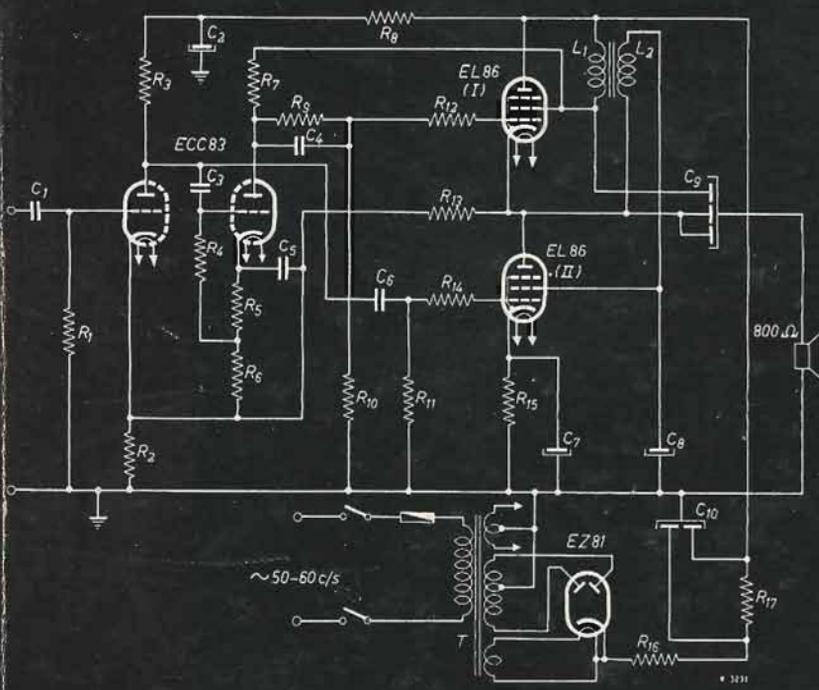


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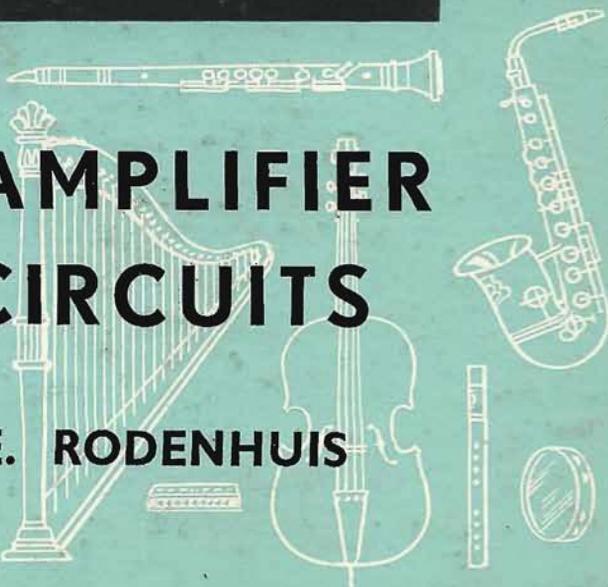


HI-FI AMPLIFIER CIRCUITS

E. RODENHUIS

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HI-FI AMPLIFIER
CIRCUITS

HI-FI AMPLIFIER CIRCUITS

Compiled by

E. RODENHUIS

1965

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Translated by G. Du Cloux, Exmouth, Devon (England)
This book contains ix + 125 pages and 76 illustrations
Other editions of this book in Philips Technical Library in French, German, Spanish
and Dutch languages
U.D.C. Nr. 621.375.2.029.4

Original Dutch edition:

© N.V. Philips' Gloeilampenfabrieken, Eindhoven, The Netherlands 1959

English edition:

© N.V. Philips' Gloeilampenfabrieken, Eindhoven, The Netherlands 1965

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First published 1960

Second revised and enlarged edition 1965

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Printed in The Netherlands.

PREFACE

Until a few years ago, "Hi-Fi" (high fidelity) reproduction was an ideal which could only be approached by the few amateurs who could afford to pay large sums of money for their equipment. The situation has now completely changed.

The quality of the available "signal sources" (long-playing records, F.M. broadcasts and magnetic recording tape) and reproduction equipment (amplifiers, pick-ups, turntables, loudspeakers, etc.) has improved considerably in recent years. The growing interest in this equipment led to a further rapid increase in quality accompanied by a decrease in prices so that "Hi-Fi" is now within the reach of everyone.

This book is intended for those people who are interested in building and experimenting with high-quality amplifiers, for example small manufacturers of "Hi-Fi" equipment, audio engineers and service engineers, students and amateurs: in short, all those whose business or pleasure it is to make use of these exciting techniques.

In a sense it is a continuation of the previous publication "Valves for A. F. amplifiers". The hints given there for assembling and wiring amplifiers still hold; some of the amplifier circuits given there can be replaced by those given in this book, which give a qualitative improvement in performance at a slightly higher cost.

This book also gives a number of pre-amplifier circuits.

All the circuits described here give high-quality reproduction, and can be built at a reasonable price by anyone who has acquired the minimum skills stipulated in the previous book. The three-tube pre-amplifier is an exception to this rule, however. It requires great skill in assembling the components and the wiring, and may be regarded as a challenge to those readers whose experience is already considerable.

In contrast to previous books of this nature, this book gives no tube characteristics. These can be found in the Philips Electron Tube Manual,

which may be obtained from any reputable radio tube dealer. This book would become needlessly expensive if all these figures and tables were included.

In the preparation of "Hi-Fi Amplifier Circuits" use has been made of work done by R. S. Babbs, D. H. W. Busby, P. S. Dallosso, P. R. Everaars, C. Hardcastle, Ir. J. Rodriques de Miranda, J. J. Schuring, C. Vonk, and the late W. A. Ferguson, who published a slightly different version of the 20-W amplifier in "Wireless World", May and June 1955.

I wish to express my hearty thanks to them for the information and advice which they so readily gave me.

Finally, I hope that this book will give its readers great pleasure and that it will arouse as much interest as previous publications in this series.

December 1959

E. Rodenhuis

PREFACE TO THE SECOND EDITION

The second edition of Hi-Fi Amplifier Circuits has been enlarged to include two new circuits both designed around the triode-power pentode ECL86. This tube, which was developed after the first edition of this book was finished, has attractive characteristics for amplifier design. One of the new circuits operates with only one tube, but nevertheless its performance is surprisingly good. The other is of a high-quality amplifier with two ECL86's in push-pull. I am indebted to Mr. Jansen and Mr. Blok who developed these circuits in their laboratory.

Several readers wrote to me asking for additional information on one circuit or another. This gave me the impression that I had been insufficiently clear in some of the circuit descriptions. In this second edition the text has been modified accordingly. I wish to express my gratitude to those readers who also brought some printing errors to my attention. The obvious interest shown in my writing has encouraged to start writing another book on the inexhaustible subject of audio.

That the present book may give its readers as much pleasure as its predecessor is the wish of

December 1964

E. Rodenhuis

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CHAPTER I

GENERAL CONSIDERATIONS ON THE DESIGN OF HI-FI AMPLIFIERS

The equipment used for high fidelity reproduction usually consists of a gramophone motor and turntable, a pickup, a pre-amplifier with appropriate equalising networks, a power amplifier and one or more loudspeakers. If desired, a tape recorder and an f.m./a.m. tuner unit can also be incorporated. Each one of these component parts contributes to the quality of the reproduction and to the amount of distortion introduced into it. At best they can do no more than reproduce faithfully the signal obtained from the source of sound, whether this be a recording on disk or tape, or a radio broadcast, in which a certain amount of distortion is certain to be present. Where this distortion is linear it can be wholly compensated by equalisation or by a tone control in the pre-amplifier or by both of these methods, but there is nothing that can be done to overcome all other forms of distortion at the source.

In recent years more and more progress has been made in meeting the demands of the discerning listener to both broadcast and recorded music, and those responsible have done all in their power to reduce distortion at the source to a minimum. In general this has met with considerable success and has in turn fostered increased interest on the part of the general public in equipment capable of doing full justice to the high quality of the different sources of sound.

The amplifiers, which form part of the link between the source and the listener, can have a very marked effect on the quality of the reproduction and, at the same time, they provide the only means, among the various component parts of the equipment, of varying the more important features such as the

amount of amplification and the relative frequency response. It is not possible to vary the characteristics of loudspeakers, pickups, or recording and playback heads, but the controls in the amplifier enable us to match the whole of the reproducing system not only to the source, but also to the acoustical properties of the room, the personal tastes of the listener, and his auditory characteristics.

Now, as the amplifier is the only means of control or adjustment in any Hi-Fi equipment and is therefore the most important part of it from the point of view of the designer and constructor, we shall commence with some general points on this subject. This will be followed by circuit design details of complete power amplifiers and pre-amplifiers with equalisation, representing the latest of their kind in the range of medium- and low-priced equipment. Universal, or d.c./a.c., circuits have not been included, for in some countries there are safety regulations which prohibit the use of any apparatus in which a peak current of more than 0.7 mA can pass between any of the terminals and earth. This means, for instance, that the earth capacitor, on 220 V a.c. mains, may not be larger than 7000 pF, and that loudspeakers cannot be included in the negative feedback circuit. These limitations rule out the possibility of series supplied heaters, although it is possible to obtain quite reasonable results using a sensitive crystal pickup and an input transformer.

Quality requirements for Hi-Fi amplifiers

Briefly the requirements to be imposed on a Hi-Fi amplifier are

1. Low harmonic distortion (0.5%).
2. Low intermodulation distortion (2%).
3. Low beat-note distortion (0.8%).
4. Linear response curve up to at least 1 octave above the audible frequency range, i.e. linear from 10 c/s to approximately 30,000 c/s.
5. Little phase shift within this range.
6. Low hum and noise levels (— 52 to — 62 dB).
7. Sufficient output power to reproduce the loudest passages of music without noticeable signs of overloading.
8. Low output impedance to ensure electrical damping of the loudspeaker.

All these things of course affect the design of the amplifier as a whole, but they are of particular importance in the output stage. This stage should

be capable of sufficient power to guarantee realistic reproduction of orchestral music. 5 to 10 W (peak) is generally considered suitable for the average living room; in very large rooms or small halls 15 W should be regarded as essential for satisfactory reproduction of the peak sounds.

Two main types of output circuit are capable of providing an effective output of 10 to 15 W: class AB push-pull stage using pentodes and class A or AB with triodes. Each of these circuits has its own advantages and disadvantages as well as its advocates and opponents, but the ultimate choice is determined mainly by considerations of quality and cost.

Pentode output stage

The pentodes EL 34 and EL 84, which are well-established types, with rated anode dissipations of 25 and 12 W respectively, lend themselves to class AB output stages giving appreciably more than the rated 25 or 12 W, assuming that transformers of some 80% efficiency are used, which is quite a reasonable average for a present-day good quality transformer.

The efficiency of such an output stage is quite high, being something like 40 to 50%. The harmonic distortion, however, is rather high, being 3 to 5% on full load and so necessitating considerable negative feedback to attain the required low percentage of distortion of 0.5% (full load).

Operating conditions for class AB output stages as published by the tube manufacturers are based on the use of sinusoidal inputs, and a value is chosen for the biasing resistor such that the output stage will operate in class A without an input signal but the working point is shifted to class B conditions on full load. The load resistance between the anodes is then chosen for optimum matching under the class B conditions. This shifting of the working point is occasioned by the increasing voltage drop across the biasing resistor as a result of the rise in anode and screen currents when the input signal increases. In a typical output stage with two EL 84 pentodes operating from a 310 V supply the cathode currents and hence also the grid bias voltages rise by about 40% at full drive with a sinusoidal input.

However, when an output stage of this kind is used for speech or music a very different situation arises. The average amplitude of the input signal is very small compared with the peaks, which occur only occasionally, and the variations in the cathode current are accordingly also small. Now, owing

to the relatively high time constant of the particular combination of biasing resistor and decoupling capacitor, the displacement of the working point, even on peaks in the signal, is so small that the output stage can be regarded as operating with fixed bias.

Actual measurement of the corresponding fixed bias in conventional class AB output stages (with cathode resistor) shows that, with a sinusoidal input, more distortion is present with fixed bias than with the cathode resistor.

The difference between the two modes of operation is seen from Fig. 1. In both cases the quiescent bias is the same. Curve *a* shows the normal working conditions with cathode resistor, as quoted by the tube manufacturers, and curve *b* the conditions with fixed bias. It will be seen from these curves that an output stage based on the use of a sinusoidal input will introduce a disproportionate amount of distortion in the louder passages of either speech or music, because of the high time constant of the cathode circuit.

In practice the distortion can be reduced by adjusting the quiescent current of the output tubes, by means of the cathode resistor, to the optimum value required for fixed bias. This implies that both the quiescent current and the optimum load resistance must be made smaller, which in turn means that the instantaneous values of both the anode current and the screen current will vary to a greater extent when the stage is fully driven. The effect, however, is partially compensated by the higher value of the cathode resistor used and hence also of the time constant of the cathode circuit. The shift in the working point then remains only slight when the output valves are driven by either speech or music signals.

Practical tests on the 10 W amplifier described on p. 46 incorporating a circuit of this kind have shown that, with large electrolytic capacitors in the supply section, sudden peaks in the input signal result in so little reduction in the anode voltage that the instantaneous output power is not affected.

Another advantage of this arrangement is that these 12 W output pentodes are then operated at an average dissipation of only 7.5 W, which may be an important factor in prolonging their life.

Curve *c* in Fig. 1 shows the corresponding working conditions with fixed bias; it should be noted, however, that these conditions are suitable only for speech and music. The use of a sinusoidal input is accompanied by considerable distortion, which cannot therefore be measured in the usual way, but only observed aurally.

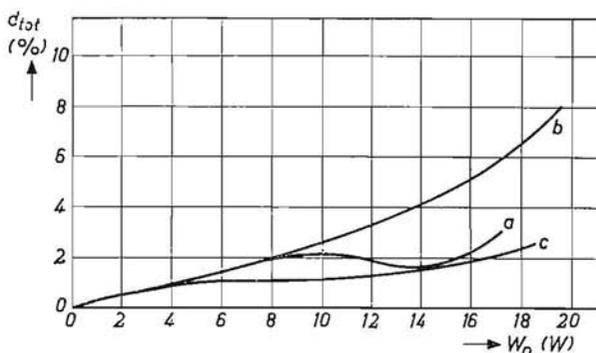


Fig. 1. Distortion curves: two EL 84 output pentodes in class AB push-pull *a*) with cathode resistors and *b*) with fixed grid bias under otherwise similar conditions, *c*) with reduced or low loading for optimum conditions with fixed bias.

Another method of improving the quality of a pentode output stage is by means of what is known as the “ultra-linear” or distributed load circuit. Incidentally, “ultra-linear” is a strange name for it, implying as it does that it surpasses the ordinary efforts to achieve linearity which are usual in amplifier design; all the same, this type of output stage, in which the load is divided or “distributed” between the anode and the screen grid, has gained much justifiable popularity. Details of the circuit are given later.

Triode output stage

Very low distortion can be achieved by using an output stage with two triodes in class A push-pull. Using 25 W output triodes, or tetrodes connected as triodes, with a 425 V supply, 12 to 15 W output can easily be obtained with less than 1% distortion.

The attainable output power and relative distortion are very dependent on the load impedance, as will be seen from Fig. 2, plotted from data obtained from the high-slope output pentode EL 34. The tubes were connected as triodes with the screen grid connected to the anode, and operated at a point somewhat below the maximum permissible anode dissipation of 25 W.

With load impedances of less than 7 k Ω either a common biasing (cathode) resistor or separate resistors may be used, but decoupling is essential in both

cases. Above $7\text{ k}\Omega$ a push-pull stage functions better with an unbypassed cathode resistor. The working point approaches class A conditions with increasing load impedance, optimum conditions occurring at about $10\text{ k}\Omega$.

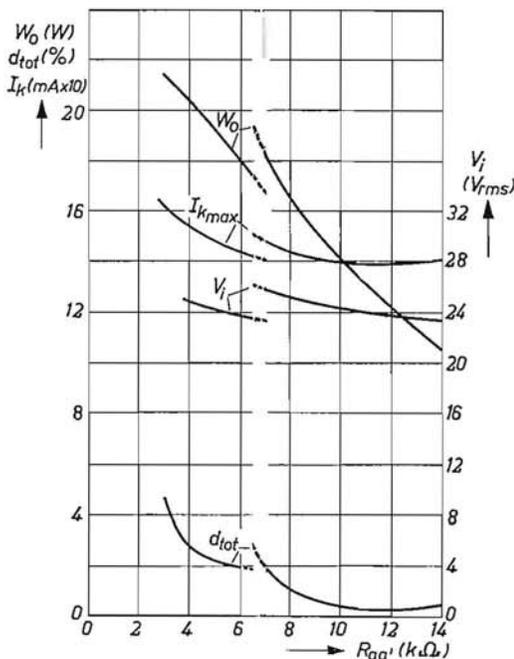


Fig. 2. Output power (W_o) distortion (d_{tot}), cathode current (I_k) and input alternating voltage (V_i) plotted against the load impedance for two EL 34 pentodes connected as triodes under class A push-pull conditions. Supply: 425 V. The curves on the left of the diagram relate to a bypassed cathode resistor; for those on the right a common, unbypassed cathode resistor of $235\ \Omega$ is used.

The output is then about 14 W, with a total harmonic distortion of considerably less than 0.5%.

Output stages of this kind were very popular a few years ago (Williamson c.s.) and many Hi-Fi amplifiers delivering about 12 W at the secondary of the output transformer have been designed along these lines. As the inherent

distortion is small, less negative feedback is needed to ensure reasonable linearity than for pentodes or tetrodes giving the same output. Moreover, in three- or four-stage amplifiers, in which the greater part of the feedback is applied from the output transformer secondary to the input, there is much less risk of instability than with other arrangements.

The main disadvantage of this type of circuit is its low efficiency, necessitating power supply units and output transformers which are both heavy and expensive, for a relatively low output.

The "ultra-linear" output stage

Although triode push-pull output stages have excellent qualities from the point of view of distortion, the above-mentioned disadvantage does constitute a serious objection, for which reason the "ultra-linear" output stage is becoming more and more popular. In this arrangement negative feedback is applied in the output stage itself; the screen grids of the output tubes are fed from suitable tapings on the primary winding of the output transformer, as shown in Fig. 3. In effect

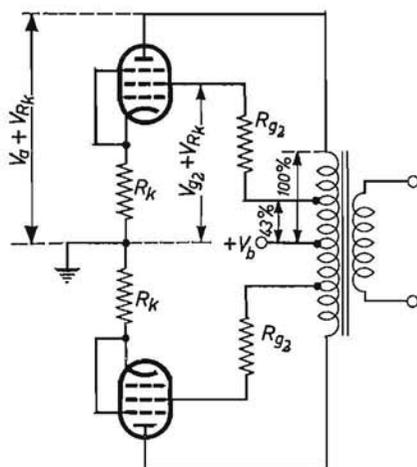


Fig. 3. Simplified circuit diagram of the ultra-linear push-pull output stage.

the stage can be regarded as one in which negative feedback is applied through the screen grids. The characteristics of the output stage are thus a compromise between triode and pentode output conditions, approaching the former according as the number of turns common to the anode and screen circuits is increased.

Under optimum conditions the ultra-linear output stage will deliver about 65% of the maximum output of the equivalent pentode circuit, but considerably less distortion proportionately. At the same time the internal resistance is reduced to the

level of a triode output stage. Using two pentodes of the 25 W class in an ultra-linear circuit it is possible to construct a Hi-Fi amplifier to give appreciably more than 20 W with very little distortion, which is a great deal better than can be obtained from the same tubes connected as triodes.

A great improvement in quality can also be achieved with 12 W pentodes when used in this way, allowing for a slight reduction in the available output power. In any case 10 or 12 W at the secondary of the output transformer would be attainable. Table 1 shows comparative figures for triode, pentode and ultra-linear operating conditions using either two EL 34 or two EL 84 pentodes. In the case of the former, the comparison between the ultra-linear and the triode conditions are the most striking, the output power with the tapped output transformer being just twice as much as that obtainable from a triode push-pull stage, with very low distortion.

Although tests have shown that a tapping ratio of 0.2, (i.e. 20% of the winding common to anode and screen grid circuits) gives a distortion comparable with the triode arrangement, it is also clear that an increase in the number of common turns yields an appreciable improvement in the reproduction at higher output powers. The best compromise in the overall performance of the output stage is found to be with 40 to 45% of the winding common. Although the output was found to drop slightly, 35 W was easily obtained in the tests, with a distortion of 2.5% at the point where grid current starts to flow.

With the EL 84 as the output tube the comparison between the ultra-linear and pentode push-pull circuits is even more interesting, considerably less distortion again being the feature of the former arrangement. With a common tapping at 20% the tubes will deliver 15 W with 1.5% distortion. As can be seen from Table 1, however, little is gained by pushing the operating conditions further towards those of a triode, i.e. to a higher tapping ratio. Nevertheless, there are two advantages in this type of circuit, particularly in the case of the pentode EL 34, from which a reasonable output is attainable. The characteristics of this tube are given in Fig. 4 for an ultra-linear output with a common tapping at 43%; the values given for the output power W_o were measured at the secondary side of the transformer, with a suitable load.

Under such conditions the performance of the circuit with fixed bias is roughly the same as with a cathode resistor. Since the conditions approximate

to those appertaining to triodes in class A, the variations in anode and screen grid current are only small when the stage is driven; the other advantage is that, as with triodes, neither the output power nor the distortion depend appreciably on the value of the load resistance; it will be seen from

TABLE I

Tube type	Circuit	Operating conditions					Total % distortion at $W_0 =$			
		V_a (V)	V_{g2} (V)	R_k (Ω)	$R_{aa'}$ (k Ω)	R_{g2} (Ω)	10 W	14 W	20 W	30 W
2 x EL 34	Triode push-pull	400	1)	470 per tube	10	1)	0.5	0.7		
	Ultra-linear 43 % com. turns	400	400	470 per tube	6.6	1000 per tube	0.6	0.7	0.8	1
	Pentode push-pull	375	375	130 com.	3.4	470 com.	1.5	1.9	2.5	3.8
2 x EL 84	Triode push-pull	300	1)	150 com.	10	1)	5 W	10 W	15 W	
							1			
	Ultra-linear 20 % com. turns	300	300	270 per tube	6.6		0.8	1.0	1.5	
	Ultra-linear 43 % com. turns	300	300	270 per tube	8.0		0.7	0.9		
Pentode push-pull	300	300	270 per tube	8.0		1.5	2.0	2.0		

1) Screen grid connected to anode.

the figure that there is little difference when the load resistance is varied between 6 and 10 k Ω .

Negative feedback

The distortion figures shown in Table 1 can certainly be considered as too high for a Hi-Fi amplifier, so the distortion is further reduced by applying negative feedback, usually from the output of the amplifier to the input. Now,

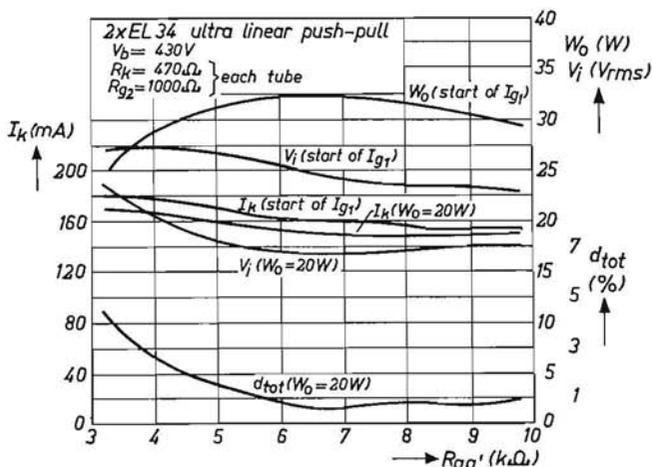


Fig. 4. Output power W_o , cathode current I_k , total distortion d_{tot} and input alternating voltage V_i as a function of the load impedance $R_{aa'}$, for two EL 34 tubes in an ultra-linear push-pull circuit with screen grid tapping at 43 % of the number of turns of the output transformer. Supply voltage $V_b = 430$ V. Each tube is provided with its own decoupled 470 Ω cathode resistor and 1 k Ω screen grid resistor.

the conventional amplifier will, as a rule, comprise a pre-amplifier, a phase inverter and an output stage. The feedback circuit includes the output transformer as well, and, as this component introduces most of the phase shift at the extreme ends of the frequency range, the amount of negative feedback that can be applied is determined almost wholly by the quality of the transformer.

In amplifiers with negative feedback, instability sets in if the loop gain, i.e. the product of the amplification without feedback and the attenuation

due to the feedback itself, is greater than unity at frequencies at which the total phase shift in the circuit is either 0° or 360° ; the voltage feedback will then be in phase with the input signal. As the conditions for negative feedback imply a phase shift of 180° , instability must necessarily occur if the additional phase shift in the amplifier and the feedback circuit approximate to 180° .

Whereas phase shift is very difficult to measure, amplification or attenuation can be measured quite simply; hence the relationship between phase shift and attenuation can be very useful to the amplifier designer.

A simple high- or low-pass filter consisting of a resistor and capacitor will produce an ultimate phase shift of 90° , the attenuation approaching asymptotically to 6 dB per octave. It follows, then, that an ultimate phase shift of 180° must produce 12 dB attenuation per octave. Thus, so long as care is taken in the design of the amplifier that the curve representing the loop gain versus frequency does not drop more than 10 dB per octave in that part where it passes from $+10$ dB to -10 dB through the point where the gain is unity, the amplifier will be adequately stable. Consequently, the response curve of an amplifier must be known well beyond the audible part of the frequency spectrum.

Stability becomes increasingly difficult to obtain as the amount of negative feedback is increased and, as 25 to 29 dB is quite usual nowadays, the phase shift due to the output transformer must of necessity impose a limit on the attainable gain and quality.

In three- or four-stage amplifiers it is practically impossible to achieve constant, and at the same time strong, negative feedback throughout the whole of the audible range of frequencies if the feedback circuit covers the whole amplifier, including the output transformer. The stability would be inadequate. Therefore, the effective negative feedback is usually reduced at the higher and lower frequencies. At the same time, there must be sufficient negative feedback in the region of the frequency of natural resonance of the loudspeaker to ensure the low internal impedance necessary for effective electrical damping. There should also be enough feedback at the highest frequency whose harmonics lie in the audible zone, 10 kc/s being taken as the limit.

Output transformer

As shown above, the output transformer can be regarded as the most critical

component in the whole amplifier; a badly designed transformer will often be the source of distortion which is being looked for in some other part of the amplifier. Distortion introduced by the output transformer may be of the following kinds:

(1) *Frequency distortion*, caused by too small a primary inductance, too much leakage inductance, or resonance effects.

(2) *Phase distortion* caused by phase shift when the feedback voltage is taken from the secondary of the output transformer. Usually this form of distortion shows itself as parasitic oscillation at the higher frequencies, due to phase shift resulting from high leakage inductance and self-capacitance.

(3) *Intermodulation and harmonic distortion in the output stage*, due to overloading at low frequencies when the primary inductance is too low. This tends to reduce the effective load impedance and also to result in reactive loading at the lower frequencies; the otherwise elliptical load curve becomes more of a circle, which does not match the anode current anode voltage curves of the output tubes.

(4) *Intermodulation and harmonic distortion*, resulting from non-linearity between the magnetising current and the magnetic field in the transformer core. This is of course always present, but can be greatly reduced by keeping B_{max} below a certain value (about 7000 gauss for standard transformer sheet).

(5) *Harmonic distortion*, due to too high a resistance of the primary winding, which at the same time affects the efficiency.

Summarising, then, a good output transformer should satisfy the following conditions:

- (1) high primary inductance;
- (2) low leakage inductance and low capacitance of both primary and secondary;
- (3) B_{max} within safe limits;
- (4) high efficiency (low losses in windings and core);
- (5) accurate matching.

Clearly, however, the output transformer will have to be a carefully balanced compromise of all these, often conflicting, factors. High inductance can be attained only by providing a large number of turns, but this is directly opposed to the need for low resistance and low selfcapacitance. Again, a low value of B_{max} means a large cross-section of the core, but this involves

large dimensions, which in turn has the same result as high primary inductance.

The leakage inductance and self-capacitance can be kept within reasonable limits by employing special winding methods, which also tend to make the leakage resonant frequency higher. This resonant frequency determines to a large extent the quality of the transformer.

It may be useful here to look more closely at the qualitative requirements for Hi-Fi amplifiers, since these depend to such an extent on the output transformer.

Why is it that these amplifiers should be capable of reproducing frequencies up to 60 kc/s and even higher when the human ear is not sensitive to anything above 16 kc/s, or in exceptional cases, possibly 20 kc/s? This is related to the fact that the initial note of many sounds is of an explosive or almost eruptive character. If measurements are taken on an amplifier with the aid of a sinusoidal signal, the voltage is seen to rise gradually to a maximum and then also gradually drop to zero, this being repeated in the opposite direction. In the case of a sudden sound, however, the rise time is very short and the voltage is anything but sinusoidal. It is for this reason that a square-wave voltage is more useful for such measurements than a sinusoidal one. The wave-front of a square-wave voltage has a rise time that can only be expressed in microseconds. Now, as is known, a square-wave voltage can be broken down into a number of sinusoidal voltages, all of which are odd harmonics of the fundamental. The most important of these is the third harmonic, whose amplitude is equal to one-third of the fundamental wave. The amplitude of the fifth harmonic is similarly equal to one-fifth of the fundamental, and so on.

If the bandwidth of the amplifier is too narrow, the third harmonic of a given fundamental in the higher frequencies will not be reproduced strongly enough; this applies all the more to the fifth harmonic. The amplifier thus makes the square-wave voltage start to resemble a sine wave, and the result is considerable distortion.

There are, of course, no square-wave frequencies in music, but some very short rise times may nevertheless be expected, especially from the percussion instruments, comparable to a square-wave voltage of say 10 kc/s. If such sounds are to be reproduced faithfully the third harmonic must suffer little or no attenuation, and this at once requires a bandwidth of 30 kc/s.

If good response up to 20 kc/s is required the bandwidth must be at least 60 kc/s, that is to say, at 60 kc/s the attenuation must not exceed 3 dB.

To return to our output transformer, if it became necessary to employ one or more RC filters to suppress parasitic oscillation in the amplifier, caused by phase shift in the negative feedback circuit, without affecting the response curve up to 60 kc/s, the leakage resonance of the transformer would have to be at least 200 kc/s (it should be remembered that the secondary is included in the feedback circuit).

Now the leakage inductance can be kept low by taking the negative feedback from a separate winding tightly coupled to the primary of the transformer, and this can be achieved by employing a special but quite simple winding method. Experiments have been carried out on the 20 W amplifier with two EL 34 tubes, described on p. 67. It was found, however, that although the stability was excellent under all kinds of loads, the response curve dropped by 17 dB at 20 kc/s owing to the fact that the loudspeaker was not included in the feedback circuit. A transformer, the primary of which was divided into a number of windings in parallel, with the secondary windings also in parallel and wound between them, gave very much better results. Even so, it was not possible to couple the primary windings tightly enough, and the leakage inductance was therefore on the high side. An improvement was found to be possible using a very complicated winding method which would however be too difficult and too costly for normal production purposes. Capacitances can be neutralised by reversing the direction of the various windings, but the least departure from the symmetry or precision of winding tends to make any improvement illusory.

The best results have been obtained by dividing the primary into a number of windings in series, with a series-parallel secondary between them ¹⁾. In effect the former is divided in two halves; in each half the first primary section is started at the outer end, this being followed by a secondary coil on each side, with the next primary section above it. In all there may be say 10 primary sections, 5 on each side with 4 secondary sections between them (8 in all), again on each side. The primaries may be interconnected internally, but this should preferably be done externally, the secondaries being connected partly in parallel and partly in series to suit the required

¹⁾ See page 80.

matching. The result is a very versatile transformer offering a wide range of matching possibilities, with facilities for connecting screen grids in ultra-linear circuits to any desired tapping on the primary winding.

It will be clear from the above, in any case, that the output transformer is a critical and also expensive item where high fidelity is required.

Single-ended push-pull output stage

It will not come as a surprise that attempts have been made to match one or more loudspeakers to conventional types of amplifiers without any intermediate transformer at all. Many years ago output stages appeared with a steep-slope triode with a matching impedance of some 2000 ohms connected direct to an electro-dynamic loudspeaker whose speech coil had a similar impedance. This coil could not, of course, carry the direct anode current of about 60 mA and it was accordingly shunted by a very large choke of such a low resistance, that most of the d.c. passed through it. This could not be called a satisfactory solution to the problem, but it proved the feasibility of winding high impedance speech coils.

An experimental centre-tapped speech coil of 2 x 2000 ohms was designed for direct connection to a push-pull output circuit, but this was not a great success, the coupling between the two halves of the coil being far too loose. The effect may be seen from the output circuit depicted in Fig. 5a and its equivalent circuit, Fig. 5b. The latter in effect consists of 2 separate generators coupled only magnetically by the output transformer. Some of the distortion inherent in this circuit must be attributed to inadequate coupling, and the coupling of the two sections of the speech coil in a high-impedance loudspeaker is necessarily much inferior to that of a transformer, so that there is more, rather than less, distortion.

It is seen from Fig. 5a that the output tubes are in parallel as regards the d.c. supply, but in series from the point of view of the a.c. load. Now, if we interchange the components, we have the circuit shown in Fig. 6a, with the tubes in series as far as the d.c. supply is concerned, but in parallel with respect to the load R_L . For the same output power as before, twice the supply voltage is needed, but the anode direct current is halved; the d.c. power is therefore the same. The load resistor, which is connected to the

centretap of the supply source is one-quarter of the resistance (R_{int}) in the conventional output stage (Fig. 5b) where:

$$R_{aa} = \frac{2 V_a}{I_a}$$

whilst, in the single-ended push-pull circuit (Fig. 6c):

$$R_L = \frac{V_a}{2 I_a} = \frac{1}{4} R_{aa}$$

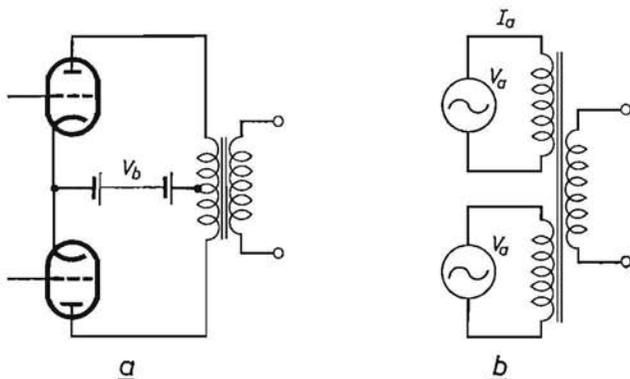


Fig. 5. Circuit diagram (a) and equivalent diagram (b) of conventional push-pull output stage.

The disadvantage of the arrangement in Fig. 6a is that the loudspeaker is at a high potential with respect to earth. In Fig. 6b the speaker is isolated from the d.c. source by a capacitor and is earthed. There is also no need for a centre-tap on the d.c. supply. The optimum matching impedance for a direct-coupled output stage can be further reduced by using such tubes as types EL 86, PL 84, or UL 84. These tubes were specially designed for this purpose, the internal resistance and hence also the optimum load resistance being considerably lower than with the usual type of output pentode. The single-ended push-pull circuit may still be developed further and this will yield, on the one hand, reproducers of exceptionally high quality and, on the other, an appreciable reduction in the cost of amplifiers and radio sets; for example, the output stage might be made to incorporate the phase inversion stage as well. Such circuits can be constructed with a relatively small number of components and, although an additional output tube is required, are vastly

superior in quality and power to the conventional class A single tube with transformer.

Details of these circuits would fall outside the scope of this book, in which only circuits ensuring the highest possible quality of reproduction with maximum efficiency for a given tube complement will be dealt with.

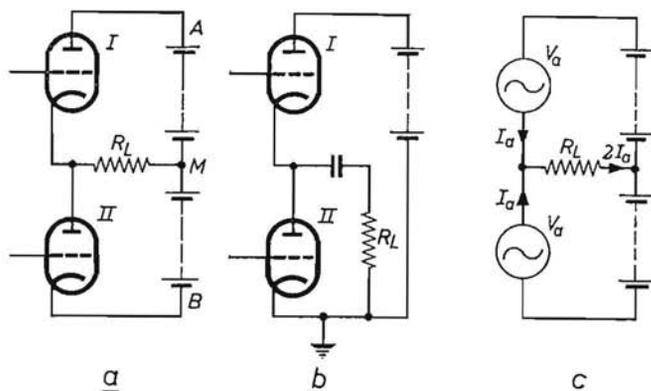


Fig. 6. Single-ended push-pull output stage: (a) with load resistor R_L connected to centre-tap on supply, (b) load resistor earthed; (c) equivalent circuit.

Typical problems inherent in single-ended push-pull output stages

The single-ended push-pull circuit as shown in Fig. 6 gives excellent results with triodes, or pentodes used as triodes. Pentodes, as such, with their much greater efficiency, give rise to certain problems centred round the screen grid. This has to be fed in such a way (Fig. 7) as to be at the same d.c. potential as the respective anode, but it must carry no a.c. voltage with respect to the cathode in the same tube. In the case of tube II the simplest way of achieving this is by feeding the screen grid from the H.T. supply line by way of a resistor R_1 , which is decoupled from the cathode with a capacitor C_3 . The value of the resistor should be so chosen that the screen grid voltage is about one-half the no-signal supply voltage. Even so, the screen-grid current rises with increasing signal voltage, and the screen-grid voltage then drops, thus shifting the working point and consequently increasing the distortion.

The voltage supply for the screen grid of tube I is not quite so straightforward, for the difference in voltage between the screen grid and the anode of the tube must be negligibly small. A resistor in series with the screen grid would be in parallel with the load and would absorb too much of the output power. The best solution is to feed the screen grid via the load resistor itself, i.e. by connecting the high-impedance loudspeaker direct between the anode and screen grid of tube I. The relatively low screen grid direct current has very little effect on the performance of the $800\ \Omega$ loudspeaker, which handles the whole of the output power. The only objection to this arrangement is that the loudspeaker is only earthed capacitively, and is itself at a high potential.

To avoid flashover between the speech coil and the metal chassis the latter must also be connected to the positive H.T. line, which is usually contrary to the safety regulations. As a compromise, a $6.8\ \text{k}\Omega$ resistor may be placed in series with the screen grid. The loudspeaker is then fed through a capacitor and can accordingly be earthed. An advantage of this is the reasonably good working point which it provides for tube I, although some 12% of the output power is lost in the resistor. Alternatively, a choke may be used in place of a resistor; this in any case takes up considerably less room than a transformer, and a narrow air-gap may be provided, seeing that little direct current flows. A choke of 5 H would be quite large enough for an $800\ \Omega$ loudspeaker with a lower frequency limit of 30 c/s. The screen-

grid voltage for tube II can be stabilised with the aid of a stabiliser tube or a voltage dependant resistor, and the half-value voltage can be taken from the power supply, especially if this is fitted with four semiconductor rectifiers

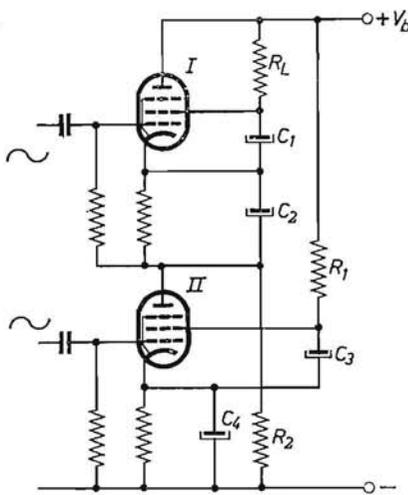


Fig. 7. Single-ended push-pull stage for using pentodes. The screen grid of tube II is fed from R_1 , that of tube I from the load resistor R_L . The anode current of tube II is higher than that of tube I because the tubes are in series. R_2 is accordingly connected in parallel with tube II to restore the d.c. balance.

connected in a bridge circuit as shown in Fig. 8. This voltage is obtained from the centre-tap of the mains transformer. The smoothing choke should be on the negative side of the d.c. supply so that the screen-grid current of tube II is also smoothed. A certain amount of asymmetry is introduced in the supply by the voltage drop across the choke, with the result that the output power is reduced slightly, but if this is not acceptable the choke can be

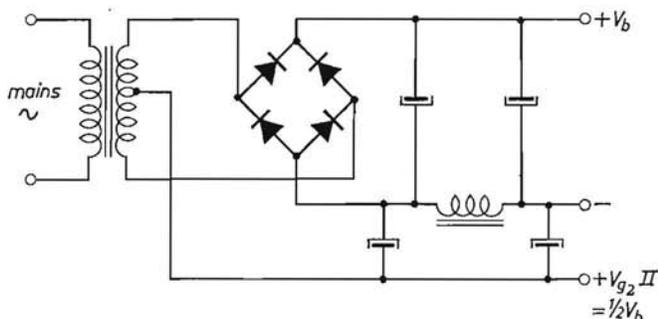


Fig. 8. Power supply circuit using metal rectifiers. The screen grid voltage for tube II is taken from the centre-tap of the power transformer.

divided between H.T. positive and negative. Magnetisation of the core can be kept low by arranging the direction of the current accordingly.

As the two output tubes are in series as far as the direct current is concerned, the anode current of tube II is greater than that of tube I, since the screen-grid current of the latter also flows through the former. Consequently the working points of the tubes are not the same, even though the cathode resistors are of the same value. In order to meet the push-pull conditions, therefore, a resistor (R_2 , Fig. 7) is connected in parallel with tube II to compensate for the effect of the screen-grid current of tube I; the cathode current and therefore the working points of the two tubes will then be the same. It is true that this extra resistor will be in parallel with the load but, as its value is fairly high, there is very little loss of output power.

The most effective solution to this difficulty of the screen-grid supply is shown in the basic output circuit diagram in Fig. 9. Here each screen grid is fed from its associated anode through one half of a double choke. The advantage of this is that the screen grids are then practically at the same d.c.

potential as their respective anodes, so that the cathode current is the same in each case. As the direct currents flow through the windings of the choke in opposite directions, the magnetic fields neutralise each other and the core is thus not magnetised; there is thus no need for an air-gap, resulting in a component which is small but which nevertheless has a high inductance.

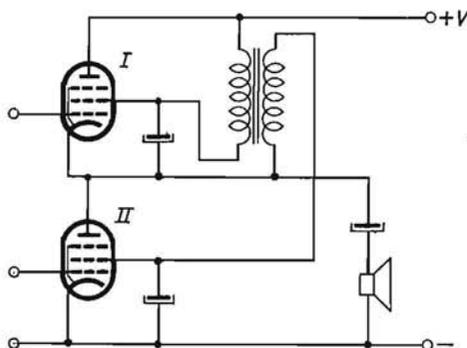


Fig. 9. Method of feeding screen grids in a Single-ended push-pull output stage, using a double choke. The screen grids are at the same potential as the respective anodes. Magnetisation of the core of the choke is neutralised by making the currents in the two halves flow in opposite directions; this ensures high inductance in a relatively small choke.

The power supply need not be centre-tapped and sufficient smoothing can be obtained from a double electrolytic capacitor and a resistor.

This arrangement, for use with a high-impedance loudspeaker has been adopted for the super high quality amplifier described on p. 85.

Phase inversion

The two tubes in a push-pull output stage are driven by signals of the same amplitude but of opposite phase, with as little distortion as possible. If a reasonable amount of amplification can be obtained at this stage, so much the better, since the negative feedback, which is applied to the whole amplifier, imposes a limit on the number of stages that can be used, as the phase shift must be reduced to a minimum if the required stability is to be guaranteed.

Some designers have a preference for the maximum possible amplification, even at the cost of increased distortion, and the circuit shown in Fig. 10 is frequently encountered. The maximum amount of negative feedback is applied over the whole amplifier to bring the distortion to within reasonable limits.

Other designers are content with a moderate amount of gain with less

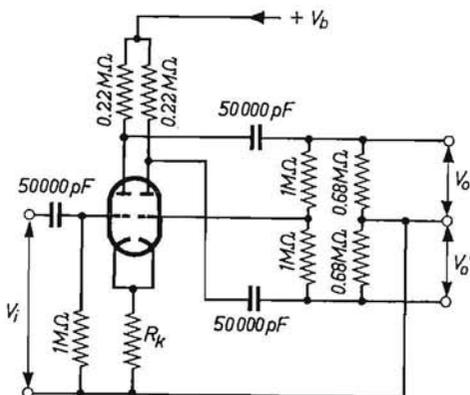


Fig. 10. Phase inverter stage with fairly high amplification but with a correspondingly large amount of distortion.

distortion, in which case the circuit in Fig. 11 will be preferred. Here there is no need for so much negative feedback, and the risk of instability is also reduced.

In both types of circuit the high-gain double triode ECC 83 is generally used; the operating conditions and other data for this tube will be found in Table 2.

Now, with triodes the distortion decreases in direct proportion to the output voltage so that, with comparable values of 20 and 35 V_{rms} for the output voltage at d.c. supplies of 250 and 350 V respectively, the distortion for the circuit in Fig. 10 would drop to 3.2 respectively 2.7% whilst the gain is more than twice that of Fig. 11. The second circuit has the advantage, however, of much lower distortion and that there is very little phase shift because, owing to the high cathode voltage, the anode of the preceding

TABLE 2

Circuit	Fig. 10		Fig. 11		
Supply voltage	250	350	250	350	V
Cathode resistor	1.2	0.82	68	82	k Ω
Anode resistors	0.22	0.22	0.1	0.15	M Ω
Output voltage	35	20	45	35	V _{rms}
Gain	58	62	25	27	—
Total distortion	5.5	3.2	3.5	2.7	%

tube can be connected direct to the grid of the phase inverter tube. As the gain obtained from the ECC 83 when used as a cathode-coupled phase inverter is nevertheless quite reasonable, this circuit has become very popular for use in high-fidelity amplifiers.

The signal voltage required to drive fully two EL 34 pentodes is about

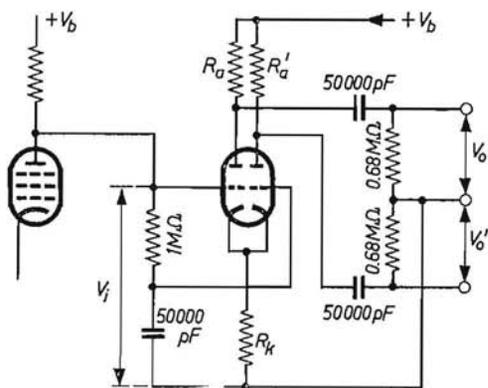


Fig. 11. Cathode coupled phase inverter circuit.

The amount of gain is about one half of that obtained with the circuit shown in Fig. 10, but the distortion is much less.

$2 \times 25 V_{rms}$, whereas two EL 84's require only $2 \times 10 V_{rms}$. In either case it makes no difference whether the tubes are used as pentodes, as triodes or in an ultra-linear output stage. This implies that in an amplifier without negative feedback the input to the ECC 83 should be about $0.4 V_{rms}$ when preceding the EL 34 push-pull output stage, using the circuit shown in Fig. 10

on 350 V. For the cathode coupled circuit an input of $0.9 V_{\text{rms}}$ would be necessary, and the values for a push-pull output stage using two EL 84 valves are then $0.17 V_{\text{rms}}$ or $0.4 V_{\text{rms}}$ respectively on a 250 V supply. This implies that the phase-inverter should be preceded by a pre-amplifier stage for which usually the low-noise, low-microphony A.F. pentode EF 86 is chosen.¹⁾

If the EF 86 is operated to give maximum amplification—as is usually the case—the overall sensitivity of the amplifier will generally be found to be too high for use with a crystal pick-up. In the case of the 10 W amplifier described on p. 42 therefore, the surplus sensitivity is utilised in a very effective tone control system. At the same time, this circuit is not sufficiently sensitive for use with a magneto-dynamic pick-up, or a microphone. Present-day amplifying systems, for the same reason, are usually provided with a separate (but often built-in) pre-amplifier equipped with all the necessary controls and equalising networks.

It follows, then, that the input signal to such power-amplifiers may be fairly large, and this has led to the experimental development of phase inverter circuits giving so much gain that the preceding stage with the EF 86 can be omitted. The circuit employs a combination of positive and negative feedback, which makes for very high amplification with adequate stability and low distortion. This circuit was found to give a gain of 800 quite easily, although the Miller effect then tends to attenuate the treble frequencies to such an extent that the gain in the practical version of the circuit (Fig. 12) is reduced to about 220.

It will be seen that the cathode resistor in the left hand triode is not decoupled, resulting in negative feedback. The earthy end of the cathode resistor in the right-hand triode is connected to the “top” end of the first-

¹⁾ It is perhaps worth-while mentioning here that the ECC 83 is sometimes also used for the pre-amplifier stage of very sensitive amplifiers. This tube is guaranteed free from microphony when used at a sensitivity of 50 mV for 5 W output, using a loudspeaker of 5 % efficiency, placed in the immediate vicinity of the tube. Cases are known of the ECC 83 being used for sensitivities as low as 1 mV and, although here and there the tube will certainly give good results under such conditions, this should not be regarded as typical. There can of course be no objection to selecting those tubes which are exceptionally free from microphony, but for mass production of, say, tape recorders with their built-in loudspeaker, the use of the ECC 83 in highly sensitive pre-stages is inadvisable. The EF 86, which was specially designed for such applications, should be used instead.

mentioned resistor, and the negative feedback is thus not only neutralised but converted to positive feedback.

Regeneration through R_2 would very soon produce instability if it were not for the fact that the load resistor of the first triode is connected to the cathode of the second triode; the cathode voltage is in phase with the anode voltage of the first triode and this results in considerable negative feedback which serves to keep the circuit stable.

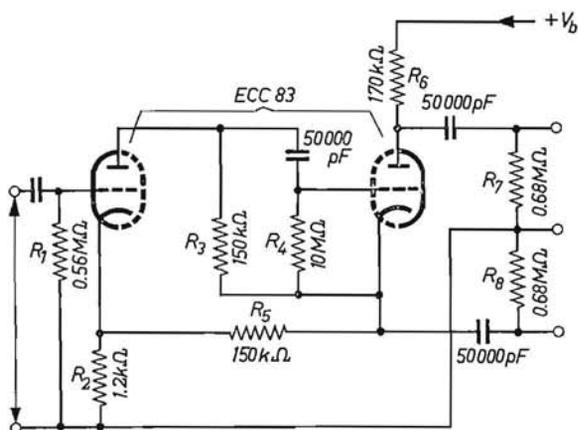


Fig. 12. Phase inverter circuit employing both positive and negative feedback. An amplification of 220 can be obtained without loss of stability. With still more gain, Miller effect cuts the treble too much.

With this circuit followed by a conventional output stage, and 26 to 30 dB negative feedback across the whole amplifier, an integral pre-amplifier stage is not necessary if a separate pre-amplifier incorporating all the necessary controls is provided.

Input stage and phase inverter circuit for a single-ended push-pull output stage

A great advantage of an output stage which is coupled direct to the loud-speaker is that, owing to the absence of the transformer, the phase shift in the amplifier is very small and considerable negative feedback can be

applied without involving any risk of instability. In fact, in the driver stage the feedback can be applied almost to the point of oscillation, the negative feedback over the complete circuit taking care of the overall stability. The gain in the input stage may therefore be very considerable.

Fig. 13 illustrates the principle on which this type of circuit operates. The left hand section of the double triode is in effect the input tube, the signal from it being passed to the grid of the right hand section and also to the grid of output tube II. Section II of the double triode is the phase inverter, the signal from which is taken to the grid of output tube I. In contrast with the conventional procedure the alternating input voltages, although of opposite phase, are not of the same amplitude. Tube II may be regarded as an ordinary output tube operating under Class A conditions and delivering a

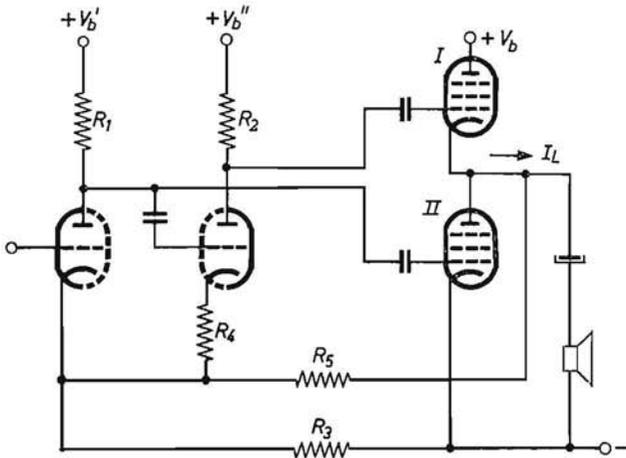


Fig. 13. Simplified diagram of phase inverter circuit with positive feedback, used with a single-ended push-pull output stage. Stability is ensured by applying negative feedback from the output to the input via R_5 .

certain alternating current to the loudspeaker. Tube I, however, is more in the nature of a cathode follower, with a load resistance which varies with the output signal (tube II) and, giving no voltage gain, passes to the loudspeaker the same amount of current, and in the same phase, as that supplied by tube II. As the input signal to a cathode follower is in phase with the out-

put signal, the drive for tube I must be opposite in phase to that of tube II. Tube I, as a cathode follower, provides an amplification less than unity and, as the cathode is at an a.c. potential equal to that of the output signal, the phase inverter must supply a voltage equal to the sum of the output voltage and the control voltage necessary for tube I to deliver the same output current as tube II.

Let us assume then, that the output voltage is 90 V and that the control grid voltage for tube I will have to be 12 V; the phase inverter would therefore have to deliver 102 V, which is more than the ECC 83 is capable of supplying.

Now, for the tube to be able to deliver an output voltage of this magnitude it must be fed from a point where the whole of the d.c. voltage is available, but with the a.c. output voltage superimposed on it. Such a point is, in fact, available, namely the screen grid of output tube I. As far as the d.c. voltage is concerned, this grid is at the same potential as the anode; from the aspect of a.c. output voltage it is at cathode potential.

However, since a certain voltage drop takes place across the anode resistor of the phase inverter tube, the amplification here should be rather more than as calculated. In any case this circuit is well able to deliver the required input signal for tube I.

Positive feedback is also applied to this combined pre-amplifier and phase

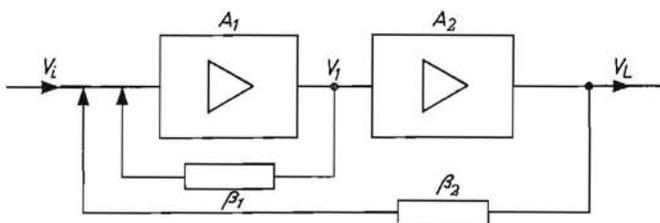


Fig. 14. Block diagram of two amplifiers A_1 and A_2 in cascade. The output voltage from A_2 is fed back to the input of A_1 via β_2 .

inverter tube in the same way as in the circuit already described, namely with the aid of a common cathode resistor. Theoretically, if this circuit is used in conjunction with a single-ended push-pull output stage, the first stage can be operated at the threshold of oscillation; it is maintained in a stable condition by heavy negative feedback derived from the output

signal. As the phase shift in the output stage is only small owing to the absence of the transformer, negative feedback can be applied to a degree that would result in serious instability in any conventional amplifier. In Fig. 13, the regenerative voltage is obtained from R_3 and the negative feedback from R_5 . The effect of this combination of feedback is illustrated in Fig. 14.

A_1 and A_2 represent 2 stages of an amplifier connected in cascade. The voltage gain in these stages is also denoted by A_1 and A_2 . A part, $\beta_1 V_1$, of the output signal from A_1 is fed back to the input, together with $\beta_2 V_L$, this being part of the output signal V_L , obtained from A_2 .

Denoting the input signal to the whole amplifier by V_i :

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

and:

$$V_L = A_2 V_1$$

The total voltage gain will then be:

$$A = \frac{V_L}{V_i} = \frac{A_1 A_2}{1 - A_1 \beta_2 - A_1 A_2 \beta_2} = \frac{A_1 A_2}{N},$$

where $N = 1 - A_1 \beta_1 - A_1 A_2 \beta_2$

It can also be shown that, if the distortion in the two stages be denoted by d_1 and d_2 and the total distortion by d , then d will be:

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

From this expression it is seen that something peculiar takes place when $A_1 \beta_1$ is equated to unity, for in this case d is reduced to d_1/N ; in other words the distortion in stage A_2 contributes nothing to the total distortion.

The distortion still present, d_1/N , is determined by the (usually small) distortion in the first stage, d_1 and the quantity N which, if $A_1 \beta_1 = 1$, is equal to $-A_1 A_2 \beta_2$. In absolute values the product of $A_1 A_2 \beta_2$ will be appreciably greater than 1, in which case the total distortion d will be even less than the already small value of d_1 . The absolute value of the overall gain is then $1/\beta_2$ and is thus independent of A_2 .

The condition that $A_1 \beta_1 = 1$ for a wide band of frequencies will be met if A_1 and β_1 are real within that range of frequencies, i.e. so long as the amplifier A_1 and the negative feedback network β_1 contain no components which

introduce phase shift. It is quite possible to approximate such conditions in a pre-amplifier stage.

To satisfy the condition $A_1\beta_1 = 1$ the positive feedback in this stage should be such that the circuit is at the threshold of oscillation; this does not imply that the combination $A_1\beta_1 - A_2\beta_2$ is unstable. Assuming that the second stage contains nothing likely to introduce an adverse phase shift, the whole circuit can be maintained in the stable state by making the feedback β_2 negative. The absence of the inevitable phase shift produced by an output transformer enables us to construct an amplifier with exceptionally low distortion and relatively high sensitivity, since the loss of gain due to the strong negative feedback is partly compensated by the positive feedback in the initial stage.

At the same time, although this theoretical proposition does lend itself to practical realisation and it is certainly possible to employ an input stage operating on the threshold of oscillation without any instability, there is an inherent disadvantage in an amplifier of this kind. As long as the input signal is such that the output stage is not overloaded, the amplifier will show no signs of instability. However, as soon as the input signal becomes strong enough to overload the output stage, the gain in this stage drops, and with it, the negative feedback voltage, which is, of course, a part of the output signal. The amplifier then becomes unstable and bursts into oscillation at the resonant frequency of the combined circuits. Such conditions could be brought about by a passage on a percussion instrument, so in practice, the input stage is thus adjusted that fortuitous overloading of the output stage will not initiate these oscillations.

Power Supply

In general the power supply for Hi-Fi amplifiers present no difficulties. Although earlier designs included a separate power supply unit with expensive smoothing chokes, practice has shown that there is no real reason why the mains transformer should not share the same chassis with the power amplifier itself, especially if a separate pre-amplifier is used in cases where the input signal is weak. Care is of course necessary in the layout of the components, but there should be no difficulties.

If electrolytic capacitors are used there is no need to use a smoothing

choke. The anodes of the output tube can be fed from the reservoir capacitor and the screen grids from the second capacitor. A resistor takes the place of the smoothing choke. Supplies for the input and phase-inverter tubes are obtained from smoothing filter networks consisting of a resistor decoupled by an 8 to 25 μF electrolytic capacitor. The hum level will then be 60 to 90 dB below full output.

Indirectly-heated rectifying tubes are generally used, the warming-up time being preferably a little longer than that of the output tubes to ensure that the voltage on the electrolytic capacitors does not exceed the rated working value every time the amplifier is switched on. Cheaper electrolytic capacitors can then be used than those required for directly-heated rectifiers or metal rectifiers.

The EZ 81 is the recommended rectifier to use for voltages up to 2×350 V_{rms} at 150 mA; for higher values, up to 2×450 V_{rms} at 250 mA the GZ 34 is recommended.

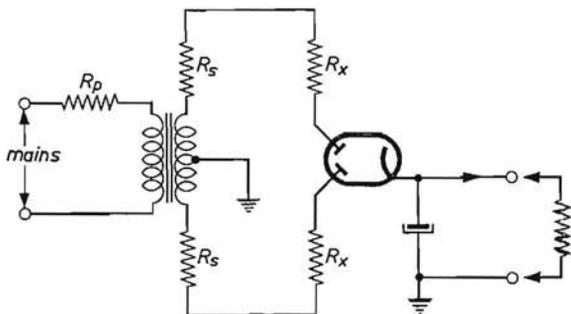


Fig. 15. Full-wave rectification circuit. The d.c. resistance of the mains transformer primary is denoted by R_p and the halves of the secondary by R_s . R_x is the extra resistance to be connected in series with the anodes to obtain the necessary minimum permissible transformer resistance.

Care should be taken that the resistance of the transformer (R_t), is not less than the limit value stated in the tube published data. This resistance may be ascertained by measuring the d.c. resistance of the primary and one of the secondaries. Let the former be R_p and the latter R_s ; then $R_t = R_s + n^2 R_p$, where n is the ratio of the primary to secondary turns (voltage

ratio). If this proves to be less than the minimum permissible resistance, a resistor R_x should be connected in series with each of the anodes of the rectifier, the value of R_x being found from the formula $R_x = R_{lmin} - R_l$. Each of these resistors carries half the d.c. output plus half the ripple, so they should be rated for 1.2 times the total d.c. current. Alternatively a single resistor of the same value may be connected in series with the cathode, but in this case it will have to carry 2.4 times the total d.c. current.

The reservoir capacitor in the H.T. section should be sufficiently large to take the ripple current, which is usually of the order of 1.4 times the d.c. current. If a separate pre-amplifier is used, and possibly also an FM/AM tuner unit, to be fed from the power amplifier, it is advisable to include the smoothing resistor and capacitors in the power amplifier chassis; allowance will of course have to be made for this in the dimensioning of the circuit elements.

CHAPTER II

POWER AMPLIFIER CIRCUITS

A SIMPLE 3 W GRAMOPHONE AMPLIFIER

Strictly speaking, this circuit should not be included under the heading of Hi-Fi amplifiers in view of its low output and restricted tone control, but the interesting circuitry and the surprisingly high quality of the response go a long way towards justifying its publication.

Although only two amplifying tubes are used, the amplification is ample for all types of crystal pickup; an equalising network is included between the pickup and amplifier. The rated output is 3 W, with 1% total harmonic distortion.

General data

Tubes: EF 86 voltage amplifying pentode

EL 84 output pentode

EZ 80 full-wave rectifier

Output: 3 W

Harmonic distortion: 1%.

Response curve: Flat to within 1 dB from 20 c/s to 40 kc/s

Tone control: maximum treble cut: 20 dB at 10 kc/s

maximum bass lift: 15 dB at 70 c/s

Sensitivity: 100 mV for 3 W output

Hum and noise level with respect to full output: — 70 dB.

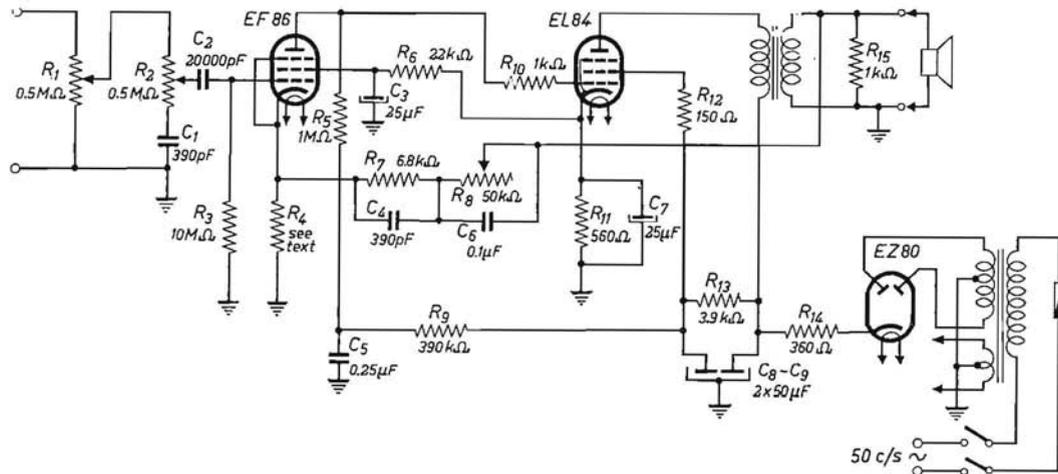


Fig. 16. Circuit diagram of 3 W amplifier using the EF 86, EL 84 and EZ 80.

List of Parts

$R_1 = 0.5 \text{ M}\Omega$, log.	$R_{10} = 1 \text{ k}\Omega \pm 20\%$, $\frac{1}{4}$ W	$C_4 = 390 \text{ pF} \pm 10\%$
$R_2 = 0.5 \text{ M}\Omega$, lin.	$R_{11} = 560 \Omega \pm 5\%$, 3 W	$C_5 = 0.25 \mu\text{F}$, 350 V wkg.
$R_3 = 10 \text{ M}\Omega \pm 20\%$, $\frac{1}{4}$ W	$R_{12} = 150 \Omega \pm 20\%$, $\frac{1}{4}$ W	$C_6 = 0.1 \mu\text{F} \pm 10\%$, 150 V wkg.
$R_4 = \text{see text}$, $\frac{1}{4}$ W	$R_{13} = 3.9 \text{ k}\Omega \pm 10\%$, $\frac{1}{4}$ W	$C_7 = 25 \mu\text{F}$, 50 V wkg.
$R_5 = 1 \text{ M}\Omega \pm 10\%$, $\frac{1}{4}$ W ¹⁾	$R_{14} = 360 \Omega$, 2 W (see text)	$C_8 \left. \vphantom{C_8} \right\} = 2 \times 50 \mu\text{F}$, 350 V wkg.
$R_6 = 22 \text{ k}\Omega \pm 10\%$, $\frac{1}{4}$ W	$R_{15} = 1 \text{ k}\Omega \pm 20\%$, $\frac{1}{4}$ W	$C_9 \left. \vphantom{C_9} \right\}$
$R_7 = 6.8 \text{ k}\Omega \pm 10\%$, $\frac{1}{4}$ W	$C_1 = 390 \text{ pF} \pm 10\%$	
$R_8 = 50 \text{ k}\Omega$, log.	$C_2 = 22\,000 \text{ pF}$	
$R_9 = 0.39 \text{ M}\Omega \pm 10\%$, $\frac{1}{4}$ W	$C_3 = 25 \mu\text{F}$, 50 V wkg.	

¹⁾ High stability.

CIRCUIT DESCRIPTION

The interesting feature of this circuit (Fig. 16) is the method of connecting the EF 86 in what is known as a "starvation circuit".

The tube is operated with a very high anode load ($1\text{ M}\Omega$) and a low screen grid voltage, derived in this case from the cathode of the output tube. Grid bias is obtained by means of a high-value grid leak ($10\text{ M}\Omega$). In this circuit the voltages and currents of the tube are very much lower than normally, but the gain is two to three times as high.

The low anode voltage makes it possible to couple the anode of the EF 86 direct to the control grid of the output tube. In order to ensure the correct grid bias for the output tube under these conditions, a fairly large cathode resistor is necessary; $560\ \Omega$, 5% tolerance, 3 W.

The screen-grid voltage for the input tube is taken from the cathode of the output tube, as already mentioned; this is done through a filter R_6, C_3 , of $22\text{ k}\Omega$ and $25\ \mu\text{F}$ respectively. This filter in conjunction with the large biasing resistor on the output tube produces considerable d.c. negative feedback, which stabilises the circuit against variations in the mains voltage and "spread" in tube characteristics within the usual production tolerances.

The serious distortion that usually accompanies single-tube output stages is for the most part neutralised by strong negative feedback, which includes the output transformer. This feedback is taken from the secondary side of the transformer to the undecoupled cathode of the input tube. The value of R_4 depends on the loudspeaker and output transformer used; it should be $82\ \Omega$ for a $15\ \Omega$ loudspeaker, $100\ \Omega$ for a $7\ \Omega$ speaker, $120\ \Omega$ for $5\ \Omega$, or $150\ \Omega$ for $3.75\ \Omega$. The primary of the transformer should match the loudspeaker to the $5\text{ k}\Omega$ load impedance of the output tube. The distortion is shown plotted against output power in Fig. 17.

The amplification attainable with these two tubes is not sufficient to permit of full tone control, but the higher frequencies can be cut by means of R_2 in series with C_1 to suppress noise. A form of bass control is included in the negative feedback network; this enables the bass response to be boosted to suit the characteristics of most gramophone recordings, in which the bass is suppressed. This control is by means of R_8 , shunted by C_6 : with R_8 set to its extreme position C_6 is short-circuited and the response curve is flat up to 1 kc/s. At the other end of R_8 the feedback at frequencies

below 1 kc/s is reduced and the bass response is boosted to a maximum of 15 dB at 70 c/s.

In general, negative feedback systems which are dependent on frequency are not recommended for high quality amplifiers, but in this instance, as the amplifier is a small one and the output not very great, it can quite successfully be used. In fact, tests carried out with a large number of amplifiers built to this circuit have given eminently satisfactory results.

The two potentiometers for volume and bass control are of the logarithmic type; the treble control is linear. If desired, the volume control and mains switch can be combined.

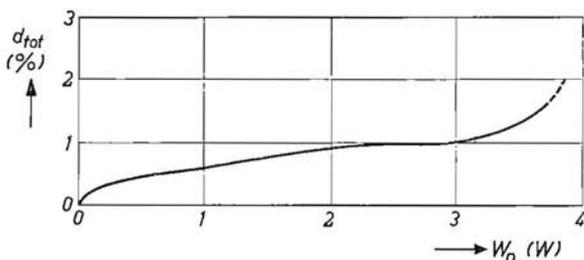


Fig. 17. Total harmonic distortion d_{tot} plotted against output power W_o , as measured on the 3 W amplifier at 400 c/s.

The response curve of this amplifier is reproduced in Fig. 18; with the two tone controls "off" the characteristic is seen to be flat (within 1 dB) from 20 c/s to 50 kc/s. Curves showing the maximum effects of the tone controls are also given; for these curves, one potentiometer was set to the minimum and the other to its maximum position.

Stability is satisfactory under all conditions; in this connection a 1 k Ω resistor is included across the secondary side of the output transformer to maintain stability should the loudspeaker be disconnected or if the load should be capacitive, e.g. with a long speaker lead.

The power supply presents no difficulties: the rectifier is the indirectly-heated EZ 80, with its heater connected to the same winding of the mains transformer as the amplifying tubes. Generally speaking, however, it is better to employ a separate winding for this purpose. If a single winding is to serve both purposes it should be capable of delivering 2 A at 6.3 V; separate windings should be rated at 1 A each. The H.T. is 2 x 300 V,

60 mA; but $2 \times 275 \text{ V}$ is also suitable, in which case R_{14} should be reduced to 100Ω . The primary should of course suit the available mains supply or may be tapped for 110, 127, 150, 220 and 240 V.

The anode voltage of the output tube is obtained from the reservoir capacitor C_9 , and the screen grid voltage from the first smoothing capacitor C_8 . The smoothing resistor is $3.9 \text{ k}\Omega$ and it need not be rated for more than $\frac{1}{2} \text{ W}$. The anode voltage for the input tube is smoothed by a $390 \text{ k}\Omega$ resistor R_9 , decoupled by a capacitor, C_5 , of $0.25 \mu\text{F}$.

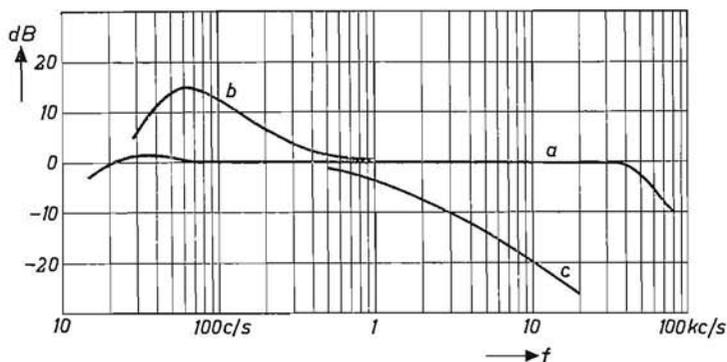


Fig. 18. Response curve (*a*), maximum bass boost (*b*) and maximum treble cut (*c*). The response is flat from 20 c/s to 30 kc/s with both controls set to minimum.

ANOTHER 3 W AMPLIFIER WITH ONLY ONE TUBE ECL 86

The gramophone amplifier using only one ECL 86 is the simplest design hitherto published, nevertheless it has remarkably good features. It is a design suitable for building into one unit together with a record player and a small loudspeaker, a so-called "electrophone". Elaborate tone controls and provision for use with magnetodynamic pick-ups would be out of place in such a simple design. The sensitivity is based on the use of either crystal or ceramic pick-up heads. The tone controls comprise treble cut and bass boost, the latter being incorporated in the feedback circuit.

General data

Tubes: ECL 86 pre-amplifier and output pentode

EZ 80 full-wave rectifier

Output: 3 W

Harmonic distortion: at 1000 c/s, output 3 W, 1.8 %

output 2 W, 0.9 %

output 1 W, 0.6 %

Frequency response: output 50 mW, — 3 dB points below 10 c/s
150 kc/s

Power response, output 3 W, — 3 dB points at 60 c/s
40 kc/s

Tone control: maximum treble cut at 10 kc/s 16 dB
at 15 kc/s 18 dB

maximum bass boost at 75 c/s 13 dB

Hum and noise levels with respect to full output — 70 dB

CIRCUIT DESCRIPTION

The typical features of this amplifier, its low distortion, its fine frequency response and its low hum and noise level, are attributable to the correct setting of the tube, the adequate smoothing and decoupling and the quality of the output transformer.

Fig. 19 shows the circuit diagram. The grid of the triode is connected to the volume control via potentiometer R_2 , which forms the treble control with C_1 . The triode section, which is equal to one section of an ECC 83, is set for maximum gain (about 70) at a supply voltage of 250 V, with an anode load resistance of 220 k Ω .

Under these conditions the total distortion is only 0.4 % at an output voltage of 3.2 V_{rms} which is the input required for the pentode section for maximum output (4 W at 10 % distortion).

The pentode section is also operated with anode and screen-grid voltages of 250 V. The load resistance is 7 k Ω , the required cathode resistor is 170 Ω . Since the latter value is not available in the standard range of resistors, one resistor of 470 Ω and one of 270 Ω (0.5 W) are connected in parallel. In the complete circuit these are the only resistors with a

tolerance of 5%; all other resistors used being of the $\pm 10\%$ tolerance range.

A negative feedback voltage is taken from the secondary of the output transformer via C_6R_9 and C_2R_3 and applied to the cathode of the triode

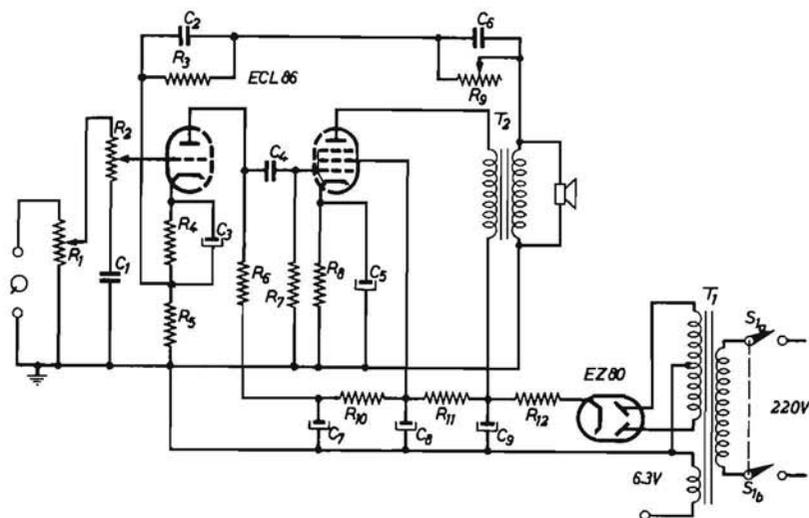


Fig. 19. Circuit diagram of the inexpensive amplifier using one ECL86.

$R_1 = 1 \text{ M}\Omega$, log. carbon potentiometer

$R_2 = 0.5 \text{ M}\Omega$, log. carbon potentiometer

$R_3 = 1.2 \text{ k}\Omega$, $\frac{1}{4} \text{ W}$

$R_4 = 1.8 \text{ k}\Omega$, $\frac{1}{4} \text{ W}$

$R_5 = 100 \Omega$, $\frac{1}{4} \text{ W}$

$R_6 = 220 \text{ k}\Omega$, $\frac{1}{4} \text{ W}$

$R_7 = 470 \text{ k}\Omega$, $\frac{1}{4} \text{ W}$

$R_8 = 470 \Omega$, $\frac{1}{2} \text{ W}$

$R_9 = 10 \text{ k}\Omega$, lin. carbon potentiometer

$R_{10} = 10 \text{ k}\Omega$, $\frac{1}{4} \text{ W}$

$R_{11} = 2.7 \text{ k}\Omega$, 5 W , wire wound

$R_{12} = 390 \Omega$, 5 W , wire wound

$C_1 = 150 \text{ pF}$, ceramic

$C_2 = 820 \text{ pF}$, ceramic

$C_3 = 100 \mu\text{F}$, electrolytic, 4 V wkg.

$C_4 = 0.1 \mu\text{F}$, polyester, 400 V wkg.

$C_5 = 100 \mu\text{F}$, electrolytic, 16 V wkg.

$C_6 = 0.22 \mu\text{F}$, polyester, 125 V wkg.

C_7

C_8 } = $3 \times 50 \mu\text{F}$, electrolytic, 350 V wkg.

C_9 }

$T_1 =$ supply transformer AD9027

$T_2 =$ output transformer AD9057

section. For this purpose, part of the cathode resistor is unbypassed. A negative feedback of 18 dB is thus obtained.

Bass boost is obtained by decreasing the feedback for the lower frequencies by means of C_6 and R_9 . If R_9 increases the low frequency feedback decreases.

Without feedback the required input signal for 3 W output would be of the order of 26 mV. With feedback the sensitivity is reduced to 300 mV, which is adequate for use with either ceramic or crystal pick-up heads. A higher sensitivity will make it easy to overdrive the amplifier when the volume control is set at maximum gain.

Output transformer

The high value of negative feedback, which is essential for obtaining the required low distortion and the wide frequency range, imposes the condition that an output transformer of outstanding quality is used. It must have low stray inductance and low winding capacitance, so that its resonant frequency is high, otherwise heavy feedback would lead to instability.

The transformer used has been specially developed for this amplifier; the type number is AD 9057. It is a transformer having a C-core of oriented laminated sheet. The primary winding is divided into two sections and the secondary is wound in between, so that the stray inductance is minimised. Moreover, the relatively high primary inductance of 10.5 H is obtained with a small number of turns so that the capacitances are small. The resulting resonant frequency is thus raised to such a value that a feedback of 18 dB, which is very high for an amplifier of this type, can be applied without risk of instability.

Supply

A standard type mains transformer type AD 9027 is used in combination with a thermionic rectifier EZ 80. The transformer is over-dimensioned, the total current being only 42.5 mA at a voltage of 276 V across C_9 . If the amplifier is to be used in an "electrophone", for example, semiconductor rectifiers in a bridge circuit may be a more attractive (but also more expensive) solution for the H.T. supply. The transformer can then be wound for about 240 V_{rms}, 50 mA, and a heater winding for 6.3 V, 0.7 A. The resistor R_{12} should be so chosen that the voltage across C_9

is 276 V.

A triple electrolytic capacitor of $3 \times 50 \mu\text{F}$ is used for smoothing and decoupling, together with the resistors R_{10} and R_{11} ; the circuit is entirely conventional. The on-off switch can be combined with the volume control without excessive hum being encountered.

MEASURED RESULTS

Voltage and currents

The following voltages and currents were measured in the experimental amplifier:

Supply

EZ80 cathode-to-earth	292 V		
V_{c9}	276 V		
V_{c8}	258 V	total current drain	42.5 mA
V_{c7}	250 V		

ECL86

pentode section

anode voltage	257 V	anode current	35.5 mA
screen-grid voltage	258 V	screen-grid current	6.0 mA
cathode voltage	7.4 V		

triode section

anode voltage	120 V	anode current	0.6 mA
cathode voltage	1.1 V		

Frequency response

The frequency response was measured at the secondary of the output transformer, (a) with the tone controls in the zero position, and (c and d) with these controls at maximum. The corresponding curves are shown in Fig. 20.

Power response

The power response at 3 W output was measured at the secondary of the output transformer, the result is shown by curve (b) in Fig. 20.

Harmonic distortion

The harmonic distortion was measured at the secondary of the output transformer, at a frequency of 1 kc/s; the result is plotted in Fig. 21. The

curve is typical for an amplifier with considerable feedback and shows that the amplifier should not be driven beyond an output power of 3.5 W.

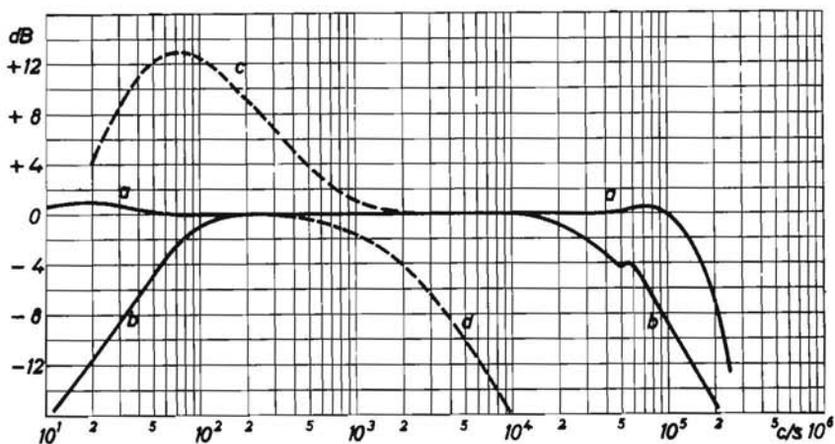


Fig. 20. Frequency response (a) and power response (b) measured at 3 W power output. Curve (c) shows the maximum bass boost, and curve (d) the maximum treble cut of the amplifier with one ECL86.

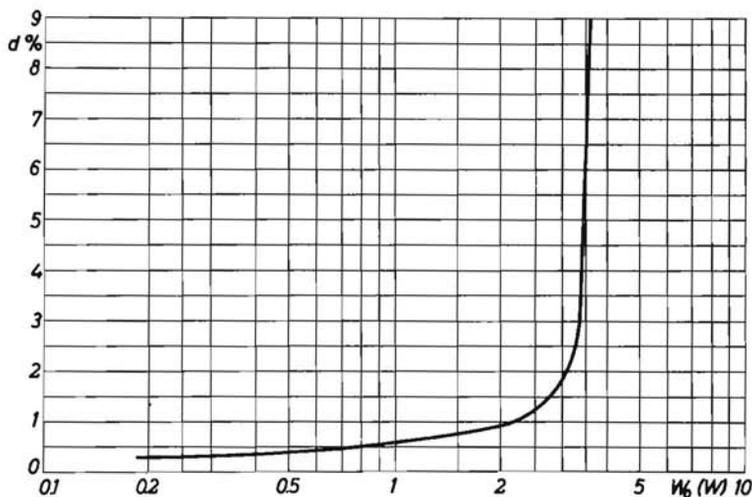


Fig. 21. Harmonic distortion of the amplifier measured at a frequency of 1 kc/s.

Stability

The loop gain and loop phase characteristics plotted in Figs 22 and 23 respectively show that the stability of the amplifier is adequate.

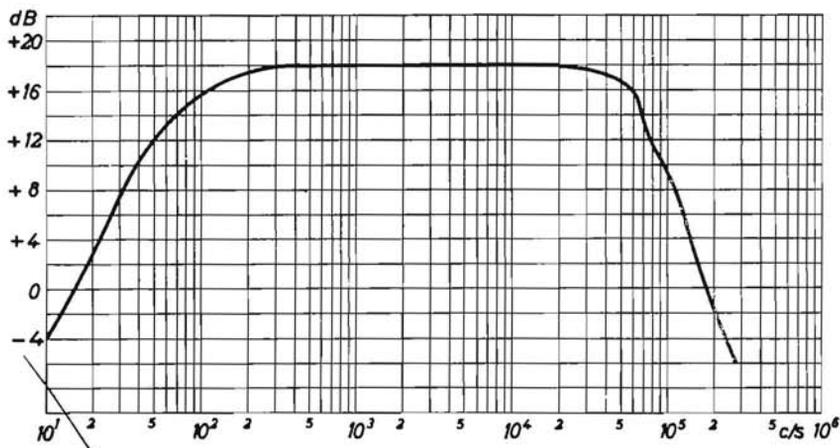


Fig. 22. Loop gain characteristic of the amplifier using one ECL86.

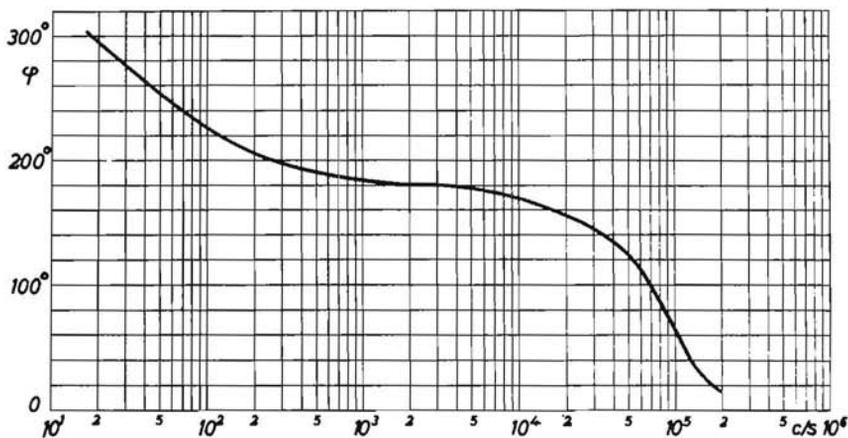


Fig. 23. Loop phase characteristic of the amplifier using one ECL86.

10 WATT HI-FI AMPLIFIER WITH TWO TYPE EL 84 PENTODES IN PUSH-PULL

The original version of this amplifier circuit*) was first published a few years ago, since when many leading radio and audio publications in the world have reprinted it, as this may be said to have been the first design to offer really high-fidelity reproduction at very reasonable cost.

Very many amplifiers based on this circuit have been built and much experience has therefore been gained, but this has not always been equally satisfactory. Some of them proved to be unstable at the higher frequencies, some at the lower and, again, a few at both ends of the frequency spectrum.

Apart from inefficient, or even careless assembly, this instability was in most cases found to be due to the use of output transformers which differed in certain respects from the original design. Some of these transformers had a high primary self-inductance; in others it was too low and in many cases the leakage inductance proved to be too high.

Tests have been made with a view to neutralising the effect of too much variation in the characteristics of the output transformer without detriment to the quality of the audio response, and these tests have proved very successful. Two modified circuits have been designed and have since gained considerable popularity; none of the prototypes has shown the least sign of instability. These two circuits will now be described.

DESCRIPTION OF CIRCUIT I

The circuit diagram is shown in Fig. 24. The output transformer used is the type AD 9009, having a primary inductance of 40 H, dropping to 28 H at 5mA d.c. This transformer has 7 Ω and 14 Ω matching for a primary load of 8000 Ω ; in the diagram the 7-ohm winding is shown in use with a 7-ohm loudspeaker type 9710 M. The circuit is intended for use with a crystal pickup, so that the tone control network precedes the input tube, the circuitry of which is conventional, giving a gain of about 200. A filter, R_{11} , C_8 is connected across the anode load resistor to reduce the amount of amplification at the higher frequencies and so ensure stability at those frequencies.

*) Cf. E. RODENHUIS, *Electronic Valves in A.F. Amplifiers*, Fig. 74; Philips Techn. Library 1954.

The phase inverter tube is the ECC 83, wired as a cathode coupled phase splitter; this arrangement ensures the least possible distortion, but low gain. It is, however, an advantage that this tube can be coupled direct to the input tube, since the grids of the double triode have to be positive in view of the considerable voltage drop across the cathode resistor, which is $68\text{ k}\Omega$. At the lower frequencies this stage operates with negative feedback, provided by R_{13} , C_{12} and the decoupling capacitor C_8 on the grid in the other section of the valve. This results in an extra load in parallel with R_{15} , which is accordingly somewhat larger than R_{14} : the respective values being $120\text{ k}\Omega$ and $100\text{ k}\Omega$.

The output stage is quite conventional, with a single decoupled biasing resistor to serve both tubes. Grid stoppers are included. A fairly small bypass capacitor ($40\text{ }\mu\text{F}$) is shunted to the biasing resistor to provide some attenuation at the lower frequencies and ensure stability at that end of the band. The negative feedback is taken from the secondary of the output transformer to the undecoupled part of the cathode resistor of the EF 86. For reasons of R.F. stability the $2.7\text{ k}\Omega$ feedback resistor is shunted by a 470 pF capacitor. If the 14-ohm matching of the loudspeaker is used, say for two 7-ohm loudspeakers, resistor R_{33} should be increased to $3.3\text{ k}\Omega$, with a 370 pF capacitor in parallel.

Sensitivity

The sensitivity of the amplifier is roughly 400 mV for 10 W output, which is suitable for a crystal pickup. That part of the circuit which is shown within the dotted lines may be replaced by a $1\text{ M}\Omega$ resistor if the amplifier is used with a separate pre-amplifier containing the controls. In this case the sensitivity may be suitable for a magneto-dynamic pickup, microphone or tape head. Without the controls the sensitivity is 40 mV to give 10 W .

Response curve and tone control

The response curve is flat from about 10 c/s to at least $30,000\text{ c/s}$ with the controls set to their central positions. With the controls set for maximum and minimum effect the curves are as shown in Fig. 25.

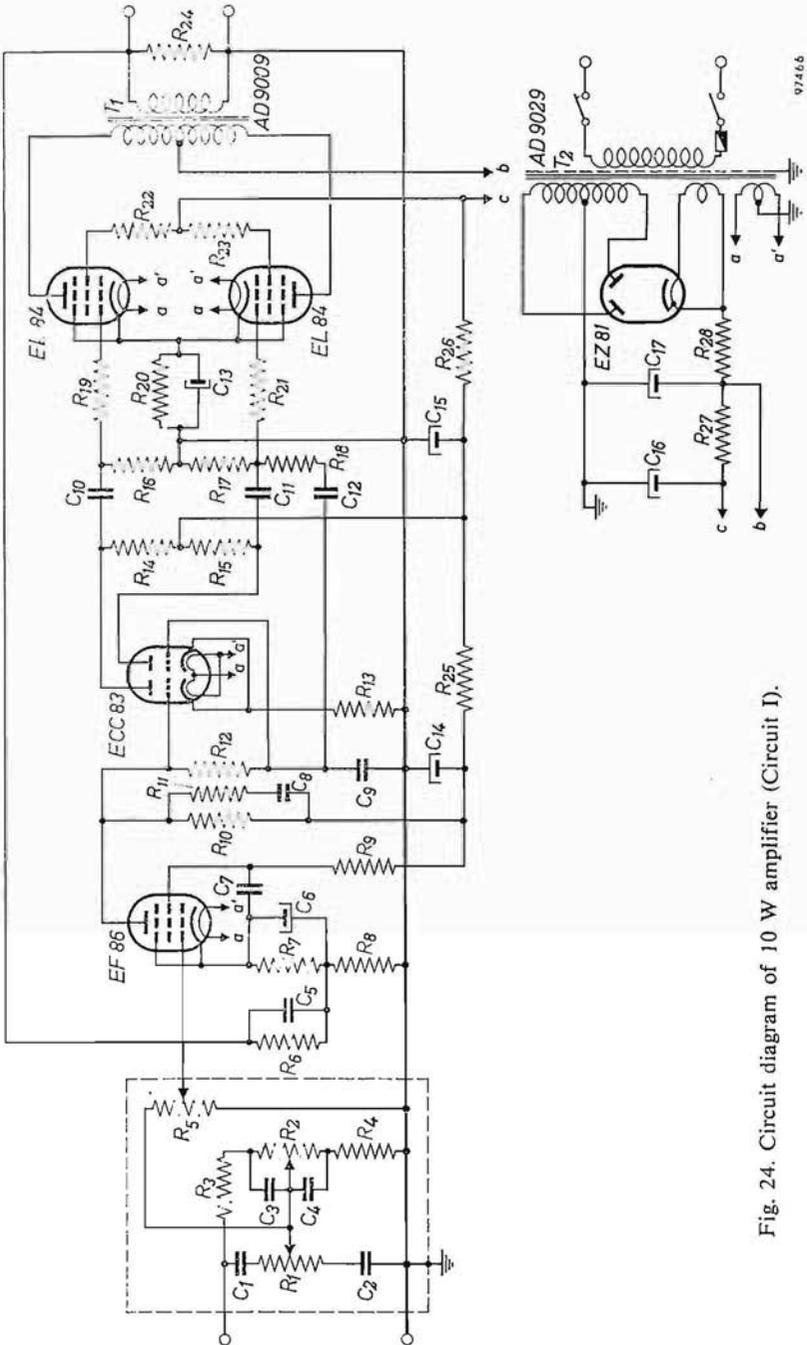


Fig. 24. Circuit diagram of 10 W amplifier (Circuit I).

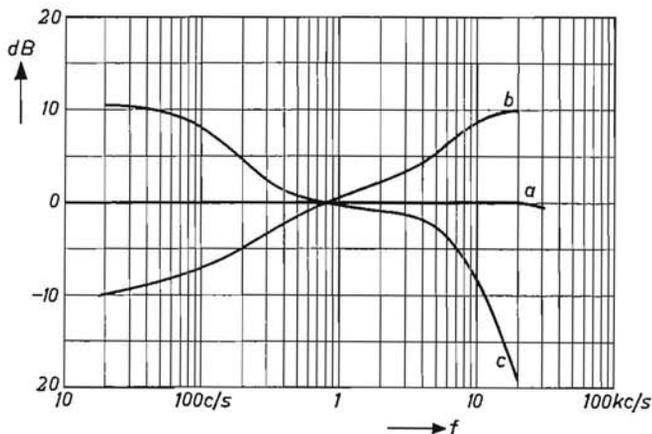


Fig. 25. Frequency response curve and effect of the tone controls of the 10 W amplifier: (a) both controls set to central position, (b) maximum bass cut and maximum treble boost, (c) maximum bass boost and maximum treble cut.

List of parts

$$R_1 = 2 \text{ M}\Omega \text{ log.}$$

$$R_2 = 2 \text{ M}\Omega \text{ log.}$$

$$R_3 = 1.5 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_4 = 0.15 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_5 = 1 \text{ M}\Omega \text{ log.}$$

$$R_6 = 2.7 \text{ k}\Omega \pm 5\%, \frac{1}{4} \text{ W}^1)$$

$$R_7 = 2.2 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_8 = 10 \Omega \pm 5\%, \frac{1}{2} \text{ W}$$

$$R_9 = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_{10} = 0.18 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}^1)$$

$$R_{11} = 22 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_{12} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_{13} = 68 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$$

$$R_{14} = 0.1 \text{ M}\Omega \pm 10\%, \frac{1}{2} \text{ W}$$

$$R_{15} = 0.12 \text{ M}\Omega \pm 10\%, \frac{1}{2} \text{ W}$$

$$R_{16} = 0.33 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_{17} = 0.33 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_{18} = 0.56 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$$

$$R_{19} = 1 \text{ k}\Omega \pm 20\%, \frac{1}{4} \text{ W}$$

$$R_{20} = 135 \Omega \pm 5\%, 1 \text{ W}$$

$$R_{21} = 1 \text{ k}\Omega \pm 20\%, \frac{1}{4} \text{ W}$$

$$R_{22} = 220 \Omega \pm 20\%, \frac{1}{2} \text{ W}$$

$$R_{23} = 220 \Omega \pm 20\%, \frac{1}{2} \text{ W}$$

$$R_{24} = 1 \text{ k}\Omega \pm 20\%, \frac{1}{2} \text{ W}$$

$$R_{25} = 0.1 \text{ M}\Omega \pm 20\%, \frac{1}{4} \text{ W}$$

$$R_{26} = 27 \text{ k}\Omega \pm 20\%, \frac{1}{2} \text{ W}$$

$$R_{27} = 1.5 \text{ k}\Omega \pm 20\%, 2 \text{ W}^2)$$

$$R_{28} = 220 \Omega \pm 20\%, 6.5 \text{ W}^2)$$

$$C_1 = 33 \text{ pF} \pm 10\%$$

$$C_2 = 680 \text{ pF} \pm 10\%$$

$$C_3 = 270 \text{ pF} \pm 10\%$$

$$C_4 = 3300 \text{ pF} \pm 10\%$$

$$C_5 = 470 \text{ pF} \pm 10\%$$

$$C_6 = 100 \mu\text{F}, 13 \text{ V wkg}$$

$$C_7 = 47 \text{ 000 pF} \pm 20\%$$

$$C_8 = 150 \text{ pF} \pm 10\%$$

$$C_9 = 0.1 \mu\text{F} \pm 20\%$$

$$C_{10} = 0.1 \mu\text{F} \pm 20\%$$

$$C_{11} = 0.1 \mu\text{F} \pm 20\%$$

$$C_{12} = 47 \text{ 000 pF} \pm 20\%$$

$$C_{13} = 50 \mu\text{F}, 25 \text{ V wkg}$$

$$C_{14} \left. \begin{array}{l} \\ \\ \end{array} \right\} 2 \times 50 \mu\text{F}, 350 \text{ V wkg}$$

$$C_{15} \left. \begin{array}{l} \\ \\ \end{array} \right\} 2 \times 50 \mu\text{F}, 350 \text{ wkg}$$

$$C_{16} \left. \begin{array}{l} \\ \\ \end{array} \right\} 2 \times 50 \mu\text{F}, 350 \text{ wkg}$$

$$C_{17} \left. \begin{array}{l} \\ \\ \end{array} \right\} 2 \times 50 \mu\text{F}, 350 \text{ wkg}$$

1) High stability.

2) Wire-wound.

Distortion

The amount of distortion is very small, 0.5% at 400 c/s for 10 W output. Measured at 40 c/s the distortion is 0.8% at 10 W. The intermodulation distortion was measured on 40 c/s and 10 kc/s signals at an amplitude ratio of 4 : 1, and the equivalent output power was found to be 8.2 W for 2% intermodulation distortion which, to the expert ear represents the limit of audibility. Curves depicting the distortion are shown in Fig. 26.

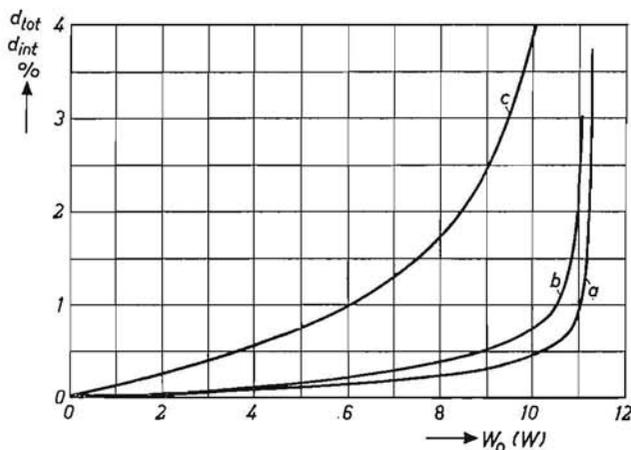


Fig. 26. Total harmonic distortion d_{tot} , measured (a) at 400 c/s and (b) 40 c/s. Intermodulation distortion d_{int} (curve c) measured at 40 c/s and 10 kc/s with an amplitude ratio of 4 : 1. Harmonic distortion is given as a function of the output power; intermodulation distortion is plotted against equivalent output power.

CIRCUIT DESCRIPTION II

Fig. 27 shows the circuit that has been developed for use in cases where output transformers are available having a higher primary inductance than that specified in the original design.

The overall gain was in many cases more than 26 dB, the value originally quoted. The average proved to be 29 dB and this involved some risk of instability. Therefore in the first place the gain in the input stage was reduced to give an overall amplification of 26 dB and thus improve stability. Further, the output tubes have been given separate biasing resistors decoupled by

40 μF capacitors, to reduce amplification at the lower frequencies, whilst the time constant of the screen-grid decoupling for the input tube has been reduced, also with a view to improving stability at the lower frequencies.

Stability at the higher frequencies, particularly when the loading is capacitive, e.g. with a long loudspeaker lead, has been increased by changing the time constant of $R_{11}C_8$ in the anode circuit and R_9C_7 in the screen-grid circuit of the input tube.

The time constant of C_5R_6 in the negative feedback line is 4×10^{-6} . Generally speaking, this will be correct but, depending upon the output transformer used, another value for C_5 might be found more suitable.

If the amplifier has been carefully assembled it should be possible to increase the negative feedback by 6 dB by reducing the value of R_6 , without any symptoms of instability. This is, in fact, a simple test of the stability of the completed amplifier.

Tests have been carried out on this amplifier using a large number of different output transformers, having primary inductance values of from 40 H to more than 100 H, with a leakage inductance of less than 30 mH. It was found that complete stability was maintained even with a capacitive load and open circuited input. Here the 1 k Ω resistor across the secondary winding of the output transformer plays an important part.

In this amplifier, too, the cathode-coupled phase inverter circuit is used, but with the negative feedback omitted; this necessitates resistors R_{14} and R_{15} being matched to within 5%, the one with the higher value being used for R_{15} . It is also important to keep wiring capacitances in the anode circuits of the phase inverter equal as far as possible, which means that the resistors, and coupling capacitors C_{10} and C_{11} , as well as the grid resistors on the output tubes, R_{16} , R_{17} , R_{18} and R_{21} should be placed at about the same distance from the chassis and should have leads of the same length. If this is done, the balance of the control signals for the output stage will not be upset at the higher frequencies, and distortion will be kept to a minimum.

We have already mentioned that separate biasing resistors are used for the output tubes. The advantage of this is that the d.c. negative feedback in the output stage will be exactly the same for each tube and any discrepancies between the characteristics of these tubes compensated, particularly in the event of differences arising in their performance during life. There is no need to select matched tubes, either from the start or when renewing them.

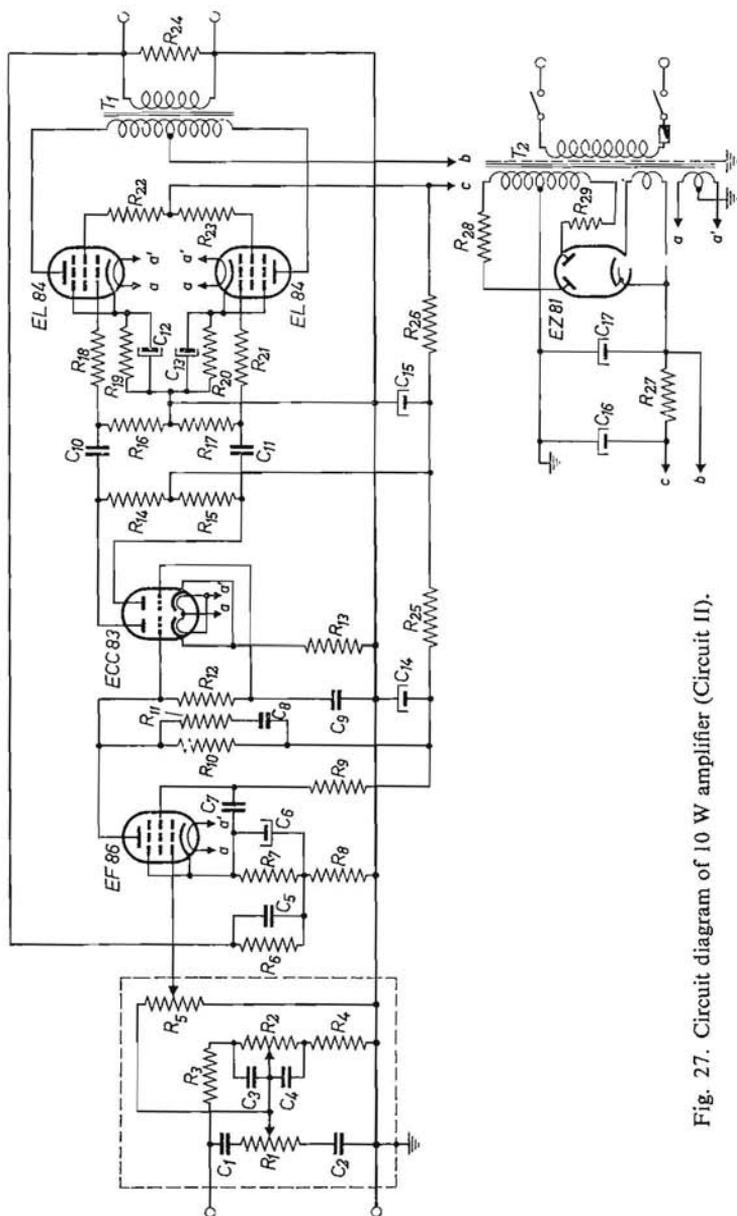


Fig. 27. Circuit diagram of 10 W amplifier (Circuit II).

The amount of compensation is sufficient to cancel out the usual differences, within the tolerances, likely to occur in normal tube production.

The sensitivity of this amplifier is the same as that of the circuit shown in Fig. 24; the curve is given in Fig. 28.

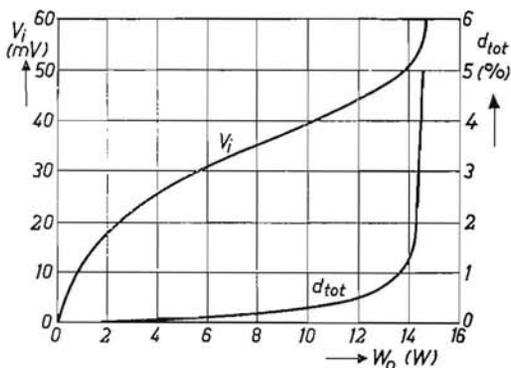


Fig. 28. Total distortion d_{tot} and input voltage V_i plotted against the output W_o of the modified 10 W amplifier at a frequency of 400 c/s.

List of parts

$R_1 = 2 \text{ M}\Omega$, log.	$R_{23} = 47 \Omega \pm 20\%$, $\frac{1}{4} \text{ W}$
$R_2 = 2 \text{ M}\Omega$, log.	$R_{24} = 1 \text{ k}\Omega \pm 20\%$, $\frac{1}{4} \text{ W}$
$R_3 = 1.5 \text{ M}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$R_{25} = 100 \text{ k}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$
$R_4 = 150 \text{ k}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$R_{26} = 33 \text{ k}\Omega \pm 10\%$, $\frac{1}{2} \text{ W}$
$R_5 = 1 \text{ M}\Omega$, log.	$R_{27} = 1.2 \text{ k}\Omega \pm 10\%$, 1 W
$R_6 = 3.75 \Omega$ matching $15 \text{ k}\Omega$	$R_{28} = \text{dependent on } R_{tr}$
7Ω matching $27 \text{ k}\Omega$	$R_{29} = \text{dependent on } R_{tr}$
15Ω matching $33 \text{ k}\Omega$	$C_1 = 33 \text{ pF} \pm 10\%$
$R_7 = 1.8 \text{ k}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$C_2 = 680 \text{ pF} \pm 10\%$
$R_8 = 100 \Omega \pm 5\%$, $\frac{1}{4} \text{ W}^{1)}$	$C_3 = 270 \text{ pF} \pm 10\%$
$R_9 = 470 \text{ k}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}^{1)}$	$C_4 = 3300 \text{ pF} \pm 10\%$
$R_{10} = 100 \text{ k}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}^{1)}$	$C_5 = 270 \text{ pF} \pm 5\%$; $3,75$ matching
$R_{11} = 10 \text{ k}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$150 \text{ pF} \pm 5\%$; 7 matching
$R_{12} = 1 \text{ M}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$120 \text{ pF} \pm 5\text{K}$; 15 matching
$R_{13} = 68 \text{ k}\Omega \pm 10\%$, 1 W	$C_6 = 100 \mu\text{F}$, $12,5 \text{ V}$ wkg
$R_{14} = 100 \text{ k}\Omega \pm 10\%$, $\frac{1}{2} \text{ W}$	$C_7 = 47 \text{ 000 pF}$, 350 V d.c. wkg
$R_{15} = 100 \text{ k}\Omega \pm 10\%$, $\frac{1}{2} \text{ W}$	$C_8 = 330 \text{ pF} \pm 10\%$
$R_{16} = 0.82 \text{ M}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$C_9 = 0.1 \mu\text{F}$, 350 V d.c. wkg
$R_{17} = 0.82 \text{ M}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$C_{10} = 0.1 \mu\text{F}$, 350 V d.c. wkg
$R_{18} = 4.7 \text{ k}\Omega \pm 10\%$, $\frac{1}{4} \text{ W}$	$C_{11} = 0.1 \mu\text{F}$, 350 V d.c. wkg
$R_{19} = 270 \Omega \pm 5\%$, 2 W (normal loading)	$C_{12} = 50 \mu\text{F}$, 25 V wkg
$437 \Omega \pm 5\%$, 2 W (low loading)	$C_{13} = 50 \mu\text{F}$, 25 V wkg
$R_{20} = 270 \Omega \pm 5\%$, 2 W (normal loading)	$C_{14} \left\{ \begin{array}{l} 2 \times 8 \mu\text{F}, 350 \text{ V} \\ C_{15} \end{array} \right.$
$437 \Omega \pm 5\%$, 2 W (low loading)	$C_{16} \left\{ \begin{array}{l} 2 \times 50 \mu\text{F}, 350 \text{ V wkg} \\ C_{17} \end{array} \right.$
$R_{21} = 4.7 \text{ k}\Omega \pm 20\%$, $\frac{1}{4} \text{ W}$	
$R_{22} = 47 \Omega \pm 20\%$, $\frac{1}{4} \text{ W}$	

1) High stability.

Response curve and phase shift

Typical curves for the response, the loop gain and the phase shift of the amplifier are given in Fig. 29. These curves were obtained with an $8\text{ k}\Omega$

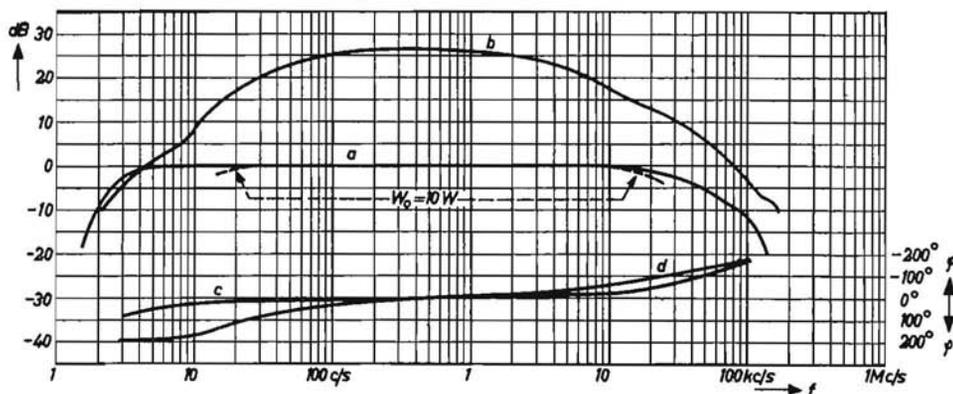


Fig. 29. Frequency response and output curve (a), loop gain (b), phase shift of the amplifier (c) and of feedback loop (d), plotted against the frequency, for the modified 10 W amplifier (circuit II).

load between the anodes of the output tubes, and they have been extended beyond 10 c/s in order to demonstrate the excellent stability.

Distortion

The total harmonic distortion plotted against the output power is shown in Fig. 28. Fig. 30 illustrates the intermodulation distortion measured with frequencies of 40 c/s and 10 kc/s, and also of 70 c/s and 7 kc/s, the amplitude ratio being 4 : 1 in both cases, with the lower frequency as the stronger of the two.

Beat-note was measured at 9 and 10 kc/s, as well as 14 and 15 kc/s, using signals of equal strength. The percentage of the 1 kc/s component in the output signal is also seen from Fig. 31.

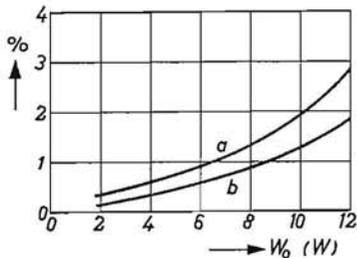


Fig. 30. Intermodulation distortion as a function of the equivalent output power as measured on signals of 40 c/s and 10 kc/s (*a*), and 70 c/s and 7 kc/s (*b*), both at an amplitude ratio of 4 : 1. 10 W amplifier; circuit II.

Low loading

In the first chapter mention is made of low loading. On normal load the working points of the output tubes are as given in the published data for class AB operating conditions in accordance with actual measurements made by the manufacturers of the tubes. The cathode resistor is 270Ω for each of the output tubes, or 130Ω in a common resistor. In either case the anode load is $8 \text{ k}\Omega$, the anode current being $2 \times 36 \text{ mA}$.

Under the low loading conditions an anode load of only $6 \text{ k}\Omega$ is used, in this case the cathode resistors are 437Ω ($390 + 47 \Omega$) the anode current then being $2 \times 24 \text{ mA}$.

On both speech and music the amplifier works almost as if fixed grid bias were used (see Fig. 1). The maximum output of an amplifier is actually made use of only for short passages; the average output power is comparatively low. Even so, the maximum amount of power available must be large enough to be able to handle the widely different sound levels occurring in music. Low loading results in less distortion of the peak sounds than does normal loading.

At the same time, the improvement in response is not readily noticeable to the ear, especially as it usually affects only brief passages. The ear needs to be trained to appreciate such subtle differences in distortion of music reproduced on Hi-Fi equipment.

The anode currents, too, are less with low loading than for normal loading, the dissipation of each tube being 7.5 W as against 11 W or well below the maximum rated dissipation of 12 W for the output tubes. There is also the advantage that a smaller output transformer can be used, and there is less ripple current and hence less hum.

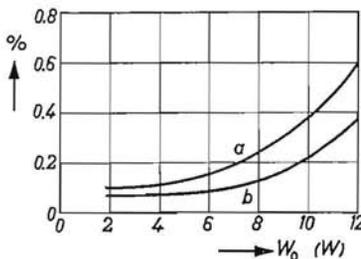


Fig. 31. Beat-note distortion plotted against equivalent output power on 14 kc/s and 15 kc/s (a), and 9 kc/s and 10 kc/s (b) measured with signals of equal amplitude.

Under the low loading conditions the peak currents are greater than ordinary class AB conditions, but on both speech and music these peaks are of very short duration and are easily taken care of by the $50 \mu\text{F}$ smoothing capacitor.

A disadvantage of low loading operation is that the amplifier cannot be tested fully driven with a sinusoidal signal, and the distortion, therefore, cannot be measured on full load. A sinusoidal signal can be employed for the purpose of plotting the response curve, but then the output must not exceed 1 to $1\frac{1}{2}$ W; beyond that considerable distortion is introduced. Although measurements can be taken with the aid of a square-wave voltage of constant amplitude, as previously mentioned, the trained and discriminating listener is well able to perceive any slight distortion present and this, after all, is the ultimate test of any amplifier.

Ultra-linear conditions

This amplifier has also been tested with an output transformer suitable for ultra-linear operation with a tapping at 43% of the total number of turns counting from the centre-tap. Capacitor C_5 (for the negative feedback) has to be 20% smaller than as given in Fig. 27. Comparable values for the working conditions of the three different types of output are given below viz. A: normal class AB push-pull, B: ultra-linear output and C: low loading.

	A	B	C
Nominal output power	10	10	10 W
Overload point	14	11	14 W
Sensitivity at volume control set for nominal output	40	40	— mV

Harmonic distortion (10 W, 400 c/s)	0.3	0.1	— %
Intermodulation distortion (10 W, 40 c/s — 10 kc/s, 4 : 1)	2	1	— %
Beat-note distortion (10 W)			
9 and 10 kc/s	0.25	0.25	— %
14 and 15 kc/s	0.4	0.33	— %
Loop gain 1000 c/s	26	20.5	— dB

Triode input for use with separate pre-amplifier

When the 10 W amplifier is to be used with a separate pre-amplifier for operation with magneto-dynamic pick-ups and/or microphones, for example, it is advantageous to use the EF 86 pre-amplifier connected as a triode. The screen-grid is strapped to the anode so that the resistor R_9 and the capacitor C_7 can be omitted (Fig. 27). Other circuit modifications are: R_{11} is reduced from 10 k Ω to 3.9 k Ω , R_{25} of 100 k Ω should be made 180 k Ω .

In the feedback circuit the following applies:

for 3.75 Ω matching	$R_6 = 3.3$ k Ω	$C_5 = 1200$ pF,
for 7 Ω matching	$R_6 = 5.6$ k Ω	$C_5 = 700$ pF,
for 15 Ω matching	$R_6 = 6.8$ k Ω	$C_5 = 580$ pF.

The input circuit with the tone and volume controls may be omitted and replaced by a resistor of 1 M Ω , provided that the volume and tone control circuits are inserted in the pre-amplifier. The sensitivity for full drive is reduced to 200 mV; the hum and noise levels are considerably lower than in the original circuit.

HI-FI AMPLIFIER USING TWO TUBES ECL 86 IN PUSH-PULL CLASS AB

The best Hi-Fi amplifier designs are those in which the successive stages are well matched to each other so that no compromises or special measures are necessary to adapt the performance of one stage to that of following. Moreover, the number of tubes should be kept small, because every tube contributes to the distortion; therefore each tube should be adjusted for a high gain at low distortion.

The overall characteristics, such as power response and frequency response, should show a gradual roll-off at frequencies well beyond the audible range. The intermodulation distortion should be low.

Such a well-balanced design is described below. With only four tubes sensitivity is sufficient to operate the amplifier with magneto-dynamic pick-ups heads; R.I.A.A. frequency correction and a "Baxandall" *) tone control circuit have been incorporated.

General data

Tubes:	EF86	pre-amplifier		
	ECC82	second pre-amplifier and tone control		
	2 x ECL86	main amplifier		
	EZ81	full-wave rectifier		
Power output			max.	10 W
			nom.	9 W
Harmonic distortion at an output of:			10 W	9 W
and a frequency of:	60 c/s		0.75	0.3 %
	1 kc/s		0.4	0.2 %
	10 kc/s		0.75	0.4 %
Intermodulation distortion				
40 c/s and 10 kc/s ratio of amplitude 4 : 1			1.8	0.7 %
Sensitivity at 10 W output				
I. tuner				350 mV
II. tape recorder reproduction				1.5 V
recording				5 mV

*) Ref. Negative-Feedback Ton Control, by P. J. Baxandall B.Sc. (Eng.); Wireless World, October 1952, p. 402-405.

III. crystal pick-up	45 mV
IV. magneto-dynamic pick-up	7 mV
Tone control: referred to 1 kc/s	
bass at 50 c/s	cut: 16 dB
	boost: 14 dB
treble at 10 kc/s	cut: 10 dB
	boost: 10 dB
Frequency response: within 1 dB from 20 c/s to 50 kc/s	
Power response: at 9 W flat from 44 c/s to 15 kc/s	
Negative feedback:	20 dB
Damping factor	14.5
Hum and noise, referred to 10 W power output	
input circuits I, II and III	— 70 dB
input circuit IV	— 55 dB
Stability:	
Capacitance across output terminals at which oscillations start, with a resistive load	56 nF
loudspeaker loading	12 nF

DESIGN CONSIDERATIONS

One of the problems in Hi-Fi amplifier design is a tone control circuit in which the "flat" frequency response can easily be set. In this respect the Baxandall circuit has many advantages: (1) bass and treble control is carried out with linear potentiometers, (2) the central point of those potentiometers is the "flat" position, (3) a control range, both for bass and treble controls, of + 15 dB and - 15 dB can easily be obtained, (4) in stereo amplifiers, where ganged potentiometers are used, superior ganging is obtained with linear potentiometers than with logarithmic types, (5) the distortion is low due to the high negative feedback.

All this seems ideal, but this circuit also has some disadvantages. The treble control should be equipped with a centre-tapped potentiometer, which is not commonly available. However, an artificial centre tapping, made with two identical resistors, shunted across a normal potentiometer, gives adequate results.

Moreover, the output impedance of the first tube should be low. A

cathode follower could be ideal in this respect, but this would require an additional tube. A high-gain tube such as the ECC83, has an output impedance of about $62\text{ k}\Omega$, so that this type is not very attractive for the purpose, notwithstanding its high gain. Preference was therefore given to the ECC82, which has an output impedance of about $7\text{ k}\Omega$ at a gain of 10 to 11.

According to the tube data, two pentode sections of the ECL86 operated at a supply voltage of 300 V , can deliver an output of 14.3 W to the primary of the output transformer at 5% distortion. Experiments with several circuits and output transformers revealed that in low-distortion feedback amplifiers the maximum secondary output that could be obtained was of the order of 10 to 11 W . Under these conditions the input signal should be approximately 6.5 V_{rms} . Investigation of various phase splitter circuits with triode sections and 20 dB overall feedback, revealed that an input signal of about 1.5 V is necessary for full drive. Such a sensitivity was considered too low, the aim being 0.5 V for full drive so that the main amplifier could be used with the tone control circuit described. The input sensitivity is then 50 mV to 60 mV which is adequate for use with modern, high quality, ceramic and crystal pick-up heads.

The required sensitivity could be obtained by applying both positive and negative feedback in the main amplifier. Due to the positive feedback in the phase splitter, the gain could be increased to 175, without effects of instability and attenuation at the high frequencies. In this way the circuit was made suitable for use with various input sources (tape recorder, radio tuner, pick-up heads). However, the sensitivity is not adequate for use with magneto-dynamic pick-up heads and/or microphones.

To comply with the requirement to use such pick-up heads, a pre-amplifier stage with an EF86 is used in a frequency-dependent feedback circuit for R.I.A.A. correction. It was considered inopportune to provide a microphone input. Microphones are usually employed with relatively long cables, so that it becomes general practice to incorporate a pre-amplifier in the stand. This makes an additional sensitive input channel in the amplifier superfluous, microphone amplifiers can be connected to one of the available inputs. This solution enables the use of a simple selector switch, which precedes the tone control stage.

CIRCUIT DESCRIPTION

The complete circuit diagram of the amplifier is shown in Fig. 32. There are four input circuits, (*M*) for magneto-dynamic pick-up heads, (*C*) for ceramic and crystal pick-up heads, (*TR*) for tape recorder input and output, and (*R*) for a tuner. The inputs are selected by means of the four-position switch, which is provided with a short-circuiting section. In position 1, the grid of the EC(C)82 is connected to the anode of the EF86 via coupling capacitor C_5 . The feedback network is also connected to this point. R_4 is the feedback resistor proper. Of the other components, C_1 introduces an attenuation which increases with the frequency in excess of 1 kc/s. C_2 increases the gain for even lower frequencies. R_5 limits the gain at the lowest frequencies.

The feedback reduces the input impedance of the EF86 to about 4 k Ω . Therefore R_1 (68 k Ω) has been connected in series, so that the correct load resistance for the majority of magneto-dynamic pick-up heads is obtained. With the selector switch in position 2, the input channel for crystal pick-up heads is connected to the grid of the EC(C)82. The load resistance is then 1 M Ω . For both pick-up inputs female plugs according to the new I.E.C. standards were used. These are in such a way that either mono pick-up heads or stereo pick-up heads can be connected to the same amplifier input.

Tape recorders are connected to the amplifier via a five-prong plug according to the DIN standard. Several modern tape recorders are provided with this type of plug and with the corresponding cable. The upper prong is the input, the lower one the output for recording. With the voltage divider R_8R_9 the sensitivity is adjusted to 1.5 V, a value in accordance with several modern tape recorders, beyond the equaliser. Should the signal at the recorder output be lower, then the correct sensitivity can be obtained by changing the ratio of R_8 to R_9 .

Finally, in the fourth position, a tuner can be connected to the input of the EC(C)82. R_{12} and R_{13} reduce the sensitivity to a value of 350 mV. C_{26} is shunted across R_{12} to compensate for the attenuation at high frequencies. The output signal of the EC(C)82 is fed to the tone-control network and to the recording terminal for the tape recorder via C_7 , R_{10} and R_{11} reduce the output voltage to about 5 mV, which is sufficient for recording purposes.

List of Parts

- $R_1 = 68 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_2 = 100 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_3 = 2.2 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_4 = 470 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_5 = 10 \text{ M}\Omega, \frac{1}{4} \text{ W}$
 $R_6 = 1.2 \text{ M}\Omega, \frac{1}{4} \text{ W}$
 $R_7 = 220 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_8 = 1 \text{ M}\Omega, \frac{1}{4} \text{ W}$
 $R_9 = 47 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{10} = 10 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{11} = 1 \text{ M}\Omega, \frac{1}{4} \text{ W}$
 $R_{12} = 820 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{13} = 100 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{14} = 1 \text{ M}\Omega, \frac{1}{4} \text{ W}$
 $R_{15} = 1.2 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{16} = 47 \text{ k}\Omega, \frac{1}{2} \text{ W}$
 $R_{17} = 390 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{18} = 100 \text{ k}\Omega,$
 $R_{19} = 1 \text{ M}\Omega, \text{ carbon potentiometer lin.}$
 $R_{20} = 100 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{21} = 220 \text{ k}\Omega,$
 $R_{22} = 1 \text{ M}\Omega, \text{ carbon potentiometer lin.}$
 $R_{23} = 390 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{24} = 2.2 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{25} = 100 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{26} = 500 \text{ k}\Omega, \text{ carbon potentiometer log.}$
 $R_{27} = 1.5 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{28} = 120 \Omega, \frac{1}{4} \text{ W}$
 $R_{29} = 220 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{30} = 15 \text{ k}\Omega, \frac{1}{2} \text{ W}$
 $R_{31} = 1 \text{ M}\Omega, \frac{1}{4} \text{ W}$
 $R_{32} = 1 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{33} = 15 \text{ k}\Omega, \frac{1}{2} \text{ W}$
 $R_{34} = 470 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{35} = 470 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{36} = 5.6 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{37} = 5.6 \text{ k}\Omega, \frac{1}{4} \text{ W}$
 $R_{38} = 270 \Omega, \frac{1}{2} \text{ W}$
 $R_{39} = 270 \Omega, \frac{1}{2} \text{ W}$
 $R_{40} = 2.2 \text{ k}\Omega, \frac{1}{4} \text{ W}, 7 \Omega \text{ matching}$
 $3.3 \text{ k}\Omega, \frac{1}{4} \text{ W}, 14 \Omega \text{ matching}$
- $R_{41} = 5.6 \text{ k}\Omega, \frac{1}{2} \text{ W}$
 $R_{42} = 5.6 \text{ k}\Omega, \frac{1}{2} \text{ W}$
 $R_{43} = 1.2 \text{ k}\Omega, 5 \text{ W, wire wound}$
 $R_{44} = 39 \Omega, 5 \text{ W, wire wound}$
- $C_1 = 200 \text{ pF, ceramic, 2\% tolerance}$
 $C_2 = 560 \text{ pF, ceramic, 2\% tolerance}$
 $C_3 = 0.1 \mu\text{F, polyester, 400 V wkg.}$
 $C_4 = 100 \mu\text{F, electrolytic, 4 V wkg.}$
 $C_5 = 22 \text{ nF, polyester, 400 V wkg.}$
 $C_6 = 100 \mu\text{F, electrolytic, 4 V wkg.}$
 $C_7 = 0.22 \mu\text{F, polyester, 400 V wkg.}$
 $C_8 = 3.3 \text{ nF, ceramic, 10\% tolerance}$
 $C_9 = 3.3 \text{ nF, ceramic, 10\% tolerance}$
 $C_{10} = 100 \text{ pF, ceramic, 10\% tolerance}$
 $C_{11} = 0.1 \mu\text{F, polyester, 125 V wkg.}$
 $C_{12} = 0.1 \mu\text{F, polyester, 400 V wkg.}$
 $C_{13} = 100 \mu\text{F, electrolytic, 4 V wkg.}$
 $C_{14} = 25 \mu\text{F, electrolytic, 25 V wkg.}$
 $C_{15} = 0.1 \mu\text{F, polyester, 400 V wkg.}$
 $C_{16} = 150 \text{ pF, ceramic, 10\% tolerance}$
 $C_{17} = 0.22 \mu\text{F, polyester, 400 V wkg.}$
 $C_{18} = 0.22 \mu\text{F, polyester, 125 V wkg.}$
 $C_{19} = 100 \mu\text{F, electrolytic, 16 V wkg.}$
 $C_{20} = 100 \mu\text{F, electrolytic, 16 V wkg.}$
 $C_{21} = 680 \text{ pF, } 7 \Omega \text{ matching}$
 $390 \text{ pF, } 14 \Omega \text{ matching}$
 $C_{22} = 16 \mu\text{F, electrolytic, 350 V wkg.}$
 $C_{23} \left. \vphantom{C_{23}} \right\}$
 $C_{24} \left. \vphantom{C_{24}} \right\} = 3 \times 50 \mu\text{F, electrolytic, 350 V wkg.}$
 $C_{25} \left. \vphantom{C_{25}} \right\}$
 $C_{26} = 10 \text{ pF, ceramic}$
- Transformers*
 $T_1: \text{AD9026}$
 $T_2: \text{AD9058}$
- Input terminals*
 $M. C. R.: \text{standard I.E.C. (five prong)}$
 $TR : \text{standard DIN (five prong)}$

The tone-control network is so dimensioned that the gain of the E(C)C82 is roughly unity. As in all sections of a pre-amplifier, the wiring of this stage of the circuit influences the performance. Long screened leads or screened leads with a high capacitance between core and screen should be avoided, especially to the grids of the E(C)C 82 and to the volume control. The coupling capacitor C_{11} is connected to the junction of R_{20} , C_{12} , R_{22} and R_{23} . No direct voltage thus being across the capacitor, a type with a low working voltage and small dimensions can be used.

The volume control is at the input of the main amplifier. This has the advantage that the component which is the main cause of noise in amplifiers, is connected at a stage where the signal level is already fairly high, so that its influence on the signal-to-noise ratio is small.

The triode sections of the two ECL 86 tubes in the main amplifier are used as an amplifier and a phase splitter. Positive feedback is introduced by R_{28} , which is common to both triodes; as a result a signal with zero phase shift (360°) is fed to the grid circuit of the first triode. With such a circuit a gain in excess of 700 can be obtained, albeit over a narrow frequency band, and with a tendency to instability. With the values of resistors chosen, the gain is 175, the high-frequency response is still unaffected, and stability is ensured. The output impedances of the phase splitter are higher at the anode side than at the cathode side. This asymmetry might give rise to a considerable difference in frequency response of the two driver signals and this phenomena is avoided by using resistors of a low value for R_{30} and R_{33} (15 k Ω). For identical amplitudes of the driver signals, the sum of the values of resistors R_{28} , R_{30} and R_{32} should be equal to that of R_{33} . Because R_{28} and R_{32} have low values compared with R_{30} , the latter resistor could be kept equal to R_{33} without the balance of the signal being upset.

The pentode sections of the triode-pentodes ECL86 are used in a distributed load circuit (also called an ultra-linear or screen-grid feedback circuit). The advantages of this circuit are described on page 7, so that there is no need to enlarge upon the subject here. Especially in this amplifier, in which positive and negative feedback are used, the requirements set to the output transformer are high. The necessary stability and frequency response could not be obtained with the currently available types. There-

fore a new transformer, with C core, was designed. The type number of this transformer is AD9058. If a random transformer is used, it is not immaterial which of the triodes is used as the phase splitter and which as the amplifier. The signal at the anode of the pentode should be in phase with the input signal of the same tube. Hence, the phase splitter is combined with the upper pentode in the circuit diagram and the amplifier triode with the lower pentode. If the triodes are connected the other way round and a random transformer is used, considerable attenuation at the upper end of the response curve might occur. With transformer AD9058 this attenuation is negligible (-1.5 dB at 250 kc/s).

The negative feedback voltage is taken from the secondary of the output transformer via R_{40} , R_{28} ; 20 dB feedback is applied. C_{21} is shunted across R_{40} for phase correction at high frequencies, stability thus being improved. The quoted values of R_{40} and C_{21} apply to the 7 Ω loudspeaker tapping. With a 14 Ω loudspeaker combination, the complete secondary should be used, and the feedback voltage should be taken from the loudspeaker connections. In that case R_{40} should be increased from 2.2 k Ω to 3.3 k Ω , and C_{21} decreased from 680 pF to 390 pF.

The frequency response of this circuit proved to be very good. Without C_{16} , the attenuation at 250 kc/s was only 5 dB. However, when the output stage was driven into the grid current region, oscillations occurred at a frequency of about 600 kc/s. This could be prevented by applying some negative feedback from the anode of the phase splitter to the cathode of the amplifier tube via C_{16} . This measure increases the attenuation at frequencies in excess of 100 kc/s.

The standard supply transformer type AD9026 is used. The total H.T. current drain of the amplifier is 79 mA, the heater current is 2.9 A, with the rectifier EZ81 included. The AD9026, however, can supply the following outputs 2 x 280 V, 90 mA; 6.3 V, 3.5 A; 6.3 V, 1.1 mA and 5 V, 1.1 A. Therefore this transformer is slightly overrated.

The heaters of the ECL86 tubes and of the rectifier are connected to the 3.5 A winding, pin 4 of the heaters being connected to chassis. The pre-amplifier tubes EF86 and ECC82 are fed from the 1.1 A winding, which is connected to chassis via the slider of a potentiometer R_{44} , giving the possibility of reducing the hum level to a minimum.

There is no objection to ganging the mains switch with the volume control; the signal level at the latter is sufficiently high so that no noticeable hum is introduced.

MEASURED RESULTS

Voltages and currents

Voltage at C_{25}	317 V			
C_{24}	307 V		Total current drain	79 mA
C_{23}	275 V			
C_{22}	270 V			
Pentode section				
	ECL86 (I)		ECL86 (II)	
Anode	310 V	30 mA	310 V	30 mA
Screen grid	314 V	4.9 mA	314 V	4.9 mA
Cathode	9.5 V		9.5 V	
Triode section				
Anode	280 V	1.8 mA	150 V	0.71 mA
Cathode	30 V		1.4 V	
ECC82				
Second section				
Anode		95 V		1.8 mA
Cathode		4 V		
First section				
Anode		212 V		3.3 mA
Cathode		4 V		
EF86				
Anode		107 V		0.74 mA
Screen grid		93 V		0.15 mA
Cathode		2 V		

Frequency response

The frequency response was measured at the secondary of the output transformer, (a) with the tone controls in the centre position; (b) with the bass control at maximum and the treble control at minimum; (c) with the bass control at minimum and the treble control at maximum (see

Fig. 33). These curves were measured with the volume control in the maximum position.

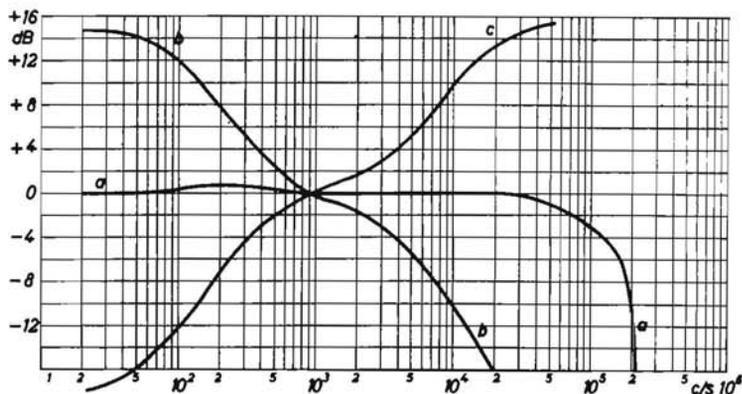


Fig. 33. Frequency response of the Hi-Fi amplifier with 2 x ECL86 in push-pull class AB; (a) with the tone controls in the "flat" position, (b) with maximum bass boost and maximum treble cut, (c) with maximum bass cut and maximum treble boost.

The slight deviation of curve (a) at about 200 c/s remains within one dB. It can be remedied by decreasing the value of the bypass capacitor of R_{24} ; however, this will also decrease the maximum bass boost by a few decibels. When C_{13} is given a value of 0.1 μF , the response curve will remain flat between 50 and 1000 c/s.

Power response

The power response at 1% distortion is plotted in Fig. 34. It is seen that an output of 10 W is obtained between 60 c/s and 10 kc/s. At an output power of 5 W the range from 45 c/s to 15 kc/s is covered.

Harmonic distortion and intermodulation distortion

The harmonic distortion was measured at frequencies of 1 kc/s, 60 c/s and 10 kc/s. The curves thus obtained are shown in Fig. 35. It is seen that at 10 W output power the distortion remains below 0.75%, but this

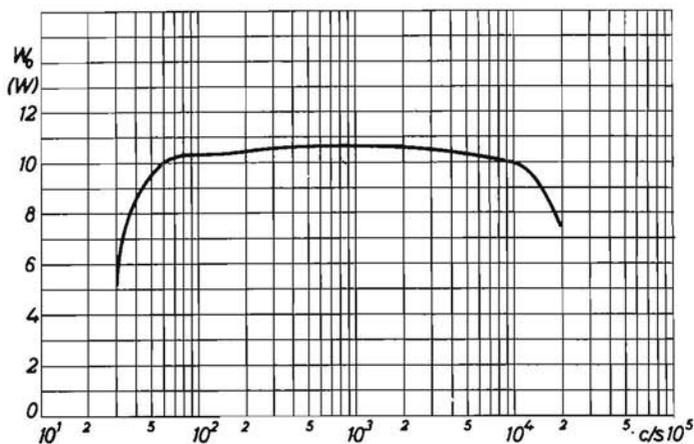


Fig. 34. Power response of the Hi-Fi amplifier with 2 x ECL86, measured at 1% harmonic distortion.

point is very close to the sudden increase in distortion which is characteristic to all heavily feedback amplifiers. An output power of 9 W is a safe value.

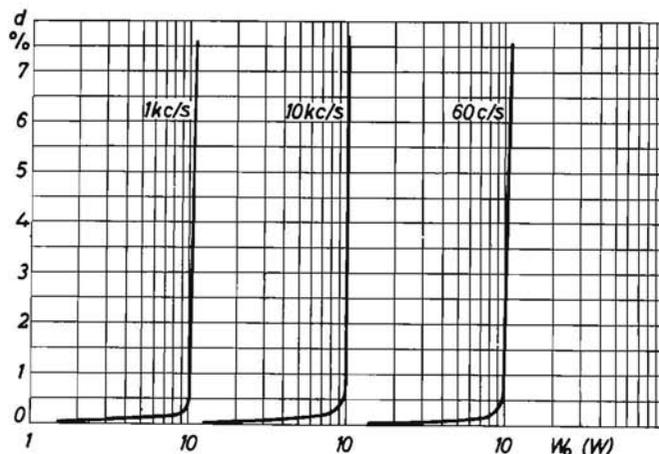


Fig. 35. Harmonic distortion of the Hi-Fi amplifier measured at frequencies of 1 kc/s, 60 c/s and 10 kc/s.

The curve of the intermodulation distortion is plotted in Fig. 36. It is seen that at 9 W output power the I.M. distortion remains below 0.5%, but at 10 W it increases to 2%.

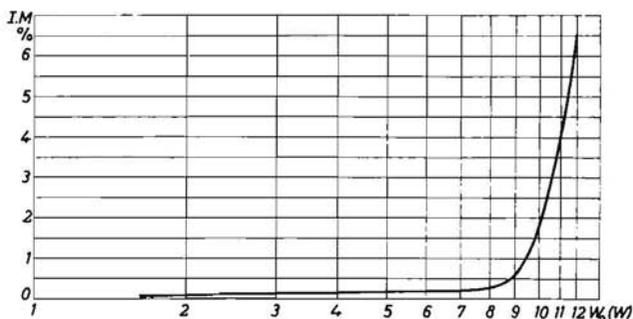


Fig. 36. Intermodulation distortion measured at frequencies of 40 c/s and 10 kc/s with an amplitude ratio of 4 : 1. The equivalent output power is indicated.

Since the power response, the harmonic distortion and the intermodulation distortion comply with stringent requirements at an output power of 9 W, it is on the safe side to call this amplifier a 9 W amplifier.

Phase shift

Fig. 37 shows the phase shift between the input and the output signal. It is seen that the phase shift remains below 25° between 25 c/s and 25 kc/s.

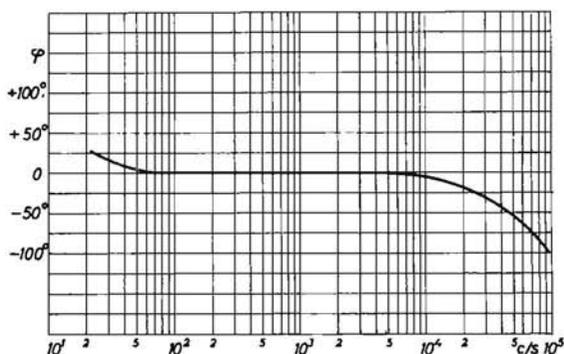


Fig. 37. Phase shift between the input and the output of the Hi-Fi amplifier as a function of the frequency.

R.I.A.A. correction

The correction characteristic for magneto dynamic pick-up heads, shown in Fig. 38, reveals that the correction is practically according to R.I.A.A. standards.

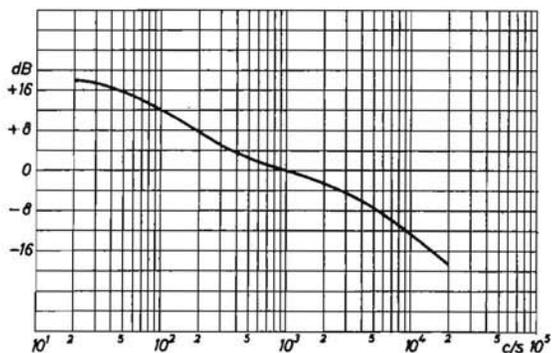


Fig. 38. Curve showing the correction characteristic of the pre-amplifier stage with an EF86, for use with magneto-dynamic pick-up heads.

20 W HI-FI AMPLIFIER WITH TWO PAIRS OF EL 84 PENTODES IN PUSH-PULL

Should a larger output than 20 W be required, a design can be chosen using either two EL 34 pentodes in push-pull or two EL 84 pentodes for each half of the push-pull circuit, i.e. four output tubes in all. The latter arrangement certainly entails rather more complicated wiring and a slightly larger chassis, but there are advantages which make this circuit well worth considering:

- 1) The output stage works with a lower anode voltage than would otherwise be necessary, involving a less expensive power transformer and electrolytic capacitors with lower working voltage.

- 2) The control voltage is the same as for the 10 W amplifier, i.e. relatively low, which is an advantage from the point of view of distortion.

- 3) The optimum load resistance is appreciably less than is necessary for the EL 34 pentode. This makes it a simpler matter to maintain a low leakage inductance in the output transformer and allows considerable negative feedback without any risk of instability.

This circuit should be of great interest to the experimenter, as it offers him much scope in selecting both the matching and the operating conditions of the output tubes.

CIRCUIT DESCRIPTION

The circuit diagram is shown in Fig. 39. In all there are 6 tubes, namely the ECC 83 double triode, used as amplifier and phase inverter with positive feedback, four EL 84 output pentodes and the GZ 34 rectifier.

As the pairs of output tubes are connected in parallel, the internal resistance and matching impedance are halved and the slope and output power doubled in comparison with the 10 W output stage using only two tubes.

The output tubes are operated with fixed grid bias. If individual biasing resistors were used any differences in the characteristics of the tubes within the usual manufacturing tolerances would certainly be overcome by the d.c. feedback, but the fixed bias, which can be controlled by means of R_{12} , with R_{13} to control the d.c. balance, provides an interesting arrangement from the aspect of the working point of the output stage.

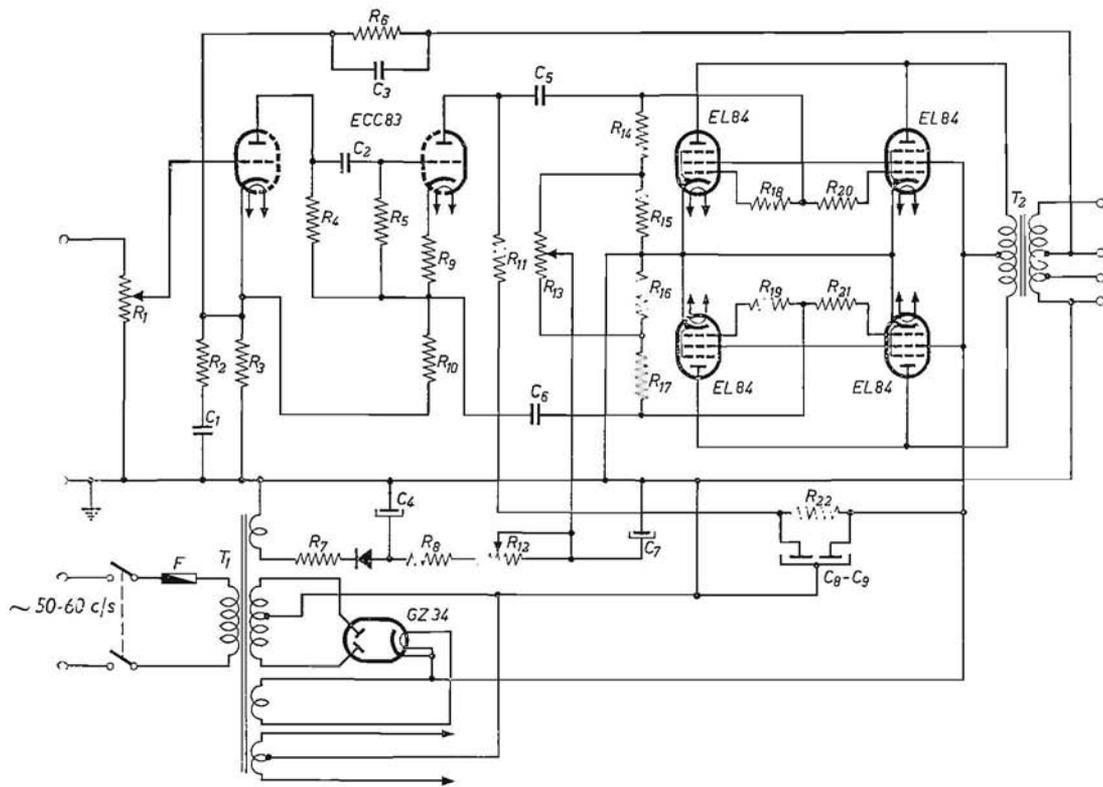


Fig. 39. Circuit diagram of 20 W push-pull amplifier using 2 x 2 EL 84 output pentodes.

The output stage can thus be adjusted to class AB with either normal or low loading or, if required, to class B by setting the grid bias; the numerous tapings on the output transformer secondary also provide facilities for varying the loading of the output stage over a wide range. The actual output transformer used for the prototype was type 4-U-63 marketed by Unitran, which happened to be available at the time, and details of which are as follows:

Maximum load	35 W
Primary matching impedance	3.6 k Ω
Impedance between screen grid tapings	670 Ω
Secondary matching impedance	3—5—7—15—400 Ω
Response curve flat within 1 dB	from 15 c/s to 50 kc/s
Maximum primary d.c.	2 \times 175 mA
Max. permissible unbalance	7 mA
Primary d.c. resistance	2 \times 95 Ω
Primary inductance (anode-to-anode)	90 H
Leakage inductance between primaries	3.5 mH
Leakage inductance between primary and secondary	10 mH
Leakage resonant frequency	100 kc/s

List of parts

R ₁ = 0.5 M Ω , linear	R ₁₇ = 330 k Ω \pm 10%, $\frac{1}{4}$ W
R ₂ = 3.9 k Ω \pm 5%, $\frac{1}{4}$ W ¹⁾	R ₁₈ = 1 k Ω \pm 20%, $\frac{1}{4}$ W
R ₃ = 1.2 k Ω \pm 5%, $\frac{1}{2}$ W ¹⁾	R ₁₉ = 1 k Ω \pm 20%, $\frac{1}{4}$ W
R ₄ = 150 k Ω \pm 10%, $\frac{1}{2}$ W	R ₂₀ = 1 k Ω \pm 20%, $\frac{1}{4}$ W
R ₅ = 1 M Ω \pm 20%, $\frac{1}{4}$ W	R ₂₁ = 1 k Ω \pm 20%, $\frac{1}{4}$ W
R ₆ = 27 k Ω \pm 5%, $\frac{1}{4}$ W ¹⁾	R ₂₂ = 27 k Ω \pm 20%, $\frac{1}{2}$ W
R ₇ = 2.7 k Ω \pm 20%, $\frac{1}{4}$ W	C ₁ = 2700 pF \pm 5%
R ₈ = 27 k Ω \pm 20%, $\frac{1}{4}$ W	C ₂ = 0.1 μ F \pm 20%, 350 V wkg.
R ₉ = 3.3 k Ω \pm 5%, $\frac{1}{2}$ W	C ₃ = 56 pF \pm 5%
R ₁₀ = 150 k Ω \pm 5%, $\frac{1}{2}$ W ¹⁾	C ₄ = 25 μ F, 50 V wkg.
R ₁₁ = 160 k Ω \pm 5%, $\frac{1}{2}$ W ¹⁾	C ₅ = 0.1 μ F \pm 10%, 350 V wkg.
R ₁₂ = 25 k Ω , linear	C ₆ = 0.1 μ F \pm 10%, 350 V wkg.
R ₁₃ = 50 k Ω , linear	C ₇ = 50 μ F, 25 V wkg.
R ₁₄ = 330 k Ω \pm 10%, $\frac{1}{4}$ W	C ₈ = 32 μ F, 350 V wkg.
R ₁₅ = 56 k Ω \pm 10%, $\frac{1}{4}$ W	C ₉ = 64 μ F, 350 V wkg.
R ₁₆ = 56 k Ω \pm 10%, $\frac{1}{4}$ W	

¹⁾ High stability

This transformer is suitable for ultra-linear operation with the screen-grid tapping at about 23% of the number of turns, which corresponds to the second set of data given in Table I (p. 9) for two EL 84 valves. No experimental results are available, however.

The transformer is wound on a high-permeability core, and can so be used for impedances differing by $\pm 30\%$ from the nominal values. It therefore lends itself well to experimental work. Some idea of the results it may be expected to yield, and the values of the associated components, may be obtained from the details of the amplifier previously described on p. 46.

Details of the phase inverter circuit and pre-amplification are already described in the previous chapter. The ECC 83 is operated at a gain of about 220, which means relatively high sensitivity with only a small number of tubes.

Negative feedback is obtained from the 15-ohm secondary tapping of the output transformer and the loop gain is reduced in the higher frequency range by connecting C_3 (56 pF) in parallel with the feedback resistor R_6 , and a high-pass filter R_2C_1 (3.9 k Ω and 2700 pF) in parallel with the biasing resistor of the first triode. No instability was found in the experimental model or in the many amplifiers subsequently made to this specification.

It may be useful at this point to examine the difference in the behaviour of amplifiers with fixed grid bias and automatic bias using a cathode resistor.

With fixed bias, when the volume control is turned up slowly, a point is reached where considerable distortion suddenly becomes evident. In amplifiers with automatic bias the transition from distortionless reproduction to clearly audible distortion is much more gradual, from which it would appear that, for amplifiers having the same electrical characteristics, those using automatic bias are capable of larger distortionfree outputs. The explanation is as follows.

Serious distortion begins to be noticeable only when grid current starts to flow in the output stage and, in an amplifier with fixed bias, a flow of grid current is accompanied by an increase in the grid bias, owing to the voltage drop across the grid resistors of the output tubes. Now, if the resultant displacement of the working point on full load is in the direction of class B operation, any increase in the grid bias due to the grid current may just tip the scales beyond the point of cut-off, at which much distortion occurs. The

effect is aggravated by the fact that the high grid bias causes the anode and screen-grid currents to drop, in consequence of which the voltage on the h.t. line rises. The voltages on the preceding tubes, and thereby also their amplification, are increased and the distortion becomes even worse. Also, when the output signal drops, the negative feedback on the preceding tubes is reduced and again the gain is increased. Thus grid current sets in motion a train of cumulative events, all of which result in distortion.

On the other hand, automatic grid bias, which is in effect a positive cathode potential, produces entirely different results. Grid current in this case also causes an increase in the bias, but now, in consequence of the corresponding drop in anode and screen grid currents, the positive cathode potential drops, so there can be no question of the tubes being biased so close to the cut-off point. And, as the current does not drop to such an extent, the supply voltage remains more constant and the effect on the preceding tubes is not so serious. The transition to noticeable distortion is accordingly more gradual.

There is yet another effect of fixed bias in that the amplifier exhibits a kind of "backlash", i.e. with a sinusoidal signal applied to the input and gradually increased in strength until its oscillogram reveals symptoms of cut-off, the signal strength has to be reduced to a considerably lower level before these symptoms disappear. In other words, sudden peaks in the reproduction of music are followed by distortion in the following passage as well. This does not happen when automatic bias is employed.

In any case, if the amplifier is designed along such generous lines that there is never any likelihood of grid current, it makes no difference how the bias is obtained.

Power supply

For normal loading the power transformer should be able to deliver 2 x 300 V_{rms} , 225 mA; 25 V_{rms} , 5 mA; 6.3 V_{rms} , 3.5 A and 5 V_{rms} , 2A, but it is advisable to have an extra 6.3 V available for a pre-amplifier if this is to be fed from the same transformer; the additional 3 to 5 mA drain on the H.T. for this purpose is well within the permissible load. If an FM/AM tuner unit is to be used with the amplifier it is better to use a separate power unit. It is not generally realised that these tuners, which comprise only r.f. and i.f. sections, operate quite well on 150 V anode potential, and the

necessary 30 to 50 mA, if taken from the amplifier transformer, renders this component unnecessarily expensive and imposes heavy requirements on the reservoir capacitor from the point of view of ripple current.

The power supply can be designed even more economically if the amplifier is built for low loading only. The transformer then only need deliver 160 mA and, as the mean anode current is less than 150 mA, the EZ 81 can be used as rectifier, being intended for an average output of 150 mA with 450 mA peak current; the short surges which accompany peaks in the music can easily be met.

Grid bias for the output tubes is obtained from a 25 V winding on the mains transformer and rectified by a type OA 85 germanium diode, which is capable of carrying relatively high inverse voltages. The potential is so adjusted by means of R_{12} that each output tube takes 36 mA on normal loading, or 24 mA on low loading. R_{13} is used for equalising the anode currents.

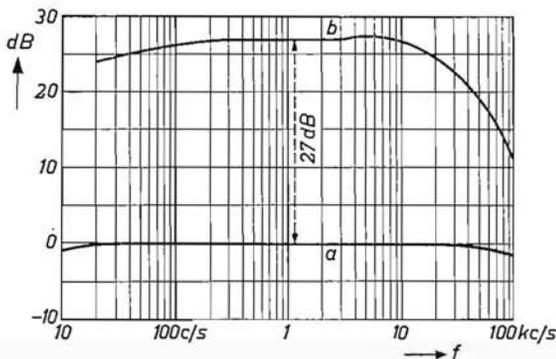


Fig. 40. Frequency response curves: (a) with negative feedback, (b) without feedback, as plotted for the 20 W amplifier using 2 x 2 type EL 84 pentodes. The feedback is 27 dB at 1 kc/s.

Frequency response and distortion

The response curve, with and without negative feedback, is shown in Fig. 40. Harmonic distortion was measured at 40 c/s and 1 kc/s, and the results are given in Fig. 41. Both diagrams relate to the normal loading conditions.

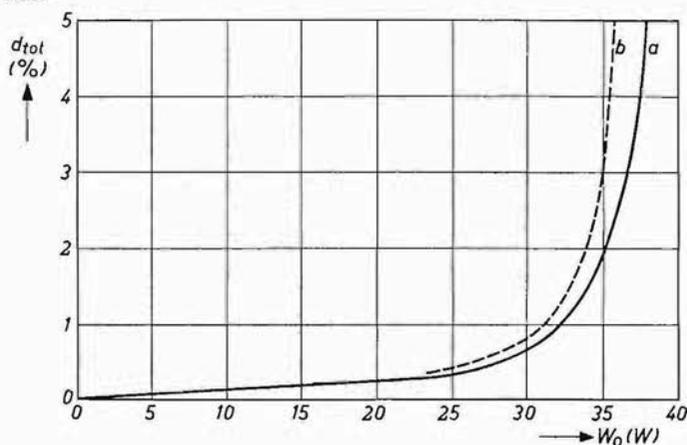


Fig. 41. Total harmonic distortion d_{tot} as a function of the output power at 1 kc/s (a) and 40 c/s (b) for the 20 W Hi-Fi amplifier with 2 x 2 type EL 84 pentodes in push-pull.

20 W HI-FI AMPLIFIER WITH 2 TYPE EL 34 PENTODES IN ULTRA-LINEAR PUSH-PULL

The original design of this amplifier was based on the measurements taken from two EL 34 pentodes in ultra-linear push-pull mentioned in the first chapter; the results agree very well indeed with the best figures obtained at the time. At the nominal output power the distortion is exceptionally low; i.e. less than five hundredth of one per cent ($< 0.05\%$) at 400 c/s for 20 W output. The measured intermodulation distortion is only 0.7% for an equivalent output of 20 W measured with frequencies of 40 c/s and 10 kc/s, with an amplitude ratio of 4 : 1. With 29 W equivalent output power this distortion is not more than 1%. Compared with the 1 kc/s level for 20 W output, the output curve is flat to within 0.5 dB from 30 c/s to 20 kc/s, and the response measured at 1 W is flat to within 1 dB from 2 c/s to 100 kc/s. The output impedance is only 0.3Ω , measured at 40 c/s, 1 kc/s and 20 kc/s, which means that the loudspeakers are very effectively damped.

The sensitivity is 220 mV for 20 W output, this necessitating the use of a pre-amplifier, preferably with equalising network, volume and tone controls. Details of a suitable pre-amplifier are given in the next chapter.

CIRCUIT DESCRIPTION

The tube complement comprises EF 86, ECC 83, 2 x EL 34 and GZ 34 rectifier (see Fig. 42).

Ultra-linear push-pull output is employed, with screen-grid tappings at 40% from the centre-tap of the primary of the output transformer. The optimum load for the output tubes is $6.6 \text{ k}\Omega$, with 440 V at the centre-tap of the transformer, and a common anode and screen-grid dissipation of 28 W for each tube. With this particular anode-to-screen-grid load ratio it has been found that improved linearity is obtained at power levels above 15 W when resistors of the order of $1 \text{ k}\Omega$ are inserted in the screen-grid connections; the consequent slight reduction in the peak output power is negligible.

Separate biasing resistors are used for the output stage in order to minimise mismatch of the direct currents flowing in the primary of the output transformer.

Further d.c. balancing is not considered necessary, but it is just possible that it may give improved response at the lower frequencies.

In this circuit the biasing resistors must be decoupled (if a common resistor is used this is also decoupled). A low time-constant for the cathode circuit is essential in the case of automatic grid bias.

As in the 10 W amplifier, the combined driver and phase inverter is the ECC 83. The high supply voltage available ensures the required drive with only 0.4% distortion. Resistors R_{12} and R_{13} (180 k Ω) should be matched within 5%, the higher of the two being used for R_{13} .

Optimum balance is obtained with 3% difference between the effective anode loads of the ECC 83 triodes. The grid leaks R_{15} and R_{16} should therefore not differ too much, seeing that they constitute a part of the anode load in the output stage. Matching of the control signal at the higher frequencies is governed entirely by the wiring and layout, so the parallel capacitances in the anode circuit should be as far as possible equal.

Matching at the lower frequencies is dependent on the time constant of R_9C_6 in the grid circuit of the ECC 83, and the selected value of 0.22 ensures adequate balance down to the lowest frequencies.

It is a disadvantage of cathode-coupled phase inverter circuits that the amplification is so low, being only one half of what it would be with the tube operated as a normal voltage amplifier. However, the amplification in the circuit is in any case 25, which ensures adequate sensitivity with the EF 86 as the input tube.

In this circuit the input tube provides an amplification of about 120. The use of high stability resistors in this stage is recommended, as these go a long way towards ensuring little noise and intermodulation distortion. This stage is coupled direct to the phase inverter to give the grids of the tube the required positive potential and avoid phase shift at the lower frequencies; in this way stability is maintained even with considerable negative feedback.

Negative feedback

Without negative feedback the sensitivity of the amplifier is 6.5 mV for 20 W; with feedback it is about 220 mV for the same output, the average loop gain being roughly 30 dB. The loop gain, response curve and phase shift are shown in Fig. 43

In spite of the large amount of negative feedback the prototype made to

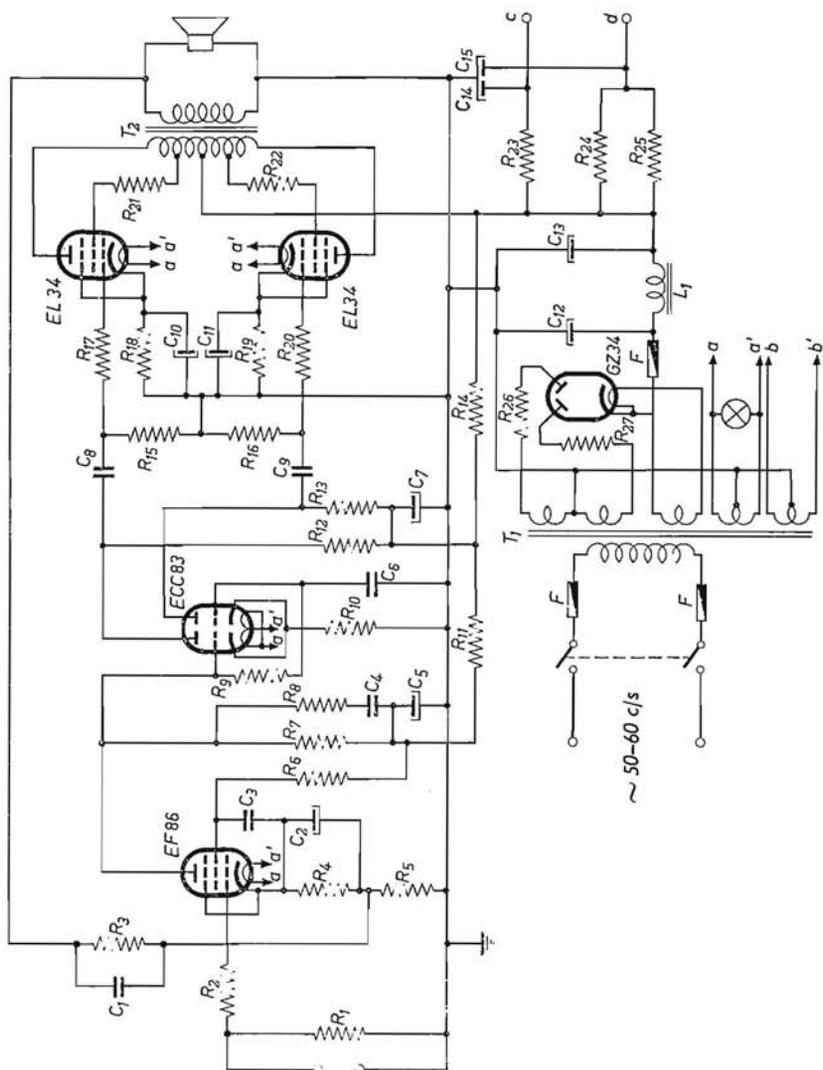


Fig. 42. Circuit diagram of 20 W amplifier using 2 x EL 34 pentodes.

this design proved perfectly stable; the negative feedback had to be increased to 40 dB by reducing the value of R_3 before signs of instability at the higher frequencies became apparent. In a circuit of this type symptoms of instability first appear with capacitive loads but, even with long leads to the loudspeaker, no trace of oscillation was observed.

Distortion

The total harmonic distortion as measured at 400 c/s, with and without negative feedback, into a resistive load is shown in Fig. 44, together with the distortion up to the overload point with 30 dB feedback (curve b). With rated output (20 W) the distortion without negative feedback is considerably less than 1%, with feedback it is less than 0.05%. Harmonic distortion at 400 c/s is only 0.1% for 27 W output.

The loop gain characteristic shows that at least 20 dB negative feedback is maintained from 15 c/s to 25 c/s, with 26 dB down to 30 c/s.

The intermodulation distortion, measured at with frequencies of 40 c/s and

List of parts

$R_1 = 1 \text{ M}\Omega \pm 20\%, \frac{1}{4} \text{ W}$	$R_{22} = 1 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_2 = 4.7 \text{ k}\Omega \pm 20\%, \frac{1}{4} \text{ W}$	$R_{23} = 56 \text{ k}\Omega \pm 10\%, 1 \text{ W}$
$R_3 = \text{high stab.} \pm 5\%$ 12-16 Ω matching: 8.2 k Ω 6-8 Ω matching: 5.6 k Ω	$R_{24} = 12 \text{ k}\Omega \pm 20\%, 6 \text{ W}^3)$
$R_4 = 2.2 \text{ k}\Omega \pm 10\%^{1)}$	$R_{25} = 12 \text{ k}\Omega \pm 20\%, 6 \text{ W}^3)$
$R_5 = 100 \Omega \pm 5\%^{1)}$	$R_{26} = \text{dependent on } R_{tr}$
$R_6 = 390 \text{ k}\Omega \pm 10\%^{1)}$	$R_{27} = \text{dependent on } R_{tr}$
$R_7 = 100 \text{ k}\Omega \pm 10\%^{1)}$	$C_1 = \pm 5\%$ 12-16 Ω matching: 220 pF 6-8 Ω matching: 330 pF
$R_8 = 4.7 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_2 = 50 \mu\text{F}, 12 \text{ V working}$
$R_9 = 1 \text{ M}\Omega \pm 20\%, \frac{1}{4} \text{ W}$	$C_3 = 56 \text{ 000 pF}, 350 \text{ V working}$
$R_{10} = 82 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_4 = 47 \text{ pF} \pm 10\%$
$R_{11} = 270 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_5 = 8 \mu\text{F}, 450 \text{ V wkg.}$
$R_{12} = 180 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}^2)$	$C_6 = 0.22 \mu\text{F}, 350 \text{ V wkg.}$
$R_{13} = 180 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^2)$	$C_7 = 8 \mu\text{F}, 450 \text{ V wkg.}$
$R_{14} = 15 \text{ k}\Omega \pm 20\%, \frac{1}{2} \text{ W}$	$C_8 = 0.47 \mu\text{F}, 350 \text{ V wkg.}$
$R_{15} = 470 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_9 = 0.47 \mu\text{F}, 350 \text{ V wkg.}$
$R_{16} = 470 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{10} = 50 \mu\text{F}, 50 \text{ V wkg.}$
$R_{17} = 2.2 \text{ k}\Omega \pm 20\%, \frac{1}{4} \text{ W}$	$C_{11} = 50 \mu\text{F}, 50 \text{ V wkg.}$
$R_{18} = 470 \Omega \pm 5\%, 3 \text{ W}^3)$	$C_{12} = 50 \mu\text{F}, 450 \text{ V wkg.}$
$R_{19} = 470 \Omega \pm 5\%, 3 \text{ W}^3)$	$C_{13} = 50 \mu\text{F}, 450 \text{ V wkg.}$
$R_{20} = 2.2 \text{ k}\Omega \pm 20\%, \frac{1}{4} \text{ W}$	$C_{14} \left\{ \begin{array}{l} 2 \times 8 \mu\text{F}, 450 \text{ V wkg.} \\ C_{15} \end{array} \right.$
$R_{21} = 1 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	

1) High stability.

2) Matched within 5%; the larger as R_{13} .

3) Wire-wound.

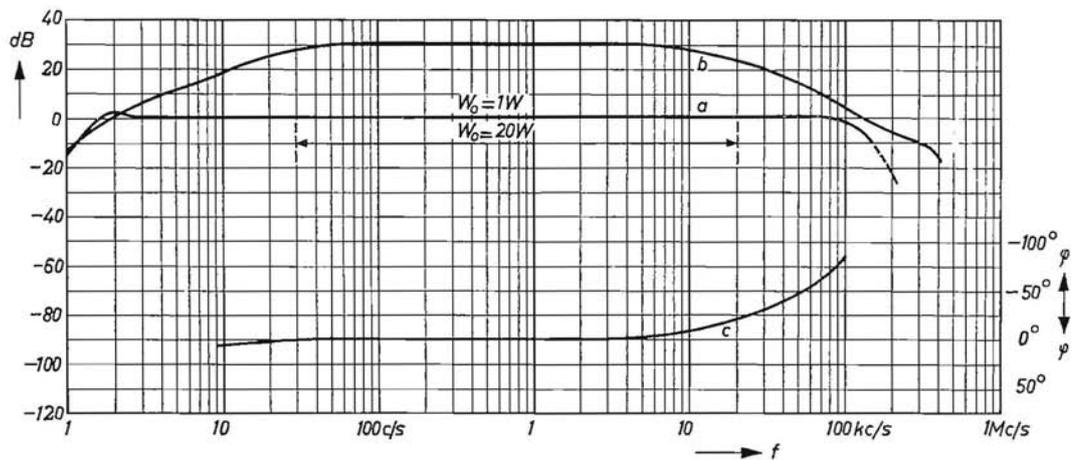


Fig. 43. Response curve (a), loop gain (b), and phase shift characteristic (c) for the 20 W ultra-linear amplifier using two EL 34 pentodes.

10 kc/s with an amplitude ratio of 4 : 1, was found to be 0.7% of the 10 kc/s amplitude for 20 W, and about 1% for 29 W at peak amplitude of the combined signal, corresponding to the maximum value of a sinusoidal signal.

Fig. 45 shows the input voltages as a function of the output voltage, measured across a 15 Ω resistor. It will be seen that excellent linearity is obtained up to 20 V, corresponding to 27 W output.

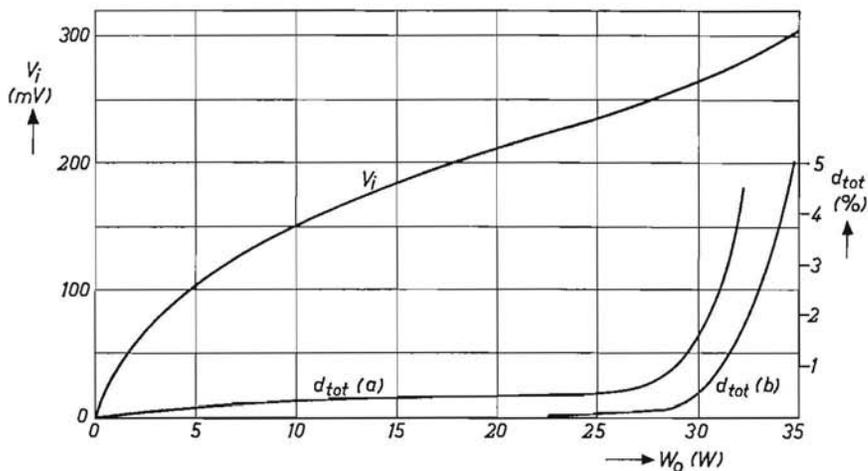


Fig. 44. Total harmonic distortion, (a) without negative feedback, (b) with feedback, and input voltage (V_i) plotted against the output power of the 20 W amplifier at 400 c/s.

Sensitivity

This amplifier has a sensitivity of about 220 mV for 20 W output, or 300 mV at the overload point, at 1000 c/s. Background noise in the case of the prototype was found to be 89 dB below 20 W, using an input resistance of 10 k Ω this is equivalent to about 5.5 μ V at the input terminals. It is possible to increase the sensitivity by a further 6 dB while still maintaining a low noise level, high loop gain and complete stability but the problem of designing a suitable pre-amplifier, particularly with a satisfactory signal-to-noise ratio, makes it doubtful whether there is any advantage in so doing.

Power response

The power handling capacity at the low frequency end of the audible range is governed mainly by the quality of the output transformer. It is desirable in Hi-Fi work for the pre-amplifier to attenuate sufficiently the very low frequencies as the power amplifier is in any case unable to handle

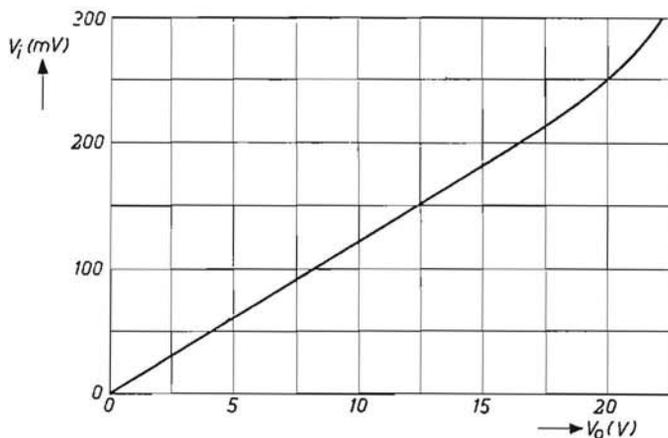


Fig. 45. Input voltage against output voltage on 15 ohm load, measured at the output transformer secondary.

these at full power without running into distortion. Using the output transformer described below, 20 W is easily available down to 20 c/s, and the frequency response at the 29 W level is flat from 30 c/s to 20 kc/s. This transformer is expensive and not easy to produce, but a less costly one is available, type AD 9021, having a lower primary inductance, but nevertheless yielding the same results at the higher frequencies. With this transformer the response curve drops rather sharply at below 10 c/s while the output curve is flat down to 40 c/s. The output power is of course slightly reduced.

The output transformer

The output transformer is suitable for 6—8 Ω and 12—16 Ω matching. At 5 V, 50 c/s the primary self-inductance is 72 H and at 10 V, 50 c/s 120 H. With the secondary short-circuited, the primary leakage inductance is 8 mH

or, with one-half of the primary short-circuited, 6 mH. The total primary resistance is 310Ω , and that of the secondary 0.45Ω for the 12 — 16 Ω matching, or 0.18Ω for the 6 — 8 Ω matching. The maximum flux density (B_{max}) is 5800 gauss at 20 c/s, 500 V peak.

The transformer is wound on a former with an accurately centred partition, each section carrying one half of the primary. Each of these is subdivided into five sections, connected in series with the secondary windings in between. Hence there are ten primary coils and eight secondary coils, the latter connected partly in series and partly in parallel. Other details are as follows:

Core:	Shell type, standard dynamo sheet
Overall dimensions:	150 x 125 mm. (see Fig. 46)
Width of core:	50 mm.
Stacking:	Without air-gap
Height of stacking:	50 mm.
Cross section:	25 cm ²

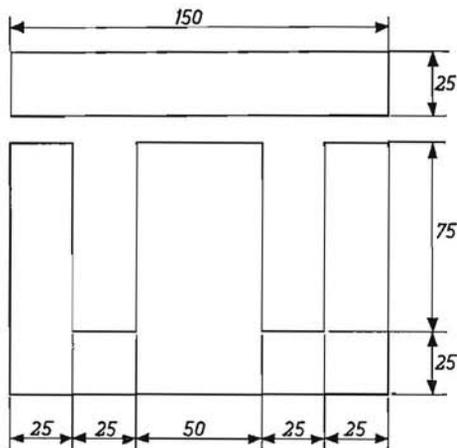


Fig. 46. Dimensions in mm of the output transformer laminations.

The dimensions of the laminations are given in Fig. 46. The coil former has double flanges at the ends and a single flange in the centre. Winding is commenced at the outer flanges and proceeds in the manner indicated in the following table.

Winding	Number of turns	Wire thickness	Winding length	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 model the windings were insulated from each other with one covering of fuller board 0.1 mm. thick and two layers of 60 μ paper. The start and finish of each winding are against the end flanges, and the following connections are made internally:

end of P ₁ to start of P ₂	end of P ₁₀ to start of P ₉
end of P ₂ to start of P ₃	end of P ₉ to start of P ₈
end of P ₃ to start of P ₄	end of P ₈ to start of P ₇
end of P ₄ to start of P ₅	end of P ₇ to start of P ₆
end of P ₅ to end of P ₆	

The starts and ends of the secondary windings S₁, S₂, S₃ and also of S₆, S₇, S₈ are joined to make two groups of three windings in parallel (see Fig. 47), and the following exterior connections are made. Start of P₁ to anode of EL 34 (I) and start of P₁₀ to anode of EL 34 (II). Junction of P₃ — P₄ to the screen grid of the EL 34 (I) and that of P₇ — P₈ to the screen grid of EL 34 (II). Junction P₅ — P₆ to H.T. positive. The secondary connections (Fig. 35) are as follows: For 6 — 8 ohm matching the starts of S₁S₂S₃ are connected to the start of S₄, the ends being similarly connected to the end of S₄, i.e. W to Y and X to Z. The same is done with W' — Y' and X' — Z', and the parallel groups thus obtained are connected in series; the loud-speaker is connected to W and W'.

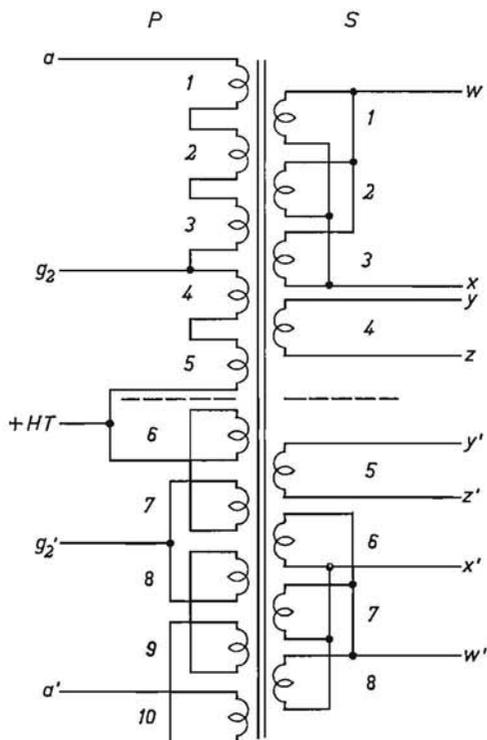


Fig. 47. Primary and secondary connections of the output transformer.

For 12 — 16 ohm matching S_4 and S_5 are connected in parallel (Y to Y' and Z to Z'); the three groups are then connected in series, namely Y and Y' to X, and Z and Z' to X', with the loudspeakers to W and W'.

The power supply

The mains transformer should be capable of delivering 2 x 410 V at 180 mA if the amplifier is to be used with an FM/AM tuner unit as well. If not, 2 x 150 mA is sufficient. 5 V, 3 A is required for the heater of the GZ 34 rectifier; 6.3 V, 4 A for the amplifying tube heaters and 6.3 V, 3 A with centre-

tapping for the pre-amplifier and radio tuner. If the latter has its own power supply, 1 A is sufficient for the pre-amplifier alone.

The values of R_{26} and R_{27} will depend on the resistance of the transformer, which for the GZ 34 should be at least 110Ω .

10 W SINGLE-ENDED PUSH-PULL AMPLIFIER WITH TWO EL 86 PENTODES

Single-ended push-pull amplifier circuits were originally designed for use in broadcast receivers in conjunction with a high impedance loudspeaker. This arrangement gives a small saving in cost, greater output power and higher quality of reproduction than the usual single output tube under Class A conditions.

The amount of distortion does not represent the ultimate minimum, but in comparison with the conventional receiver the improvement in quality is at once apparent to the listener.

With a little more expenditure on additional components and some modification of the circuit, however, the distortion can be reduced to extremely low limits, rendering the amplifier eminently suitable for Hi-Fi reproduction. Tone controls and equalising networks can be housed in a separate pre-amplifier, which may take its power supplies from the main amplifier.

In the tests made on the prototype of this amplifier the unusual difficulty was encountered that the distortion could not be measured with the aid of the conventional A.F. signal generators and distortion meters, because the distortion in the generator is greater than that inherent in the amplifier itself, particularly at low output powers. In this case, therefore, tuned A. F. filters were used between the signal generator and the amplifier, the distortion being measured with a wave analyser.

Summary of performance

Nominal output power:	10	W
Maximum output power (1% distortion):	11	W
Harmonic distortion: for 11 W	< 1	%
for 10 W	< 0.3	%
for 2 W	< 0.02	%
Intermodulation distortion (10 W equivalent output):	< 1.2	%
Frequency response: flat from 7 c/s to 40 kc/s; — 3.8 dB at 300 kc/s.		
Power response: (1% distortion) flat from 30 c/s to 20 kc/s. (It was not possible to take measurements at higher frequencies owing to the limited frequency range of the distortion meter).		

Sensitivity:	40 mV for 50 mW 0.6 V for full output (11 W)
Output impedance:	800 Ω
Internal resistance	50 Ω

CIRCUIT DESCRIPTION

The circuit diagram of this amplifier is shown in Fig. 48. The output stage is equipped with two type EL 86 pentodes, and the double triode ECC 83 is used for pre-amplification and phase inversion in the manner described in Chapter I.

One of the problems arising in single-ended push-pull amplifiers is the method of supplying the screen grids of the output tubes, and more particularly that of the "upper" tube (I). In the amplifier shown here this difficulty has been overcome by using a double choke. Each of the screen grids is fed from a separate winding on the choke, in which the current flows in opposite directions through the two sections, thus eliminating magnetisation of the core and permitting of small dimensions and high inductance. Choke Type AD 9025 has been specially designed for use in this amplifier, the inductance being 60 H per section.¹⁾

¹⁾ If the double choke type AD9025 is not available the following coil data may be useful for construction:

Core: normal dynamo-sheet.

Laminations: see Fig.

Stacking height: 16 mm.

Cross section: 1.6 cm².

Stacking: without air gap.

Windings: $S_1 = S_2 = 3200$ turn, enamelled copper wire of 0.11 mm diameter

Width of windings: 21 mm.

Number of layers: each coil 22 layers.

Number of turns per layer: 146.

Resistance: $S_1 = 480 \Omega$; $S_2 = 655 \Omega$.

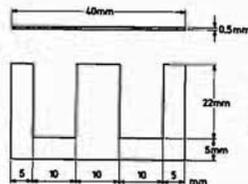
Insulation between windings: 1 layer 0.06 mm plastic folium

1 layer 0.03 mm paper

2 layers 0.06 mm plastic folium.

Inductance: $L = 60$ H; $-12 + 32\%$.

Measured at S_1 with 50 V, 50 c/s, without D.C. magnetisation.



Connect (1) to screen grid EL 86 (I) and (4) to screen grid EL 86 (II).



The use of the choke makes that one side of the speech coil can be earthed, and the H.T. need not be centre-tapped. The loss of power is only small; the output curve drops sharply below 30 c/s, but this makes little difference since the resonant frequency of most loudspeakers is above this value.

The EL 86 (II) output tube is connected in the usual way, with cathode resistor for biasing purposes and earthed grid leak. The $1\text{ M}\Omega$ grid leak of EL 86 (I) is also earthed, producing strong negative feedback and hence low distortion. However, in this circuit the grid of the EL 86 (I) must be at a positive potential which is below that of the anode of the EL 86 (II). To achieve this capacitor C_4 between the phase inverter and the grid of EL 86 (I) is shunted by a $620\text{ k}\Omega$ resistor.

Amplifiers of this type are designed more or less experimentally. Of course care is taken that the limiting values of the tubes are not exceeded, but after the design has been committed to paper and the values of the components have been worked out, the values of the resistors are changed where necessary and measurements are taken in order to achieve the minimum amount of distortion. Tubes with characteristics at the limits of their manufacturing tolerances are tested in the actual circuit and all measurements are finally repeated in order to be quite certain that none of the results is the outcome of fortuitous circumstances. In many cases readjustments may have to be made to arrive at the optimum average conditions to meet the inevitable variations in the characteristics of tubes and components.

After all this it was found that the least distortion occurred in this amplifier with the anode potential of the EL 86 (II) slightly higher than that of the other output tube. On a 325 V supply the anode potential of EL 86 (I) was then 154 V and that of EL 86 (II) 160 V, with 11 V across the biasing resistor R_{15} .

The output stage was found to be taking a total of 73 mA. From the above it follows that the grid should have a potential above earth of 160 V, which means a current of 0.16 mA through the grid leak R_{10} , the same current passing through R_9 ; the voltage on the anode of the phase inverter will thus be 260 V.

Now, the resistance of the screen-grid choke is about $500\ \Omega$ and the screen-grid current is 6 mA, so the supply voltage for the phase inverter is 322 V. In order to obtain 260 V the anode resistor ($R_7 = 100\text{ k}\Omega$) must pass a current of 0.62 mA, of which 0.16 mA flows through the grid potentiometer

and 0.46 mA through the tube. The potential difference across the tube is about 240 V, and the amount of current in question will be obtained with a grid bias of 2.3 V. This gives the biasing resistor a value of 5.1 k Ω . The value of R_6 is determined by the potential which will have to be available at the anode to ensure the correct grid voltage for EL 86 (I) and, at the same time, to provide the correct amount of gain to obtain the required drive for this tube.

The anode current taken by the first triode of the ECC 83 is about 0.6 mA and the grid bias is 1.75 V. The biasing resistor R_2 also carries the direct

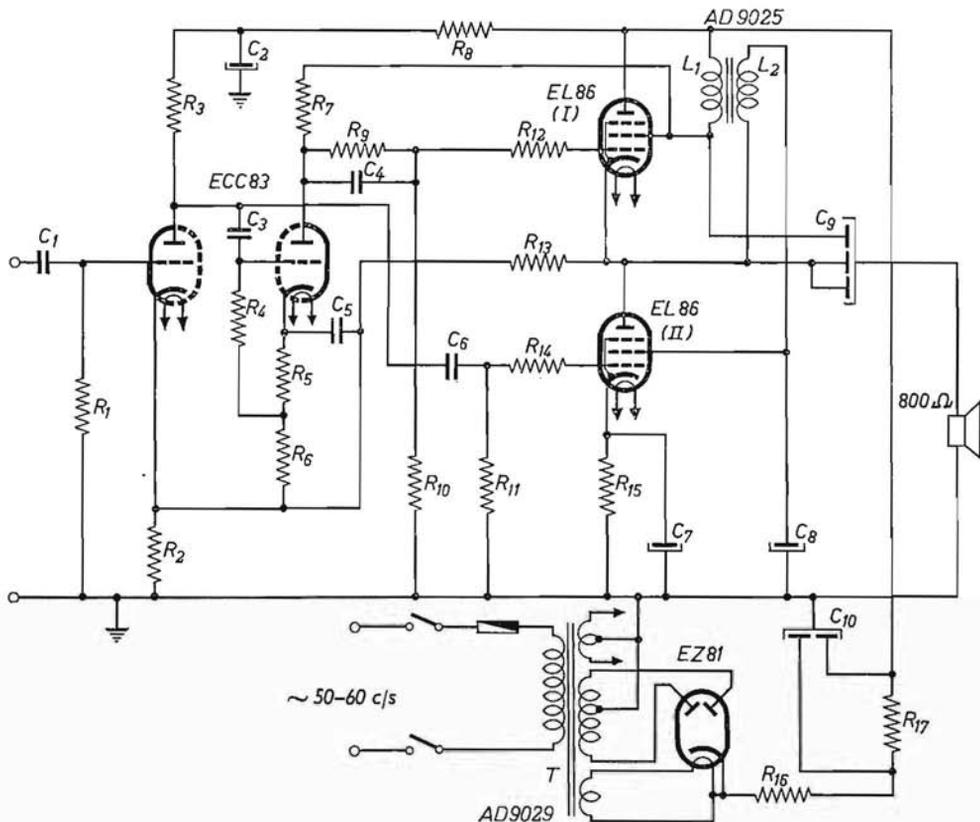


Fig. 48. Circuit diagram of the single-ended push-pull amplifier with two EL86 power pentodes.

current flowing through R_{13} together with the cathode current of the phase inverter. Originally a higher value was chosen for this resistor, but positive feedback brought the tube so close to the point of oscillation that the signs of instability mentioned in Chapter I were evident. The present value just maintains stability, even without negative feedback.

This subject has been dealt with at some length in order to stress the point that, in circuits such as this, if the value of only one single component is changed, the working conditions of almost all the tubes in the amplifier are modified. It is hardly necessary to add that many of the resistors will have to be in the 5% tolerance range.

The same thing applies to the signal ratios. Mention has already been made in Chapter I of the high control voltage required for EL 86 (I). The control signal to be delivered by the phase inverter is only 2.5 times greater than that required for the input tube, but the choice of anode load resistor is in itself not sufficient to provide such a wide difference in amplification. A correct signal ratio is ensured, however, by the fact that the undecoupled biasing resistor for the phase inverter is very much larger than that of the input tube, so that the negative feedback to the phase inverter is also much greater.

A glance at the circuit diagram will show that, although the coupling capacitor C_4 to the grid of EL 86 (I) is $0.1\mu\text{F}$, that of EL 86 (II) is only 47,000 pF. This is done in order to maintain a balance at the lower frequencies; the last mentioned tube being operated direct from the input stage whilst

List of Parts

$R_1 = 1\text{ M}\Omega \pm 20\%, \frac{1}{4}\text{ W}$	$R_{16} = \text{see text}$
$R_2 = 680\ \Omega \pm 5\%, \frac{1}{4}\text{ W}^1)$	$R_{17} = 150 \pm 20\%, 6\text{ W wire-wound}$
$R_3 = 220\text{ k}\Omega \pm 10\%, \frac{1}{4}\text{ W}^1)$	
$R_4 = 1\text{ M}\Omega \pm 20\%, \frac{1}{4}\text{ W}$	$C_1 = 0.1\ \mu\text{F}$
$R_5 = 5.1\text{ k}\Omega \pm 5\%, \frac{1}{4}\text{ W}^1)$	$C_2 = 8\ \mu\text{F}, 350\text{ V wkg}$
$R_6 = 33\text{ k}\Omega \pm 5\%, \frac{1}{4}\text{ W}^1)$	$C_3 = 0.1\ \mu\text{F}, 350\text{ V wkg}$
$R_7 = 100\text{ k}\Omega \pm 5\%, \frac{1}{4}\text{ W}^1)$	$C_4 = 0.1\ \mu\text{F}, 350\text{ V wkg}$
$R_8 = 100\text{ k}\Omega \pm 10\%, \frac{1}{4}\text{ W}$	$C_5 = 47\text{ pF}$
$R_9 = 620\text{ k}\Omega \pm 5\%, \frac{1}{4}\text{ W}^1)$	$C_4 = 47\ 000\text{ pF}, 350\text{ V wkg}$
$R_{10} = 1\text{ M}\Omega \pm 5\%, \frac{1}{4}\text{ W}^1)$	$C_7 = 100\ \mu\text{F}, 25\text{ V wkg}$
$R_{11} = 1\text{ M}\Omega \pm 10\%, \frac{1}{4}\text{ W}$	$C_8 = 8\ \mu\text{F}, 350\text{ V wkg}$
$R_{12} = 1\text{ k}\Omega \pm 20\%, \frac{1}{4}\text{ W}$	$C_9 = 3 \times 50\ \mu\text{F}, 350\text{ V wkg}$
$R_{13} = 120\text{ k}\Omega \pm 5\%, \frac{1}{4}\text{ W}^1)$	$C_{10} = 2 \times 50\ \mu\text{F}, 350\text{ V wkg}$
$R_{14} = 1\text{ k}\Omega \pm 20\%, \frac{1}{4}\text{ W}$	
$R_{15} = 150\ \Omega \pm 5\% 1\text{ W}$	

¹⁾ High stability.

EL 86 (I) is controlled from the phase inverter, which is in turn coupled to the input stage by an $0.1 \mu\text{F}$ capacitor. Tube EL 86 (I), then, works with high-pass filters in cascade, so the capacitors have to be twice as large.

A small capacitor ($C_5 = 47 \text{ pF}$) is connected across the resistors in the cathode circuit of the phase inverter to compensate for the Miller effect, which tends to attenuate the higher frequencies. This capacitor reduces the negative feedback at those frequencies and so the gain increases.

The power supply section is very simple. A transformer giving $2 \times 300 \text{ V}$, 90 mA ; 6.3 V , 2.4 A and 6.3 V 1 A will be sufficient to feed the amplifier and a pre-amplifier as well. The rectifier EZ 81 is used, but it is by no means fully loaded. The smoothing filter consists of a double $50 \mu\text{F}$ electrolytic capacitor with a 150Ω resistor, which must be wire-wound, rated at 6 W . The EZ 81 rectifier requires a transformer resistance of $2 \times 200 \Omega$. Such simple smoothing arrangements are possible because the screen grids of the output tubes are very effectively decoupled by the choke and the input stage is decoupled by means of a $100 \text{ k}\Omega$ resistor with an $8 \mu\text{F}$ capacitor. This capacitor shares

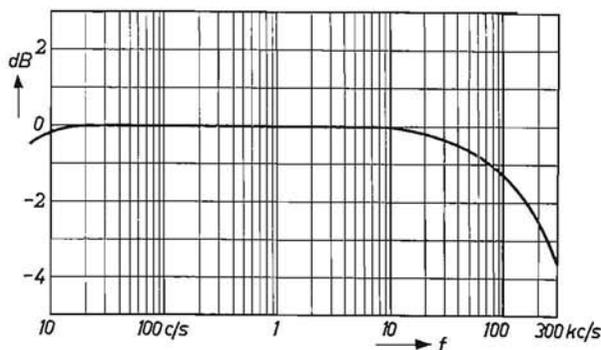


Fig. 49. Frequency response curve of the single-ended push-pull amplifier.

the same can as the $8 \mu\text{F}$ capacitor for the screen grid of EL 86 (II). The $50 \mu\text{F}$ capacitor for the screen grid of EL 86 (I) is similarly combined with the $100 \mu\text{F}$ in series with the high-impedance loudspeaker. In this case a $3 \times 50 \mu\text{F}$ electrolytic capacitor is used, as this saves space on the chassis and requires only one insulating ring.

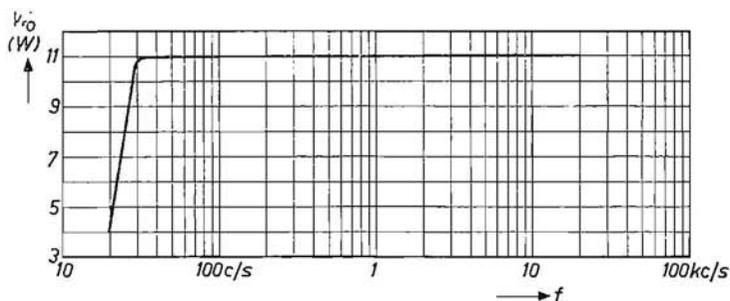


Fig. 50. Output curve of the amplifier measured at 1% distortion.

This power amplifier is intended for use with a separate pre-amplifier which includes the volume and tone controls and equalising network. The circuit shown on p.104 is recommended.

If desired, the pre-amplifier can be made as a separate unit, but it can also be combined with the main amplifier, in which case it is advisable to

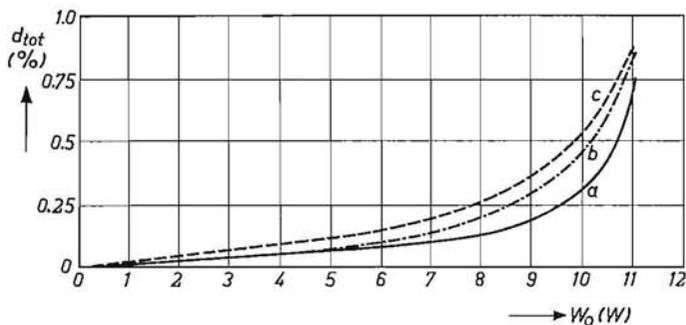


Fig. 51. Total harmonic distortion plotted against the output power (a) at 400 c/s and 1 kc/s (b) at 4 kc/s, (c) at 90 c/s.

screen the two sections from each other; in this respect a great deal depends on the layout of the components.

Measured results

The response curve is given in Fig. 49 and is seen to be practically flat (less

than 0.5 dB drop) from 7 c/s to 40 kc/s. The 3-dB points occur at about

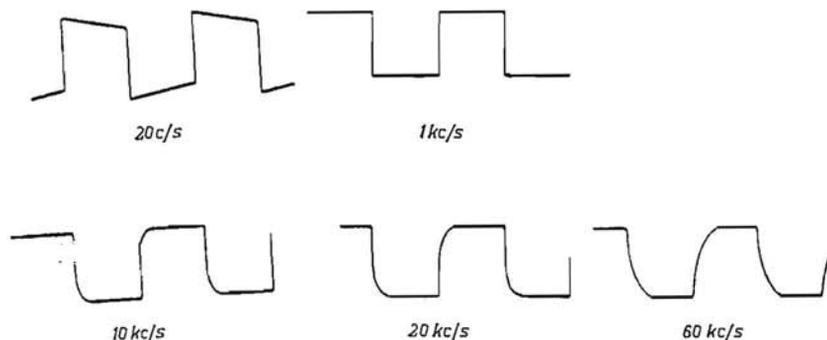


Fig. 52. Square-wave voltages of various frequencies as reproduced by the amplifier on a cathode ray oscilloscope.

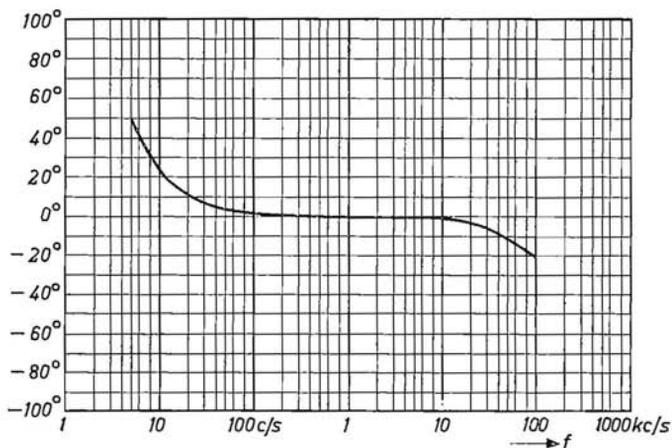


Fig. 53. Phase shift of the 10 W single-ended push-pull amplifier.

3 c/s and 230 kc/s. At 300 kc/s the line drops by only 3.8 dB compared with 1 kc/s.

Fig. 50 shows the power characteristic of the amplifier with 1% distortion. The curve drops sharply at 30 c/s owing to the action of the choke in the

screen grid circuit of EL 86 (I). It was not possible to take measurements above 20 kc/s in view of the limited range of the distortion meter, but at that point no signs of a drop in the curve were evident. The harmonic distortion plotted against output power at 90,400, 1000, and 4000 c/s is shown in Fig. 51. This distortion was also measured at 10 and 20 kc/s at full power, with the following results: less than 1% at both frequencies for 11 W and 0.23% at 10 kc/s and 0.33% at 20 kc/s for 10 W.

The amplifier was also subjected to tests on a square-wave voltage at frequencies of 20 c/s, 1, 10, 20, and 60 kc/s; the results are shown in Fig. 52, from which it is seen that the reproduction is ideal at 1 kc/s and very good at all the other frequencies. The phase shift is shown in Fig. 53.

If no high impedance loudspeakers can be obtained the circuit may be used to drive 15 Ω speakers via a matching transformer. This transformer carries no d.c. and therefore it may be constructed as an auto-transformer with a ratio of 7 : 1. Thus a high efficiency and low stray inductance is obtained, so that its influence on the quality of reproduction is scarcely noticeable. For 3 dB drop at 25 c/s the primary inductance should be about 5 H.

CHAPTER III

PRE-AMPLIFIERS

Introduction

Pre-amplifiers for use with small power amplifiers are usually assembled on the same chassis as the power amplifier, and the wiring etc. is perfectly straightforward. In larger installations, however, a separate pre-amplifier is generally used, combining an equalising network to suit various recording characteristics and broadcast programmes. Nowadays there is a tendency also to include an FM/AM radio tuner unit in the pre-amplifier, which is then provided with its own power supply. If this procedure is adopted the cabinet need not be much higher and wider than the tuning dial and it can be conveniently housed in, say, a bookcase, while the power amplifier, which requires no adjustments, can be hidden from view.

Recording characteristics as employed today are shown in Fig. 54; the dotted line represents the British (R.I.A.A.) and American (A.E.S.) standard for microgroove recordings. The full line is the Decca F.F.R.R. curve, as employed in their test record LXT 2695, and the chain-dotted line is the latest R.I.A.A. 78 r.p.m. recording characteristic. The system on which the older records was based was very simple, dropping by some 5 dB per octave below 500 c/s, above which point it was flat.

When incorporated in pre-amplifiers, equalisation is often achieved by means of negative feedback which is dependent on frequency, the various combinations being switched to suit the recording characteristic.

The pre-amplifier will also include the necessary controls for volume, treble and bass, and sometimes also noise and rumble filters. Very complicated units are obtainable with all sorts of refinements, but only relatively

simple circuits will be dealt with in this chapter. The more complex equipment demands much care in the layout and assembly of the components; careful screening may be necessary, and the use of makes of components other than those used in the prototype may well end in unsatisfactory results, including instability. It is therefore not practicable to build such equipment

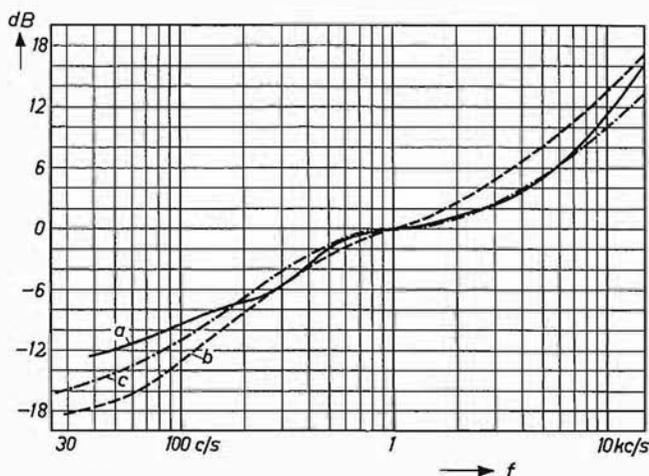


Fig. 54. Disc recording characteristics according to the latest international agreements. (a) Decca F.F.R.R. system, as used for their test record LXT 2695; (b) R.I.A.A. and A.E.S. standard for microgroove records; (c) R.I.A.A. standard for 78 r.p.m. recordings.

using only a simple circuit diagram and a list of components; comprehensive instructions, however, fall outside the scope of this book.

PRE-AMPLIFIER USING TWO EF 86 PENTODES

This design is intended for use with either the 10 W or the 20 W amplifier described in Chapter II and will operate with a crystal or magneto-dynamic pickup, a tape-deck, a microphone and an F.M./A.M. tuner. An auxiliary input socket is provided as to match any desired signal source.

A selector switch on the front panel is used for the various input channels; both crystal and magnetic pickups cannot be used simultaneously, as the same equalising network serves both channels. From left to right the switch positions are: (1) Crystal or magneto-dynamic pickup, 78 r.p.m.; (2) same for $33\frac{1}{2}$ r.p.m.; (3) microphone; (4) tape; (5) radio; (6) auxiliary.

The pickup equalisation is based on the latest R.I.A.A. or A.E.S. recording characteristics, as adopted by most manufacturers of gramophone records today, and the playback characteristic for magnetic tape recordings is matched for a speed of $7\frac{1}{2}$ in./sec.

A low-impedance, wide-range tone control is provided, covering the most divergent conditions of reproduction.

General data

Output voltage	: 40 mV for 10 W amplifier 250 mV for 20 W amplifier
Sensitivity:	crystal pickup, $33\frac{1}{2}$ r.p.m. 50 mV 78 r.p.m. 150 mV magneto-dynamic pickup $33\frac{1}{2}$ r.p.m. 3 mV 78 r.p.m. 9 mV microphone: 6 mV tape: 3 mV at 5 kc/s radio: 250 mV auxiliary: 250 mV
Hum and noise:	measured with 10 W amplifier: — 55 dB at full output for both types of pickup at $33\frac{1}{2}$ r.p.m.; — 57 dB for both types of pickup at 78 r.p.m. microphone: — 44 dB tape: — 53 dB

- Input impedance: 100 k Ω at all pickup settings of the switch
1 M Ω for microphone
80 k Ω for tape
2 M Ω for radio and auxiliary.
- Distortion: less than 0.15% total harmonic distortion at nominal output; 0.24% at 10 times nominal output.
- Tone control: max. bass boost: + 17 dB at 50 c/s
max. bass cut — 14 dB at 50 c/s
max. treble boost: + 14 dB at 10 kc/s
max. treble cut — 15 dB at 10 kc/s
all with respect to the 1 kc/s level.

CIRCUIT DESCRIPTION

This pre-amplifier, the circuit diagram of which is shown in Fig. 55, has two stages, each using the A.F. pentode EF 86. Equalisation is effected wholly in the first stage by means of negative feedback between the anode and control grid of the input tube. Negative feedback is not applied to the output tube, from which the signal is taken direct to a tone control circuit. the grid circuit of the first tube and so reduce the risk of hum being induced at the input terminals. At the same time the gain in the input stage is not too high, an advantage from the point of view of Miller effect, which can be troublesome if high-value resistors are used in the grid circuit of the input tube.

Resistors are connected in series with all the input sockets for matching of the input impedances of the various channels; the values suggested in the list of components are suitable for normal inputs, but the sensitivity, as well as the impedance, can be modified quite simply by using resistors of different values from those given.

The impedance of each input channel comprises the impedance of the EF 86 as modified by the negative feedback, plus that of the input network.

The sensitivity of the pre-amplifier itself can be modified by changing the output voltage, this being effected by altering the value of resistors R_{21} and R_{23} in the anode circuit; the sum of the two values should remain at 100 k Ω . The values of 18 k Ω and 82 k Ω indicated in the circuit are

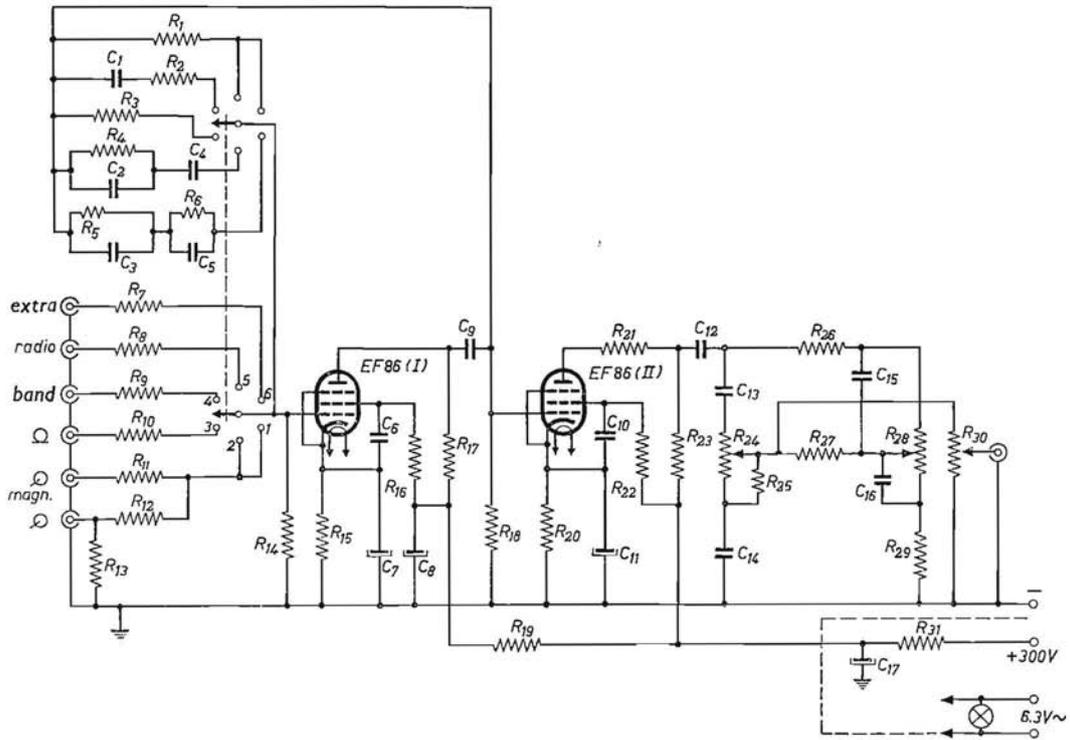


Fig. 55. Circuit diagram of pre-amplifier using two type EF 86 A.F. pentodes.

intended for use with the 10 W amplifier, in which the tone control network has been replaced by a resistor. For use with the 20 W amplifier the output signal should be taken direct from the anode of the tube.

The smoothing filter $R_{31} C_{17}$ should preferably be included in the power amplifier. The total anode current taken by the pre-amplifier is not more than 3 mA at 300 V.

The pickup input channels

Equalisation curves for the two pickup input channels are given in Fig. 56 and it can be seen that these are suitable for the present-day recording characteristics shown in Fig. 54. The input channel for the magneto-dynamic type of pickup can be modified so as to operate satisfactorily with an electrodynamic pickup delivering a larger output voltage, by using a higher value for R_{11} than that given in the figure. The necessary differences between the sensitivities for the $33\frac{1}{3}$ and 78 r.p.m. positions is

List of parts

$R_1 = 330 \text{ k}\Omega \pm 5\%, \frac{1}{4} \text{ W}$	$R_{26} = 68 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_2 = 560 \text{ k}\Omega \pm 5\%, \frac{1}{4} \text{ W}$	$R_{27} = 39 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_3 = 10 \text{ M}\Omega \pm 5\%, \frac{1}{4} \text{ W}$	$R_{28} = 250 \text{ k}\Omega \text{ log.}$
$R_4 = 560 \text{ k}\Omega \pm 5\%, \frac{1}{4} \text{ W}$	$R_{29} = 6.8 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_5 = 5.6 \text{ M}\Omega \pm 5\%, \frac{1}{4} \text{ W}$	$R_{30} = 250 \text{ k}\Omega \text{ log.}$
$R_6 = 220 \text{ k}\Omega \pm 5\%, \frac{1}{4} \text{ W}$	$R_{31} = 30 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$
$R_7 = 2.2 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	
$R_8 = 2.2 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_1 = 390 \text{ pF} \pm 5\%$
$R_9 = 56 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_2 = 150 \text{ pF} \pm 5\%$
$R_{10} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_3 = 2200 \text{ pF} \pm 5\%$
$R_{11} = 68 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_4 = 560 \text{ pF} \pm 5\%$
$R_{12} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_5 = 220 \text{ pF} \pm 5\%$
$R_{13} = 100 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_6 = 0.1 \mu\text{F}, 350 \text{ V wkg.}$
$R_{14} = 100 \text{ k}\Omega \pm 5\%, \frac{1}{4} \text{ W}^1)$	$C_7 = 25 \mu\text{F}, 12 \text{ V wkg.}$
$R_{15} = 2.2 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_8 = 8 \mu\text{F}, 350 \text{ V wkg.}$
$R_{16} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{2} \text{ W}^1)$	$C_9 = 0.1 \mu\text{F}, 350 \text{ V wkg.}$
$R_{17} = 220 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^1)$	$C_{10} = 0.1 \mu\text{F}, 350 \text{ V wkg.}$
$R_{18} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{11} = 25 \mu\text{F}, 12 \text{ V wkg.}$
$R_{19} = 33 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{12} = 0.1 \mu\text{F}, 350 \text{ V wkg.}$
$R_{20} = 1.2 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{13} = 560 \text{ pF} \pm 10\%$
$R_{21} = 82 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^1)$	$C_{14} = 8200 \text{ pF} \pm 10\%$
$R_{22} = 390 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^1)$	$C_{15} = 2200 \text{ pF} \pm 10\%$
$R_{23} = 18 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^1)$	$C_{16} = 20\,000 \text{ pF} \pm 10\%$
$R_{24} = 250 \text{ k}\Omega \text{ log.}$	$C_{17} = 16 \mu\text{F}, 350 \text{ V wkg.}$
$R_{25} = 47 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	

¹⁾ High stability.

achieved partly by using different values for the negative feedback, but mainly by means of an attenuating resistor R_{13} .

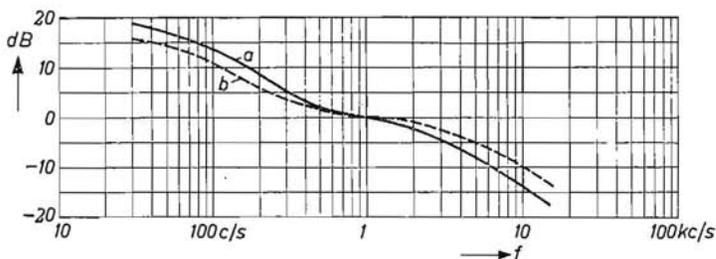


Fig. 56. Equalisation curves: (a) $33\frac{1}{3}$ r.p.m. (b) 78 r.p.m.

The low impedance of the crystal pickup channel ($100\text{ k}\Omega$) results in attenuation of the lower frequencies and so renders the response curve similar to that of a magneto-dynamic pickup. Hence the same components can be employed for both channels to produce the negative feedback, so long as a crystal pickup with low or moderate output is used. If the pickup is not suitable for this form of loading or delivers a relatively high voltage, it can be connected to the auxiliary socket, the object of which will be discussed later.

For both the pickup channels the hum and noise level, measured with the 10 W amplifier of Fig. 27 is 55 dB below 10 W at $33\frac{1}{3}$ r.p.m. or 57 dB below at 78 r.p.m. In both cases the input impedance is approximately $100\text{ k}\Omega$. The sensitivity with a magneto-dynamic pickup is 3 mV and 9 mV at 1 kc/s as for micro-groove and 78 r.p.m. respectively. For crystal pickups the sensitivity is 50 mV and 150 mV respectively, in both cases for full output of the power amplifier.

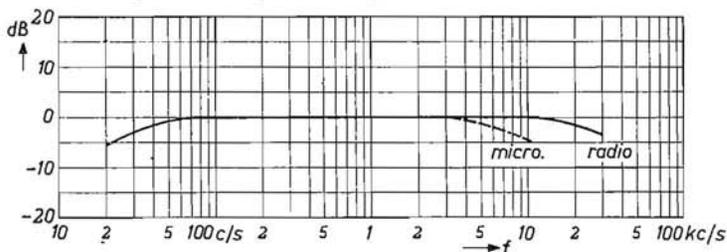


Fig. 57. Response curve for the radio and auxiliary channels (full line) and microphone (dotted line).

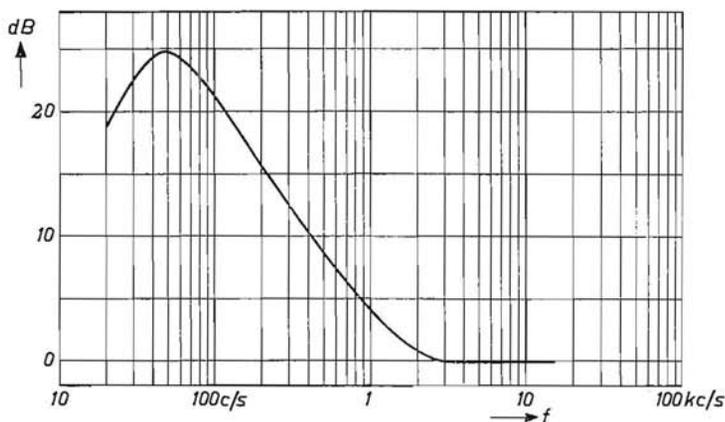


Fig. 58. Equalisation curve for magnetic tape playback.

Microphone

The microphone channel is intended for use with a crystal microphone, or a magnetic microphone with transformer. The sensitivity is 6 mV and the input impedance 1 M Ω , with a hum and noise level 44 dB below 10 W. Fig. 57 shows the response curve.

Tape playback

The response curve of the pre-amplifier on tape playback is shown in Fig. 58; this follows the C.C.I.R. curve down to 100 c/s, but below that frequency less boost is employed. The input impedance is about 80 k Ω , the sensitivity 3 mV at 5 kc/s; hum and noise 52 dB below 10 W.

This channel is suitable for the replay of pre-recorded tapes, when using a high-impedance playback head, and the response characteristic chosen will guarantee excellent and well-balanced results. Should the sensitivity prove too low for a given playback head, however, this can be increased by reducing the value of R_9 (56 k Ω) until the required sensitivity is reached.

Radio input

The frequency characteristic of the radio input channel can be seen from Fig. 57. With de-emphasis incorporated in the tuner unit the circuit ensures

good results from F.M. broadcasts. For reception of ordinary A.M. medium- and long-wave broadcasts from other than local stations it may be necessary to attenuate the higher frequencies to some extent. In this channel the sensitivity is 250 mV with an input impedance of $2\text{ M}\Omega$; this will usually be sufficient, but other values may be obtained by changing the value of the negative feedback resistors R_1 ($330\text{ k}\Omega$) and R_8 ($2.2\text{ M}\Omega$). Should the input impedance be too high, however, it can be reduced by connecting a resistor of the required value between the input socket and earth.

The auxiliary channel

It will be seen from the circuit diagram that the auxiliary channel is identical to that for radio input. This channel is suitable for use with a tape deck working at speeds other than $7\frac{1}{2}$ in/sec. or with a characteristic other than the C.C.I.R. It can also be used for very sensitive crystal pickups; if the input is less than 250 mV, however, the resistor R_7 may be reduced to about $1\text{ M}\Omega$.

Volume and tone controls

The volume and tone controls are in the anode circuit of the second tube. Low-value potentiometers ($250\text{ k}\Omega$) are used to ensure that the capacitance of long coaxial cables between this unit and the power amplifier will not affect the treble response.

The effect of the tone controls at their extreme settings may be seen from Fig. 59 and should be quite adequate for all ordinary purposes.

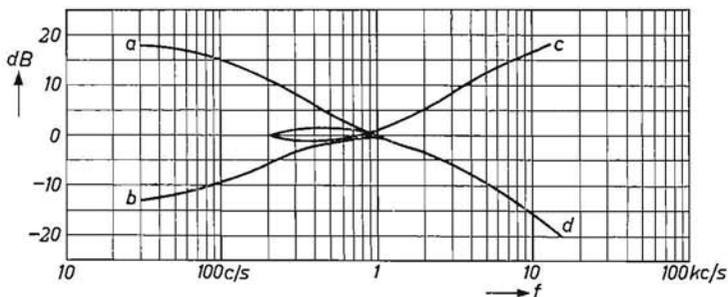


Fig. 59. Tone controls: (a) bass maximum, treble flat; (b) bass minimum, treble flat; (c) treble maximum, bass flat, (d) treble minimum, bass flat.

Output voltage

The values of R_{21} and R_{23} were chosen for use with the 10 W amplifiers described in Chapter II. With the tone and volume controls in these amplifiers replaced by a 1 M Ω resistor, the sensitivity will be 40 mV for 10 W output, thus matching the output of the pre-amplifier.

If the pre-amplifier is to be used with the 20 W amplifier using two EL 34 pentodes in push-pull, however, the output voltage of the pre-amplifier must be increased to 250 mV for full output; to do this it is only necessary to replace resistors R_{21} and R_{23} by a single 100 k Ω resistor and to connect C_{12} to the anode of the second EF 86. This applies also for use with the 10 W amplifier having the EF 86 connected as a triode.

If still higher output voltages are required, for use with the 4 x EL 84 20 W amplifier, the values of the series resistors in the input circuits can be reduced; this of course also reduces the input impedance.

TWO-TUBE PRE-AMPLIFIER FOR USE WITH A MAGNETO-DYNAMIC PICKUP

This amplifier was designed for use with all types of magneto-dynamic pickups and, in particular, to drive the single-ended push-pull amplifier described on page 85; it can be included on the same chassis, provided it is properly screened from the power amplifier. Needless to say, it is also suitable for use with any other power amplifier.

The input tube operates with frequency dependent negative feedback, to provide equalisation for $33\frac{1}{3}$ r.p.m. recordings on the basis of R.I.A.A., S.A.E. or Decca F.F.R.R. characteristics as well as the older 78 r.p.m. recordings. A switch is provided for this purpose, having three positions for equalisation and a fourth position in which the first stage is cut out, the grid of the second tube being then connected direct to the volume control for use with a radio tuner unit or equalised tape, etc.

There are two volume controls, one of which is built-in and pre-set, the other providing manual control from the front panel. Ample bass and treble control has been provided, the result is a wide range of control which renders the amplifier very suitable for various signal sources, combinations of loudspeakers and local acoustical conditions.

General data

- Tubes: EF 86 and ECC 83.
- Amplification: Input I 100
Input II 8
- Distortion: Total harmonic distortion for $0.5 V_{\text{rms}}$ output voltage on maximum input ($V_i = 200$ mV):
at 1 kc/s, $d_{\text{tot}} < 0.02\%$
80 c/s, $d_{\text{tot}} < 0.06\%$
- Hum and noise: measured at the output of the single-ended push-pull amplifier: - 60 dB on full output, and with volume control set to minimum: 20 mV.
- Tone control: bass boost max. + 12 dB at 25 c/s
bass cut max. - 18 dB at 25 c/s
treble boost max. + 8 dB at 15 kc/s
treble cut max. - 14 dB at 15 kc/s
compared with the 1 kc/s level.

CIRCUIT DESCRIPTION

The circuit diagram is given in Fig. 60; the two tubes used are the EF 86 and the ECC 83. Normal values are chosen for the anode, cathode and screen grid resistors, and the circuit employs frequency-dependent negative feedback. A three-pole, four-position switch gives the following positions: First equalisation for $33\frac{1}{3}$ r.p.m. recordings with R.I.A.A. or S.A.E. characteristics. The output signal from the tube is taken by way of a coupling capacitor C_7 and C_4 to R_3 which is shunted by C_6 . The grid of the input tube is earthed via R_2 in series with a magneto-dynamic pickup. This circuit is not specially adapted to any particular make of pickup, but is equally suitable for all current types; the pickup must remain connected if the other input socket is used, however.

The second position of the switch provides equalisation for Decca F.F.R.R. recordings. Capacitor C_4 is now replaced by C_3 and R_1 in parallel, and C_5 is connected across R_3 . The third position is for equalising the old 78 r.p.m. recordings, the response curve of which is flat to 500 c/s, thence falling off at about 5 dB per octave. Compensation is provided by a simple filter consisting of C_2 in series with R_9 .

In the fourth position the input stage is switched off and the second input socket is connected to the volume control; this channel can be used for an FM/AM radio tuner, or for the playback of equalised tape recordings. A crystal pickup can also be used here, but there is then no equalisation and the tone control is used instead.

The two sections of the ECC 83 are connected in cascade, with a high degree of negative feedback from the anode of the second section to the cathode of the first, by way of C_{14} and R_{14} . This feedback reduces the output impedance of the amplifier and a long coaxial cable can therefore be used to connect the pre-amplifier to the power amplifier without loss of treble. Two volume controls are provided, one of which is operated from the front panel (R_9) and the other (R_{13}) from the rear, by means of a screwdriver. The first of these (R_9) is of the manual type, and is of $1\text{ M}\Omega$ with tappings at 0.1 and $0.2\text{ M}\Omega$ from the earth end. These tappings are connected to earth through high-pass filters so that, when the control is turned back, the middle frequencies are attenuated to a relatively greater extent than the lower frequencies. With considerable attenuation the treble is given a

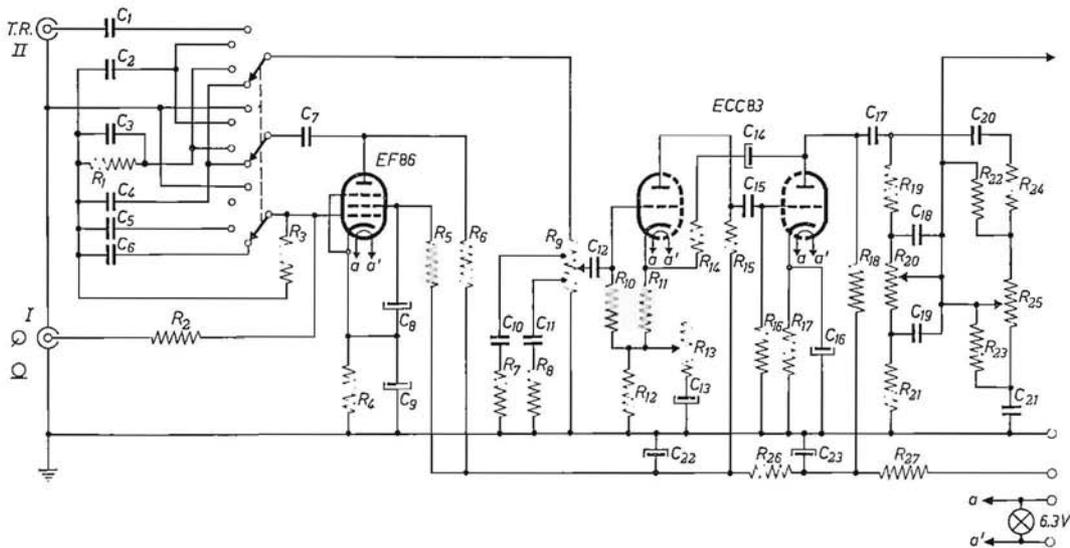


Fig. 60. Circuit diagram of pre-amplifier using EF 86 en ECC 83, for use with the single-ended push-pull amplifier.

certain amount of boost. The attenuation curves are shown in Fig. 61; these are converse of the Fletcher-Munson curves for the average sensitivity of the human ear to frequency at various sound levels.

With this type of volume control, the effect depends considerably on the acoustics of the room in which the amplifier is used. For this reason a second volume control is provided, pre-set to an average level, so that when the manual control is operated it is not necessary to readjust the tone controls.

Owing to the considerable negative feedback (29 dB) in the anode circuit of the EF 86 and about 38 dB on the ECC 83, the amount of distortion is extremely low. This was measured in conjunction with the 10 W single-ended push-pull amplifier and was found to be 0.02% at 1 kc/s, with 0.06% at 80 c/s.

The ECC 83 is followed by a conventional tone control network, the characteristics of which, for the extreme settings, may be seen from Fig. 62.

List of parts

$R_1 = 8.2 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_1 = 22\,000 \text{ pF}$
$R_2 = 68 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_2 = 470 \text{ pF}$
$R_3 = 680 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_3 = 330 \text{ pF}$
$R_4 = 2.2 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_4 = 330 \text{ pF}$
$R_5 = 390 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^1)$	$C_5 = 100 \text{ pF}$
$R_6 = 100 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^1)$	$C_6 = 120 \text{ pF}$
$R_7 = 47 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_7 = 10\,000 \text{ pF}$
$R_8 = 800 + 100 + 100 \text{ k}\Omega, \text{ log.}$	$C_8 = 8 \mu\text{F}, 300 \text{ V wkg.}$
$R_{10} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_9 = 100 \mu\text{F}, 3 \text{ V wkg.}$
$R_{11} = 2.2 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{10} = 5600 \text{ pF}$
$R_{12} = 33 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{11} = 33\,000 \text{ pF}$
$R_{13} = 100 \text{ k}\Omega, \text{ lin.}$	$C_{12} = 0.1 \mu\text{F}, 125 \text{ V wkg.}$
$R_{14} = 120 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{13} = 10 \mu\text{F}, 70 \text{ V wkg.}$
$R_{15} = 220 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{14} = 8 \mu\text{F}, 300 \text{ V wkg.}$
$R_{16} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{15} = 10\,000 \text{ pF}$
$R_{17} = 1 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{16} = 100 \mu\text{F}, 3 \text{ V wkg.}$
$R_{18} = 100 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{17} = 0.1 \mu\text{F}, 400 \text{ V wkg.}$
$R_{19} = 100 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{18} = 4700 \text{ pF}$
$R_{20} = 1 \text{ M}\Omega, \text{ log.}$	$C_{19} = 22\,000 \text{ pF}$
$R_{21} = 22 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{20} = 390 \text{ pF}$
$R_{22} = 470 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{21} = 4700 \text{ pF}$
$R_{23} = 100 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{22} \{ 2 \times 16 \mu\text{F}, 300 \text{ V wkg.}$
$R_{24} = 10 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{23} \}$
$R_{25} = 1 \text{ M}\Omega, \text{ log.}$	
$R_{26} = 47 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	
$R_{27} = 22 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	

¹⁾ High stability.

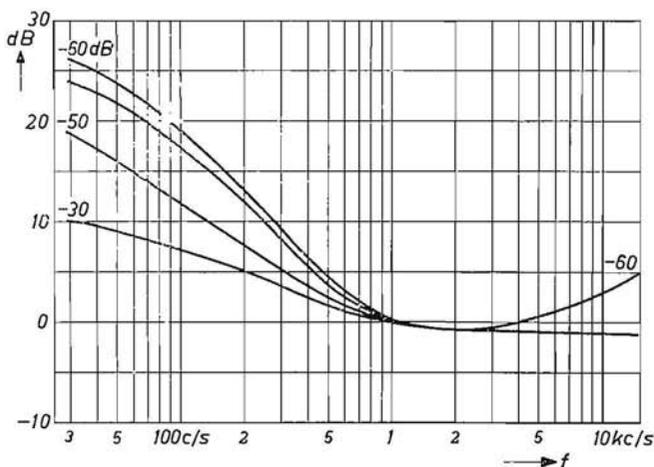


Fig. 61. Relative attenuation of the manual volume control.

The equalisation curves are given in Fig. 63 and the response curve obtained with Decca Test Record LXT 2695 in Fig. 64. The equalisation is seen to be highly effective, the curves being the complements of those of Fig. 54.

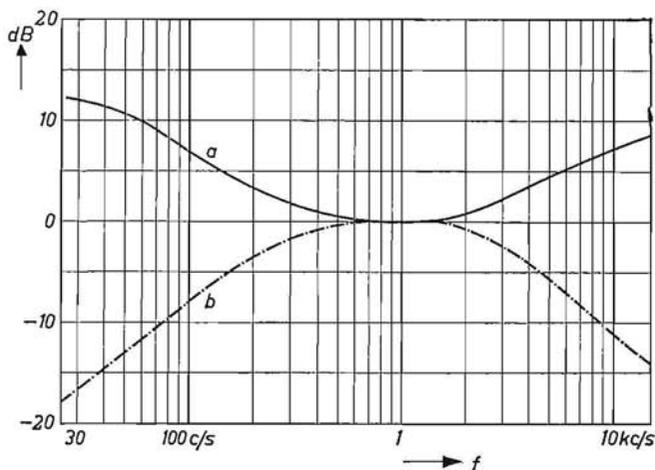


Fig. 62. Tone control: (a) bass and treble control at maximum, (b) minimum.

The pre-amplifier takes about 4 mA at 300 V, which can easily be supplied by the power amplifier; the 5 k Ω series resistor with the usual smoothing capacitor is best included in the power amplifier. The heater current (0.5 A at 6.3 V) can also be supplied by the main power transformer, preferably by a separate winding with centre-tapping.

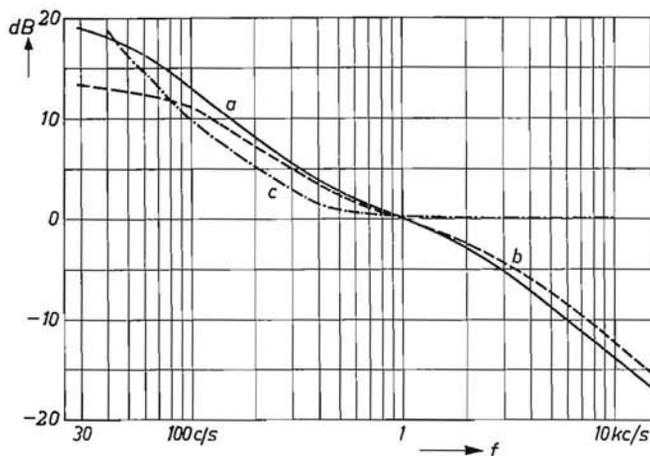


Fig. 63. Equalisation curves (a) 33 $\frac{1}{3}$ r.p.m. recordings, R.I.A.A. and A.E.S. system, (b) Decca F.F.R.R. (c) 78 r.p.m. recordings.

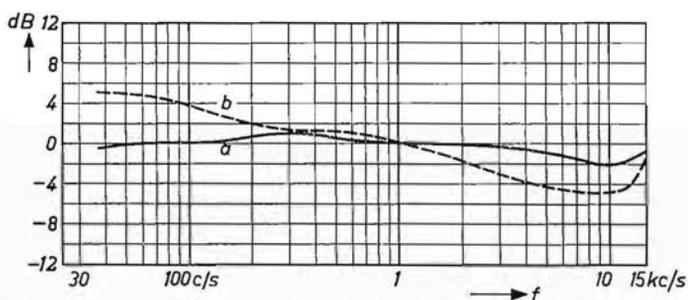


Fig. 64. Response curves: (a) played back with Decca test record LXT 2695, (b) equalized according to the A.E.S. and R.I.A.A. micro-groove recording characteristics.

PRE-AMPLIFIER WITH NOISE AND RUMBLE FILTERS,
USING TWO EF 86 PENTODES AND THE ECC 83
DOUBLE TRIODE

Although expressly designed for use with the 20 W amplifier with two EL 34 pentodes in push-pull, this pre-amplifier is quite suitable for driving other power amplifiers as well. Input sockets are provided for magneto-dynamic and crystal pickups, radio tuners and tape recorder play-back heads; there is also an auxiliary input which can be used for tape amplifiers or a microphone, but this necessitates a separate microphone amplifier.

Apart from the usual output terminals for connection to the power amplifier a special outlet is provided for use with a tape recorder for recording on tape any programme applied to the input. This extra channel is not influenced by the tone control.

It is convenient to have the auxiliary input and additional output sockets on the front panel for easy access. All the other sockets would be at the rear.

Very effective noise and rumble filters are provided, each of which can be set to three different cut-off frequencies. The ECC 83 double triode is used here as an extra amplifying tube to compensate the attenuation of these filters, and for this reason it is possible to employ rather more negative feedback than in the pre-amplifier of Fig. 55, from which this circuit was developed. There is therefore even less distortion than in the previous circuit.

General data

Tubes:	2 x EF 86, ECC 83.	
Output voltage:	for power amplifier	250 mV
	for tape recorder	300 mV
Sensitivity:	crystal pickup 33 $\frac{1}{3}$ r.p.m.	150 mV
	78 r.p.m.	270 mV
	magneto-dynamic	
	pickup 33 $\frac{1}{3}$ r.p.m.	7 mV
	78 r.p.m.	12 mV
	tape playback	2.5 mV at 5 kc/s
radio input	250 mV	
auxiliary	250 mV	
Frequency response:	flat within 1 dB from 25 c/s to 20 kc/s.	

Hum and noise:	measured with 20 W amplifier (nominal output power)	
	33 $\frac{1}{3}$ r.p.m. recordings	— 53 dB
	78 r.p.m. recordings	— 58 dB
	tape play-back	— 47 dB
	radio and auxiliary	— 63 dB
Input impedance:	pickup sockets	100 k Ω
	tape play-back	80 k Ω (approx.)
	radio and auxiliary	1 M Ω
Distortion:	250 mV output:	< 0.1%
	2.5 V output:	0.65%
Tone control:	bass boost maximum	+ 18 dB at 50 c/s
	bass cut maximum	— 10 dB at 50 c/s
	treble boost maximum	+ 17 dB at 10 kc/s
	treble cut maximum	— 17 dB at 10 kc/s
	all referred to the 1 kc/s level	
Noise filter:	cut-off frequencies 9, 7 and 5 kc/s; attenuation 20 dB per octave.	
Rumble filter:	cut-off frequencies 200, 100 and 60 c/s; attenuation 12 dB per octave.	

CIRCUIT OF DESCRIPTION

The circuit diagram is given in Fig. 65 from which it will be seen that the input stage is very similar to that of the pre-amplifier of Fig. 55; it is in fact based on the latter, and the remarks concerning the EF 86 (I) under that heading apply here as well. In this design no special input socket is provided for a microphone, instead, the auxiliary socket can be used, this being placed on the front panel for convenience. A separate microphone amplifier is necessary in this case and a small transistorised amplifier will serve this purpose.

The main difference compared with the previous design lies in the EF 86 (II). As this tube also operates with negative feedback (via R_{19}) there is even less distortion. The output signal from this tube is taken to a conventional tone control network. It is also taken to a socket which can be used for a tape recorder, so that a programme can be listened to and recorded at the same time without being affected by the tone control or filters. This socket could

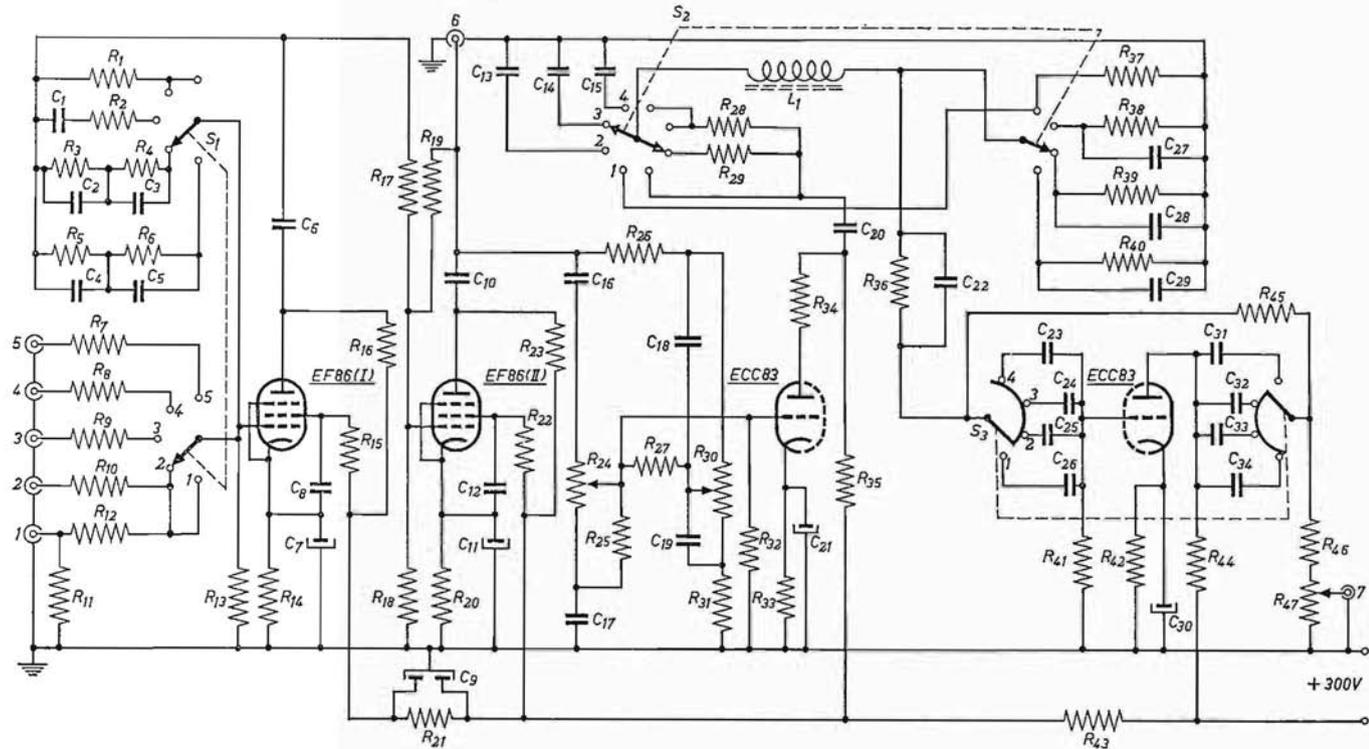


Fig. 65. Circuit diagram of pre-amplifier using 2 x EF 86 and the ECC 83 with switched rumble and noise filters.

also be mounted on the front panel, together with the auxiliary input socket. The output signal for recording purposes is 300 mV.

The tone control circuit is followed by one section of the double triode ECC 83 which functions as an ordinary amplifier without negative feedback. In order to reduce possible interaction between the noise filter and the tone control by way of the tube and wiring capacitances the coupling capacitor C_{20} is connected to a tap on the loading resistor R_{34} R_{35} .

Noise and rumble filters

The noise filters are operated with a three pole, four-position switch, the first of which is the "off" position with flat response; Fig. 66 is a simplified diagram of the circuit between the anode of the first section of the ECC 83 and the grid of the second. The signal is passed by way of the coupling capacitor C_{20} to the junction of R_{28} and R_{37} and thence by a filter R_{36} - C_{22} and coupling capacitor C_{23-26} to the grid of the second section of the tube. The potential divider R_{28} R_{37} serves to keep the overall effect of the signal constant when the noise filters are switched, and the high-pass filter R_{36} C_{22} to compensate for the attenuation at the higher frequencies caused by tube and wiring

List of parts

$R_1 = 100 \text{ k}\Omega$	$R_{22} = 470 \text{ k}\Omega^1)$	$R_{43} = 12 \text{ k}\Omega$	$C_{16} = 560 \text{ pF}$
$R_2 = 390 \text{ k}\Omega$	$R_{23} = 100 \text{ k}\Omega^1)$	$R_{44} = 47 \text{ k}\Omega$	$C_{17} = 8200 \text{ pF}$
$R_3 = 8.2 \text{ M}\Omega$	$R_{24} = 250 \text{ k}\Omega \text{ log.}$	$R_{45} = 1.5 \text{ M}\Omega$	$C_{18} = 2200 \text{ pF}$
$R_4 = 270 \text{ k}\Omega$	$R_{25} = 47 \text{ k}\Omega$	$R_{46} = 220 \text{ k}\Omega$	$C_{19} = 0.02 \mu\text{F}$
$R_5 = 2.2 \text{ M}\Omega$	$R_{26} = 68 \text{ k}\Omega$	$R_{47} = 50 \text{ k}\Omega \text{ log.}$	$C_{20} = 0.25 \mu\text{F}$
$R_6 = 180 \text{ k}\Omega$	$R_{27} = 39 \text{ k}\Omega$		$C_{21} = 50 \mu\text{F}$
$R_7 = 1 \text{ M}\Omega$	$R_{28} = 22 \text{ k}\Omega$	$C_1 = 330 \text{ pF}$	$C_{22} = 33 \text{ pF}$
$R_8 = 1 \text{ M}\Omega$	$R_{29} = 10 \text{ k}\Omega$	$C_2 = 820 \text{ pF}$	$C_{23} = 820 \text{ pF}$
$R_9 = 56 \text{ k}\Omega$	$R_{30} = 250 \text{ k}\Omega \text{ log.}$	$C_3 = 330 \text{ pF}$	$C_{24} = 820 \text{ pF}$
$R_{10} = 100 \text{ k}\Omega$	$R_{31} = 6.8 \text{ k}\Omega$	$C_4 = 2700 \text{ pF}$	$C_{25} = 1500 \text{ pF}$
$R_{11} = 100 \text{ k}\Omega$	$R_{32} = 270 \text{ k}\Omega$	$C_5 = 330 \text{ pF}$	$C_{26} = 3300 \text{ pF}$
$R_{12} = 2.2 \text{ M}\Omega$	$R_{33} = 1.2 \text{ k}\Omega$	$C_6 = 0.1 \mu\text{F}$	$C_{27} = 390 \text{ pF}$
$R_{13} = 100 \text{ k}\Omega$	$R_{34} = 22 \text{ k}\Omega$	$C_7 = 50 \mu\text{F}$	$C_{28} = 820 \text{ pF}$
$R_{14} = 3.9 \text{ k}\Omega$	$R_{35} = 33 \text{ k}\Omega$	$C_8 = 0.1 \mu\text{F}$	$C_{29} = 1800 \text{ pF}$
$R_{15} = 1.5 \text{ M}\Omega^1)$	$R_{36} = 270 \text{ k}\Omega$	$C_9 = 2 \times 8 \mu\text{F}$	$C_{30} = 50 \mu\text{F}$
$R_{16} = 270 \text{ k}\Omega^1)$	$R_{37} = 56 \text{ k}\Omega$	$C_{10} = 0.1 \mu\text{F}$	$C_{31} = 0.01 \mu\text{F}$
$R_{17} = 220 \text{ k}\Omega$	$R_{38} = 56 \text{ k}\Omega$	$C_{11} = 50 \mu\text{F}$	$C_{32} = 4700 \text{ pF}$
$R_{18} = 1 \text{ M}\Omega$	$R_{39} = 47 \text{ k}\Omega$	$C_{12} = 0.1 \mu\text{F}$	$C_{33} = 2700 \text{ pF}$
$R_{19} = 3.9 \text{ M}\Omega$	$R_{40} = 33 \text{ k}\Omega$	$C_{13} = 390 \text{ pF}$	$C_{34} = 2700 \text{ pF}$
$R_{20} = 1.2 \text{ k}\Omega$	$R_{41} = 220 \text{ k}\Omega$	$C_{14} = 820 \text{ pF}$	
$R_{21} = 33 \text{ k}\Omega$	$R_{42} = 1.5 \text{ k}\Omega$	$C_{15} = 1800 \text{ pF}$	
			$L_1 = 1.2 \text{ H}$

¹⁾ High stability.

capacitances. At the other positions of the switch a π -filter having an inductance of 1.2 H is placed in the horizontal line (Fig. 67), with a capacitor in the first vertical leg and a resistor shunted by a capacitor in the second. In the fourth position of the switch the series resistor R_{28} R_{29} is

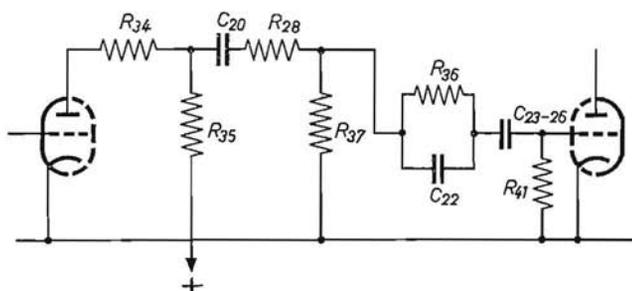


Fig. 66. Basic circuit of the network between the anode of the first section of the ECC 83 and the grid of the second section, with the filter inoperative.

excluded. The filter R_{36} C_{22} remains in circuit however; this has a time constant of 9μ s and thus does not affect the noise filter.

The rumble filter is of very simple construction. The attenuation due to an RC filter approaches asymptotically to 6 dB per octave, and two such filters in series therefore give an attenuation of 12 dB/octave. Now the coupling capacitor and grid leak of a tube constitute a high-pass filter and, if a smaller

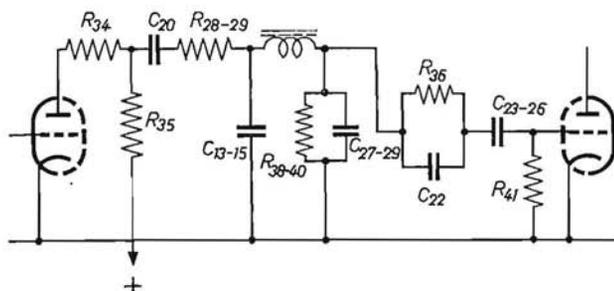


Fig. 67. As Fig. 66 but with the filter in circuit.

coupling capacitor is used, the frequency up to which the characteristic is straight is increased. Below this point the curve drops by about 6 dB/octave. A similar filter in the anode circuit of the tube thus results in an attenuation of 12 dB/octave and this system, which is illustrated in Fig. 68, has been applied to the ECC 83. In the practical form of the filter the coupling capacitors are connected in parallel by a continuous segment in the double-pole four-position switch in such a way as to avoid switch clicks. The pass-curves of the two filters are shown in Fig. 69.

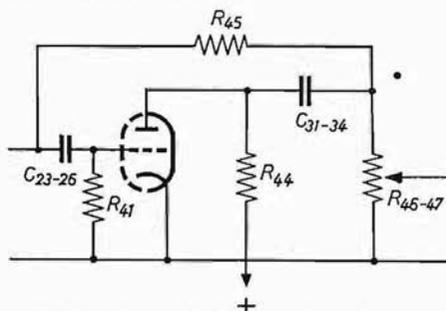


Fig. 68. Simplified circuit of rumble filter.

The output circuit

The output circuit is also quite simple. It comprises a 50 k Ω potentiometer for the volume control in series with a 220 k Ω resistor. Together these form the resistive side of a high-pass filter in the anode circuit of the output tube.

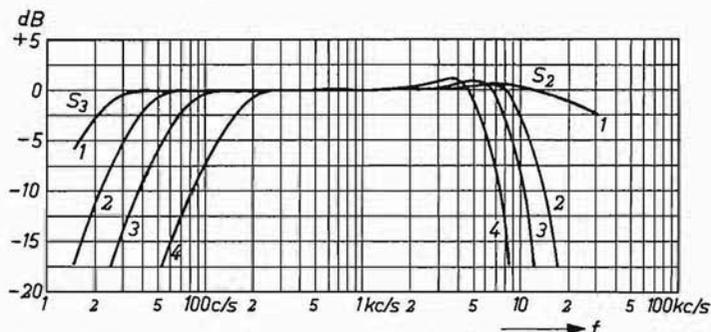


Fig. 69. Pass-curves of the noise and rumble filters. The switch positions S_2 and S_3 correspond to those in the circuit diagram, Fig. 65.

Negative feedback is applied to the tube by way of R_{45} . This has the effect of reducing the internal resistance, which in turn is an advantage at the filter input, apart from the reduction in distortion.

The values as given for the hum and noise levels were not measured at the output of the pre-amplifier, but at the output terminals of a 20 W power amplifier used in conjunction with it, with open input channels. Response and tone control curves were also measured in this way. The sensitivity values as quoted relate to 250 mV output from the pre-amplifier, with less than 0.1% distortion. At ten times greater output voltage the distortion was found to be only 0.65%. A rapid increase in distortion takes place only when the pre-amplifier is heavily overloaded.

Equalisation

Fig. 70 shows the equalisation curves in respect of the input channels for magneto-dynamic and crystal pickups. The difference in sensitivity between the switch positions for $33\frac{1}{3}$ and 78 r.p.m. (positions 1 and 2) is achieved by varying the amount of negative feedback.

For the first type of pickup the input impedance is 100 k Ω ; the sensitivity at 1 kc/s for $33\frac{1}{3}$ r.p.m. recordings is 7 mV and that for 78 r.p.m. recordings 12 mV. The hum and noise levels are then 53 and 58 dB below 20 W respectively. Although this input is intended for a magneto-dynamic pickup, it can also be used for the electrodynamic type, which usually delivers a larger output. It is then advisable to increase the value of R_{10} by a suitable amount.

For a crystal pickup (socket 1) the input impedance is also 100 k Ω , but the sensitivities for $33\frac{1}{3}$ and 78 r.p.m. recordings are 150 and 270 mV respectively.

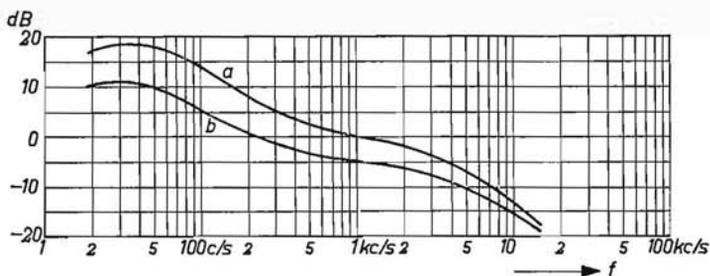


Fig. 70. Equalisation curves for the pickup channels (a) $33\frac{1}{3}$ r.p.m., (b) 78 r.p.m.

The hum and noise level is the same as at the input for magneto-dynamic pickups.

This input is suitable for use with a pickup giving a low or medium output, to be loaded with $100\text{ k}\Omega$ (R_{11}) across the input. This resistance, which is in series with the capacitance of the pickup, causes some loss of bass notes, with the result that the response curve resembles that of a magneto-dynamic pickup, so the same negative feedback network is suitable for both types of pickup; the values specified represent the best possible compromise for present-day makes of pickup.

If, however the pickup to be used is of a type that should not be loaded in this manner, or if its output is too great, the auxiliary channel (5) can be used; here the load is $1\text{ M}\Omega$ and the recording characteristic is automatically equalised.

Tape playback

Input socket No. 3 is intended for playing back tape. The input impedance is $80\text{ k}\Omega$, the sensitivity 2.5 mV at 5 kc/s and the hum and noise level — 47 dB at 20 W . The equalisation curve is shown in Fig. 71.

This channel is intended for use with playback heads having a medium or high resistance. The response curve as obtained with the E.M.I. test tape T.B.T.1. is shown in Fig. 72.

Radio input

The sensitivity of this channel (No. 4) is 250 mV and the input impedance $1\text{ M}\Omega$, with hum and noise level — 63 dB with respect to 20 W . These values should give good results with almost any tuning unit, but if a lower input impedance is required it is only necessary to connect a resistor between the input side of R_6 and earth.

Auxiliary input

It will be seen from the circuit diagram that this input channel is identical with the radio channel. It can be used for sensitive crystal pickups, or for the output signal from a tape recorder; in the prototype of the amplifier this socket

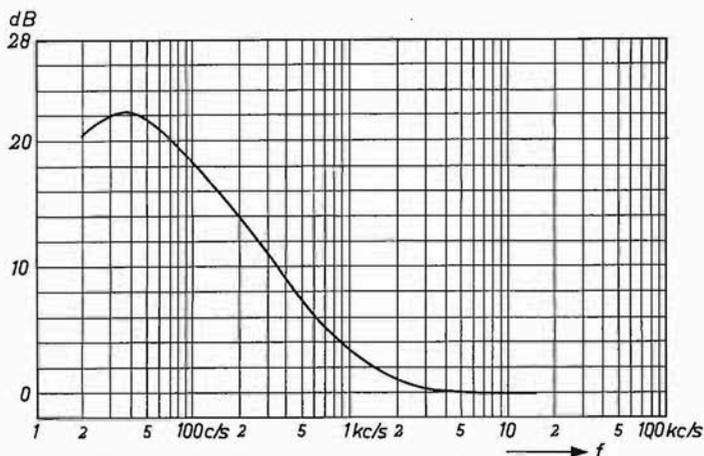


Fig. 71. Equalising curve for magnetic tape playback.

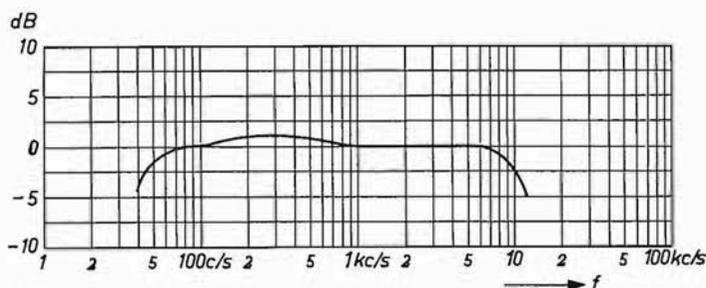


Fig. 72. Response curve of the tape playback channel, for the E.M.I. test tape T.B.T.I.

was placed on the front panel to provide an easily accessible connection for a portable tape recorder.

A microphone with separate pre-amplifier can also be plugged into this socket, or alternatively a mixing amplifier such as the one described in the next section.

The suggested arrangement offers a wide range of possibilities and, if a five-way switch is provided, an extra segment can be provided for short circuiting those channels which are not actually in use. This is not shown in the circuit diagram.

Tone control

Fig. 73 shows the tone control curves for (a) maximum bass, treble flat; (b) minimum bass, treble flat; (c) maximum treble, bass flat; (d) minimum treble, bass flat, and (e) both controls in their central position.

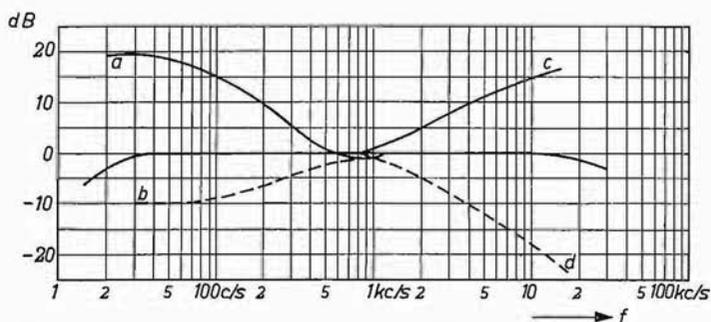


Fig. 73. Response curves of the pre-amplifier with tone controls set to the central and extreme positions. (a) max. bass, treble flat; (b) min. bass, treble flat; (c) max. treble, bass flat; (d) min. treble, bass flat.

FOUR-CHANNEL MIXING AMPLIFIER

By means of the following amplifier, up to four signals can be mixed to produce a single output signal suitable for feeding any ordinary single-channel amplifier or pre-amplifier. Two of the input channels are suitable for microphones, the third is intended for radio or equalised tape, and the fourth can be used for a pickup.

This amplifier has been designed for an output of 40 mV and it is accordingly suitable for use with the 10 W amplifier described on p. 42 provided the tone control network is omitted. With a minor modification, however, the amplifier can be made to deliver 800 mV.

Simplicity is the keynote of the circuit, with a minimum of components and tubes. It is typically a circuit for amateur constructors, and will give excellent service in making of tape recordings, so long as no professional standards are imposed.

General data

Tubes:	2 × EF 86 and the ECC 83.	
Output voltage:	for low gain:	40 mV
	for high gain:	800 mV
Sensitivity:	microphone input:	3 mV
	radio input:	230 mV
	pickup input:	250 mV
Response curve:	microphone: flat within ± 3 dB (with respect to 1 kc/s level) from 20 c/s to 20 kc/s.	
	Radio and pickup: flat within ± 2 dB from 15 c/s to 20 kc/s.	
Hum and noise:	50 dB below full output.	

CIRCUIT DESCRIPTION

The circuit diagram is given in Fig. 74. The two microphone channels are identical, each making use of the A.F. pentode EF 86, biased by means of a high-resistance grid leak ($R_1 = R_6 = 10 \text{ M}\Omega$).

Crystal microphones generally have an internal impedance that is mainly capacitive (about 2000 pF) so in order to ensure that there will be no loss of output voltage at the lower frequencies these microphones should be

connected to a high-impedance input stage. For example, if R_1 were 1.5 M Ω , the combination of the series capacitance of the microphone itself, the capacitor C_1 in the grid circuit, and the grid leak, would result in a loss of one-third of the signal voltage at 100 c/s. The value of the grid leak chosen, however, is high enough to prevent this loss of bass.

If the microphone is of either the electro-dynamic or the ribbon type, the amplifier can be matched with its low impedance by means of an input transformer to step up the low output voltage. The grid leak, which also provides the bias, is then placed in series with the transformer; this grid leak is bypassed by a 5000 pF capacitor as shown in Fig. 75, which replaces the section A — B in Fig. 74.

The leads from the transformer to the control grid of the EF 86 should be as short as possible, to avoid loss of treble and to eliminate possible hum pick-up.

Each of the microphone amplifier tubes is coupled capacitively to one section of the ECC 83 double triode, the radio and pickup input being taken to the same grid by way of resistors R_{11} and R_{12} . This section of the tube operates as a voltage amplifier.

Potentiometers R_4 , R_9 and R_{13} are for adjustment of the signal strength and for mixing the microphone and pickup signals. In the fourth channel the necessary adjustments can be made with the aid of the volume control incorporated in the radio receiver or tape recorder itself. If desired, however, the radio signal can be controlled in the same way as the pickup signal, by means of R_{13} , but in that case a potentiometer has to be connected across the input and the slider taken to R_{11} .

The values of R_6 , R_{10} , R_{11} and R_{12} are such as to ensure that the various signals will not affect each other and also that the grid of the ECC 83 will not be short circuited when one or more of the potentiometers are turned off.

The output signal from the first half of the ECC 83 is taken through a potential divider R_{16} — R_{18} and C_{10} to the grid of the second section, which operates as a cathode follower; this will be referred to again later. Negative feedback is applied to the first section by way of C_9 , R_{14} , and serves to reduce the input impedance of the tube and also to limit high-note loss due to Miller effect.

The output voltage of the amplifier is taken from the cathode of the second section of the ECC 83. With the large amount of negative feedback the

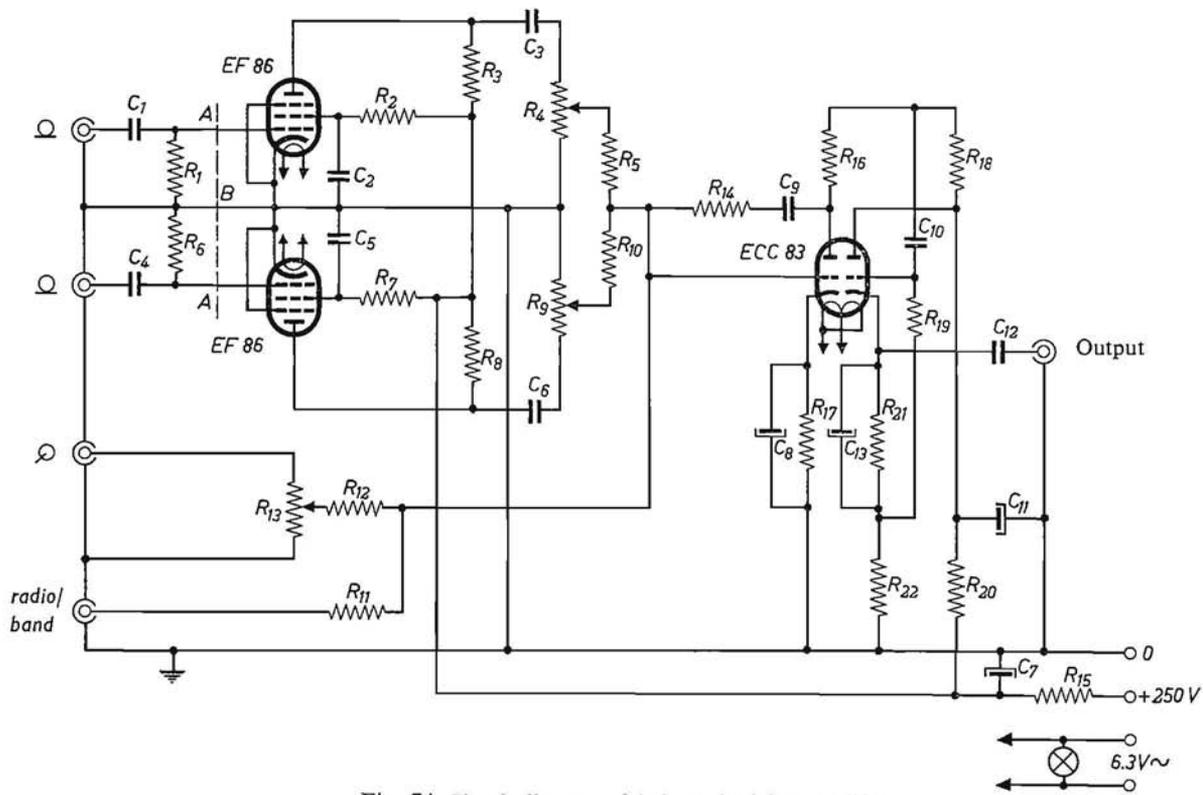


Fig. 74. Circuit diagram of 4-channel mixing amplifier.

output impedance of a cathode follower is very low, in this case about 600Ω , which is low enough to ensure that any long cables between the pre-amplifier and the power amplifier will not cause any loss of treble. The input impedance of the power amplifier, however, should be more than $100 \text{ k}\Omega$ if attenuation of the bass due to the coupling capacitor C_{12} is to be avoided.

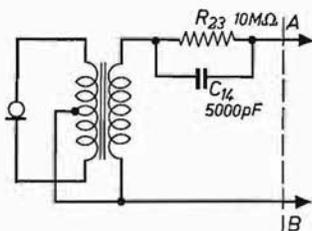


Fig. 75. Method of connecting low impedance microphones. This circuit replaces the section on the left of the line A-B in Fig. 62.

Output voltage and sensitivity

The maximum output voltage delivered by the amplifier is normally 40 mV for an input of 3 mV at the microphone socket or 230 mV at the radio socket. As the input sensitivity for the pickup is 250 mV the amplifier is suitable for use with a crystal pickup giving a medium output.

Should it be desired to use this unit with a power amplifier requiring a

List of parts

$R_1 = 10 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}^1)$	$R_{19} = 1 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_2 = 390 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}^1)$	$R_{20} = 27 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_3 = 100 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}^1)$	$R_{21} = 1.5 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_4 = 500 \text{ k}\Omega, \text{ log.}$	$R_{22} = 47 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$
$R_5 = 470 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	
$R_6 = 10 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}^1)$	$C_1 = 0.05 \mu\text{F}, 250 \text{ V wkg.}$
$R_7 = 390 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}^1)$	$C_2 = 0.25 \mu\text{F}, 250 \text{ V wkg.}$
$R_8 = 100 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}^1)$	$C_3 = 0.1 \mu\text{F}, 250 \text{ V wkg.}$
$R_9 = 500 \text{ k}\Omega, \text{ log.}$	$C_4 = 0.05 \mu\text{F}, 250 \text{ V wkg.}$
$R_{10} = 500 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_5 = 0.25 \mu\text{F}, 250 \text{ V wkg.}$
$R_{11} = 390 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_6 = 0.1 \mu\text{F}, 250 \text{ V wkg.}$
$R_{12} = 470 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_7 = 16 \mu\text{F}, 350 \text{ V wkg.}$
$R_{13} = 500 \text{ k}\Omega, \text{ log.}$	$C_8 = 50 \mu\text{F}, 12 \text{ V wkg.}$
$R_{14} = 1.5 \text{ M}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_9 = 0.05 \mu\text{F}, 250 \text{ V wkg.}$
$R_{15} = 22 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}$	$C_{10} = 0.01 \mu\text{F}, 250 \text{ V wkg.}$
$R_{16} = 100 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^2)$	$C_{11} = 16 \mu\text{F}, 350 \text{ V wkg.}$
$R_{17} = 2.2 \text{ k}\Omega \pm 10\%, \frac{1}{4} \text{ W}$	$C_{12} = 0.1 \mu\text{F}, 250 \text{ V wkg.}$
$R_{18} = 5.6 \text{ k}\Omega \pm 10\%, \frac{1}{2} \text{ W}^2)$	$C_{13} = 50 \mu\text{F}, 12 \text{ V wkg.}$

1) High stability.

2) Other values may be used for higher output voltages.

larger input signal, e.g. with tone correction network, the output voltage can be increased simply by modifying the coupling between the two sections of the ECC 83. By connecting the capacitor C_{10} direct to the first section an

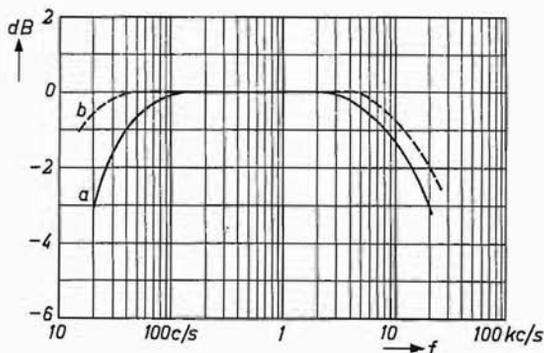


Fig. 76. Response curve of the mixing amplifier: (a) microphone input, (b) radio and pickup input.

output voltage of 800 mV can be made available. Intermediate values for the output voltage can be obtained by varying the values of R_{16} and R_{18} ; if both of these are made 47 k Ω , for example, the output voltage will be 400 mV.

Adjustment of the output voltage by means of a potentiometer across the output stage would neutralise the advantage of the low input impedance of the cathode-follower.

Response curve

Measured between the microphone input socket and the output the response curve of this mixing amplifier is as shown in Fig. 76; with respect to the relative 1 kc/s level it is flat within 3 dB from 20 c/s to 20 kc/s. The response curves measured from the radio and pickup sockets are also given in this diagram. As the EF 86 stages are not included in these latter measurements the curve remains flat, within 2 dB, further into the lower frequency region, down to 15 c/s.

Hum and noise

Hum and noise tests were carried out with $100\text{ k}\Omega$ resistors across the microphone and pickup input sockets and with the potentiometers set to maximum to simulate average practical conditions. The output signal from the mixing amplifier was applied to a 10 W power amplifier for measurement of the hum and noise, with $7\ \Omega$ across the output, and the voltage across this resistor was 26 mV.

Nominal output power of the amplifier (10 W) corresponds to 8.4 V across this resistor, so the hum and noise was — 50 dB for 10 W. The background noise of the 10 W amplifier alone is better than — 70 dB for 10 W, so the — 50 dB may be attributed to the mixing amplifier itself.

