AMPLIFICATION OF LOW ALTERNATING VOLTAGES WITH THE MAGNETIC AMPLIFIER

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INTRODUCTION

The magnetic amplifier has hitherto been used almost exclusively for the amplification of currents which vary only slowly in relation to the time occupied by a period of the mains frequency. In the following remarks, the conditions will be investigated under which pre-magnetised chokes are also suitable for the amplification of alternating voltages. In this connection we shall be thinking mainly of its application as zero current amplifier in a.c. bridges, though it can equally well be employed as a measured value amplifier in control circuits.

COMPARISON OF a.c. VOLTAGE INDICATORS

The accuracy of a bridge measurement is critically dependent upon the sensitivity and zero point precision of the zero indicator. The most widely used zero apparatus for a.c. bridges is the vibration galvanometer. Its properties correspond in many respects with the requirements of an indicator apparatus in the case of a.c. bridges. The indication is proportional to the amount of the bridge voltage and therefore independent of the phase. The moving system can be tuned simply to the working frequency. As the mechanical vibration system composed of re-setting force and mass has a steep resonance curve owing to its extremely low damping, the harmonics of the measured voltage scarcely affect the balance. A further advantage is the low input resistance of the instrument. This fact is important in relation to the interference field sensitivity. The higher the resistance of the input circuit, the more difficult it is to protect the indicating instrument from the strong leakage fields, which are always present in the heavy current test station, by screening or other measures.

In spite of these advantageous features, there has been no lack of attempts to replace the vibration galvanometer by other measuring devices. These efforts were mainly due to the requirements of the works test station, which regularly carries out control and acceptance tests of cables etc. Conditions of space are here usually very unfavourable. For example, neighbouring machinery sets up vibration to which a vibration galvanometer is particularly sensitive. Moreover, in view of the large number of measurements to be carried out in one day the low-brightness of the screen picture is a disturbing factor. A further disadvantage is the low resistance to overload of the vibration galvanometer, which if the electrical values of the cable to be measured differ considerably from the intended value, due to some defect of construction, may easily lead to destruction of the vibration system.

In the endeavour to replace the sensitive vibration galvanometer by the robust rectifier instrument it is an obvious matter to connect an amplifier between the latter and the bridge. Appropriate electronic amplifiers were constructed for the purpose but for several reasons they were used in practice only to a slight extent. In order to suppress the disturbing influence of the harmonics, the amplifier had to be constructed in the form of a resonance amplifier. It is very difficult to build resonance circuits for 50 c/s with adequate selectivity so that the indicating instrument mainly shows the harmonic voltage, especially in the neighbourhood of the balancing point. Another disadvantageous effect is that the thermionic valve represents a high ohmic amplifier component and is consequently highly sensitive to disturbing fields. The expedient of increasing the tuning frequency for the subsequent resonance amplifier stages, on the principle of the heterodyne receiver, by mixing with a fixed intermediate or high frequency oscillator frequency, does not introduce any substantial advantages.

The rectifier instrument alone is also useless as a zero current indicator when a sensitive galvanometer is being employed instead of the normal moving coil instrument, since below a certain threshold voltage practically no rectifier effect occurs in the metal rectifier. This difficulty can be overcome by the aid of special modulation circuits. Figure 1 shows one of the many circuit modifications. Eight resistances, of which $R_1 = R_4 = R_5 = R_8$ and $R_2 = R_3 = R_6 = R_7$, were connected to form a bridge. At the points a, b they are connected to the measured voltage and at the points c, d to an auxiliary voltage U_h which is taken off the voltage U_m feeding the measuring bridge. U_h and the resistances are so selected that the rectifiers carry rated currents. No current flows through the indicating instrument, since the bridge and all the voltages are symmetrical. If the measured voltage U_m is applied to the points a, b, the potentials of the points e and f are displaced by the additional voltage drops across the resistances, so that a current flows through the indicating instrument. Since the rectifiers carry approximately rated current, independently of U_m rectification is linear even in the case of very low signal voltages. The circuit may also be extended by the rectifiers shown in faint lines, so that the same behaviour is obtained in both half-waves of the alternating voltage.

With this arrangement, approximately the sensitivity of a vibration galvanometer may be attained. The deflection of the indicating instrument, for constant U_h, is proportional to the bridge voltage and to the phase angle between U_h and U_m. This property makes balancing of the bridge extremely difficult, as the phase of the bridge voltage is liable to undergo much alteration in the vicinity of the point of balance and thus the deflection of the indicating instrument may increase in the course of balancing, although the absolute value of the measured voltage decreases, and vice versa.

The uncertainty of the indications does not occur when the frequencies of U_h and U_m differ by a small amount, e.g. 1 c/s. The indicated current fluctuates in this case with the differential frequency. Quite apart from the fact that it is difficult to keep to a given small frequency difference, the sensitivity of the zero indicator is reduced, since the natural frequency of the sensitive indicating instrument is in general not high in comparison with the difference in frequency. Balancing is also hardly any easier than in the case of the vibration galvanometer, because the amplitudes of oscillation have to be read.

VOLTAGE AMPLIFICATION OF THE REACTOR CIRCUITS WITHOUT FEEDBACK

The disadvantages of the indicating methods described suggest investigating the feasibility of using the magnetic amplifier for the purpose under consideration [1]*. The fact that this amplifier is extremely insensitive to interference voltages is a favourable circumstance in this connection. It also suppresses of its own accord the higher harmonics of the measured magnitude owing to its natural low-pass characteristic, and can be made so robust as to withstand the roughest treatment. The operation of the various magnetic amplifier circuits of interest in this connection cannot be discussed within the scope of the present paper, but reference must be made to the extensive literature on the subject [2, 3].

The simplest amplifier arrangement (Figure 2) consists of two identical reactors each with two windings, a load or working winding and a control winding. The working windings W_b are connected to the working voltage U with the frequency f_b. The windings are connected together in such a way that no current can flow in the short circuited control circuit if there is no pre-magnetisation.

The parallel reactor connection (Figure 2a) is out of the question for the amplification of alternating currents. If in fact an alternating voltage is applied to the control winding, alternating voltages are also induced in the working windings and these voltages are in phase with each other. Parallel connection of the working windings thus represents a short circuit for the control alternating voltage and neither of the two reactors is pre-magnetised. If the alternating control voltage is replaced by a d.c. voltage, the short circuit causes a slow increase in the control current I_s and thus also of the working current I_b.

In the case of the series reactor connection (Figure 2b) there is no short-circuit, so that a variation of the control current will not induce any compensating current in the working circuit. The dynamic properties of this circuit are readily understood with a few assumptions. It will thus be assumed that the magnetisation curve is rectangular and that the frequency of the control alternating current (f_s) is small in comparison with the working frequency f_b of the amplifier. If the control magnitude is a direct current, only the ohmic resistance R_s of the control circuit acts as input magnitude. The attainable voltage amplification is then

$$V_{uo} = \frac{\delta I_b R_b}{\delta I_s R_s} \quad \ldots\ldots (1)$$

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* For references, see end.

Owing to the high permeability of the reactors, the alternating current, in the absence of pre-magnetisation, is negligibly small. With pre-magnetised reactor, the alternating current set up is such that the mean value of the alternating flux equals the pre-magnetisation flux. Consequently

$$w_s I_s = w_b I_b . \qquad \text{...... (2)}$$

Substituting in equation (1), we get:

$$V_{uo} = \frac{\delta I_s \frac{w_s}{w_b} R_b}{\delta I_s R_s} = \frac{w_s}{w_b} \frac{R_b}{R_s} . \qquad \text{...... (3)}$$

If an alternating voltage with the frequency f_s is applied to the control winding the input resistance also contains an inductive component. The voltage amplification is now:

$$V_u = \frac{w_s R_b}{w_b \sqrt{(\Omega_s L_s)^2 + R_s^2}}, \quad \Omega_s = 2\pi f_s . \qquad \text{...... (4)}$$

According to Gale and Atkinson[4], on the assumption of a rectangular magnetisation characteristic, the control circuit inductance is

$$L_s = \frac{w_s^2 R_b}{w_b^2 4 f_b} \qquad \text{...... (5)}$$

Due to the periodically varying permeability, L_s is not a constant but a mean value which is associated with the mean flux Φ_M of the reactors. If equation (5) is substituted in equation (4) and taking equation (1) into account, we have after a few transformations:

$$V_u = \frac{V_{uo}}{\sqrt{\left(\frac{\pi}{2} \frac{f_s w_s}{f_b w_b} V_{uo}\right)^2 + 1}} \qquad \text{...... (6)}$$

If the abbreviation $f_b w_b / f_s w_s = m$ is introduced, where m denotes the inertia factor, we have

$$\frac{V_u}{V_{uo}} = \frac{1}{\sqrt{\left(\frac{\pi}{2} \frac{V_{uo}}{m}\right)^2 + 1}} . \qquad \text{...... (7)}$$

Figure 3 represents equation (7).
It supplies the factor by which the voltage amplification for a.c. control is less than for d.c. control. In order to be able to estimate the amplification to be obtained at f_s = 50 c/s, values such as the following, which may also be attained in practice, are assumed:

$$\frac{R_b}{R_s} = 10, \quad \frac{f_b}{f_s} = 30.$$

Substituting these values in equation (6) we have:

$$V_u = \frac{w_s}{w_b} \frac{10}{\sqrt{0.28 \ (w_s/w_b)^4 + 1}}$$

As shown in Figure 4 the greatest amplification is obtained when w_s/w_b = 1.3 but even then it is less than 10. For comparison, the voltage amplification in the case of direct current control V_{uo} is included. While V_{uo} can be increased by a large number of turns ratio, the a.c. amplification drops steeply for a large w_s/w_b. The series reactor connection without feedback is not, therefore, suitable for the amplification of alternating voltages, since its amplification factor is too small.

According to equation (2) control flux and working flux are equal for the series reactor connection. If the control and working numbers of turns only differ, as in the present case, to a slight extent, the control current is approximately equal to the alternating current and no current amplification takes place. The voltage and thus the power only are amplified owing to the fact that on account of the small control frequency f_s in comparison with f_b for the same current a lower voltage need be applied to the control coil than that which drops across the load resistance R_b.

VOLTAGE AMPLIFICATION IN THE CASE OF REACTOR CONNECTIONS WITH FEEDBACK

The control flux demand may be substantially reduced when a large part of the pre-magnetisation flux is covered by the a.c. circuit. Figure 5 shows the two most usual feedback circuits. In the case of circuit (a), the working current flowing through the load resistance is rectified and fed to a special feedback winding w_k. In circuit (b), a rectifier is connected in series with each of the two parallel connected reactors. As the rectifier suppresses the current in one direction, only half of the magnetisation characteristic curve will be used. If the whole of the magnetisation characteristic curve has to be used, an additional flux is required which displaces the lower limit induction up to the opposite saturation induction. Since the series reactor circuit, owing to the additional feedback coil, requires more winding space, it has in the course of time been increasingly replaced by the saturation circuit (b).

Figure 6 shows the variation of flux and current in the auto-saturation circuit when the control flux Θ_s or the flux Φ_s pre-magnetises the reactors. The instantaneous value of the flux (Φ_I, Φ_{II}) cannot be less than Φ_s or greater than the saturation flux $\Phi_{s\phi}$. For the case of fully compensated feedback [circuit (a): $w_k = w_b$, circuit (b): ideal rectifiers] and assuming a rectangular magnetisation characteristic, the whole of the pre-magnetising flux is provided by the a.c. circuit. An infinitely small control flux is sufficient to alter the degree of saturation of the reactors. In reality, the magnetisation characteristic departs more or less from the ideal form and the control current source has to provide a flux which is to some extent the equivalent of the flux along the rising section of the magnetisation characteristic. Accordingly, the magnitude of the amplification attainable depends to a considerable extent on the properties of the core material, though the latter play a subordinate part in the circuit without feedback.

In order to ascertain voltage amplification the control circuit inductance L_s must again first be determined. As Hedstroem and Borg [3] show, it will be sufficient for this purpose to consider the mean flux Φ_m. If the frequency of the control current f_s is small in comparison with the working frequency f_b, the control current I_s and hence the mean flux during one period of the working voltage can be regarded as approximately constant. In the case of a control current discontinuity δI_s which varies the mean flux by $\delta \Phi_M$ an inductance

$$L_s = 2\, w_s\, \delta \Phi_M / (\delta I_s) \qquad \text{...... (8)}$$

is on the average effective. The flux fluctuates before the control instant between Φ_{s1} and $\Phi_{s\phi}$ [mean value $\Phi_{M1} = 0.5\,(\Phi_{s\phi} + \Phi_{s1})$] and afterwards between Φ_{s2} and $\Phi_{s\phi}$ [mean value $\Phi_{M2} = 0.5\,(\Phi_{s\phi} + \Phi_{s2})$]. The mean value therefore varies by

$$\delta \Phi_M = 0.5\,(\Phi_{s1} - \Phi_{s2}). \qquad \text{...... (9)}$$

The variation of the mean voltage at the load resistance during half a period of the working frequency [$\delta t = 1/2 f_b$] is:

$$\delta U_M = U_{M1} - U_{M2} = U - w_b \frac{\delta \Phi_1}{1/(2 f_b)} - \left[U - w_b \frac{\delta \Phi_2}{1/(2 f_b)} \right] =$$

$$= \frac{w_b}{1/(2 f_b)} (\delta \Phi_2 - \delta \Phi_1) = 2\, w_b\, f_b\, (\Phi_{s\phi} - \Phi_{s2} - \Phi_{s\phi} + \Phi_{s1}), \qquad \text{..... (10)}$$

$$\delta U_M = 2\, w_b\, f_b\, (\Phi_{s1} - \Phi_{s2}) = \delta I_b\, R_b, \qquad \text{..... (11)}$$

so that

$$\Phi_{s1} - \Phi_{s2} = \delta I_b\, R_b / (2\, w_b\, f_b)\,. \qquad \text{..... (12)}$$

Equation (12) substituted in equation (9) and the latter in equation (8) give

$$L_s = \frac{w_s}{\delta I_s} \frac{\delta I_b\, R_b}{2\, w_b\, f_b}\,. \qquad \text{..... (13)}$$

The voltage amplification in the case of alternating current control is then, as in the case of the series connection without feedback

$$V_u = \frac{\delta U_{Rb}}{\delta U_s} = \frac{\delta I_b R_b}{\delta I_s \sqrt{(2\pi f_s L_s)^2 + R_s^2}} \; . \qquad \ldots\ldots (14)$$

Equation (13) substituted in equation (14), if V_{uo} is the voltage amplification in the case of direct current control, gives:

$$\frac{V_u}{V_{uo}} = 1 \Bigg/ \sqrt{\left(\pi \frac{f_s w_s}{f_b w_b} V_{uo}\right)^2 + 1} \qquad \ldots\ldots (15)$$

The inertia factor m again comes into consideration, so that the final result is

$$\frac{V_u}{V_{uo}} = 1 \Big/ \sqrt{(\pi V_{uo}/m)^2 + 1} \; . \qquad \ldots\ldots (16)$$

If V_{uo} is large in comparison with m, equation (16) may be simplified as follows:

$$\frac{V_u}{V_{uo}} = \frac{1}{\pi} \frac{m}{V_{uo}} \, , \qquad \ldots\ldots (17)$$

$$V_u = \frac{m}{\pi} = \frac{f_b w_b}{f_s w_s} \frac{1}{\pi} \, , \qquad \ldots\ldots (18)$$

i.e. in the case of a.c. control the voltage amplification is independent of the voltage amplification in the case of direct current control and proportional to the inertia factor m. Equations (16) and (17) are represented in Figure 7.

Finally, we have still to bring voltage amplification in the case of direct current control into relation with the reactor data.

$$V_{uo} = \frac{\delta U_{Rb}}{\delta U_s} = \frac{\delta U_{Rb}}{\delta I_s R_s} = \frac{\delta U_{Rb}}{R_s \, \delta \Theta_s} w_s \; . \qquad \ldots\ldots (19)$$

The working voltage U applied to the reactors must be measured in such a way that in the absence of pre-magnetisation, the magnetisation characteristic will be rectilinear from one saturation inflection to the other. Then

$$U = \frac{2\pi}{\sqrt{2}} \cdot f_b w_b \Phi_{sg} \, , \qquad (\Phi_{sg} = \text{saturation flux}) \qquad \ldots\ldots (20)$$

If, for $\Phi_s = 0$, the voltage U_{Rb} applied to the load resistance is also 0, then

$$U_{Rb} = 2 \frac{\sqrt{2}}{\pi} U \frac{\Phi_s}{\Phi_{sg}} \; . \qquad \ldots\ldots (21)$$

Equation (20) substituted in equation (21) and the latter in equation (19) give

$$V_{uo} = 4 \frac{f_b w_b w_s}{R_s} \frac{\delta \Phi_s}{\delta \Theta_s} \; . \qquad \ldots\ldots (22)$$

In order to obtain high a.c. amplification, the number of control turns w_s must be made low. This will of course reduce V_{uo}, but this loss can be compensated by increasing the number of working turns w_b or the working frequency f_b. There are upper limits for both w_b and f_b, on the one hand due to the available winding space and on the other to the eddy current losses occurring in the core. The latter also smooth out the steep section of the magnetisation characteristic curve, so that for high working frequencies the quotient $\delta \Phi_s / \delta \Theta_s$ drops sharply.

In contrast with the properties of thermionic valves, amplification with the magnetic amplifier decreases with increasing control frequency. The harmonics of the control current are amplified less than the fundamental wave. If it is desired to avoid these linear distortions, it will be necessary to arrange for the inertia factor m, applicable to the control frequency range concerned, to be substantially greater than V_{uo}.

THE RESONANCE AMPLIFIER

The foregoing section showed that voltage amplification depends to a great extent on the inertia factor m. As m will nearly always be less than 100, it appears advisable to consider some other means of increasing V_u.

The sensitivity of the parallel reactor connection with self-saturation can be considerably increased if, as Figure 8 shows, a capacitance is applied in series with the reactors. We then obtain a series resonance circuit in which inductance varies in rhythm with the control current. The fact that each of the two reactors is connected in series with a rectifier is not important so far as resonance is concerned, since the rectifiers only regulate the current distribution between the two reactors and do not in the least affect the inductive character of the parallel reactor connection. Owing to the series capacitance, the alternating voltage applied to the reactors is no longer constant but varies with the output. The stability of the amplifier therefore becomes decidedly more critical. The danger of instability is, however, diminished by the fact that the load resistance damps the series resonance circuit.

The measurements described below were carried out with ring-core reactors. Iron cross-section 1 sq.cm.: mean length iron path 16 cm.: core of mumetal, w_b = 3000, w_s = 500, f_b = 500 c/s.

Figure 9 shows the working characteristics $\Theta_b = f(\Theta_s)$ or $U_{Rb} = f(\Theta_s)$ of the resonance amplifier. Only the steep section of the characteristic is shown. Owing to the series capacitance, the shape of the characteristic is retained; only the steepness increases with diminished C. In this case there is no point in reducing the capacitance beyond 0.7 μF, since voltage fluctuations or variations of the control circuit resistance may then set up natural oscillations.

Figure 10 shows very well how sensitivity is increased by series capacitance. In this figure, the slope of the characteristics of Figure 9 has been plotted against series capacitance. In this example, therefore, the resonance amplifier is approximately 2.5 times more sensitive than the amplifier without series capacitance.

Between the stable range and the instability limit there is a transition state in which the arrangement is excited to natural oscillations. The origin of such oscillations, which have their source in the curvature of the magnetisation characteristic of the reactors, is extremely complex on account of the non-linear relationships. In the case of the resonance amplifier, natural oscillation expresses itself in periodic fluctuations of the total flux, and thus of the load current i_b of the reactors. The load current appears to have superposed on it a lower frequency which may be designated as f_{sub}. In the oscillogram of Figure 11 the working frequency and subharmonic frequency are as 1:10.

The natural oscillations are unwanted in the resonance amplifier. It is therefore important to know under what limit conditions they occur. In the present amplifier, therefore, the subharmonic frequency f_{sub} was measured in its dependence upon the individual parameters, such as C, R_s, R_b and Θ_s the measurements being made by means of a cathode-ray oscillograph and valve generator, using Lissajous figures, in order to eliminate any reactions in the amplifier.

The frequency of natural oscillation is lower, the smaller is the series capacitance C. Figure 12 shows the dependence of the frequency f_{sub} upon the load resistance R_b and the control circuit resistance R_s. Both resistances influence it in the same way. If R_b and R_s increase, f_{sub} also increases. The dependence on the control flux Θ_s is also important. In Figure 12 the latter influence is expressed by the fact that for each curve $f_{sub} = f\ (R_s)$ = constant two frequencies are always shown. The upper frequency is related to the lower limit flux, i.e. the smallest control flux at which oscillation first just sets in. The lower frequency accordingly relates to the upper limit flux. Consequently, the greater the control flux, the lower is the frequency of the oscillation excited. The control flux now represents only a small part of the entire pre-magnetisation flux, the greater part of it being supplied by the self-saturation rectifiers. All factors influencing the feedback ratio also influence f_{sub}. Here the control circuit resistance R_s is of particular importance. If it is reduced an additional flux may be formed across the control circuit, increasing the pre-magnetisation and thus varying f_{sub} in the same way as with an increase in the control flux.

Since the load resistance determines the magnitude of the current in the saturation phase, it is also a decisive factor in charging the capacity in the separate half-waves. Since the alternating current i_b is inversely proportional to the load resistance for a given pre-magnetisation, a reduction of R_b must have precisely the same effect on the frequency f_{sub} as a reduction of R_s or an increase of Θ_s.

Figure 13 indicates the oscillation range limits. Measurements were carried out for two series capacitances $C = 0.7\ \mu F$ and $0.6\ \mu F$. The voltage U_{Rb} across the load resistance which gives a measure of the output of the amplifier, is plotted as abscissae. It is clear that the oscillation range limits depend only to a slight extent on the control circuit resistance R_S unless the control circuit is opened. The influence of the load resistance R_b is all the greater. With increasing R_b, the oscillation range becomes steadily more restricted. It is therefore expedient to operate the amplifier with the maximum possible load resistance. A comparison of the oscillation ranges for the two capacitances shows that with decreasing series capacitance, the oscillation range widens considerably.

CONSTRUCTION OF THE AMPLIFIER

With the ring-core reactors described, a two-stage zero-current magnetic amplifier was constructed with the circuit shown in Figure 14. The signal voltage is fed to the control coil of the first reactor pair. The alternating current, after rectification in Gl_1, passes through the load resistance R_{b1}. Between R_{b1} and the control coil of the following stage lie a low-pass filter and an additional series capacitor which keep all frequencies over 400 c/s, as well as the direct-current component away from the control coil of the second stage of the amplifier. Suppression of all higher frequencies (especially the 1000 c/s voltage) is necessary, as these have an incomparably greater amplitude than the signal voltage, and would considerably overload the second stage and lead to instability. A band pass with high selectivity, which contains chokes instead of resistances, is also connected to the load resistance of the second reactor pair. The upper limit frequency is here considerably lower, since the indicating instrument only serves to detect the 50 cycle components. The auxiliary control circuit, which gives the reactors a constant basic magnetisation, displaces the working point to the centre of the linear part of the working characteristic. The voltage amplification of a stage with direct current control, amounts to $V_{uo} = 260$. On the other hand, the inertia factor is small: $m = 60$. The simplified equation (18) can therefore be used to ascertain the voltage amplification. From this $V_u = 19.2$. Having regard to the voltage loss occurring in the intermediate circuits, the result is a total amplification of

$$V_{u\ \mathrm{total}} = 19.2 \times 19.2 \times 0.35 = 129.$$

So that the indicating instrument of the amplifier will show a clearly readable deflection corresponding to $5\ \mu A$, a 50 cycle voltage of 0.15V must be applied to the load resistance of the second amplifier stage. The corresponding amplifier input voltage is equal to 1.2 mV when $V_{u\ \mathrm{total}} = 129$. For the case of the resonance amplifier, pre-calculation of the amplifier factor gives rise to considerable difficulties, and it must therefore be measured. Figure 15 shows the dependence of the instrument current on the input voltage of the amplifier. The control voltage source had an internal resistance of $2000\ \Omega$. With an initial voltage of 0.25 mV a deflection appropriate to $10\ \mu A$ is obtained. A 50 cycle voltage of 0.19 V must therefore be available at the load resistance of the second stage of the amplifier. The voltage amplification is now $V_{u\ \mathrm{total}} = 760$. This figure lies far above that attainable with amplifiers without series capacitance.

Figure 16 shows the complete set. To facilitate balancing, the sensitivity can be varied by series resistances in the first control circuit. To provide optimum matching of the input resistance of the amplifier with the measurement bridge, the first reactor pair was made with three different control windings, which can be selected by a switch. A small instrument permits control of the correct working point of both stages of the amplifier.

SUMMARY

The object of this paper is the investigation of the conditions under which premagnetised reactors are suitable for the amplification of small alternating currents, for example in place of the vibration galvanometer. The parallel reactor connection with self-saturation proves to be the most suitable for the purpose. Simple expressions are derived for the voltage amplification V_u on the assumption of a rectangular magnetisation characteristic. The amplification factor may be increased by connecting a capacitance in series with the reactors. A two-stage amplifier is then described, which operates on a working frequency of 500 c/s. A voltage of 0.25 mV can be measured with this apparatus. At the amplifier output, the set includes an ordinary moving coil instrument with point bearings.

REFERENCES

[1] F. KÜMMEL. The application of magnetic amplifiers for zero-current indications at 50 cycles. Dissertation Technische Hochschule Darmstadt, 1952.

[2] T. Buchhold. Theory of the magnetic amplifier. Arch. Elektrotechn., 37 (1943) 197-211.

[3] S.E. HEDSTROEM and L.F. BORG. Transductor fundamentals. Electronics, 21 (1948) 88-93.

[4] H.M. GALE and P.D. ATKINSON. A theoretical and experimental study of the series-connected magnetic amplifier. Proc. I.E.E., 96 (1949) 99-124.

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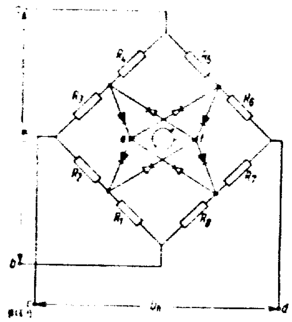


U_m = measured voltage

U_h = auxiliary voltage

$U_h \gg U_m$

Fig.1: Rectifier-modulation circuit. The deflection of the indicating instrument is proportion to U_m and to the cosine of the angle between U_m and U_h.

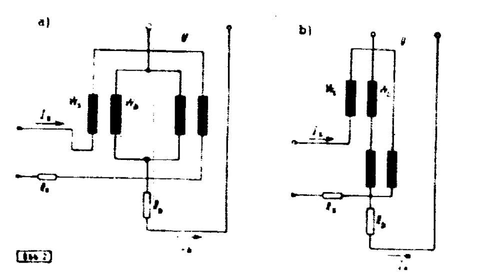


Fig.2: Basic circuit of the magnetic amplifier without feedback.

(a) Parallel reactor connection
(b) Series reactor connection.

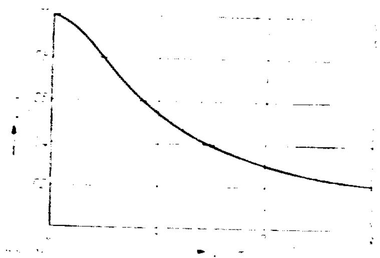


Fig.3: Ratio of voltage amplification for a.c. output V_u to voltage amplification for d.c. output V_{uo}, plotted against the quotient V_{uo}/m (m = inertia factor) for the magnetic amplifier without feedback.

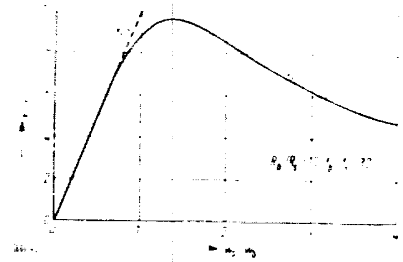


Fig.4: Voltage amplifications V_u and V_{uo} plotted against ratio of number of terms w_s/w_b. The maximum voltage amplification V_u = 9.5 is obtained for w_s/w_b = 1.3.

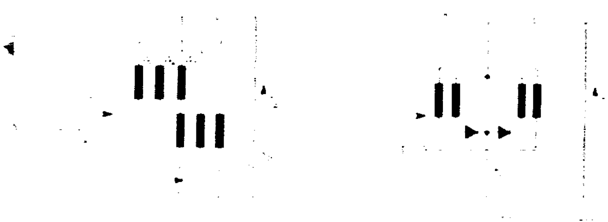
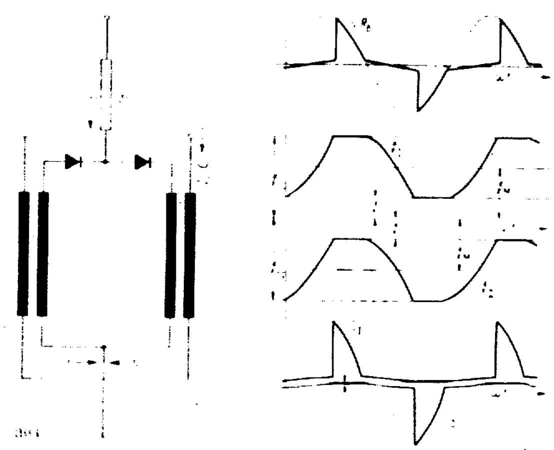
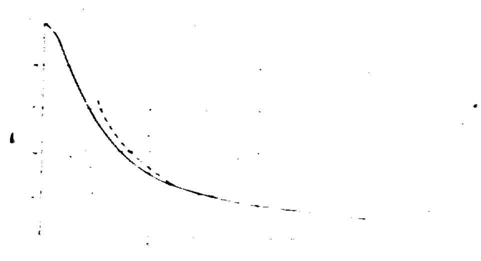


Fig.5: Basic circuits of the magnetic amplifier with feedback. (a) Series reactor connection with separate feedback winding, (b) Parallel reactor connection with self-saturation. The output voltage drops across the load resistance R_b (U_{Rb}).



Φ_M - mean reactor flux

Fig.6: Parallel reactor connection with self-saturation. The alternating flux of the two reactors I and II fluctuates between the values of the control flux Φ_s and saturation flux Φ_{sg}.



Solid curve: Equation (16)
Dashed curve: Equation (17)

Fig.7: Ratio of voltage amplification with a.c. output V_u to voltage amplification with d.c. output V_{uo} as a function of the quotient V_{uo}/m for the magnetic amplifier with self-saturation.

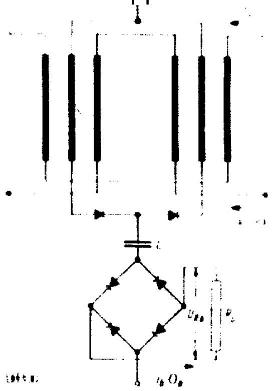


Fig.8: Resonance amplifier with parallel reactor connection and self-saturation. The capacitor C forms a series resonance circuit with the amplifier reactors.

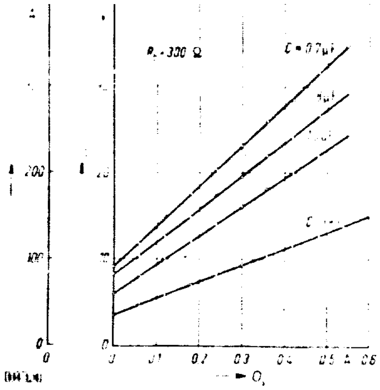


Fig.9: Linear range of working characteristics of the resonance amplifier for different series capacitances.

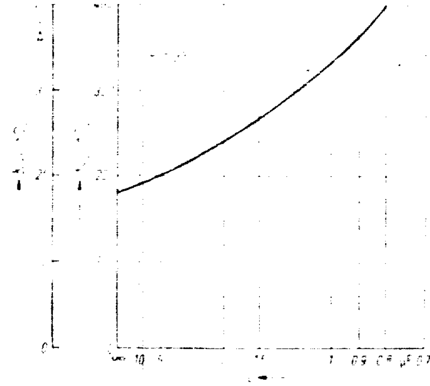


Fig.10: Slope of working characteristics in the linear range plotted against series capacitance.

Θ_s - control flux, Θ_b - working flux, U_{Rb} - voltage drop across load resistance

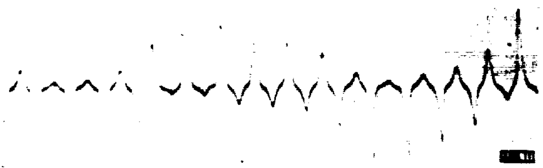


Fig.11: Natural oscillation of the resonance amplifier. The subharmonic frequency f_{sub} is to the working frequency f_b as 1:10.

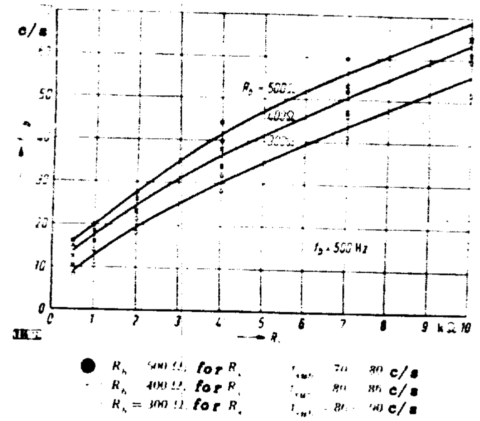


Fig.12: The natural frequency f_{sub} excited in the resonance amplifier against control circuit resistance (R_s) and load resistance (R_b).

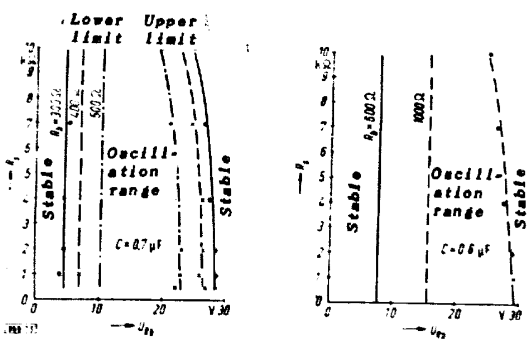


Fig.13: Oscillation range limits of the resonance amplifier for two series capacitances. The amplifier is stable in operation outside the limits.

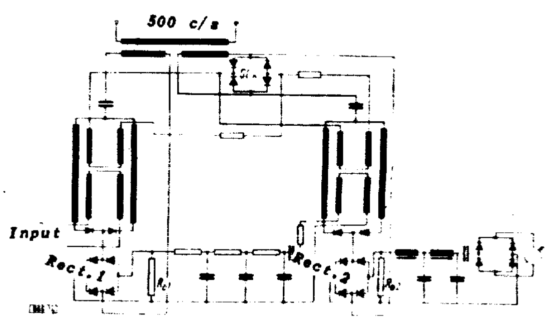


Fig.14: Complete circuit diagram of magnetic zero-voltage amplifier.

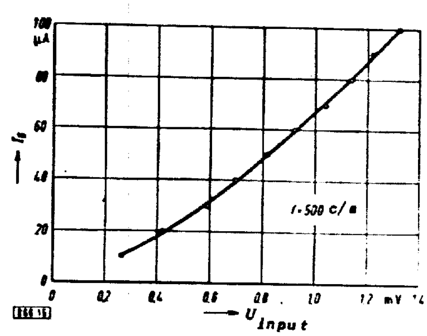


Fig.15: Instrument current I_G plotted against alternating input voltage U_{input} of the two-stage zero voltage amplifier.

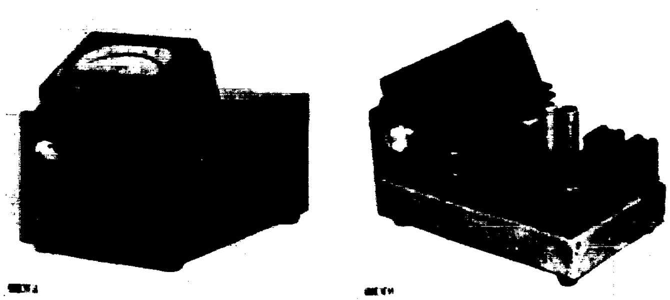


Fig.16: View of final set. (a) Front view, (b) side view with cover removed. Between the ring core reactors are the rectifiers.