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Editorial

The Effect of Screening Cans on the Effective Inductance and Resistance of Coils

THIS was the title of our editorial article in March, 1934, which dealt with the work of Kaden, who had published two papers on the subject.* We returned to the subject in July, 1934, in an editorial article entitled "Electromagnetic Screening." Kaden brought the subject within the range of accurate mathematical treatment by assuming that the exact shape of the screen was not important, and that therefore the cylindrical can could be replaced by a spherical screen which he assumed to have a diameter equal to the geometrical mean of the three co-ordinate dimensions of the actual cylindrical can. To calculate the current distribution in the spherical screen he idealised the problem still further by assuming the coil to be replaced by a dipole, that is, by a very small solenoid carrying such a current that the field at a distance would be the same as that of the actual coil. From this current distribution he calculated the effect on the inductance and resistance of the coil and also on the external field, which, if the screening were perfect, would be reduced to zero. Although Kaden evolved relatively simple formulae for the calculation of these quantities, his mathematical treatment was somewhat advanced—it involved Hankel functions—and in the July 1934 editorial article we therefore developed an approximate treatment of an elementary character in which the screening can was replaced by a short-circuited solenoid of the same diameter and radial thickness with a length somewhat greater than that of the can to make up for the absence of ends. This would only differ from a tube in that the current distribution would necessarily be uniform along the length of the cylinder. It was shown, however, that this difference was important and that, although the solenoid gave the same magnetic screening effect as the can, its effect on the resistance and inductance of the coil was much less.

In the current number of the *Journal of the I.E.E.* (Sept., 1940, p. 299) there is published an interesting paper by A. G. Bogle entitled "The effective inductance and resistance of screened coils" in which no reference whatever is made to Kaden's earlier work on the subject. He gives the results of a large number of experiments on various coils placed in copper cylinders of different dimensions. To obtain a calculated basis of comparison he begins by assuming a long solenoid in a long tubular

E.N.T., July, 1933, p. 277.
Zeitschr. f. Hochf : tech., 40 pp. 92-97, 1932.

screen, the eddy currents in which will produce a uniform magnetic field in approximate opposition to that of the solenoid. If the coil is very close to the screen Bogle points out that the relation between the currents in the coil and screen approximates to that which would exist between a plane band of wires at a short distance from an infinite conducting plane. Kaden had shown that if L_0 is the unscreened, and L the screened inductance of the coil, the effect of the screening could be represented by the formula

$$\frac{L_0 - L}{L_0} = \frac{2}{3} \cdot \frac{V_c}{V_s} \cdot \frac{\alpha}{K}$$

where V_c and V_s are the volumes of the coil and sphere respectively, K is the ratio of the inductance L_0 to the value calculated from the formula for very long solenoids, and α is a factor depending on the frequency and on the dimensions, permeability, and resistivity of the screen. For the usual non-magnetic screens α is nearly unity. Bogle's simple assumption of a very long solenoid and screen leads at once to the formula $\frac{L_0 - L}{L_0} = \frac{r_c^2}{r_s^2}$ which has to be multiplied by a factor $1/(1 + x^2)$ to allow for the resistivity, etc. of the screen which factor, like Kaden's α, is very nearly unity in most practical cases. r_c and r_s are the radii of the coil and screen respectively. It may be noted that for coils having a length equal to the diameter, the value of K is 0.7, so that for such coils Kaden's formula reduces to

$$\frac{2}{3} \cdot \frac{\pi r_c^2 l_c}{\frac{4}{3}\pi r_s^2 \frac{l_s}{2}} \cdot \frac{1}{0.7} = \frac{1}{0.7} \frac{r_c^2}{r_s^2} \cdot \frac{l_c}{l_s}$$

which agrees exactly with Bogle's formula if the length of the coil is 0.7 of that of the screen. Bogle's formula was based, however, on the assumption of very long coils and screens and he made a number of experiments to determine how it should be modified in the case of actual coils and screens. As one would expect, in all actual cases the decrease of inductance due to the screen was somewhat less than that calculated for a very long coil and screen. Bogle found that if the tubular screen was not less than twice the length of the coil, adding end plates had very little effect. He also found that the length l'_s of the screening tube beyond which any increase in length had practically no effect on the inductance (which he regarded as a criterion of complete screening), could be expressed by the formula

$$\left(\frac{l'_s}{l_c} - 1\right)\left(\frac{l_c}{r_c} - 0.38\right) = \frac{r_s}{r_c} - 0.84.$$

If the length of the screening tube was not less than this value it was found that

$$\frac{L_0 - L}{L_0} = \frac{l_c/g}{l_c/g + 1.55} \cdot \left(\frac{r_c}{r_s}\right)^2$$

where $g = r_s - r_c$, that is, the gap between the coil and the screen. Adding end plates to the screen made little difference if the screen length exceeded the coil length by $2g$, so that the end gaps were at least equal to the radial gap.

We may compare Bogle's formula with that of Kaden by applying both to an example. If the coil is 3.5 cm. diameter and 3.5 cm. long (for which $K = 0.7$) and the screening can is 7.5 cm. diameter and 7.5 cm. long, the volume V_c of the coil is 34 cm.^{3} and the volume V_s of a sphere of 7.5 cm. diameter is 220 cm.^{3}; hence, taking α as 1, we have

$$\frac{L_0 - L}{L_0} = \frac{2}{3} \cdot \frac{34}{220} \cdot \frac{1}{0.7} = 0.15$$

In Bogle's formula $l_c/g = 3.5/2 = 1.75$

and $\frac{L_0 - L}{L_0} = \frac{1.75}{1.75 + 1.55} \times \left(\frac{3.5}{7.5}\right)^2 = 0.115$

Hence, according to Kaden, the inductance would be reduced to 85 per cent. of its unscreened value, whereas according to Bogle it would be reduced to 88.5 per cent., which is perhaps as close agreement as one could expect.

Effect on Resistance

Kaden gave two formulae for the added effective resistance of the coil due to the screen. If $t\sqrt{f/\rho}$ is much less than 5,000, where t is the thickness of the screen and ρ its specific resistance, skin effect is negligible and the added resistance is given by the formula

$$R_s = \frac{3}{2\pi} T^2 A^2 \frac{\rho}{t r_s^4}$$

whereas if it is much greater than 5,000, skin effect is not negligible and the formula becomes

$$R_s = \frac{3}{\pi} 10^{-4} T^2 A^2 \frac{\sqrt{f\rho}}{r_s^4}$$

In these formulae A is area of cross-section of the coil, i.e. πr_c^2, T the number of turns, and r_s the radius of the supposedly spherical screen. There will be a region in which either formula may be employed to give an approximate result. The frequency does not enter into the first formula and the thickness t does not appear in the second.

Bogle establishes the formula

$$R_s = \left(\frac{L_0 - L}{L_0}\right)\left(\frac{T}{l_c} \cdot \frac{r_c}{r_s}\right)^2 \times 2\pi r_s \rho \sqrt{\frac{2\pi\omega}{\rho}} \times l_c.$$

This again is based on the assumption of a very long coil in a very long tube, which gives a uniform current distribution along the tube, as in the case of a solenoidal screen, but Bogle then investigates the effect of the finite length and the consequent distribution of current in the screen. He concludes that the above formula gives a sufficiently correct value for practical purposes.

When the frequency is so high that the current in the screen does not penetrate to the outer surface, its equivalent depth of penetration is given by the formula $t = \frac{1}{2\pi}\sqrt{\frac{10^9\rho}{f}}$, where ρ is in ohms per cm. cube*, i.e. the dissipation of power is the same as if the current were uniformly distributed over this depth. Kaden's second formula is obtained from the first by making this substitution for t. To find the relation between these formulæ and that of Bogle, we put as an approximation

$$\frac{L_0 - L}{L_0} = \frac{V_c}{V_s} = \frac{r_c^2 l_c}{0.7 r_s^2 l_s};$$

this reduces Bogle's formula to

$$R_s = \frac{A^2T^2}{r_s^4} \cdot \frac{5.7 r_s}{l_s} \cdot \sqrt{f\rho}$$

In his paper Bogle sometimes expresses the specific resistance in c.g.s. units and sometimes in microhms per cm. cube; in this formula it is in c.g.s. units; if ρ were in ohms per cm. cube the formula would be

$$R_s = \frac{A^2T^2}{r_s^4} \cdot \frac{5.7 r_s}{\sqrt{10}\, l_s} \cdot 10^{-4}\sqrt{f\rho}$$

which would agree exactly with Kaden's formula if

$$\frac{r_s}{l_s} = \frac{3\sqrt{10}}{\pi\, 5.7} = 0.53$$

that is, if the length of the screen were about the same as its diameter, which it would be approximately for a coil such as we assumed when we put $K = 0.7$, i.e. a coil, the length of which is equal to its diameter.

G. W. O. H.

*"Application of telephone transmission formulae to skin-effect problems." *Journ. I.E.E.*, 54, 1916, p. 475.

Index to Abstracts

WITH this, the last issue of the volume, we publish again, as has been our custom in the past, the Index to the Abstracts and References published during the year.

The value of these Abstracts is greatly enhanced by the existence of this index. The type of index is unusual in that instead of each title being alphabetically arranged, the key word in the title is set in heavier type and the title then appears alphabetically in the index under this word. Not only does this method of preparing the index avoid much unnecessary clerical work but it has been accepted by many of our readers as an index system of great convenience for reference.

Phase Focusing in Velocity Modulated Beams*

By W. E. Benham

THE details of the various devices[1, 2, 3, 4] proposed for the utilisation of phase focusing will in the present article be assumed known, though all we need really attend to are fundamental principles.

An interesting graphical method has been proposed by Tombs,[5] and various published analyses include those of Brüche and Recknagel,[6] of Webster,[7] and of Kömpfner[8] and Strachey.[9] Tombs' method seems to be open to criticism; the other authors are not always lucid. The sequel provides an elementary calculation, and curves, showing *where* and *when* bunching occurs.

We imagine as our data, first, a beam of particles of charge e, mass m, travelling with uniform speed u_0 parallel to some fixed direction. The influence of these particles on one another in the radial direction can be calculated by a method due to Watson.[10] Provided, however, that the consequent enlargement of cross-section does not cause the beam to become too large for the holes, or grid-electrodes, through which the beam has to travel, the radial repulsions are of advantage rather than otherwise since the further electrons are apart radially the less important will be the effect of longitudinal mutual repulsions. This is because there will be a smaller chance of any electron in the beam lying on exactly the same line (parallel to the axis of the beam) as another electron just ahead of, or just behind, this electron. Longitudinal mutual repulsions will be neglected in the present paper. Webster[7] has shown that they are usually small; if they are not small in any specified case, such as that of the amplifier with long drift tube, or low voltage detector, then further examination must be made.

Secondly, when our electron beam meets a certain fixed plane at right angles to the beam axis, it finds itself under the influence of an alternating field. Let this field be $E(t)$ volts per cm. acting across a gap of b cm. width. In other words, the beam is under the influence of a modulating field for the time τ necessary for the beam to cover the distance b, which is assumed such that τ is small, compared with the time period $2\pi/\omega$ of the fundamental component[11] of $E(t)$.

Electrons entering the modulating field at instant t_0 leave with velocity u_1 at instant t_1, where, writing $eE = mF$,

$$u_1 = u_0 + \int_0^1 F(t)dt \quad \ldots \quad (1)$$

Later electrons, entering the modulating

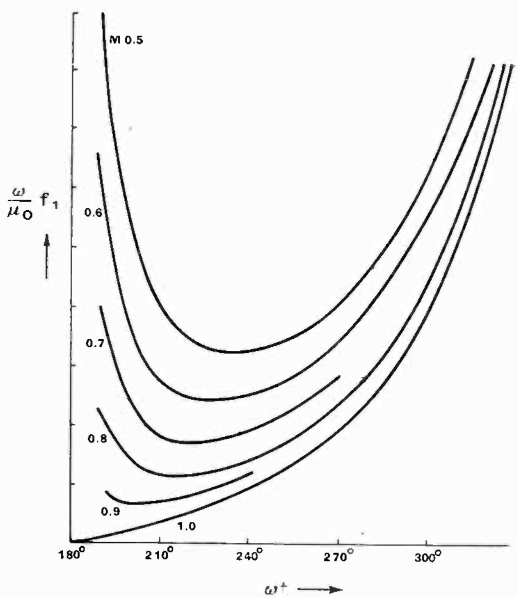

Fig. 1.

field at, say, $t_0 + \Delta t_0$, leave with velocity $u_1 + \Delta u_1$ at instant $t_1 + \Delta t_1$ where

$$u_1 + \Delta u_1 = u_0 + \int_0^1 F(t + \Delta t)dt \quad \ldots \quad (2)$$

We shall suppose that Δt_0 is so small that not only can we neglect such products as $\Delta t_0 \Delta u_1$ and Δt_0^2, but also that $\Delta t_0 = \Delta t_1$ and that Δt_0 is a much smaller interval even

* MS. accepted by the Editor, July, 1940.

than τ. Then, by subtraction

$$\Delta u_1 = \int_0^1 \{F(t + \Delta t_0) - F(t)\}dt = \int_0^1 \Delta t_0 d(F(t))$$

$$= \Delta t_0\{F(t_1) - F(t_0)\} \quad \ldots \quad \ldots \quad (3)$$

Neglecting all mutual repulsions, the second set of electrons catches up to the first in a distance f_1 if the transit time f_1/u_1 of the first satisfies the condition

$$(u_1 + \Delta u_1)(f_1/u_1 - \Delta t_0) = u_1 f_1/u_1$$

whence $$f_1 = u_1^2 \frac{\Delta t_0}{\Delta u_1} = \frac{u_1^2}{F(t_1) - F(t_0)} \quad \ldots \quad (4)$$

Since after leaving the modulating field the two sets of electrons are travelling in a field-free space they will never "bunch" again after passing f_1. If the modulation depth is very small, $u_1^2 = u_0^2 = 2eV_0/m$. Then

$$f_1 = \frac{2V_0}{E(t_1) - E(t_0)} \quad \ldots \quad \ldots \quad (5)$$

The form (5), the equivalent of which was given by Brüche and Recknagel,[6] gains interest by comparison with the well-known Davisson formula for the electron optical focal length of an ideal "cylindrical lens" constituted by a long narrow slit irradiated by electrons on the side where the field is E_1, emerging on the side where the field is E_2; namely[12]

$$f = \frac{2V_0}{E_2 - E_1} \quad \ldots \quad \ldots \quad (6)$$

The focusing here depends, of course, on a spatial discontinuity of field whereas (5) depends, in effect, on a temporal discontinuity. The analogy may, however, be coincidental, since in the case of a circular aperture the coefficient in (6) changes from 2 to 4, owing to the focusing field becoming radial, whereas (5) is, to a fair approximation, unaffected, e.g. by changing over from a grid to an apertured disc. This illustrates the danger of pushing an analogy too far.

Now (5) applies only when the modulating field is small. The calculation for M large is a little intricate.* The results are shown on Fig. 1, M being the modulation depth and $\omega/2\pi$ the frequency of the modulation, assumed sinusoidal. By "modulation" we merely mean the radio-frequency modulation of a direct current beam; speech, or even television frequency, modulation presents no special problems which can be considered as within the scope of the present article. M will thus be the ratio $\hat{E}b/V_0$, where b is the gap across which the field acts, and may be, for example, the distance between the two grids of a "rhumbatron" electrode. When $M \to 0$ the minimum f_1 occurs at $\omega t = 270°$, the general formulg then agreeing with the result of writian $E(t) = E \cos \omega t$ in equation (5).

*For reasons connected with a patent application (No. 9544/40) it has been found necessary, with regret, to omit certain mathematical details from the present article.

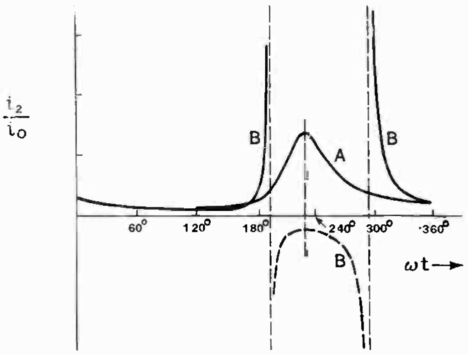

Fig. 2.—Curve A $x = 0.4$ cm. Curve B $x = 1.0$ cm. } $M = 0.5$

Fig. 1 brings out the fact that as M tends towards unity the bunching, as might, for example, be reckoned in terms of the time during which $\omega f_1/u_0$ remains in the neighbourhood of its minimum, deteriorates in a marked manner. The *minimum* values will now be referred to as f_1. These are compared with values for the bunching distance, f_2, as quoted by Strachey[14] in a recent note, in the following Table:

TABLE I

	$M \to 0.25$	0.5	0.8	0.9	1
$\omega M f_1/2u_0$ ..	0.94	0.80	0.47	0.30	0
$\omega M f^2/2u_0$ ..	0.925	0.73	0.34	0.18	0

Now Strachey has calculated the distance down the beam at which the charge passing *per second* is a maximum, whereas my figure corresponds to the distance down the beam at which the *actual charge density* is a maximum. The bunching distance for maximum charge density, f_1, is seen from the table to exceed that for maximum current density, to which Strachey's f_2 may be regarded as applying. The analysis presented by Strachey would appear to be acceptable and will not be repeated here. His symbols make it clear that his focal length corresponds to

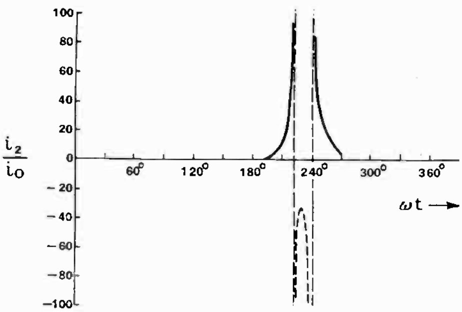

Fig. 3.—$x = 0.5$ *cm.* $M = 0.5$.

maximum *current* density, though this is perhaps less clear from the accompanying discourse. As to which formula must be used, this depends largely on the application, and on the view taken as to the mechanism of the second, or " catcher " rhumbatron, or other "collecting" mechanism. For M values up to 0.5 the formulae differ by less than 10 per cent.

The beam current amplification $\frac{i_2}{i_0}$ according to Strachey's analysis will conform to the curves given here in Figs. 2 and 3. Attention is directed to the duple peaks which occur when the bunching distance x is made greater than f_2, which in this example corresponds to 0.48 cm. These peaks have been predicted on slightly different assumptions by Webster,[7] who finds that the *reciprocal* of the current waveform is trochoidal ; when $x > f_2$ this trochoid has a loop below the axis, hence the asymptotes and discontinuities in the current curves. Note that the scale of Fig. 3 is 10 times contracted as compared with Fig. 2. Strachey's calculation of ωt for maximum i_2 corresponds to the *mathematical* maximum in each case, i.e. to the minimum negative value for that part of the curve which may occur below the ωt axis.

References

[1] A good general description of their design of tube, known as the Klystron, is given by Varian R. F. and Varian, S. F., in the *Journ. of Applied Phys.*, 1939, Vol. 10, No. 5, pp. 321-327. The advantages of this tube are stressed by Wheelock, M. ; *QST*, 1939, Vol. 23, No. 5, p. 37.

[2] For this and other types see Fink, D. G., *Electronics*, April, 1939, Vol. 12, No. 4, p. 9.

[3] Haeff, A. V. and Nergaard, L. S., have just issued further particulars relating to a form of inductive amplifier, described earlier by Haeff (*Electronics*, Feb., 1939, Vol. 12, No. 2, pp. 30-32), which is claimed to yield 10 watts, in a circuit loaded to give 10-Mc/s band width, with a power gain of 10 and an efficiency of 25%. Frequency 500 Mc/s.

[4] Hahn, W. C., and Metcalf, G. F., *Proc. Inst. Rad. Eng.*, 1939, Vol. 27, No. 2, pp. 106-116.

[5] Tombs, D. M., *Wireless Engineer*, Feb., 1940. Vol. 17, No. 197, pp. 54-60.

[6] Brüche, E., and Recknagel, A., *Zeitschr. f. Physik*, Mar., 1938, Vol. 108, Nos. 7 and 8, p. 459.

[7] Webster, D. L., *Journ. of Applied Phys.*, July, 1939, Vol. 10, No. 7, pp. 501-508, *ibid.*, Dec., 1939, Vol. 10, No. 12, pp. 864–872.

[8] *Wireless Engineer*, May, 1940, Vol. 17, No. 200, p. 202.

[9] *Ibid.*, pp. 202, 3.

[10] Watson, E. E., *Phil. Mag.*, April, 1927, pp. 849-853.

[11] With conventional forms of tuned circuit, such as might be used, if desired, to supply the modulating signal, the selectiveness is usually such that harmonics are largely suppressed. With the special forms [due to Hansen, W. W. : see *Journ. of Applied Physics*, 1938, Vol. 9, No. 10, pp. 654-663 ; see also Hansen, W. W., and Richtmyer, R. D., *Journ. of Applied Physics*, 1939, Vol. 10, No. 3, pp. 189-199] known as the Rhumbatron, a whole series of allowed frequencies, harmonics of the fundamental, are present.

[12] Davisson, C. J., and Calbick, C. J., *Phys. Review*, 1931, Vol. 38, p. 585.

[13] Bedford, L. H., *Proc. Phys. Soc.*, 1934, Vol. 46, pp. 882-888.

[14] Strachey, C., *Wireless Engineer*, May, 1940, Vol. 17, No. 20, pp. 202, 203. Kömpfner, R., now agrees with Strachey (*ibid.* p. 202.)

The Industry

LONDEX, LTD., 207, Anerley Road, London, S.E.20, have introduced a new series of magnetically-operated double-acting relays. These are fitted with armatures and pole pieces suitable for AC or DC and are available for excitation voltages from 2 to 500 volts. The consumption is 2 VA. Two types are available, the LQA/Double fitted with a mercury switch, and the LF/Double which may be obtained with a variety of silver or tungsten spring contacts.

Barimar, Ltd., 14/18, Lambs Conduit Street, London, W.C.1, have evolved a new process for welding ferrous and non-ferrous metals such as steel and aluminium. This should prove of value in research work as well as in industrial production.

The Performance of Commutator Inverters*

By L. C. Stenning, B.Sc., Grad. I.E.E.

(Communication from the Staff of the Research Laboratories of The General Electric Company, Limited, England)

SUMMARY.—Inverters which consist essentially of switches which periodically reverse the input direct current have a field of usefulness which has been rapidly widened in the last few years.

This paper deals with the circuit design of commutators and inverters of the vibrator, gas discharge rotary commutator type, and the calculation of the performance of these circuits on the assumption that their transformers do not distort the square waveform of the input alternating voltage and current. The conditions of low flux density, adequate copper current density, etc., under which this is true, and the allied requirements for minimum sparking and good commutation, are briefly discussed.

Given the above assumption, equivalent circuits are deduced from which it is possible to calculate the output, regulation and efficiency near the fully loaded condition.

The scope of the paper is indicated by the following list of sections.

LIST OF SECTIONS.

1. Introduction

DIRECT current can be "inverted" into alternating current of square wave form by a rapidly operated reversing switch. This alternating current can be used for a number of purposes, where the load is of good power factor. In particular, it has been found possible to use it for supplying wireless receivers and transmitters if the square wave A.C. is filtered to remove its radio and audio-frequency components. Very simple filters suffice for most purposes.

There are two slightly different main types of circuit. Type A, shown in Fig. 1, uses a straightforward reversing switch, and obviously causes an alternating voltage of the waveform shown in Fig. 2 to be applied to the transformer primary; whereas Type B uses a single pole switch in conjunction with a centre tapped transformer primary, and has the same effect, opposite half waves being applied to opposite halves of the primary winding. All calculations regarding external effects of the transformer can be carried out on the assumption that opposite half waves are applied to the same half primary winding, i.e. formulae deduced for the Type A circuit can be used for the Type B circuit if the half winding is taken as the primary. The same simplification can be used with the biphase rectifying circuit, which is again of Type B.

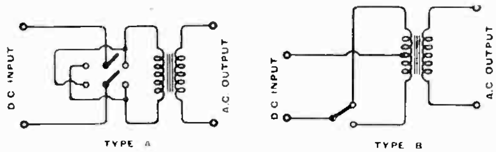

Fig. 1.—*The two types of commutating circuit.*

Type A may be used without a transformer, if an output alternating voltage slightly less than the input direct voltage is required. With Type B, the transformer is indispensable.

There are several means of carrying out this principle illustrated in Figs. 3 and 4. Vibrators which consist of one or more tuned reeds carrying tungsten contacts may be

* MS. accepted by the Editor, July, 1940.

applied to either type up to about 1 kW. output. Fig. 3 illustrates a sample of these with some typical circuits. The reed is maintained in vibration by an electric bell circuit as shown in Fig. 3 (b). Fig. 4 (a)

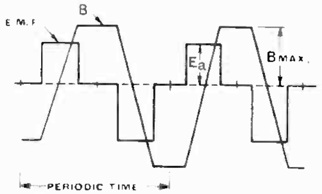

Fig. 2.—*The flux density wave in the transformer (ignoring D.C.).*

gives a commutator circuit belonging to Type A. Fig. 4 (b) shows a gas discharge relay circuit of Type B without the exciting circuit. Hereafter the term commutator will be used generally of any kind of switching means. The term converter is used of any commutator plus transformer arrangement and may include a rectifier and smoothing system.

The condensers across the contacts in Fig. 3 (b) have a value 0.1 to 0.5 μF. and are for the purpose of suppressing sparking when the contacts open. A decrease of sparking when the contacts close is sometimes obtained by putting resistances of the order of 100 ohms in series with these condensers. If the converter supplies a short wave receiver, and the current carried by the contacts is less than about one ampere, it might be worth while to supply it through the sparking condenser foils, as in Fig. 3 (c), one pair of ends being connected to the contacts, and the other pair to the rest of the circuit. One object of this is to suppress interference by connecting the resistive characteristic impedance of the foils across the spark instead of the reactive condenser impedance. Probably more important is the attenuation of upwards of 20 db., provided by the transmission line formed by the foils. In the case of vibrating reed commutators some interference suppression may be secured by adding an extra short circuited winding on the magnet.

2. Factors Governing the Performance of Commutator Type Inverters

The frequency of alternation is usually chosen within the range of 50 to 120 c/s. A low frequency necessitates a larger trans-

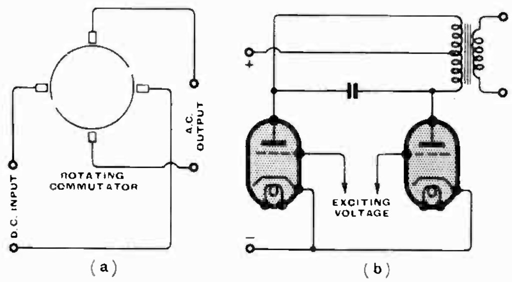

Fig. 4.—(*a*) *Rotating commutator inverter;* (*b*) *Gas discharge relay circuit.*

former and too high a frequency will give trouble due to mechanical difficulties.

The resistance of the contacts is almost

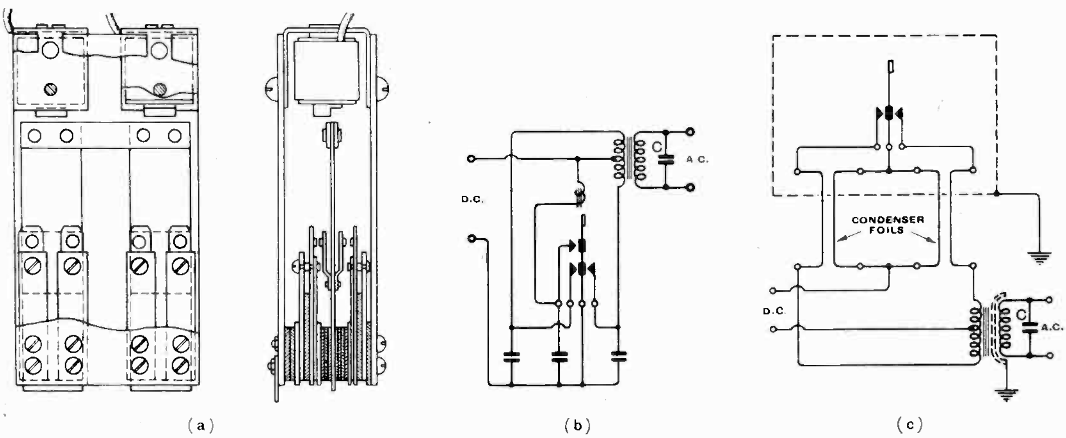

Fig. 3.—(*a*) *Vibrator (double type);* (*b*) *Practical vibrator circuit;* (*c*) *Interference suppression.*

always negligible, being approximately 0.05 ohms per pair. Due to this, the regulation resistance of a vibrator converter is purely limited by the regulation of the transformer and other circuit resistances, and is very good.

Some vibrators, in consequence of the method by which they are driven, have unequal contact times and produce an asymmetrical A.C. wave. Typical figures are 40% of the total periodic time one side; and 45% the other. This will produce an unwanted steady magnetisation of the transformer core. It is not considered that the labour of calculating the effect of a small inequality of contact time would be worth the additional accuracy that would be obtained, having regard to the random asymmetry that occurs in associated rectifiers, transformers and other components. There is no doubt, however, that asymmetry should be avoided, as it causes extra loss.

3. Distortion of Square Wave Voltage

The characteristic feature of the commutator is its inability to supply current during the time its contacts are open. However, if the commutator is feeding a resistive load, no current is required to flow during these open intervals. With respect to such a resistive load the commutator and battery could be replaced by a square wave A.C. generator of open circuit peak voltage equal to that of the battery and of an internal resistance equal to that of the battery, leads and closed commutator contacts. The fact that the actual configuration has an infinite internal impedance during the open contact intervals does not become apparent in this case at all.

If this equivalent generator were to feed a reactive load, current would flow in the zero voltage intervals corresponding to the open circuit intervals of the commutator, and in the case of the real commutator, sparking and transients would occur at make and break. The problem of the elimination of sparking may therefore be treated as a question of ensuring that the commutator works into a resistive impedance. The same necessity is not so apparent in the case of a gas discharge tube inverter, but this also does not run well when the waveform is not nearly square.

Consider the e.m.f. waveform shown in Fig. 2. By the application of Fourier analysis we find the amplitude of the nth harmonic a_n is given by

$$a_n = \frac{4E_a}{n\pi} \cdot \sin\frac{nk\pi}{2}$$

where E_a is the peak voltage of the flat topped wave (equal to the D.C. generator voltage), and k is the fraction of the periodic time during which the e.m.f. has its peak value.

The first few harmonics for $k = 0.8$ and $E_a = 100$ are :—

Order ...	1	3	7	9	11	13	17	19
Amplitude ..	121	24.4	11.3	13.7	10.8	5.1	5.0	6.6

Thus harmonics above the 11th have not more than 6% of the fundamental amplitude. We have, therefore, to consider the behaviour of the impedance into which the commutator works over roughly the above range of frequency.

As a simple example let us take the case of A.C. output into a resistive load r.

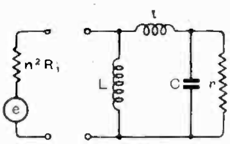

Fig. 5.

The equivalent circuit is given in Fig. 5. C is an anti-sparking condenser usually placed across the secondary as in Fig. 3 (b).

L is the secondary inductance with primary open circuited. l is the leakage inductance with primary short circuited; let f be the frequency of alternation of the commutator.

Considering the highest harmonic (say the 11th), the half section low pass filter formed by l and C should be matched by r and should cut off above $11f$.

$$\text{Thus } C = \frac{2l}{r^2} \text{ and } 11f < \frac{1}{\pi\sqrt{2lC}}.$$

Considering the fundamental harmonic $2\pi fL/1-4\pi^2f^2CL$ must be great compared with r for small input reactance, i.e. small lagging component of the input current, as L, r and C are in parallel at low frequencies.

The ratio l/L depends on the construction of the transformer and on the maximum flux density. C is given by the first equation. L/r depends on construction, flux density and on the current density in the copper, which should be above a certain minimum.

The above theory cannot be considered complete as it does not include the condensers across the contacts, nor is it so easy to apply to the Type B circuit, but it is put forward to account for the circuit values which have been found to work satisfactorily without sparking or arcing. It also discloses possibilities of further improvement, for instance by the addition of *m*-derived matching sections. Both the above conditions require a low flux density in the transformer, and it is usual to limit B_{max} to 7,000 lines per square cm. at 60 c/s.

If these conditions are satisfied the primary and secondary voltages and currents in the transformer will be of substantially square waveform and we may assume that the transformer is perfect, i.e. it adds resistance in the primary and secondary circuits, multiplies voltages and divides currents by its turns ratio and multiplies impedances by the square of its turns ratio.

4. Calculation of the Characteristics of a Converter

(a) *List of Symbols.*

k = the fraction of the total period during which power is being transmitted (taken as twice the shortest half period).

$I_a E_a V_a$ = peak alternating current, e.m.f. and voltage.

$E'_a E''_a$ = peak alternating e.m.f.'s in primary and secondary of the transformer.

$I_d E_d V_d$ = direct current, e.m.f. and voltage.

$I_2 V_2$ = direct current and voltage output.

R_1 = total resistance in primary circuit with switch held to one side.

R_2 = resistance in secondary circuit carrying A.C. or chopped D.C. (including half that of a centre-tapped transformer).

R_3 = resistance in secondary circuit carrying smoothed D.C.

n = turns ratio of transformer (counting half of a centre-tapped winding).

R = overall regulation resistance of the converter.

R_T = smoothed—D.C.—equivalent resistance of the rectifier = e_T/I_2.

e_T = voltage drop across rectifier when passing I_2/k.

E_T = voltage drop across gasfilled rectifier.

W_1 = watts output from the battery less driving power.

W = watts input to core of transformer.

η_1 = primary efficiency (= W/W_1).

x = $4\ WR_1/k\ E^2{}_d$.

(b) *Measurements.*

The proportional contact time of a pair of contacts in motion may be measured by placing them in series with a battery, a resistance and a linear ammeter. If the deflection is i and the deflection with the contacts short circuited is I then $k = i/I$, provided that the resistance has a much higher value than that of the closed contacts. A very convenient way of doing this is to use a deflection type resistance meter with a linear 0-100 scale. The moving contacts are placed across it and the value of k can be read straight off this scale. This method is not easily applicable to a vibrator without a separate driving contact. An oscillograph method must be used in this case. In the case of a vibrator which made contact 40% of the time on one side and 45% on the other side, we would take the fractional contact time as 40% + 40%, i.e. = 0.8.

In the case of a converter with D.C. output, it is necessary to examine the circuit carefully to find the effective value of k of the A.C. waveform.

If the rectifier is a pair of vacuum diodes, the value is that of the primary switch as before. However, if the rectifier is a "synchronous" switch or vibrator, it will usually have a contact time less than the primary switch, and if this is so, the proportional time of both primary and secondary should be measured and the shorter value used. Gasfilled rectifiers particularly the cold type, may introduce some delay in "switching" on and thereby slightly reduce the k of the circuit.

A rough estimation of k should be sufficient, as it does not have much effect on the calculation.

Low resistances are most easily measured by passing a high current through the circuit and measuring the separate voltage drops, as this method lends itself to easy evaluation of the separate resistances in the circuit such as contacts, pins and sockets and leads.

In measuring the total primary circuit resistance it should be remembered that the resistance of leads and batteries is appreciable in converters with low input voltage and high output. In this connection it may be mentioned that a converter of this type will be maltreated by supplying it from a high voltage source through a resistance in lieu of the proper low voltage source, since the contacts will have to break under the open circuit voltage.

The rectifier is easily dealt with so long as we can assume that the smoothing circuit has a very low input impedance to A.C. In this case the alternating voltage across the input of the filter may be assumed to be zero, and there will be a square wave current through the rectifier because there is a square wave voltage across it. This condition requires a large reservoir condenser at the input of the smoothing filter and is associated with maximum rectifier output, because the rectifier current is more uniform than with sine wave input.

If the rectifier is a pair of vacuum diodes, it is necessary to know the voltage drop e_T across one diode when it is passing a current equal to the peak value A.C. ($= I_a$).

The anode dissipation in one diode is given by $e_T I_2/2$ watts. The voltage varies with the 3/2 power of the current, but as the output current required is usually specified, this does not cause much difficulty. It is convenient to evaluate the equivalent resistance of the diode at the particular current value calling it $R_T = e_T/I_2$ (having divided by k as explained in Sec. 4 (I)—D.C. Output). This will not be strictly correct for calculating the regulation resistance of the converter, but the accuracy required in this is not usually high. Gasfilled rectifiers usually introduce a reverse voltage E_T, but no resistance into the circuit.

If the rectifier is a synchronous set of contacts, its resistance and voltage drop will be negligible, but it will generally reduce the overall contact time of the converter as previously mentioned.

The frequency should be known for the purpose of calculating the flux density in the transformer.

(c) *Properties of a Square Wave.*

The e.m.f. waveform illustrated in Fig. 2 has the following constants:—

Contact time ratio	k	
Peak values	E_a	I_a
Mean values	kE_a	kI_a
R.M.S. values	$\sqrt{k}E_a$	$\sqrt{k}I_a$
Power	kE_aI_a	

In order to avoid confusion, all subsequent calculations deal with peak voltage and current unless otherwise stated.

(d) *Equivalent Circuits.*

The following theory applies to both A and B types of circuit equally, if half the input winding of type B is taken as the primary in all the equivalent circuits. Similarly, in the case of a rectifying circuit of the usual kind with centre tapped secondary, half the secondary is taken as the secondary in the equivalent circuits.

There are two methods of dealing with the circuit given in Parts (I) and (II). The treatment throughout is carried out bearing in mind the problem of designing a converter for a given duty.

Part (i) deals with a method of calculating the performance of a given converter circuit with A.C. or with D.C. output. This method is very simple, and predicts immediately the voltage output and regulation resistance of any complete converter circuit. The hysteresis loss in the transformer core which it ignores has little effect on the above two quantities, and, together with the reed or commutator driving power, can be allowed for somewhat laboriously in the calculation of the overall efficiency.

This method of calculation can be used to design a converter for a given duty, i.e. when the desired output voltage and current and the input voltage are known. However, if the output required is a large fraction of the maximum possible, it will be necessary to calculate the exact value of the transformer ratio by successive approximation.

In order to analyse the circuit in more detail, the method of Part (II) has been used. This starts at the output terminals and uses the technique of the first part to estimate the power input to the core of the transformer. A few simple operations then produce all the quantities which are of interest, and show up several interesting factors which are of importance in the functioning of chopping inverters.

If the total circuit resistances R_1, R_2 are substantially independent of the transformer ratio, the method of the second part serves as a more elegant design method than that of Part (I). In general, an explicit design formula can always be worked out if the resistances R_1 and R_2 can be expressed as a function of n.

(e) *The Transformer.*

T = turns.
e = the induced or inducing e.m.f.
B = flux density in the core,
A = area of the core.

$$B = \frac{1}{AT}\int e\,dt$$

this curve is given in Fig. 2,

if $\hat{B}$ = maximum flux density.
k = proportional contact time.
E_a = peak induced e.m.f.

Then the turns per peak volt are given by the expression

$$\frac{T}{E_a} = \frac{k10^8}{4ABf}$$

The hysteresis loss is calculated as usual.

R_1 and R_2 partly depend on the size of the transformer which should be chosen by the usual methods of designing small transformers. It is usual to keep the maximum flux density low, not more than 7,000 lines/sq. cm.

If both windings are of Type A or of Type B, they should have equal volumes and losses. If one is Type A and the other Type B, the latter should have 1.4 times the volume and losses of the former. With these conditions the copper loss density is equal in primary and secondary and the maximum efficiency is obtained.

Part (I) Single Mesh. Equivalent Circuit

Both Type A and B inverter circuits can be resolved into the equivalent circuit of Fig. 6 in which E_d is the battery e.m.f. R_1 is the total resistance in the primary circuit when the switch is held to one side.

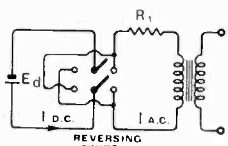

Fig. 6.

A.C. Output.

Turning our attention to the secondary side of the circuit, we first consider the case where the output is alternating current. Fig. 7 shows the first equivalent circuit. R_2 is the total resistance of the secondary circuit with the output terminals short circuited. Fig. 8 is equivalent to Fig. 7.

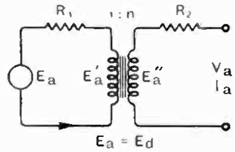

Fig. 7.

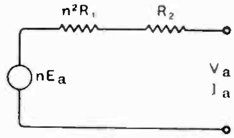

Fig. 8.

We can immediately deduce the regulation resistance

$$R = n^2R_1 + R_2$$

and the output voltage

$$V_a = nE_d - RI_a$$

Due to the conditions mentioned elsewhere being violated at low loads, these values will only be correct near full load. Referring back to Figs. 6 and 7 we see that the primary alternating current is nI_a and hence the D.C. component of the input current is knI_a, which would be indicated by a moving coil meter.

The output power is kV_aI_a.

The input power is knI_aE_d.

For all ordinary purposes it is sufficiently accurate to add the hysteresis loss to the input watts calculated above.

Otherwise we can use the method of Part (II) and evaluate x, read off η_1 from Fig. 11, evaluate primary voltage $E'_a = \eta_1 E_d$ and substitute in the equation of Section 4 (e) to find B_{max}, remembering to use half the primary turns if the inverter is of Type B (Fig. 1). The usual formulae can then be used to find the hysteresis loss, which can be added to the primary input, although strictly the method of Part (II) should be used for the whole input to the core.

The driving power must not be omitted in calculating the overall efficiency.

The maximum output power of this circuit is $kE^2_d/4R$. Normally the output voltage and current are specified and also the input voltage. Then n is calculated from

$$n = \frac{E''}{E'} = \frac{V_a + I_aR_2}{E_d - nI_aR_1}$$

D.C. Output.

The typical circuit is given in Fig. 9.

In some parts of this circuit we have A.C. or chopped D.C., and in others smooth D.C. Since we have D.C. output, it is convenient to find the voltage drop and loss effect of resistances carrying chopped or alternating current in terms of their effect on the D.C. output.

This is done by *dividing their resistance by k and assuming that it carries the equivalent smooth D.C.* Thus R carrying I_a amperes A.C. or chopped D.C. (peak value) will

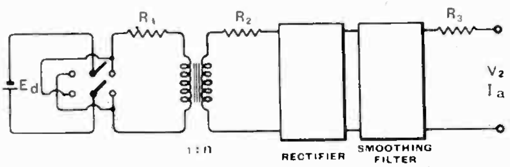

Fig. 9.

dissipate kI^2_aR watts and produce a volt drop of I_aR. If I_d is the equivalent rectified and smoothed D.C.

$$I_d = kI_a$$

A resistance R/k carrying I_d dissipates $RI^2{}_d/k$ and produces RI_d/k voltage drop.

Similar reasoning applied to a constant e.m.f. generator such as a battery or gas discharge shows that it is immaterial whether it carries smooth or chopped current, though of course it is multiplied by n if transferred across a transformer.

Using the above equivalences and eliminating the transformer, we get Fig. 10 (a) or 10 (b), according to whether the rectifier has a constant voltage drop E_T or an equivalent resistance R_T at the particular current value at which it is working, as dealt with in the section on measurements.

In the case of a gasfilled rectifier, we get immediately from Fig. 10 (a) :—

Regulation resistance $R = \frac{n^2R_1}{k} + \frac{R_2}{k} + R_3$

Output voltage $V_2 = nE_d - E_T - I_2R$.

Output power $= V_2I_2$

$$\left.\begin{array}{l}\text{Input power} \\ \text{less hysteresis and driving power}\end{array}\right\} = nI_2E_d$$

Maximum possible output

$$= (nE_d - E_T)^2/4R$$

The transformer ratio

$$n = \frac{V_2 + E_T' + I_2\left(\frac{R_2}{k} + R_3\right)}{E_d - nI_2R_1} = \frac{E''}{E'}$$

In the case of a thermionic rectifier, we get from Fig. 10 (b) :—

Regulation resistance

$$R = \frac{n^2}{k}R_1 + \frac{R_2}{k} + R_T + R_3$$

Output voltage $V_2 = nE_d - I_2R$

Output power $= V_2I_2$

$$\left.\begin{array}{l}\text{Input power} \\ \text{less hysteresis} \\ \text{and driving power}\end{array}\right\} = nI_2E_d$$

Maximum possible output $\doteqdot n^2E^2{}_d/4R$.

The transformer ratio

$$n = \frac{V_2 + I_2\left\{\frac{R_2}{k} + R_T + R_3\right\}}{E_d - nI_2R_1} = \frac{E''}{E'}$$

In all the cases treated, the values of the separate components of R show in what

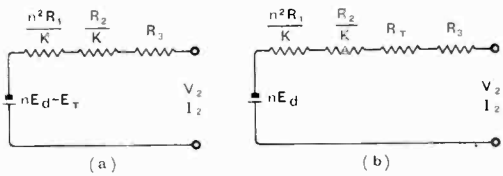

Fig. 10.

proportion they contribute to the total losses and regulation resistance.

Part (II) Double Mesh Equivalent Circuit

In all the above cases it is possible to evaluate the total secondary circuit power, which is a function of the output power only. Adding the hysteresis loss we can put W = total power input to the core.

Let $\eta_1 = \frac{W}{W_1}$

$$= \frac{\text{power input to core}}{\text{output from battery less driving power}}$$

or primary efficiency.

Let

$$\left.\begin{array}{l} E'_a = \text{transformer primary e.m.f.} \\ I'_a = \text{primary A.C.} \\ R_1 = \text{total primary circuit resistance} \end{array}\right\} \text{as before.}$$

$I_d E_d$ = D.C. component of battery current and e.m.f., then

$$E'_a = E_d - I'_a R_1 \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

$$W = kI'_a E'_a \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

$$W_1 = E_d I_d = kE_d I'_a \quad \ldots \quad \ldots \quad (3)$$

$$\text{Therefore } \eta_1 = \frac{W}{W_1} = \frac{E'_a}{E_d} = \frac{E_d - I'_a R_1}{E_d} \quad \ldots \quad \ldots \quad (4)$$

Substituting in (4) for I

$$\eta_1 = 1 - \frac{WR_1}{\eta_1 kE^2_d}$$

$$\text{Put } x = \frac{4WR_1}{kE^2_d}$$

$$\underline{\eta_1 = \tfrac{1}{2} + \tfrac{1}{2}\sqrt{1 - x}} \quad \ldots \quad \ldots \quad (5)$$

This function is plotted in Fig. 11.

Several interesting facts emerge from this equation. The overall efficiency of the converter depends on $4WR_1/kE^2_d$ and is large when this quantity is small.

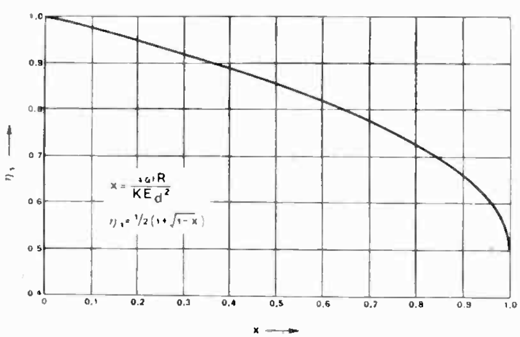

Fig. 11.

There is an upper limit to W, given by $W = kE^2_d/4R_1$, which actually corresponds to the well-known optimum transformer matching equation

$$n = \sqrt{\frac{R_s}{R_p}}$$

It is obviously deleterious to the efficiency to add resistance to limit the current which flows if the vibrator sticks to one side.

Design, Calculation of Efficiency, Transformer Ratio, etc.

Given the output, we must choose a transformer size with regard to the wattage output and maximum flux density. The primary and secondary resistance must be estimated.

Hence W is known

then $x = 4WR_1/kE^2_d$

$$\eta_1 = \tfrac{1}{2} + \tfrac{1}{2}\sqrt{1 - x}$$

use curve of Fig. 11.

$$\text{Input watts (less drive)} = \frac{W}{\eta_1}$$

Transformer primary $E_a = \eta_1 E_d$.

Hence transformer primary turns can be derived.

$$\text{Peak primary A.C. } I_a = \frac{W}{kE_a}$$

D.C. component of input current (less drive) $I_d = kI_a$.

5. Examples

Example I.

Converter giving 180 volts 35 mA. D.C. nominal output with 6 volts D.C. input using a synchronous 90 cycle vibrator for rectification. Primary and secondary circuits Type B.

Take the measured input voltage 5.85 and assuming 35 mA. output.

R_1 = 0.65 ohms (measured at input terminals with vibrator held to one side, neglecting battery resistance probably 0.02 ohms).

R_2 = 330 ohms (half of secondary winding).

R_3 = 500 ohms (smoothing choke).

Transformer turns ratio = 56.3 (primary 80 turns centre tapped, secondary 4500 turns centre tapped).

Square of turns ratio = 3160

rectifying contacts k = 0.7 approximately.

The calculated regulation resistance due to primary, secondary and smoothing choke respectively

$$= \frac{0.65 \times 3160}{0.7} + \frac{330}{0.7} + 500 \text{ ohms.}$$

i.e. 2940 + 480 + 500 = 3920 ohms.

This value was not measured in this case.

Note that the primary loss is much larger than that of the secondary, indicating incorrect division of space between primary and secondary.

The calculated output voltage

$$= 5.85 \times 56.3 - 3920 \times 0.035$$
$$= 329 - 137 = 192 \text{ volts.}$$

Measured value

$$= 171 \text{ volts.}$$

Note that the regulation resistance is rather high for the output. The output of this converter fell off very badly when the vibrator contacts became worn, giving a k of 0.5 and a measured regulation resistance of 5300 ohms, the characteristic near full load being—volts output $= 320 - 5300\ I_2$. The first term compares very well with that in the calculated output voltage.

Going back to the case of the unworn vibrator, the calculated input current

$$= 56.3 \times 0.035 + 0.15$$
$$= 2.12 \text{ amps.};$$

the second term is the driving current of the vibrator. Hysteresis and other core loss has, so far, been neglected. The measured input current was 2.15 amps.

As a matter of interest, using the method given in Section 4 (ii) we find $x = 0.84$, $\eta_1 = 0.70$ $E'_a = 4.1$ volts, $\hat{B} = 3240$ lines/sq. cm. and hysteresis loss $= 0.4$ watts which would bring the input current to 2.19 amperes.

Example II.

Inverter of Type B. Input 25 volts D.C. Output 230 volts nominal, 33 watts A.C. of 90 cycles/sec.

Half primary resistance $R = 0.175$ ohms.
Secondary resistance $R = 14$ ohms.
Core No. 28 stalloy square section.
Half primary turns $= 115$
secondary turns $= 1260$
turns ratio $= 11$.
Proportional contact time $k = 0.8$
∴ regulation resistance

$$= 121 \times 0.175 + 14$$
$$= 21 + 14 = 35 \text{ ohms.}$$

the ratio of primary to secondary losses $= 1.5$, which agrees well with the requirement of Section 4 (e).

Assuming an output current of 0.14 amperes R.M.S. then the peak output current

$$= \frac{0.14}{\sqrt{0.8}} = 0.157 \text{ amps.}$$

∴ Output voltage

$$= 25 \times 11 - 0.157 \times 35$$
$$= 275 - 5.5 \doteqdot 270 \text{ volts peak.}$$

that is $270 \times \sqrt{0.8} = 240$ volts R.M.S.

Input current $= 0.8 \times 11 \times 0.157 + 0.106 + 0.1 = 1.59$ amps.; the second term is the measured drive current, and the third term the estimated hysteresis loss ($B_{max} = 4900$) of 2.5 watts.

At 0.14 amps. R.M.S. the measured output voltage was 235 volts and the input current 1.55 amperes D.C.

Example III.

(Hypothetical case to bring out some points not mentioned in the previous cases.)

Converter giving 390 volts 0.2 amperes nominal with 60 c/s. vibrator and cold cathode rectifier. Primary and secondary Type; B; D.C. input 6.5 volts nominal. Open circuit battery voltage $= 7.0$.

The resistance of R_1 is composed of :—

Half of transformer primary	0.0075	ohms.
Vibrator contacts and pins	0.021	,,
Battery	0.01	,,
Leads, etc.	0.024	,,
Total R_1	$= 0.063$	,,

Half of secondary resistance	R_1	$= 40$ ohms.
Smoothing choke resistance	R_2	$= 60$,,
Transformer turns ratio	n	$= 80.7$
	n^2	$= 6500$
Voltage drop of rectifier	E_T	$= 90$ volts.
Proportional contact time	k	$= 0.75$
Hysteresis loss		$= 5$ watts.

Equivalent circuit Fig. 10 (a).

Regulation resistance

$$= \frac{0.063 \times 6500}{0.75} + \frac{40}{0.75} + 60$$
$$= 546 + 53 + 60 = 660 \text{ ohms.}$$

Note the enormous effect of incidental primary circuit resistances.

Voltage output

$$= 80.7 \times 7.0 - 90 - 0.2 \times 660$$
$$= 564 - 90 - 132 = 342 \text{ volts.}$$

Using the method of Part ii :—

Total secondary power

$$= 0.2\,(342 + 90) + 0.04\left(60 + \frac{40}{0.75}\right)$$
$$= 90.9 \text{ watts.}$$

$\therefore$ Total power input to core including hysteresis

$=96$ watts.

$$\therefore\ x = \frac{4 \times 96.0 \times 0.063}{0.75 \times 49} = 0.66.$$

$$\therefore\ \eta_1 = 0.79$$

$$\therefore \text{ D.C. Input power} = \frac{96}{0.79} = 122 \text{ watts.}$$

$+$ Driving power 6 watts $=$ 128 watts.

Input D.C. $=$ 18.3 amperes.

Introducing Radio Receiver Servicing

By E. M. Squire. Pp. 97+vi, 106 Figs. Published by Sir Isaac Pitman and Sons. Price 6s.

Described in the preface as a concise introductory guide to the practical operation of a radio receiver, this book is obviously intended for those who wish to fit themselves for carrying on the work of the many service-men, radio mechanics and testers who have joined the Services. An introductory chapter on the fundamentals of the electrical circuit and the wireless signal is followed by a general survey of a receiver. The construction and functions of the various components are then described, and a chapter is devoted to valves and their operation. Subsequent chapters deal with the reading of circuit diagrams and the practical problems of receiver maintenance and fault-tracing.

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

Frequency Modulation

To the Editor, The Wireless Engineer.

Sir,—The distinction between "frequency modulation *per se*" and auxiliary devices whose use is rendered possible by frequency modulation is difficult to define. Dr. Carson's classification of the amplitude limiter as an auxiliary device can perhaps be further justified by the fact that it must partially distort the R.F. wave-form of the frequency-modulated signal: anything which distorts the signal cannot be inherent in the form of signal.

On the other hand, it is arguable that the characteristics of the demodulator used in the receiver are an intrinsic part of the modulation system, since the signal cannot be observed without a demodulator. Now the balanced type of demodulator (described by Armstrong, *Proc. I.R.E.*, Vol. 24, p. 689) should give a considerable reduction of transient interference, regardless of the ratio of signal and interference amplitudes, provided the pulse of interference is so short that it may be represented by a Fourier series whose components are of practically uniform amplitude over the whole band of frequencies accepted by the receiver. Since the demodulator of a frequency modulation system is necessarily a device for detecting differences in the instantaneous frequency distribution of the received energy, it might be argued that this immunity from interference whose energy is uniformly distributed over the frequency band is a function of frequency modulation itself. This argument on transient interference cannot, however, be applied to the more general type of noise made up of numerous closely-spaced but random components ("fluctuation noise"), which has only a uniform *average* distribution of energy over the frequency band, with instantaneous variations. On this basis, frequency modulation can be said to minimise transient disturbances, but not other types of noise.

D. A. Bell.

London, N.21.

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