620 kΩ ± 5 %, high stab. ¼ W	C_5	= 47 pF
R_{10}	= 1 MΩ ± 5 %, high stab. ¼ W	C_6	= 47 000 pF, 350 V wkg.
R_{11}	= 1 MΩ ± 10 %, ¼ W	C_7	= 100 μF, 25 V wkg.
R_{12}	= 1 kΩ ± 20 %, ¼ W	C_8	= 8 μF, 350 V wkg.
R_{13}	= 120 kΩ ± 5 %, high stab. ¼ W	C_9	= 3 x 50 μF, 350 V wkg.
R_{14}	= 1 kΩ ± 20 %, ¼ W	C_{10}	= 2 x 50 μF, 350 V wkg.

The loudspeaker carries only audio currents and can be connected to earth. The power loss is, moreover, negligible; the power response curve falls rapidly below 30 c/s.

The lower EL 86 is biased by a bypassed cathode resistor of 150 Ω, the 1 MΩ grid leak being connected to earth. The 1 mΩ grid leak of the upper EL 86 is also connected to earth, so that considerable feedback is applied which contributes to the low distortion, but involves the necessity of applying a positive direct voltage to the grid of the tube. This has been achieved by connecting a resistor of well-defined value across the coupling capacitor between the anode of the right-hand section of the ECC 83 double triode and the grid of the upper EL 86. A 5 % tolerance resistor of 0.62 MΩ serves the purpose.

It will be clear that the best balance is obtained when the d.c. setting of the two pentodes is identical, so that with a supply voltage of 320 V and a grid bias of -11 V for each pentode, the anode voltage of the lower EL 86 is 165.5 V with respect to earth. This implies that the voltage at the grid of the upper EL 86 should be 154.5 V with respect to earth. This voltage now depends heavily on the adjustment of the right-hand section of the ECC 83 double triode, which takes its supply voltage from the screen grid of the upper EL 86. The resistance of each winding of the double choke is about 400 Ω, and the screen-grid current about 4 mA. The voltage drop in the choke will thus be roughly 1.6 V, so that the supply voltage of the triode may be taken to be 318 V. The voltage drop in the anode load resistor of 100 kΩ should be 67 V, to ensure that the correct bias is applied to the upper EL 86. The current through the voltage divider R_9R_{10} is 0.555 mA, that through R_7, 0.67 mA, which means that the anode current of the triode should be 0.51 mA. Assuming the voltage between anode and cathode to be about 230 V, the grid bias of the tube should be 2.4 V for obtaining this current. Hence the cathode resistor should be 4.7 kΩ; In practice a 5 % tolerance resistor of 5.1 kΩ is recommended.

This exposition demonstrates that in the design of this type of circuit all circuit elements are related to each other, and that alteration of one circuit element may greatly influence the operating conditions in a different part of the circuit.

This applies also to the audio frequency conditions. The upper EL 86 may be considered as a cathode follower. Its cathode resistor is formed by the lower EL 86 on which an a.f. voltage is superimposed which is equal to the output voltage of the amplifier. Now the gain of a cathode follower is smaller than unity, and hence the input voltage of the upper EL 86 is equal to the output voltage plus the required driving voltage for obtaining that output voltage in a normally earthed circuit, that is about 7 V_{rms} for full drive. Suppose the output voltage is 100 V_{rms}, then the driving voltage of the upper EL 86 should be 107 V_{rms}. Such a voltage cannot be delivered by an ECC 83 in a normal circuit. Therefore the anode of the second tube section, which operates as a phase splitter, is fed from the screen grid of the corresponding output pentode, the potential of which fluctuates with the output voltage of the amplifier. The anode load resistor of the phase splitter is 100 kΩ, and the apparent resistance of the phase splitter in series with the resistors in the cathode circuit are approximately 800 kΩ, so that 89 % of the output voltage appears at the anode of the phase splitter. Hence only 18 V_{rms} need be delivered by the phase splitter to obtain the required input voltage of 107 V_{rms}.

The left-hand section of the ECC 83 double triode serves as a prestage amplifier. Its output signal is fed to the grid of the lower

EL 86 via C_6, and to the grid of the phase splitter via C_3. The pre-stage is operated with a higher load resistance than the phase splitter. The output signal of the latter should be about 2.5 times that of the former, but at the chosen values of the load resistors the difference in gain is insufficient to obtain this ratio. The value of the unbypassed cathode resistor of the pre-stage amplifier is therefore made much smaller than that of the phase splitter, so that the latter has a higher negative feedback.

The function of this cathode resistor (R_2) is rather peculiar. It carries the d.c. of the pre-stage amplifier, of the phase splitter and the current flowing through the feedback resistor R_{13}. The total current is about 2.35 mA so that the voltage across R_2 (680 Ω) is about 1.6 V, which serves as a cathode bias for the pre-stage amplifier. In addition to the direct current, the alternating currents of the pre-stage and the phase splitter, which are in anti-phase, flow through R_2. The alternating current of the phase splitter predominates, so that positive feedback is applied to the pre-stage which increases the gain considerably, but may also lead to instability. The tendency to instability, however, is completely suppressed by the overall negative feedback from the output of the amplifier to the input via R_{13}.

The positive feedback in the pre-stage may not be so high that the tube is set at the verge of oscillating. When in that case the output stage is over-driven by too high an input signal, the distorted output signal and hence the negative feedback would no longer have the correct ratio to the input signal, so that the pre-stage would start to oscillate. When the given values of circuit elements, stability is ensured.

It will be noticed that the coupling capacitor C_4 to the grid of the upper EL 86 has a value of 0.1 μF, whereas the coupling capacitor to the grid of the lower output tube is only 47 000 pF. In this way the grid drive at low frequencies is kept correct. The grid of the lower EL 86 is fed directly from the pre-stage, but that of the upper EL 86 from the phase splitter, which is in turn coupled to the anode of the pre-stage amplifier via a 0.1 μF capacitor. The time constants in the grid circuits of the output tubes are thus made equal, and unbalance at low frequencies with inherent distortion is avoided.

The capacitor C_5 (47 pF) is inserted to compensate the Miller effect in the phase splitter, which would lead to attenuation of the drive of the upper EL 86 at the higher frequencies.

The power supply is rather simple. A supply transformer giving 2 x 300 V_{rms}, 90 mA; 6.3 V, 2.4 A and 6.3 V, 1 A, is sufficient to feed not only the amplifier but also a pre-amplifier-equalizer. The EZ 81 double anode rectifier is used. Smoothing is obtained with a double electrolytic capacitor of 2 x 50 μF in combination with a 150 Ω, 6 W wire-wound resistor. Under these conditions the transformer resistance should be 2 x 200 Ω. The smoothing circuit could be kept simple because the screen grids of EL 86's are adequately decoupled by means of the double choke, whilst the anode supply of the pre-stage is decoupled by means of a 100 kΩ resistor and an electrolytic capacitor of 8 μF. The latter is combined with the 8 μF capacitor used for bypassing the screen grid of the lower EL 86. The 50 μF electrolytic capacitor which decouples the screen grid of the upper EL 86 is combined with the 2 x 50 μF series capacitor for the output circuit, which has the advantage that space is saved and only one insulation ring is needed between the can and the chassis.

PRE-AMPLIFIER

The single-ended push-pull amplifier requires a pre-amplifier-equalizer, for which the circuit given on page 57 is recommended. The pre-amplifier may be used either separately or built on the same chassis as the power amplifier. In the latter case adequate screening between the two sections is essential.

MEASURED RESULTS

FREQUENCY RESPONSE

The frequency response curve is given in Fig. 38. The curve is virtually flat (less than 0.5 dB down) between 7 c/s and 40 kc/s. The 3 dB point is at about 230 kc/s; at 300 kc/s the curve is about 3.8 dB down.

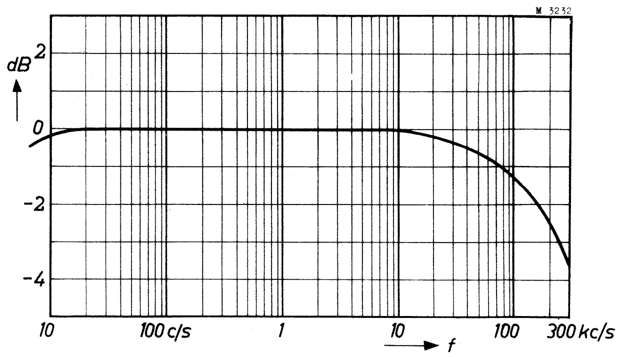


Fig.38. Frequency response curve of the single-ended push-pull amplifier with 2 x EL 86.

POWER RESPONSE

Fig. 39 shows the power response curve measured at 1 % harmonic distortion. The curve falls steeply below 30 c/s due to the choke in the screen-grid circuit of the upper EL 86. Power response could not be measured in excess of 20 kc/s due to the limited range of the distortion meter.

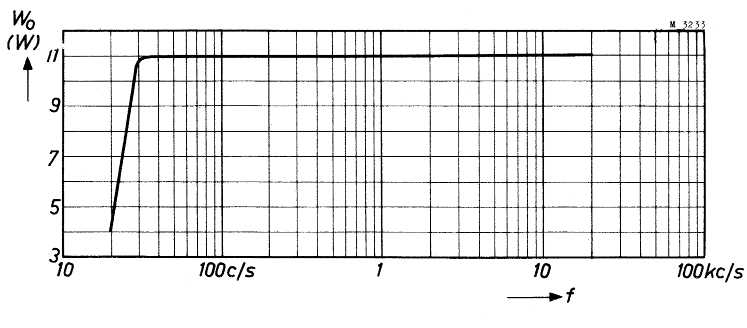


Fig.39. Power response curve for 1 % harmonic distortion.

HARMONIC DISTORTION

Fig. 40 gives the harmonic distortion as a function of the power output for frequencies of 90 c/s, 400 c/s, 1 kc/s and 4 kc/s.

The harmonic distortion was measured at full output with frequencies of 10 kc/s and 20 kc/s. At 11 W output the distortion was in both cases below 1 %; at 10 W output 0.23 % distortion was measured at 10 kc/s, and 0.33 % at 20 kc/s.

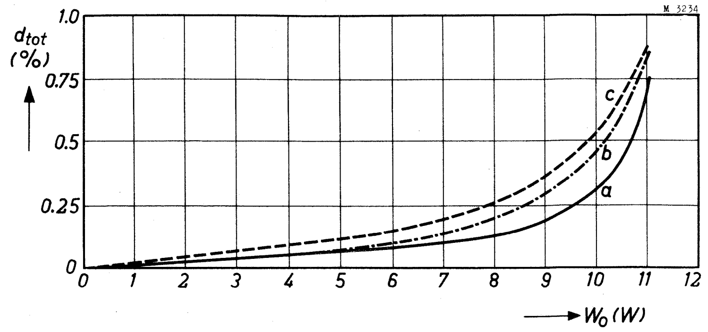


Fig. 40. Total harmonic distortion as a function of the output power, measured at 400 c/s, and 1 kc/s (a), at 4 kc/s (b) and at 90 c/s (c).

SQUARE-WAVE RESPONSE

Fig. 41 shows the square-wave response oscillograms of the amplifier at frequencies of 20 c/s, 1 kc/s, 10 kc/s, 20 kc/s and 60 kc/s. It is seen that the square-wave response is ideal at 1 ks/s and reasonably good at the other frequencies.

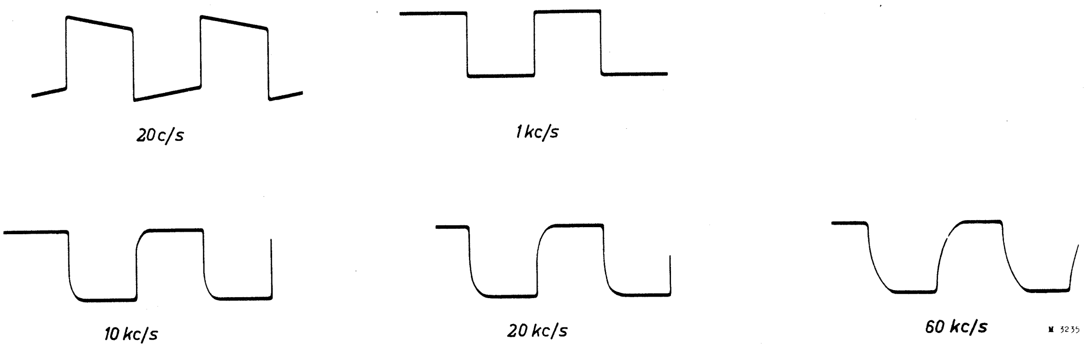


Fig.41. Square-wave response of the single-ended push-pull amplifier, at different frequencies.

PHASE SHIFT

The phase response curve is shown in Fig. 42. It is seen that the phase leads 50° at 5 c/s, and lags 20° at 100 kc/s, so that stability is ensured.

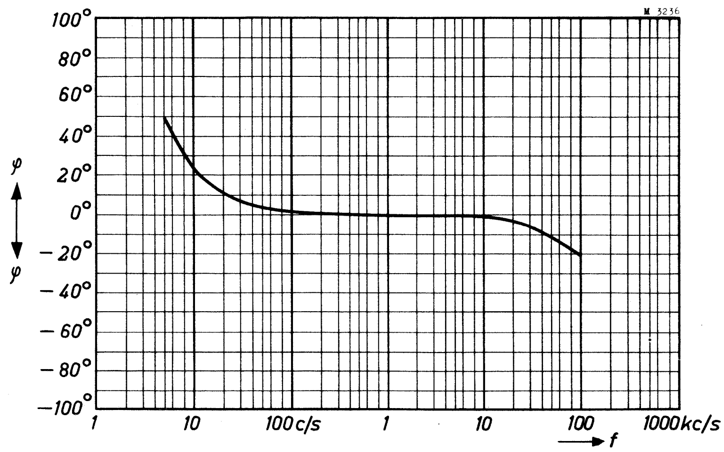


Fig.42. Phase shift of the single-ended push-pull amplifier.

Chapter III

Pre-Amplifiers

INTRODUCTION

In small amplifiers the pre-amplifier is usually built on the same chassis as the power amplifier, and the lay-out is very simple. More elaborate installations have a separate pre-amplifier with built-in equalizers for various recording characteristics and for broadcast reception.

The recording characteristics used at present are given in Fig.43, viz. the standard curve for microgroove recordings used in the United States, the A.E.S. curve, which is identical to the British standard R.I.A.A. characteristic. This curve is represented by the broken line. The dot-dash line is the latest 78 r.p.s. recording characteristic as advised by the R.I.A.A.; the full line is the Decca ffrr curve as used on the test record LXT 2695.

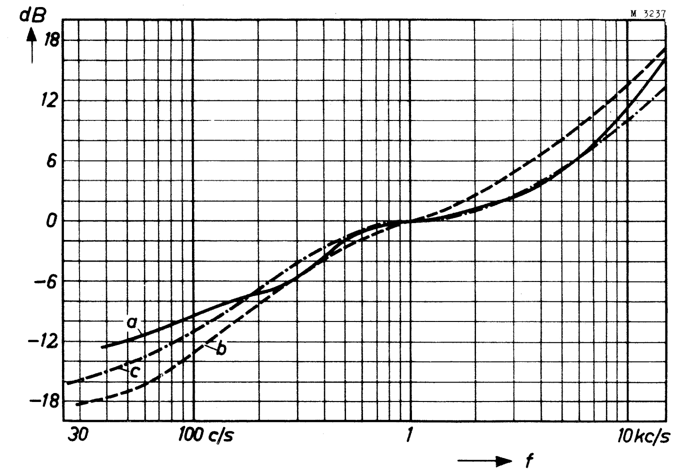


Fig.43. Disc recording characteristics according to the present A.E.S. and R.I.A.A. conventions.
(a) Decca ffrr curve as on test record LXT 2695, (b) microgroove, (c) 78 r.p.m. records.

Old 78 r.p.m. records are made according to a very simple characteristic; below 500 c/s the curve drops by about 5 dB per octave.

Pre-amplifiers are often provided with an equalizer consisting of frequency-dependent feedback circuits the response characteristics of which may be selected to match the recording characteristic, so that a linear response can be obtained from the record plus the equalizer circuit.

Apart from the equalizer networks, the pre-amplifiers contain all necessary controls, such as volume, bass and treble response controls, sometimes rumble filters, noise filters, etc. Although very comprehensive pre-amplifiers are available, only relatively simple circuits are dealt with here, because elaborate circuits require much engineering with respect to circuit lay-out, wiring, screening and choise of components. As a result, they can hardly be realized, with a fair chance of stable operation, from the circuit diagram only.

TWO-TUBE PRE-AMPLIFIER

The two-tube pre-amplifier discussed below is intended for use with the 10 W and 20 W amplifiers described in Chapter II. Provision has been made in this equipment to connect magneto-dynamic and crystal pick-ups, tape-recorder play-back heads, microphone and radio inputs. An auxiliary socket is provided for any other convenient input source.

Selection of an input channel is achieved by a switch on the front panel. The sockets for magneto-dynamic and crystal pick-ups should not be used simultaneously. The switch positions, from left to right, are: (1) Crystal pick-up or magneto-dynamic pick-up for 78 r.p.m. records; (2) the same for microgroove records; (3) microphone; (4) tape; (5) radio and (6) auxiliary.

The equalization for disc recordings conforms to the latest R.I.A.A. or A.E.S. standards, which have been adopted by most large recording companies. The tape play-back characteristic is intended to play back pre-recorded tapes at a speed of 19 cm (7½") per second. Low-impedance tone controls covering a wide frequency range are used in the amplifier and provide adequate control for most applications.

SPECIFICATION

Tubes:	2 x EF 86 low-noise amplifier pentode.
Output voltage:	40 mV for 10 W amplifiers, 250 mV for 20 W amplifiers.
Sensitivity:	crystal pick-up { microgroove 50 mV 78 r.p.m. 150 mV magneto-dynamic pick-up { micro-groove 3 mV, 78 r.p.m. 9 mV. microphone: 6 mV tape play-back: 3 mV at 5 kc/s radio input: 250 mV auxiliary: 250 mV
Hum and noise:	55 dB below 10 W for both pick-up microgroove positions and 57 dB below 10 W for the 78 r.p.m. position. microphone: 44 dB below 10 W tape: 53 dB below 10 W
Input impedance:	100 kΩ for all pick-up positions; 1 MΩ for the microphone, 80 kΩ for the tape, and 2 MΩ for the radio and auxiliary positions.
Distortion	< 0.15 % total harmonics at the nominal output level and 0.24 % at ten times the nominal output level.
Tone control:	bass boost maximum + 17 dB at 50 c/s, bass cut maximum - 14 dB at 50 c/s, treble boost maximum + 14 dB at 10 kc/s, treble cut maximum - 15 dB at 10 kc/s, all referred to 1 kc/s.

CIRCUIT DESCRIPTION

The pre-amplifier (Fig.44) is made up of two stages each of which uses a high-gain pentode, type EF 86. All the equalization takes place in the first stage, and is achieved by means of frequency-selective feedback between the anode and grid of the first EF 86. There is no feedback in the second stage, and the output from the second EF 86 is taken directly to a passive tone-control network.

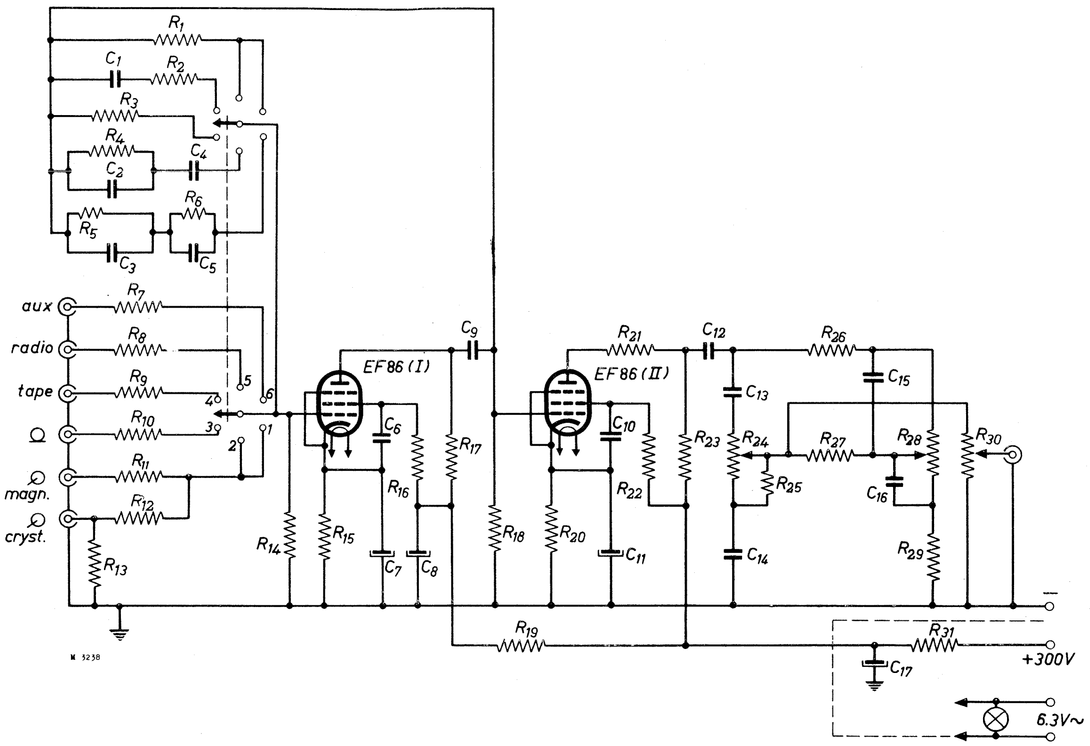


Fig.44. Circuit diagram of the two-tube pre-amplifier with 2 x EF 86.

R_1 =	330 kΩ ± 5 %, ¼ W	R_{25} =	47 kΩ ± 10 %, ¼ W
R_2 =	560 kΩ ± 5 %, ¼ W	R_{26} =	68 kΩ ± 10 %, ¼ W
R_3 =	10 MΩ ± 5 %, ¼ W	R_{27} =	39 kΩ ± 10 %, ¼ W
R_4 =	560 kΩ ± 5 %, ¼ W	R_{28} =	250 kΩ, log.
R_5 =	5.6 MΩ ± 5 %, ¼ W	R_{29} =	6.8 kΩ ± 10 %, ¼ W
R_6 =	220 kΩ ± 5 %, ¼ W	R_{30} =	250 kΩ, log.
R_7 =	2.2 MΩ ± 10 %, ¼ W	R_{31} =	30 kΩ ± 10 %, ½ W
R_8 =	2.2 MΩ ± 10 %, ¼ W	C_1 =	390 pF ± 5 %
R_9 =	56 kΩ ± 10 %, ¼ W	C_2 =	150 pF ± 5 %
R_{10} =	1 MΩ ± 10 %, ¼ W	C_3 =	2200 pF ± 5 %
R_{11} =	68 kΩ ± 10 %, ¼ W	C_4 =	560 pF ± 5 %
R_{12} =	1 MΩ ± 10 %, ¼ W	C_5 =	220 pF ± 5 %
R_{13} =	100 kΩ ± 10 %, ¼ W	C_6 =	0.1 μF, 350 V wkg.
R_{14} =	100 kΩ ± 5 %, high stab. ¼ W	C_7 =	25 μF, 12 V wkg.
R_{15} =	2.2 kΩ ± 10 %, ½ W	C_8 =	8 μF, 350 V wkg.
R_{16} =	1 MΩ ± 10 %, high stab. ½ W	C_9 =	0.1 μF, 350 V wkg.
R_{17} =	220 kΩ ± 10 %, high stab. ½ W	C_{10} =	0.1 μF, 350 V wkg.
R_{18} =	1 MΩ ± 10 %, ¼ W	C_{11} =	25 μF, 12 V wkg.
R_{19} =	33 kΩ ± 10 %, ½ W	C_{12} =	0.1 μF, 350 V wkg.
R_{20} =	1.2 kΩ ± 10 %, ½ W	C_{13} =	560 pF ± 10 %
R_{21} =	82 kΩ ± 10 %, high stab. ½ W *)	C_{14} =	8200 pF ± 10 %
R_{22} =	390 kΩ ± 10 %, high stab. ½ W *)	C_{15} =	2200 pF ± 10 %
R_{23} =	18 kΩ ± 10 %, high stab. ½ W	C_{16} =	20 000 pF ± 10 %
R_{24} =	250 kΩ, log.	C_{17} =	16 μF, 350 V wkg.

This arrangement was chosen so that the grid-circuit impedance of the first stage should be low. A low impedance at this grid lessens hum pick-up and reduces the effect of plugging-in external low-impedance circuits. Furthermore, the arrangement also results in low gain in the first stage. Hence, Miller effect between the anode and grid of the first EF 86, which can be troublesome when high values of resistance are used in series with the grid, is reduced.

Series resistors are used in the input channels so that the sensitivity and impedance of any channel can be adjusted accurately. The component values given in Fig.44 are intended for sources encountered most frequently, but the sensitivity and impedance*) of each channel can be altered by changing the value of the appropriate series resistor.

The sensitivity of the pre-amplifier can be altered for all the input channels by varying the output from the second EF 86. This is achieved by altering the ratio of the resistors R_{21} and R_{23}. (The sum of these two resistors should be maintained at 100 kΩ). The values of 18 kΩ and 82 kΩ shown in Fig.44 are appropriate for use with the 10 W amplifier. With the 20 W amplifier, the full output is taken directly from the anode of the EF 86.

The smoothing components R_{31} and C_{17} shown in the H.T. line should be included in the main amplifier rather than in this pre-amplifier. The H.T. current drawn by the pre-amplifier is 3 mA at 300 V.

PICK-UP POSITIONS

The equalization curves for both pick-up positions are given in Fig.45. It is seen that equalization is adapted to the present recording curves as given in Fig.43. The magneto-dynamic pick-up

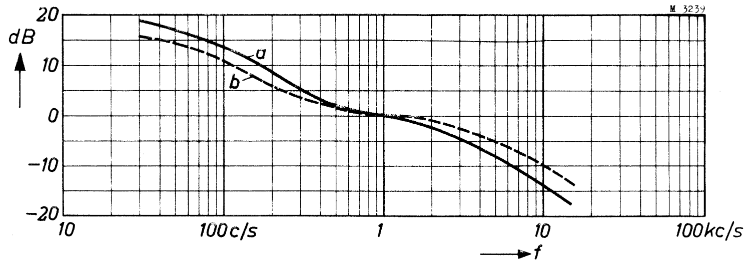


Fig.45. Equalization (a) for microgroove recordings and (b) for 78 r.p.m. recordings.

input channel is also suitable for pick-ups of the moving-coil type with higher outputs, provided a larger value of series resistor R_{11} is included. The change in sensitivity between 78 r.p.m. and microgroove positions is achieved partly by different amounts of feedback but mostly by insertion of the attenuating resistor R_{13}.

The low input impedance of the crystal pick-up position (100 kΩ) causes attenuation in the low frequencies with the result that the characteristics become similar to those of a magnetic pick-up. Under these conditions the same feedback elements can be used for both channels, that is, when low and medium output crystal pick-ups

*) The impedance of the input channels includes the grid impedance of the EF 86 modified by the feedback components, as well as the impedance of the input network.

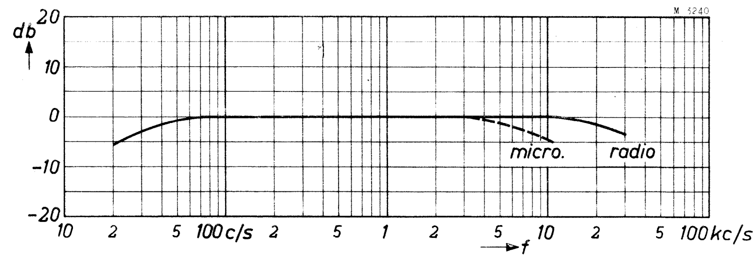
are used. If, however, the pick-up is not suitable for this form of loading or gives too much output, it may be plugged into the auxiliary socket, whose function will be described later.

The hum and noise are 55 dB below 10 W for the microgroove and 57 dB below 10 W for the 78 r.p.m. position, for both input sockets. In both cases the input impedance is approximately 100 kΩ. For magnetic pick-ups the sensitivities at 1 kc/s are 3 mV and 9 mV for microgroove and 78 r.p.m. respectively, and 50 mV, respectively 150 mV for crystal pick-ups, in both cases for full output of the amplifier.

MICROPHONE POSITION

The microphone input channel has been designed for use with crystal microphones and magnetic types with input transformer. The sensitivity is 6 mV and the input impedance 1 MΩ. The hum and noise levels are 44 dB below 10 W. The frequency response is given in Fig.46.

Fig. 46. Frequency response for the radio-input and auxiliary terminals (full line), and for the microphone terminal (broken line).



TAPE PLAYBACK POSITION

The playback characteristic for tape is shown in Fig.47. It follows the C.C.I.R. characteristic down to 100 c/s, but below that frequency somewhat less boost is employed. The input impedance on this channel is approximately 80 kΩ, the sensitivity 3 mV at 5 kc/s. The hum and noise levels are 52 dB below 10 W.

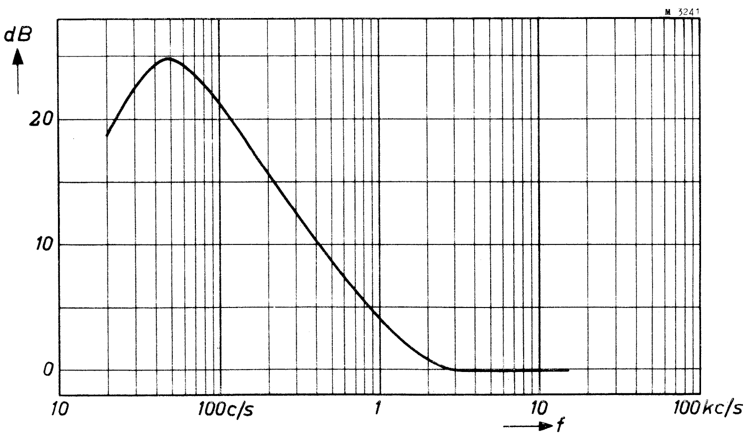


Fig. 47. Tape play-back characteristics of the two-tube pre-amplifier.

This channel is intended for the replay of pre-recorded tapes, when using high-impedance play-back heads. The curve adopted will then result in very good performance. If the sensitivity is insufficient for a certain head, it may be increased by decreasing the value of R_9 (56 kΩ) until the required sensitivity is reached.

RADIO INPUT POSITION

The frequency characteristic of the radio input channel is plotted in Fig.46. It ensures very good reproduction of FM broadcast, when

de-emphasis is applied in the tuner. For AM broadcast it may be necessary to attenuate the high frequencies at other than local transmitters.
The sensitivity of this channel of 250 mV at an input impedance of 2 MΩ is sufficient to meet most requirements. Other values, however, can easily be obtained by changing the feedback resistor R_1 (330 kΩ) and the series resistor R_8 (2.2 MΩ). If the input impedance is too high, it can be decreased by connecting a resistor of the required value between the input contact and earth.

AUXILIARY INPUT POSITION

It follows from the circuit diagram that the auxiliary input position is identical to the radio input. With these characteristics it is suitable for use with tape amplifiers or alternatively with high-output crystal pick-ups. If it is required to work at a lower input voltage, R_7 may be reduced in value from 2.2 MΩ to 1 MΩ.

TONE AND VOLUME CONTROLS

The tone control characteristics are shown in Fig.48. Both the bass and the treble control provide an adequate range of control for most uses.

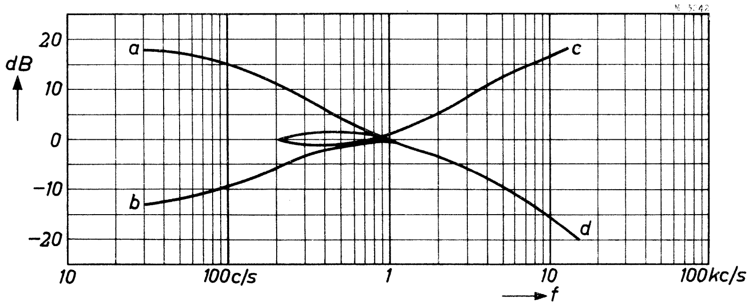


Fig. 48. Tone control characteristics:
(a) Bass maximum, treble flat
(b) Bass minimum, treble flat
(c) Treble maximum, bass flat
(d) Treble minimum, bass flat

The values of all the control potentiometers are relatively low, the volume control included (0.25 MΩ), so that the capacitance due to long co-axial leads for connecting the pre-amplifier to the power amplifier should have minimum effect on the high-frequency response.

OUTPUT VOLTAGE

The values of R_{21} and R_{23} are chosen for an output signal sufficiently high for driving the 10 W amplifiers described in Chapter II. The tone and volume controls are then left out the power amplifier and are replaced by a 1 MΩ grid resistor. The sensitivity of the power amplifier is then 40 mV for 10 W output, which conforms with the quoted sensitivity figures of the pre-amplifier described here.

When the pre-amplifier is to be used with the 20 W amplifier with two EL 34 output pentodes, the output voltage for 20 W should be 250 mV for full drive. This is obtained by replacing R_{21} and R_{23} by one single 100 kΩ resistor, and connecting C_{12} to the anode of the second EF 86.
If still higher output voltages are required at equal sensitivities, e.g. for use with the 4 x EL 84 20 W amplifier, the values of the series resistors in the input circuits may be reduced. It should be borne in mind, however, that the input impedances are reduced simultaneously.

TWO-TUBE PRE-AMPLIFIER FOR USE WITH MAGNETO-DYNAMIC PICK-UPS

This pre-amplifier has been designed for use with magneto-dynamic pick-ups of all types. It is particularly suitable to be used in combination with the single-ended push-pull amplifier described on page 45, with which it may be built on the same chassis provided it is screened from the power amplifier. The circuit can, of course, also drive other power amplifiers.

The input tube is used in a frequency-dependent feedback circuit which may be switched either for equalization of micro-groove recordings with S.A.E. or R.I.A.A. characteristics, for recordings with Decca f.f.r.r. characteristics, or for old 78 r.p.m. recordings. In the fourth switch position the first stage is disconnected from the circuit and a second input socket is connected to the top of the volume control in the grid circuit of the second stage. This socket may be used for radio, equalized tape, etc.

The pre-amplifier is provided with two volume controls; one of them is built-in and pre-set, the other, which is of the physiological type, is for manual operation. Ample bass and treble control has been provided, so that the amplifier may be adapted to various input circuits, loudspeaker combinations and auditorium conditions.

SUMMARY OF PERFORMANCE

Sensitivity

Terminal I	at 1 kc/s	gain V_o/V_i	= 100
Terminal II	at 1 kc/s	gain V_o/V_i	= 8.

Distortion

At an output voltage of 0.5 V_{rms} and maximum input voltage (V_i = 200 mV_{rms}) the distortion d_{tot} is:

Frequency f = 1 kc/s, $d_{tot} < 0.02$ %

f = 80 kc/s, $d_{tot} < 0.06$ %.

Hum and noise

Hum and noise, measured at the output of the single-ended push-pull power amplifier in combination with the pre-amplifier, proved to be lower than -60 dB at full output and about 20 mV with the volume control at its minimum position.

Tone control

Bass compared to 1 kc/s level at 25 c/s	max. cut	-18 dB
	max. lift	+12 dB.
Treble compared to 1 kc/s level at 15 kc/s	max. cut	-14 dB
	max. lift	+ 8 dB.

CIRCUIT DESCRIPTION

Only two tubes are employed in this pre-amplifier (see Fig.49), viz. a low-noise A.F. pentode type EF 86, and a high-mu double triode ECC 83. The first tube is used with a conventional load,

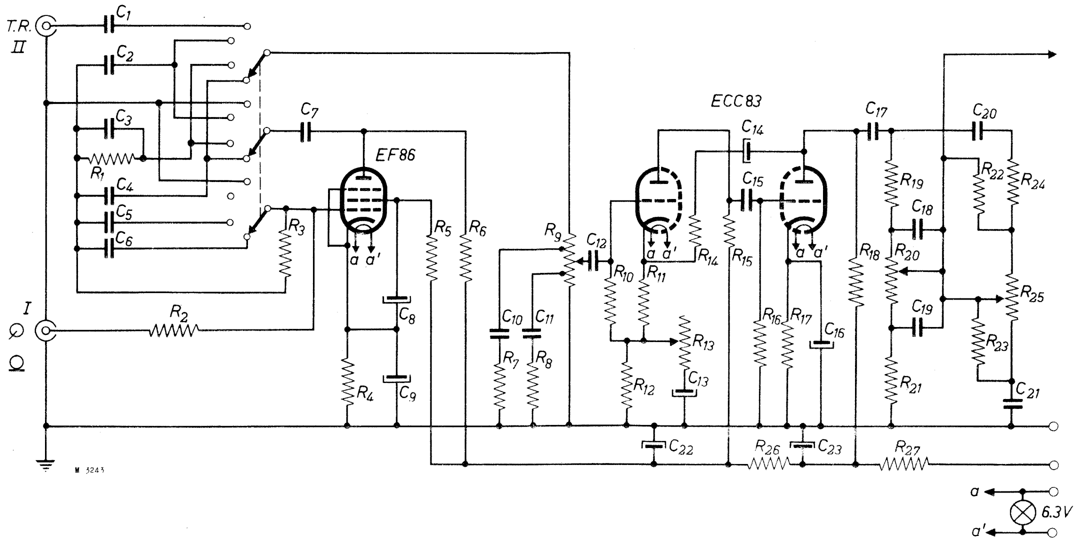


Fig.49. Circuit diagram of the two-tube pre-amplifier with EF 86 and ECC 83.

R_1 = 8.2 MΩ ± 10 %, ¼ W
R_2 = 68 kΩ ± 10 %, ¼ W
R_3 = 680 kΩ ± 10 %, ¼ W
R_4 = 2.2 kΩ ± 10 %, ½ W
R_5 = 390 kΩ ± 10 %, high stab. ½ W
R_6 = 100 kΩ ± 10 %, high stab. ½ W
R_7 = 47 kΩ ± 10 %, ¼ W
R_8 = 10 kΩ ± 10 %, ¼ W
R_9 = 800 + 100 + 100 kΩ, log.
R_{10} = 1 MΩ ± 10 %, ¼ W
R_{11} = 2.2 kΩ ± 10 %, ½ W
R_{12} = 33 kΩ ± 10 %, ½ W
R_{13} = 100 kΩ, lin.
R_{14} = 120 kΩ ± 10 %, ¼ W
R_{15} = 220 kΩ ± 10 %, ½ W
R_{16} = 1 MΩ ± 10 %, ¼ W
R_{17} = 1 kΩ ± 10 %, ½ W
R_{18} = 100 kΩ ± 10 %, ½ W
R_{19} = 100 kΩ ± 10 %, ¼ W
R_{20} = 1 MΩ, log.
R_{21} = 22 kΩ ± 10 %, ¼ W
R_{22} = 470 kΩ ± 10 %, ¼ W
R_{23} = 100 kΩ ± 10 %, ¼ W
R_{24} = 10 kΩ ± 10 %, ¼ W
R_{25} = 1 MΩ, log.

R_{26} = 47 kΩ ± 10 %, ½ W
R_{27} = 22 kΩ ± 10 %, ½ W
C_1 = 22 000 pF
C_2 = 470 pF
C_3 = 330 pF
C_4 = 330 pF
C_5 = 100 pF
C_6 = 120 pF
C_7 = 10 000 pF
C_8 = 8 μF, 300 V wkg.
C_9 = 100 μF, 3 V wkg.
C_{10} = 5600 pF
C_{11} = 33 000 pF
C_{12} = 0.1 μF, 125 V wkg.
C_{13} = 10 μF, 70 V wkg.
C_{14} = 8 μF, 300 V wkg.
C_{15} = 10 000 μF
C_{16} = 100 μF, 3 V wkg.
C_{17} = 0.1 μF, 400 V wkg.
C_{18} = 4700 pF
C_{19} = 22 000 pF
C_{20} = 390 pF
C_{21} = 4700 pF
C_{22}, C_{23} } 2 x 16 μF, 300 V wkg.

screen-grid dropping resistor and cathode biasing in a frequency-dependent feedback circuit which allows for three different equalization characteristics. A three-pole four-position selector switch offers successively the following possibilities.

In position *1*, equalization is provided for micro-groove recordings according to R.I.A.A. or A.E.S. characteristics. The output signal of the tube is fed via the coupling capacitor C_7 via C_4 to the top of R_3. The latter resistor is shunted by C_6.

In position *2*, equalization is obtained for Decca ffrr recordings. The capacitor C_4 is now replaced by C_3 in parallel with R_1, and R_3 is shunted by C_5 of 100 pF.

Position 3 equalizes old 78 r.p.m. recordings, the characteristics of which drop about 5 dB/octave below 500 c/s. For this purpose only C_2 is connected in series with R_3.

In position 4 the EF 86 is disconnected from the volume control, and the input socket II is connected instead. This socket may be used for connecting an AM/FM tuner unit to the amplifier, but may also serve for the use with crystal pick-ups or equalized tape.

The two sections of the ECC 83 double triode are connected in cascade, heavy feedback being applied from the anode of the second section to the cathode of the first via C_{14} and R_{14}. The cathode of the first section is earthed via the unbypassed cathode resistor R_{11}, which provides also the grid bias, and via R_{12}. The latter resistor is shunted by the pre-set volume control R_{13} in series with an electrolytic capacitor.

The manual volume control R_9 has a total resistance of 1 MΩ with tappings at 0.1 and 0.2 MΩ from the earthed end. These tappings are connected to earth via high-pass filters consisting of C_{11} (33 000 pF) and R_8 (10 kΩ) in series, and of C_{10} (5600 pF) in series with R_7 (47 kΩ) respectively. The result is that with the volume control R_9 set in a lower position, the low frequencies are less attenuated than the midfrequency level. Hence the frequency response at different settings of the volume control changes with the attenuation. The relative increase in bass and treble levels, as shown in Fig.50, is opposite to the aural sensitivity curves,

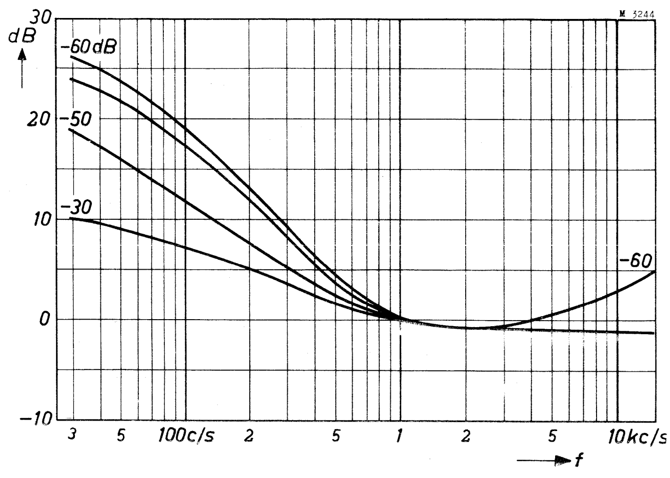


Fig.50. Attenuation curves of the physiological volume control.

showing reduced sensitivity at a decreased sound level (Flecher-Munson curves). However, since much depends on the room in which the amplifier is installed, the average sound level is pre-adjusted by means of R_{13}, the volume being controlled by R_9, so that at low sound levels resetting of the tone controls is superfluous.

The total feedback in the ECC 83 circuits amounts to about 38 dB, that in the anode circuit of the EF 86 to about 29 dB. The ECC 83 is followed by a conventional tone-control circuit which requires no further explaination; the curves are shown in Fig.51.

The power supply of this pre-amplifier requires about 4 mA at 300 V. This can easily be delivered by the H.T. supply unit of the power amplifier; the dropping resistor of 5 kΩ should then be mounted in the power amplifier, together with the usual by-pass capacitor.

The equalization curves are shown in Fig.52, and the playback curves measured with Decca test record LXT 2695 in Fig.53. These curves clearly show that good equalization has been obtained.

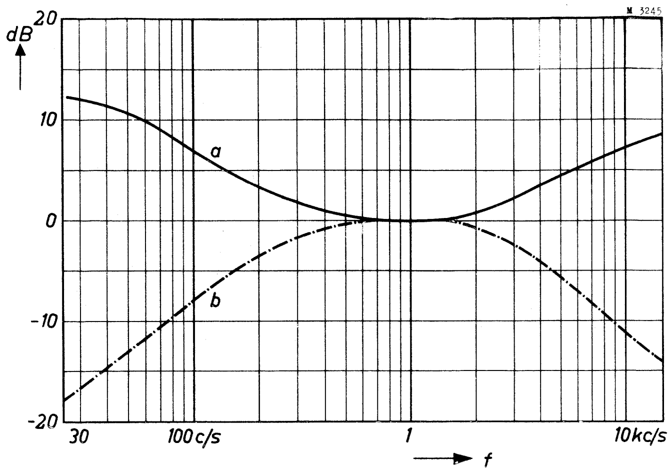


Fig.51.Tone control characteristics of the two-tube pre-amplifier with EF 86 and ECC 83.(a):bass maximum and treble maximum; (b):bass minimum and treble minimum.

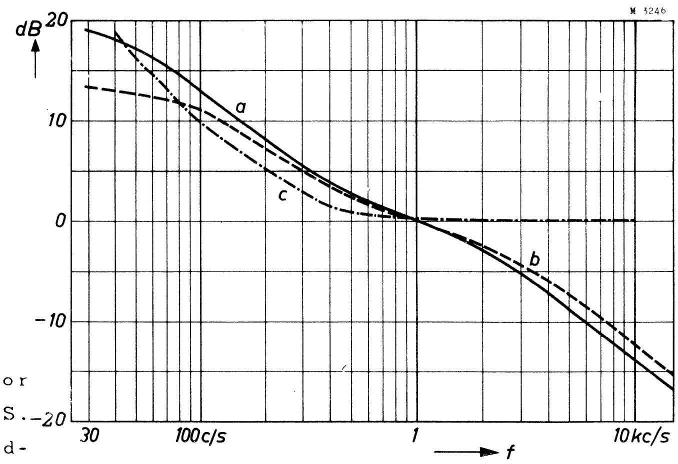


Fig.52. Equalization curves for several recordings. (a): A.E.S. and R.I.A.A.microgroove recordings; (b): Decca ffrr recordings; (c): old 78 r.p.m. recordings.

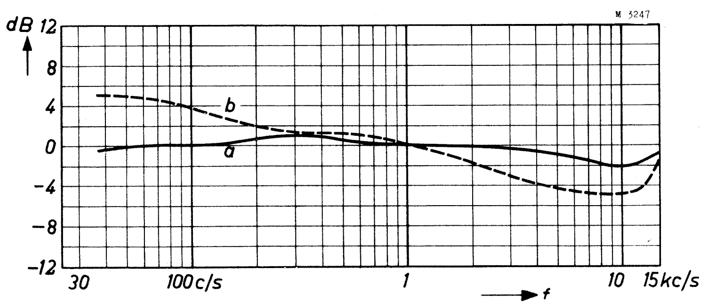


Fig.53. Play-back curves measured on the two-tube pre-amplifier with EF 86 and ECC 83 in combination with the 10 W single-ended push-pull amplifier.

(a): equalized according to the A.E.S. and R.I.A.A. microgroove recording characteristics;

(b): played back with Decca test record LXT 2659

THREE-TUBE PRE-AMPLIFIER

The pre-amplifier described below has been designed principally for use with amplifiers built to the 20, 10 or 3 W circuits. It can, however, be used with any power amplifier which does not require an input signal greater than 250 mV for full output.

Facilities are provided for using magnetic and crystal pick-ups, tape-recorder playback heads and radio tuner units. An auxiliary input position is provided for ancillary equipment such as tape amplifiers. An additional output position, which is independent of the tone controls, is also provided, enabling programmes to be taken to a tape amplifier while they are being fed from the normal output position to a power amplifier.

All the input sockets in the circuit are connected to one switch which selects any one channel and short-circuits the unused sockets to earth. This arrangement reduces considerably the amount of "break-through" between channels. For the sake of clearness this arrangement has not been drawn in the circuit diagram. The equalization for disc recordings conforms to the latest R.I.A.A. characteristics which have been adopted by most of the major recording companies. The tape playback characteristic is intended for use with high-impedance heads when replaying pre-recorded tapes at a speed of 19 cm per second.

The tone controls used cover a wide range of frequency and provide boost and cut for high and low frequencies. Switched high and low band-pass filters are included in the circuit so that unwanted signals such as rumble and record scratch can be eliminated.

SPECIFICATION

Tubes:	2 x EF 86 low-noise amplifier pentode		
	1 x ECC 83 high-mu double triode		
Sensitivity:	crystal pick-up	micro-groove	150 mV
		78 r.p.m.	270 mV
	magneto-dynamic pick-up	micro-groove	7 mV
		78 r.p.m.	12 mV
	tape play-back:	2.5 mV at 5 kc/s	
	radio input:	250 mV	
	auxiliary:	250 mV	
Hum and noise:	pick-up positions:	micro-groove	53 dB below 20 W
		78 r.p.m.	58 dB below 20 W
	tape play-back:	47 dB below 20 W	
	radio input:	63 dB below 20 W	
	auxiliary input:	63 dB below 20 W	
Input impedance:	pick-up positions:	100 kΩ	
	tape play-back:	80 kΩ (approx.)	
	radio input:	1 MΩ	
	auxiliary:	1 MΩ	
Output voltage:	nominal:	250 mV	
	auxiliary recording output:	300 mV	

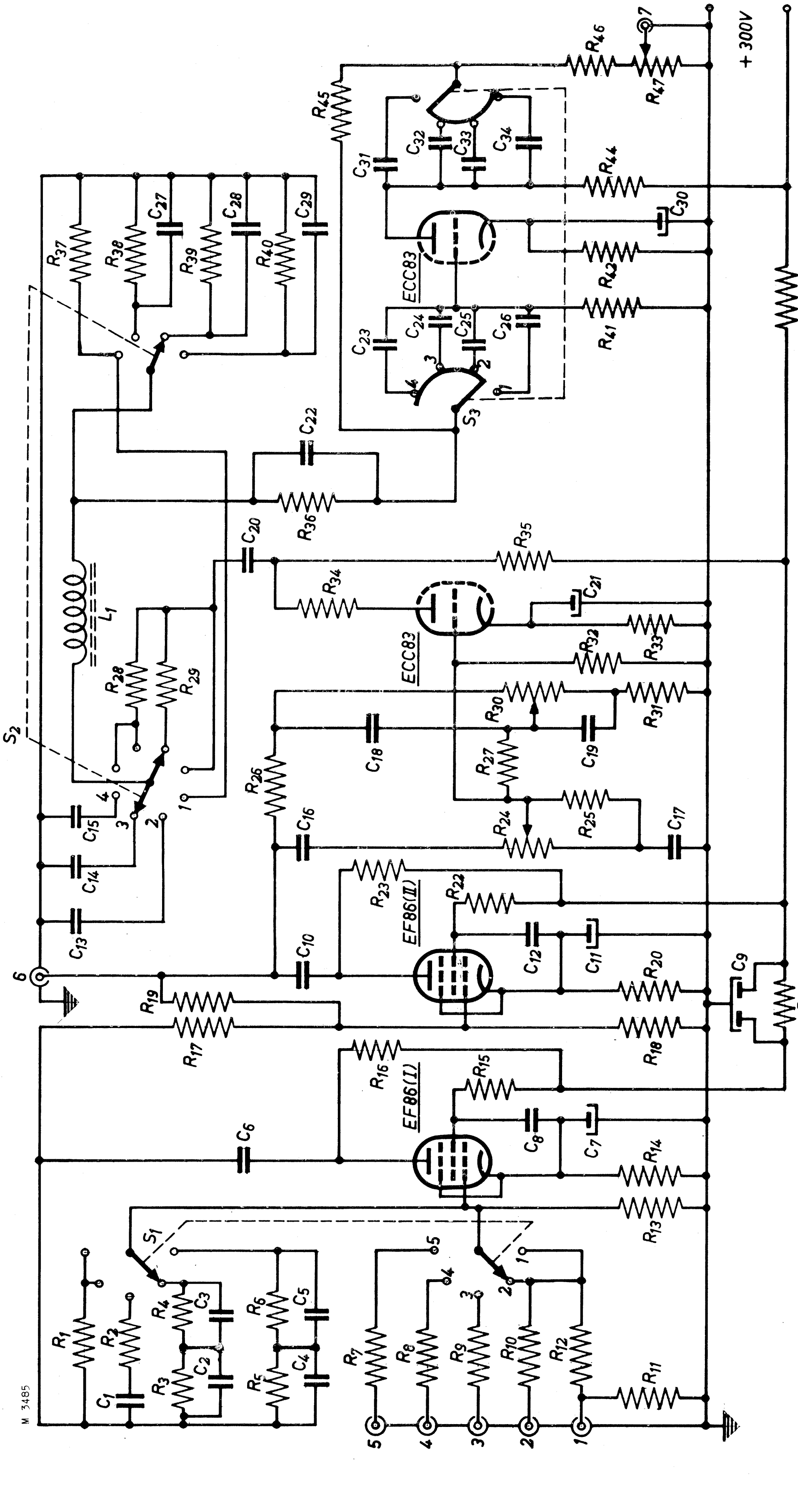


Fig.54 Circuit diagram of the three-tube pre-amplifier.

PARTS LIST

R_1	= 100 kΩ	R_{25}	= 47 kΩ
R_2	= 390 kΩ	R_{26}	= 68 kΩ
R_3	= 8.2 MΩ	R_{27}	= 39 kΩ
R_4	= 270 kΩ	R_{28}	= 22 kΩ
R_5	= 2.2 MΩ	R_{29}	= 10 kΩ
R_6	= 180 kΩ	R_{30}	= 250 kΩ log.
R_7	= 1 MΩ	R_{31}	= 6.8 kΩ
R_8	= 1 MΩ	R_{32}	= 270 kΩ
R_9	= 56 kΩ	R_{33}	= 1.2 kΩ
R_{10}	= 100 kΩ	R_{34}	= 22 kΩ
R_{11}	= 100 kΩ	R_{35}	= 33 kΩ
R_{12}	= 2.2 MΩ	R_{36}	= 270 kΩ
R_{13}	= 100 kΩ	R_{37}	= 56 kΩ
R_{14}	= 3.9 kΩ	R_{38}	= 56 kΩ
R_{15}	= 1.5 MΩ*)	R_{39}	= 47 kΩ
R_{16}	= 270 kΩ*)	R_{40}	= 33 kΩ
R_{17}	= 220 kΩ	R_{41}	= 220 kΩ
R_{18}	= 1 MΩ	R_{42}	= 1.5 kΩ
R_{19}	= 3.9 MΩ	R_{43}	= 12 kΩ
R_{20}	= 1.2 kΩ	R_{44}	= 47 kΩ
R_{21}	= 33 kΩ	R_{45}	= 1.5 MΩ
R_{22}	= 470 kΩ*)	R_{46}	= 220 kΩ
R_{23}	= 100 kΩ*)	R_{47}	= 50 kΩ log.
R_{24}	= 250 kΩ log.		

*) These resistors should be of high stability.

C_1	= 330 pF	C_{18}	= 2200 pF
C_2	= 820 pF	C_{19}	= 0.02 μF
C_3	= 330 pF	C_{20}	= 0.25 μF
C_4	= 2700 pF	C_{21}	= 50 μF
C_5	= 330 pF	C_{22}	= 33 pF
C_6	= 0.1 μF	C_{23}	= 820 pF
C_7	= 50 μF	C_{24}	= 820 pF
C_8	= 0.1 μF	C_{25}	= 1500 pF
C_9	= 2x8 μF	C_{26}	= 3300 pF
C_{10}	= 0.1 μF	C_{27}	= 390 pF
C_{11}	= 50 μF	C_{28}	= 820 pF
C_{12}	= 0.1 μF	C_{29}	= 1800 pF
C_{13}	= 390 pF	C_{30}	= 50 μF
C_{14}	= 820 pF	C_{31}	= 0.01 μF
C_{15}	= 1800 pF	C_{32}	= 4700 pF
C_{16}	= 560 pF	C_{33}	= 2700 pF
C_{17}	= 8200 pF	C_{34}	= 2700 pF
		L_1	= 1.2 H

Switch positions

S_1	1. 78 r.p.m.	S_2	1. flat	S_3	1. 20 c/s
	2. microgroove		2. 9 kc/s		2. 40 c/s
	3. tape		3. 7 kc/s		3. 80 c/s
	4. radio		4. 5 kc/s		4. 160 c/s
	5. auxiliary				

Input and output terminals.

1. Crystal pick-up
2. Magneto-dynamic pick-up
3. Tape lead-in
4. Radio
5. Auxiliary
6. Programme recording out
7. Output

Distortion:	at rated output:	<0.1 %
	at ten times rated output:	0.65 %
	(measured with the 20 W amplifier)	
Tone control:	bass boost maximum	+ 18 dB at 50 c/s
	bass cut maximum	- 10 dB at 50 c/s
	treble boost maximum	+ 14 dB at 10 kc/s
	treble cut maximum	- 18 dB at 10 kc/s
	all referred to the 1 kc/s level.	

CIRCUIT DESCRIPTION

The pre-amplifier circuit drawn in Fig.54 is made up of three stages. The first and second stages both use a low-noise high-gain pentode, type EF 86, and the output stage uses a double triode type ECC 83.

All the equalization takes place in the first stage and is achieved by means of frequency-selective feedback between the anode and grid of the first EF 86. This arrangement has been chosen so that the grid-circuit impedance of the first stage should be low. A low impedance at this grid lessens hum pick-up and reduces the effect of plugging-in external low-impedance circuits. Furthermore, the arrangement also results in low gain in the first stage. Hence, Miller effect between the anode and grid of the first EF 86, which can be troublesome when high values of resistance are used in series with the grid, is reduced. The arrangement also means that the other stages of the pre-amplifier are independent of the equalization characteristics.

Series resistors are used in the input circuit so that the sensitivity and impedance of any channel can be adjusted accurately. The component values given in Fig.54 are intended for sources encountered most frequently, but the sensitivity and impedance[1]) of each channel can be altered by changing the value of the appropriate series resistor.

The second stage of the pre-amplifier has a linear frequency-response characteristic and its gain is reduced by a small amount of negative feedback taken to the control grid by way of the resistor R_{19}. The output of this stage is taken from the anode to the tone-control network, and also to the programme-recording output socket. Because of the negative feedback, the tone controls have very little effect on the frequency response at this anode.

The output stage is made up of the band-pass filter circuits. The low-pass filter is situated between the two sections of the ECC 83, and is arranged on the switch S_2. It incorporates a potcore inductor, type *D* 18/12 in a π-type network. The high pass filter is arranged on the switch S_3 and consists of two RC networks in a feedback loop around the second triode section of the tube.

The level of output signal from the pre-amplifier will have to be adjusted if it is to be used with power amplifiers having different sensitivities from the 20-watt circuit. The signal should be attenuated by using a simple potential divider as shown in Fig.55.

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[1]) The impedance of the input channels includes the grid impedance of the EF 86 modified by the feedback components as well as the impedance of the input network.

The values of resistance in Fig.55*a* are for an attenuator suitable for the 5-tube, 10-watt amplifier, and those in Fig.55*b* are suitable for the 3-tube, 3-watt amplifier, both these amplifiers having the volume and tone controls disconnected.

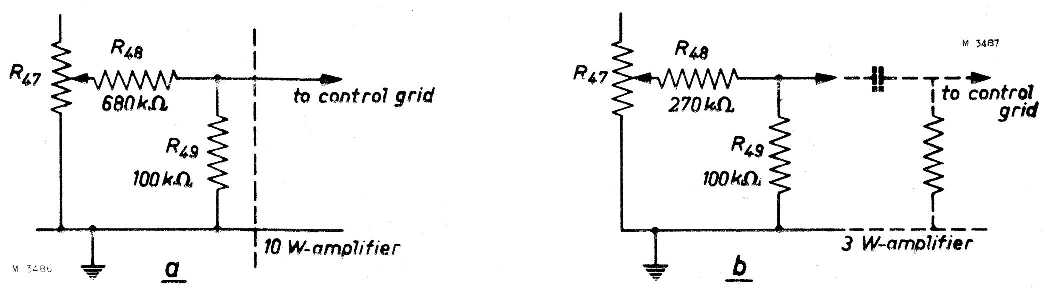

Fig.55. Output attenuating networks.

The H.T. voltage for the unit should be adequately smoothed, and the method of achieving this will depend to some extent on the power supply. The H.T. current drawn by the pre-amplifier is approximately 6 mA at 300 V, and the L.T. current is 0.7 A at 6.3 V.

PERFORMANCE

The values for hum and noise in the pre-amplifier which are quoted below for each input channel have been measured with the pre-amplifier connected to a 20-watt power amplifier. The measurements were made at the output socket of the power amplifier when the input terminals of the pre-amplifier were open-circuited. The frequency-response curves were also obtained with this combination of pre-amplifier and power amplifier.

The sensitivity figures given below are for an output of 250 mV from the pre-amplifier. The total harmonic distortion of the unit is less than 0.1 % at the rated output, and at an output of ten times this, the distortion is only 0.65 %. A rapid increase in distortion does not occur until the pre-amplifier is considerably overloaded.

PICK-UP INPUT CHANNELS

Equalization curves for the magnetic and the crystal pick-up channels are drawn in Fig.56.

The difference in sensitivities between the positions for long-playing and 78 r.p.m. records is achieved with the different levels of feedback provided at positions 2 and 1 of the switch S_1. The change is necessary because standard discs are recorded at a higher level than micro-groove discs.

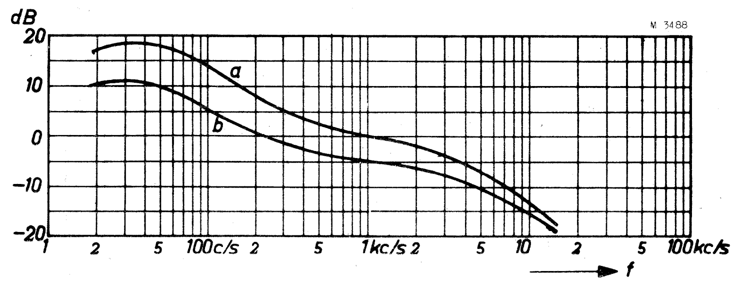

Fig.56. Equalization characteristic for pick-up input;
a = microgroove, b = 78 r.p.m.

Magneto-dynamic pick-up

Input impedance:	100 kΩ
Sensitivity at 1 kc/s	
(*a*) micro-groove	7 mV
(*b*) 78 r.p.m.	12 mV
Hum and noise	
(*a*) micro-groove	53 dB below 20 W
(*b*) 78 r.p.m.	58 dB below 20 W

This input channel is most suitable for pick-ups of the variable-reluctance type, but moving-coil types which have higher outputs can be used if a larger value of series resistance R_{10} is included.

Crystal Pick-up

Input impedance:	100 kΩ
Sensitivity at 1 kc/s	
(*a*) micro-groove:	150 mV
(*b*) 78 r.p.m.:	270 mV
Hum and Noise	
(*a*) micro-groove:	53 dB below 20 W
(*b*) 78 r.p.m.:	58 dB below 20 W

Low- and medium-output crystal pick-ups can be used for this input channel. The input is loaded with the 100 kΩ resistor R_{11} (causing bass loss) in order that its characteristic shall approximate to that of a magnetic cartridge, and to allow the same feedback network to be used. This produces the best compromise with most types of pick-up. However, if the pick-up is not suitable for this form of loading, or if its output is too high, then it can be connected to the auxiliary input socket, the function of which is discussed below. With this channel, the output is fed into a 1 MΩ load which compensates automatically for the recording characteristic.

Tape Playback input Channel

Input Impedance:	80 kΩ (approx.)
Sensitivity at 5 kc/s:	2.5 mV
Hum and Noise:	47 dB below 20 W

The equalization characteristic used in this channel is shown in Fig.57. The channel is intended for replaying pre-recorded tapes using medium- or high-impedance heads and the frequency-response

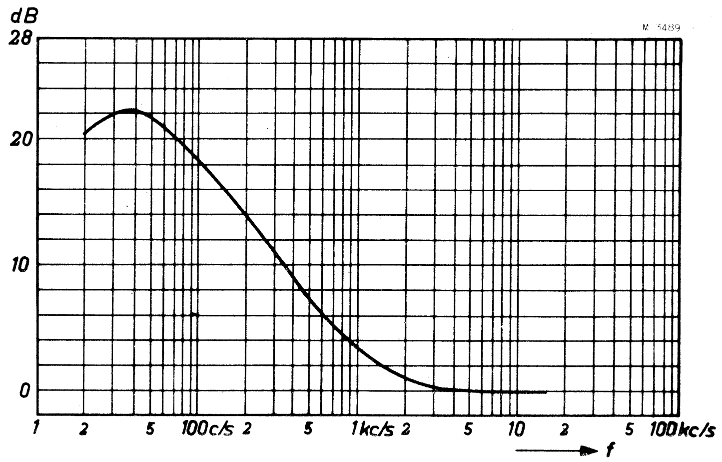

Fig.57. Equalization characteristic for tape play-back.

curve for playback with the E.M.I. test tape T.B.T.1. is drawn in Fig.58. Variation of the sensitivity can be achieved by altering the value of the resistance R_9.

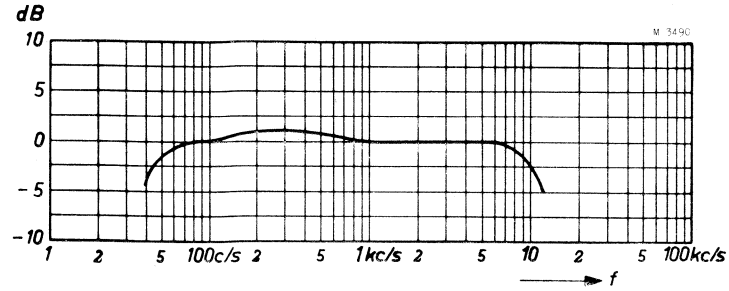

Fig.58. Tape play-back frequency response characteristic using E.M.I. test tape T.B.T.1.

RADIO INPUT CHANNEL

Input impedance:	1 MΩ
Sensitivity:	250 mV
Hum and Noise:	63 dB below 20 W

The frequency-response characteristic of this channel is drawn in Fig.59. With the above values of impedance and sensitivity the channel should meet most requirements. The impedance of the channel can be reduced by connecting a resistor between the input end of R_8 and the chassis.

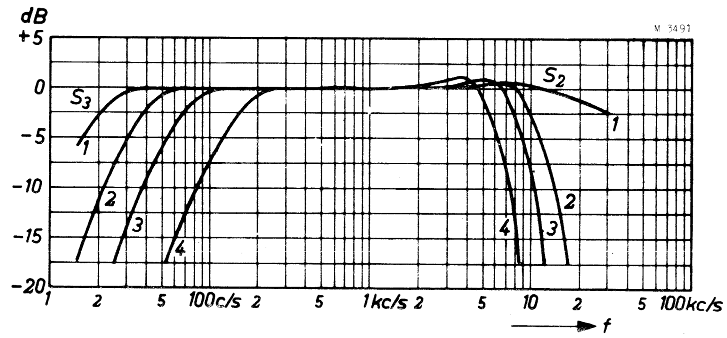

Fig.59 Filter characteristics and flat-position frequency response characteristic.

AUXILIARY INPUT CHANNEL

It can be seen from the circuit of Fig.54 that the auxiliary channel is identical with the radio input channel. With these characteristics it is suitable for use with high-output crystal pick-ups or with a tape playback pre-amplifier. In the prototype equipment, the auxiliary input is taken to a jack socket at the front of the chassis. This is to facilitate connection of a portable tape recorder.

If it is desired to use a microphone, a microphone pre-amplifier or input mixing pre-amplifier can be connected to this socket. This gives greater flexibility of arrangement and also, by reducing the number of positions on the selector switch, allows the short-circuiting wafer to be incorporated.

AUXILIARY OUTPUT POSITION

The output from the anode of the second EF 86 is available at this auxiliary position enabling a record of programme material to be made with tape equipment. This additional output is taken to a jack socket at the front of the chassis. Excellent recordings can be made even when the input signal is derived from a low-output magnetic pick-up.

An output of 300 mV is available at this output socket, and the impedance is low. Recording equipment plugged into this socket should not have an input impedance less than 500 kΩ. The tone controls and the filter networks are inoperative when this output is used.

TONE CONTROLS

The treble and bass tone-control characteristics are shown in Fig.60. These indicate that an adequate measure of control is provided in the unit for most applications.

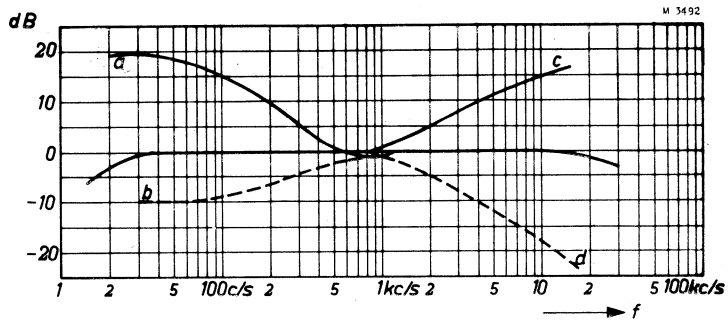

Fig.60. Tone control characteristics.

a = bass maximum, treble flat;

b = bass minimum, treble flat;

c = treble maximum, bass flat

d = treble minimum, bass flat

FILTER NETWORKS

The characteristic (Fig.59) of the low-pass filter has a slope of approximately 20 dB per octave. The components of the network are arranged for operation at 5, 7 and 9 kc/s in positions 4, 3 and 2 respectively of the switch S_2 in Fig.54. Position 1 of the switch gives a flat characteristic.

The characteristic (Fig.59) of the high-pass filter has a slope of approximately 12 dB per octave. Operation is at 160, 80 and 40 c/s respectively at positions 4, 3 and 2 of the switch S_3. Position 1 of the switch cuts at 20 c/s. This is considered preferable to allowing the response to continue to lower frequencies because this could result in the amplifier or loudspeaker being overloaded.

FOUR-CHANNEL INPUT MIXING AMPLIFIER

It may often be desirable to use several signal sources with an amplifying unit which has provision for only one input signal. The circuit described below is capable of handling four input signals and of supplying a mixed output voltage suitable for driving a single-input amplifier. Two of the input channels of the mixer are suitable for microphone signals, a third for a radio or equalized tape input, and a fourth for pick-up signals.

The circuit as it stands provides an output voltage of 40 mV and was intended to be used with the 10-W amplifier in which the tone-control network is disconnected. With a simple modification to the output stage, the mixer will provide an output voltage of up to 800 mV.

The equipment has not been designed to professional standards. The number of components has been limited to the minimum compatible with a performance which will be satisfactory in conjunction with a high-quality audio amplifier of the 10-W type. It is felt that this input mixing amplifier will therefore be of interest principally to the home constructor.

SUMMARY OF PERFORMANCE

Output voltage:	Low-gain connection:	40 mV
	High-gain connection:	800 mV

Sensitivity (for full output in either low or high-gain connection):

	Microphone input:	3 mV
	Radio input:	230 mV
	Pick-up input:	250 mV

Frequency response:	For the microphone channels flat to within ± 3 dB (relative to level at 1 kc/s) from 20 c/s to 20 kc/s. For the radio or pick-up channels flat to within ± 2 dB (relative to level at 1 kc/s) from 15 c/s to 20 kc/s.
Hum and noise level:	50 dB below full output.

CIRCUIT DESCRIPTION

Fig. 61 shows the circuit diagram. It can be seen that both microphone input stages of the input mixing circuit are identical. Each is equipped with the low-noise pentode, type EF 86, operating with grid current bias obtained by means of a high-valued grid resistor (R_1, R_6).

The internal impedance of a crystal microphone is predominantly capacitive, and the capacitance is of the order or 2000 pF. Therefore, to avoid loss in terminal voltage at low audio frequencies, the microphone should be connected to a high-impedance input stage. If a low value of resistance of 1.5 MΩ is chosen for R_1, for example, the combination of the series capacitive elements of the microphone, the grid circuit capacitance C_1 and the grid resistance

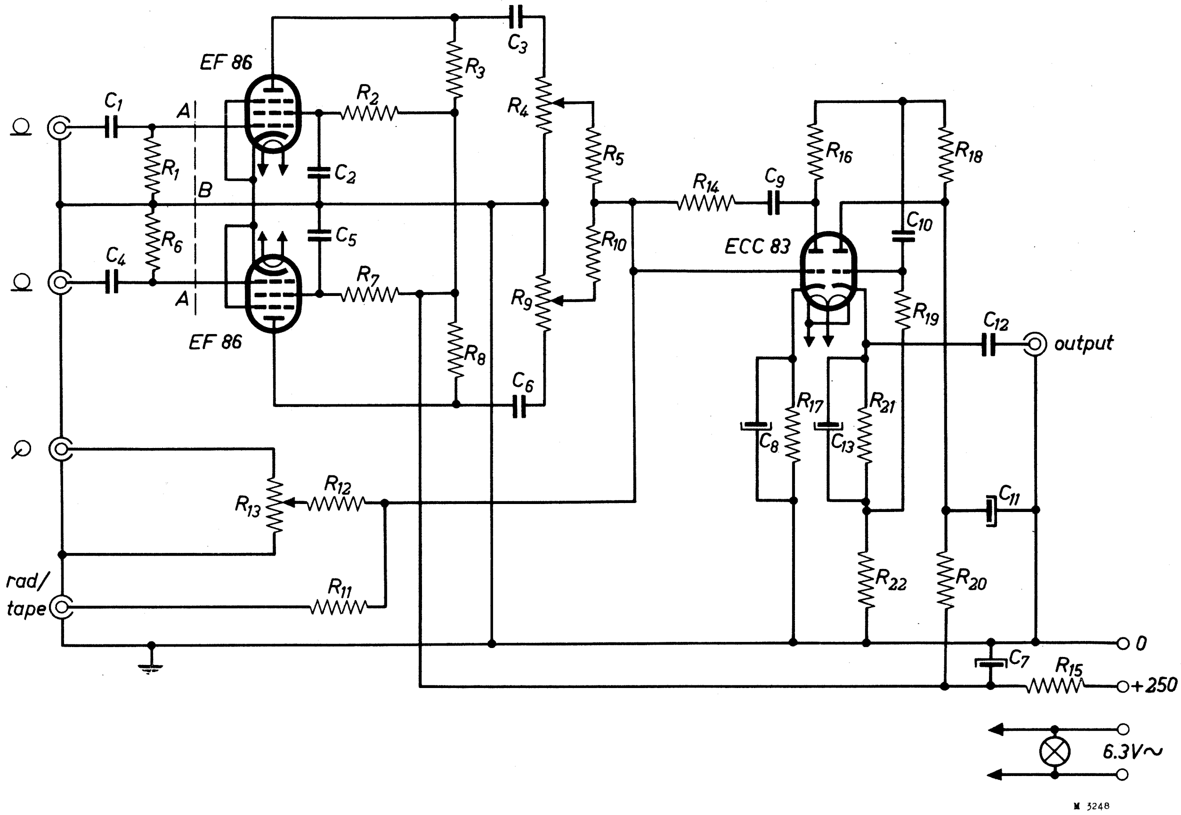

Fig. 61. Circuit diagram of the four-channel input mixing amplifier.

R_1 = 10 MΩ ± 10 %, high stab. ¼ W
R_2 = 390 kΩ ± 10 %, high stab. ¼ W
R_3 = 100 kΩ ± 10 %, high stab. ¼ W
R_4 = 500 kΩ, log.
R_5 = 470 kΩ ± 10 %, ¼ W
R_6 = 10 MΩ ± 10 %, high stab. ¼ W
R_7 = 390 kΩ ± 10 %, high stab. ¼ W
R_8 = 100 kΩ ± 10 %, high stab. ¼ W
R_9 = 500 kΩ, log.
R_{10} = 500 kΩ ± 10 %, ¼ W
R_{11} = 390 kΩ ± 10 %, ¼ W
R_{12} = 470 kΩ ± 10 %, ¼ W
R_{13} = 500 kΩ, log.
R_{14} = 1.5 MΩ ± 10 %, ¼ W
R_{15} = 22 kΩ ± 10 %, ½ W
R_{16} = 100 kΩ ± 10 %, ½ W *)
R_{17} = 2.2 kΩ ± 10 %, ½ W
R_{18} = 5.6 kΩ ± 10 %, ½ W

R_{19} = 1 MΩ ± 10 %, ¼ W
R_{20} = 27 kΩ ± 10 %, ½ W
R_{21} = 1.5 kΩ ± 10 %, ½ W
R_{22} = 47 kΩ ± 10 %, ½ W

C_1 = 0.05 μF, 250 V wkg.
C_2 = 0.25 μF, 250 V wkg.
C_3 = 0.1 μF, 250 V wkg.
C_4 = 0.05 μF, 250 V wkg.
C_5 = 0.25 μF, 250 V wkg.
C_6 = 0.1 μF, 250 V wkg.
C_7 = 16 μF, 350 V wkg.
C_8 = 50 μF, 12 V wkg.
C_9 = 0.05 μF, 250 V wkg.
C_{10} = 0.01 μF, 250 V wkg.
C_{11} = 16 μF, 350 V wkg.
C_{12} = 0.1 μF, 250 V wkg.
C_{13} = 50 μF, 12 V wkg.

*) This values may be altered when higher output voltages are required.

R_1 will result in a loss of about one third of the output voltage from the microphone at a frequency of 100 c/s. Consequently, a value of 10 MΩ has been selected for R_1 to provide the high impedance input for the crystal microphone and to prevent the loss in bass output voltage.

If it is required to use a low-impedance microphone such as a moving-coil or ribbon type, the mixer can be made suitable by using a step-up transformer in the grid circuit of either EF 86.

The arrangement for the low-impedance microphone is shown in Fig. 62. The connection marked A and B in Fig. 62 should replace those similarly marked in Fig. 61. The leads of the microphone transformer should be made as short as possible to avoid hum pick-up and loss of treble response.

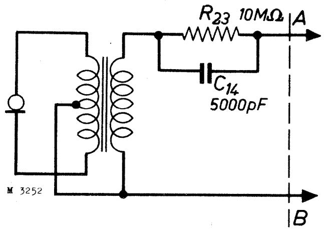

Fig. 62. Alternative arrangement for low-impedance microphones. This circuit replaces the part of the diagram in Fig.61 to the left of the line A-B.

The output from each microphone input stage is RC-coupled to the grid of one half of the high-mu double triode ECC 83. The radio and pick-up input stages are also connected to this grid by way of the resistors R_{11} and R_{12}, R_{13}. This half of the ECC 83 is arranged as a voltage amplifying stage.

The potentiometers R_4, R_9, and R_{13} serve for the adjustment of signal level and the mixing of the microphone and pick-up input channels. Adjustment for the fourth channel will be achieved by means of the gain control incorporated in the radio unit used at the radio input terminals. (If it is required, control of the radio input can be achieved by connecting a potentiometer to the radio input socket in the way that R_{13} is joined to the pick-up socket.) The value of the resistors R_5, R_{10}, R_{11}, and R_{12} have been chosen so that, in combination with the potentiometers R_4, R_9 and R_{13}, they prevent any interaction between the channels. Of course, these fixed resistors also ensure that the grid of the ECC 83 will not be short-circuited when any one of the potentiometers is set at a maximum.

The output stage of the mixer unit is suitable as it is shown in Fig. 61 for use with the 10 W Hi-Fi amplifier in which the tone control network is non-operative. The circuit has been arranged so that the sensitivity at the microphone inputs is 3 mV for an output voltage of 40 mV, and this is sufficient for crystal microphones. The sensitivities of the other input stages, for the same output voltage, are 230 mV and 250 mV for the radio and pick-up terminals respectively.

Feedback is taken from the anode to the grid of the first half of the ECC 83 by way of the components R_{14} and C_9. The purpose of this is to provide a low impedance at the grid and hence minimize the loss in response at treble frequencies caused by the Miller effect between the anode and the grid of the triode.

The output voltage is obtained from the second half of the ECC 83 which has been connected as a cathode follower. This type of connection provides a low output impedance, which has the approximate value of 600 Ω. Because of this low impedance, no attenuation of high notes will occur in consequence of cable capacitance if a long cable is required between the mixer unit and the main amplifier. But the input impedance of the main amplifier must be greater than 100 kΩ to ensure that the bass frequencies will not be attenuated by the coupling capacitor C_{12}.

PERFORMANCE

OUTPUT AND SENSITIVITY

The maximum output voltage of the mixer unit is 40 mV. This output is obtained with an input signal voltage of 3 mV at either microphone socket, 230 mV at the radio terminals or 250 mV at the pick-up terminals. If greater outputs from the unit are required to drive, for example, an amplifier incorporating a tone-control network, these can be achieved simply by adjusting the coupling between the anode of the first triode of the ECC 83 and the grid of the second. If the capacitor C_{10} is joined directly to the anode of the first triode, an output of 800 mV will be available. Intermediate values of output voltage can be achieved by altering the value of R_{16} and R_{18}. If for example, R_{16} and R_{18} were each 47 kΩ, the output voltage of the unit would be about 400 mV.

If the required output voltage had been obtained by attenuating the output voltage from the cathode load in the final stage, the low output impedance resulting from the cathode follower action would have been lost.

FREQUENCY RESPONSE

The response curve of the mixer unit, measured between microphone input and output terminals, is shown in Fig. 63. The curve is flat to within ± 3 dB, relative to the response level at 1 kc/s, from

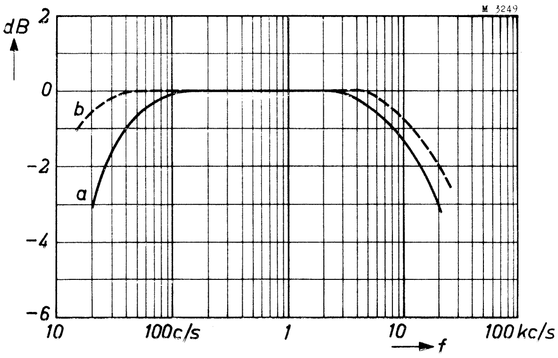

Fig. 63. Frequency response curves, (a) for microphone signals and (b) for the radio and pick-up channels.

20 c/s to 20 kc/s. The response curve measured between either the radio or the pick-up socket and the output terminals is also shown in Fig. 63. Because the EF 86 stages are not included for this second position of measurements, the bass response is slightly extended and the curve is flat to within ± 2 dB, compared with the 1 kc/s level from 15 c/s to 20 kc/s.

HUM AND NOISE

Measurements of hum and noise were made with 100 kΩ resistors connected across the microphone and pick-up terminals and with the potentiometers fully advanced. These arrangements simulate a reasonable practical condition. The mixer was connected to a 10 W amplifier, and measurements were made across a 15 Ω load resistor. The voltage measured in this way was 38 mV.

The full output of the amplifier (10 W) corresponds to a voltage of 12.3 V across the 15 Ω load resistor. Consequently the hum and noise level is 50 dB below 10 W. The background level of the 10 W amplifier alone is better than 70 dB below 10 W, so the figure of 50 dB must be attributed to hum and noise in the mixer unit.