DIFFERENCE AMPLIFIERS WITH A REJECTION FACTOR GREATER THAN ONE MILLION

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In the various Philips laboratories there is a constant demand for special electronic measuring instruments capable of meeting widely diverse requirements. The above authors are members of a research group whose task it is to meet this demand. Among the numerous circuits which they have developed in the course of the years, there are several that give surprising results with simple means. A number of these circuits will be dealt with in a series of articles in this journal, the first of which follows below. It is a sequel to the articles on difference amplifiers that appeared in volumes **22** *and* **23**, *and shows the way in which the exceptionally high rejection factor required in a special case was obtained.*

The rejection factor of a difference amplifier

In electrical measuring techniques a frequently encountered problem is the measurement of a voltage between two points both of which have potentials with respect to earth which are much larger than the voltage to be measured. This problem is not so easy to solve if the voltage under measurement (the input signal) is so small that it has to be amplified. One of the difficulties is then that the voltage between the output terminals of the amplifier is in general not only a function of the input signal but also a function of the common voltage on the input terminals with respect to earth. This difficulty is overcome by the use of special amplifiers. These *difference amplifiers*, as they are called, have been the subject of several articles in this journal [1][2].

The nomenclature used in those articles will also be used here; see *fig. 1*. The input signal is $2E_{\mathrm{it}}$. The voltage $2E_{\mathrm{ut}}$ between the output terminals is that to which the meter responds. E_{if} and E_{uf} are the average voltages with respect to earth on the input and output terminals, respectively. E_{if} and E_{uf} will be called the in-phase signals, E_{it} and E_{ut} the anti-phase signals. The letter E may denote either a DC or an AC voltage. The frequencies of E_{it} and E_{if} need not be identical.

In general we can write:

$$E_{\mathrm{ut}} = AE_{\mathrm{it}} + BE_{\mathrm{if}}. \quad \ldots \quad (1)$$

In this expression A is the amplification which E_{it} undergoes, and the term BE_{if} represents the interfering effect due to E_{if}. In a good difference amplifier B is much smaller than A. The ratio $H = A/B$ is called the *rejection factor* and is one of the most important characteristics of a difference amplifier.

Article I, mentioned under reference [2], gives examples of difference amplifiers where the *guaranteed* [3] value H_{min} of the rejection factor is approximately 2×10^4. This applies to balanced amplifiers in which special measures have been taken to obtain a high degree of symmetry (given perfect symmetry the factor B in eq. (1) is zero and H is therefore infinitely high). The measures consist in using valves having a high amplification factor μ, and incorporating in the common cathode lead an element or circuit having a very high differential resistance R_{d}

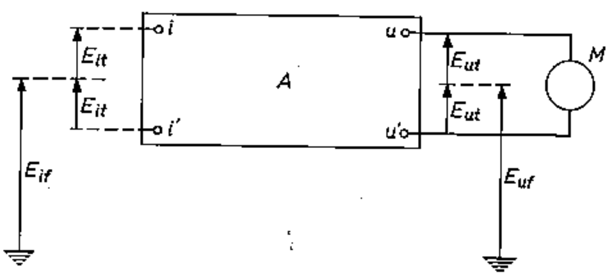

Fig. 1. Difference amplifier A. The signal to be measured, $2E_{\mathrm{it}}$, is applied between the input terminals i-i'; the amplified voltage $2E_{\mathrm{ut}}$ appears between the output terminals u-u', to which a meter M is connected. The voltages on the terminals i, i', u and u' with respect to earth are respectively $E_{\mathrm{if}} + E_{\mathrm{it}}$, $E_{\mathrm{if}} - E_{\mathrm{it}}$, $E_{\mathrm{uf}} + E_{\mathrm{ut}}$ and $E_{\mathrm{uf}} - E_{\mathrm{ut}}$. The voltages E_{it} and E_{ut} are the anti-phase signals; E_{if} and E_{uf} are the (interfering) in-phase signals. E_{if} can be very much larger than E_{it} and have a different frequency.

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[1] G. Klein and J. J. Zaalberg van Zelst, General considerations on difference amplifiers, Philips tech. Rev. 22, 345-351, 1960/61.

[2] G. Klein and J. J. Zaalberg van Zelst, Circuits for difference amplifiers, I and II, Philips tech. Rev. 23, 142-150 and 173-180, 1961/62 (Nos. 5 and 6).

[3] A distinction must be made between the guaranteed value and the actual value of the rejection factor. By the guaranteed value H_{min} is understood the calculated value of H for the case where all parameters of the circuit that influence the rejection differ about 10% in the most unfavourable sense from the nominal value. In reality the most unfavourable set of circumstances will seldom be encountered. The actual (measured) value of H will therefore nearly always be appreciably higher than H_{min}; see, for example, figs 13, 15 and 17 in article I [2].

(this can be achieved in various ways). The mentioned value 2×10^4 is in most cases amply sufficient.

A measurement problem that arose some time ago nevertheless prompted us to study the question whether difference amplifiers could be developed with an even higher rejection factor. The problem concerned a set-up for measuring the Hall effect on test bars of various materials. The set-up is shown schematically in *fig. 2* and explained in the caption. The Hall voltage to be measured, which varied in frequency from 50 to 200 c/s, was extremely small (of the order of a microvolt) and called for a millionfold amplification. This would not have presented any special difficulties if the potential drop in some test bars had not been so large as to give the electrodes a voltage E_{if} with respect to earth that was 10^5 to 10^6 times higher than the voltage to be measured; moreover E_{if} had the same frequency as the Hall voltage. This made it necessary to give the amplifier a guaranteed minimum rejection factor with the exceptionally high value of more than one million. In the most unfavourable case, E_{if} could reach an amplitude of about 200 V.

To be able to guarantee a higher rejection factor than 2×10^4, it is desirable to increase not only the amplification factor μ but also the differential resistance R_d in the common cathode lead. However, the latter is only meaningful as long as the impedance, formed by the stray capacitance bypassing R_d, is higher than R_d. One can thus obtain better rejection the smaller the stray capacitance and the lower

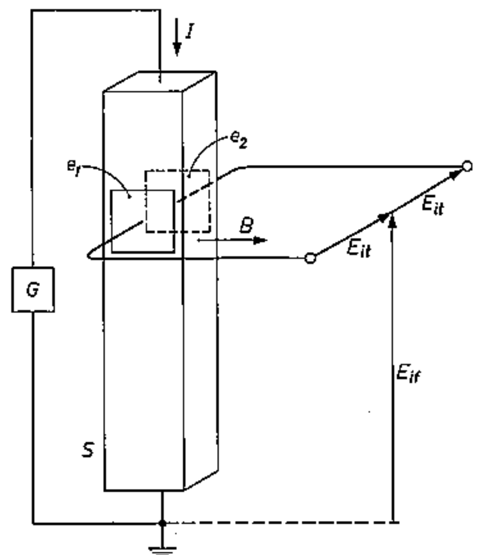


Fig. 2. Measurement of Hall effect. The generator G sends a current **I** through the test bar S, which is situated in a magnetic field with induction **B** perpendicular to the direction of **I**. The Hall voltage to be measured appears between the electrodes e_1 and e_2, perpendicular to the vectors **I** and **B**. Following the nomenclature in fig. 1, the Hall voltage is the input anti-phase signal $2E_{it}$, and the potential drop across the test bar gives rise to an input in-phase voltage E_{if}, which is much larger than E_{it}.

the frequency. Along these lines it might perhaps be possible, at low frequencies and given an extremely careful circuit design, to guarantee a rejection factor of 10^5. In the case just mentioned, however, the rejection factor required was at least 10 times higher. To meet this requirement we therefore had to look for a solution in another direction.

If an AC signal is to be measured — as in this case — we might in principle look for a solution in the use of an isolating transformer between the signal source and the input of the amplifier (which need not then be a balanced amplifier). Two reasons, however, stopped us from adopting this method:

1) The input impedance of the transformer should not be much lower than the internal resistance of the signal source. If this resistance is high (as it is in a Hall voltage generator) and if moreover the frequency is low, it follows from this condition that the primary of the transformer would have to have an impractically high inductance.
2) It is difficult to screen the windings of the transformer sufficiently from each other to keep the influence of a large in-phase voltage E_{if} small enough.

Combination of a "floating" and an ordinary difference amplifier

In article I [2]), an electrically "floating" amplifier was mentioned in passing as a possible solution of the problem (page 142), i.e. a (non-earthed) amplifier which as a whole closely "follows" the interfering in-phase voltage. It was mentioned that this method leads as a rule to complicated and unwieldy constructions, and that in nearly all cases a balanced amplifier with a very large cathode resistance offers a simpler solution.

This most certainly applies if the rejection factor to be guaranteed does not exceed a value of several times 10^4. If, however, it is to be of the order of 10^6, then this is an instance where a floating amplifier offers advantages.

The circuit decided upon consists of two difference amplifiers in cascade (*fig. 3*), the first being a floating amplifier. This passes E_{if} as the in-phase signal with virtually no change ($E_{uf} \approx E_{if}$), but considerably amplifies the anti-phase signal $2E_{if}$ which is to be measured. At its output the ratio of the anti-phase to the in-phase voltage is thus very much better, making it possible to use as the second stage (henceforth referred to as the output stage) a normal difference amplifier having a guaranteed rejection factor of the order of 100.

As will be shown presently, it is in fact possible to design the amplifier in such a way that it can follow the in-phase voltage up to a small fraction $1/K$. In such an amplifier the in-phase voltage at the input terminals is effectively reduced to E_{if}/K. Let A_1 be the amplification of the anti-phase signal, and

H_1 be the rejection factor; we then find at the output of the amplifier the following signals:

anti-phase signal . . $E_{ut} = A_1 E_{it} + \dfrac{A_1}{H_1}\dfrac{E_{if}}{K}$,

in-phase signal . . . $E_{uf} = \left(1 - \dfrac{1}{K}\right) E_{if} \approx E_{if}$.

Denoting the amplification of the output stage as A_2 and its rejection factor as H_2, we can write for the anti-phase signal at the output of this amplifier:

$$A_2\left(A_1 E_{it} + \frac{A_1}{H_1}\frac{E_{if}}{K}\right) + \frac{A_2}{H_2} E_{if} =$$

$$A_1 A_2 \left[E_{it} + \left(\frac{1}{H_1 K} + \frac{1}{A_1 H_2}\right) E_{if}\right].$$

The total rejection factor of the two amplifiers in cascade is thus:

$$H_{tot} = \frac{1}{\dfrac{1}{H_1 K} + \dfrac{1}{A_1 H_2}} . \quad . \quad . \quad (2)$$

From this it appears that both $H_1 K$ and $A_1 H_2$ must be greater than the required rejection factor H_{min}, even when H_1 and H_2, as a consequence of parameter values differing by $\pm 10\%$ [3]), have their minimum values:

$$H_{1\,min} K > H_{min} \text{ and } A_1 H_{2\,min} > H_{min} . \quad . \quad (3)$$

If we suppose the first amplifier to consist of two stages and the second of one stage, then values such as $H_{1\,min} = 10^4$ and $H_{2\,min} = 10^2$ are readily obtainable. With these values it follows from the conditions in (3) that where $H_{min} = 2\times10^6$:

$$K > 200 \text{ and } A_1 \geqq 2\times10^4 .$$

In other words, the floating amplifier must amplify the anti-phase signal at least 20 000 times and must follow the in-phase voltage closely to within 0.5%. For the amplifiers presently to be described, $A_1 \approx$ 50 000 and H_2 is greater than 500; consequently $A_1 H_2$ is greater than 25×10^6. Further, K is indeed roughly 200, while $H_{1\,min}$ is roughly 2×10^4. *A value of roughly 4 million can therefore be guaranteed for the total rejection factor*, which was more than enough for our purpose.

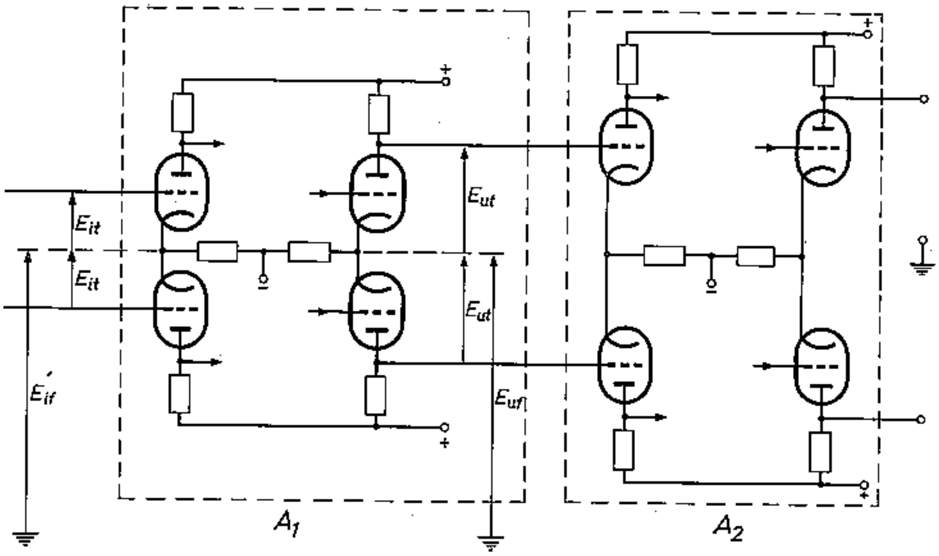


Fig. 3. Cascade circuit of an electrically "floating" difference amplifier A_1 and a normal difference amplifier A_2; only the first and last stages are shown in both cases. A_1 has a high gain for the measured signal $2E_{it}$ ($E_{ut} = A_1 E_{it}$) and "follows" as a whole the interfering in-phase voltage E_{if}, so that $E_{uf} \approx E_{if}$. A moderately high rejection factor is therefore sufficient for amplifier A_2.

In the following we shall discuss the difference amplifier designed by us for measuring Hall voltages. The problems examined will mainly concern the floating amplifier. At the end of the article we shall touch briefly on the difficulties involved when the signal to be measured is a DC voltage or has a frequency very much higher than 200 c/s.

The floating difference amplifier

The floating difference-amplifier consists of three stages. The first and second stages have identical circuits and together amplify the signals about 50 000 times. Each of the stages consists of two cascodes in balanced configuration and have a triode with cathode resistor in the common cathode lead. A simplified diagram of one stage can be seen in *fig. 4*. For the most part it is a combination of the

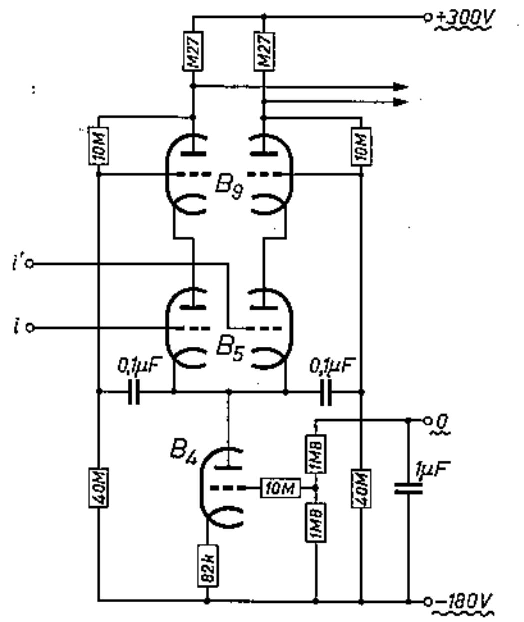


Fig. 4 [4]). Simplified circuit of the first stage of the floating difference amplifier. The double triodes B_5 and B_9 (type UCC 85) form a balanced configuration of two cascodes having a high effective amplification factor. Incorporated in the common cathode lead is a triode B_4 ($\frac{1}{2}$ UCC 85) with cathode resistor; this combination offers a high differential resistance. A high amplification factor and high (differential) resistance in the cathode lead are prerequisites for high rejection.
Correct biasing of the "upper" cascodes valves is ensured by voltage dividers. The wavy line under +300 V, 0 and −180 V indicates that these DC voltages are "floating", i.e. that an AC voltage with respect to earth (in-phase signal) is superimposed on them.

circuits shown in figures 6 and 8 in article I, where a description will be found of its operation. We shall only recall here that each cascode behaves as a triode having a very high amplification factor μ, and that the triode with cathode resistor provides the high differential resistance in the cathode lead. In this way both conditions for a high rejection factor are fulfilled.

As the successive stages are coupled together capacitively, they do not pass DC signals. At the output it is therefore not directly perceptible whether in the long run the biasing of the balanced cascodes shows disparities. In that case the valves would be operating in the non-linear part of their characteristic, resulting in distortion and hence in an error of measurement. For this reason identical (and constant) biasing of the individual stages is ensured by connecting a voltage divider to the grids of each of the "upper" cascode valves (unlike the situation shown in fig. 6 in article I). As can be seen in fig. 4, these voltage dividers are not connected to the positive terminal of the power supply but to the anode of the relevant valve. This has the effect of producing strong negative feedback for direct voltage, which makes the biasing of the cascodes practically equal. Decoupling capacitors between the above-mentioned grids and the common cathode ensure that there is no significant feedback for alternating voltage of the signal frequency. Another result of the decoupling capacitors is that the grids closely follow the voltage on the common cathode, which is a prerequisite for good difference amplifiers.

We shall deal presently with the third stage of the floating amplifier. First we shall consider the measures needed to make this amplifier "float", that is to make it follow the in-phase voltage as closely as possible.

The guiding principle in this connection is that *no point of the circuit should have any perceptible capacitance with respect to earth.* Otherwise, of course, the large in-phase voltage on such a point would give rise via this capacitance to a current that would upset the balance, at the expense of the guaranteed rejection factor. This requirement becomes all the more important the more "sensitive" is the point in question, in other words, electrically speaking, the closer it is to the input terminals and the farther away from the middle of the stages.

To satisfy the above principle the first thing to do is to enclose the amplifier in a metal can (*S*, *fig. 5*), which itself follows the in-phase signal with respect to earth. Although the can possesses capacitance (and indeed a fairly high capacitance) with respect to earth, the stray currents flowing via this capacitance are bypassed through an auxiliary cascode (see below) which keeps them out of the amplifier. The can is mounted on elastic strips, which keep it electrically insulated from the earthed chassis and also counteract microphony.

The circuit inside the can has of course various external connections. There is consequently a danger that certain points of the circuit will still "see" earth. These points are: the input terminals, the output terminals, the DC supply terminals, and the cathodes, which possess capacitance with respect to the (externally fed) heaters.

Input terminals. The input terminals are extremely sensitive. They are therefore arranged in such a way (see fig. 5) that their capacitance to earth is only 0.35 pF. They are connected to the object under measurement by cables whose screening is connected to the can, as a result of which the capacitance of 0.35 pF is increased by 0.6 pF per metre length of cable. Using 1 metre of cable, which will usually be long enough, the input capacitance thus remains below 1 pF.

Output terminals. If the (non-floating) output stage were directly connected to the second stage, the rejection of the floating amplifier would be spoilt by the fairly considerable stray capacitance

[4]) In figs 4, 6, 7 and 8 use is made of a space-saving notation for resistance values. For example,

270 means	270 Ω,
82 k	82 kΩ,
10 M	10 MΩ,
1 k 2	1.2 kΩ,
M 27	0.27 MΩ.

of the connections between the two amplifiers. Interference voltages might also be induced in these connections. Both harmful effects are smaller the lower are the output resistances in the last stage of the floating amplifier. To this end the amplifier is provided with a balanced cathode follower as a (non-amplifying) third stage with relatively low output resistances ($\approx 1/S$, where S is the transconductance of the valves).

DC supply voltages. The floating amplifier requires two DC supply voltages with respect to the cathodes of B_5 (fig. 4): +300 V and —180 V. The sources of these voltages must follow the in-phase signal $E_{\rm if}$. A simple and adequate, though not particularly practical solution would be to take the voltages from batteries mounted inside the can. In principle the voltages might also be taken from two floating power packs, but in that case the screening between the primary and other windings of the power transformer would have to meet extremely severe demands. It will be shown below how this difficulty was circumvented by using *earthed* power packs in conjunction with a few auxiliary valves.

Cathodes. The effect of the cathode heater capacitances can be eliminated by preventing alternating voltage from appearing between the cathodes and the heaters. A circuit for the heater supply has been designed which meets this requirement most satisfactorily. It is simpler than the circuit supplying the floating voltages of +300 V and —180 V, and will therefore be discussed first.

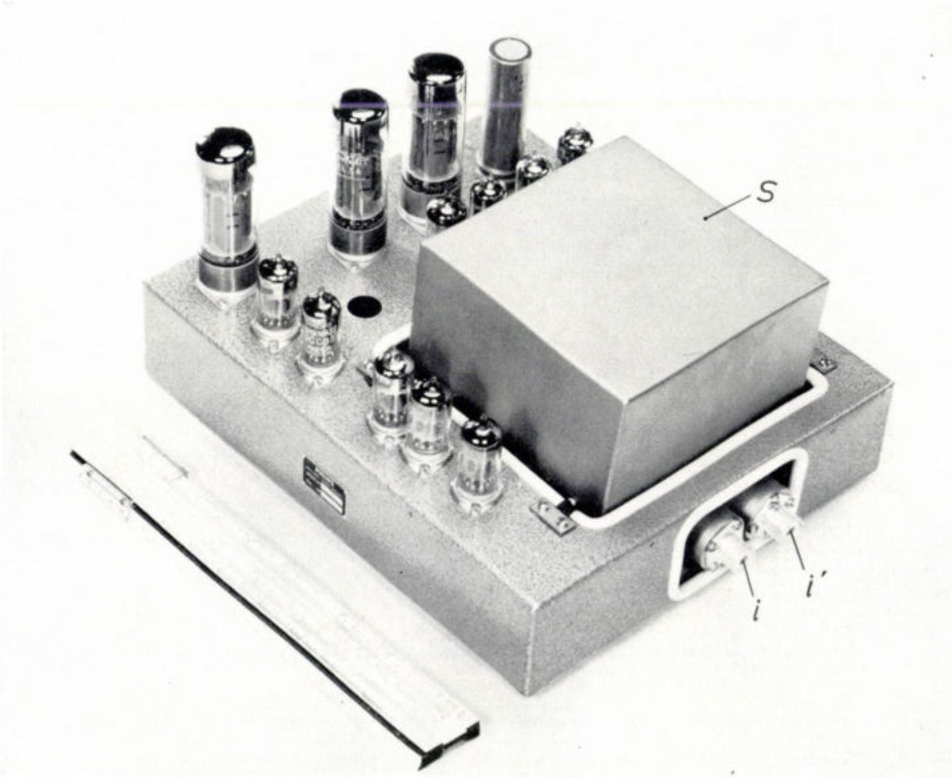


Fig. 5. The difference amplifier with an amplification of about 3.5 million and a guaranteed rejection factor of 4 million. A floating and an earthed difference amplifier are connected in cascade. The floating amplifier is enclosed in a metal screening can S and has screened input terminals i-i'. The can carries alternating voltage (maximum approx. 140 V r.m.s.) with respect to the earthed chassis, and is mounted on elastic, insulating strips.

Floating heater-current supply

The circuit shown in *fig. 6* neutralizes the cathode heater capacitances by ensuring that the heaters of the relevant valves receive the same alternating voltage ($E_{\rm if}$) with respect to earth as the cathodes. All these valves are double triodes of the type UCC85, which requires a heater current of 100 mA. This current here is direct current. This is necessary in the first place because it avoids the hum interference which is present to some extent if the indirect cathodes are heated with AC. In the second place, direct current is needed because the floating supply can then benefit from a useful property of pentodes, which is that in their normal operating region their differential resistance is much higher than their DC resistance.

The heaters concerned are connected in series and are fed from two power-supply units, one of which delivers +550 V and the other —525 V with respect to earth. Connected in series with the heaters there is at one end a pentode B_{15} (fig. 6), which acts as a current source and is biased to an anode current of 100 mA. At the other end is a pentode B_{16}, which works as a cathode follower. The control grid of B_{16} is connected to a point x elsewhere in the circuit (see fig. 8), which is not only at the DC potential that gives the valve its proper operating point but also carries with respect to earth the alternating voltage E_{if}. As a result of the latter the whole chain from the cathode of B_{16} to the anode of B_{15} acquires practically this same alternating voltage: owing to the high differential resistance of B_{15} and the high μ of B_{16}, the latter valve works as a cathode follower, the cathode of which follows the grid voltage very closely and thus acquires almost the alternating voltage E_{if} with respect to earth. Moreover, a further result of the high differential resistance of B_{15} is that the AC component in the chain remains limited to a very small value, so that all points of the chain show roughly the same alternating potential with respect to earth. This alternating potential is that which is applied to the cathode of B_{16}, i.e. the signal E_{if}. Since the cathodes of the valves in the floating amplifier also follow E_{if}, the cathode-heater capacitances have no effect.

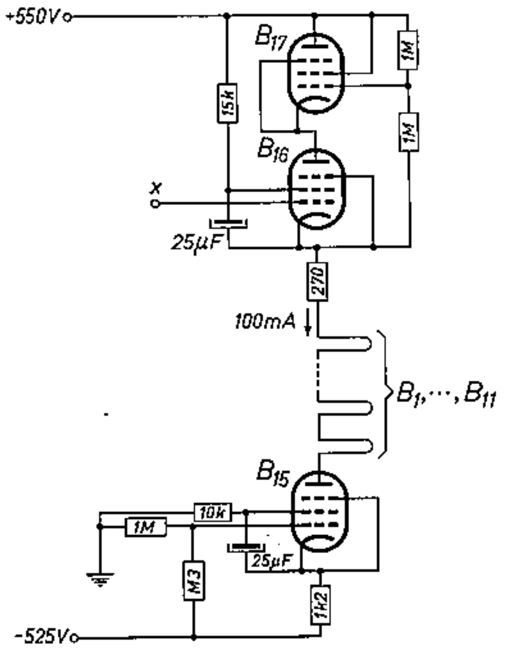


Fig. 6 [4]). Circuit for supplying in series the heaters of eleven valves type UCC 85 ($B_1, \ldots, B_{11}$) with DC current (100 mA) and for neutralizing cathode-heater capacitances in the amplifier valves.

Pentode B_{15} with cathode resistance functions as current source and is biased for an anode current of 100 mA. Pentode B_{16} works as a cathode follower, owing to its high μ and the high differential resistance of B_{15}. The cathode of B_{16} thus follows the voltage on the control grid; this is connected to a suitable point x elsewhere in the circuit (see fig. 8) which carries the alternating voltage E_{if} with respect to earth. The high differential resistance of B_{15} prevents any appreciable alternating current flowing in the circuit; all eleven heaters therefore receive practically the same AC voltage E_{if}. Since this is also on the cathodes of the valves in the amplifier, no stray currents flow between the heaters and cathodes.

Valve B_{17} is needed to prevent the dissipation of B_{16} from becoming excessive. B_{15}, B_{16} and B_{17} are power pentodes, type EL 34.

The DC voltage supply for the circuit is 550 + 525 = 1075 V, a value which is needed for the amplifiers because of the high amplitude of E_{if}. To prevent the permissible dissipation of B_{16} thereby being exceeded, a second valve is connected in series with it (the pentode B_{17}, circuited as a triode) which takes part of the voltage. The valves B_{15}, B_{16} and B_{17} are type EL 34 power pentodes.

There are altogether eleven UCC 85 valves, whose heaters are series-fed in this way. Seven of them belong to the floating amplifier and two to the output stage; the remaining two are auxiliary valves, which will be discussed presently. The other valves used in the amplifier are E types, the heaters of which are fed with alternating voltage.

Floating DC supply voltages

Fig. 7 shows only the "lower" triodes (B_5) of the first stage of the floating amplifier. The rest of the diagram gives the circuit (simplified) which is needed to make it possible to use earthed power-supply units. These are here the same units from which the heater current is obtained: they deliver +540 V (10 V lower than the point where the heater current is derived, owing to extra smoothing) and —525 V with respect to earth; the latter voltage is highly stabilized.

We shall now show how the appropriate DC voltages are obtained with the circuit in fig. 7, and then discuss the way in which the whole floating amplifier is made to follow the in-phase voltage.

As mentioned, the floating amplifier requires +300 V and —180 V with respect to the cathodes of B_5, which carry, in addition to a small DC voltage, the in-phase voltage E_{if} with respect to earth. At first sight it might seem obvious to connect that point (the cathodes of B_5) with the screening can. If that were done, however, the capacitive currents flowing via the can would not remain outside the amplifier. For this reason we connect between the points +300 V and —180 V an auxiliary cascode (B_3a-B_3b), the middle of which (point 0) can safely be connected to the can. Since B_3a is a very faithful

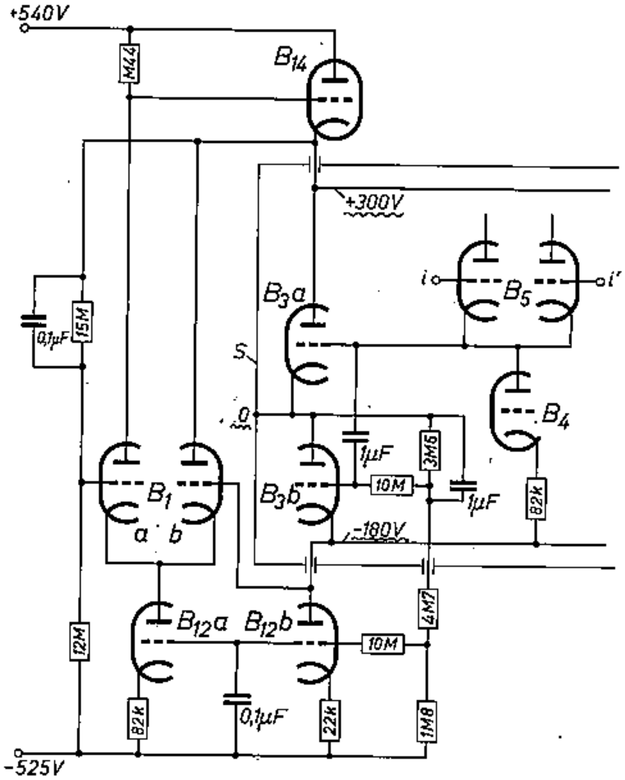


Fig. 7 [4]). Circuit for delivering the floating supply voltages +300 V and —180 V to the floating difference amplifier. B_5 belongs to the first stage of this amplifier (cf. fig. 4). *i-i'* input terminals. B_3a-B_3b auxiliary cascode with current source $B_{12}b$ in the cathode lead. Since B_3a is a cathode follower, its cathode (point 0 connected to the can *S*) has virtually the same DC voltage ($\approx$ 0) and alternating voltage (E_{if}) as the cathodes of B_5. A voltage divider 3.6 + 4.7 + 1.8 MΩ gives a DC voltage of —180 V to the grid of B_3b, which also receives the AC signal E_{if} via a 1 μF capacitor. Since B_3b is also a cathode follower, both these voltages appear on the cathode of B_3b (point —180 V).

The series triode B_{14}, controlled by the auxiliary amplifier B_1a-B_1b-$B_{12}a$, ensures that the point +300 V acquires both the DC voltage (determined by the voltage divider 15 + 12 MΩ) and the alternating voltage E_{if}.

cathode follower (its cathode lead contains the very high differential resistance of the triodes B_3b and $B_{12}b$, see fig. 7), the point 0 has virtually the same potential as the cathodes of B_5.

A voltage divider 3.6 + 4.7 + 1.8 MΩ between the points 0 and —525 V ensures that the point —180 V in fact carries this DC potential. This is because B_3b is also a good cathode follower, so that its cathode has practically the same potential as its grid, which is connected to this voltage divider.

The required DC potential at the point +300 V is obtained by using the well-known principle of stabilized power supplies: from the point +540 V the supply is effected through a series triode (B_{14}) which is controlled by an auxiliary amplifier, using the above derived DC voltage of —180 V as reference voltage. This auxiliary amplifier is a difference amplifier, consisting of a balanced arrangement of the double triode B_1a-B_1b with the triode $B_{12}a$ in the cathode lead. The grid of B_1b is connected to the point —180 V, the grid of B_1a to the tap of a voltage divider (12 + 15 MΩ) between the points +300 V and —525 V. The voltage division ratio is so chosen that when the point +300 V has the correct potential, the grids of B_1a and B_1b carry the same voltage. If, for example, the potential of the point +300 V increases, the anode current in B_1a increases, the anode voltage of B_1a (hence the grid voltage of B_{14}) consequently decreases, and since B_{14} is also a cathode follower, the cathode voltage of B_{14} — i.e. the potential of the point +300 V — likewise drops. In this way, then, the potential of this point is automatically kept constant.

It is now easily seen that the points 0, —180 V and +300 V closely follow the in-phase voltage E_{if}. This is present at the cathodes of B_5 and therefore also at the grid of B_3a and, via a 1 μF capacitor, at the grid of B_3b. (A 10 MΩ resistor between the grid of B_3b and the tap on the voltage divider of 3.6 + 4.7 + 1.8 MΩ prevents undesired loading of the voltage divider via the capacitor mentioned.) Because of the cathode-follower properties of B_3a and B_3b, their cathodes too will follow the in-phase voltage, and these cathodes are connected to point 0 and point —180 V respectively. The reference voltage on the grid of B_3b is therefore the sum of a DC and an AC voltage. Consequently the cathode of B_{14} (the point +300 V) carries with respect to earth not only the correct DC voltage but also the correct AC voltage (the in-phase signal E_{if}).

The complete diagram of the amplifier (without the earthed power-supply units and the heater-supply circuit) is shown in *fig. 8*. The thin line represents the screening can. The valves have the same numbering as in the previous figures. As can be seen, the amplifier that drives the triode B_{14} in fact consists of a balanced arrangement of two cascodes, giving a greater amplification than the balanced arrangement of two triodes in fig. 7. The control grid of B_{16} (fig. 6) is connected to the point *x* (fig. 8).

Measurements of the above-mentioned factor $1/K$ at 150 c/s have shown that the point +300 V follows the in-phase voltage to within 0.53%, while the points 0 and —180 V follow this voltage to within 0.36%.

Further details

The output stage

As stated (see fig. 3), the floating difference amplifier (A_1) is followed by a non-floating difference amplifier (A_2) as output stage. The circuit of A_2 is shown on the right in fig. 8. Here too, a balanced configuration of two cascodes is used (B_8-B_{11}, both

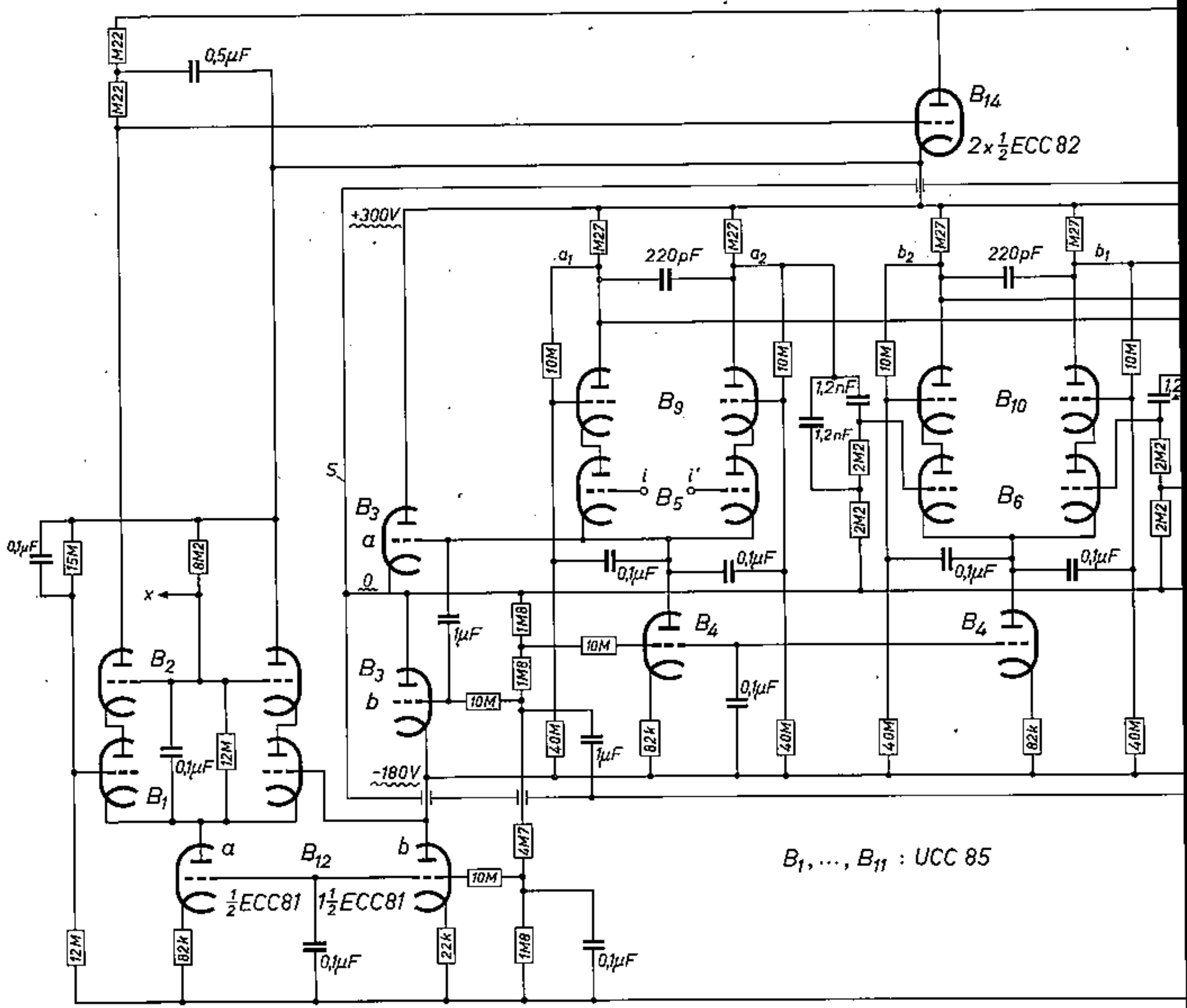


Fig. 8 [4]). Complete circuit (but excluding the power supply for the heaters) of the difference amplifier shown in fig. 5. The valves are numbered in the same way ($B_1 \ldots B_{14}$) as in the previous figures. *i-i'* input. *u-u'* output. *S* screening can enclosing the floating amplifier. The control grid of valve B_{16} is connected to the point *x* (fig. 6).

types UCC 85), having in the cathode lead a current source consisting of a triode (one half of an ECC81) with cathode resistor. The circuit was designed to deal with the high amplitude that E_{if} can acquire (approx. 200 V), which called among other things for the high supply voltages. At the output the amplified input signal appears at sufficient amplitude, and sufficiently free from in-phase voltage, that it can be further processed by conventional means.

Stepwise gain control using photoresistors

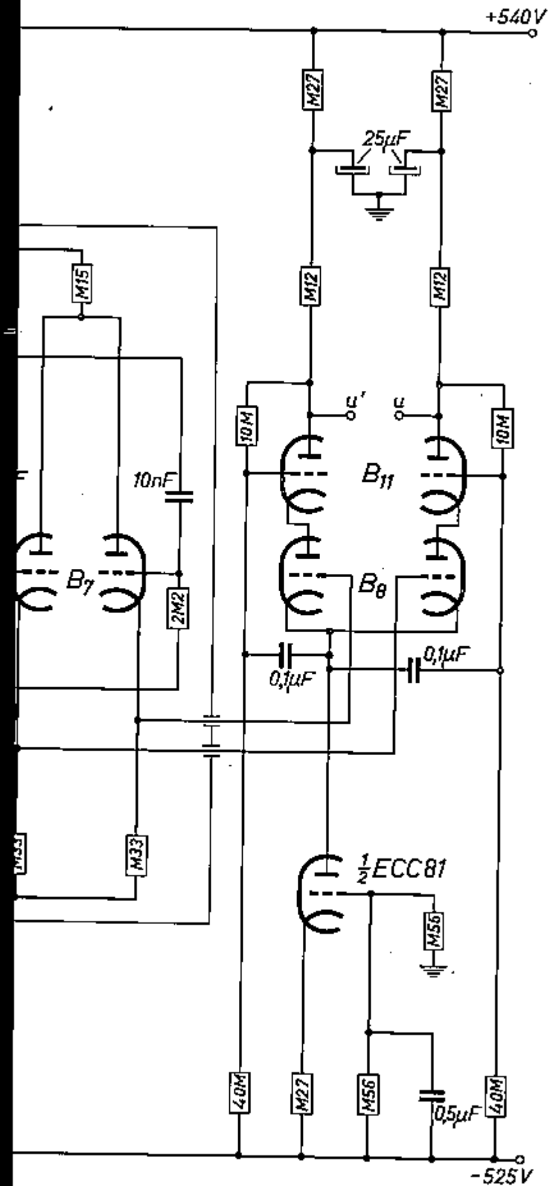
The complete amplification of several times 10^6 is not always needed. Often a gain 10 or 100 times smaller will be adequate, provided the rejection factor is not reduced to a greater extent.

A suitable method of thus reducing the gain is to introduce a resistor between a_1 and a_2 or between b_1 and b_2 (fig. 8) [5]): for a 10 times as small a gain a 47 kΩ resistor between a_1 and a_2, and for a 100 times as small a gain a 4.7 kΩ resistor between b_1 and b_2.

These resistors are switched in and out of circuit by means of *photoresistors* [6]), which do not have the drawbacks of ordinary switches or relays mounted in a can that electrically and mechanically "floats".

[5]) For methods of gain control, see article II, mentioned in footnote [2]), p. 175 *et seq.*

[6]) N. A. de Gier, W. van Gool and J. G. van Santen, Photoresistors made of compressed and sintered cadmium sulphide, Philips tech. Rev. **20**, 277-287, 1958/59.



The type of photoresistor used has a dark resistance of more than 100 MΩ, and when illuminated by a small bulb a resistance of only about 150 Ω. Photoresistors of this type can thus serve as switches for resistors with values much higher than 100 Ω and also much lower than 100 MΩ. This condition is fulfilled by the 4.7 and 47 kΩ resistors required in this case. The (earthed) electric bulbs are so positioned that they do not appreciably increase the capacitance with respect to earth of the components inside the can (*fig. 9*).

Results

The total amplification of the anti-phase signal is about 3.5×10^6. Despite this very high figure, there is not the least tendency towards oscillation [7]. During a period of many hours the amplification changes by no more than 0.5%. (If necessary this change can be made still smaller by replacing the carbon resistors, used for the anode load, by metallic resistors.)

As mentioned, a value of 4×10^6 can be guaranteed for the rejection factor. The values measured at 150 c/s on five UCC 85 double triodes at the position B_5 (fig. 8) varied from 8×10^6 to 15×10^6.

In order to neutralize part of the various interfering voltages (including noise and hum), measures were taken to reduce the amplification outside the signal frequency range (50 to 200 c/s). Below 50 c/s this is done by the *RC* networks between points a_1-a_2 and the grids of B_6; above 200 c/s by the 220 pF capacitors between a_1 and a_2 and between b_1 and b_2 (fig. 8). The total interference voltage, derived at the input, varied from 1.2 to 2.0 μV on the five UCC 85 valves tested. The frequency pass band can be given a sharper cut-off by conventional methods behind the amplifier (high-*Q* filters, selective detection).

Owing to the exceptionally low input capacitance (less than 1 pF per metre of cable), even considerable unbalance in the signal source is not able to spoil the rejection factor. If, for example, two signal sources with identical e.m.f.'s are connected, one of which has zero internal resistance and the other an internal resistance of 2000 Ω, it is still possible to guarantee the very high rejection factor of 1.5×10^6 (at 50 c/s).

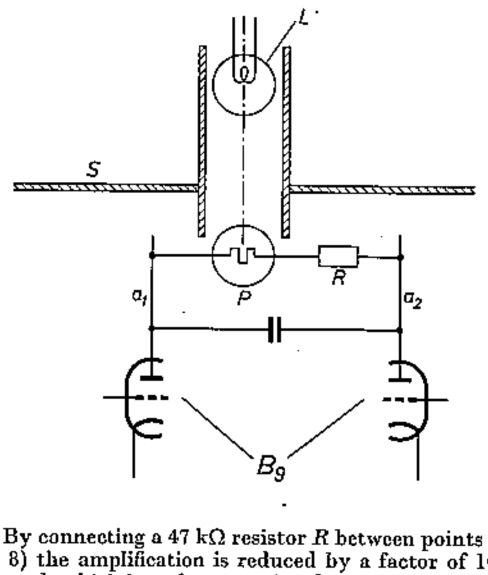


Fig. 9. By connecting a 47 kΩ resistor R between points a_1 and a_2 (fig. 8) the amplification is reduced by a factor of 10. The switch used, which introduces no significant stray capacitance, is a photoresistor P, which has a dark resistance of more than 100 MΩ, and, upon illumination by the electric bulb L, a resistance of only about 150 Ω.

In the same way it is possible to connect between b_1 and b_2 (fig. 8) a 4.7 kΩ resistor, which reduces the gain by a factor of 100.

[7]) The explanation will be found in article II, mentioned in footnote [3]), p. 177.

Other signal frequencies

To conclude we shall touch briefly on the difficulties to be expected if the principles described are to be applied at higher or lower signal frequencies.

The higher the frequencies the greater is the influence of stray capacitances. This means that the floating difference amplifier will not follow the in-phase voltage so closely, and that the rejection factor will therefore be lower. On the other hand, it is pointless to ask for a particularly high rejection factor at high frequencies, for asymmetries in the connections between the amplifier and the signal source make a very high rejection factor in any case impossible [8]).

The lower the signal frequency the better the floating amplifier follows the in-phase voltage, but the more substantial are the ordinary drawbacks encountered at very low frequencies, namely that coupling and decoupling capacitors of very high capacitance then have to be used. Where DC signals are involved the difficulty may arise, especially if the in-phase voltage (and hence the supply voltage) is high, that the dissipation of some valves, which are required to pass continuously a large current at high voltage, will become excessive. In that case each of the valves involved will have to be replaced by two valves in series, so that the voltage drop — and thus the power to be dissipated — will be divided over the two valves. A similar case arose in the heater power supply (fig. 6): here it was necessary to connect a valve B_{17} in series with B_{16}.

[8]) See figs 18 and 19 of article II, mentioned in footnote [2]).

Summary. For measuring Hall voltages (minimum value of the order of 1 μV) in the frequency range from 50 to 200 c/s, an amplifier with a gain of one million was needed. The measuring electrodes could acquire such a high potential with respect to earth (the in-phase voltage) that it was necessary to give the amplifier a guaranteed rejection factor of at least one million. This problem has been solved by using two difference amplifiers in cascade, the first of which is electrically "floating", i.e. closely follows the in-phase voltage (to within about 0.5%). Among the measures discussed are the circuit that neutralizes the cathode-heater capacitance of the valves in the floating amplifier, and the circuit from which the floating DC supply voltages are obtained. The guaranteed rejection factor is 4 million, the amplification about 3.5 million. Resistors for reducing the gain by a factor of 10 or 100 are switched in and out of circuit by means of photoresistors.