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Editorial

The Temperature Coefficient of Inductance

THE maintenance and measurement of radio frequencies to an ever-increasing degree of accuracy has compelled attention during the last few years to the variation of the inductance of coils with variation of temperature, and to the design of coils in which this variation is reduced to a minimum. Readers will be familiar with the pioneer work of Mr. W. H. F. Griffiths in this connection, the results of which have led to the design and manufacture of coils with a greatly increased constancy of inductance. The principle involved is simple; an increase of temperature causes both the diameter and the length to increase; an increase of diameter causes an increase of inductance, whereas an increase in coil-length causes a decrease of inductance; if the materials and construction are such that one counterbalances the other, the inductance will be independent of temperature.

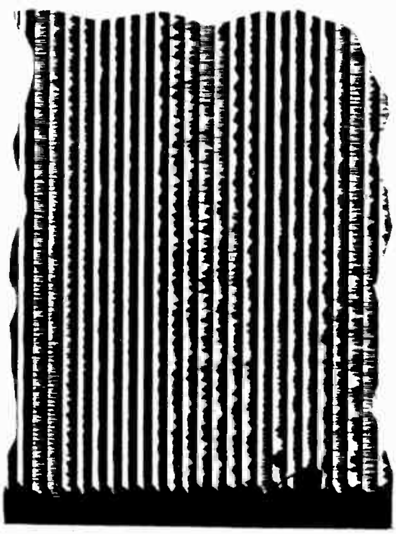
On page 650 of this number, we publish a letter in which it is stated that the measured value of the change of inductance for a given change of temperature does not agree with the value calculated from the change of dimensions, and putting forward what is virtually skin-effect as the cause. That skin-effect, or rather, a variation of skin-effect, causes a change in the effective inductance as well as in the effective resistance is well known, but whether its variation is sufficient to give results of the magnitude quoted by Prof. Groszkowski is a matter for investigation. We use the term skin-effect in its wider sense, embracing not only the concentration of current towards the surface, but also its asymmetrical distribution over the surface, especially, in the case of a coil, its crowding towards the inner surface. In the case of a straight wire of circular section the matter lends itself to relatively simple calculation; as the frequency is raised the current is confined more and more to the surface with the result that the flux is excluded from the copper, and the term in the expression for the inductance which arises from the internal flux gradually disappears. For a given frequency the depth of penetration of the current increases with the specific resistance; therefore, as the temperature increases, the current penetrates further into the conductor and there is an increased flux in the conductor. The effect will be small, but when one is looking for parts in a million it may not be negligible.

When the frequency is so high that the penetration into the conductor is relatively small, it can be shown that that part of the inductance which is due to the internal flux is proportional to $\sqrt{(\mu\rho/f)}$*. The higher the frequency the less it becomes, until at extremely high frequencies no flux penetrates the conductor and there is therefore no contribution from this source to the total inductance. The important point for our present purpose is that this internal inductance varies as the square root of the specific resistance, due, of course, to the change in penetration. For copper ρ increases 4 parts in 1,000 per degree centigrade and the internal inductance would therefore increase 2 parts in 1,000. What effect this will have on the total inductance depends entirely on the proportion of the total inductance represented by the internal inductance at the given frequency. If it amounted to 1 per cent. the variation of inductance would be about 20 parts in a million per degree centigrade, and this is of the same order as the experimental results quoted by Prof. Groszkowski. One per cent. appears a high value at which to estimate the internal component of the total inductance at very high frequencies, but one cannot be more precise without going into the matter more fully and that is not possible without knowing, at least, approximately, the frequency at which the measurements were made. At low frequency a long solenoid 8 cm diameter, closely wound with wire 1 mm square, has an internal inductance of about 1.6 per cent. Such a coil can be calculated for any frequency, but Prof. Groszkowski refers to the danger of applying such calculations based on long idealised solenoids to short coils such as he employed. G. W. O. H.

* See "The Application of Telephone Transmission Formulae to Skin-effect Problems." *Journ. I.E.E.*, 54, p. 473, 1916.

The Paper Record: A Practical Scheme

A new system of recording sound on paper was recently demonstrated in London. The illustrations show a sample of the paper record and the reproducing machine known as the "Fotoliptofono." The equipment is the product of a South American firm with an agency in London.



Modifications of the Push-Pull Output Stage

Part II.*

By K. A. Macfadyen, M.Sc.

(From the Staff of the Research Laboratories of The General Electric Company, Limited, Wembley)

SUMMARY.—With the help of the graphical methods discussed in Part I (October, 1935, issue), an analysis of the familiar "Class B" amplifier is given. A new driver circuit for push-pull output stages which employ the grid-current region during the louder passages of programme-matter is described and analysed graphically. A new system of symbols to represent the various types of output stage is proposed.

1. Summary of the Results of Part I

IT was explained in Part I that the increasing use of certain types of push-pull output stage to which simple theory could not be applied (*e.g.* the "quiescent" push-pull stage) had made it necessary to resort to graphical methods of analysis. It was shown how, by this means, a very complete idea of the functioning of these circuits could be gained. The method consisted, briefly, in deriving a new set of characteristics for the pair of valves, called "composite characteristics," upon which a straight load-line could be drawn to represent the case of a resistive load. By geometric means, the course of the load-lines representing the behaviour of the individual valves could be deduced. These lines, it was shown, were almost straight when the anode-to-anode load-resistance was high and the operating-point about normal for the type of valve employed, but became increasingly curved as the grid-bias was increased in value or the A.C. load-resistance was reduced.

The only types of output-stage considered up to the present have been those in which the grid-potential is always negative with respect to the filament-potential. In the following sections we shall consider two commonly used systems in which the grid-potential is caused to become positive for a part of each cycle. The first of these is what we shall call the "Class B" amplifier, in which the mean grid-potential is zero or, at any rate, only slightly negative. The valves employed are high-impedance triodes and are worked in such a way that grid-current flows in each triode for about a half of each cycle. This type is exemplified by the B21 valve in which a pair of triodes are mounted in the same bulb.

The other kind of output stage in common use is that in which low-impedance valves with negatively biased grids are fitted. The application of a signal-voltage does not cause any flow of grid-current until the valves are giving a considerable output. This system can be applied to single-valve operation, but we shall consider its application to the push-pull stage in greater detail.

2. The "Class B" Amplifier

Since this type of amplifier always works with a relatively small undisturbed anode-current, we shall refer once more to the "quiescent" state described in Part I. When the value of the bias-potential is increased so that the quiescent state is reached, the composite characteristics, over a large part of their length, coincide with the characteristics of the individual valves. Also, the curved load-lines become identical, over most of their "operative" portion, with the straight load-line AC (Fig. 1) drawn through the point on the E_a axis representing the H.T. voltage E_s. Fig. 1 reproduces the upper half of the composite diagram for the quiescent case, the lower half being identical with it but rotated through 180°. It was shown in Part I that the slope of AC corresponded to a

* MS. accepted by the Editor, October, 1935.

resistance of one-quarter of the anode-to-anode load, but it will easily be seen that AC also represents the load imposed on each valve during its "operative" half-

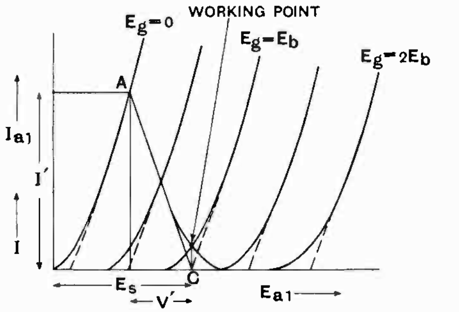


Fig. 1.—*Upper half of composite diagram for the "quiescent" case. The five composite curves (dotted) represent signal-voltages (e) equal to* E_b, $\frac{1}{2}E_b$, 0, $-\frac{1}{2}E_b$ *and* E_b *respectively.*

E_{g1} = *Instantaneous grid potential of valve No.* 1.
E_b = *Bias potential.*
I = *Effective primary current (see Part I).*
E_{a1} = *Anode potential* } *Valve* 1.
I_{a1} = *Anode current* }

cycle. Referring to Fig. 1, which explains the symbols V' and I', we have these approximate results :—

Average full-load feed-current, per pair of valves	$\frac{2}{\pi}I'$
Maximum power output ..	$\frac{1}{2}V'I'$
Power supplied by H.T. unit..	$\frac{2}{\pi}I'E_s$
(E_s is the H.T. supply-voltage) Efficiency	$\frac{\pi V'}{4E_s}$

The cases giving rise to these results have consisted of pairs of triodes or pentodes arranged in the conventional push-pull circuit, but, subject to a small correction, the reasoning is also valid for the so-called "Class B" amplifier. In this case, the fact that appreciable power is dissipated by the grids of the valves makes it necessary to provide a fairly powerful "driver-valve" in the preceding stage, associated with an efficient coupling-transformer. An output-transformer is, of course, necessary also. We shall consider the analysis of the problem with a view to arriving at correct designs for these two transformers.

Fig. 2 shows the $I_a - E_a$ characteristics of a "Class B" triode. This diagram, which we shall suppose to have been constructed from static measurements on the valve, forms the starting-point of our calculations. The composite characteristics should first be added to the diagram by the method given in Part I. These are indicated in Fig. 2 by heavy dotted lines, but it will be noticed that, as in Fig. 1, they coincide for the most part with the ordinary I_a—E_a curves. We assume that the H.T. voltage E_s and the grid bias-potential E_g are known. It is next necessary to determine the optimum load-resistance, *i.e.* the most favourable slope for the load-line AC (Fig. 2). Here we are met with a difficulty, for there is no "boundary-curve" as there was in the Q.P.P. case, in which the load-line always ended at the curve $E_{g1} = 0$. There are, in fact, two unknown quantities, the slope and the length of the load-line AC, for it is not certain to what positive potential the grids may be driven. This, of course, depends on the driver-valve, which must next be examined.

It is supposed that, by means of the usual measurements, it has been found that this is capable of supplying a power W_d (usually expressed in milliwatts) to the pair of output

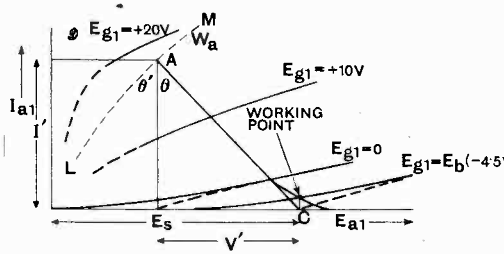


Fig. 2.—*Characteristics of "Class B" Triode.*

valves. Then, assuming the grid circuits to act as linear conductors in one direction,

$$\frac{1}{2} I'_{g1} (E'_{g1} - E_b) = W_d$$

since the peak voltage across each half of the secondary of the coupling transformer is the algebraic difference between E'_{g1}, the peak positive grid-potential, and E_b, the bias-potential (usually negative). W_d is the power being fed into both grids.

It will be shown later that the assumption of linearity in the grid-circuit does not affect the final result even though this circuit does not in fact behave as a linear conductor.

The figure W_d is introduced into the I_a—E_a diagram in the following manner :—

When the measurements of anode-current are being made from which the diagram is constructed, the values of the corresponding grid-currents are simultaneously noted. The diagram can then be covered with a large number of values of the product $\frac{1}{2}\ I_{g1}$ $(E_{g1} - E_b)$. The meaning of each of these figures will be that, if the load-line ends at one of them, the figure will then show the power-input required to cause the representative point to traverse the full length of this load-line. It is possible, by interpolation, to draw a contour, LM, passing through all points characterised by the input power W_d. The load-line, then, must

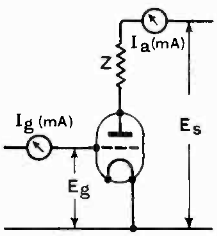


Fig. 3.—*Circuit for investigating variation of grid-current along load-line.*

end on this contour, which, in consequence, assumes the rôle of a " boundary-curve " to which the " equal-slope " theorem (see Part I) can be applied. By making the angles θ, θ' equal, a unique load line is obtained representing the maximum possible output to be secured from the given valves. The anode-to-anode load is four times the resistance represented by the slope of this load-line. As in a previous case, if the load-line lies in non-linear parts of the diagram it may be necessary to revise some of the original data. If the final load-line represents a resistance Z, and the speaker-impedance is R, then the ratio of the output transformer should be :—

$$\frac{\text{Total primary turns}}{\text{Total secondary turns}} = \sqrt{\frac{4Z}{R}}$$

The next matter to be considered is the intervalve transformer. Remembering that only half of the secondary turns are " loaded " at any given moment, it will be seen that the correct ratio $\frac{\text{Total primary turns}}{\text{Total secondary turns}}$ will be $\frac{1}{2}\sqrt{\frac{S}{(E'_{g1} - E_b)/I'_{g1}}}$ since the fraction $\frac{E'_{g1} - E_b}{I'}$ equals the resistance with which the " operative " side of the secondary is closed. S represents the correct load required by the driver valve. The values of E'_{g1} and I'_{g1} are simply the values of grid-voltage and current at A, in Fig. 2, and can presumably be gathered from that figure. A more instructive method, however, and one which will have to be considered in connection with the question of non-linearity, is to construct a grid-voltage—grid-current curve in the following way :—

The " Class B " triode is set up for static testing in the usual way, but a resistance equal to the " load per valve," Z, is inserted in the anode-circuit (Fig. 3). If, now, a curve of grid-current is plotted, using grid-voltage as the independent variable and maintaining the total H.T. voltage at the appropriate value, E_s, this curve will represent the variations of grid-current along that part of the curved load-line where it coincides with the " composite " load-line. Since this coincidence is very rigorous near the end A (Fig. 2), we can be assured that the newly constructed curve can be relied on in this region. The new curve, which will look like Fig. 4 can, of course, be used for an accurate determination of E'_{g1} and I'_{g1} as already described, but its particular interest lies in the fact that it shows the non-linearity of

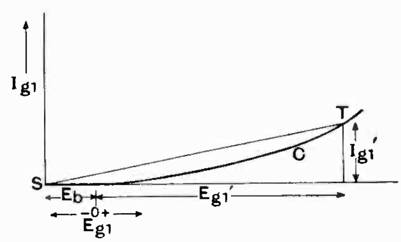


Fig. 4.—*Variation of grid-current along load-line.*

the grid-circuit. Let T represent the limiting point to which the driver-valve will cause the grid-potential to swing.

3. Effects of Non-linearity

The peculiar shape of the curve in Fig. 4 has given rise to some doubt as to the correctness of employing the peak values of grid-current and voltage when assessing the power absorbed by the grids of the Class B triodes. Nevertheless, in performing the calculations just described, these peak values should be used for the following reason. The fact that the driver-valve is loaded with this particular type of conductor results in the formation of a peculiarly shaped load-line on the I_a—E_a characteristics of this valve. Fig. 5 shows the type of load-line caused by a slightly biased Class B output valve. To calculate the actual power given out by the driver under these conditions would be laborious and of theoretical interest only, for it is very easy to see that, apart from small errors due to imperfections in the coupling transformer, the positions of the points A and B (Fig. 5) are decided by the peak value of the grid-current in the output triodes, and by the ratio of the intervalve transformer. In other words, if the transformer-secondary were closed with two rectifying conductors characterised by the straight line ST (Fig. 4) the driver-valve would operate according to the straight load-line AB (Fig. 5). The effect of replacing the conductor ST by the conductor SCT

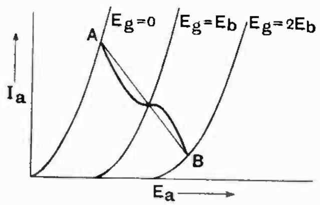


Fig. 5.—*Load-line on characteristic of driver-valve.*

merely alters the shape of the load-line of the driver-valve into the curve indicated in Fig. 5.

Hence if the intervalve transformer is designed on the assumption that it is to be loaded with conductors of the type ST, the points A and B (Fig. 5) will be in what is considered the "optimum" positions when the actual grid circuits (of the type SCT) are connected to it. This is the principle used in section 2.

4. Grid-Current Operation of Power Triodes

This is the name which we shall, for the moment, apply to the second of the two systems described in section 1, for utilising the "positive-grid" region of the valve-characteristic. It possesses the advantage that at fairly low volume-levels, the amplifier functions without the flow of grid-current, so that any form of distortion or extraneous noise accompanying this current does not appear until the reproduced sounds are loud, in which case the distortion is less objectionable than at low volume levels. Another advantage is that existing types

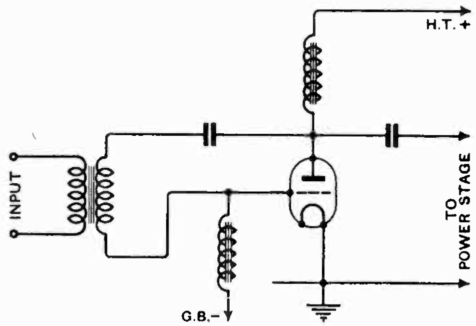


Fig. 6.—*Driver stage with grid-anode input.*

of valve can be used, but it is not advisable to assume that grid-current operation can be applied to any valve, because in some cases the emission of the filament is quickly destroyed by gas disengaged from the grid by electronic bombardment.

The main difficulty in the system consists in providing a preceding stage ("driver" stage) of sufficiently low impedance to avoid grid-current distortion and at the same time capable of providing the large voltage-swing necessary to elicit the full power from the output stage. Baggally* drew attention to this difficulty and showed how it could be surmounted without recourse to an exceedingly powerful driver-stage.

If we confine our attention for the moment to the case of a *single* output valve, the circuit employed by Baggally assumes the form given in Fig. 6. This arrangement reduces the effective internal resistance of the stage to $\frac{R_a}{M+1}$ where R_a is the anode-impedance of the valve and M its amplification factor. This property is also possessed by the "cathode-load" circuit illustrated in Fig. 7. We shall examine this in greater detail, but in passing, the reader's attention is drawn to an ingenious method patented by the Telefunken Gesellschaft† in which the distortion due to grid-current is compensated by a feed-back circuit which causes the

* *W.E. & E.W.*, February, 1933, page 65.
† British Patent No. 407,542.

application of an amplified signal during the grid-current period.

The working of the "cathode-load" driver stage is best investigated graphically. Starting with the I_a—E_a characteristics of the valve (a triode, see Fig. 8) we assume an H.T. supply voltage E_s and a bias-potential E_b, both being measured from the filament-potential (see Fig. 7). As before, the anode-cathode and grid-cathode voltage-displacements will be denoted by E and e respectively. If the signal-voltage applied to the input-terminals is S and the load-voltage L we have

$$S = L + e \qquad \ldots \qquad \ldots \qquad (1)$$

and since $L = - E$ $\qquad \ldots \qquad \ldots \qquad (2)$

we can draw an I_a versus L characteristic for any chosen value of signal voltage (see Fig. 8). Thus if we choose $S = 0$ we must keep e equal to $- L$ and plot the value of the load-current over a suitable range of load-voltages. The latter voltage may conveniently be measured in the negative direction along the E_a axis with the point E_s as the origin, so as to satisfy equation (2).

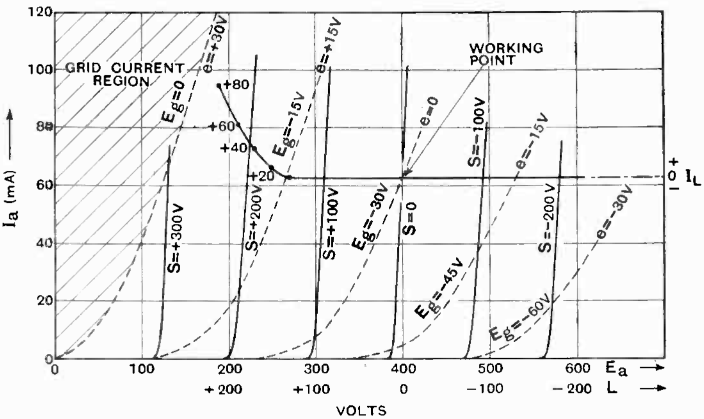


Fig. 8.—*Ordinary characteristics (dotted) and "cathode-load characteristics" of PX25 triode. The load-line due to a directly-coupled, positively-driven DA60 is superimposed.*

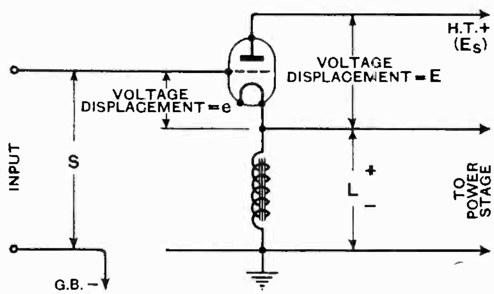


Fig. 7.—*Driver stage with cathode-load, showing instantaneous potentials.*

By choosing other values of signal-voltage S, a whole set of "cathode load" characteristics may be plotted. Reference to Fig. 8 will show that these characteristics are very steep, *i.e.* they represent a very low impedance. A load-line, characteristic of the load connected to the output terminals can be drawn through the working-point, and in this way the exact working-conditions of the driver stage are made clear.

It can easily be proved that the effective impedance and amplification-factor of this circuit are respectively $\dfrac{R_a}{M+1}$ and $\dfrac{M}{M+1}$, but it is most important to notice that the useful portion of the newly-derived set of characteristics is bounded by the old "zero-grid-potential" line and not by one of the new S-lines. Thus the reduction in impedance is attended by limitations though it is interesting to notice that by virtue of the "negative feed-back" of the output-voltage, a reduction in second-harmonic due to curvature of characteristic is to be expected.

By means of this circuit an internal resistance as low as 150 ohms can be realised. This is of great importance in the present problem on account of the fact that the ear is extraordinarily sensitive to the type of distortion, involving harmonics of high order, brought about by the internal impedance of the driver stage. Even when the condition (visualised by Baggally*) of ordinarily weak second and third harmonic is fulfilled, the resulting reproduction is often unsatisfactory by reason of these high harmonics. An exceedingly low internal impedance is essential for good-quality reproduction.

* Baggally, *loc. cit.*

By way of example, a load-line corresponding to the case of a single DA60 triode driven to a peak grid voltage of +80 by means of a cathode-loaded PX25 has been drawn on Fig. 8.

It was once supposed that in order to combine grid-current operation and the push-pull circuit, it would merely be necessary to interpose a suitable double-secondary transformer between the cathode-loaded driver-valve and the output-stage. Unfortunately this plan is quite unsatisfactory for high-grade reproduction On each occasion when grid current ceases to flow in either secondary winding, the oscillatory circuit formed by the transformer leakage-inductance and the stray capacity is thrown into a state of temporary oscillation. The audible effect of this is a very ugly rasping sound. The grid-voltage waveform in these circumstances assumes the appearance of the curve in Fig. 9.

We are thus obliged to adopt a direct coupling system if it is required to run the output valves in push-pull. Thus, a pair of cathode-loaded drivers, one for each output valve, could be used but a better plan is to combine the cathode coupling-chokes as shown in Fig. 10. This has the advantage that both driver valves are brought simultaneously into play on each occasion when grid-current flows in either output-valve

To analyse the way in which this cathode-loaded, push-pull driver-stage works would be difficult if we were not able to avail ourselves of two of the graphical methods already described. Thus, if we combine, by the construction given in Part I (section 2), the cathode-load characteristics of each of the driver-valve, a set of curves is obtained

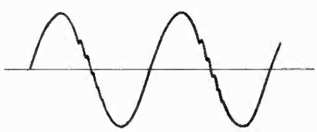


Fig. 9.—*Grid voltage waveform with transformer coupling.*

connecting the cathode voltage (L) with the *effective load-current, i.e.*, the difference of the cathode-currents. It is easier to visualise the centre-tapped choke as the primary of a 2 : 1 (primary : secondary) transformer loaded on its secondary. Then, by the same arguments as were used in Part I, we imagine the split primary to be replaced by a winding having half the number of turns and traversed by this effective load-current, $I_{a1} - I_{a2}$. The construction is illustrated in Fig. 11. It will now be clear that if the secondary load-resistance is Z (or, in the case of the choke, if the load connected across half the choke has resistance Z) we must draw a load-line characteristic of the resistance Z on the composite characteristics. In the case under consideration, this load is the grid-cathode circuit of one of the output valves

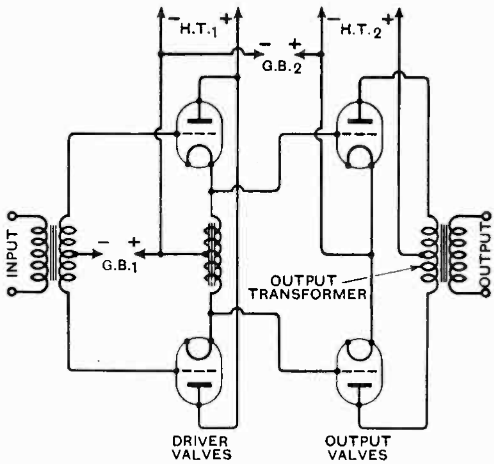


Fig. 10.—*Cathode-loaded push-pull driver stage.*

(only one at a time is in the grid-current condition), hence the grid-voltage—grid-current curve of the output valves—DA100s in this case—is superimposed on Fig. 11. The extraordinary steepness of the combined characteristics accounts for the small degree of grid-current distortion observed when this driver stage is used. The internal resistance is about 70 ohms.

The simple circuit illustrated in Fig. 10 has to be modified slightly for practical application, and it may not be out of place to mention here that the construction of a grid-bias unit with an internal resistance low enough to permit the flow of grid-current without serious voltage-variation is not the only practical difficulty likely to be encountered, though it is in fact a considerable obstacle. However, the difficulties are not insurmountable, and the results—as shown in Fig. 12—are encouraging. The two DA100 triodes, which in the ordinary way, furnish an output of 90 watts, when used with the driver circuit described above, give 180 watts with a negligible degree of harmonic distortion. The development of this form of circuit must, however, be regarded as still in the experimental stage.

5. Suggested New Nomenclature for Output Stages

The foregoing review of the different types of output stage leads us to consider the matter of nomenclature. It is obviously inconvenient, when describing an amplifier, to refer to all its properties by name. Thus, the expression "push-pull output stage utilising the grid-current region and biased to the 'quiescent' condition" would be descriptive, but clumsy. To avoid this, the "Class" system (Class A, Class B, etc.) was used, but owing to misunderstanding, these terms have become meaningless. There is a complete lack of agreement as to whether the "Class" refers to the condition of bias and load, or to the presence of grid-current.

These are two of the three essential properties which should be specified by the symbol in any "shorthand" system of nomenclature. The third essential feature is the form of the circuit, *i.e.* whether a single output valve (including the case of paralleled valves) or the push-pull circuit is in use. It is because the present "Class" system attempts to describe all these three qualities in one symbol that it has proved unsatisfactory. It is hoped that, even in spite of the recently created "Class D" and "Class A-prime" it is not too late to evolve a more acceptable system.

The three salient features referred to above will be described in greater detail, and letters suggested to form a descriptive symbol for any type of power stage.

(1) *Circuit.*

This is almost invariably either of the single-valve type, or the push-pull type. The former includes valves in parallel. If we choose V as the letter to indicate the single-valve stage, it is very easy to remember and has the advantage that when two V's are placed "back to back" they form the letter X, which can conveniently be used for the push-pull type. If desired, a number can be prefixed to indicate the number of valves in parallel, or the number of pairs in the "parallel push-pull" arrangement.

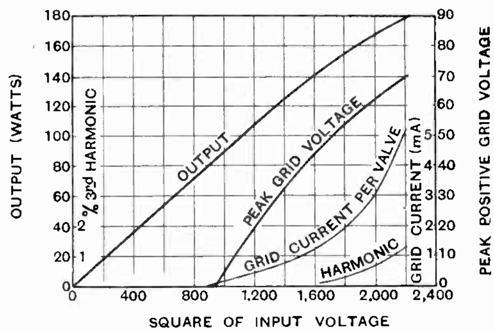


Fig. 12.—*Results of test on DA100 push-pull "positive drive" output stage.*

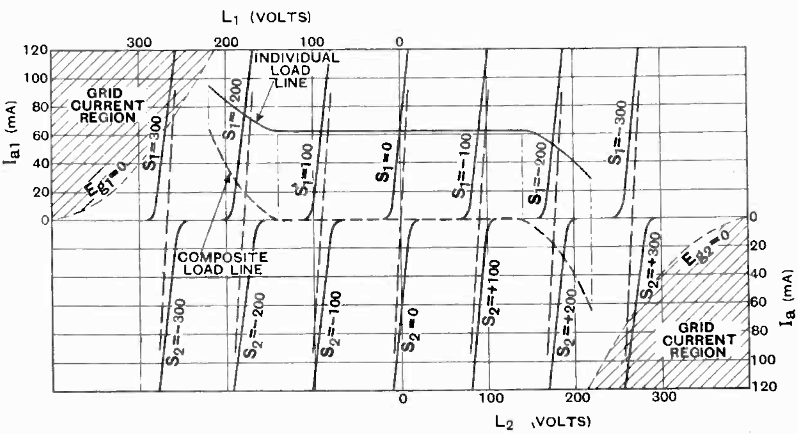


Fig. 11.—*The composite characteristics (dotted) are constructed in this case from the cathode-load characteristics (full lines) of the PX25, derived in Fig. 8. The grid-cathode impedance curve of a DA100 ("composite load line") is superimposed, and the resulting "individual load lines" deduced from this.*

(2) *Conditions of Bias and Load.*

From what was said in Part I it will be remembered that the properties of push-pull circuits were largely determined by the curvature of the load-lines. The constancy of anode current, for instance, depends on this curvature, which is decided by the value of bias, the value of load-resistance and, of course, the type of valve and its operating voltages. The name " quiescent " has been applied to the kind of output-stage in which the magnitude of the H.T. current increases with the loudness of the programme-matter ; in this type, the load-lines are curved. When this current is sensibly constant (apart from the audio-frequency components), the word " pure " has been used to describe the conditions of load, bias, etc. ; here the load-lines are straight.

It is suggested that, in conformity with current practice in this country, the letter *Q* be used to designate the former (quiescent) type. The other type need have no letter : the absence of it will indicate the constancy of anode feed-current and the other properties long associated with the ordinary (" pure ") push-pull system.

The difficulty of deciding the boundary between the quiescent and pure may be removed by reference to a distinction which many designers have had to consider, namely, whether automatic bias can be used, or is precluded by the fluctuations in anode-current. Those amplifiers in which the ordinary automatic-bias circuit is, or can be, used must not be regarded as quiescent.

(3) *Presence of Grid-Current.*

It would seem a very easy matter to specify this property by merely adding the letter *G* to the symbol in those cases where special arrangements are made to utilise the grid-current region of the characteristics. Owing, however, to the great difference between the driver-stage required to feed the kind of output stage described in section 4 from that used before a high-impedance " Class B " valve (see section 2) a further distinction seems necessary. The wide difference between the two types of output valve makes this distinction all the more desirable. A moment's reflection will show that it is the fraction of the total grid-swing (in each output-valve) lying in the grid-current region, which decides the category to which the amplifier belongs. In the case of the " Class B " amplifier (section 2) about one-half of the voltage-swing of each grid takes place in the positive region, while for the other type, utilising power triodes, the fraction is only about ¼.

These suggestions may be summarised as follows :—

First letter	Single Valve	V
	Pair of valves in push-pull	X

Prefix the number of *valves* or *pairs* in parallel if more than 1.

Second letter	Constant anode current (" pure ")	—
	Variable anode current (" Quiescent ") ..	Q

Third letter	Negative region only ..	—
	Grid-current region used..	G

Prefix the fraction of the total grid-swing of each valve lying in the positive region.

These letters form a symbol descriptive of the *type of amplifier* and it is suggested that they should be preceded by the word " type " to avoid confusion with the present " Class " system.

The following examples show something of the wide range of the proposed system :—

Description.	Symbol.
Pentode output stage in most of the average commercial radio receiving-sets	Type V.
Push-pull triode output stage of the ordinary type	Type X.
Pair of DA60 triodes in parallel, driven into grid-current as described by W. Baggally[1] ..	Type 2V¼G.
" Class B " amplifier using L21 and B21 valves, as fitted in battery receivers..	Type XQ½G.
" Triple Twin " (2B6) valve (zero bias, single output triode, amplifying both half-cycles)[2]	Type V½G.
180-watt amplifier described in section 4	Type XQ¼G.
" Quiescent push-pull " amplifier (*e.g.*, one using the QP21 double pentode)	Type XQ.
PX25A triodes connected to give 30 watts as described in Part I (the so-called " low loading " condition)	Type XQ.

[1] Baggally *loc. cit.*
[2] Stromeyer, *Proc. I.R.E.* 20, p. 1149 (1932.)

In many cases (as, for instance, where the type of valve is specified) it may be convenient to omit the fraction in the symbol. The third example given above would then become " Type 2VG," the fourth " Type XQG " and so on.

An Improved Carrier Interference Eliminator*

By W. Baggally

IN a previous communication (*Wireless Engineer*, July, 1932) the writer described how the Campbell sifter circuit could be so designed as to attenuate strongly the frequency corresponding to a carrier whistle without interfering with the reproduction of frequencies both above and below the whistle frequency.

This circuit suffers from the following disadvantages.

(1) Complete elimination of the whistle is never obtained, owing to the presence of losses in the condenser and mutual induction.

(2) From a commercial standpoint, the arrangement is costly since it involves the use of very good quality coils and condenser in order that the above losses may be kept down to a minimum.

The circuit about to be described does not suffer from the above disadvantages since, as will be shown, the losses are completely cancelled out and may therefore be allowed to become quite large, thus enabling simple and cheap components to be used.

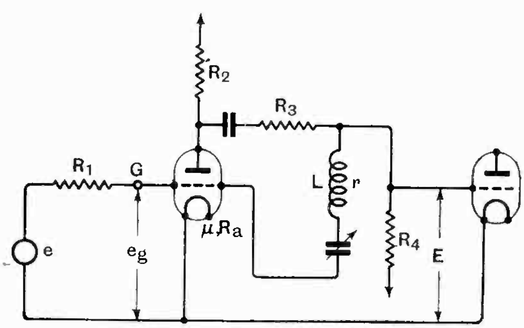


Fig. 1a.

The cancelling of the coil and condenser losses means also that in every case the heterodyne whistle can be completely eliminated.

Fig. 1a shows the circuit arrangement, and it will be seen that the filter is associated with a resistance-coupled audio-frequency stage. The equivalent circuit is shown in Fig. 1b.

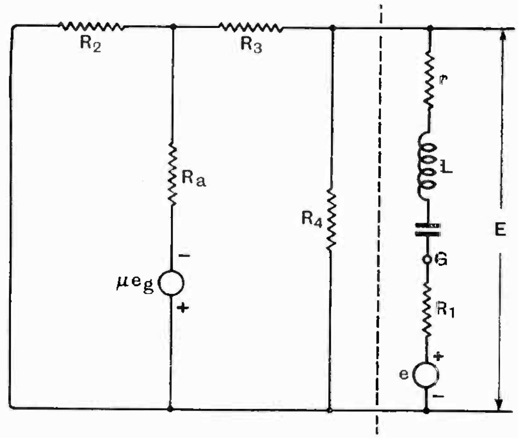


Fig. 1b.

By Thévenin's theorem that part of the circuit lying to the left of the dotted line is equivalent to a generator of constant voltage in series with a resistance. The resistance is made up of R_2 and R_a in parallel, in series with R_3; the whole in parallel with R_4. Calling its value R, we find by application of the ordinary laws for series and parallel resistances

$$R = \frac{s^2 R_4}{s^2 + R_4(R_2 + R_a)} \quad \ldots \quad \ldots \quad (1)$$

wherein

$$s^2 = R_2R_a + R_aR_3 + R_3R_2 \quad \ldots \quad (2)$$

The voltage of the fictitious generator is that across R_4 when everything to the right of the dotted line is disconnected. Its value is found, by applying Kirchhoff's laws or merely by inspection of the circuit, to be

$$\frac{\mu \,.\, e_g \,.\, R_2 \,.\, R_4}{s^2 + R_4(R_2 + R_a)}$$

* MS. accepted by the Editor, July, 1935.

so that we have

$$m = \frac{\mu \,.\, R_2 \,.\, R_4}{s^2 + R_4(R_2 + R_a)} \quad \ldots \quad \ldots \quad (3)$$

It will be observed that m would be the magnification of the stage if the impedance of the coil and condenser branch were infinitely great, which will be approximately fulfilled in practice for frequencies far removed from the resonant frequency.

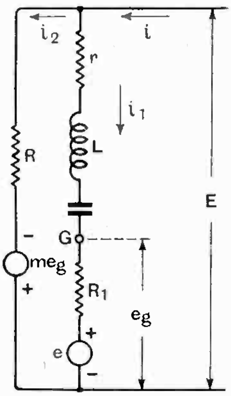
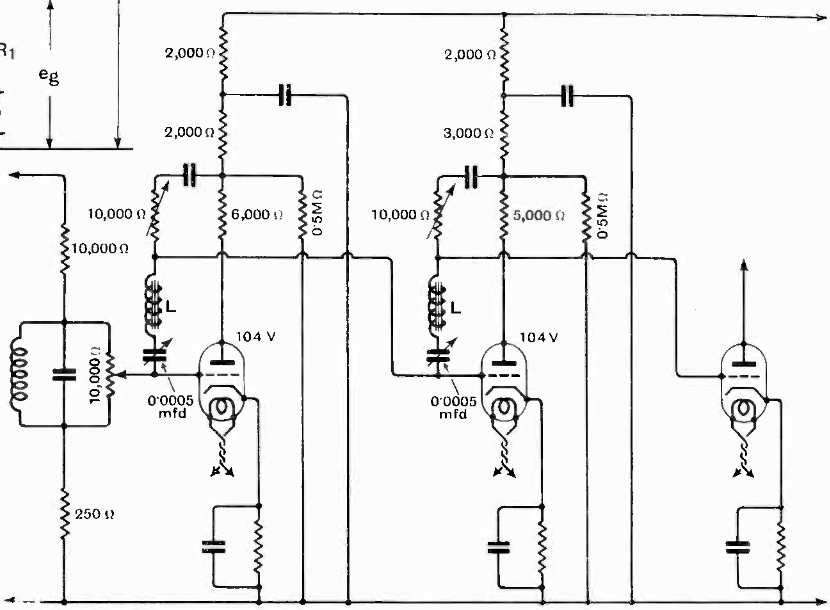


Fig. 1c shows the equivalent circuit transformed by application of (1) and (3). Putting $i = 0$, inspection of this circuit

(*Left*) Fig. 1c.

(*Below*) Fig. 2.



leads to the following results:

$$i_1 = -\,i_2$$

$$e_g(m + 1) = i_2(R + r + jx) \quad \ldots \quad (4)$$

$$e_g = e - i_2R_1 \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

$$E = e - i_2(R_1 + r + jx) \quad \ldots \quad \ldots \quad (6)$$

wherein x is the reactance of the coil and condenser in series: and from these last, putting $E/e = n$ we have

$$n = \frac{R - mr - jmx}{(m + 1)R_1 + R + r + jx} \quad \ldots \quad (7)$$

In (7) let us put

$$R = mr \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

giving

$$n = \frac{-\,jmx}{(m + 1)(R_1 + r) + jx} \quad \ldots \quad (9)$$

the scaler form of which is

$$|\,n\,| = \frac{mx}{\sqrt{(m + 1)^2(R_1 + r)^2 + x^2}} \quad \ldots \quad (10)$$

The condition expressed in (8) is equivalent to cancelling the resistance of the coil and condenser out of the numerator of (7), and at resonance the reactances also cancel, so that we have $x = 0$ and therefore $n = 0$. The frequency corresponding to $x = 0$ does not, therefore appear in the output, and if this frequency corresponds to that of a heterodyne whistle, complete suppression of the whistle is secured.

By a mathematical process similar to that given in the appendix to the above-mentioned paper describing the original form of filter, it is found that the width of the band of frequencies within which the attenuation is greater than three decibels is given by

$$B = \frac{(m + 1)(R_1 + r)}{2\pi L} \quad \ldots \quad \ldots \quad (11)$$

where B is the band width in cycles per second and L is the inductance of the coil in henries.

It is in practice frequently desirable to operate two of these filters simultaneously so as to eliminate the heterodyne notes produced by the two transmitters occupying adjacent channels to the wanted transmitter, one above it and one below it in frequency.

Hence it may be required to connect two filters, the filters being associated with successive stages of the amplifier in such a way that the output of the first feeds into the input of the second.

To determine the performance of the entire system it is necessary to know the input and output impedances of the filters. We assume for simplicity that the terminating impedances are purely resistive; R_4 will consist of the grid leak in parallel with the input resistance of the second filter if one is used; in any case the modification in R_4 will be taken care of by assuming the right values for R and m, which we will denote by R' and m'. Referring to Fig. 1c, the current is given by equation (4). Dividing the grid voltage by this we have the input impedance z given by

$$z = \frac{R' + r + jx}{m' + 1} \quad \ldots \quad \ldots \quad (12)$$

which when combined with (8) leads to

$$z = r + jx/(m' + 1) \quad \ldots \quad \ldots \quad (13)$$

which shows that the input impedance is independent of the output terminating impedance except for the effect of the latter in altering m, the effective magnification of the stage.

We proceed to find the output impedance Z as follows. Putting $e = 0$ in Fig. 1c we find

$$i = i_1 + i_2 = \frac{E}{R_1 + r + jx} + \frac{E + m \cdot e_g}{R} \quad (14)$$

also

$$e_g = E \cdot \frac{R_1}{R_1 + r + jx} \quad \ldots \quad \ldots \quad (15)$$

from which we find on applying (8),

$$Z = \frac{mr(R_1 + r + jx)}{(m + 1)(R_1 + r) + jx} \quad \ldots \quad (16)$$

in which m has its original open-circuit value.

These formulae are sufficient to enable a design to be worked out to meet any given set of requirements. The requirements are in general:—band width, input impedance, output impedance, amount of magnification required from the stages incorporating the filters, and considerations of cost of manufacture. They vary so widely from case to case that no useful purpose would be served by stating anything but the broad principles of design as set out above.

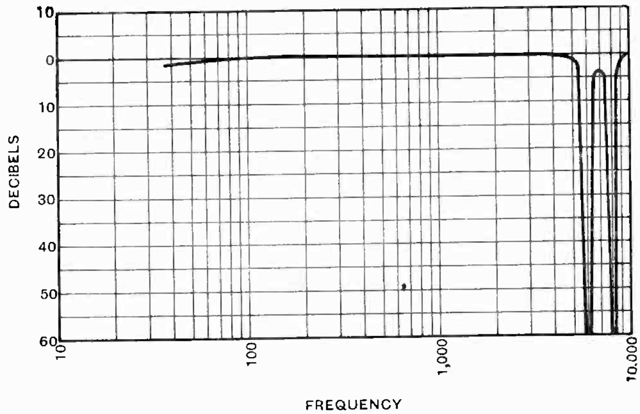


Fig. 3.

As an example of a design which is being used in practice, we may take the arrangement shown in Fig. 2, which illustrates the heterodyne-eliminator circuits of one type of Nuvolion high-fidelity broadcast receiver designed for use in radio relay systems.

The coils marked L are wound to an inductance of about 1.8 henries and D.C resistance of about 2,000 ohms; they employ Ferrocart cores.

The overall response of this receiver with one filter tuned to 6,000 c.p.s. and the other to 8,000 c.p.s. is shown in Fig. 3.

The writer wishes to thank Messrs. Nuvolion, Ltd., on whose behalf this work was undertaken, for their permission to publish this paper.

Correspondence

Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

The Temperature Coefficient of Inductance

To the Editor, The Wireless Engineer

SIR,—The problem of the temperature coefficient (t.c.) of inductance is very important at present in connection with frequency stabilisation. The results of many experiments show that the t.c.—in spite of the theory—is not equal to the coefficient of linear expansion of the metal of which the coil is made. It can even exceed a small multiple of the theoretical value. Thus, for instance, the t.c. of a coil made of copper (in such a way that all dimensions can change freely in accordance with the linear expansion) may exceed 50 parts in 10^6 per degree centigrade (p. in m.p.d.c.), while the coefficient of linear expansion of copper is only of the order of 17 p. in m.p.d.c.*

The investigations of the causes of this discrepancy have so far had little success. Recently Mr. E. B. Moullin† has examined some factors influencing the t.c. of inductance (internal inductance at high frequencies and its temperature coefficient, t.c. of self-capacitance, difference of radial and axial expansion, deforming of turns, etc.).

However, taking into consideration these factors and making necessary corrections by means of available formulae it was possible to justify the discrepancy up to about 50 per cent. but in no case to 200 or 300 per cent.

In the light of the present knowledge it may be suggested that if a valve generator consisting of a coil of varied temperature, a condenser of constant temperature and a valve, is found to have a t.c. of frequency which is considerably greater than half the coefficient of linear expansion of the material of the coil, then the responsibility for this discrepancy is imputed not to the change of inductance of the coil, but to other factors, probably to the valve.

As this accusation of the valve seemed to be unfair, I have tried to acquit the valve, by finding the real culprit.

I am glad to be able to state that—I hope—I have found it at last. The following is a short résumé of my investigations and experiments.

Two coils, of copper and of constantan wire, were made identical (with regard to the dimensions and construction).

In order to eliminate the effects of deformation, the coils were built without formers, the "self-supporting" ones. They consisted of 16 turns of enamelled wire (copper or constantan) 1 mm. thick. The diameter of the coil was about 8 cms., the turns were wound close to each other. The beginning and the end of the wire was used for the binding of the coil (in two diametrically opposite points).

The data of materials used are as follows (in p. in m.p.d.c.):

Temp. coefficient of linear expansion:

copper	16.5	constantan	15.2

Temperature coefficient of resistance:

copper	4,200	constantan	8

(We see that temperature coefficient of linear exp. of copper and of constantan are nearly identical, while their temperature coefficient of resistance differ considerably.)

The oscillatory circuit consisted of the coil under test (placed in a thermostat), of a condenser and of an additional resistance (making constant the dynamic resistance of the circuit) and was excited by means of a dynatron tube working near the limit of oscillation-generation.

The t.c. of inductance calculated on the basis of the measurement of the frequency variation was as follows: for the copper coil 45 p. in m.p.d.c., for the constantan coil 17 p. in m.p.d.c.

Thus, for the copper coil it exceeds considerably the theoretical value (16.5) while for the constantan coil it is very close to this value (15.2). As the resistance of the constantan coil remained almost unchanged during the heating of the coil, we may deduce that—in the case of the copper coil—only the resistance variations are responsible for this discrepancy. Now comes the question: Are these frequency variations produced *via* the valve (as a pure frequency change) or *via* the coil (as an inductance change)?

In order to answer this question a pure resistance (consisting of thin copper wire) was added in series with the circuit; this resistance was changed (by heating) in the same degree as the (internal) resistance of the copper coil under test (the temperature of the coil was then constant).

The observed frequency drift due to the (external) resistance variation was of the order of 1 p. in m.p.d.c. (according to the theory of the dynatron oscillator). Therefore we can answer our question: the resistance variation of the coil is acting not *via* the valve but *via* the coil itself.

Such an effect we will consider as an inductance change and we ought to examine its contribution to the increase of the temperature coefficient of the inductance of the coil.

Probably this effect is caused by the eddy currents in the winding of the coil itself. The mass of metal can be roughly considered as acting as the short circuited secondary of a transformer which diminishes the inductance of the coil; as the temperature increases, the resistivity of the metal increases also and the reaction of the secondary on the primary diminishes: thus we obtain the increase of the inductance.

* I am indebted to Mr. F. M. Colebrook of The N.P.L. who was the first to draw my attention to this fact—several years ago.

† E. B. Moullin, The Temperature Coefficient of Inductance with Special Reference to the Valve Generator. *Proc. Inst. Rad. Eng.*, v. 23, N.1, 1935.

Obviously there exists a relation between the internal inductance at high frequencies and this effect of eddy currents. But if the calculation of this effect with the help of existing formulae‡ does not give the proper value of the correction, it is caused probably by the fact that these formulae relate to a special type of coil (to an infinitely long cylindrical coil with a uniform sheet of h.f. current.

In order to prove that the hypothesis of eddy currents is right, the following experiment was made. To the copper coil was added a number of turns of the same copper wire. These turns were cut asunder and placed close to the winding of the coil but not connected electrically with it. The t.c. then increased considerably: for instance, the addition of 40 per cent. of turns increased the t.c. by about 50 per cent., *i.e.*, from 45 to 76 p. in m.p.d.c. Then, in the same way constantan wire was added. The increase of the t.c. was very small —from 45 to 47 p. Now, the addition of the copper wire to the constantan coil increased the t.c. from 17 to 28 while the increase due to the added constantan wire was only from 17 to 18. I think, therefore, that the influence of the eddy currents, effect on the t.c. of the coil is evident.

Of course, further theoretical as well as experimental investigations are necessary and will be carried out.

JANUSZ GROSZKOWSKI.

State Institute of Telecommunications,
Warsaw, Poland.

‡ J. G. Coffin, *Bull. B. of St.* 2, 1906, p. 275.

"Plane Magnetron Diode"

To the Editor, The Wireless Engineer

SIR,—A paper by Braude (Abstract 3437, October issue) has recently appeared in which it is claimed that there is no return path for the electrons, however great the magnetic field. The equations given by Braude are correct as far as they go, but the interpretation of them cannot be accepted.

Even if we allow for one moment his conclusion (p. 569) that all electrons reach the anode, Fig. 2 seems quite incomprehensible, possibly, on account of his unfortunate choice of units. From his equations (11) for x and y it appears that as ωt goes from 0 to 2π, the change in y is π times the change in x. Thus electrons would escape altogether from the system shown.

Using an experimental plane diode I found that the effect of a magnetic field was to reduce the anode current, due largely to lateral escape of electrons, and to a rise in the space charge density near the cathode. A certain number of electrons appeared also to return to the cathode. Such electrons were allowed for in my paper to the Physical Society 1.1.35. For returning electrons it is no longer possible to use the same equations for the path, but the value of x must be obtained by reversing the sign of $\frac{dx}{dt}$ and integrating between suitable limits. From the value of x thus found y is readily obtained. Actually, however, it is not necessary to obtain x and y since $\frac{dx}{dt}$ simply reverses sign on the return journey. Remembering also $\frac{dy}{dt} = -\left(\frac{e}{m}H\right)x$ it is evident that the return path would be a mirror image of the outward path (the "mirror" being the xz plane passing through $y = y_{max.}$). In the event of the electron having appreciable tangential components of emission velocity the path will be distorted somewhat.

Actually any suggestion that the electrons continue their outward journey, in spite of a magnetic field in excess of the "critical" value, is refuted by considering the third equation on p. 17 of my paper, which may be written $\left(\omega = \frac{e}{m}H,\ e>0.\right)$:—

$$U_x = \pm\sqrt{2\frac{e}{m}V + \left(\frac{dx}{dt}\right)_0^2 - \omega^2x^2}$$

The velocity U_x becomes imaginary when ωx exceeds a certain value, moreover it is seen that U_x may be either positive or negative. It is evident from the above that a critical condition exists for magnetic fields in excess of a certain value, so that the electrons cannot travel more than a certain distance from the cathode, given by

$$x_c = \frac{\sqrt{2\frac{e}{m}V + \left(\frac{dx}{dt}\right)_0^2}}{\omega}$$

At this value of x the electrons reverse their direction in general. Mathematically this means that we must reverse the sign of i_a as well as of $\frac{dx}{dt}$ in the equations

$$i_a = \rho\frac{dx}{dt}$$

$$\frac{dx}{dt} = \frac{4\pi\frac{e}{m}c^2 i_a}{\omega^2}(1 - \cos\omega t)$$

as t passes through the values $2\pi, 4\pi, 6\pi \ldots n\pi$, where n is even.

Thus it appears that repeated cycloidal paths are possible in the space charge limited case, as well as in the case studied by Sir J. J. Thomson where space charges were unimportant.

Cambridge. W. E. BENHAM.

Practical Radio Communication

To the Editor, The Wireless Engineer

SIR,—I would like to bring to your notice a rather important matter relating to my review of "Practical Radio Communication" by Nilson and Hornung, published in the November issue of *The Wireless Engineer*, Whereas I praised the book very highly, a subsequent detailed examination has revealed that such high commendation is not merited.

Although the work is well planned and very complete, and no fault is to be found with the treatment of the fundamental principles of electricity and electric circuits, it is now evident that there are many serious mistakes in the detailed treatment particularly of thermionic valve operation, amplifiers, coupled circuits, and the theory of aerials and wave propagation.

No useful purpose would be served by citing specific cases here, but such errors are so numerous that I feel it a duty to advise you of my now qualified opinion of the book. Perhaps, for the benefit of readers, you may care to publish this letter in the next issue. S. O. PEARSON.

Bromley, Kent.

Electrical Standards for Research and Industry

THIS new catalogue of the instruments made by Messrs. H. W. Sullivan, Ltd., is something more than a mere trade list. It does not of course claim to be a handbook of radio-frequency measurements written with entirely disinterested motives, but it does give a great deal of information about the accuracy now obtainable in various classes of measurement, the qualities required in precision instruments, and the extent to which they have been achieved in instruments of various grades and prices. It is therefore a volume quite as useful to the wireless engineer as many a purely technical book, and may well be considered as such.

The instruments are in the main very highly specialised. Probably no firm in the world has specialised in standards of the highest precision to a greater extent. For each type of instrument the factors conducive to high accuracy of reading and adjustment, low temperature-coefficient, and great permanence are exhaustively discussed. The enormous accuracy now attained in the measurement of frequency has made the question of temperature-coefficient much more important than it used to be, and the attention devoted to this point is one of the outstanding features of the book. There are the Sullivan-Griffiths inductance standards with temperature compensation obtained by constructing the frame of materials with suitably proportioned coefficients of expansion: mica condensers with a temperature coefficient less than 1 part in 100 000 per degree F; and variable air condensers with a temperature coefficient of the same order. These components form essential parts of the precision wavemeters and the Ryall-Sullivan beat-tone oscillator, instruments of outstanding quality which, like the primary frequency standards, have evidently been built more or less regardless of expense, a practice which is probably economical in the long run to the large organisations who can afford it. Those who are forced to work on a less lavish scale will probably be more interested in the instruments of the "second grade" which have recently been introduced. These include inductance standards, fixed mica condensers and variable air condensers made on the lines of the high precision apparatus, but designed for economical production. They are extremely convenient for measurements requiring something distinctly better than the ordinary mass-produced component, but limited in price to one or two pounds.

A very interesting recent development is the dual-range variable condenser. These condensers have two ranges, say 500 to 1 000, and 1 000 to 2 000, and the scale-law (whether linear in capacity, frequency or wave-length, or logarithmic) is the same for both ranges. The principle of construction is not explained, but those who are sufficiently inquisitive will be able to puzzle it out from the illustrations. The reviewer spent a good half-hour on the job.

The volume contains a great deal of useful technical information concerning such matters as the self-capacities of coils, the power factors and decrements of condensers and coils, and the order of accuracy attainable in various measurements. For the most part these matters are not controversial but an estimate of the permanence of the inductance standards at 1 part in 100 000 over three years is truly staggering. An accuracy ten times worse would be impressive, and by no means easy to verify. Is it possible that reason was tempered by enthusiasm at this point? Such things are not entirely unknown in catalogues for the radio industry.

A word should certainly be said in praise of the format of this volume. It marks a complete break from previous tradition. There are no loose leaves; the size of page is a very convenient one (that of *The Wireless Engineer*) and in general the prices are clearly stated in conjunction with the specification of the instrument. As usual it is fully illustrated. The result is that the catalogue is by far the most convenient one that the firm has issued and may be commended to all who are interested in electrical measurements of any kind. L.H.

Mullard Transmitting Valves

AS a result of recent research on the special requirements of valves for short-wave transmitters, Mullard has developed and has now introduced a series of three special valves with ratings of from 75 to 750 watts power output. The largest is designated the TX4–750 and operates at 4,000 volts on the anode with a maximum dissipation of 750 watts.

Similar in construction, but rated at about half the power output, is the TX4–400, which operates also at 4,000 volts but having a maximum anode dissipation of 400 watts.

For transmitters not exceeding 75 watts power output there is a smaller valve described as the TZ1–75. As it would be suitable for use in amateur sets, the following details might prove of more than usual interest. This valve can be operated at 1,500 volts at wavelengths above 45 metres but the anode volts must be reduced at shorter wavelengths, thus 1,200 volts is the maximum at 25 metres, 1,000 at 14 metres, and 800 at 5 metres. A base cap with filament pin is fitted but both anode and grid connections are brought out through seals on the top of the bulb. The filament requires 10 volts at 1.6 amps.

The A.C. resistance is 5,000 ohms and the amplification factor 25, giving a mutual conductance of 5 mA per volt. Other features of the valve are low anode-grid capacity and very sturdy construction. H. B. D.

Some Recent Patents

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each

FREQUENCY MODULATION

Convention date (U.S.A.) 24th January, 1933. No. 426227

A system of frequency modulation is characterised by the use of a pair of push-pull amplifiers for converting the varying amplitude of a microphone current into corresponding phase-displacements of a carrier-wave in an aperiodic circuit.

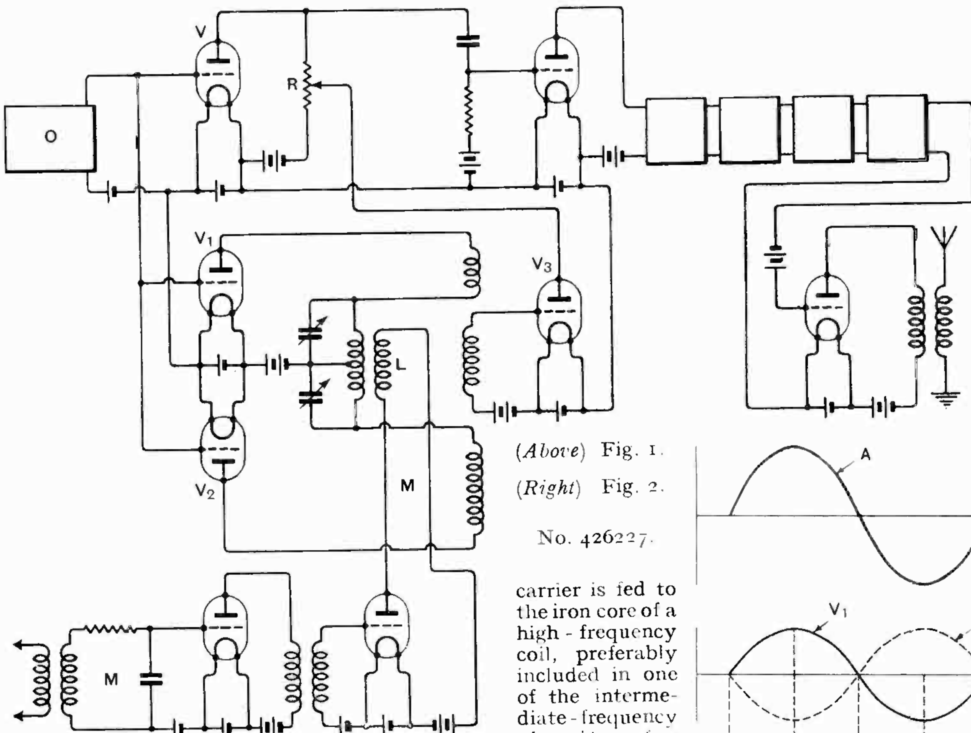
As shown in Fig. 1, oscillations from a constant-frequency source O are fed to an aperiodic amplifier V. A part of the output from the source O is also fed to the push-pull amplifiers $V1$, $V2$, the output through a tapping on the output resistance R of the amplifier V, and the resultant frequency-modulated wave is passed through a series of amplifiers, limiters, and filters to the transmitting aerial.

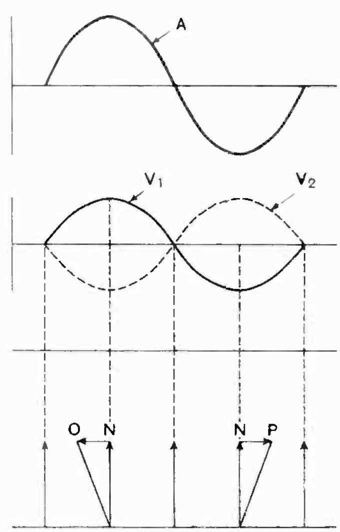
Patent issued to E. H. Armstrong.

AUTOMATIC VOLUME CONTROL

Application date 2nd October, 1933. No. 426347

In order to maintain signal strength at constant volume, without using a valve of the variable-mu type, a rectified current derived from the incoming





(*Above*) Fig. 1.

(*Right*) Fig. 2.

No. 426227.

from which is normally balanced so far as any transfer to the input of amplifier $V3$ is concerned. The effect of applying signals from the microphone line M upsets the existing balance.

A signal voltage in the coil L of the form shown at A in Fig. 2 raises the plate voltage of the amplifier $V1$ and lowers that of the amplifier $V2$, as indicated by the curves so marked. This transfers to the input of valve $V3$ signal-energy which is out of phase with the carrier-wave to an extent indicated by the vectors $N\ O$, $N\ P$ at the bottom of Fig. 2. The out-of-phase signal component from the amplifier $V3$ is superposed on the original carrier frequency

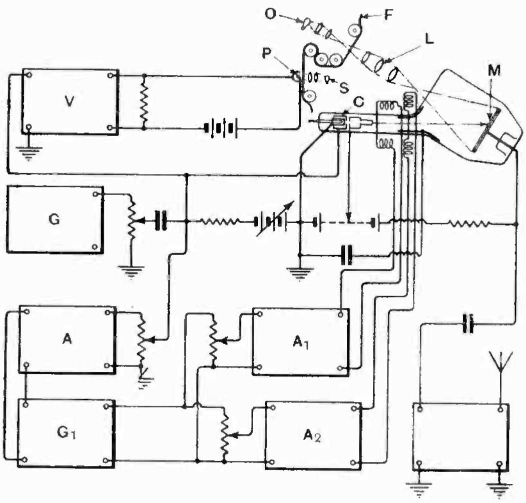
carrier is fed to the iron core of a high-frequency coil, preferably included in one of the intermediate-frequency circuits of a superhet receiver. The D.C. current varies the effective inductance of the coil, and therefore the gain of the set, in the sense necessary to keep the loudspeaker volume uniform. The varying magnetic saturation of the iron core also serves to introduce losses which are beneficial in adjusting the overall selectivity—as well as the sensitivity—of the set in accordance with changes in the strength of the incoming carrier-wave.

Patent issued to Marconi's Wireless Telegraph Co., Ltd.; and A. A. Linsell.

TELEVISION TRANSMITTERS

Application date 2nd October, 1933. No. 426672

Light from a source *O* is projected through a film *F* of the picture to be transmitted and is then focused by lenses *L* on to a photo-electric "cell" electrode *M* in a cathode-ray tube. The



No. 426672.

resulting charges set up on the P.E. electrode are scanned by the stream from the gun part of the tube. Simultaneously the intensity of the cathode-ray stream is varied at carrier frequency by oscillations fed from a generator *G* on to the control grid *C* of the tube, so that the resulting picture-signals fed to the aerial are modulated at that frequency.

Line and frame scanning frequencies are applied to the control grid from a generator *G1* through an amplifier *A* so that the outgoing signals include these frequencies, too. The same scanning frequencies are applied locally to control the electron stream through the tube from the generator *G1* via separate amplifiers *A1*, *A2*.

The energy level of the transmitted signals is varied in accordance with the light-and-shade values of the film picture by means of an auxiliary photo-electric cell *P* receiving light from a lamp *S* and feeding corresponding control voltages to the grid *C* of the cathode-ray transmitter through a direct-current amplifier *V*.

Patent issued to Marconi's Wireless Telegraph Co., Ltd.

Application date 6th December, 1933. No. 426505

Relates to the type of cathode-ray transmitter in which the picture is first focused on to a mosaic of photo-electric cells, and is then scanned by the electron stream flowing through the tube. In practice it is found that a certain amount of gas is given off by the cathode of the tube, and this in course of time tends to damage the mosaic-cell electrode.

Such damage is avoided by mounting this electrode in a separate tube or vessel from that generating the electron stream. The stream is, however, still indirectly used for scanning. It impinges as usual upon a fluorescent screen, and the resulting spot of light is passed through a lens and so made to traverse the photo-sensitive electrode in the second tube. The capacity charges set up by the photo-electric effect of the picture focused on cell electrode are thus discharged by the action of a ray of light, instead of by the direct impact of the electron stream, and the resulting voltages in the output circuit of the second tube are amplified and passed to the modulator.

Patent issued to W. F. Tedham.

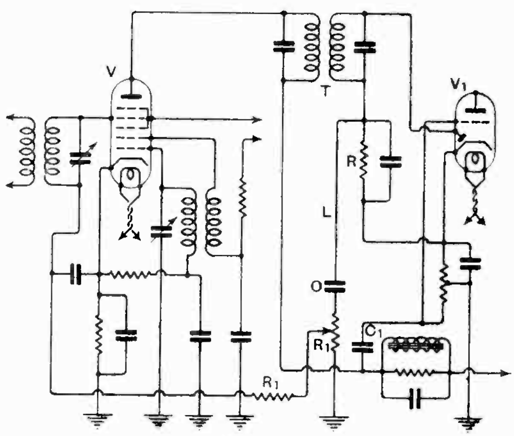
A "REFLEXED" SUPERHET

Application date 12th October, 1933. No. 426802

A valve *V* of the pentagrid-converter type serves as a combined local-oscillator, mixer, and I.F. amplifier. In addition it amplifies the L.F. current "reflexed" back to it from the diode-triode *V1*, which serves primarily as a first detector and subsequently as a second stage of L.F. amplification.

The intermediate-frequency output from the valve *V* passes through a transformer *T* to the diode part of the valve *V1*. The rectified voltage built up across the resistance *R* is then fed back through a lead *L*, condenser *C*, and resistances *R1* to the control grid of the first valve *V*, where it is amplified between that electrode and the anode. The resulting signal passes via the primary winding of the transformer *T* through a condenser *C1* on to the grid of the valve *V1*, which now acts as a low-frequency amplifier.

Patent issued to E. K. Cole, Ltd., and G. Bradfield.



No. 426802.