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AUDIO AMPLIFIERS WITH SINGLE-ENDED PUSH-PULL OUTPUT

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For some time now radio receivers have been available in which the loudspeaker is directly driven by the output valves, without the use of an output transformer. This means the omission of a component that, apart from adding to the losses and distortion, is quite expensive even to satisfy only moderate demands. The phase shift it introduces, moreover, prevents feedback being applied to its full advantage. This article deals with a new transformerless circuit for the last two stages of audio-frequency amplifiers which differs appreciably from the conventional one and can satisfy far higher demands. No expensive components are required to replace the transformer.

Music reproduction in the home of gramophone records, tape recordings or F.M. transmissions, may nowadays reach a very high quality indeed. If the excellent properties of such "music sources" are to be done full justice, the reproduction installation must come up to very high standards. Particularly in the last few years, therefore, there has been a general trend towards improvement of the reproduction system. This may consist of one or more loudspeakers and an audio-frequency amplifier, connected to a pick-up, the play-back head of a tape recorder, or to the R.F. and I.F. stages of an F.M. receiver.

Confining ourselves to the audio amplifier and the loudspeaker alone, there are three aspects to be considered, viz. the electronic circuitry, the loudspeaker design, and the acoustics.

The last two aspects were dealt with at some length in a recent article in this Review on a high-fidelity loudspeaker installation [1]). The present article is mainly concerned with the electronic aspects, i.e. with circuit development. As we shall see presently, this work has been influenced by certain innovations in the loudspeaker field.

The demands made on audio-frequency amplifiers involve the output power, frequency response, non-linear distortion and output resistance.

As regards frequency-response characteristics, it is not sufficient if merely the *amplitude* characteristic is flat within the audible range (20-18 000 c/s), since then the *phase* characteristic would possess the appropriate form only within a far narrower range; by the "appropriate" form is meant that the phase angle is proportional to the frequency (the proportionality factor may also be zero). With a poor phase characteristic a sudden burst of music, sharp consonants, etc. cannot be reproduced without distortion [2]). A good phase characteristic can be obtained if the amplitude characteristic remains flat far beyond the audible range, e.g. up to 40 kc/s. However, it is usually considered desirable to be able to vary the amplitude characteristic at either end of the audio band, i.e. have a tone control — preferably independent — of both the high and low notes.

Non-linear distortion in an amplifier is due to the non-linear characteristics of valves and magnetic materials. It gives rise to harmonics, which alter the timbre of the sound, and, worse, when two or more tones are produced simultaneously, to non-harmonic overtones (intermodulation). Non-linear distortion, rapidly increasing with amplitude above a certain signal strength, puts a definite limit to the maximum output power at which the amplifier can be satisfactorily operated.

[1]) G. J. Bleeksma and J. J. Schurink, A loudspeaker installation for high-fidelity reproduction in the home, Philips tech. Rev. **18**, 304-315, 1956/57 (No. 10).

[2]) See e.g. J. Haantjes, Judging an amplifier by means of the transient characteristic, Philips tech. Rev. **6**, 193-201, 1941.

Finally, the output resistance of the amplifier should be low, in order to provide a substantial damping effect on the loudspeaker, the latter having a mechanical resonance in the region of the very low notes [3]).

The designer of the amplifier has to try to satisfy these demands on characteristics, distortion and output resistance, whilst keeping the costs as low as possible. Considerations of price may carry considerable weight where mass production, e.g. of radio sets, is concerned. A low cost price is closely related to a high efficiency, for a poor efficiency involves a larger power pack and unnecessarily high-power output valves.

As output valve, the pentode is generally preferred to the triode, the former having a higher efficiency and requiring a smaller input signal. With a pentode, maximum efficiency (theoretically 50%) is obtained if the load resistance R_0 equals the D.C. resistance R_{ak} of the valve ($R_{ak} = V_{a(w)}/I_{a(w)}$, $V_{a(w)}$ and $I_{a(w)}$ being the anode voltage and the anode current at the working point; R_{ak} should not be confused with the (much higher) internal resistance $R_i = \partial V_a/\partial I_a$ at constant control-grid voltage). For most output pentodes R_{ak} has a value of several thousands of ohms.

A moving-coil loudspeaker of conventional design has a speech-coil of relatively few turns of fairly thick wire. Its impedance (virtually a resistance) is accordingly low, e.g. 7 Ω. This low impedance is matched to the far higher value of R_{ak} via a step-down output transformer. This component requires a great deal of care and material in its construction. Its stray self-inductance forms together with stray capacitances one or more oscillatory circuits, which impose an upper limit to the frequency range that can be reproduced. This cut-off frequency is the higher as the stray field and the winding capacitance are smaller, and therefore as the number of turns is smaller. One essential requirement for a good reproduction of the *low* notes, however, is a large primary self-inductance, which means a large number of turns. Another unfavourable factor is the fact that the anode direct current pre-magnetizes the core of the transformer.

The latter difficulty is avoided if a push-pull output stage is employed; D.C. magnetization of the core is then absent and a more favourable compromise between large primary self-inductance and small stray field and capacitance can be reached. A push-pull arrangement has furthermore the well-known advantage of producing less distortion: if the circuit is symmetrical, no even harmonics occur in the output signal. The complications incurred by using a second output valve (with the necessary phase-inverting element for getting two equal driving voltages in phase opposition), make the circuit more expensive, but for better-quality radio receivers and amplifiers the above advantages of push-pull operation outweigh the difference in price.

Apart from limiting the frequency range at both ends, the output transformer has the drawback of causing a loss of output power. Its efficiency is often as low as 50%, and seldom better than 80%, dependent upon the amount of material (and hence the cost) spent on it. The same applies to the distortion it introduces as a result of the non-linear B-H-curve of the iron core; here too, improvement can be reached only at the expenditure of more material.

Most serious of all, perhaps, is the fact that the output transformer prevents *negative feedback* being applied to its full advantage. With the aid of negative feedback, as is well known, the amplitude characteristic can be improved, the non-linear distortion reduced, and the output resistance lowered. If the circuit contains a phase-shifting element, however, such as an output transformer, which may act either inductively or capacitatively, there is a risk that for a given frequency the negative feedback aimed at becomes a positive feedback. To prevent instability, the feedback thus cannot be made as strong as might be desired.

As a consequence of all the difficulties mentioned above one is compelled to use a heavy and large (and hence expensive) output transformer, unless lower demands are made.

Some years ago Philips started working in another direction, viz. avoiding an output transformer altogether. If this attempt were to succeed, it would be possible, not only to eliminate all these drawbacks, but also to get rid of an expensive, bulky and heavy component. That these attempts have, in fact, been successful will appear from the following.

High resistance loudspeakers

Loudspeaker speech-coils having few turns of fairly thick wire have the advantage of being quickly wound, with little risk of breaking the wire, and also that the insulation of the wire takes up relatively little winding space (high filling factor).

The first investigations were aimed at establishing the largest number of turns and the smallest wire diameter that could be accommodated in the available space without unduly complicating manufacture or making the filling factor too small. Supported by their experience with all kinds of miniature

[3]) See the article quoted in [1]), page 314.

winding jobs, the Eindhoven loudspeaker factory succeeded in 1953 in mass-producing 4000 Ω coils using 40 μ copper wire. These coils were provided with a central tapping, so that they could be directly incorporated in a push-pull circuit (*fig. 1*). The

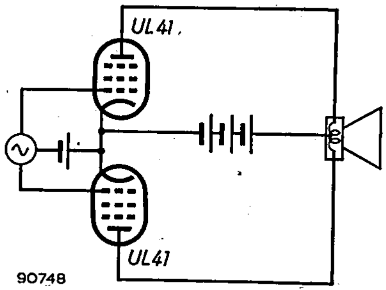

Fig. 1. Output stage with two pentodes UL 41 and a split-coil loudspeaker (type 9754) without output transformer. The resistance of the speech coil is 2×2000 ohms.

resistance of 2000 Ω for each half coil is fairly low for ordinary output valves, but is a practicable value for the UL 41 valve with its rather low D.C. resistance [4]). Both direct tests and life tests proved satisfactory, and part of the production of the radio receiver BX 521 U was equipped with such a loudspeaker, so that practical experience on a large scale could be gained. As regards the loudspeakers, no complaints were received. A remarkable feature of this set was the improved reproduction of the low notes.

This solution, however, has its drawback. The output valves are operated as a Class A amplifier, i.e. with a large anode direct current. This current flows through the high-resistance loudspeaker coil, causing a considerable voltage drop (far greater than in an output transformer), and a considerable development of heat. In a new circuit, described below, these disadvantages have been completely eliminated.

The single-ended push-pull circuit

Fig. 2a shows a normal push-pull circuit. Each of the valves has been biassed to give a quiescent anode current $I_{a(w)}$. The supply, at a voltage V_b, therefore delivers a power $2V_bI_{a(w)}$. Fig. 2*b* shows a somewhat modified circuit, in which each valve again has the supply voltage V_b and the current $I_{a(w)}$, the total D.C. power being likewise $2V_bI_{a(w)}$. If the load resistance R_0 is the same in either case, the efficiency is also the same and the circuit of fig. 2*b* can supply the same A.C. power as that of fig. 2*a*.

This remains valid if PP' and QQ' in fig. 2*b* are interconnected. The two resistances R_0 are then connected in parallel and are equivalent to a single resistance $\frac{1}{2}R_0$ (fig. 2*c*), i.e. $\frac{1}{4}$ of the total resistance $2R_0$ of the circuit in fig. 2*a*. If the two valves are equally biassed, direct current flows only through the circuit formed by the supply source (voltage $2V_b$) and the two series-connected valves, none passing through the resistance $\frac{1}{2}R_0$.

This type of push-pull circuit was already well known in 1951 [5]). Clearly, it is ideally suited to the system of direct power transfer: it requires a 4 times lower load impedance than the conventional push-pull circuit, whilst owing to the fact that the load is free of direct current, all drawbacks such as large

[4]) Valves of the U-series are designed for universal A.C./D.C. receivers without mains transformer. They therefore have a low anode voltage (100-170 V), so that output valves of the U-series have to be designed for handling a large current. Their D.C. resistance R_{ak} is accordingly low.

[5]) A. Peterson and D. B. Sinclair, A single-ended push-pull audio amplifier, Proc. Inst. Rad. Engrs. **40**, 7-11, Jan. 1952.

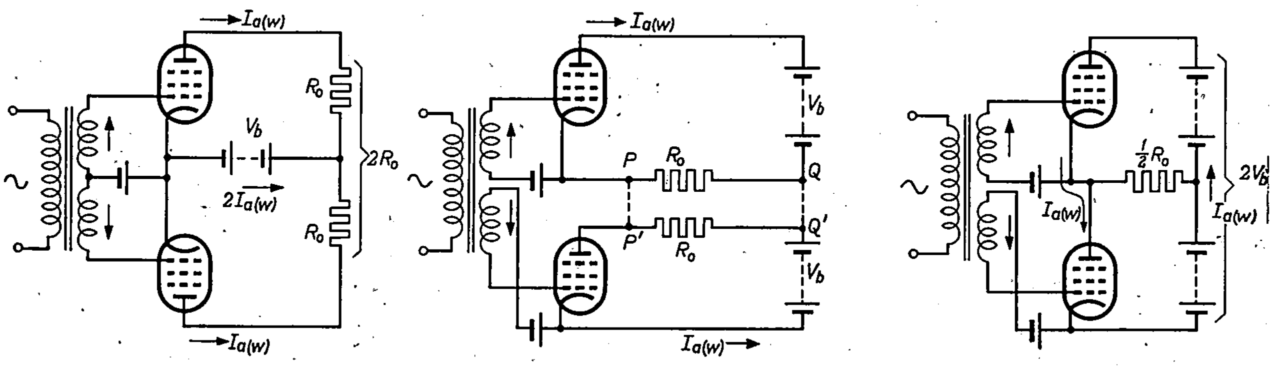

Fig. 2. *a*) Conventional push-pull circuit. *b*) Circuit which is electrically equivalent to (*a*). By interconnecting *P* and *P'* and *Q* and *Q'*, the single-ended push-pull circuit (*c*) is obtained, which matches to a 4 times lower load resistance ($\frac{1}{2}R_0$ instead of $2R_0$), whilst no D.C. flows through the load.

voltage losses and heat development in the loudspeaker coil are eliminated. The latter, moreover, does not require a central tapping, which simplifies manufacture. The central tapping of the supply source can likewise be eliminated, e.g. by connecting the loudspeaker L in series with a capacitor C_0 of sufficiently large capacitance according to *fig. 3*. Another method will be discussed presently.

The above principle has been applied in an increasing number of Philips receivers during the last few years. A new output pentode (type EL 86) has been developed for it, which can operate at low anode voltage and large anode current [6]); at maximum anode dissipation, the D.C. resistance R_{ak} amounts to only 1600 Ω. For the two valves, this requires matching by a loudspeaker impedance (mainly resistive) of 800 Ω, which presents no difficulties in manufacture. A single-ended push-pull

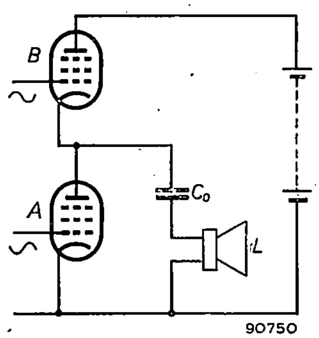

Fig. 3. Single-ended push-pull circuit with two pentodes A, B (EL 86) requiring no central tapping of the H.T. supply. A large capacitor C_0 is placed in series with the speech coil of loudspeaker L.

circuit with two EL 86 valves can produce a maximum A.C. power of 10 W; for 10-20 W two EL 86 valves may be connected in parallel on either side, the required load resistance then being 400 Ω, which may be realized by using one 400 Ω loudspeaker or two 800 Ω loudspeakers in parallel. Nearly all Philips loudspeakers with a cone diameter of 12.5 cm and larger are now available in 400 or 800 Ω versions [7]).

The relations between load resistance R_0, output power P_0 and the position of the operating point $V_{a(w)}$, $I_{a(w)}$ can be determined from the following simplified considerations. *Fig. 4* again shows the circuit of fig. 3 but with voltage and current indications added. The load resistance is here denoted R_0; the reactance of C_0 will be neglected. *Fig. 5* shows two idealized I_a-V_a characteristics of the pentodes employed. One characteristic is valid for a control-grid voltage $V_{g1} = 0$ and the other for the value of V_{g1} where $I_{a(w)}$ is half as large. XY is the load line intersecting the knee of the upper characteristic and subtending an angle α with the V_a axis such that $\cot\alpha = 2R_0$, this being the most favourable load line (i.e. a load of $2R_0$ for each valve; being in parallel, this gives a net load of

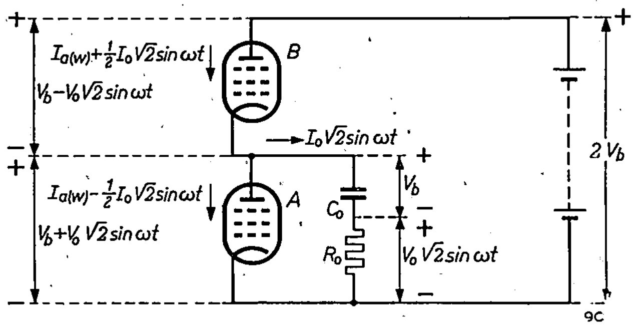

Fig. 4. Circuit of fig. 3 showing currents and voltages. A load resistance R_0 represents the loudspeaker.

R_0 as in fig. 4). It intersects the V_a axis at the point Y, corresponding to the voltage $2V_b$ of the supply source. The working point is W. We see that the relation $\cot\alpha = (2V_b - V_{a(w)})/I_{a(w)}$ is also valid; by neglecting the "knee voltage" $V_{a_{min}}$ below which V_a must not drop, we have $2V_b = 2V_{a(w)}$ and therefore $\cot\alpha = V_{a(w)}/I_{a(w)}$. By definition, $\cot\alpha = 2R_0$ and $V_{a(w)}/I_{a(w)} = R_{ak}$, so that the matching condition $2R_0 = R_{ak}$ is satisfied.

The r.m.s. values of the alternating current I_0 and the alternating voltage V_0 are given by:

$$I_0 = \sqrt{P_0/R_0}, \quad V_0 = \sqrt{P_0 R_0}. \quad \quad (1)$$

Either valve contributes half the amount of I_0. The valve currents are $I_{a(w)} \pm \frac{1}{2}I_0\sqrt{2}\sin\omega t$ (see fig. 4). Under conditions of full drive (maximum signal), the entire load line XY is traversed, I_a swinging from zero to $2I_{a(w)}$. Therefore

$$I_{a(w)} = \tfrac{1}{2}I_0\sqrt{2}. \quad \quad (2)$$

V_a swings from $V_{a_{min}}$ (approx. zero) to $2V_b$, so that

$$V_{a(w)} = V_0\sqrt{2}. \quad \quad (3)$$

For the voltage $2V_b$ of the H.T. supply, we thus have

$$2V_b = 2V_{a(w)} = 2V_0\sqrt{2}. \quad \quad (4)$$

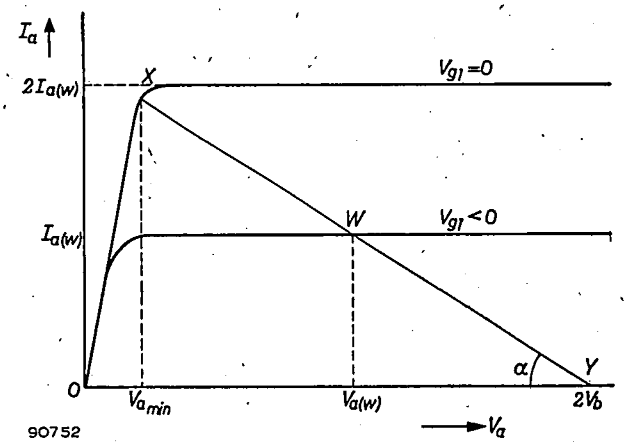

Fig. 5. Idealized pentode characteristics and load line XY. The working point is W.

[6]) As was recently pointed out in this Review (K. Rodenhuis *et al.*, Philips tech. Rev. **18**, 185, 1956/57, No. 7), such conditions are more favourable to the working life of the valve than when the same dissipation is obtained at a higher voltage and a smaller current.

[7]) 800 Ω speakers are designated by the suffix A, 400 Ω speakers by the suffix B to the type number, e.g. 9710 A, 9766 BM; the suffix M indicating a double-cone loudspeaker (see J. J. Schurink, Philips tech. Rev. **16**, 241-249, 1954/55, or the article quoted by [1]), pp. 305-306).

The power P_b actually delivered by the source is

$$P_b = 2V_bI_{a(w)} = 2V_0\sqrt{2} \times \tfrac{1}{2}I_0\sqrt{2} = 2V_0I_0 = 2P_0,$$

which demonstrates that the efficiency is 50%.

In the *absence* of a signal the supply delivers the same power $P_b = 2P_0$, which is now entirely dissipated in the valves, which means for each valve a dissipated power $P_0 = V_bI_{a(w)}$ (twice as much as under conditions of full drive). This means that the maximum A.C. power the circuit is permitted to deliver, is equal to the maximum anode dissipation in the absence of a signal. This quantity is a parameter of the valve, and amounts to 12 W for the EL 86. Owing to the simplifications introduced here, P_0 is actually somewhat lower, viz. 10 W.

For $P_0 = 10$ W and $R_0 = 800\ \Omega$, we arrive, according to (1), at $I_0 = 112$ mA and $V_0 = 90$ V. The working point of the valve now follows from (2) and (3):

$$I_{a(w)} = 79 \text{ mA}, \quad V_{a(w)} = 127 \text{ V}.$$

According to (4), the supply voltage must amount to $2V_b = 254$ V, to which must be added the amount $2V_{a_{\min}}$ (and, if the negative grid bias of valve A in fig. 4 is produced across a cathode resistor, an additional $|V_{g1}|$).

For maximum output powers lower than 10 W the valve may be biassed to give a smaller quiescent anode current at a proportionally lower supply voltage, the working point being evaluated in the above manner.

The signal to be amplified generally consists of a single alternating voltage with respect to earth. For driving the valves of a push-pull circuit in opposite phase, however, two control voltages are required. This could be obtained by means of a transformer with two secondary windings as shown in fig. 2*a* for the ordinary push-pull circuit and in fig. 2*c* for the single-ended push-pull circuit. For some time however, a valve (phase inverter) has been preferred to obtain the opposed control voltages for a conventional push-pull circuit. A valve can also be used for this purpose with a single-ended push-pull circuit, as is shown in *fig. 6*. Section I of a double-triode ECC 83 functions here as amplifier and is connected to the control grid of the output valve A (for the sake of clarity, coupling capacitors are not shown in the diagram) and also to the grid of section II. At the anode of II a signal in opposite phase is produced, which, by an appropriate choice of resistances, is given an amplitude suitable for driving output valve B.

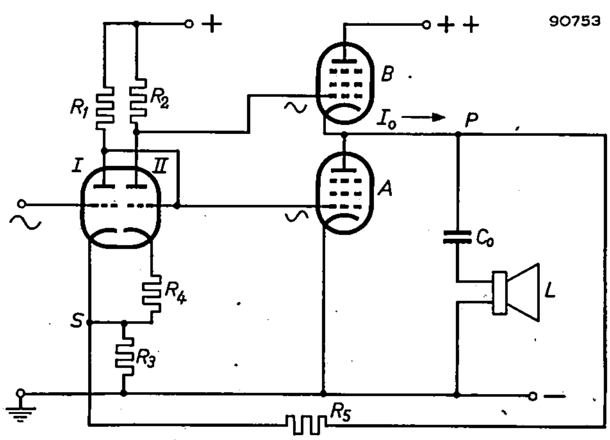

Fig. 6. *I-II* double-triode ECC 83. *A*, *B* pentodes EL 86 in single-ended push-pull arrangement. The first stage of the amplifier, consisting of triode *I*, drives the output valve *A* and the phase inverter *II* producing the control voltage for *B*. (Components not essential to the principle, such as coupling capacitors, are not shown.) Positive feedback in the first-amplifier - phase-inverter stage is effected via the common cathode resistor R_3 of *I* and *II*. Negative feedback, from *P* to *S*, is obtained via R_5.

Feedback

As mentioned earlier, one of the advantages of omitting an output transformer is the greater freedom in applying negative feedback. Without running the risk of instability, certain combinations of negative and positive feedback can even be made, by which, as we shall demonstrate presently, distortion can be drastically reduced. In the circuit of fig. 6 [8]) negative feedback is realized by passing part of the output current I_0 (being proportional to the output voltage), via resistor R_5, through cathode resistor R_3; at the same time there is positive feedback in the phase inverter stage in the form of valves I and II having resistor R_3 as a common cathode resistance.

The effect of combined positive and negative feedback is clearly demonstrated by a hypothetical example of greater simplicity than that of fig. 6. A_1 and A_2 in *fig. 7* represent two amplifiers in cascade, e.g. a pre-amplifying stage and an output stage, whose amplifications will likewise be called A_1 and A_2 respectively. A portion B_1V_1 of the output signal V_1 of A_1 is fed back to the input of A_1, and a portion B_2V_0 of the output signal V_0 of A_2 is likewise fed back to the input of A_1. If V_i is the signal to be amplified, we may write, quite generally,

$$V_1 = A_1(V_i + B_1V_1 + B_2V_0)$$

and

$$V_0 = A_2V_1.$$

The overall amplification A is accordingly:

$$A = \frac{V_0}{V_i} = \frac{A_1A_2}{1 - A_1B_1 - A_1A_2B_2} = \frac{A_1A_2}{N},$$

where

$$N = 1 - A_1B_1 - A_1A_2B_2.$$

Similarly, the total distortion D is given by

$$D = \frac{1}{N}D_1 + \frac{1 - A_1B_1}{N}D_2 + \frac{1 - A_1B_1}{N}D_1D_2,$$

[8]) Designed by E. H. Nielsen, formerly a member of the laboratory of the Philips factory in Copenhagen.

where D_1 and D_2 are respectively the distortion of either amplifier.

It may be seen that if A_1B_1 is made unity, a very special situation arises: D is reduced to D_1/N, i.e. the distortion of amplifier A_2 does not contribute at all to the total distortion. The remaining distortion, D_1/N, is determined by the (slight) distortion D_1 of the first amplifier and by the quantity N, the latter becoming $-A_1A_2B_2$ for $A_1B_1 = 1$. The absolute value of $A_1A_2B_2$ may be considerable larger than unity, in which case the total distortion D will even be appreciably smaller than D_1. Also, the absolute value of the total amplification A becomes equal to $1/B_2$ (and therefore independent of A_2).

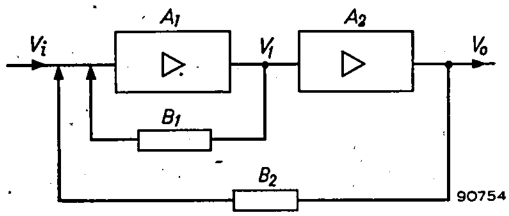

Fig. 7. Cascade connection of two amplifiers A_1 and A_2, the former having a feedback loop via B_1, and the combination a feedback via B_2.

The condition $A_1B_1 = 1$ can be fulfilled for a wide frequency range if A_1 and B_1 are real in that range, i.e. if the amplifier A_1 and the feedback circuit B_1 contain no phase-shifting elements. In the first amplifying stage this condition may be closely approached without any difficulty. This condition being fulfilled means that the amplifier A_1 is given such a positive feedback B_1 that it is on the verge of oscillating. This does not necessarily mean that the combination A_1-B_1-A_2-B_2 (fig. 7) is unstable; if the second amplifier and the feedback circuit are free of any elements causing adverse phase shifts, the whole circuit can be kept stable by applying a certain negative feedback B_2. An output transformer is an element which inevitably causes such phase shifts that the stability would be seriously jeopardized [9]). Only by its elimination is it possible to realize circuits of the type considered here.

Practical examples

A.F. amplifier for a radio receiver

Fig. 8 shows a further developed version of the circuit of fig. 6, as it might be used for a radio receiver. Here the feedback resistor R_5 of fig. 6 has been replaced by two networks, C_1-R_6-R_7 and C_2-R_8-R_9. These incorporate the tone controls: R_6 for the low and R_9 for the high notes.

[9]) The output transformer can be kept outside the feedback loop B_2 (see e.g. fig. 6 of the article quoted by [5])), but the distortion introduced by it remains undiminished in the output signal; also the degree of negative feedback is necessarily limited owing to the phase shift between primary voltage and primary current.

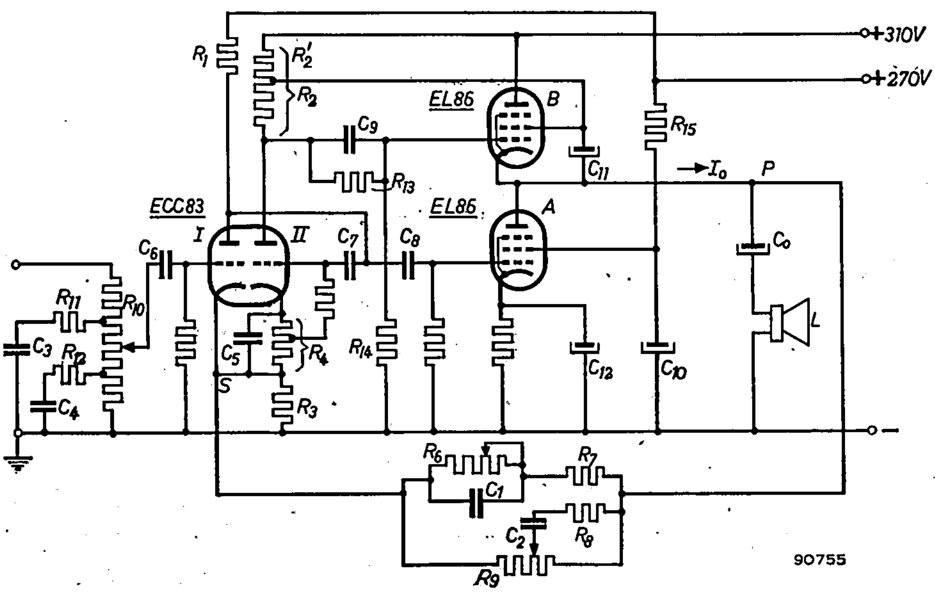

Fig. 8. More detailed version of the schematic diagram of fig. 6. Instead of the feedback resistor R_5, the networks C_1-R_6-R_7 and C_2-R_8-R_9 have been incorporated; R_6 low-note control, R_9 high-note control. R_{10} volume control with branches R_{11}-C_3 and R_{12}-C_4 for low-note compensation at low levels of volume. C_5 reduces the negative feedback through R_4 for the high audio frequencies. C_6, C_7, C_8, C_9 coupling capacitors. The voltage divider R_2-R_{13}-R_{14} provides the correct bias for the control grid of output valve B. The screen grid of the output valve A is fed via resistor R_{15} and the screen grid of B from a tapping on R_2 (for other methods of screen-grid supply, see figs. 11 and 12).

Low notes. Capacitor C_1 reduces the negative feedback at low frequencies, so that the latter are relatively more amplified. For a radio set this is a desirable feature, e.g. for compensating the drop in loudspeaker radiation at frequencies below those at which the cabinet forms a sufficiently large baffle. With a cabinet of average size, the drop in loudspeaker response at descending frequency (6 dB per octave) begins at about 1000 c/s. As desired, the effect of C_1 can be decreased (increasing the negative feedback and thus increasing low-note attenuation) by lowering the resistance of R_6. The non-variable resistor R_7 prevents the negative feedback becoming too strong.

has been compensated by placing a capacitor C_5 across cathode resistor R_4 of the phase inverter. The effect of this is a reduction of the negative feedback in the phase-inverter stage, thus increasing the gain, particularly at the higher frequencies.

Figs. 9 and *10* give some idea of the results attained with an amplifier of the above type. The diagrams of fig. 9 show the distortion as a function of the output power for the frequencies 90, 1000 and 8000 c/s, whilst fig. 10 shows the amplitude-frequency characteristic with both tone controls set to maximum. It can been seen that the low and high notes are boosted with respect to the medium range, as is desirable for a radio receiver.

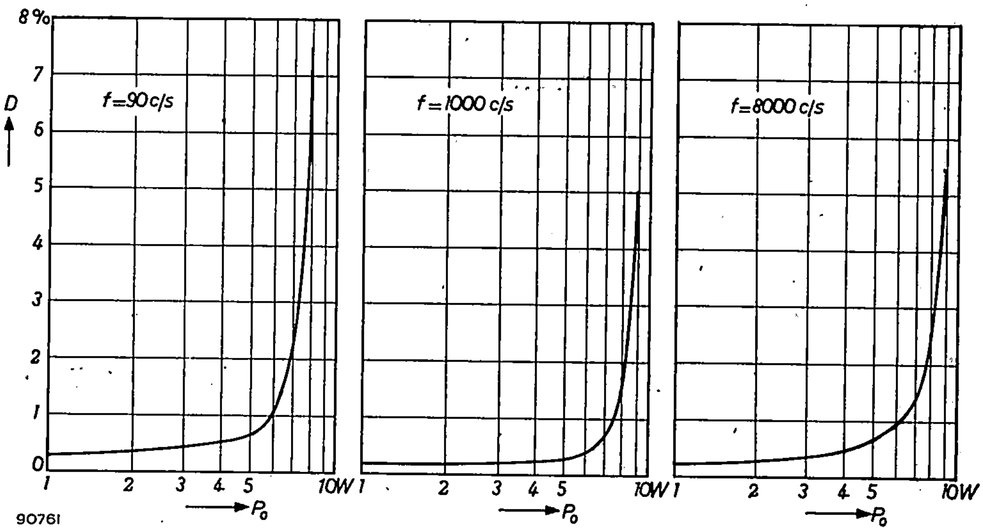

Fig. 9. Distortion D, measured on an amplifier of the type shown in fig. 8, plotted against the output power P_0, at the frequencies 90, 1000 and 8000 c/s. Both tone controls set to maximum.

High notes. The values of C_2 and R_8 are so chosen that the higher audio frequencies are mainly fed back through this branch. If the sliding contact on R_9 in fig. 8 is moved to the right, a larger part of R_9 is connected in series with this branch, the negative feedback diminishes and the high notes become accordingly stronger.

As is customary in many Philips receivers, automatic low-note compensation is provided at low levels of volume: volume control R_{10} is provided with the branches R_{11}-C_3 and R_{12}-C_4, which ensure that with a reduction of the total volume the low notes are relatively less attenuated, thus compensating for the property of the human ear of becoming insensitive to a steadily increasing portion of the low-note range when the volume is reduced.

Near the upper limit of the audio spectrum the amplification is reduced as a result of the anode capacitance of the output valves and of the Miller effect (reaction of anode upon control grid). This

As a measure of the sensitivity of an amplifier, it is customary to give the input r.m.s. voltage required to produce an output power of 50 mW at 1000 c/s. For this particular example this voltage amounts to 24 mV, which is normal for the audio section of a radio receiver.

Negative feedback reduces the internal resistance of the output stage to about 20 Ω, a value that is

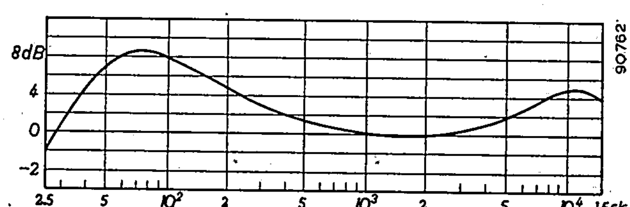

Fig. 10. Output voltage with constant input voltage versus frequency f, measured on an amplifier of the type shown in fig. 8, loaded with 800 ohms. Both tone controls set to maximum (i.e. minimum negative feedback). Boosting of the low and high notes as shown is desirable in the audio amplifier of a radio receiver.

low compared with the load resistance (800 Ω). This is conducive to the uniform reproduction of the very low notes (damping of the loudspeaker resonance).

With a single-ended push-pull circuit a question requiring attention is the supply for the screen grids of the output valves. The screen grid voltage should be at about the same D.C. level as the anode, but there must be no alternating voltage between screen grid and cathode.

Let us first investigate how this ideal condition can be approached with the output valve A (fig. 8). Here the screen grid is supplied via resistor R_{15}, whilst the electrolytic capacitor C_{10} (in series with C_{12}) constitutes a low impedance between screen grid and cathode. The screen-grid current (about 5 mA) forms only a minor portion of the average cathode current. This portion is not constant however: it varies with the amplitude of the signal on the control grid. The average value of the screen-grid current therefore varies with the signal strength, so that the presence of the series resistor R_{15} makes the screen-grid voltage vary with signal strength: the stronger the signal, the lower the screen-grid voltage. This effect reduces the maximum output power of the valve. To minimize the effect, the resistance R_{15} should be made as small as possible and the screen grid must therefore be fed from a source whose voltage exceeds the required value by as little as possible. For this reason the screen grid in fig. 8 is not fed from the 310 V source, but from the 270 V source (which also supplies the valve-half I and the R.F. part of the receiver). A supply direct from a voltage of approx. 155 V would be even better, and this can be easily realized if the 310 V supply is derived from a full-wave bridge circuit rectifier (*fig. 11*): if the secondary of the transformer is provided with a central tapping, half the direct voltage can be tapped off from this point.

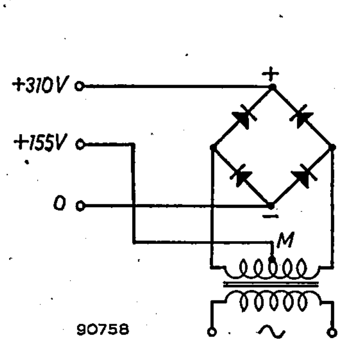

Fig. 11. Full-wave bridge circuit rectifier. Half the direct voltage can be obtained from the centre-tapping M on the secondary of the transformer. This can be used for feeding the screen grid of valve A in fig. 8.

As for the output valve B (fig. 8), its anode has a fixed potential, whilst the output signal voltage is taken off at the cathode. If the screen grid were connected directly to the anode, then the loudspeaker would be nearly short-circuited (for signal currents) via capacitor C_{11} and the H.T. supply. A certain impedance between screen grid and anode must therefore always be provided (unless a separate power supply is to be provided for this screen grid). This impedance should be high with respect to the (parallel) loudspeaker impedance, but, on the other hand, it should not have a high D.C. resistance, as otherwise the screen-grid voltage would be too much below the anode voltage and also would vary too much with the signal strength (see above). In some respects a choke has some advantages over a resistor, but this would again introduce a phase-shifting element. In fig. 8 a compromise has been struck between a not too high and not too low resistance by connecting the screen grid to a suitable tapping on the anode resistor R_2 of the phase inverter.

A more elegant solution is shown in *fig. 12*. Here the loudspeaker itself is connected between screen grid and anode. The output signal current I_0 passes

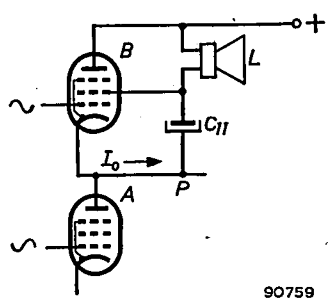

Fig. 12. Supply to the screen grid of valve B in fig. 8 via the loudspeaker L. The output current I_0 (less the negative-feedback current) flows through C_{11} and L.

without losses (with the exception of the part used for feedback) via C_{11} through the loudspeaker. The function of capacitor C_0 is now taken over by C_{11} and it can therefore be omitted. The fact that the (small) screen-grid current now flows through the loudspeaker forms no objection.

Further improved circuit

The reader may have noted that in fig. 8, from which the output transformer with its phase-shifting effect has been eliminated, other phase-shifting elements have after all been included, viz. the tone control networks in the feedback circuit. In principle such a solution is indeed not ideal. That it is nevertheless used is justified by the fact that otherwise an additional amplifying stage would be required (the additional amplification of low and high notes is effected here by reducing the negative feedback), and by the fact that even with the smallest amount of negative feedback, distortion remains appreciably smaller than in a conventional circuit with the average output transformer.

If near-perfection is aimed at, then the negative feedback loop should be kept free from any phase shift by effecting the tone control in an additional stage preceding valve *I*. For an amplifier of this type the following distortion figures were measured:

Output power (in W)	10.4	9.25	4	1
Distortion (in %)	0.30	0.11	0.05	0.03

In order to obtain for this measurement an input signal itself sufficiently free of distortion, a filter for suppressing harmonics should be incorporated between the audio generator and the amplifier. Any harmonics in the output signal of the amplifier are measured with a "wave analyzer", by means of which each harmonic can be separately determined.

Fig. 13 shows the amplitude and phase characteristics of this amplifier. We see that the point where the amplification has dropped by 3 dB lies well into the ultrasonic range, namely at 250 kc/s, whilst the phase shift in the audible range is restricted to a few degrees only. For an output power of 10 W the efficiency of the output stage is a good 42%, which is very satisfactory. The only critical resistance value is that of the cathode resistor of the phase inverter (R_4), which must be accurate within 2%; for all other resistors a 10% tolerance is adequate.

We shall confine ourselves to the above two examples. It may be mentioned, however, that in many other cases the greater freedom in the application of negative feedback which single-ended push-pull makes possible will be welcome: it permits the realization of various types of special circuits for which the output transformer has hitherto formed the great stumbling block.

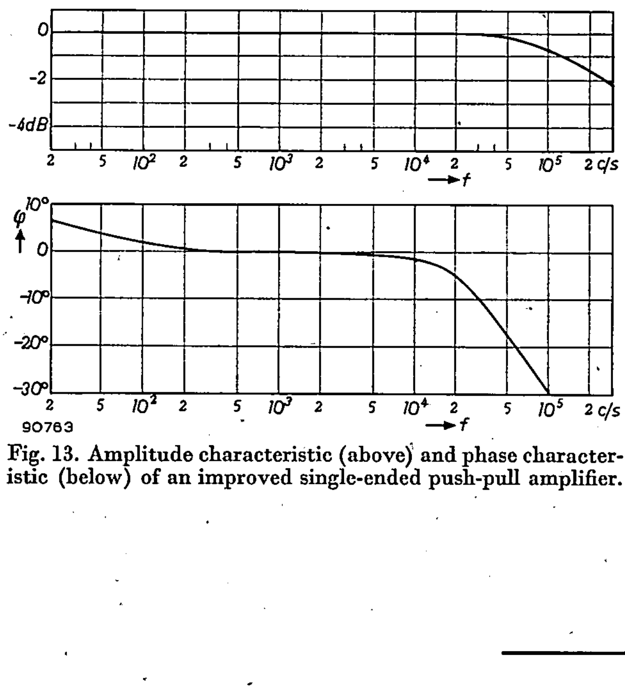

Fig. 13. Amplitude characteristic (above) and phase characteristic (below) of an improved single-ended push-pull amplifier.

Summary. The low resistance of the speech coils of conventional moving-coil loudspeakers requires a transformer for matching to the far higher D.C. resistance of the output pentode. This output transformer has many disadvantages: it limits the reproduced audio spectrum on either end, causes power losses, introduces distortion, and by the phase-shift it introduces, prevents negative feedback being employed to its full advantage. Any transformer in which these drawbacks are confined within acceptable limits is necessarily a heavy, bulky and expensive component.

In 1953 Philips manufactured loudspeakers with a speech coil of 2× 2000 ohms, which could be incorporated in a normal push-pull circuit with two pentodes UL 41. For such a "direct power transfer", however, the "single-ended push-pull circuit" is more suitable: here the load (loudspeaker) can have a resistance 4 times as low as with the conventional push-pull circuit, whilst no direct current flows through it. This led to the development of the output pentode EL 86, with a D.C. resistance of 1600 ohms at full load, and of associated loudspeakers of 800 and 400 ohms. Two valves EL 86 in single-ended push-pull can deliver 10 W.

The absence of an output transformer makes it possible to realize circuits in which the first amplifying stage has positive feedback up to the verge of oscillation, whilst the first amplifying stage and the output stage together have negative feedback. Such a circuit is perfectly stable. The total distortion can be reduced to only a small fraction of the distortion of the first amplifying stage alone and made independent of the distortion in the output stage. Two practical examples are discussed, one with the tone controls incorporated in the negative feedback circuit and one with the tone controls in a separate preceding stage. In the latter case the distortion at 10 W is only 0.30%; the amplitude characteristic remains straight well into the ultrasonic range before showing a 3 dB drop at 250 kc/s, so that the phase characteristic does not show deviations greater than a few degrees within the audible range.