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Editorial

Directive Acoustic Pick-ups

IN a recent paper* Olson has described a number of what he calls line microphones. These are acoustic pick-ups with highly developed directional characteristics, so that they have a large response to sound coming from a desired source, but a small response to undesired sounds coming from other directions. A further important point that is discussed is the desirability of making the response ratio in the various directions independent of the frequency, otherwise the character of the sound will be modified. The basic idea of these line microphones will be evident from Fig. 1 where a number of small pipes with their open ends in a vertical line end in a common pipe in which is a ribbon element. Beyond the ribbon element is a heavily damped pipe which forms an acoustic terminating resistance ; its cross-section is equal to the combined section of all the small pipes. By means of bends all the small tubes are made of the same length, so that impulses that are in phase at the pick-up points arrive in phase at the junction. It has been found that the damping in the small pipes is sufficient to prevent resonance effects and interference between the separate pipes. If the acoustic terminating resistance is large compared with the mechanical reactance of the ribbon, its velocity, and therefore the induced e.m.f., will be proportional to the pressure and independent of the frequency just as in the electrical analogy $I = E/R$ if R is big compared with ωL.

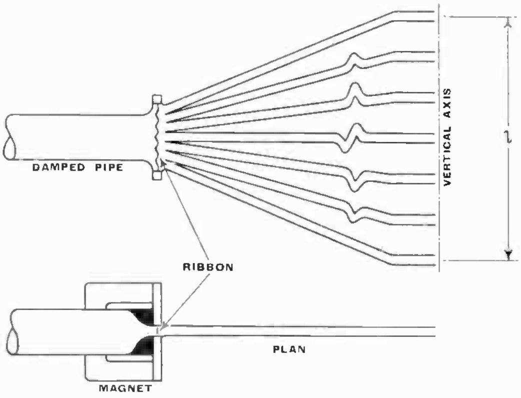

Fig. 1.

With the arrangement shown in Fig. 1 the response will be a maximum for any sound arriving in a horizontal direction, no matter from which point of the compass it may come, but for sound in any other direction the impulses at the open ends will have a phase displacement and will therefore give a reduced resultant at the junction. For sound arriving vertically the resultant will be zero if the length l is exactly a wavelength or a multiple of the wavelength. Assuming a large number of pipes close together it is easy to see that the ratio of the response in any direction making an angle θ with the

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* *Proc. I.R.E.*, July, 1939, p. 438.

horizontal to the maximum response is given by the formula

$$\frac{\sin\phi}{\phi} \quad \text{where } \phi = \frac{\pi l}{\lambda}\sin\theta.$$

The response curves so calculated are shown in Fig. 2 for three different values of l/λ. To obtain the characteristic surface these curves must be rotated about the vertical axis; Fig. 3 indicates the nature of the surface for $l = 2\lambda$, but omitting the small lobes.

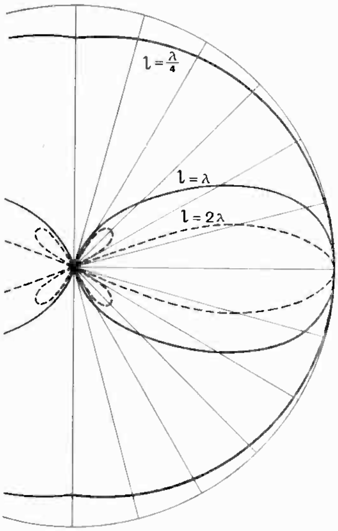


Fig. 2.

A more useful type is that having a maximum response in one direction only; such a pick-up is shown in Fig. 4 (a). Here the openings are approximately in a line in the direction of the pipes but with their distances from the common junction decreasing uniformly. Sound arriving along the axis from the front will arrive at the ribbon element just as if the pipes were not there. For sound arriving in the opposite direction there will be a maximum path difference of $2l$ between the shortest and the longest pipes, and if this is equal to λ the response in this direction will be zero. The outer curve in Fig. 5 shows the response for $l = \lambda/2$; that for $l = 4\lambda$ is shown by the inner curve, again omitting a number of small lobes. In this case the ratio is given by the formula $\frac{\sin\phi}{\phi}$ where $\phi = \frac{\pi l}{\lambda}(1 - \cos\theta)$. These curves must be rotated about the horizontal axis; Fig. 6 indicates the nature of the surface for $l = 4\lambda$. It will be seen that when this relation holds there is little response to sounds arriving at more than 30 degrees off the axis.

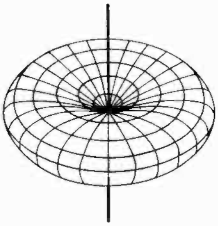
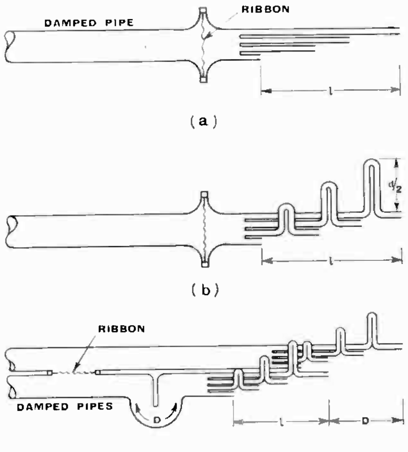


Fig. 3.

A further development is illustrated in Fig. 4 (b). Here delays are introduced into the pipes, the amount of the delay being proportional to the length of the pipe measured from the opening of the shortest. If the longest delay path is d, then the extreme difference in path between the longest and shortest pipes is $l + d$, and the ratio of response in any direction making an angle θ with the axis to what it would be for $\theta = 0$ *and no delay* is given by the formula

$$\frac{\sin\phi}{\phi} \quad \text{where } \phi = \frac{\pi}{\lambda}(l - l\cos\theta + d)$$

As d is increased the maximum response decreases, that is, the microphone loses sensitivity, but for small values of d this effect is small and is offset by the increased directivity obtained by using the delay, a line of length 2λ giving a curve very similar to that for a length of 8λ without the delay.



(a)

(b)

(c)

Fig. 4.

Fig. 4 (c) represents a radical departure from the foregoing types in that the ribbon element is operated on by the difference of pressure in two sets of pipes each similar to Fig. 4 (b), one set being nearer to the ribbon element by an amount D. Were this all, a sound arriving axially would not affect the ribbon, but a delay D is introduced into the shorter set of pipes which causes a phase difference at the ribbon which would give a maximum effect if D were made equal to $\lambda/2$. Here again, values of D are chosen which cause some loss of sensitivity but which give enhanced directivity. If D is small compared with λ the ratio of response in any direction making an angle θ with the axis to what it would be for $\theta = 0$ and no delay, is given by the formula $\frac{\sin\phi}{\phi}\cos\theta$ where ϕ has the value given above for the corresponding single set of pipes.

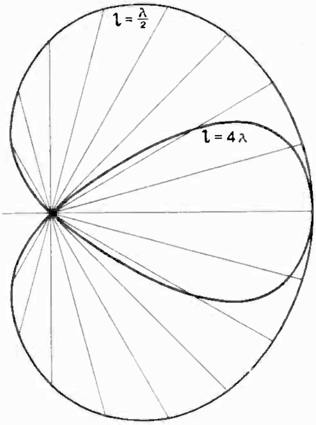


Fig. 5.

The object of all these expedients is to obtain a high degree of directivity for low notes, i.e. for long waves, with a small value of l, i.e. with a compact apparatus. Whereas a surface like Fig. 6 requires l in Fig. 4 (a) to be equal to 4λ, a similar surface can be obtained with the device shown in Fig. 4 (c) with $l = \lambda$. The importance of this is realised when it is remembered that a low note with a frequency of 100 cycles per second has a wavelength of about 11 feet.

Olson defines the directional efficiency of such a device as the ratio of the energy response of a non-directional pick-up to that of the directional one on the assumption that sound is arriving uniformly in every direction; the former is assumed to have a sensitivity in every direction equal to the optimum sensitivity of the directional pick-up. This figure—it is hardly an efficiency in the ordinary sense of the term—is about 14 for a pick-up with a characteristic surface similar to Fig. 6.

Figs. 2 and 5 show that a pick-up with a given length l has directive properties that are very dependent on the frequency. One of the applications of a directive pick-up is in the recording of sound-motion pictures, but it is obvious that for this purpose the directivity should be approximately independent of frequency over the essential speech and music range.

This has been attained by using a battery of five pick-ups each designed to have the desired directivity over a certain frequency range, and each ribbon element working through a filter, cutting out all frequencies outside this range. The five filters feed into a common output. The three highest frequency pick-ups are of the simple type shown in Fig. 4 (a), having 9, 13 and 17 pipes respectively, the length l of the last being about 5 times that of the first; their optimum frequencies are about 6,000, 2,000 and 700. For the two lower frequency bands pick-ups of the type shown in Fig. 4 (c) are employed with about 2 × 13 and 2 × 17 pipes respectively and with optimum frequencies of 250 and 85. By suitable adjustment the response curves for all frequencies between 85 and 8,000 lie within the shaded area of Fig. 7, except for small secondary lobes for angles greater than 90 degrees. The complete apparatus has an overall length of about 10 feet and is in the form of a horizontal beam carried on a tripod near its midpoint—somewhat cumbersome perhaps, but one cannot justifiably criticise it on this score unless he has a less cumbersome alternative that will give the same service.

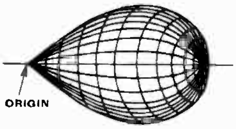
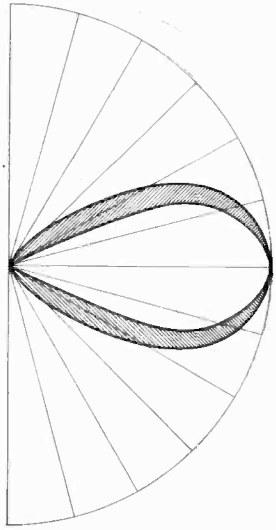


Fig. 6.

(Right) Fig. 7.



G. W. O. H.

B.

Distortion in Valves with Resistive Loads*

Graphical Methods for Its Determination

By Dr.-Ing. A. Bloch, M.Sc., F.Inst.P.

(Communication from the Research Staff of the M.O. Valve Co., Ltd., at the G.E.C. Research Laboratories, Wembley, England)

SUMMARY.—Simplified graphical methods are shown for the derivation of D.C., fundamental, 2nd, 3rd and 4th harmonic components in the anode current of a valve with sinusoidal grid excitation, if the anode current is known for 5 values of grid voltage spaced evenly over the range of grid swing. The methods follow directly from a rewriting of the well-known formulae for a "5-point analysis" and form at the same time an easy method for memorising these formulae.

Introduction

THE amplitude of fundamental and harmonic components in the anode current of a valve with sinusoidal grid excitation is, for resistive loads, usually determined by means of the formulae quoted below. These formulae are based on the approximation of the valve characteristic by a polynomial of the fourth order and they utilise a representation of this characteristic by 5 significant values of anode current (i_1 to i_5), spaced at equal intervals over the total swing of grid potential $2 \times E_g$ (see Fig. 1).

The formulae are as follows:

If the output of the valve is to be represented by

$$i_a = C_0 + C_1 \cos \omega t + C_2 \cos 2\omega t + C_3 \cos 3\omega t + C_4 \cos 4\omega t \quad \ldots \quad (1)$$

then

$$\begin{aligned} C_0 &= \tfrac{1}{6}[+ i_1 + 2i_2 + 2i_4 + i_5] \\ C_1 &= \tfrac{1}{3}[- i_1 - i_2 + i_4 + i_5] \\ C_2 &= \tfrac{1}{4}[+ i_1 - 2i_3 + i_5] \\ C_3 &= \tfrac{1}{6}[-i_1 + 2i_2 - 2i_4 + i_5] \\ C_4 &= \tfrac{1}{12}[+i_1 - 4i_2 + 6i_3 - 4i_4 + i_5] \end{aligned} \quad \ldots \quad \ldots \quad (2)$$

These equations can be derived by putting on the right hand side of equation (1) $\omega t = 0, \frac{\pi}{3}, \frac{\pi}{2}, \frac{2\pi}{3}, \pi$ and on the left-hand side $i_a = i_5, i_4, \ldots i_1$.†

It seems to have been overlooked that these formulae can be rewritten in a way which allows an extremely simple graphical determination of the various harmonics. The rearranged forms are as follows:

$$\begin{aligned} C_0 - i_3 &= -\tfrac{1}{3}\left[\left(i_3 - \frac{i_1 + i_5}{2}\right) + 2\left(i_3 - \frac{i_2 + i_4}{2}\right)\right] \\ C_1 &= \tfrac{2}{3}\left[\frac{i_4 + i_5}{2} - \frac{i_1 + i_2}{2}\right] \\ C_2 &= -\tfrac{1}{2}\left[i_3 - \frac{i_1 + i_5}{2}\right] \\ C_3 &= \tfrac{1}{3}\left[\left(i_2 - \frac{i_1 + i_3}{2}\right) - \left(i_4 - \frac{i_3 + i_5}{2}\right)\right] \\ C_4 &= -\tfrac{1}{6}\left[\left(i_3 - \frac{i_1 + i_5}{2}\right) - 4\left(i_3 - \frac{i_2 + i_4}{2}\right)\right] \end{aligned} \quad \ldots \quad \ldots \quad (3)$$

The formula for C_3 is obtained by adding and subtracting the term i_3 on the right-hand side of the original formula. Obviously any value i_x could have been used instead of i_3 and it will be seen that certain values of i_x lead to special simple constructions. The formula for C_4 is derived similarly to that for C_3. The D.C. term $C_0 - i_3$ is the change in direct current due to non-linearity of the valve.

These formulae lead to the graphical constructions described below.

Determination of Responses from the Anode Current/Grid Volts Characteristics

Fundamental response (Fig. 1).

Determine F midway between A and B, i.e. $i_F = \frac{1}{2}(i_1 + i_2)$, and similarly G midway

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* MS. accepted by the Editor, June, 1939.

† See Appendix. Espley and Farren, "Direct Reading Harmonic Scales," *Wireless Engineer*, April, 1934, use a similar method of development.

between D and E, i.e. $i_G = \frac{1}{2}(i_4 + i_5)$. The difference t between the ordinates of F and G multiplied by $\frac{2}{3}$ gives the amplitude of the fundamental response.

$$C_1 = \tfrac{2}{3} \times t \quad \ldots \quad \ldots \quad (4)$$

Note: Fig. 1 is useful for remembering the formula. As the peak response $\frac{1}{2}(i_5 - i_1)$ differs from the fundamental response only due to the presence of the 3rd harmonic,* the construction shown here is only necessary if the amplitude of the 3rd harmonic is appreciable; otherwise the more easily constructed expression $\frac{1}{2}(i_5 - i_1)$ is accurate enough.

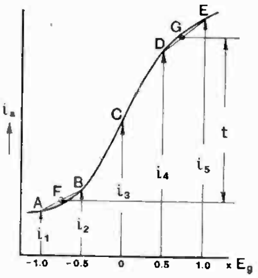

Fig. 1.—*Determination of fundamental response* $C_1 = \frac{2}{3} \times t$.

If the amplitude of the 3rd harmonic is known, it is also possible to find the magnitude of the fundamental response by subtracting the 3rd harmonic amplitude (taking consideration of its sign) from the peak response.

Second harmonic (Fig. 2).

Draw the straight line AE. The intercept HC formed on the vertical line through C, multiplied by $\frac{1}{2}$ gives the amplitude of the second harmonic. $i_H = \frac{1}{2}(i_1 + i_5)$

$$C_2 = -\tfrac{1}{2}(i_3 - i_H) = -\tfrac{1}{2}h \quad \ldots \quad (5)$$

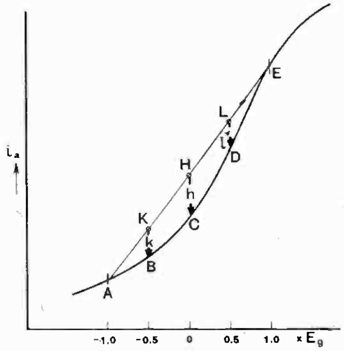

Fig. 2.—*Determination of 2nd and 3rd harmonic*

$C_2 = -\frac{1}{2}h$

$C_3 = \frac{1}{3}(k - l)$.

This construction coincides with the well-known construction in the case of the "3-point analysis"; it should be realised from the above derivation that it is valid in the presence of 3rd and 4th harmonic.

In the case illustrated the sign of h is negative, giving a positive amplitude of second harmonic.

It is useful to remember as a sign rule, that h is reckoned positive if C is above the line AE which acts as a zero line. Similar conventions are adopted for the intercepts used in the following constructions.

Third harmonic (Fig. 3).

Draw the straight lines AC and CE. The difference between the intercepts $K'B = k'$ and $L'D = l'$, multiplied by $\frac{1}{3}$ gives the amplitude of the 3rd harmonic.

$$C_3 = \tfrac{1}{3}(k' - l') \quad \ldots \quad \ldots \quad (6)$$

In the case illustrated the sign of k' is negative giving a third harmonic of negative sign.

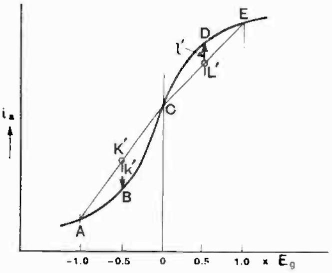

Fig. 3.—*Determination of 3rd harmonic* $C_3 = \frac{1}{3}(k' - l')$. *In this case k' is of negative sign.*

It will be noticed that instead of drawing the straight lines AC and CE as in Fig. 3 one could have used straight lines AC' and $C'E$, where C' is an arbitrarily selected point on the vertical through C (not illustrated). The magnitude of the significant intercepts, now denoted by k and l, would have been different. However, as the change is the same for each of them their difference $k - l$ remains equal to $k' - l'$, so that equation (6) may be rewritten $C_3 = \frac{1}{3}(k - l)$. This procedure corresponds to the possibility mentioned above, of writing i_x instead of i_3 in the formula for C_3.

The following cases are of special interest:
(1) The position of C' can be selected so that it falls on the straight line AE (as drawn in Fig. 2, C' coinciding there with H). The difference in the intercepts $KB = k$ and

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* The second and fourth harmonics each add equal amounts to the currents i_1 and i_5, leaving their difference unchanged. See the equations given in the appendix.

$LD = l$, multiplied by $\frac{1}{3}$ gives again the amplitude of the 3rd harmonic.

$$C_3 = \tfrac{1}{3}(k - l)$$

(2) The position of C' can be selected so that the magnitude of one of the intercepts (e.g. k) becomes zero (see Fig. 4). This is done by drawing the line AC' through B.

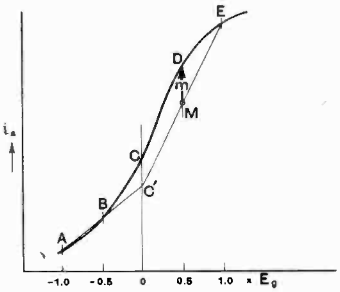

Fig. 4.—*Determination of 3rd harmonic*

$C_3 = -\frac{1}{3}m.$

Then the full magnitude of the 3rd harmonic (apart from the factor $\frac{1}{3}$) is shown on the other intercept $MD = m$.

$$C_3 = -\tfrac{1}{3}m$$

It will be shown that this procedure is useful when the valve characteristic is given by a load line drawn through the family of anode current/anode voltage curves.

Fourth harmonic and D.C. component. (Fig. 5).

Draw the straight line AE, defining the intercept h as in Fig. 2, and the line BD, defining in a similar way the intercept $n = i_3 - \frac{1}{2}(i_2 + i_4)$

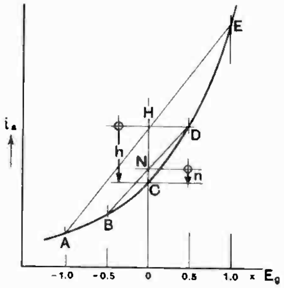

Fig. 5.—*Determination of D.C. component and of 4th harmonic*

$C_0 - i_3 = -\frac{1}{3}(h + 2n)$
$C_4 = -\frac{1}{6}(h - 4n).$

Then

$$C_0 - i_3 = -\tfrac{1}{3}(h + 2n)$$
$$C_4 = -\tfrac{1}{6}(h - 4n)$$

It will be seen from these formulae, that in the absence of 4th harmonic ($n = \frac{1}{4}h$), the expression for the D.C. component becomes

$$C_0 - i_3 = -\tfrac{1}{2}h$$

i.e. equal to that derived from a 3-point analysis.

Determination of Valve Responses from a Load Line Drawn on Anode Current/Anode Volts Characteristics

If the valve characteristic is given by the family of anode current/anode voltage curves, it is easily seen how the rules given above have to be applied. Fig. 6 shows a " load line " on such an anode voltage/anode current diagram, on which the points A, B, C, D, E (corresponding to the points $A - E$ of Fig. 1) are determined by the intersection of this load line with the anode current curves for 5 equally spaced values of grid voltage.

As a rule only the relative magnitude of the harmonic components is wanted, so for convenience the distances measured along the load line are spoken of as " amplitudes," in place of their projection on the i_a axis.

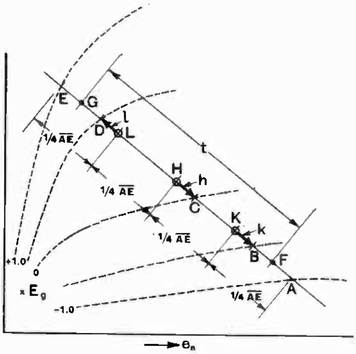

Fig. 6.—*Determination of fundamental and harmonic responses from a load line on an anode current/anode voltage diagram.*

$C_1 = \frac{2}{3}xt \quad C_2 = -\frac{1}{2}h \quad C_3 = \frac{1}{3}(k - l).$

Intercepts used for the determination of harmonics are counted as positive if their projection on the i_a axis counts as positive (i.e. is pointing in the direction of increasing current).

(*a*) Fundamental response: Point F bisects the distance AB, point G bisects the distance DE; $\frac{2}{3}$ of the distance $FG = t$ represents the amplitude of the fundamental response.

(*b*) Second harmonic: Point H is midway between the points A and E. $\frac{1}{2}$ of the distance $HC = h$ represents the second harmonic. (In the case illustrated h is negative, therefore C_2 positive, cf. eq. 5.)

(*c*) Third harmonic: Point k is midway between the points A and H, point L is midway between the points H and E. Thus the intercepts $KB = k$ and $LD = l$ are defined as above. (In the case illustrated k is negative and l is positive.)

The constructions required under (*b*) and (*c*) lead to the design of an extremely simple scale for the direct determination of the harmonic content. It consists simply of 5 parallel lines drawn at equal distances. A number of subsidiary lines can be drawn, subdividing the total width of the scale into 20 parts. This scale and its use are illustrated in Fig. 7. This scale is placed over the load line as shown :

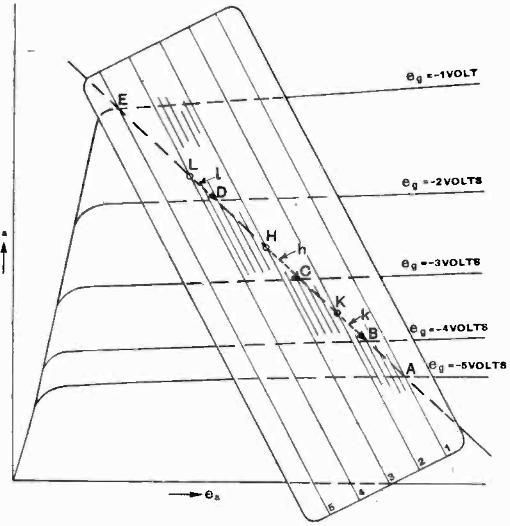

Fig. 7.—*Use of a transparent scale for the determination of 2nd and 3rd harmonic.*

The 5 main lines are numbered 1 – 5 ; line 1 of the scale passes through point A of the load line, line 5 through point E.

Line 3 will then indicate on the load line the position of point H, and the lines 2 and 4 the position of the points K and L. If the fundamental response is assumed to be accurately enough equal to $\frac{1}{2}(i_5 - i_1)$ then the spacing of the subsidiary dividing lines amounts to 10 per cent. of the fundamental amplitude and thus it is possible to read the significant intercepts directly in per cent. of the fundamental response.* In order to avoid the crowding of the Figure by the subsidiary lines, these are indicated only on the places required for a reading.

In the case illustrated the second harmonic content will be $\frac{1}{2} \times 25 = 12.5$ per cent. and the 3rd harmonic content will be $\frac{1}{3}(-21 + 18) = 1$ per cent.

The scale illustrated can also be used for the determination of the 4th harmonic C_4 and of the D.C. component $C_0 - i_3$. The procedure follows from Fig. 5. The intercept h is determined as was shown in Fig. 7, the intercept n is similarly determined from NC where N bisects the distance BD (not illustrated).

$$C_4 = -\tfrac{1}{6}(h - 4n)$$

$$C_0 - i_3 = -\tfrac{1}{3}(h + 2n)$$

If a scale as just described is not available, then the following method will be found very convenient for the determination of the 3rd harmonic. It is based on the special selection of the position of point C' which has been illustrated above in Fig. 4. The procedure is illustrated in Fig. 8.

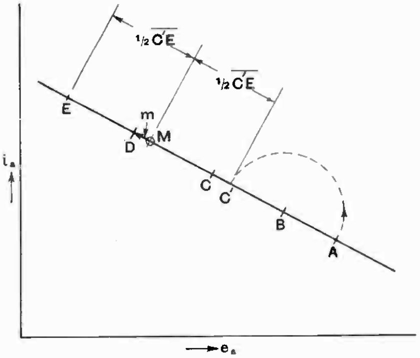

Fig. 8.—*Determination of 3rd harmonic* $C_3 = -\frac{1}{3}m$.

Make BC' equal to AB (indicated by the dotted circle) and select M midway between C' and E. The amplitude of 3rd harmonic is then represented by $\frac{1}{3}$ of the distance $MD = m$.

$$C_3 = -\tfrac{1}{3}m$$

In conclusion the author desires to tender his acknowledgments to the Marconiphone Co., Ltd., and The General Electric Co., Ltd.

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* If there should be an appreciable amount of 3rd harmonic present, say a fraction p of the fundamental, then the distance AE will differ from the value which it is assumed here to possess (twice the amplitude of the fundamental) by the same fraction. The true value p follows then from the value p' read off the scale according to the relation $p = \frac{p'}{1 - p'}$.

on whose behalf the work was done which has led to this publication.

APPENDIX

If the grid potential is given by $e_g = E_g \cdot \cos \omega t$ then the anode current of the valve can be represented for resistive loads by

$$i_a = C_0 + C_1 \cos \omega t + C_2 \cos 2\omega t + C_3 \cos 3\omega t + C_4 \cos 4\omega t$$

(There can be no sine terms in this series, for the following reasons: The input function is symmetrical with respect to $t = 0$; the output function in the case of a resistive load—being entirely in phase with the input—must possess the same symmetry with respect to $t = 0$; the presence of sine terms would interfere with this symmetry.)

For the values ωt indicated in the text, the following equations are obtained:

$$i_5 = C_0 + C_1 + C_2 + C_3 + C_4$$

$$i_4 = C_0 + \frac{C_1}{2} - \frac{C_2}{2} - C_3 - \frac{C_4}{2}$$

$$i_3 = C_0 \qquad - C_2 \qquad + C_4$$

$$i_2 = C_0 - \frac{C_1}{2} - \frac{C_2}{2} + C_3 - \frac{C_4}{2}$$

$$i_1 = C_0 - C_1 + C_2 - C_3 + C_4$$

The solution of these 5 equations gives the expressions for C_0, C_1, C_2, C_3 and C_4 quoted in the text.

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

Permeability at Very High Frequencies

To the Editor, The Wireless Engineer.

Sir,—I have read with great interest Prof. Howe's Editorial in the November issue of *The Wireless Engineer*, but do not agree with his statement that the effect of eddy currents can be allowed for in the measurement of the apparent permeability of a magnetic material at high frequencies or even at low frequencies in the case of permalloy and mumetal. In a paper on "Magnetic Characteristics of Nickel-Iron Alloys with Alternating Magnetising Forces" (*Journ. I.E.E.*, 79, p. 213, 1936), results are given in which the average flux density in ring laminations of mumetal, 0.038 cm. thick, tested at 30 c/s, is less than an eighth of the value with the corresponding D.C. excitation. Taking the permeability as 60,000—namely, the maximum D.C. value for the specimen—and a resistivity of 42 microhm-cm. it was found that the usual mathematical formula for sheet material (derived on similar assumptions to those made by Prof. Howe) does not account for more than about 12 per cent. of the decrease in flux density obtained experimentally. In other words, the skin effect indicated by the mathematical formula ought to be comparatively small even though the maximum D.C. value of permeability was used. But why should we take this value in preference to, say, the value corresponding to the maximum magnetising force? As far as I am aware, in no book or paper dealing with this subject mathematically is it suggested how the value of the permeability should be determined. Furthermore, it is assumed that the permeability is constant throughout the section of the specimen—an assumption that has no justification except its mathematical convenience.

In the above-mentioned paper, I showed that it was possible to account for at least 70 per cent. of the decrease in the apparent flux density by using the differential permeability obtainable from the D.C. hysteresis loops.

Finally, is it correct to say that iron loses its magnetic property at high frequencies merely because the depth of penetration is extremely small? For instance, would an iron wire carrying a very high frequency current cease to be attracted by a magnet? I doubt it. I suggest that the iron at the centre of the wire is as magnetic as when the wire is not carrying any current.

Technical College, Edward Hughes.
Brighton.

Dr. Hughes says that he does not agree with my statement "that the effect of eddy currents can be allowed for in the measurement of the apparent permeability." I have searched the Editorial in vain for this statement. I said that "skin effect can be employed to determine the value of the permeability" and the whole Editorial is a proof of this statement. I also said "even if one assumes that the permeability is constant throughout the material and throughout the cycle" showing that I fully appreciated that it is really neither, but varies throughout the material depending on variations of magnetic induction, and from moment to moment as the material at any point goes round its hysteresis loop. I feel sure that every reader would understand perfectly well that any value of permeability calculated from skin-effect observations on an iron wire could be nothing but an effective value covering a multitude of complexities.

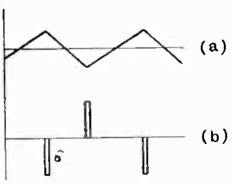
I am not sure whether his concluding paragraph is intended to be taken seriously. Obviously only iron which is subjected to the high-frequency magnetic force can exhibit this phenomenon. Other iron in the neighbourhood, whether outside or inside the wire, which is not subjected to the magnetic force, will take no part in the phenomenon, and will retain its normal properties. If someone told Dr. Hughes that ice lost its solidity when subjected to heat I can hardly imagine him saying, "Would an ice block brought into a warm room cease to be solid? I doubt it. I suggest that the ice at the centre of the block is as solid as when the block was not in the warm room." Similarly the change of magnetic properties under discussion only occurs in the thin surface layer into which the magnetic field has penetrated.

G. W. O. H.

Frequency or Phase Modulation?

To the Editor, The Wireless Engineer.

Sir,—In a short note headed "Frequency or Phase Modulation?" appearing in the November *Wireless Engineer*, a method of illustrating the differences between the two systems is described. It may be of interest to examine further the implications of the type of modulation dealt with there, particularly as it tended to obscure the important fact that a phase modulated wave is *ipso facto* also frequency modulated and *vice versa*. The form of modulation is, however, different according as to whether it is regarded as phase or frequency modulated. Thus, in the case quoted by G. W. O. H., rectangular frequency modulation is indistinguishable from phase modulation of the type represented by curve (*a*), and rectangular phase modulation is the same as frequency modulation in a series of alternate positive and negative pulses, somewhat as in (*b*).



The argument is not materially affected by the fact that, in the theoretical case chosen, the pulses are of zero duration and infinite amplitude. The curves (*a*) and (*b*) are, of course, the integral and differential of the rectangular wave.

G. W. O. H. suggests that in the case of rectangular frequency modulation the accompanying change of phase is a "secondary phenomenon," but in view of the fact that the particular wave form can be accurately and completely described in terms of either system, the supposition that any one is more fundamental must be purely conventional. Finally, it is worth pointing out that this relation between the two systems is of more than theoretical interest as it enables frequency modulation to be produced by a phase modulating device, provided the modulation is initially integrated as in the Armstrong system.

Cambridge. D. I. Lawson.
D. Weighton.

The above brings out very clearly and emphasises the point which I discussed in the note referred to. With regard to "*secondary* phenomena," if Mr. Lawson said "I am tired of crawling along like this at 30 miles an hour, I am going at 50 for a bit," and his friend, Mr. Weighton, presently said "We are leaving our friends in that other old car a long way behind," I can imagine Mr. Lawson replying "Oh, that's a *secondary* matter." Conversely, if, on another occasion, he remarked that he could not stand the smell of that other fellow's exhaust and was going to let him get farther ahead, Mr. Weighton might presently draw his attention to the fact that they were moving very slowly, to which Mr. Lawson might again reply that that was a *secondary* matter. I agree that the supposition that speed or frequency is more or less fundamental than displacement whether of cars or phase is purely conventional.

G. W. O. H.

Tables for Converting Rectangular to Polar Co-ordinates

By J. C. P. Miller, Ph.D. Pp. 16. Scientific Computing Service Ltd., 23 Bedford Square, London, W.C.1. Price 2s.

These tables will be a perfect godsend to all those who have to make alternating current calculations. When two or more vector quantities have to be added or subtracted they must be expressed in rectangular co-ordinates, but when they have to be multiplied or divided it is simpler if they are in polar co-ordinates, and the continual transformation from one form to the other often involves much irksome calculation. Since the triangles are always right-angled the shape is determined by the ratio of the shorter to the longer rectangular co-ordinate and this is taken as the "argument" of the tables, or, in other words, the longer co-ordinate is taken as unity. The tables then give the hypotenuse and the two angles in degrees; they also give one of the angles in radians. The argument is given in steps of 0.001 which gives an accuracy sufficient for all practical purposes. A table is also given for reducing angles to the first quadrant, that is, for telling you exactly where you would be after rotation through, say, 200 radians.

The tables are well printed in a very convenient form and can be unreservedly recommended.

G. W. O. H.

The I.E.E.

The retirement of Mr. P. F. Rowell, who has been secretary of the Institution of Electrical Engineers for thirty-eight years, is referred to by Mr. Johnstone Wright in his inaugural address as President of the Institution for the 1939–40 Session which, because of the suspension of meetings, has been circularised to all members. Mr. Wright says "To most members Mr. Rowell and the Institution have become synonymous, and his unique and splendid contribution to its development will be of permanent value." The new secretary is Mr. W. K. Brasher.

"The History of the Institution of Electrical Engineers" is the title of a book just published by the I.E.E., which is available to the public at the price of 18s. 6d. The publication relates chiefly to the sixty years from 1871, when the Society of Telegraph Engineers was founded, to 1931, which marked the centenary of the discoveries of Michael Faraday.

Electron Transit Time*

Effects in Cathode-Ray Tubes and Diodes

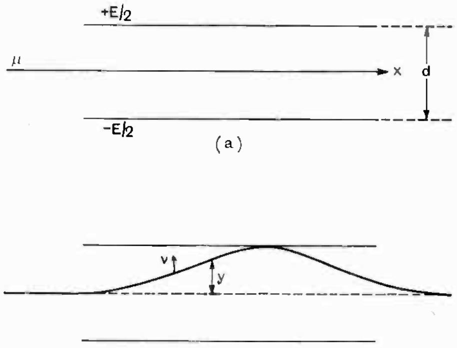
By W. E. Benham

THE conductivity endowed upon a temperature limited, or voltage saturated, diode at sufficiently high frequencies by virtue of electron inertia has been studied by a number of writers on transit time topics, usually in connection with the retarding field triode, or "transitron." More recently the interest in the magnetron has prompted extension to magnetic fields. Hollmann and Thoma[1] have also attempted the case where the electrons in a cathode-ray beam are subjected to combined magnetic and electrostatic forces.

The conclusions (not always correctly) arrived at at different times by different workers may best be discussed in terms of two transit angle functions Y_2 and Y_4. The function Y_3 will also be mentioned. All we need to worry about in the preliminary discussion in regard to these functions is that the in-phase (real) part of these complex functions is for all frequencies positive in the case of Y_2, while in the case of Y_4, Y_3 negative as well as positive values are possible. All functions are oscillatory, in the physical sense that the curves are wavilinear, but in the case of Y_2 the troughs of the curve never extend beyond zero into negative values. This means that this function does not give negative resistance effects, whereas the other two do.

As there is a difference according as whether the electrons are moving normally to or parallel to the plates (diode or deflecting condenser), we will start by considering the simplest case—that of the deflecting condenser originally treated by Sven Benner in 1929. The following analysis gives Benner's results more simply than any method I know. The effect of stray field will be neglected at the outset, as in all cases where the distance apart of the deflector plates is small compared with their length the effect of stray field is very small.

Let v denote the component normal to the plates† of the beam velocity, said component arising as a result of a deflecting voltage $E \cos \omega t$. By well-known convention we replace $\cos\omega t$ by $exp(i\omega t)$ to simplify analysis. The rate of change of momentum normal to the plates of a beam electron of charge $-q$ which entered the deflecting condenser at arbitrary instant t_0 is then given in well-known manner by



(a)

(b)

Fig. 1.

$$\frac{d}{dt}(mv) = q\frac{E}{d}e^{i\omega t} \quad .. \quad .. \quad (1)$$

Since we are neglecting stray fields no transverse momentum is acquired by the charge $-q$ before the instant t_0, so that integration with respect to t we obtain the momentum

$$mv = \frac{qE}{d}(e^{i\omega t} - e^{i\omega t_0})/i\omega \quad .. \quad .. \quad (2)$$

where t_0 makes its first appearance as part of an arbitrary constant chosen so that $v = 0$ when $t = t_0$. If I_B is the beam current an infinitesimal length dx of beam contains $I_B dx/qv_B$ electrons, where v_B is the beam velocity parallel to the plates at instant t_0. These give rise by electrostatic induction

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* MS. accepted by the Editor, July, 1939.

† See Fig. 1.

to the current

$$di = \frac{I_B dx}{q v_B} \frac{qv}{d}$$
$$= I_B q E\, dx(e^{i\omega t} - e^{i\omega t_0})/i\omega m d^2 v_B \quad .. \quad (3)$$

which flows in the circuit attached to the deflector plates. Replacing dx by $v_B d\tau'$, where $\tau' = t - t_0$, we obtain by integration with respect to τ', at t constant, the current corresponding to the length l of beam included momentarily between the deflecting plates :—

$$i = \frac{I_B q E}{d^2 m} e^{i\omega t} \int_0^\tau \frac{1 - e^{-i\omega\tau}}{i\omega} d\tau'$$
$$= \frac{I_B q \tau^2}{2md^2} (E e^{i\omega t}) Y_2(i\omega\tau) \quad .. \quad .. \quad (4)$$

where

$$Y_2(x) = 2(x - 1 + e^{-x})/x^2 \quad .. \quad .. \quad (4a)$$
$$= 1 - x/3 + x^2/12 - x^3/60 + x^4/360 - .... \quad .. \quad .. \quad (4b)$$

Separate calculation for the correction to be applied for the effect of wide angle deflection shows equation (4) to be subject to an error usually below 5 per cent. The apparent impedance of the plate pair to sinusoidal currents is obtained by dividing by $Ee^{i\omega t}$, giving (in real notation) after slight rearrangement :—

$$1/Z = 1/R + i\omega C = \frac{I_B q \tau^2}{2md^2}\left[\frac{2}{(\omega\tau)^2}(1 - \cos\omega\tau) - \frac{2i}{(\omega\tau)^2}(\omega\tau - \sin\omega\tau)\right] + i\omega C_0 \quad .. \quad (5)$$

where C_0 = cold capacitance between plates. It is convenient to replace τ by $1/v_B$ or, better, by $(ml^2/2qV_B)$ where V is the beam velocity in volts. We then obtain

$$R = \frac{V_B}{I_B}\left(\frac{2d}{l}\right)^2 \frac{(\omega\tau)^2}{2(1 - \cos\omega\tau)} \quad .. \quad (6)$$

$$C - C_0 = -\frac{I_B}{V_B}\left(\frac{l}{2d}\right)^2 \frac{\tau}{3}\left[\frac{6}{(\omega\tau)^3}(\omega\tau - \sin\omega\tau)\right] \quad .. \quad .. \quad (7)$$

the quantities in square brackets tending to unity for $\omega\tau$ small. We thus see that the apparent parallel resistance at low frequencies is equal to the D.C. resistance of the beam (V_B/I_B) times the square of the ratio : distance between deflector plates : half deflector plate length.

As a concrete example consider a television cathode-ray tube with beam current 50 μA. and gun voltage 5000. Let $2d = 1.2$ cms., $l = 3.6$ cms. making $(2d/l)^2 = 1/9$, giving for low frequencies

$$R = 1.1 \times 10^7 \text{ ohms.}$$

Since $(1 - \cos\omega\tau)$ vanishes for $\omega\tau = 2n\pi$, where $n = 0, 1, 2, 3 \ldots\ldots$ we have except in case $n = 0$, an infinite resistance and zero power consumption. In practice, however, the deflection is also zero* under the same conditions. Deflection is thus always* accompanied by expenditure of energy by the deflecting means, as rightly pointed out by Colebrook and Vigoureux[2] in their reply to Hollmann and Thoma.

For the sake of completeness we give for small $(\omega\tau)$ the value of the capacitance change in the above example :—

$$-\Delta C = C_0 - C = \frac{I_B}{V_B}\left(\frac{1}{2d}\right)^2 \frac{1/3}{\sqrt{2qV_B/m}}$$
$$= 2.6 \times 10^5 \ \mu\mu\text{F}.$$

In the above example both damping and capacitance changes are small. They are smaller still at larger values of $(\omega\tau)$.

Now we are in a position to see how the temperature limited diode must be treated. In the above treatment of an electron beam variations in τ over the cycle are zero. This however, is not true in the diode case, where the velocities are aligned, even if $v \ll v_B$. In the case of the deflecting condenser these velocity components being at right angles it is possible to show that there is no such change in τ not only for low frequencies, but for any frequency. Accordingly, for the temperature limited diode the procedure is more involved, though not unduly so. If V_0 be the D.C. potential across the diode the equation corresponding to (2) is

$$mv - mv_0 = \frac{qV_0}{d}[\tau + N(e^{i\omega t} - e^{i\omega t_0})/i\omega] \quad .. \quad (8)$$

where N is the ratio of A.C. to D.C. potential on the diode and is of the first order, while τ is composed of a zero order term τ_0 and a

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* The deflection y is obtained by a further integration, starting with equation 2. The function Y_3 (compare equation 11a) is involved, the *real* part only of which is zero when $\omega\tau = 2n\pi$. The deflection y is thus not actually zero, but *in phase quadrature* with the deflecting potential under those conditions. Such deflection is, however, to be distinguished from the deflection registered on the cathode-ray fluorescent screen which is substantially proportional to v.

first order term τ_1. There are also higher order components of τ but while Sloane and James showed how important these are in determining waveform we will for simplicity here restrict ourselves to the case of first order solution for which it is sufficient to write

$$\tau\ (= t - t_0) = \tau_0 + \tau_1.$$

τ_1 is known as the first order variation time.

In order to determine τ_1 a further integration of (8) is necessary. Since we are only seeking essential differences the initial velocities are neglected ($v_0 = 0$). Then, taking $x = 0$ at the cathode, we have

$$mx = \frac{qV}{2d}\left[\tau^2 + 2N(e^{i\omega t} - e^{i\omega t_0} - i\omega\tau e^{i\omega t_0})/i^2\omega^2\right]$$

$$= \frac{qV}{2d}\left[\tau_0{}^2 + 2\tau_0\tau_1 + \frac{2N\tau_0{}^2e^{i\omega t}}{\alpha^2}(1 - e^{-\alpha} - e^{-\alpha}\,.\,\alpha)\right], \quad \ldots \quad (9)$$

to first order, where $\alpha \equiv i\omega\tau_0$.

Now since the time taken by an electron to travel a given distance x is τ_0 in the absence of alternating potentials and τ when alternating potentials are present we have a second equation for x in terms of τ_0, namely :

$$mx = qV\tau_0{}^2/2d \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

Eliminating x equations (9) and (10) give, writing $\tau_1 = \alpha_1/i\omega$ and

$$Y_3(\alpha) = \frac{2}{\alpha^2}(1 - e^{-\alpha} - \alpha e^{-\alpha}) \quad \ldots \quad \ldots \quad (11a)$$

$$= 1 - 2\alpha/3 + \alpha^2/4 - \alpha^3/15 + \alpha^4/72 - \ldots$$

$$\alpha_1 = \frac{N\alpha Y_3}{2}e^{i\omega t} \quad \ldots \quad \ldots \quad (11b)$$

being $i\omega$ × the first order variation time. We now give τ its full value in (8) which becomes (dropping out the velocity v_0 as we did in reaching (9)) :—

$$v = \frac{qV\tau_0}{md}\left[1 + Ne^{t}(2 - 2e^{-\alpha} - \alpha Y_3)/2\alpha\right]$$

$$= \frac{qV}{md}\tau_0\left(1 + \frac{N}{2}Y_2e^{i\omega t}\right) \quad \ldots \quad \ldots \quad (12)$$

where $Y_n \equiv Y_n(\alpha)$; see definition in (4a).

Now each electron induces in the electrodes a current proportional to v. Since however there is not a uniform distribution of space charge we must effect summation of all contributions obtained throughout the space charge at a given instant. Note that while we still talk of space charge in the temperature limited diode this does not mean that all the effects of space charge are equally important. We are here, for example, purposely neglecting the effects of the electrons in the space on one another. The words "in the space" should be particularly noted—as will be seen we take into account the action of the space electrons on the electrode charges, and since the positive charged particles in the metal of the electrodes are by common consent practically immobile it must be the movement of the free electrons away from the surface layer which brings about the effect of the (positive) charge induced by the space electrons by electrostatic "influence."

Since the fluctuating electron convection current is zero (apart from Schroteffekt) at the cathode, all we need to do is to take the induced current (due to electrostatic influence of the space electrons) as[3]

$$i_e = \frac{I_0}{d}\int_0^\tau v\,d\tau', \quad \ldots \quad \ldots \quad (13)$$

where I_0 is the D.C. current and v is the velocity. For the purposes of this integral, which must be carried out at t constant, we use equation (8).

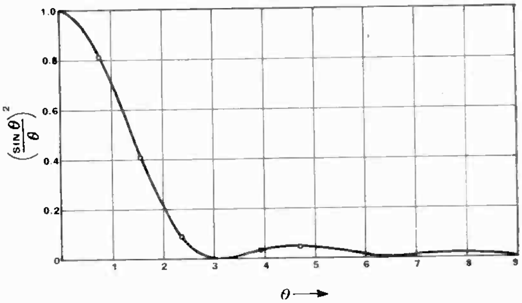


Fig. 2.—*Deflection system. Inter-plate conductance of deflector plates of cathode-ray tube—transit time fixed, frequency varied.*

Equation (12) would not be suitable as it is expressed in terms of τ_0 whereas in (13) we are concerned with the instantaneous value τ :—

$$i_e = \frac{I_0qV_0}{md^2}\int_0^\tau\left[\tau' + \frac{N}{i\omega}(e^{i\omega t})(1 - e^{-\alpha'})\right]d\tau'$$

$$= \frac{I_0qV_0}{md^2}\frac{\tau^2}{2}\left[1 + N(Y_2)e^{i\omega t}\right] \quad \ldots \quad \ldots \quad (14)$$

Equation (14) gives the induced current in a temperature limited diode in terms of the instantaneous transit time. To separate out the fluctuating component of τ we use equation (11b). In the study of these transit angle functions we come across a number of useful relations, and the one which gives us the solution sought is as follows :—

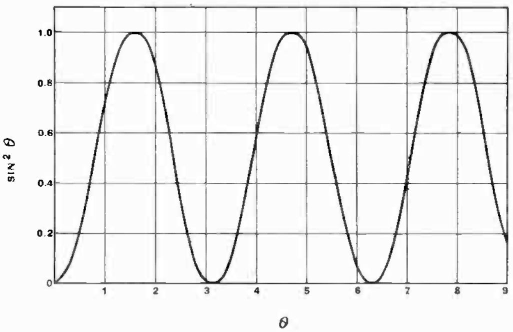


Fig. 3.—*As Fig. 2, but for fixed frequency and variable transit time.*

$$Y_2 - Y_3 = \frac{\alpha}{3} Y_4 \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

Thus in terms of D.C. transit times we have

$$\left(\text{writing now } d = \tfrac{1}{2}\frac{qV}{md}\tau_0{}^2 = \tfrac{1}{2}g\tau_0{}^2\right)$$

$$i_e = I_0\left(1 + N\frac{\alpha}{3}Y_4e^{i\omega t}\right) \quad \ldots \quad \ldots \quad (16)$$

Thus the omission to separate out the variation time would have led us to conclude that no negative resistance was to be expected in the temperature limited diode, since (14) contains Y_2, while (16) tells us that negative resistance effects are in fact possible, since Y_4 is involved. Conversely, in the case of the deflecting condenser, had a variation time which does not in fact exist been included in error we would have obtained a solution in Y_4 instead of the correct solution (equation (4)), which contains Y_2. It thus becomes of importance to study these two functions a little more closely.

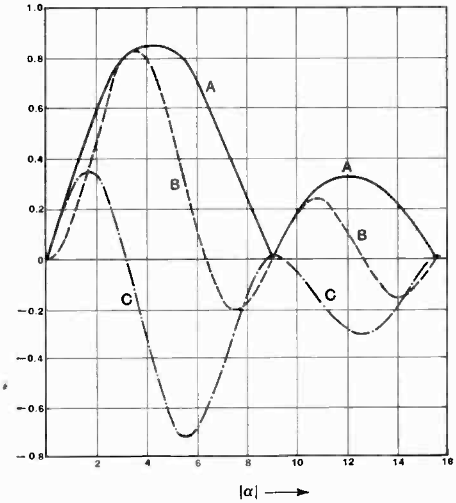
Equations (4) and (14) show that to obtain the in-phase component we take the real part of Y_2 and write $exp\ (i\omega t) = \cos \omega t$. Writing in either case

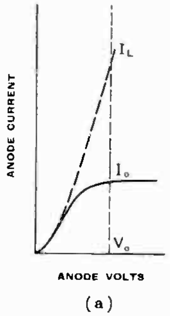
$$Y_2 = Y'_2 + jY''_2 \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

we plot Y'_2 if we want to examine the effect of frequency and $\tau_0{}^2Y'_2$ if we want to study the effect of increasing transit time, the frequency being held constant. In the case of (16), however, owing to the presence of α the real part is given by taking the imaginary part of Y_4, and is ($\xi = |\alpha|$) :—

$$Re[\alpha Y_4/3] = \frac{6}{\xi^2}[2(1 - \cos \xi) - \xi \sin \xi]$$

In this case it is necessary only to plot one curve for studying the effects of frequency and of transit time. This is because τ_0 occurs associated with each ω. We turn the function given in equation (6) upside down to give the transit angle correcting factor for the inter-deflector plate conductance. It is more convenient to write $\omega\tau_0 = 2\theta$,





(a)

Fig. 4.—*Temperature limited diode. Curves of the function* $\frac{\alpha}{3}Y_4\,(a)$; $|F|$, *curve A*; $Re[F]$, *curve B*; $Im[F]$, *curve C. Key to use of diagram. To obtain added admittance due to presence of electrons between plates of diode multiply each curve by* $I_0 I V_0$. *Then read from A the admittance, B the conductance and C the susceptance. That part of the admittance which exists even when the diode is cold is given by broken line A. In case of A multiply by* I_L/V_0 *instead of* I_0/V_0 *(see 4a). Since A gives a susceptance, that given by C must be added to obtain resultant susceptance. We see then that for* $|\alpha| > \pi$ *the capacitance is less than the cold value (as in space charge limited diode).*

when the function becomes

$$Y'_2 = Re\,[Y_2] = \frac{\sin^2\theta}{\theta^2}$$

We then plot this function (Fig. 2), which is actually already well known. The case where transit time alone (e.g. by varying gun volts) is varied (ω = constant)—requires the form $(\sin\theta)^2$ which is still better known (Fig. 3). The function Y_4, given by

$$Y_4 = \frac{6}{\alpha^3}[(\alpha - 2) + (\alpha + 2)e^{-\alpha}]$$

$$= 1 - \frac{\alpha}{2} + \frac{3\alpha^2}{20} - \frac{\alpha^3}{30} + \frac{\alpha^4}{168} \quad .. \quad (18)$$

will not be dealt with, but the function $\left(\frac{\alpha Y_4}{3}\right)$ is considered. Fig. 4 shows the real part of this function, as well as the imaginary part and the modulus—both of interest.

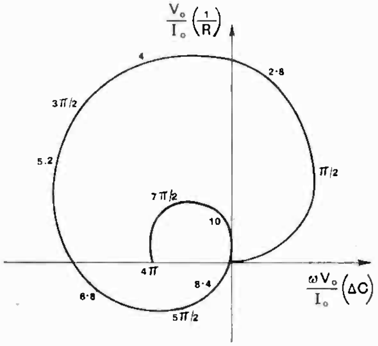


Fig. 5.—*Temperature limited diode curve of conductance vs. added susceptance.*

Fig. 5 gives a sort of circle diagram in which conductance is plotted as ordinates and susceptance as abscissae with $\omega\tau_0$ as parameter. These diagrams are self-explanatory with the help of the key on Fig. 4.

APPENDIX

Energetics of Deflecting Condenser

An electron (charge $-q$) induces on the plates a positive charge

$$q\frac{d \pm 2y}{2d},$$

the positive sign being taken for the plate nearest the electron. Assuming push-pull operation, with plates at potentials $\pm E/2$ (Fig. 1) the potential energy gained by the deflecting condenser due to a single electron is

$$+\frac{Eq}{2}\frac{d + 2y}{2d} - \frac{Eq}{2}\frac{d - 2y}{2d} = \frac{y}{d}qE \quad .. \quad (a),$$

which is also the potential energy lost by the given electron in moving from the median plane to a point distant y therefrom.

Thus, during the transit of the electron between any two points within the confines of the deflecting condenser there is no resultant change in the *potential* energy of the system: electron + condenser.

By equations (2) and (6) it may be shown that the *kinetic* energy acquired by an electron by the time it is ready to emerge from the deflecting condenser is equal to the energy "dissipated" in the effective resistance R divided by the number of beam electrons momentarily present between the plates. Thus, for the passage from entry to exit of the condenser there is no resultant change in the *kinetic* energy of the system: electron + condenser; the loss in the "leak" of the condenser being regarded as kinetic energy.

From the above it follows that as a result of a passage of an electron between entry and exit of the deflecting condenser no resultant change takes place in the *total* energy (potential + kinetic) of the system: electron + condenser.

There is, however, a resultant change in the total energy of the electron such that the latter gains energy on the whole, at the expense of the condenser, for an infinite number of frequency bands of which the first is specified by ($\xi = \omega\tau$):—

$$0 < \xi < 2\pi \quad .. \quad .. \quad .. \quad (b)$$

while, for an infinite number of bands, the first of which is

$$2\pi < \xi < 9 \quad .. \quad .. \quad .. \quad (c)$$

the electron does work on the field of the deflecting condenser. The function to which such energy is proportional for given τ is that depicted by curve B of Fig. 4. It is thought possible that here we have an explanation of absorption and radiation of electromagnetic energy on the part of a moving particle. If ω is constant and τ variable the ordinates of curve B when multiplied by $|\alpha^2|$ give the required variation of this energy with $|\alpha|$, a remark which applies to the deflecting condenser but not to the diode case.

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[1] Hollmann and Thoma, *Hochf:tech. u. Elek:akus.*, 1937, Vol. 49, pp. 145-162.

[2] Colebrook and Vigoureux, *Wireless Engineer* (correspondence), August, 1938, Vol. 15, p. 441.

[3] Bakker and de Vries, *Physica*, July, 1935, Vol. 2, p. 683.

For other references see Benham, *Proc. Inst. Rad. Eng.*, 1938, Vol. 26, p. 1093.

Applied Acoustics

By H. F. Olson and F. Massa.

The second edition of this book, which was reviewed on p. 350 of the July, 1939, issue of this journal, is now published in this country by Constable and Co., Ltd., 10, Orange Street, London, W.C.2. The price is 25s.

Wireless Patents

A Summary of Recently Accepted Specifications

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

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

511 294.—Low-frequency network for relaying broadcast programmes, with variable regulation according to the load taken.

Standard Telephones and Cables; J. R. Vezey; and A. L. Long. Application date 15th February, 1938.

512 118.—Amplifier with automatic "contrast" control which is regulated by two independent voltages derived from the higher and lower regions of a low-frequency signal.

Philips' Lamp Co. Convention date (*Germany*) *19th January,* 1938.

DIRECTIONAL WIRELESS

509 842.—Indicator device for direction finding in which provision is made for the correction of quadrantal and similar errors.

Marconi's W.T. Co.; G.M. Wright; and C. S. Cockerell. Application date 22nd January, 1938.

510 394.—Means for sharpening the critical response of a direction finder using a cathode-ray indicator.

Telefunken Co. (*addition to* 508139). *Convention date* (*Germany*) *15th September,* 1937.

511 322.—Means for preventing the radiation of horizontally-polarised waves from a directional beacon transmitter.

R. J. Berry (*communicated by C. Lorenz Akt.*). *Application date 28th June,* 1938.

512 745.—Radio-navigational system in which each of the overlapping beams is doubly modulated in order to provide both an audible and visual indication to the pilot when he strays off his course.

C. Lorenz Akt. Convention date (*Germany*) *19th January,* 1938.

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

510 360.—Means for accurately adjusting tuning-elements by push-button control.

The General Electric Co.; A. Bloch; and G. M. Wells. Application date 1st February, 1938.

510 503.—Amplifying circuit designed to give uniform gain over a wide band of frequencies, including very-low frequencies.

Philips' Lamp Co. Convention date (*Germany*) *4th January,* 1938.

510 535.—Wide-band amplifier with negative feed-back and with means for preventing undesired positive feed-back.

Telefunken Co. Convention date (*Germany*) 2nd *February,* 1937.

510 819.—Push-button tuning system arranged so that the selecting-devices can only be operated after the wave-change switch has been set.

The General Electric Co. and W. H. Peters. Application date, 18th March, 1938.

510 831.—A.V.C. system specially designed to prevent "fading" on amplitude-modulated short-wave signals.

Philips' Lamp Co. Convention date (*Germany*) *10th September,* 1937.

510 895.—Push-button tuning system in which the wave-change switch is "interlocked" with the station-selecting means so as to ensure correct operation.

Radio Gramophone Development Co. and W. R. Parkinson. Application date 11th February, 1938.

510 897.—Wireless receiver with push-button tuning, and with an alternative hand-control on a more-selective circuit arrangement.

E. K. Cole and H. A. Brooke. Application date 12th February, 1938.

510 919.—Means for regulating the damping, say of a band-pass filter, without affecting its tuning.

Philips' Lamp Co. Convention date (*Germany*) *7th June,* 1937.

511 121.—Means for reducing the power taken by the driving-motor on a push-button control system combined with automatic fine tuning.

Ideal Werke Akt. Convention date (*Germany*) *17th December,* 1937.

511 329.—Cutting-out interference from a receiver of the superhet type by utilising the automatic effect of the interference on the local-oscillator stage.

Magyar Wolframlampa Co. Convention date (*Hungary*) *21st September,* 1937.

511 386.—Visual tuning indicator of the C.R. tube type in which the fluorescent effect is produced on the side of the anode turned away from the cathode, so that it can be more clearly seen.

The M-O Valve Co. and H. S. Smith. Application date 20th May, 1938.

511 503.—Safeguarding arrangement for the driving motor of a set designed both for press-button selection and for manual tuning.

E. K. Cole and A. W. Martin. Application date 18th February, 1938.

TELEVISION CIRCUITS AND APPARATUS

(For Transmission and Reception)

510 530.—Arrangement for providing a number of selectively-operated optical paths between a film to be transmitted and a scanning device.

Baird Television; V. A. Jones; and T. C. Nuttall. Application date 2nd February, 1938.

510 531.—Method of synchronizing saw-toothed oscillation-generators which are back-coupled through a transformer.

Baird Television and L. R. Merdler. Application date 2nd February, 1938.

510 696.—Cathode-ray tube for generating television signals from a grid-structure or " screen " on which storage charges are built up by the action of a scanning stream.

J. D. McGee and H. G. Lubszynski. Application date 4th November, 1937.

510 699.—Reducing the length of the " gun " structure in a cathode-ray tube, and preventing " aberration " in the electron-focusing system.

O. Klemperer and F. H. Nicoll. Application date 4th December, 1937.

510 715.—Use of negative feed-back to provide " contrast-control " in a television receiver.

Kolster-Brandes. Application date 4th February 1938.

510 881.—Frequency-divider in which a constant phase-relation is maintained between the fundamental and a lower derived frequency, as in television scanning.

Baird Television and T. C. Nuttall. Application date 7th February, 1938.

511 048.—Controlling the " flyback " stroke in scanning so as to prevent undesired disturbances.

Baird Television (communicated by Fernseh Akt.). Application date 11th February, 1938.

511 362.—Cathode-ray tube in which a mosaic screen is scanned by an electron beam and in which the charges produced by the scanning action are restricted by an earthed grid.

Baird Television and V. A. Jones. Application date 16th February, 1938.

511 363.—Means for regulating or standardising the effective brightness of the spot formed on the fluorescent screen, particularly for interlaced systems of scanning.

Fernseh Akt. Convention date (Germany) 17th February, 1937.

511 519.—Shunt circuit designed to protect a television receiver from excessive modulation voltages due to strong pulses of interference.

Kolster Brandes ; C. N. Smyth ; and R. J. Berry. Application date 18th February, 1938.

511 600.—Increasing the available " peak " voltage in the saw-toothed form of oscillation used for scanning.

Murphy Radio and G. F. Hawkins. Application date 20th December, 1937.

511 733.—Method for producing the saw-toothed synchronising voltages used in an interlaced system of scanning.

Hazeltine Corporation (assignees of M. Cawein). Convention date (U.S.A.) 29th November, 1937.

TRANSMITTING CIRCUITS AND APPARATUS

(See also under Television)

511 014.—Separating and modulating the different types of waves which can be propagated along a transmission-line of the " dielectric-guide " type.

Telefunken Co. Convention date (Germany) 11th January, 1938.

511 079.—Method of coupling and " matching " a transmitter and receiver to a common aerial so as to prevent mutual interaction.

The General Electric Co. ; N. R. Bligh ; J. B. L. Foot ; and R. F. Proctor. Application date 25th March, 1938.

512 028.—Frequency-variable circuit, particularly suitable for the production of frequency-modulated signals, or for automatic tuning control.

W. S. Percival. Application date 24th February, 1938.

512 042.—Multiple-valve circuit with means for switching the valves on or off in a predetermined sequence, particularly for producing frequency-modulated signals.

W. S. Percival (divided out of 512 028). *Application date 24th February*, 1938.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

510 967.—Multi-grid electron-beam devices for detecting, amplifying, and generating ultra-short waves.

Standard Telephones and Cables. Convention date (U.S.A.) 31st March, 1937.

511 444.—Electron-optical focusing arrangement for the electron stream of a cathode-ray tube.

L. F. Broadway and O. Klemperer. Application date 17th February, 1938.

511 449.—Electron multiplier in which " lens " electrodes are used to direct the primary stream to the secondary-emission " targets."

Marconi's W.T. Co. and G. B. Banks. Application dates 17th February and 21st June, 1938.

511 468.—Short-wave generator of the magnetron type arranged to produce " electron " as distinct from " dynatron " oscillations.

Ideal Werke Akt. Convention date (Germany) 29th October, 1937.

511 681.—Arrangement of the photo - electric cathode and " target " electrodes, and the method of applying biasing voltages, in an electron-multiplier tube.

Radio-Akt. D. S. Loewe. Convention date (Germany) 20th November, 1936.

512 040.—Electron-multiplier tube in which the target or secondary-emission electrodes are arranged in the form of a truncated cone.

Farnsworth Television Inc. Convention date (U.S.A.) 22nd March, 1937.

SUBSIDIARY APPARATUS AND MATERIALS

510 265.—Thermionic valve circuit arranged so that the anode current decreases as the grid is made more positive.

Marconi's W.T. Co. (assignees of B. J. Thompson). Convention date (U.S.A.) 29th January, 1937.

510 707.—Loud-speaker diaphragm made from an artificial resin which has been brought into the condition of a " foam."

L. G. Farbindustrie Akt. Convention date (Germany) 15th February, 1937.

511 413.—Reinforced spindle - fitting used for mounting the operating-knobs of tuning condensers and the like.

Magneti Marelli Fabrica. Convention date (Italy) 10th March, 1938.

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