Electronic Engineering

INCORPORATING ELECTRONICS, TELEVISION AND SHORT WAVE WORLD

PRINCIPAL CONTENTS

2/- MAR., 1943

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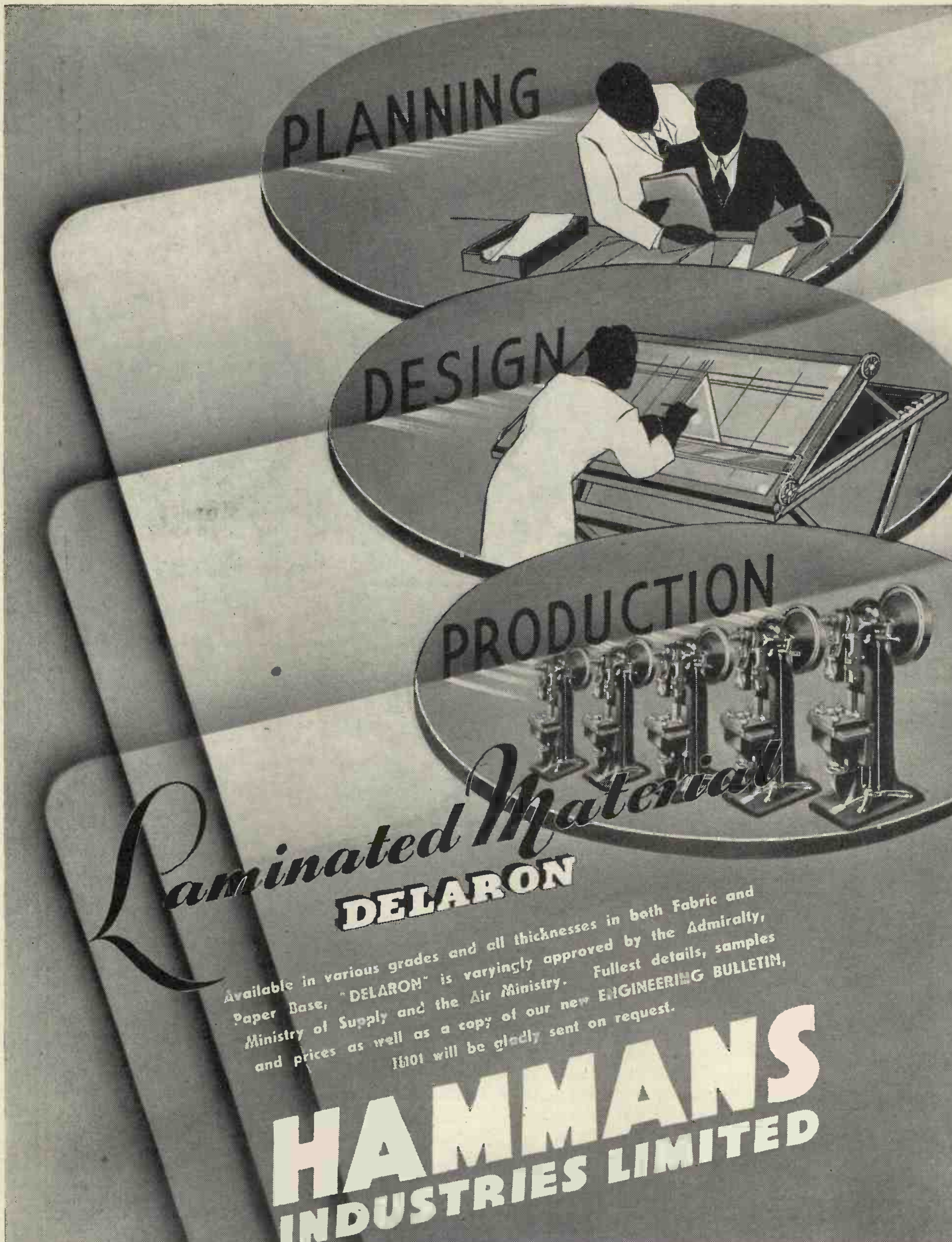

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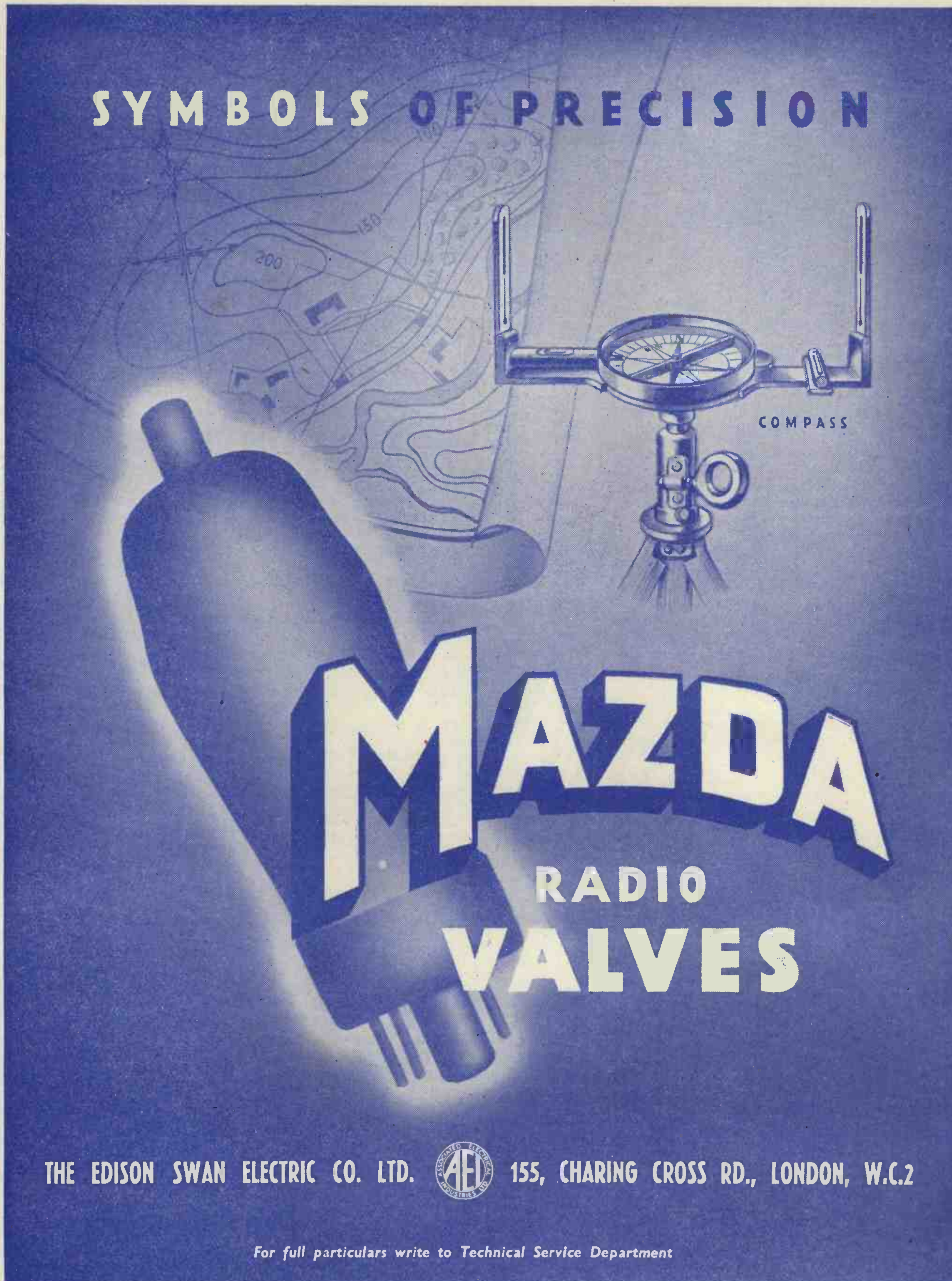
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Glossaries

TWO official glossaries of electrical terms and definitions have recently appeared, the one published by the American Institute of Electrical Engineers* and the other by the British Standards Institution† (Part 1 only).

While it is obviously unfair to draw any comparison between the publications—the American one produced without restriction on size and content, and the British one limited by war conditions—it is permissible to compare the definitions as drawn up by engineers on both sides of the Atlantic.

What is the object of a glossary? It is obviously not intended to take the place of a technical dictionary, but to crystallise the meaning of certain technical terms which are liable to confusion or to regularise others which have crept into the language. It is not always necessary to explain a term in order to define it, and acting on this dictum, the B.S.I. glossary has confined itself to as few words as possible and deals only with essential terms. This lays the Glossary open to the usual criticism that what is non-essential to some readers is of importance to others. For example, it is assumed that Ohm's Law is so familiar that it is not found among the fundamental electric and magnetic terms. But the American glossary has it between Lenz and Joule, with a rider that it does not apply to all circuits.

In the fundamental units the American glossary gives three or four pages of explanatory definitions, an example which could well be followed. These may be found in any text-book, but it is convenient to have a master reference book which will set them out clearly in relation to one another. As a mental exercise, readers might define the following without reference to a book:

Abampere, Gilbert, Maxwell, Newton, and if these are too easy, try Phot and Apostilb.

It is not possible to draw a full comparison in the electrical definitions as the B.S.I. Glossary is published in separate parts. The one of most interest to radio engineers (Terms relating to Telecommunication) which comprised sections 9 and 10 of the 1936 edition is being published separately under B.S. 204, 1943.

It is, however, embarrassing to find that the British electron is actually fatter than its American counterpart, even under war-time conditions. Here are the figures:

American definition:

An electron is the natural elementary quality of negative electricity.

The quantity of electricity on an electron is 1.592×10^{-19} coulomb, or 4.774×10^{-10} electrostatic unit.

The mass of the electron at rest is 9.00×10^{-28} gm.

British definition:

An elementary particle containing the smallest negative electric charge (4.803×10^{-10} E.S. unit) and having a mass of 9.11×10^{-28} gm. at low velocities.

In these days of brains trusts and quizzes the B.S.I. Glossary is a useful book to carry round where scientists gather. It is pretty sure to start an argument and it has the great advantage that the topics are not on the secret list.

* American Standard Definitions of Electrical Terms (*The American Institute of Electrical Engineers*, N.Y., U.S.A.) $1.25.

† Glossary of Terms used in Electrical Engineering (Section 1), (*The British Standards Institution*), BS.205, Part 1. 2/- net.

Dielectric or Puncture Strength of Porcelain and Other Ceramic Materials

By Dr. Ing. E. ROSENTHAL

THE dielectric or breakdown strength of an insulating material is that property which determines its suitability for use as a high tension insulator. The dielectric strength may be defined as the voltage gradient at which the electrical breakdown occurs. The dielectric strength of porcelain and other ceramic insulating materials—as well as that of all other solid insulating materials—is, to a very high degree, dependent upon the test conditions. It is calculated by dividing the breakdown voltage by the thickness of the test specimen between the electrodes and is commonly expressed in volts per mil or kilovolts per millimetre (1 volt per mil corresponds approximately to 25 kV per millimetre).

The test values for dielectric strength of an insulating material vary to an extent not generally appreciated with :—

(1) The thickness of material.
(2) The duration and rate of increase of the voltage applied.
(3) The characteristics of the voltage applied (frequency and wave shape).
(4) Electrostatic field distribution (edge effects, surrounding media).
(5) Temperature of the material.

In order to obtain comparable values for the breakdown characteristics of the dielectric, the conditions enumerated above must be exactly the same for the material tested.

Tests on specimens of different thicknesses, tests made with different electrodes, tests made with different rates of voltage increase, or in different surrounding media, are not comparable.

The test methods for ascertaining the dielectric strength of electrical insulating materials at power frequencies are, therefore, standardised in various countries. In the United States, for instance, there are Specifications designated D. 149/40 T. and D. 116 (Standard Methods of Testing Electrical Porcelain) and in Germany V.D.E. 0303/1929. But since wall thickness, rate of increase of voltage and the nature of electrodes are not the same, the values obtained in accordance with the A.S.T.M. Methods and the V.D.E. Methods are not comparable—the V.D.E. Method giving much higher Test Values.

The main difference between the American and V.D.E. Methods is as follows :—The American Tentative Methods specify (for porcelain) electrodes having the shape of a metal disk 0.75 in. diameter, with edges rounded to a radius of $\frac{1}{8}$ in. Plain, unrecessed, test pieces of uniform thickness are used. The V.D.E. recommends electrodes formed by metal coatings deposited on both surfaces of a recessed test disk. The test disk has, therefore, not a uniform thickness since one or two hemispherical cavities are inserted in the centre of one or both of the faces of the disk, the smallest wall thickness of the porcelain between the two metal coatings being 2 mm. (Fig. 1). On the other hand, the American Test Standards for electrical porcelain provide for a thickness of a test specimen being 0.250 in. (6.35 mm.), 0.4 in. (10.16 mm.), 0.75 in. (19.05 mm.) or 1 in. (2.54 mm.).

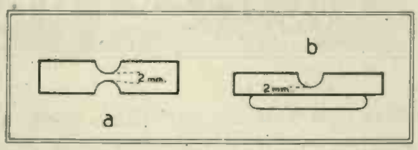


Fig 1. Test Specimens. (a) with the electrodes formed by metal coatings provided on the surface of two spherical cavities. (b) with one metal coated cavity and a metal disk with rounded edges.

The breakdown in the case of the German test specimen generally occurs across the shortest distance between the two hemispherical metal coatings inserted into the disk. Field concentrations which may occur in air or under oil around the peripheral edges of the two metal coatings have no influence on the breakdown strength because the porcelain thickness between these two edges is so much greater than the shortest distance between the two hemispherical electrodes that breakdown between these electrodes occurs before any edge effects can influence the dielectric properties of the test specimen.

In the case of test arrangements, as specified by the A.S.T.M. Methods, edge effects under oil develop between the rounded edges of the electrode and the test specimen, resulting in premature breakdown of the specimen.

Relation of Breakdown Strength to Thickness of Material

Fig. 2 illustrates the breakdown voltage of porcelain disks of varying thicknesses in different surrounding media, and illustrates the considerable influence which the surrounding media have on the breakdown voltage owing to edge effects caused by surrounding media having higher breakdown strength and lower dielectric constant than porcelain.

The same Fig. (Curve 1) shows the breakdown voltage of porcelain disks of varying thicknesses when edge effects are eliminated, that is to say, the actual breakdown strength of porcelain.

Curve 2 shows the breakdown strength of porcelain disks of various thicknesses in transformer oil of best insulating quality. Curve 3 shows the same plates in used transformer oil, and Curve 4 the same plates in low resistance oil.

In explanation of the considerable decrease in dielectric strength per mil, with increasing thickness of solid materials, many theories have been advanced. In the early days this phenomenon was explained by the assumption that it is much more difficult to produce thick-walled porcelain disks than thin-walled ones, and that the decreasing puncture strength is therefore caused by the lower quality of thicker specimens. This explanation is, however, unsatisfactory because the decrease in puncture strength can be ascertained with almost any kind of solid insulating material and even with those where thick-walled articles are easier to manufacture than thin ones. Furthermore, thin slices cut out of an insulator disk have often a higher puncture strength per mil than that of the original complete disk.

Within the last few years physicists have made pronounced progress in the study of the dielectric failure, but in spite of this progress our knowledge of the phenomena involved is still very incomplete so far as solid insulating materials are concerned.

In order to form an idea as to why the dielectric strength of solid insulating materials decreases with increasing thickness, it may be as well to discuss at this stage the theory of breakdown.

Theory of Breakdown

Dielectric failure of solid insulating materials may occur in one of the following ways, or in a combination of both :—

(a) In disruptive breakdown; and
(b) In thermal breakdown.

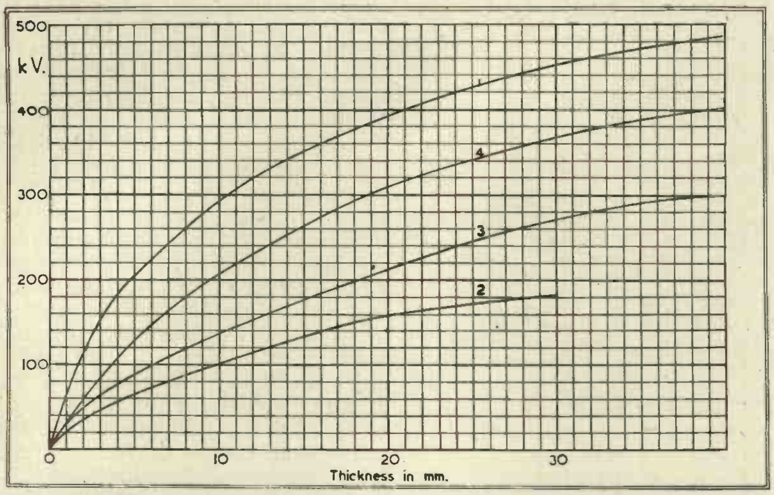


Fig. 2. Breakdown voltage of porcelain disks. (1) With no edge effects (Wagner and Weicker). (2) Plates in transformer oil (pure). (3) Plates in transformer oil (used). (4) Plates in low resistance oil.

Disruptive failure is one which results directly from an electrical overstress of the dielectric material without perceptible internal temperature rise. It is caused by ionisation and collision within the molecular structure of the material.

Disruptive failure only occurs in the case of solid insulating materials under special conditions, and accurate test values of pure disruptive dielectric strength for such materials are not easily obtained. If tests are made on very thin specimens so that heat may easily be dissipated by the electrodes and a sufficiently high voltage is applied to cause instantaneous breakdown, that there is no time for heat to develop in the thin section, the failure is purely disruptive. Impulse tests on thin ceramic sections cause a breakdown which is predominantly, or almost purely disruptive. With increasing thickness internal temperature effects modify the characteristics of disruptive failure. The breakdown strength per unit wall thickness at first decreases slowly with increasing wall thickness and then more rapidly when a certain wall thickness is reached, the breakdown showing more and more the characteristics of thermal breakdown with increasing wall thickness. Heat developed under the influence of the alternating electric field between the electrodes can less easily be dissipated when the wall thickness increases. When a certain thickness of the specimen is reached, a further increase in wall thickness will then no longer result in an increase of breakdown voltage. The curves (Fig. 2) show that this critical wall thickness, the increase of which would not cause a further increase in breakdown voltage, depends not only on the dielectric properties of the test specimen, but also on the nature of the surrounding medium. Theoretically, the disruptive breakdown voltage is proportional to the thickness of the test specimen* The curves show that for commercial frequencies and for thicknesses such as are used in actual insulation design, disruptive breakdown is the smaller and thermal breakdown the larger of the two components causing the actual breakdown. Disruptive breakdown strength is the higher the more homogeneous the structure of the insulating materials. Thermal breakdown strength is determined:

(1) By electrical conductivity (volume resistance).
(2) By thermal conductivity.
(3) Power factor.
(4) Dielectric constant of the insulating material (specific inductive capacity, or permitivity).

The higher the dielectric constant and the higher the power factor of the test specimen, the higher will be the electrical losses and more heat will be developed. The heat developed decreases the volume resistance of the material and more current will develop further heat until breakdown occurs.

It is well known that every insulator coming within the influence of an electric alternating field consumes a certain amount of electric energy and transforms it into heat. The electrical energy lost in this way is given nearly enough by the following equation:—

$$N = V^2 2\pi f C \tan \delta$$

where V is the voltage
f the frequency
C the capacity of the test specimen
$\tan \delta$ the tangent of the loss angle, or the power factor of the insulating material.

It can be concluded from this formula that insulating materials having a higher power factor possess a lower dielectric breakdown strength, particularly at high frequencies and if the voltage is applied for a long period, or increased at a small rate, *i.e.*, if the thermal component of the actual breakdown is predominant. The disruptive breakdown, however, seems to be less dependent on the power factor.

The curves illustrating the dependence of temperature on the breakdown strength of various ceramic materials show the influence of the power factor on the thermal breakdown strength. (Fig. 5).

Although the phenomena causing breakdown cannot be attributed exclusively to the two factors "disruptive" and "thermal" breakdown, there is no doubt that these are the two most important factors and the breakdowns which occur in practice are in most cases the result of these two factors.

Short-time (or impulse) tests are predominately disruptive in their nature and long-time tests, high frequency tests and tests at elevated temperatures are predominately thermal. In most cases, disruptive and thermal effects combine to produce failure. If the breakdown is purely disruptive the breakdown current increases from a substantial steady value to breakdown in a fraction of a microsecond.

Electrostatic Field Distribution and Edge Effects

Both in the actual application of insulation and in the testing of it for breakdown strength, it is not easy to avoid conditions which lead to local field concentrations and similar effects which reduce the breakdown voltage.* These are generally referred to as "edge effects" because they are observed at the edges of electrodes, although field concentrations are, of course, not limited to the edges of the electrodes. Unequal field distribution, concentration of the field, edge effects, etc., etc., have a great influence on breakdown voltage.

* E. B. Shand : Dielectric Strength of Glass. *Electrical Engineering Transactions*—August, 1941.

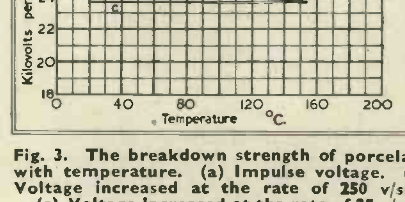


Fig. 3. The breakdown strength of porcelain with temperature. (a) Impulse voltage. (b) Voltage increased at the rate of 250 v/sec. (c) Voltage increased at the rate of 25 v/sec. *(Rosenthal-Mitteilungen 1926)*

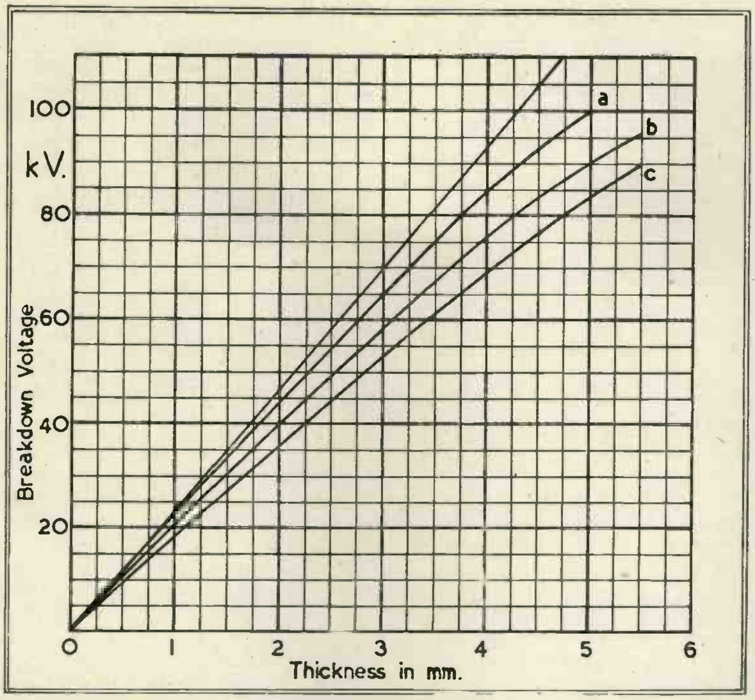


Fig. 4. Variation of breakdown voltage of Porcelain with wall thickness and rate of increase of voltage. (a) Impulse voltage. (b) Voltage increased by 250 v/sec. (c) Voltage increased by 25 v/sec.

(Rosenthal-Mitteilungen 1926)

Even in elaborate tests it is not easy to eliminate electric field concentrations completely. Many discrepancies in published data on the breakdown strengths of solid materials result from small differences in the test arrangements causing edge effects of various strengths. In addition to the concentrations at the edges of the electrodes, these effects are produced by imperfect contact between the electrodes and the insulation, by scratches in the electrode surface, when the electrode is in the form of a thin metal foil or metal coating, by included air-pockets or pores in the test specimen, and by scratches and other imperfections of the surface of the dielectric itself. A special type of edge effect is that connected with the influence of the ambient medium on the field distribution, such as insulating oil.

Dielectric tests are frequently made under oil in order to eliminate flashover of the test specimen. Insulating oils have a dielectric constant much lower than that of porcelain and glass so that the electrical stress in the oil will be correspondingly higher than in the material having the higher dielectric constant. Under these conditions, Corona discharges will form in the oil long before the breakdown voltage of the test specimen is approached. Streamers develop on the surface of the test specimen, commencing on the edges of the electrode. The higher the resistance of the insulating oil, the more concentrated are these discharges. These streamers cause intense voltage gradients along the surface of the insulation rapidly causing disintegration. If this action continues for a certain time a hole may be bored into the surface of the test specimen causing dielectric failure before the actual breakdown strength of the test specimen has been approached. This type of failure is dependent on the dielectric resistance of the oil, and, to a lesser degree only on the dielectric properties of the test specimen. (Fig. 2) shows the influence of various oils on the puncture strength of porcelain. It can be seen that the higher the dielectric strength of the surrounding medium, the lower the breakdown strength of the test specimen. These curves show very clearly the great influence which the dielectric properties of the surrounding medium have on the dielectric properties of the test specimen and that tests made in different surrounding media are not comparable.

Edge effects, also, play a very important part in actual breakdown under normal service conditions. Imperfections between the electrode and the insulator surface produce field concentration which very often is largely responsible for breakdown at voltages far below the actual breakdown voltage of the insulating material in question.

In the design of electrical insulators and components like bushings, etc., where metal parts have to be affixed to the porcelain insulator, these conditions have to be carefully considered and field concentrations avoided as far as possible. This can be achieved by suitable design of both the electrodes and the insulator and by suitable assembly methods. The application of conductive and semiconductive coatings on the surface of the insulator (particularly on curvatures with generous radii and by connecting the metal coatings with the metal work) has been used to an increasing extent during recent years in order to improve field distribution.

In this connexion reference may be made to the importance of the hard and non-attackable surface of porcelain and porcelain glazes. Surface irregularities and scratches produced on the surfaces of the insulator by metal parts during assembly, may cause edge effects which may be responsible for early breakdown. Fortunately, both porcelain and porcelain glazes are so hard that scratches caused by metal parts are practically excluded. Any imperfections of the surface (whether they be due to scratches caused by a tool or by the live metal-work under service conditions, or whether they be cavities invisible to the naked eye caused by atmospheric conditions and resulting in a matt surface) are bound to cause edge effects with the inherent detrimental influence on breakdown strength. Very few insulating materials possess the same hardness and 'unattackability' by chemical and atmospheric influences as does porcelain.

Duration and Rate of Increase of Applied Voltage

Fig. 4 shows the dependence of breakdown voltage of porcelain on wall thickness and the rate of increase of voltage. It can be seen that for thin sections the influence of different rates of voltage increase is very small. The influence of rate of voltage increase becomes more pronounced with increasing wall thickness. The breakdown strength of test specimens with great wall thickness is lower when the voltage is slowly increased, compared with breakdown strength measured at more rapid rates of voltage increase. In the case of impulse voltage, the breakdown strength is higher than in the case of tests carried out at commercial frequencies—the more so the thicker the test specimen. It can be seen, therefore, that the rate of voltage increase, or in other words the time during which the voltage is applied, is an important factor in determining breakdown voltage.

Curve (c) in this Fig. shows the breakdown voltage of porcelain disks if the voltage is increased by 25 volts per second; curve (b) if the voltage is increased by 250 volts per second, and curve (a) under a steep wave impulse. The tests at commercial frequency were made in oil, whereas the impulse tests were made in air. The results of impulse tests in oil and in air do not, however, differ very much if the tests are made under conditions which exclude edge effects. The effect of rate of increase of voltage on dielectric strength is more pronounced in the case of bakelised laminated insulating material (Fig. 5a).

The A.S.T.M. Standards provide for short-time tests and for step-by-step tests. With regard to the short-time tests, it is provided that the voltage shall be increased from zero to breakdown at a uniform rate. The rate of rise is 0.5 or 1.0 kV per second, depending on the total test time required and the voltage time characteristic of the material. The step-by-step test provides that an initial voltage will be applied equal to 50 per cent. of the breakdown voltage in the short-time test.

The voltage will then be increased in equal increments as laid down in the various material specifications. For testing electrical porcelain the A.S.T.M. Standards, however, provide no specifications for the step-by-step test.

With regard to porcelain and other dense ceramic materials, it can be stated that it does not age under the influence of electrical stress. Tests carried out show that depreciation does not take place—at least during a period of many years at power frequency. For instance, if the breakdown strength of a test specimen has been ascertained, a voltage of 10 per cent. below the breakdown voltage can be applied to it for many years without causing breakdown (of course if no irregularities are present in the test specimen).

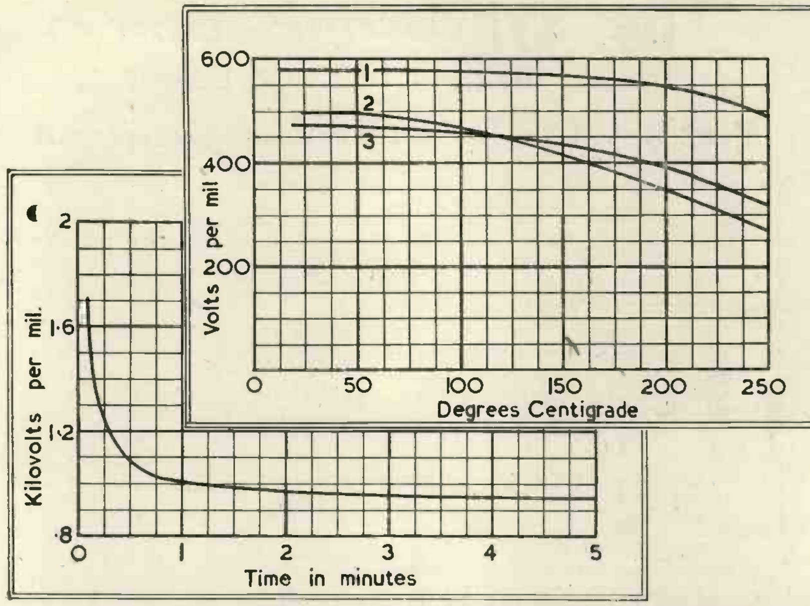


Fig. 5. Upper curves. Variation of breakdown strength of three Clinoenstatite materials with temperature. (1) AlSiMag 196 (p.f. 0.14 per cent. at 60 c/s). (2) AlSiMag 35 (p.f. 0.30 per cent. at c/s). (3) AlSiMag 192 (p.f. 0.20 per cent. at 60 c/s).

Fig. 5a. Lower Curve. The effect of rate of increase in voltage on the dielectric strength of black varnished cambric.

(*A.S.T.M. Standards*, p.69).

Effect of Voltage Characteristics on Breakdown Strength of Ceramics

With direct current the breakdown voltage of porcelain is 20-30 per cent. higher than with alternating current at commercial frequencies.

The breakdown strength of solid materials is lower at high frequencies than at low frequencies. It has been mentioned that as the frequency is increased, the dielectric losses increase and the temperature in the test specimen increases. As a consequence of the high temperature developed in the test specimen, the breakdown voltage drops. The power factor of the dielectric plays a very important part on the dielectric strength of the materials at high frequencies. For instance, the breakdown strength of improved Clinoenstatite bodies is 35-45 kV per mm. at 50 cycles, and 25-27 kV per mm. at 1 Mc/s. The breakdown voltage of this special type of ceramic body drops, therefore, only by about 40 per cent. (breakdown voltage at 1 Mc/s. compared with that at commercial frequency). In the case of porcelain the breakdown voltage drops by 50-60 per cent. and in the case of most glasses by 70 per cent., under corresponding conditions.

The influence of wall thickness on breakdown strength is more important at high than at low frequencies. If, by employing thin sections and massive electrodes the heat developed by the high frequency field is quickly dissipated, the breakdown strength per unit is higher than otherwise. There are puzzling exceptions to the general rule that the breakdown strength is higher at low frequencies than at high frequencies. Rutile bodies, for instance, have higher breakdown strength at high frequencies than at low frequencies. This refers to certain rutile bodies, the power factor of which is 100 times higher at 800 cycles than at 1 Mc/s., and 300 times higher at commercial frequencies than at 1 Mc/s. In such very special cases the breakdown at low frequencies is thermal and at high frequencies disruptive in character.

Influence of the Temperature of the Material on Dielectric Strength

Fig. 3 illustrates the dependence of breakdown strength of porcelain on temperature variations between 20-160° C. It can be seen that within this temperature range the breakdown strength of porcelain at power frequency decreases only very slightly, and the less so the slower the voltage increase. In the case of impulse voltage the decrease with increasing temperature is more noticeable. From these curves it can be inferred that at a temperature of 160° C. the breakdown is purely thermal under all voltage conditions.

Fig. 5 shows the dependence of breakdown strength of three steatite articles at temperatures varying between 25-250° C. The decrease in breakdown strength between room temperature and about 100° C. is not considerable in the case of two of the materials both possessing a very low power factor.

The decrease in dielectric strength between room temperature and 250° C. is only about 15 per cent. in the case of the material AlSiMag 196, owing to the extremely low power factor and high volume resistance at elevated temperatures of this special material. The same breakdown values at high temperatures are attained with other Clinoenstatite type materials having the same power factor and volume resistance. Breakdown strength at temperatures higher than 80° C. is, generally speaking, better the higher the volume resistivity at elevated temperatures and the lower the power factor of the material under test.

The Synchronisation of Oscillators

By D. G. TUCKER, B.Sc. (Eng.), A.M.I.E.E.*

Part I.—The Direct Synchronisation of Feedback Oscillators

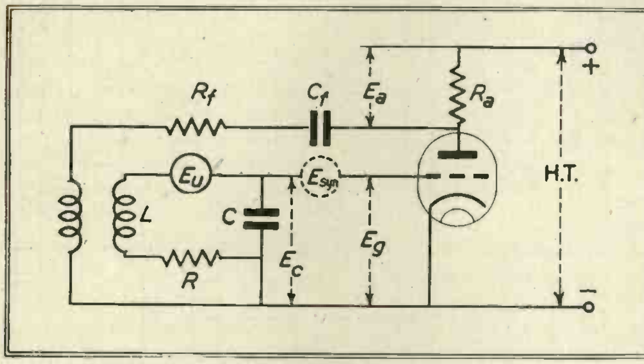


Fig. 1. Basic Circuit of Inductance-Capacitance Tuned Feedback Oscillator.

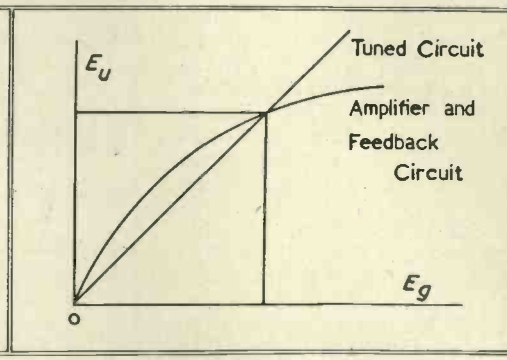


Fig. 2. Loop Characteristics of Oscillator of Fig. 1.

1. Introduction

IT is probably well-known that oscillators may be "locked" in frequency to some external controlling signal; but the theory and the practical problems encountered are comparatively little-known, in spite of the large amount of mathematical analysis that has been published on the subject. The main trouble seems to be that in recent years little experimental work has been carried out. It is hoped that the present article will contribute to the practical understanding of the subject.

The best-known example of oscillator synchronisation is perhaps that involved in the time-base circuit of a cathode-ray oscillograph. Here a considerable variety of circuit arrangements may be found,[1] but in general the time-base will be derived from a relaxation oscillator producing a "saw-tooth" wave-form. Such an oscillator is readily synchroniseed to an external frequency not greatly different from its own natural frequency, and the usual method of applying the control is, in the case of an oscillator using a gas-filled triode valve, to connect the synchronising tone or signal to the grid. If the natural (*i.e.*, unsynchronised) frequency of the oscillator is approximately $1/n$th of the external frequency, it is possible to control the oscillator so that its frequency is exactly $1/n$th of the external frequency. However, the range of control diminishes as the ratio n increases, and in practice ratios above 10 are avoided.

The operation of this control is quite simply explained by graphical methods, and involves no more than the idea of the applied signal "triggering" the gas-filled valve and thus controlling the charge-discharge sequence of the relaxation circuit. Several articles have been published, all explaining it in the same manner[2] and consequently it is not proposed to go into further detail here.

The multivibrator[3] is another type of relaxation oscillator which is very readily synchronised to an external frequency, or to a sub-multiple of it. This circuit is much used for frequency division, with ratios generally not higher than 5, although better results can now be obtained with the quasi-stable frequency divider[4] using what the Americans call "regenerative modulation." This process will be dealt with in a later part of this article. The synchronisation of a multivibrator is effected by injecting the control signal into the grid or anode circuit; the mechanism of the control of frequency is basically the same as for the single-valve relaxation oscillator, and has been adequately described in the literature.[3]

The remainder of this article will be devoted to the synchronisation of oscillators other than relaxation oscillators. Of these other types, the most important is the simple inductance-capacitance tuned feedback oscillator, and this will be considered in detail.

2. The Direct Synchronisation of the Inductance - Capacitance Tuned Feedback Oscillator

2.1. Methods of Approach to the Problem

An investigation into the mechanism of synchronisation of an oscillator may be made either by a mathematical analysis or by a graphical process. In the former case the working is usually rather complex and involves difficult differential equations. The graphical method is more useful for general purposes and enables quite simple mathematical relationships to be deduced using readily comprehensible simplifications—this method will be adopted in this article. The mathematical method has been published in various forms by a number of authors. One of the best treatments is that published by E. V. Appleton as early as 1923,[5] and the results there obtained are basically the same as those derived in this part of the present article.*

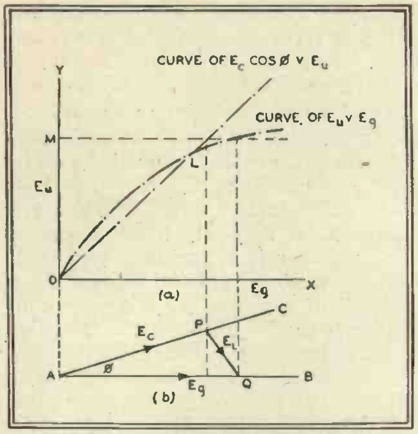
2.2. Graphical Treatment of the Problem.

The circuit arrangement of a conventional type of inductance-capacitance tuned feed-back oscillator is shown in skeleton form in Fig. 1. Such an oscillator may be synchronised by the injection of the control frequency into the grid circuit. We must first consider the operation of the unsynchronised oscillator, and then determine the changes that occur when it is synchronised. R is the resistance of the tuning inductance L, and C is the tuning capacitance. E_a is the voltage developed across the anode load R_a. E_u is the e.m.f. induced in L from the feedback circuit, and E_g is the voltage developed between grid and cathode, *i.e.*, the voltage across C. These voltages refer to a steady oscillation, and do not, of course, include any d.c. component.

It is convenient to consider the oscillator loop circuit in two parts, one, the amplifying portion from the grid of the valve to the terminals of the inductance, *i.e.*, from voltages E_g

* P.O. Research Station

*The early work on the subject of synchronisation of oscillators was concerned with the mutual effect of two wireless transmitters in relative proximity to one another.



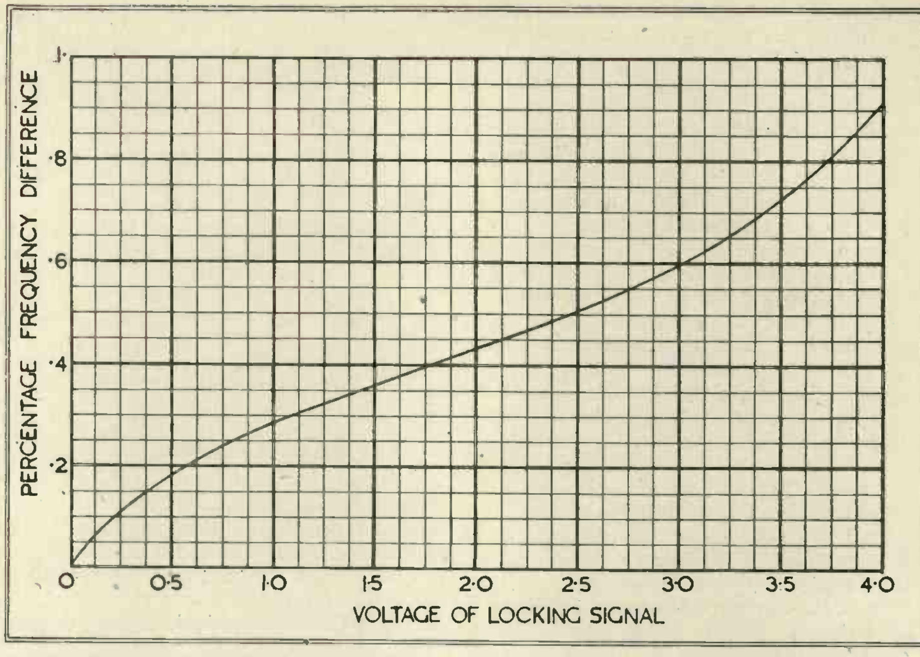


Fig. 3. Graphical Construction.
Fig. 4. Typical Locking Characteristic; 1 kc/s oscillator with Q = 15, and Eg = 1 volt.
(From P.O.E.E.J.)

to E_u; and the other, the tuned circuit, *i.e.*, from voltages E_u to E_g. The first is non-linear in that the amplification of the valve decreases as the applied voltage increases. The second may be considered linear for the present purpose, although in practice, if the inductance is iron-cored, there is a slight non-linearity here also.

Fig. 2 shows the voltage relations of the loop circuit graphically. If oscillation is to take place, it is clearly essential that the two curves should intersect at some point (representing the condition of stable oscillation) remote from the origin; in other words, the resultant loss or gain round the loop must be zero in the steady state. The greater the gain of the valve at low voltages, the greater is the voltage to which the oscillation will build up, in order that the loop gain may be reduced to zero. Another fundamental requirement is that the phase shift round the loop shall be zero or a multiple of 360°. This can only be adjusted by a slight change in frequency, which changes the phase relation between E_g and E_u in the tuned circuit. This requirement explains why oscillators rarely oscillate at the natural frequency of the tuned circuit. Evidently, it will be necessary to determine the frequency of oscillation from this phase-shift requirement before the graph of E_g against E_u for the tuned circuit can be plotted in Fig. 2. Having determined the mode of oscillation of the unsynchronised oscillator, let us next consider the circuit relationships if a synchronising or "locking" signal,* E_{syn} is injected into the grid circuit, as shown in Fig. 1. If the natural frequency of the oscillator is $\omega_0/2\pi$, then the phase difference ϕ_0 between E_u and the current in the tuned circuit, when the oscillator is unsynchronised is given by $\tan\ \phi_0 = (\omega L - 1/\omega_0 C/R$. But if the oscillator is forced by the injected signal to oscillate at a different frequency $\omega_{syn}/2\pi$, then the phase difference becomes ϕ_1 where $\tan\ \phi_1$

$$= \frac{\omega_{syn}L - 1/\omega_{syn}C}{R}$$

Referring now to Fig. 3, which shows in (a) the loop gain characteristic, we can construct a vector diagram as shown in (b). AB is taken as the direction of the vector representing the voltage E_g on the grid. By forcing the oscillator to change its frequency, we have changed the phase shift in the tuned circuit by an amount $\phi = \phi_1 - \phi_0$ and consequently the voltage E_c across the condenser differs in phase from E_g by an amount ϕ. The direction AC of the vector representing E_c can now be indicated in Fig. (3b).

It is convenient to arrange that point A of the vector diagram is in line with OY, and to replot the straight line graph to represent the relation between $E_c \cos\phi$ and E_u. Let the intersection of this line with the curve be L. If any horizontal line through a point M on OY is drawn to cut the curve and straight line above L, and from the point of intersection with the straight line a perpendicular is dropped to the E_c vector line at P, and from the point of intersection with the curve a perpendicular is dropped to the E_g vector line at Q, then complete vectors E_c = AP and E_g = AQ are obtained corresponding to a certain value of E_u. It is evident that to "lock" the oscillator stably in this particular condition at this frequency $\omega_{syn}/2\pi$, it is necessarey to add a synchronising vector E_{syn} = PQ to complete the vector triangle.

If less synchronising voltage than that corresponding to PQ is injected, then the E_c and E_g vectors adjust themselves to a smaller value, and OM becomes smaller. When the horizontal line passes through L, the vector E_{syn} is vertical. This condition represents the boundary of the synchronised range of operation; if E_{syn} is reduced further, the oscillator will not synchronise. The vertical condition where synchronisation fails is generally referred to as the "pull-out."

2.3. Conclusions from the above work

The vector diagram determined above (which was first published by U.Bab.[6]) does not, of course, explain how "pull-in" and "pull-out" occur, but it does demonstrate clearly the magnitudes and phase angles obtained in a synchronised oscillator. The following properties of the circuit will be noted:—

(a) The greater the locking voltage E_{syn}, the greater can be the angle ϕ; and since ϕ is a measure of the difference between the natural frequency of the oscillator and the control frequency, it is clear that the greater is E_{syn} the greater is the frequency range over which synchronisation is possible.

(b) Since the output of the oscillator will normally be taken from across the anode load, we can consider the phase of the output to be determined

* Throughout the article the word "synchronising" and "locking" will be regarded as synonymous.

exactly by the phase of E_g. The phase of E_g relative to E_{syn} is 90° at pull-out. If E_{syn} remains constant, and the frequency difference is reduced, then ϕ is reduced, and the angle PQA becomes acute. When the frequency difference is zero, E_{syn} and E_g are exactly in phase. As the frequency difference is increased again in the opposite direction ϕ becomes negative, and the phase difference between E_g and E_{syn} is now of opposite sign, reaching a limit of $-90°$ at pull-out.

These two properties of the synchronised oscillator generally determine the practical design of synchronised systems. It is useful to plot them for any particular oscillator as

(a) a locking characteristic; this shows the difference between the natural and control frequencies plotted against the voltage of control frequency required just to effect pull-in. A typical curve measured on a certain 1 kc/s. oscillator is shown in Fig. 4.

(b) a phase characteristic; this shows the phase difference between E_g and E_{syn} plotted against the frequency difference at a fixed voltage of control frequency. The marked points in Fig. 5 show a typical measured phase characteristic; the dotted curve is a calculated characteristic. (*See* paragraph 5.2).

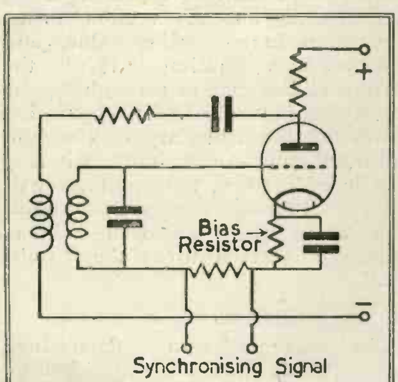


Fig. 8. Injection of Synchronising Signal in a Practical Case.

In many practical cases, it is necessary only to ensure that the oscillator never, or rarely, pulls out of synchronism. The drift of the natural frequency of the oscillator can be estimated or measured for the periods of time, and the voltage and temperature conditions involved, and the amount of control signal required to ensure continuous synchronism can then be readily estimated from the locking characteristic. In some instances, however, it is necessary to maintain phase relations within certain limits, and then it will be necessary to inject a greater voltage of control signal; the required magnitude

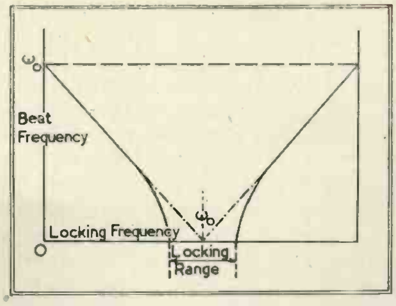


Fig. 7. Beat Frequency of Oscillator with Injected Tone.

can be determined by considering the phase characteristic as well as the locking characteristic.

3. The Effect of Pull-Out

When the voltage of the locking signal is just insufficient to synchronise the oscillator, the oscillator frequency does not assume its natural value, but varies continuously in a regular cyclic manner. A useful demonstration of this is obtained by beating the output of the oscillator with the locking signal itself. Typical beats obtained are shown in Fig. 6, which shows a series of oscillograms of the beat of a 6 kc/s. oscillator. In the first, the oscillator tuning condenser has been rotated until the oscillator just pulls out of lock. The beat obtained is just a series of slow impulses. In between these impulses the oscillator is apparently synchronised, and the impulse represents a " slip " of synchronism. As the condenser is further rotated, the beat frequency increases, although the impulse itself retains the same form. Finally, the beat note becomes almost a sine-wave, and has a frequency nearly equal to the difference between the injected frequency and the natural frequency of the oscillator. Fig. 7 shows this effect in graphical form. If the external tone were not injected as a locking signal into the oscillator, but merely used to obtain a beat-frequency in some other man-

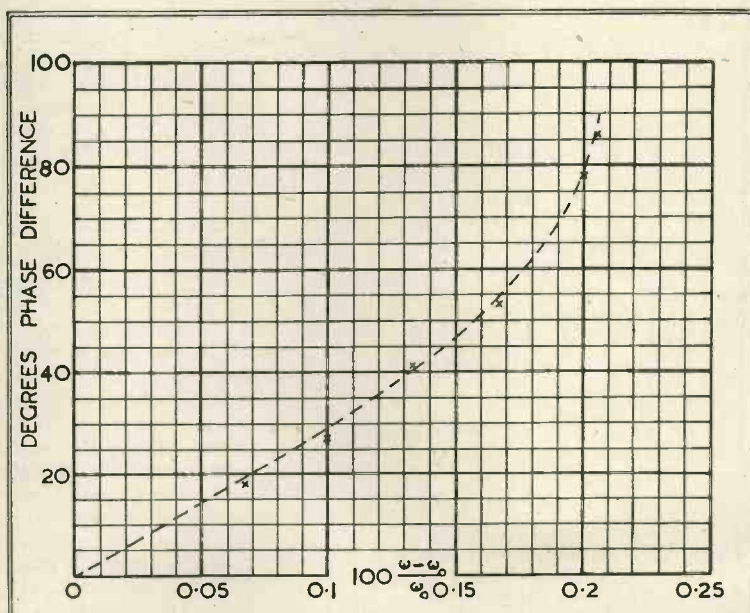


Fig. 5. Phase Characteristic. Crosses are measured values on a 6 kc/s oscillator with Q = 125. Dotted curve is calculated from eqn. 3.

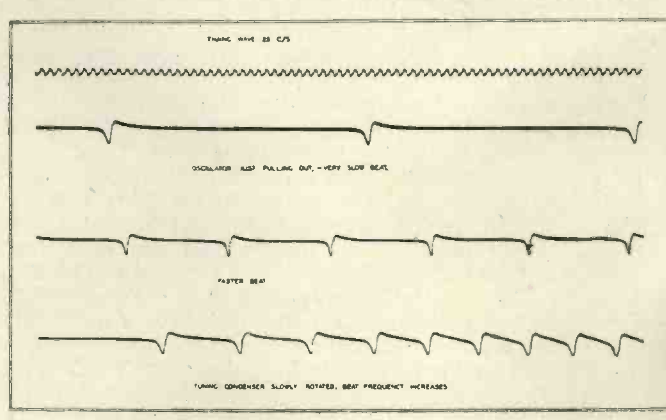


Fig. 6. Effect of Pull-Out.

(*From P.O.E.E.I.*)

ner, the curve of beat-note against external frequency would be merely two straight lines at 45° to the axes, as shown dotted. But when the external tone is used as a controlling signal, the effect is as shown in full lines, and over the locking range, the beat-note is, of course, zero. Some interesting measurements on this beat-note have been published by Subra.[7]

4. Method of Injection of Synchronising Tone

Although in Fig. 1 the synchronising voltage is shown injected directly between the tuned circuit and the grid, this connexion would not generally be desirable in practice. It is better to inject the voltage across a resistance connected between the tuned circuit and the H.T. negative, as shown in Fig. 8. The value of the resistance can be determined by the nature of the circuit supplying the synchronising signal.

There are other methods of injecting the synchronising tone; one that is sometimes convenient at high frequencies is to couple the synchronising circuit to the tuning inductor by means of a loosely coupled coil.

5. Simple Quantitative Relationships in Synchronised Feedback Oscillators

5.1. Dependence of the Locking-range Characteristics on the Circuit Parameters

Some very useful quantitative relations may be deduced from the preceding work. We have stated, $\phi_o = \dfrac{\omega_o L - 1/\omega_o C}{\omega_o}$ where $\omega_o/2\pi$ is the natural frequency, and $\tan\,\phi_1 = \dfrac{\omega_{syn} L - 1/\omega_{syn} C}{R}$ where $\omega_{syn}/2\pi$ is the controlled frequency.

Now the angle ϕ of the vector diagram (Fig. 4) is given by $\phi = \phi_1 - \phi_o$. If all these angles are small, we may assume that $\tan\,\phi = \tan\,\phi_1 - \tan\,\phi_o$ so that

$$\tan = \frac{\omega_{syn}L - 1/\omega_{syn}C}{R} - \frac{\omega_o L - 1/\omega_o C}{R}$$

$$= \frac{\omega_{syn} - \omega_o}{R}\left[L + \frac{1}{\omega_o\omega_{syn}C}\right]$$

It is usual to refer to the Q value of a coil rather than to its resistance; therefore we can substitute $1/R = Q/\omega_o L$ for convenience. Also, if $\omega_{syn} - \omega_o$ is small, we can replace $1/\omega_o\omega_{syn}C$ by $1/\omega_o^2 C$, and since $1/\omega_o^2 C = L$ very nearly, we obtain

$$\tan\,\phi = \frac{Q(\omega_{syn} - \omega_o)}{\omega_o L}\,(2L). = \frac{Q(\omega_{syn} - \omega_o)}{\omega_o} \quad \ldots \quad (1)$$

In the majority of practical cases the voltage of synchronising signal required just to lock the oscillator is the important value in the design of the synchronising system; only where great phase stability is required will any other value be a design factor. At the limit of the locking range, the

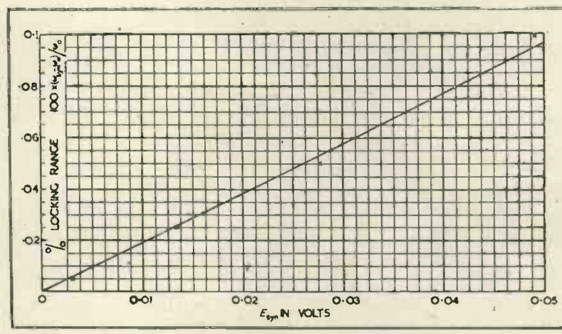
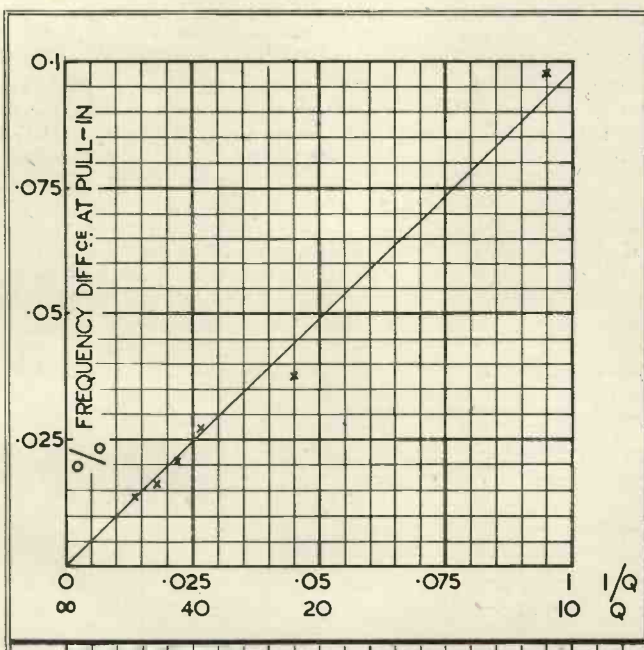
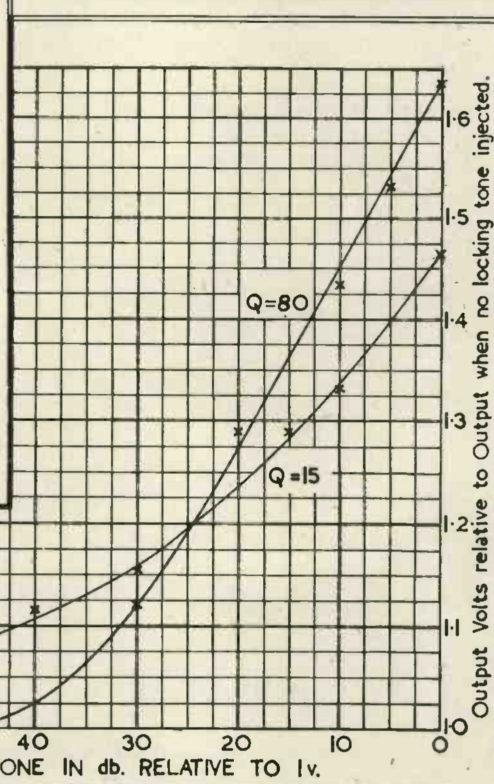


Fig. 9 (right). Locking Characteristic of 4 kc/s Oscillator Q=50, Eg=0.5 volt.

Fig. 10 (below). Locking Range in Relation to Q. Eg = 1 volt.

Fig. 11. Increase in Output as Locking Tone Increased. Eg = 1 volt.





vector E_{syn} is perpendicular to E_g; let the limiting value of E_{syn} by $E_{syn\cdot 1}$. Then

$$\frac{E_{syn\cdot 1}}{E_g} = \tan\phi$$

So that we now have

$$E_{syn\cdot 1} = 2Q\ E_g.\frac{\omega_{syn} - \omega_0}{\omega_0} \quad \dots \quad (2)$$

Q is known, but it should not be forgotten that allowance must be made for the damping of $R_f + R_a$ on the tuned circuit. E_g is known or can be determined for any particular valve and loop gain characteristic. The grid voltage is not constant over the locking range, and the value of E_g in equation (2) is strictly the value just before pull-out; but fortunately this value is related to the amplitude of free oscillation independently of E_{syn} or Q or $(\omega_{syn} - \omega_0)/\omega_0$—it is 0.707 × the free amplitude for a cube law valve characteristic (*see* Appendix). Thus E_g above is somewhat less than the r.m.s. voltage required to cause the valve to run into grid current. Thus for a given oscillator circuit $E_{syn\cdot 1} \propto (\omega_{syn} - \omega_0)$ for small locking ranges. It is, in practice, found that the locking characteristic is a straight line for the usual small ranges, say 0.1 per cent. on the frequency scale of a normal *LC* oscillator, as may be seen in Fig. 9.

Another important result to be observed from the equation above is that for a given locking voltage, but a variable Q, we obtain $(\omega_{syn} - \omega_0) \propto 1/Q$ which means that the greater the Q (*i.e.*, the lower the resistance of the coil), the smaller is the range of frequency over which the oscillator can be synchronised. Fig. 10 shows a graph of this relation actually measured on a working oscillator, in which the tuned circuit could be damped by a parallel resistance as required. For every reading the loop gain was adjusted so that the circuit just oscillated gently, and E_g was kept constant. The valve used was type SP41 (Mazda). E_{syn} was kept at 0.025 volts; E_g was about 1 volt. The measured results agree reasonably closely with equation (1) for small locking ranges. The tests were carried out at about 10 kc/s, but the locking range has been plotted as $(\omega_{syn} - \omega_0)/\omega_0 \times 100$, in order to express it in a convenient form independent of the actual frequency.

5.2. Derivation of the Equation to the Phase Characteristic

Referring again to the vector diagram of Fig. 3, it is seen that the phase angle between the locking signal and the resultant grid voltage is ∠PQA, which we may designate θ. Let $\omega/2\pi$ be any synchronised frequency and $\omega_1/2\pi$ be the frequency of locking signal at which pull-out occurs. The natural frequency of the oscillator is $\omega_0/2\pi$, as before.

From the vector triangle,

$$\frac{E_{syn}}{\sin\phi} = \frac{E_g}{\sin(\theta + \phi)}$$

since $\sin(\pi - \theta - \phi) = \sin(\theta + \phi)$ so that $\sin\theta\cos\phi + \cos\theta\sin\phi = (E_g\sin\phi)/E_{syn}$.

Since ϕ is small, $\sin\phi \doteqdot \phi$ and $\cos\phi \doteqdot 1$ thus $\sin\theta + \phi\cos\theta = E_g\phi/E_{syn}$

Using equation (1) of the preceding section, we obtain

$$\sin\theta + \cos\theta . 2Q\frac{\omega - \omega_0}{\omega_0} = \frac{E_g}{E_{syn}} . 2Q\frac{\omega - \omega_0}{\omega_0}.$$

and equation (2) gives us (changing symbols as necessary)

$$\frac{E_g}{E_{syn}} . 2Q = \frac{\omega_0}{\omega_1 - \omega_0} \quad \text{and}$$

$$2Q = \frac{E_{syn}}{E_g} . \frac{\omega_0}{\omega_1 - \omega_0}$$

so that

$$\sin\theta + \cos\theta\frac{E_{syn}}{E_g} . \frac{\omega - \omega_0}{\omega_1 - \omega_0} = \frac{\omega - \omega_0}{\omega_1 - \omega_0}$$

and since E_{syn}/E_g is generally small, and since as $\cos\theta$ increases

$$\frac{(\omega - \omega_0)}{(\omega_1 - \omega_0)}$$

decreases, we have finally (and approximately)

$$\sin\theta = \frac{\omega - \omega_0}{\omega_1 - \omega_0} \quad \dots \quad (3)$$

So that the phase characteristic as described earlier and illustrated in Fig. 5 should be very nearly a sine-wave. That this is so in the case of Fig. 5 may be seen by comparing the measured curve (marked with crosses) with the curve calculated from equation (3) (shown in dotted lines). The agreement is very close.

6. Considerations of Oscillator Output Amplitude

It was seen from the graphical analysis of Fig. 3 that the working points on the loop characteristic curves were altered by the addition of the synchronising signal. This means that the output of the oscillator varies as the synchronising tone is varied, both in magnitude and frequency, relative to the free oscillation. The exact law relating the output ampli-

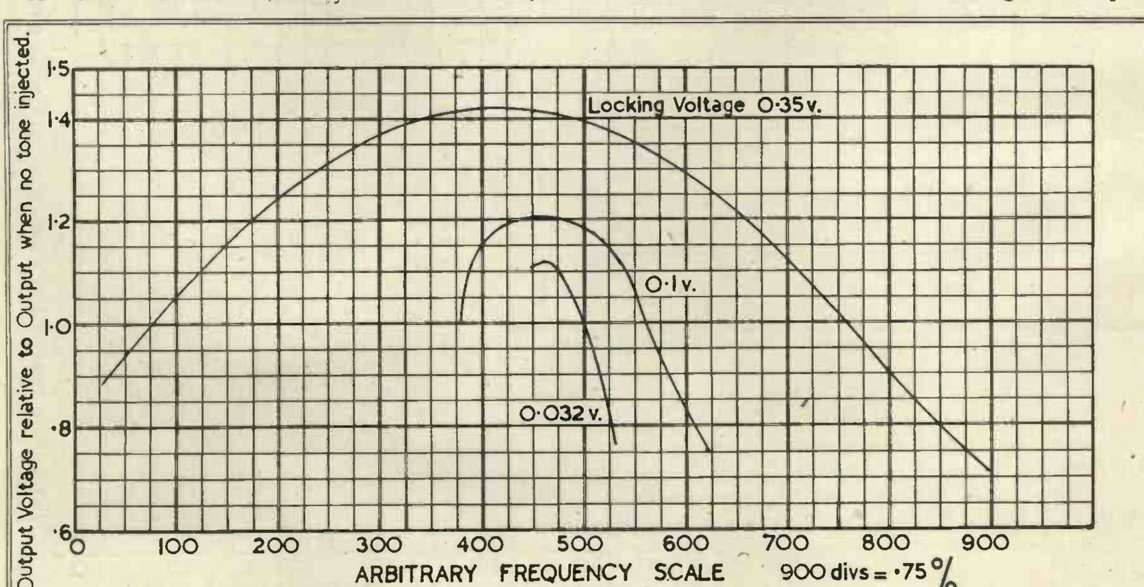


Fig. 12. Variation in Output over the Locking Range. Q = 80.

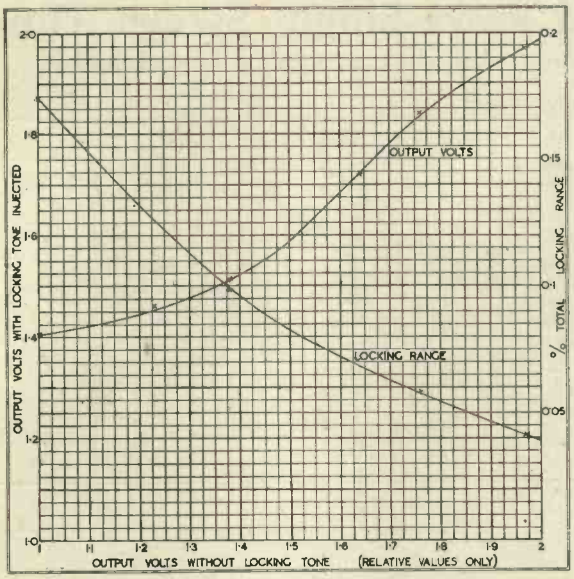


Fig. 13. Effect of Natural Amplitude on Locking Range and Output. Q = 80, Esyn = 0.1 volt, fo = fsyn.

tude with the other circuit conditions is difficult to work out in a general form, owing to its mathematical complexity, but van der Pol[8] has shown how it may be done, and has given a very valuable series of results deduced from his working. The variation of output (which below the overload point is readily related to the grid voltage E_g) can be worked out, for any given valve characteristic, by the graphical construction of Fig. 4, but a considerable amount of work is often required to obtain each separate point, and the method is therefore liable to be very laborious.

In practice, the theoretical results are rarely obtained, owing to secondary effects which occur, for instance, overloading and the change of natural frequency which occurs due to a changed harmonic production in the valve, or the change of inductance (if iron-cored) with amplitude. It will be sufficient, therefore, for present purposes, to discuss some typical experimental results, which show adequately the type of relationship to be expected. For most purposes, these variations of output are only of secondary importance. The test oscillator used a valve Mazda type SP41 with 150 volts H.T., the maximum r.m.s. grid volts for no overload being 1 v.; the anode load was such that grid and anode overload occurred very nearly at the same time.

6.1. Variation of Output as the Locking Voltage is Varied

In this case we consider the change in the output voltage of the oscillator as the locking voltage is increased from very small to large values, *the natural frequency of the oscillator being equal to the control frequency.* The measured output of the test oscillator is shown in Fig. 11, the locking voltage being plotted in decibels below 1 volt in order to accommodate a large range of values. The output is expressed in terms of the output obtained when the locking tone is absent. The value of E_g for the free oscillation was in this case very nearly 1 volt; if a smaller amplitude of free oscillation is used, the change of output is greater, even for the same ratio of E_{syn}/E_g, owing to the reduced nonlinearity of the valve at smaller amplitudes. The effect of the Q value of the tuned circuit is a secondary one only.

If secondary effects are neglected, this relation of output to locking voltage can be calculated fairly readily provided the valve characteristic can be expressed as a simple power series. Suppose the law is a cubic, thus,

$$E_c = \alpha E_g \sin \omega t - \beta (E_g \sin \omega t)^2 - \gamma (E_g \sin \omega t)^3$$

and that free oscillation takes place at the natural resonance of the tuned circuit. We are concerned only with the terms in the equation which are at the fundamental frequency. Now

$$E_g^2 \sin^2 \omega t = \tfrac{1}{2} E_g^2 - \tfrac{1}{2} E_g^2 \sin 2\omega t$$

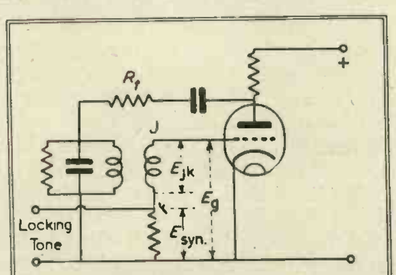


Fig. 14.

$$E_g^3 \sin^3 \omega t = \tfrac{3}{4} E_g^3 \sin \omega t - \tfrac{1}{4} E_g^3 \sin 3\omega t$$

so that, neglecting harmonics and d.c. components,

$$E_c = (\alpha E_g - \tfrac{3}{4}\gamma E_g^3) \sin \omega t.$$

We may now, for simplicity, drop the " sin ωt " and deal only in magnitudes.

If E_{go} is the grid amplitude of the free oscillation, we have

$$E_{go} = \alpha E_{go} - \tfrac{3}{4}\gamma E_{go}^3$$
$$i.e., \ \alpha - 1 = \gamma E_{go}^2$$

With the locking voltage present, since E_{syn} and E_g are in phase, we have

$$E_g = E_{syn} + E_c = E_{syn} + \alpha E_g - \tfrac{3}{4}\gamma E_g^3$$
$$i.e., \ \tfrac{3}{4}\gamma E_g^3 - \gamma E_{go}^2 . E_g - E_{syn} = 0$$

This can be solved analytically, or graphically, to give E_g in terms of E_{go} and E_{syn}. Thence, by applying once more the non-linear valve equation, the output voltage E_a can be determined. The same result would be obtained, of course, by the graphical construction described earlier.

6.2. Variation of Output over the Locking Frequency Range

If we again refer to the output of the oscillator as unity when no locking tone is injected, then in the middle of the range the output will be increased above unity by the injection of the locking signal, but will be decreased below unity at the edges of the locking range. This variation is larger for larger locking voltages, and its extent is determinable approximately from the results of the previous paragraph. The grid amplitude at pull-out is theoretically 0.707 times the free amplitude, except for minute values of E_{syn}, and since the nonlinearity is small at this reduced voltage, all curves should show a relative output of, say, 0.75 at pull-out.

Fig. 12 shows three measured curves for an oscillator with Q = 80 and natural frequency 8 kc/s. It will be seen that the maximum overall variation is less than 2 : 1 even in the case where 0.35 volt of locking signal is used; this voltage represents about one-third of the free grid amplitude, and may be considered a fairly high locking voltage. If a smaller free amplitude is used, the variation in output is larger, but the curves become more symmetrical. This suggests that the very noticeable lack of symmetry in the curves shown is due to the fact that harmonic production in the valve has caused the natural frequency to differ considerably from the resonant frequency of the tuned circuit.

6.3. Change of Locking Effect as the Natural Amplitude is Varied

In practice it is generally best to operate an oscillator with as little feedback as possible, *i.e.*, to have it just oscillating gently, in order to obtain the best inherent stability. But it is sometimes necessary to allow a much larger natural amplitude, and it is important then to see how this affects the locking performance. For convenience, the output of the oscillator may be called unity in the " gently oscillating " condition. Fig. 13 shows the output volts when a locking signal of 0.1 volt is applied to the oscillator whose natural (*i.e.*, unsynchronised) output amplitude is varied from 1 to 2; this indicates that the effect of the locking signal diminishes rapidly as the natural amplitude increases—a result only to be expected, of course. The change of the locking range with natural output amplitude is also shown. Equation (2) shows that the locking range is inversely proportional to the grid amplitude, and it will be seen that considering the valve non-linearity, the curve supports this.

7. The Effect of Taking the Output Across the Tuned Circuit instead of From the Anode

In practice it is advantageous to take the output to a buffer amplifier from the terminals of the tuned circuit. This gives a pure sine wave, although the output amplitude is less.

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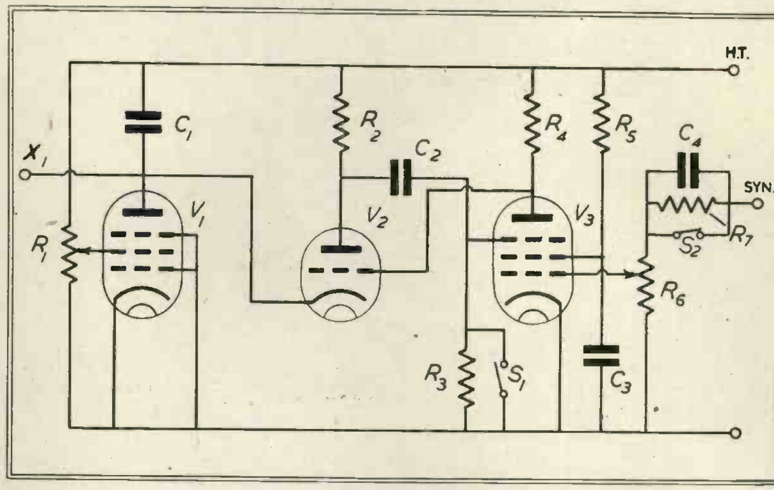
A Hard Valve Single Sweep Time Base



FOR a number of investigations with the cathode-ray tube, and particularly with transient phenomena, it is required to operate the time base once only, the start of the sweep being synchronised with the switching of the circuit under test.

After the single sweep has taken place the beam remains off the screen until the time base circuit is re-set ready for a second trace.

In a hard valve time-base circuit of the Cossor type (O. S. Puckle's patent), the single sweep can be accomplished by modifying the circuit as shown in the accompanying figure.

In normal operation the condenser C_1 is charged through the pentode V_1 the discharge valve V_2 being biased negatively by the voltage drop across the resistance R_4 in the circuit of the valve V_3.

When the condenser is charged to a pre-determined value the cathode of V_2 approaches the potential of the grid and current commences to flow in the anode circuit. This produces a voltage drop across R_2 which is applied to the suppressor grid of V_3, causing it to become increasingly negative with respect to the cathode and reducing the anode current of V_3.

This in turn causes the grid of V_2 to become positive, accelerating the discharge of the condenser through V_2. When the condenser is fully discharged the anode current in V_2 ceases and the original conditions are restored until the condenser charges once more to the point at which current commences in V_2.

From the above it will be seen that the automatic discharge of the condenser is occasioned by the application of a negative potential to the suppressor grid of V_3. If this potential is prevented from being applied to the suppressor grid the condenser will remain charged and the time base will not repeat its traverse of the screen.

The suppressor grid is maintained at a constant potential by connecting it to the negative line by the switch S_1.

The condenser voltage will then remain constant until its discharge is started by applying a negative pulse to the control grid of V_3 which has the same effect as a pulse applied to the suppressor grid.

This negative pulse needs only to be of sufficient duration to enable the condenser to discharge completely after which the grid of V_3 should be allowed to return to the potential of the H.T. –ve line in order to allow the condenser to recharge.

In practice a negative potential of 16 v. can be applied to the control grid through a fixed condenser of 0.005 μF. capacity for the slower speeds of traverse of the time base. For higher speeds, 0.0002 μF. should be used.

In the commercial Cossor oscillograph, this negative potential is applied to the synchronising terminal on the control panel, the existing condenser and resistance (C_4 and R_7) in the circuit being short-circuited by a switch S_2.

The external fixed condenser should be shunted by a resistance of 5.0 megohms to enable it to discharge between successive sweeps.

The correct value of negative pulse for full traverse of the screen can be found by trial, and in the commercial oscillograph by adjustment of the " Synchronising " control knob.

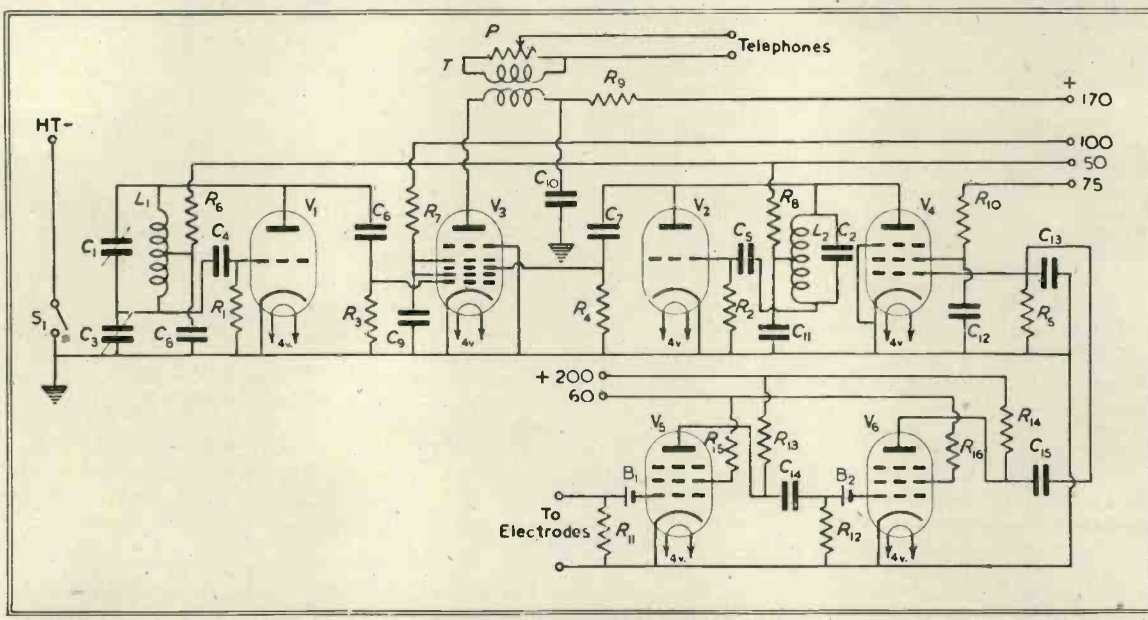
This refinement of the time base circuit is fitted to the latest models of the Cossor Oscillograph (Model 339) and the description of it is given by courtesy of Messrs. A.C. Cossor.

The Encephalophone

A New Method for Investigating Electro-Encephalographic Potentials

By C. A. BEEVERS, D.Sc., and Dr. R. FURTH*

The following is a brief description of a new electronic apparatus developed for electrobiological research, which was recently demonstrated before the Royal Society of Edinburgh.



VALUES OF COMPONENTS

C_1 Variable 0 — 200 $\mu\mu$F.	C_6 to C_{13} inclusive 5×10 $\mu\mu$F.	R_{11} and R_{12} 2 megohms
C_2 100 $\mu\mu$F.	C_{14} and C_{15} 2 μF.	R_{13} and R_{14} 15×10^4 ohms
C_3 Variable 0 — 10 $\mu\mu$F.	R_1 to R_4 inclusive 5×10^4 ohms	R_{15} and R_{16} 5×10^3 ohms
C_4 100 $\mu\mu$F.	R_5 2 megohms	
C_5 100 $\mu\mu$F.	R_6 to R_{10} inclusive 5×10^3 ohms	

Present Experimental Methods.—Two methods are in common use at the present time for the observation of the potentials of the human brain. One method uses cathode-ray oscillographs in connexion with voltage amplifiers with very high amplification, since the scalp potentials must be amplified about a million-fold before being able to produce a sufficiently large deflection of the cathode-rays in the oscillograph tube. This method is accurate, permits long watches to be made of the activity, and lends itself to photographic recording. These characteristics suggest that the cathode-ray oscillograph method is ideal for research investigations, but for clinical use, and especially for a series of clinical studies, the method is troublesome and slow.

The second method in use at the present time (developed especially in America) is that of electromagnetic oscillographs writing with ink directly on to moving paper strip. These require the use of power amplifiers as distinct from voltage amplifiers. The oscillographs themselves are made substantial, although the moving parts must be as light as possible for speed and sensitivity. The mechanical oscillograph is, therefore, necessarily a compromise which, however, is fairly satisfactory for most purposes. The great advantage of the method is that it gives immediately and cheaply a permanent record of the electrical activity during the whole period of observation. This feature has led to the adoption of this method in most EEG* laboratories. The installation, however, comprises quite a large mass of electrical machinery, and the records (generally on paper strip 3 in. wide and 1 in. long per second's observation) soon become so bulky that they are difficult to manage.

The Scope of Audio Methods

It is felt that there is scope for an apparatus mainly for clinical as distinct from purely research purposes which will convert the potential changes from the head into appropriate sounds. It would seem that such an apparatus can be made which is cheap both in its first cost and in running costs, which can be made readily portable, and in other ways possesses some convenience compared with writing methods. Such an apparatus would be suitable for surveys of large numbers of cases, appropriate ones of which could be examined by equipment giving a permanent record.

A comprehensive study of the EEG in unconscious patients would be of great interest and value, but such studies are not easily made, largely owing to practical difficulties in dealing with unconscious patients. Such patients are frequently undergoing necessary treatment which precludes their being taken off to a special EEG laboratory. A portable apparatus in these circumstances would be of considerable value. In such cases an audio method would also be more convenient for the elimination of artefacts. In the separation of extraneous potentials arising from friction, from body movements or from muscle activity or eyelid movement, it is of the greatest value to be able to watch a non-co-operative patient

* Edinburgh Royal Infirmary.

* Electroencephalograph or Electroencephalogram depending on the context.—Ed.

closely during the actual observation of potentials. In the visual methods of observation this is not easy to do, whereas in an audio method the observer's visual perceptions are left entirely free to watch the patient.

The Encephalophone. — As mention above, the idea of the new method is to make the EEG potential changes audible. It is, however, not possible to do this simply by connecting the electrodes through an amplifier to a telephone,* as the frequencies involved are in all cases far below the range of audibility. On the other hand, just because of this comparatively slow rate of potential change, a "frequency modulation" method can be used which consists in the production of an electric oscillation in the audible range which is changed in its frequency by the change of brain potential. Thus in a telephone a steady musical tone is heard as long as the potential is constant, and the pitch of the note will go up or down when the potential is increasing or diminishing.

After some preliminary experiments had been carried out on these lines with good results, an instrument was constructed which proved to be efficient for the present purpose.[1] The authors propose to call this instrument an "Encephalophone."[2]

The circuit diagram of the instrument is shown in the figure. It contains two high-frequency valve oscillators of the Hartley type, each consisting of a triode valve (V_1 and V_2), a single layer coil of ten turns with centre tapping (L_1 and L_2) and a condenser of 100 $\mu\mu$F (one of which, C_1, is variable and the other fixed). R_1 and R_2 are grid leak resistances of $5 \times 10^4\omega$ and C_4 and C_5 are coupling condensers of 100 $\mu\mu$F capacity. C_3 is a small variable condenser of 10 $\mu\mu$F max. capacity, used for the fine adjustment of the frequency of the first oscillator. The order of magnitude of the oscillation frequency is 5 Mc/sec.

The two high-frequency oscillations thus produced are electronically mixed by means of the heptode valve V_3 with the help of the two coupling condensers C_6 and C_7 of 5×10^4 $\mu\mu$F capacity and the coupling resistances R_3 and R_4 of $5 \times 10^4\omega$ each. C_8, C_9, C_{10}, C_{11} are decoupling condensers of 5×10^4 $\mu\mu$F capacity and R_6, R_7, R_8, R_9 decoupling resistances of $5 \times 10^3\omega$, meant to prevent interlocking between the two oscillators. T is an output transformer and P a potentiometer.

The anode current of V_3 contains two a.c. components with frequencies equal to the sum and to the difference of the two h.f. oscillations. The latter, the "beat frequency" can, by proper adjustment of C_1 and C_3, be easily set to a convenient value in the audible range and will consequently be heard as a tone in a telephone connected to the secondary coil of T. The intensity of this tone can be controlled with the help of P. The "summation tone" is suppressed by the impedance of the transformer coils.

The advantage of this arrangement is that evidently a very slight relative change of frequency of one of the two h.f. oscillations will result in a considerable relative change of the beat frequency and hence in an easily detectable alteration of the pitch of the telephone tone. If, for example, the two high frequencies are set to 5 Mc/s. and 500 c/s. respectively, the beat frequency is 500 c/s. If now the first oscillation is increased in frequency by 50 c/s. corresponding to a fractional change of one thousandth per cent., the beat frequency drops from 500 to 450 c/s. or about 10 per cent. and the pitch of the tone is lowered by a whole (small) tone. This also shows that the lower the beat note the higher will be the sensitivity of the instrument. Unfortunately it was not possible to lower the note as much as would have been desirable in the present arrangement, as interlocking between the two h.f. oscillators took place if the frequency difference reached a certain minimum value, in spite of the decoupling precautions; but it is hoped that this difficulty will be overcome eventually.

The frequency modulation is achieved in the usual way with the help of the valve V_4, a variable μ-pentode which, as it appears from the diagram, is connected effectively in parallel to the second oscillating circuit. A change of the control grid potential of this valve alters its impedance and hence the oscillating frequency of the circuit according to well known general principles. Thus a change of this potential is eventually converted into a change of pitch of the telephone tone as intended. R_5 the grid leak resistance of V_3 has a value of 2 megohm; the decoupling condensers C_{12} and C_{13} are of 5×10^4 $\mu\mu$F capacity and the decoupling resistance R_{10} is $5 \times 10^3\omega$.

The sensitivity of this arrangement was measured by applying known potential differences between cathode and control grid of V_4, and it was found that a voltage of 0.01 volt could just be detected by a good musical ear and a voltage of 0.1 volt could be easily recognised by anybody. But this is not nearly enough for the present purpose as the amplitude of the normal EEG effect is in the order of magnitude of 10^{-5} volts, and it therefore is desirable to be able to detect potential changes of about one microvolt.

Thus the potential changes between the electrodes must first be amplified at least in the ratio 1 : 10,000 before being applied to the frequency changer valve. The amplifier must be specially designed for the amplification of very slow oscillations since, in abnormal cases 3 c/sec. and even lower frequencies occur; the response of the amplifier should also be fairly uniform over a wide range.

In the present experimental instrument a two-stage amplifier was used, shown in the lower part of the figure. It consists of two tetrode valves V_5 and V_6 with resistance-capacity coupling. The coupling resistances (R_{13} and R_{14}) are $15 \times 10^4\omega$ each and the coupling condensers C_{14} and C_{15} are of 2 μF capacity. The control grids of the valves are biased to 1.5 volts negative by means of the two cells B_1 and B_2 in connexion with the grid leak resistances R_{11} and R_{12} of 2 Mω each. The reason for this is to reduce the average anode current of the valves as much as possible so as to be able to employ fairly high coupling resistances without being forced to increase the anode voltage over 200 volts. R_{15} and R_{16} are screen grid resistances of $5 \times 10^5\omega$.

This amplifier, although satisfying the conditions stated above, was not efficient enough, as it gave a voltage amplification of only about 500. In actual operation the instrument had therefore to be used in conjunction with a pre-amplifier of two stages for which the first two stages of one of the amplifier units of an Ediswan Electroencephalograph were used. There can, of course, be no question that a similar instrument can be built as one single compact unit.

It may be added that all the valves were indirectly heated, operated by two small accumulator cells in series as the heater battery. All the anode and grid potentials were provided by two high tension dry batteries in series, with various tapping terminals. S_1 and S_2 are the battery and heater switches. The whole set was mounted on an aluminium chassis and covered by a metal shield which also contained the dry batteries (but not the heater battery). The leads to the electrodes were also shielded by an earthed covering.

Concluded on page 442

* Some secondary effects, such as the variation of background noise due to variation of amplification-factor with EEG potential may, however, be obtained with such a straight-forward arrangement (Adrian, 1934).

1. The general idea for an audio method is due to C. A. Beevers; the principle of the present method is due to R. Fürth, who has also mainly carried out the construction of the instrument in the Department of Mathematical Physics of Edinburgh University. The necessary money was provided by the Rockefeller Foundation to whom the authors are much obliged.
A preliminary note describing the encephalophone has already been published (Fürth and Beevers, 1943)

2, following a suggestion by Dr. George Dawson.

DATA SHEETS XLV and XLVI

Performance of Resistance Capacity Coupled Amplifiers

Electronic Engineering

THE uncompensated R.C.C. Amplifier is extensively used in the lower range of the frequency spectrum for the amplification of both sinusoidal and square wave signals. The most general form of the circuit is shown in Fig. 1.

R_1 represents the coupling resistance which is unavoidably shunted by the capacity C_3 and by the capacity C_4 in series with the coupling condenser C_2. The capacity C_3 represents the anode-earth inter-electrode capacity of valve V_1 plus all stray capacities on the anode side of condenser C_2, while C_4 includes the total input capacity of valve V_2 (including Miller effects) as well as all stray capacities on the grid side of condenser C_2. The resistance R_2 represents the resistance of the grid leak of valve V_2 in parallel with the input (Miller) resistance of valve V_2.

When as is often the case $C_2 \gg C_4$ we can simplify the circuit of Fig. 1a to that shown in Fig. 1b, where $C_1 = C_3 + C_4$. The generalised solution of Fig. 1b is given by:

$$\frac{E_3}{E_1} = \frac{gR_4}{\sqrt{\left[\frac{C_1R_4}{C_2R_3} + 1\right]^2 + \left[p_4 - \frac{R_4}{p_2R_3}\right]^2}} \angle\theta \quad (1)$$

and the phase angle by:

$$\theta = \tan^{-1}\left[\frac{p_4 - \frac{R_4}{p_2R_3}}{\frac{C_1R_4}{C_2R_3} + 1}\right] \quad \ldots \quad (2)$$

where $R_3 = (1/R_1 + 1/R_a)$ and $R_4 = (1/R_1 + 1/R_a + 1/R_2)$; $p_2 = \omega C_2R_3$ and $p_4 = \omega C_1R_4$.

The time delay $-t_1$ based on phase delay is given by (θ/ω) secs.

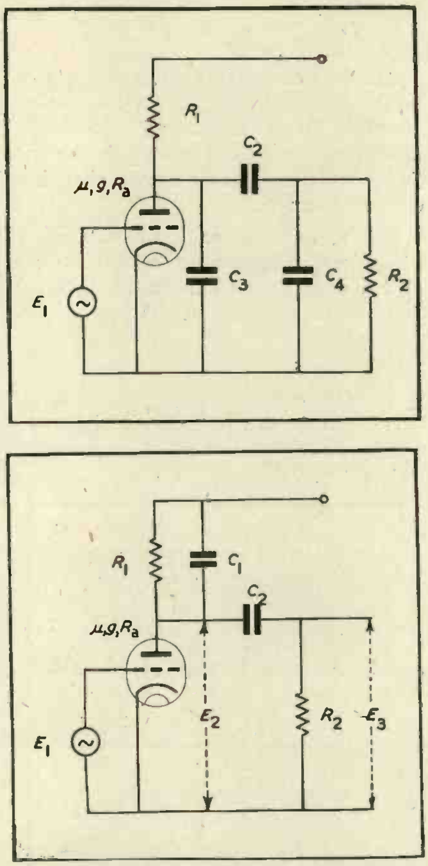
A better insight into the action of the circuit of Fig. 2 is obtained by considering separately the attenuation of the higher frequencies (given by the ratio of E_2/E_1) due to the presence of C_1. This is obtained by letting p_2 become infinite.

Similarly the attenuation of the lower frequencies (given by the ratio E_3/E_1) due to the finite value of C_2 is given by making $C_1 = 0$ or $p_3 = 0$.

Performance at Higher Frequencies

With $C_2R_3 = \infty$, $p_2 = \infty$, we have:

$$\frac{E_2}{E_1} = \frac{gR_4}{\sqrt{1 + (p_4)^2}} \angle\theta_1 \quad \ldots \quad (3)$$



Figs. 1a and 1b

and the phase angle of E_2 relative to E_1 is:

$$\theta_1 = \tan^{-1}(p_4) \quad \ldots \quad (4)$$

At the lower frequencies where the shunting action of the capacity C_1 is negligible the stage gain is

$$\frac{E_2}{E_1} = gR_4$$

so that the "Relative Gain" M or the ratio of the gain at a frequency f to the gain at very low frequencies is given by:

$$M_{\text{in db}} = 20 \log_{10} \frac{1}{\sqrt{1 + p_4^2}} \quad (5)$$

Equations (4) and (5) have been plotted on Data Sheet No. 45, where in addition is plotted a curve of time delay. In order to be able to plot a generalised time delay curve the time delay ($-t_1$) is expressed in the form

$$f_4t_1 = -\frac{\tan^{-1}p_4}{2\pi p_4} \quad \ldots \quad (6)$$

where

$$f_4 = \frac{1}{2\pi C_1R_4} \quad \ldots \quad (7)$$

On Data Sheet No. 45 the curves are shown up to a value of $p_4 = 5$. For higher values of p_4 the following simple approximations may be employed

$$M_{\text{in db}} = 20 \log_{10} \frac{1}{p_4} \quad \ldots \quad (8)$$

The bandwidth for a 1 db. attenuation can be obtained directly from the $gR_c = 0$ curve of Data Sheet 43.*

Example I

To obtain the "Relative Gain" M and the absolute gain at 15,000 c/s of an amplifier consisting of a triode having a mutual conductance g of 1 mA/V at the working point and an Anode A.C. Resistance R_a of 35,000 ohms. The anode coupling resistance has a value of 100,000 ohms with a total shunting capacity of 200 $\mu\mu$F. In addition $R_2 = 0.5$ megohm and $C_2 = 0.005$ μF.

We have therefore

$R_4 = 24{,}700$ ohms and $R_3 = 25{,}900$ ohms

and $\quad p_2 \frac{R_3}{R_4} = 247$ and $p_4 = 0.465$

therefore $\quad M = -0.85$ db.

and the amplification

$$\frac{E_2}{E_1} = 22.4$$

Performance at the Lower Frequencies

At the lower end of the frequency scale the rising reactance of the capacity C_2 will produce increased attenuation in the network C_2R_2. This attenuation is given by:

$$\frac{E_3}{E_1} = \frac{gR_4}{\sqrt{1 + \left(\frac{R_4}{p_2R_3}\right)^2}} \angle\theta_2$$

$$= \frac{gR_4}{1 + \sqrt{\left[\frac{1}{\omega C_2(R_2 + R_3)}\right]^2}} \angle\theta_2 \quad (9)$$

and the phase angle of E_3 relative to E_2 is

$$\theta_2 = \tan^{-1}\frac{R_4}{p_2R_3} \quad \ldots \quad (10)$$

and therefore the time delay $-t_1$ based phase delay can be expressed as before in the form

* See last month's Data Sheets.

$$f_5 t_1 = \frac{\tan^{-1} R_4/p_2 R_3}{2\pi p_2 (R_3/R_4)} \quad \ldots \quad (11)$$

where $$f_5 = \frac{1}{2\pi C_2(R_2 + R_3)} \quad \ldots \quad (12)$$

As the ratio of E_3/E_1 at higher frequencies again tends to the value gR_4 the expression for the "Relative Gain" M is given by

$$M_{\text{in db}} = 20 \log_{10} \frac{1}{\sqrt{1 + (R_4/p_2 R_3)^2}} \quad (13)$$

The equations (10) (11) and (13) are plotted in Data Sheet No. 46 for p_2 down to 0.2. For lower values the following simple approximation may be used.

$$M_{\text{in db}} \approx 20 \log_{10}\left(\frac{p_2 R_3}{R_4}\right) \quad \ldots \quad (14)$$

Example 2

To calculate the attenuation at 50 c/s. of the amplifier given in Example 1, we have:

$$\frac{p_2 R_3}{R_4} = 0.785 \times 1.05 = 0.825$$

and $$M = -3.95 \text{ db.}$$

To calculate the time delay we have $f_5 t_1 = 0.17$ therefore $t_1 = 2.8$ milliseconds.

At t_1 is positive it means that the output voltage of 50 c/s frequency will lead that of a higher frequency if the two are applied in phase across the anode load.

Response to Square Waves

With the aid of the amplitude response curves and phase angle or time delay curves, it is possible to calculate the shape of the output waveform from the amplifier for any given waveform of signal at the input grid. The process is, however, laborious and for the special cases where the input signal can be resolved into Heaviside's Unit-step Functions a much more convenient and simpler solution is available.

Heaviside's Unit-Step is illustrated in Fig. 2a where the applied signal is zero up to time t_0, at which time it rises instantaneously to unity value and remains there. The applied signal can always be brought back to zero by applying a second unit-step function, negative in sign, at any desired time interval after (t_0).

By the use of the Expansion Theorem we can obtain a general solution with a unit-step input for equation (1). The setting down of the equation would, however, require too much space. If, however, we consider the case frequently met in practice when $C_2R_2 \gg C_1R_4$ $R_4 \approx R_3$ than a very simple solution is available, as follows:

$$e_3 = E_1 g R_4 \left[\exp\left(-\frac{t}{C_2(R_2 + R_3)}\right) - \exp\left(-\frac{t}{C_1 R_4}\right) \right] \quad (15)$$

where e_3 is the instantaneous response at the output after a time interval t, resulting from the application of a unit-step function of magnitude E_1, and t is the time interval that has elapsed after t_0.

Just as in the case of the sine wave input we can separate the effect of the high frequency cut-off due to C_1 from the low frequency cut off in the network C_2R_2.

H.F. Attenuation

We obtain the effect of high frequency cut off alone on the response to a unit step by making $C_2R_2 \to \infty$ which gives:

$$e_2 = E_1 g R_4 \left[1 - \exp\left(-\frac{t}{C_1 R_4}\right)\right] \quad (16)$$

L.F. Attenuation

Similarly the effect of low-frequency attenuation alone can be obtained by letting $C_1 \to 0$, when

$$e_3 = E_1 g R_4 \exp\left(-\frac{t}{C_2(R_2 + R_3)}\right) \quad (17)$$

When the final amplitude of the pulse is not required to fall by more than say 10-15 per cent. of its initial value it is possible to re-write Equation (17) in a form which does not require the use of exponential tables for its solution.

If we let K denote the difference in amplitude of the pulse (expressed as a ratio) after a time "t" from its initial value at a time "t_0," we have:

$$\frac{t}{C_2(R_2 + R_3)} \approx K$$

and $$C_2(R_2 + R_3) \approx t/K.$$

The general effect of high-frequency attenuation alone is shown in Fig. 2b, low-frequency attenuation alone in Fig. 2c, and combined effects (equation 15) in Fig. 2d.

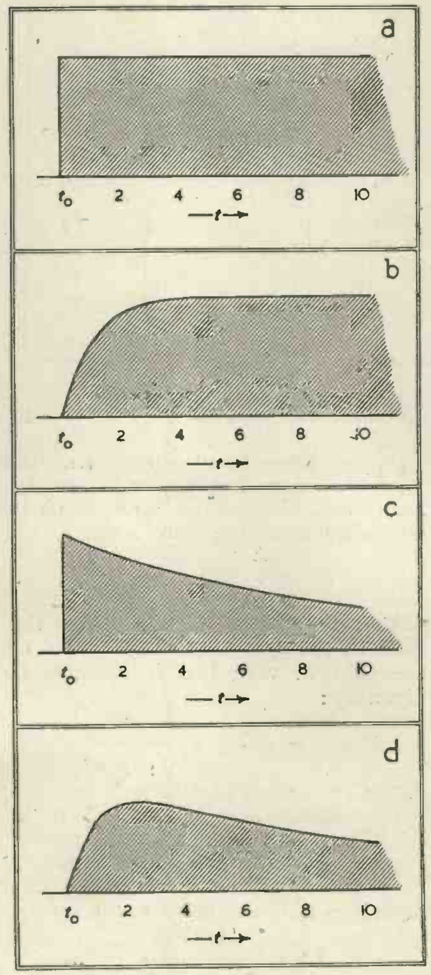


Fig. 2. Heaviside's Unit-Step Functions

Stages in Cascade

It is important to realise that in general the equivalent time constant of two similar R.–C. coupled stages in cascade is not always given by halving the CR value of one stage. Thus in the case of two cascaded stages of the type shown in Fig. 1b, with $C_1 = 0$ the output would be:

$$g^2 R_4^2 \left(1 - \frac{t}{C_2(R_2 + R_3)}\right) \exp\left[\frac{-t}{C_2(R_2 + R_3)}\right]$$

This is only equivalent to halving the time constant of a single stage when the values of K do not exceed 0.1-0.15, as

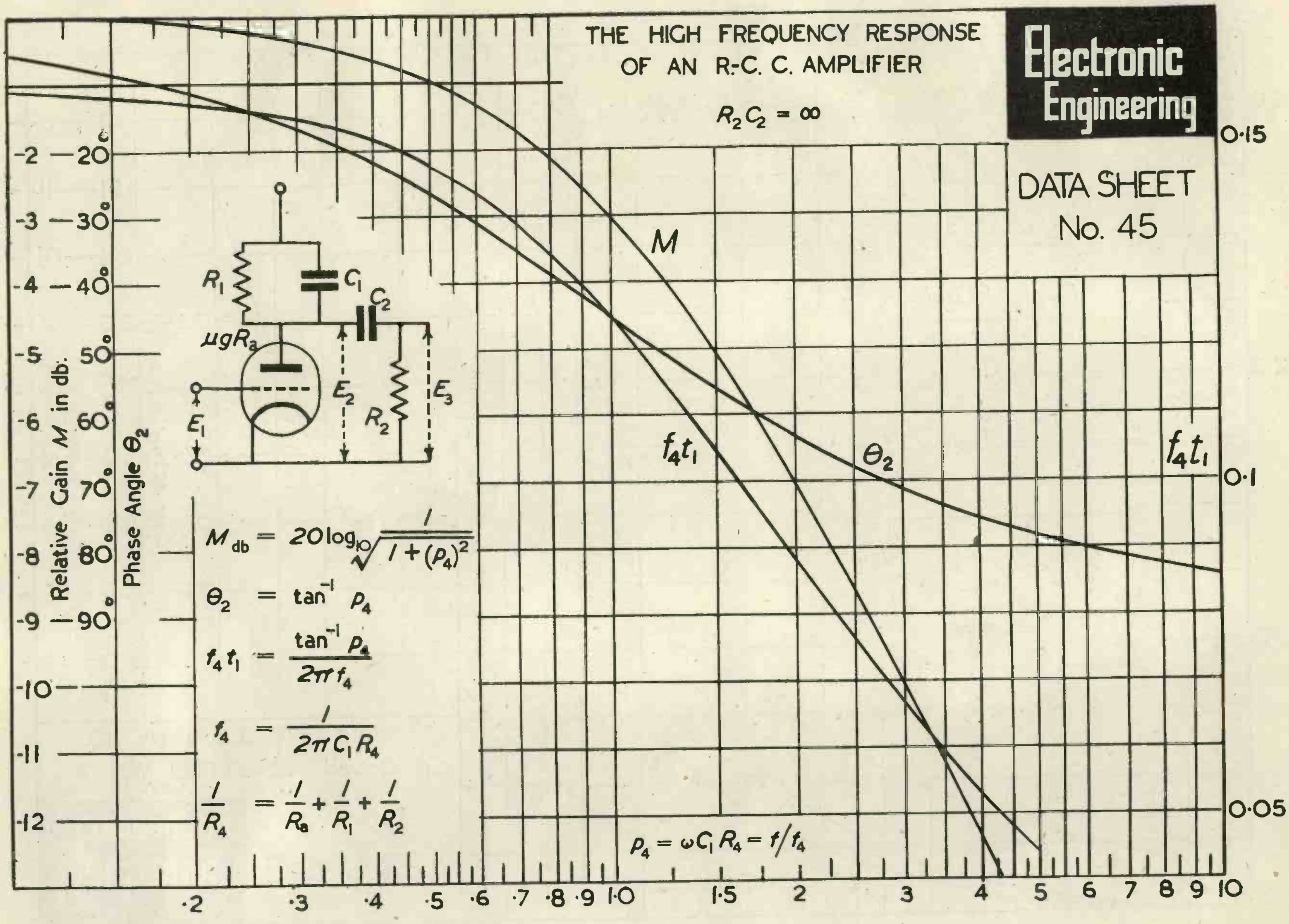
$$E.\exp(-x) \approx 1 - x + \tfrac{1}{2} x^2 -$$

THE HIGH FREQUENCY RESPONSE OF AN R-C. C. AMPLIFIER

$R_2 C_2 = \infty$

Electronic Engineering

DATA SHEET No. 45



$$M_{db} = 20\log_{10}\sqrt{\frac{1}{1+(p_4)^2}}$$

$$\theta_2 = \tan^{-1} p_4$$

$$f_4 t_1 = \frac{\tan^{-1} p_4}{2\pi f_4}$$

$$f_4 = \frac{1}{2\pi C_1 R_4}$$

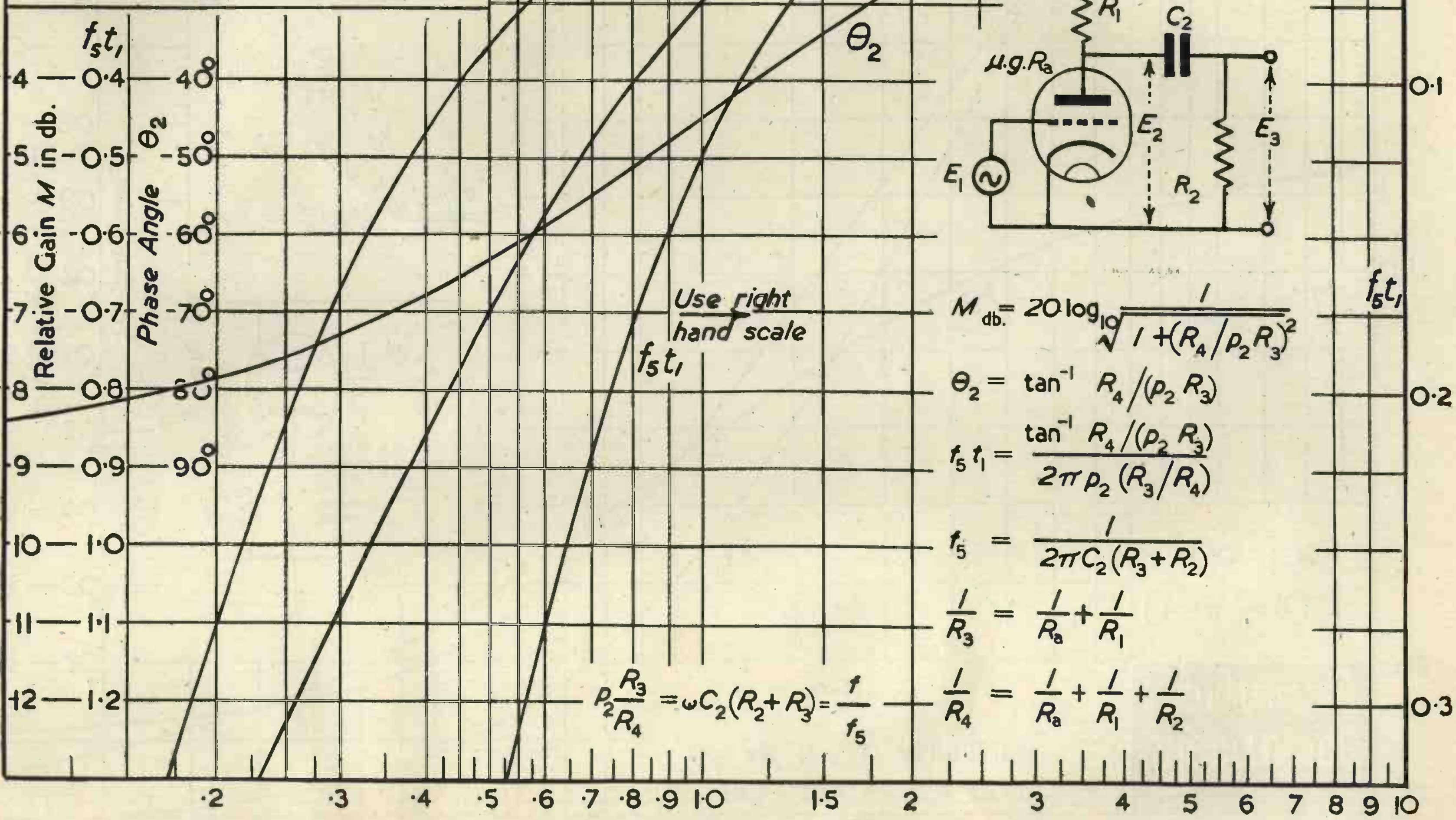
$$\frac{1}{R_4} = \frac{1}{R_a} + \frac{1}{R_1} + \frac{1}{R_2}$$

$p_4 = \omega C_1 R_4 = f/f_4$

Electronic Engineering

DATA SHEET No 46

THE LOW FREQUENCY RESPONSE OF AN R-C. C. AMPLIFIER



$$M_{db.} = 20\log_{10}\frac{1}{\sqrt{1+(R_4/p_2R_3)^2}}$$

$$\theta_2 = \tan^{-1} R_4/(p_2 R_3)$$

$$f_5 t_1 = \frac{\tan^{-1} R_4/(p_2 R_3)}{2\pi p_2 (R_3/R_4)}$$

$$f_5 = \frac{1}{2\pi C_2 (R_3 + R_2)}$$

$$\frac{1}{R_3} = \frac{1}{R_a} + \frac{1}{R_1}$$

$$\frac{1}{R_4} = \frac{1}{R_a} + \frac{1}{R_1} + \frac{1}{R_2}$$

$$p_2 \frac{R_3}{R_4} = \omega C_2 (R_2 + R_3) = \frac{f}{f_5}$$

Experimental Demonstrations for Radio Training Classes

II.—The Valve as an Amplifier

By T. J. REHFISCH, B.Sc. (Eng.)*

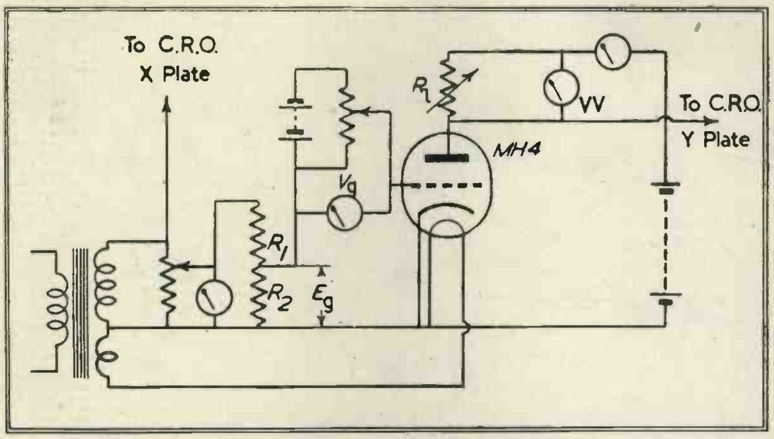


Fig. 1. Circuit for examining the voltage amplification properties of a triode, under Class A operating conditions.

THE performance of a given valve as an amplifier of low-frequency A.C. may be predicted from its D.C. characteristics. To be complete, such an analysis is laborious. An easier course is to investigate the actual performance with an A.C. input, when the results obtained may facilitate the comprehension of more advanced ideas.

The most important aspect of a thermionic valve is the fact that variations of voltage between grid and cathode (or filament) produce changes in the anode current, and hence voltage changes across a load in the anode circuit. A continuous alternating P.D. may thus be amplified into a larger P.D. across the load impedance. Further, A.C. power (= volts × amps.) is developed in the load. A valve may thus be used as a voltage or a power amplifier. Nearly always in the former, and frequently in the latter case, it is necessary that the wave-form of the output voltage should be a faithful image of the input. This condition is met by "Class A" operation. In *voltage* amplifiers the R-C coupled type preponderates, at least at low frequencies, and an ohmic resistor between H.T.+ and anode provides the load in this case. The latter, however, presents much the same resistance to D.C. as to A.C. and thus makes the effective anode potential, V_a, appreciably lower than the H.T. voltage supply. In *power* amplifiers the effective impedance of the load is always larger than the D.C. resistance between anode and H.T.+, for example, where a loudspeaker is coupled into the anode circuit through a transformer. Separate experiments to investigate voltage and power amplification are therefore required.

Voltage Amplification

It was decided to investigate the output voltage as a function of (a) input voltage, and (b) load resistance, for a triode and a pentode valve.

The circuit for the triode is shown in Fig. 1. Several practical points may be of interest. The A.C. input was derived from the 50-cycle mains via the 4v. secondary of a mains transformer of the type used in receivers. In the first part of the experiment only a *small* constant alternating input of about 0.1 v.rms was required. As the usual A.C. voltmeter (*e.g.*, an Avo-meter) is not calibrated for voltages less than 0.5 v.rms a potential divider had to be used, as indicated in Fig. 1. Here the output from the transformer is applied across resistors R_1 and R_2 in series and is measured by the Avo. Only a fraction, $(R_2/(R_1 + R_2))$, of it being applied to the valve. R_1 and R_2 are resistors of a known value, and were obtained by a ratio box. The circuit of Fig. 1 is completed by adding a bias battery, and bias potentiometer, a D.C. voltmeter, an H.T. battery, a D.C. milliammeter, and a decade resistance box R_L to act as the load. The L.T. was obtained from another low voltage secondary of the mains transformer referred to above. A high impedance valve voltmeter, of the type that is not affected by D.C. voltages, was connected across R_L. It was found necessary to insert a switch in the valve-voltmeter connexion as it was recognised that this might very

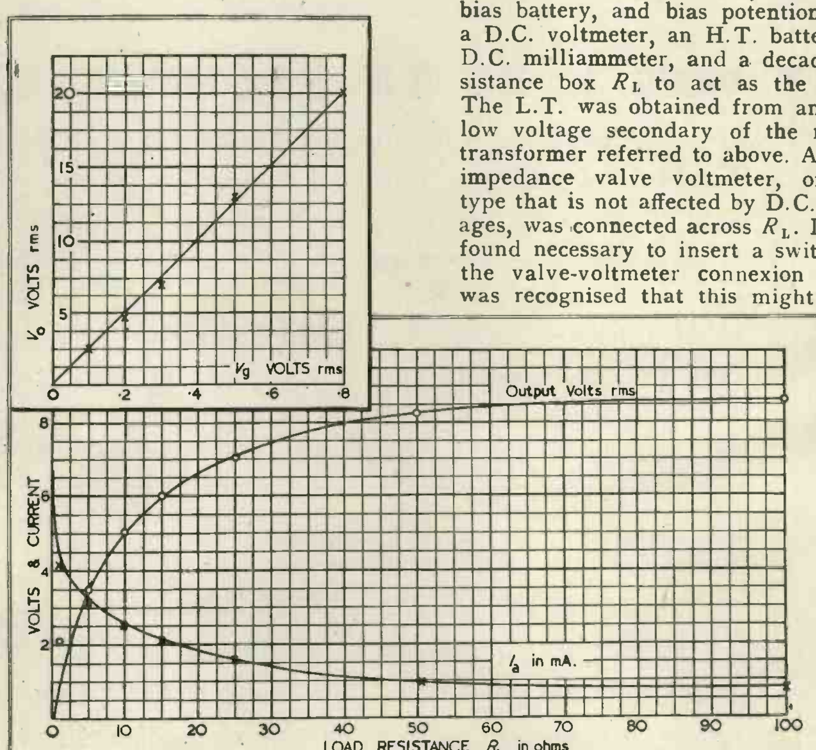


Fig. 2a. Output volts v. input volts on constant load. Circuit of Fig. 1 with valve type MH4 Anode load = 25 mΩ H.T. = 120v. Eg. = − 1v. Ia = 1.6 mA.

Fig. 2b. Output volts v. Load Resistance R_L and Anode d.c. v. R_L on constant input. (For conditions of Fig. 2a but input volts constant at .25v. r.m.s.).

* Northampton Polytechnic Institute, London.

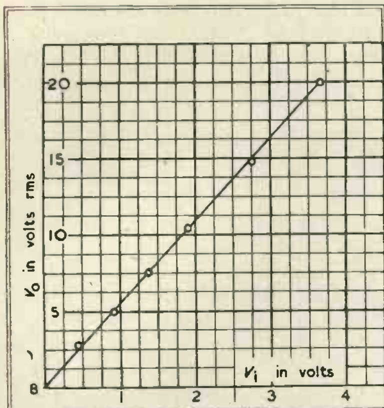


Fig. 3a. Output volts v. Input volts on constant load as for Fig. 2a but using a pentode type MSP4, with Vg. = 2v., H.T. = 150 v. Screen volts = 75 v. Suppressor grid connected to cathode.

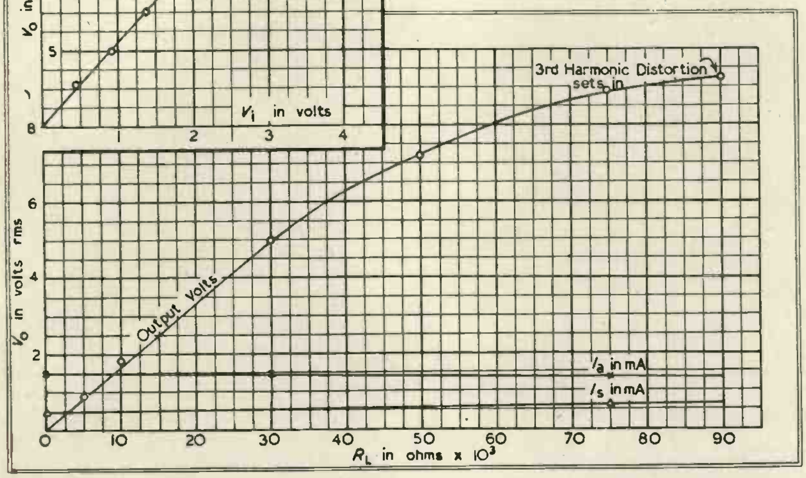


Fig. 3b. Output volts v. Load Resistance RL and anode and screen d.c. v. RL on constant input for conditions of Fig. 2b, but with constant input voltage = .075 v. r.m.s.

well distort the output voltage. A double-beam C.R.O. was available, and both the input and output voltages could be observed by connecting its A_1 and A_2 terminals to the points indicated in Fig 1. By adjustment of the amplifier controls on the C.R.O. both traces could be made equally large, and any distortion thus readily observed.

With the conditions specified in Figs. 1 and 2(a), the output voltage, V_o, was measured as the input was increased from zero. The output voltage remained undistorted up to a point well beyond the range of the valve-voltmeter. When the input voltage just exceeded 2.6v. in this experiment the negative half of the output voltage became noticeably distorted. The wave-form on the C.R.O. did not include the distorting effect which the valve voltmeter would have had throughout, as the latter was disconnected for each observation. Further, it is known from theory that the input voltage and the output voltage across a purely resistive load should have a phase difference of 180°. This is borne out by the traces on the C.R.O. screen, where they appear in phase by virtue of the construction of the deflector plate system in double-beam tubes itself producing a 180° phase displacement.

The experimental procedure outlined above was now repeated for varying load resistances the *input voltage being kept constant* at 0.25v. The results are shown in Fig. 3(a).

Both parts of the experiment were repeated for an MSP4 pentode valve, which replaced the MH4 triode. The screen was fed from a tapping on the H.T. battery, no further modification to the circuit of Fig. 1 being necessary. The results are shown in Figs. 2(b) and 3(b).

It must now be stressed that with no value of input voltage or load was it possible to obtain a perfectly undistorted output wave-form from the *pentode* circuit. The positive half was always more "compressed" than the negative half, a type of distortion which the mathematician attributes essentially to the presence of 2nd harmonics. The point at which the other half of the output wave-form also became distorted is marked in Fig. 3(b). It introduces a 3rd harmonic, and indicates an overload point.

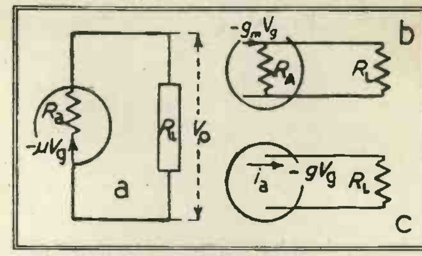


Fig. 4. A.C. equivalent circuits of Class A amplifiers. (a) and (b) are rigid equivalents, (c) a simplification which usually holds well for pentodes. (a) Is most suited for use with triodes.

Discussion of Results

An analysis of D.C. valve characteristics leads to the well-known expression

$$\text{output voltage, } V_o = \frac{\mu V_g R_L}{R_L + R_a} \qquad (1)$$

where μ and R_a are the amplification factor and internal A.C. resistance of the valve respectively. Hence the voltage amplification factor

$$|m| = \frac{V_o}{V_g} = \frac{\mu R_L}{R_L + R_a} \qquad (2)$$

= slope of Fig. 2a or 3a

Using the nominal value $R_a = 11{,}000\Omega$ for the MH4 triode and the known values of R_L and $|m|$, equation (2) gives $\mu = 36$ compared with the nominal value of $\mu = 40$ for this valve. The discrepancy may be attributed to the valve specimen or the valve-voltmeter which was used to measure V_o. At any rate, the results of Fig. 2(a) indicate that m is constant in a given circuit, over the range of input volts investigated.

Similarly, the results shown in Fig. 2(b) satisfy the requirements of equation (2). For $R_L = R_a = 11\ \text{K}\Omega$, $V_o = \frac{1}{2}\mu V_g$, and from this equation $= 36 \times 0.25/2 = 4.5$v., a value somewhat less than that actually obtained. When R_L is much greater than R_a we should expect V_o to be approximately equal to μV_g, *i.e.*, to 9v. The highest value actually observed was 8.6, which was not very far off the theoretical value.

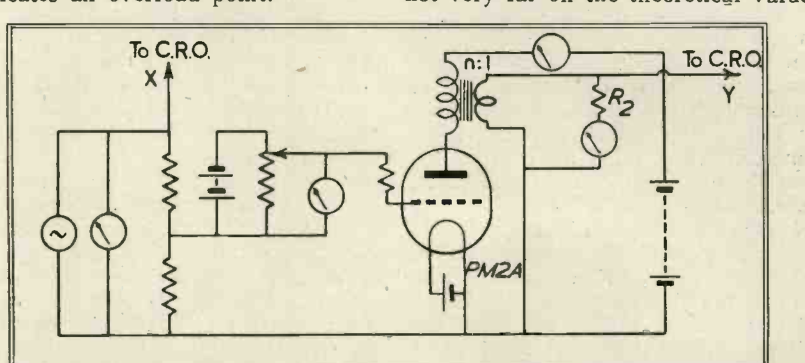
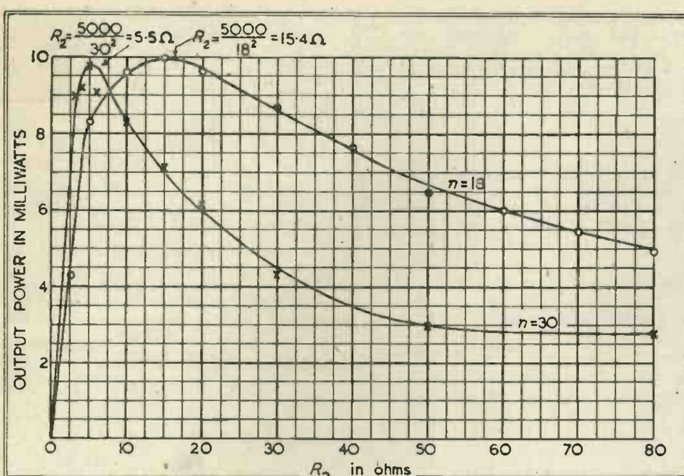


Fig. 5. Circuit for examining the properties of an amplifying circuit under Class A operating conditions.



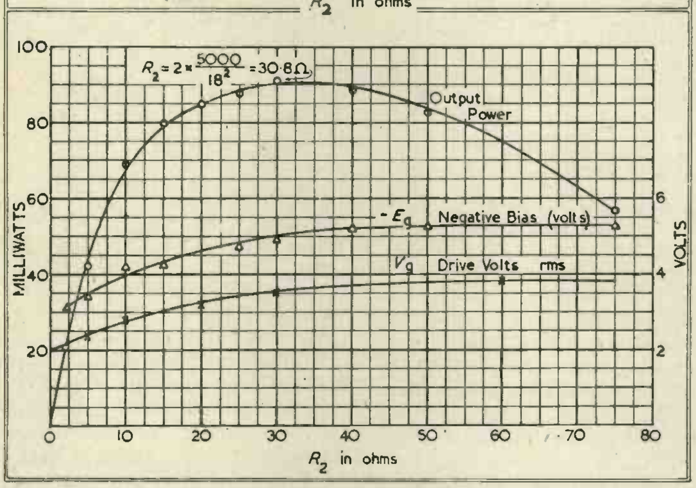


Fig. 6. Output power v. load resistance for constant grid conditions. Vg. = 1.1 volt r.m.s. Eg. = – 2v., Ea = 118, Ia = 13.2 mA.

Fig. 7. Output power, Bias Volts Eg, Drive Volts Vg v. load resistance for max. undistorted power output with each value of R_2 n = 18.

Of course, $V_o = \mu V_g$ can never be obtained in practice, because distortion will eventually set in as R_L is further increased.

The A.C. Internal resistance. The curve of Fig. 2(b) is similar to the curves of output-voltage vs. load impedance *for a source having internal resistance*. (*cf.* the curve with those in the first of these articles). Hence the A.C. properties of a Class A amplifying valve circuit may be represented by means of a valve "equivalent circuit." In the version suggested in Fig. 4 a negative sign has been attached to the internal e.m.f. so as to account for the previously noted phase difference of 180° between V_o and V_g, and a circle encloses R_a and $-\mu V_g$ to indicate their presence within the valve. No difficulties arise in regarding the valve as a simple A.C. generator if one bears in mind that the equivalent circuit refers to A.C. only. Distortion, input impedance, feedback, etc., can always be accounted for by suitable modification of the three constituents of Fig. 4. Second harmonic distortion, for example, may be introduced by inserting a generator of twice the frequency of V_g in series with the one actually shown. (See *Terman* "Radio Engineering" for details of this method).

Comparing the two valves, the outstanding feature of Fig. 2(b) is that its slope is over twice that of Fig. 2(a) although the mutual conductance of the triode (3.6 mA/v) is greater than that of the pentode (2.4 mA/v). A greater sensitivity is therefore obtainable with the pentode. This valve, however, gives greater distortion than the triode. Hence, where distortionless amplification is vital, as in audio work, pentodes may only be used if great care is taken to compensate for their inherent distortion, for example, by negative feedback and push-pull circuits.

The $V_o - R_L$ characteristic, Fig. 3(b) for the pentode is of the same general shape as for the triode. The fact that it is so obviously more linear is due to the greater value of R_a compared to R_L for the pentode. As a matter of fact, equation (2) may be rewritten as

$$V_o = \frac{-\mu V_g R_L / R_a}{1 + R_L/R_a} = \frac{-g_m R_L V_g}{1 + R_L/R_a} \approx -g_m R_L V_g$$

for the usual pentode amplifier. Thus V_o is approximately proportional to R_L. A form of A.C. equivalent circuit most suitable for pentodes is shown in Figs. 4(b) and 4(c). The current $I_A (\approx -g_m V_g)$ flows into the parallel combination of R_a and R_L, the proportion flowing into the external resistance R_L depending on the ratio R_L/R_a. Usually, the shunting effects of R_a may be neglected, and the simplified circuit of Fig. 4(c) may be used.

Power Amplification

Fig. 5 is a diagram of the circuit for investigating the properties of a Class A power amplifier. It is similar to the circuit, Fig. 1 previously used, except that the load is now of the coupled type; the small universal transformer described in the first article was used to couple the load R_2 into the anode side of the valve. A double-beam C.R.O. was again turned to good account in comparing the wave-forms of the input voltage with the voltage developed across R_L (and hence also the current in R_L). Output power was measured in the usual way (volts × amps). The alternating input, V_g, was derived from a 1,000 c/s. audio oscillator, and its magnitude could be both adjusted and measured. The only additional feature worth noting in Fig. 5 is the resistor placed in series with the grid. Its value is a few thousand ohms, and its purpose is to make noticeable the flow of grid current in the case of the applied alternating P.D. driving the grid positive. If this occurs the actual grid-cathode P.D. becomes disturbed, and hence the output voltage. This, of course, is what happens in an actual power amplifier excited by a source of appreciable internal impedance. A 2-volt battery "power" triode—actually a Mullard PM2A—was used in these experiments.

In the first part, moderate grid bias, E_g, and drive, V_g, were used and kept *constant* as the load R_L was varied. The resulting load current, I_L, was observed. The output power ($I_L^2 R_L$) was calculated, and is shown plotted against R_2 in Fig. 6. This was repeated for another transformer ratio, *n*.

In the second part of the experiment the output current and output power were again measured as R_L was varied, but this time for any given value of R_L, both V_g and E_g were ad-

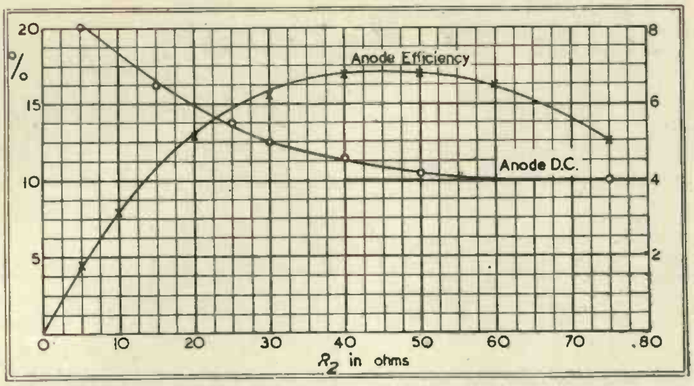


Fig. 8. Anode d.c. and Anode efficiency v. load resistance R_2 for conditions of Fig. 7.

justed for *maximum undistorted* output power. The latter condition was actually met by observing the trace of the output voltage on the C.R.O. screen, and, each time, increasing V_g and simultaneously adjusting E_g until distortion became just noticeable at both top and bottom of the output wave-form. This procedure may appear somewhat difficult, but provided a picture of the input wave-form is also obtained on the screen it becomes quite simple after a short while.

The following headings of the list of observations may help, as the number of dependent variables in the experiment are large:—

R_L ohms	V_g volts r.m.s.	$-E_g$ volts	I_L mA r.m.s.	$P_L = I_L^2 R_L$ mW	I_a mA	$I_a V_a$	η

$$\eta \left(= \frac{100\, P_L}{I_a V_a} \right) \text{ is the ``efficiency''}$$

$$\left(= \frac{\text{output power}}{\text{total power supplied}} \times 100\,\%. \right)$$

The more important results and their derivatives are plotted against R_L in Figs. 7 and 8.

Discussion of Results

The relationship between output power, P_L, and load resistance, R_L, in the case of *constant grid* conditions is analogous to the case of the simple A.C. generator with internal resistances, R_i, already referred to in the previous article. It was stated there that the output power, P_L, is a maximum when the load, R_L, is equal to R_i/n^2. With the valve, $R_i = R_a$, and the *calculated* condition for matching is shown in Fig. 6 for two different values of transformer-ratio, n. The value $R_a = 5{,}000\Omega$ was obtained from the D.C. characteristics of the valve specimen actually used. Moreover, the output power is

$$\left(\frac{\mu V_g}{R_a + R_L} \right)^2 \times R_L,$$

which, *with correct matching*, becomes a maximum of magnitude $(\mu V_g)^2/4R_a$;

here $\mu = 13$, $V_g = 1.1$v, and $R_a = 5{,}000\Omega$

$\therefore$ maximum output power,

$$(P_L)_{max} = \frac{(13 \times 1.1)^2}{4 \times 5{,}000} = 10 \text{ mW},$$

which checks up well with Fig. 6.

In the second part of this experiment, however, it was seen that more output power is obtainable if the grid be driven harder for a given load. The grid drive is limited on its positive peaks, though, by the condition $(V_g)_{peak} = E_g$, and, at the negative end, by the bottom curvature of the dynamic mutual characteristic. Inspection of Fig. 7 indicates that this danger region recedes until $R_L = 50\Omega$ as the load is increased. When $R_L = 50\Omega$ the output power decreases, and at this point the peak of the drive voltage approaches the value of the static cut-off bias.

Thus, when V_g and R_L both increase, we can state that there are two conflicting results so far as *power* is concerned. In the former case, the total A.C. power generated increases, but when R_L is increased the total A.C. power generated decreases, although a larger share of this power is then developed in R_L.

In terms of algebra, total A.C. power $= (\mu V_g)^2/R_a + R_L$. This increases when R_L is constant and V_g increases, but decreases if V_g is constant and R_L is increased. At the same time the power in R_L is a fraction of the total A.C. power given by $= R_L/R_a + R_L$ which increases as R_L increases. Thus a larger share is developed in R_L as R_L increases. When these contrasting properties are considered for an idealised valve (*i.e.*, one whose characteristics are straight lines without bottom curvature) it can be shown that $R_L = 2R_a$ is the condition for maximum output power.

The expression for the latter then becomes

$$P_L = \frac{(\mu V_g)^2 \times 2R_a}{(R_a + 2R_a)^2} = \frac{\mu^2 V_g^2}{4.5\, R_a}$$

Fig. 7 indicates that this condition obtains with $R_L = 30\Omega$.

By calculation,

$$R_L = \frac{2R_a}{n^2} = \frac{2 \times 5{,}000}{18^2} = 30.8\Omega.$$

The output power P_L, by calculation

$$= \frac{13^2 \times 3.5^2}{4.5 \times 5{,}000} = 92 \text{ mW},$$

and actually came to 91 mW on the graph. The maximum permissible grid drive was about 4v r.m.s., and could, of course, have been forecast from the D.C. characteristics of the valve.

D.C. and A.C. Efficiencies

The anode d.c. values and anode efficiency are shown in Fig. 8. The maximum efficiency reached is 17.5 per cent., which is somewhat less than the theoretical 25 per cent. for an idealised valve. It should be noted that this max. efficiency occurs at a load larger than that corresponding to max. A.C. efficiency. As R_L is increased beyond the value $2R_a/n^2$ and the grid is biased further back, the D.C. power supplied to the anode falls off more rapidly than the A.C. power output until a point is reached when no further appreciable increase of V_g drive is possible. Remembering that practically all the D.C. power not converted into A.C. power is dissipated at the anode in the form of heat, it is not surprising that much effort has been spent on devising more efficient methods of amplification such as Class B, Class C, and push-pull operation.

In conclusion, it should be noted and stressed that these considerations do not in any way indicate faults in the equivalent A.C. circuit of the valve. It is still correct for A.C. power relations in the actual valve circuit. In order, however, to obtain information on just how large V_g and how small I_a and E_a (the anode voltage) can be made the D.C. characteristics of the valve must be studied or an experiment performed. It is as absurd to maintain that this means a breakdown in the equivalent valve circuit as to say that the electrical circuit diagram of an amplifier is incorrect because the weight of its chassis cannot be deduced from it!

Thanks are due to Mr. M. Nelkon, B.Sc., A.K.C., for his help with the M.S.

Standard Values of Resistors

In an Editorial note last month it was stated that the manufacturers of moulded fixed resistors had agreed to the adoption of standard values which would cover the full range of resistance values from 10 ohms to 10 megohms if the usual tolerances were allowed. The table below gives the values of the resistances to cover the range for three tolerance figures—±20%, ±10% and ±5%.

It should be noted that the standard value ±20% is a " preferred value " and should be used wherever possible. A tolerance of ±10% is to be used only where essential, and for ±5% authorisation from the appropriate Supply Department is required.

The schedule applies only to new development projects and not to existing orders for equipment or spares.

A quick reference chart similar to the table shown is in preparation and will be supplied to firms engaged on work of national importance, price 3d. Application should be made to the manufacturers.

±20%	±10%	±5%	±20%	±10%	±5%	±20%	±10%	±5%
10	10	10	1000	1000	1000	100000	100000	100000
		11			1100			110000
	12	12		1200	1200		120000	120000
		13			1300			130000
15	15	15	1500	1500	1500	150000	150000	150000
		16			1600			160000
	18	18		1800	1800		180000	180000
		20			2000			200000
22	22	22	2200	2200	2200	220000	220000	220000
		24			2400			240000
	27	27		2700	2700		270000	270000
		30			3000			300000
33	33	33	3300	3300	3300	330000	330000	330000
		36			3600			360000
	39	39		3900	3900		390000	390000
		43			4300			430000
47	47	47	4700	4700	4700	470000	470000	470000
		51			5100			510000
	56	56		5600	5600		560000	560000
		62			6200			620000
68	68	68	6800	6800	6800	680000	680000	680000
		75			7500			750000
	82	82		8200	8200		820000	820000
		91			9100			910000
100	100	100	10000	10000	10000	1.0 Meg.	1.0 Meg.	1.0 Meg.
		110			11000			1.1 Meg.
	120	120		12000	12000		1.2 Meg.	1.2 Meg.
		130			13000			1.3 Meg.
150	150	150	15000	15000	15000	1.5 Meg.	1.5 Meg.	1.5 Meg.
		160			16000			1.6 Meg.
	180	180		18000	18000		1.8 Meg.	1.8 Meg.
		200			20000			2.0 Meg.
220	220	220	22000	22000	22000	2.2 Meg.	2.2 Meg.	2.2 Meg.
		240			24000			2.4 Meg.
	270	270		27000	27000		2.7 Meg.	2.7 Meg.
		300			30000			3.0 Meg.
330	330	330	33000	33000	33000	3.3 Meg.	3.3 Meg.	3.3 Meg.
		360			36000			3.6 Meg.
	390	390		39000	39000		3.9 Meg.	3.9 Meg.
		430			43000			4.3 Meg.
470	470	470	47000	47000	47000	4.7 Meg.	4.7 Meg.	4.7 Meg.
		510			51000			5.1 Meg.
	560	560		56000	56000		5.6 Meg.	5.6 Meg.
		620			62000			6.2 Meg.
680	680	680	68000	68000	68000	6.8 Meg.	6.8 Meg.	6.8 Meg.
		750			75000			7.5 Meg.
	820	820		82000	82000		8.2 Meg.	8.2 Meg.
		910			91000			9.1 Meg.
						10.0 Meg.	10.0 Meg.	10.0 Meg.

A Circular Aerial for U.H.F.

(From Q.S.T.—Nov. 1942)

AT a paper presented before the Summer Convention of the Institution of Radio Engineers, M. W. Sheldorf described a "circular end-loaded folded dipole" which radiates equally well in all horizontal directions and has very little vertical radiation.

Designed mainly for f.m. broadcasting, the aerial is as simple as possible in construction and can be mounted on an earthed metal pole. Individual aerials can be stacked to form a multi-unit system. While the resonance characteristic is not broad enough for television transmission it is sufficiently good for wide-band frequency modulation. The resonant frequency is adjustable after installation.

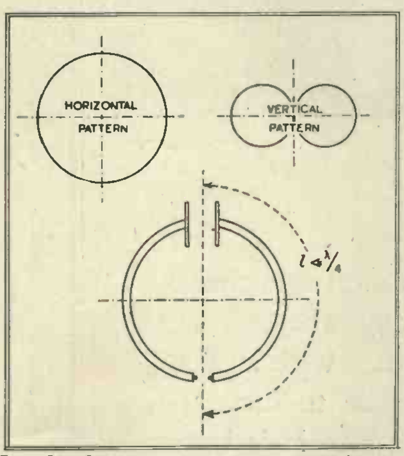
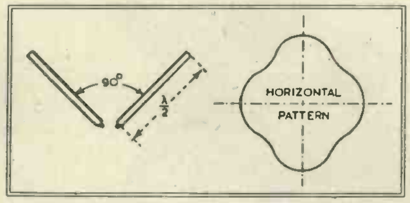


Fig. 3. Simple loop aerial. To obtain a true circular pattern in the horizontal plane the total length of loop must be small enough in comparison with the ½-wavelength so that the current is substantially the same in all parts.

The final aerial design was evolved from a cubical aerial consisting of two horizontal sets of four half-wave elements each, the elements of a set being arranged in the form of a square. Subsequent work showed that the same effect could be secured by replacing the square set of four elements by a pair of elements arranged in the form of a V having a 90-degree opening, as shown in Fig. 1. This gave the horizontal pattern also shown in Fig. 1; the shape could be controlled by altering the angle between the arms of the V, an angle smaller than 90 degrees giving an improvement over the pattern shown. However, the aerial was still bulky and the elements had to be insulated from the support.

The next step is shown in Fig. 2, where the aerial consists of two quarter-wave sections each bent in the form of a U having sides of equal length, the two sections being fitted



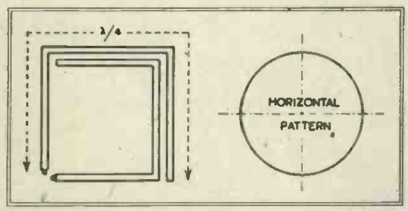


Fig. 1. 90° V-aerial and radiation pattern.
Fig. 2 (*below*). Overlapping square aerials.

together in the form of a square with two of the sides overlapping. This gives a circular radiation pattern, since the currents in the overlapping sections are in phase and the resultant "effective" current tends to be uniform around the square. This type of aerial is also obviously much smaller than the V or cubical arrangements. Because of the capacity between the adjacent sections of the aerial, the overlapping square aerial is practically the equivalent of a loop aerial having capacity loading, as shown in Fig. 3.[1]

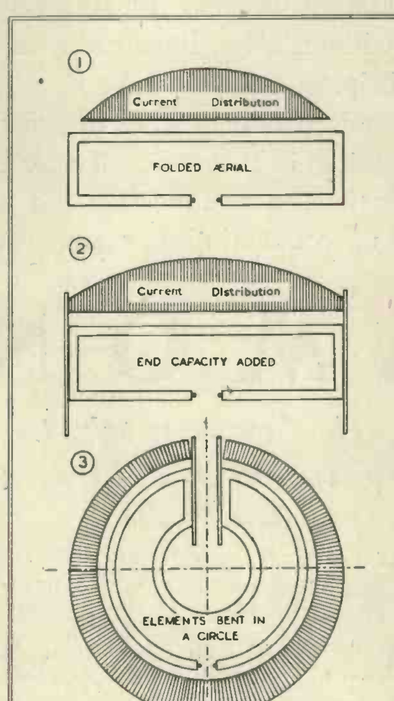


Fig. 5. The evolution of the circular aerial from a folded dipole.

The final system used is shown in Fig. 4. Because the radiation resistance of a circular aerial such as that shown in Fig. 3 is quite low, a second element was added to provide a step-up impedance transformation, using the principle of the folded dipole.[2] The effective length of the elements, including the loading of the end capacity C, is one-half wavelength overall. Point D, Fig. 4, is at earth potential and the aerial therefore can be mounted directly on a metal supporting pole at this point, without insulation. In the practical aerial the elements are made of steel pipe formed into a circle having a diameter of 33 inches, for a centre frequency of about 46 Mc/s. This compares with a length of slightly over 10 ft. for a half-wave dipole at the same frequency.

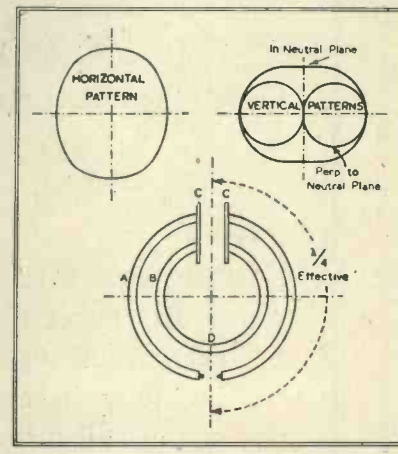


Fig. 4. The circular aerial described in the text.

Fig. 5 shows the development of the aerial from the plain folded arrangement. In the top drawing, the current distribution is close to that characteristic of an ordinary half-wave aerial. By adding end capacity, Stage 2, the current distribution is made more uniform because an appreciable current flows into the end capacitors. In the final stage the aerial system is formed into a circle with the end capacitors facing each other to form a condenser.

The relative diameters of the two elements A and B determine the magnitude of the impedance step-up. It has been found experimentally that a wide range of impedance change can be obtained. In the commercial design the terminal impedance is about 35 ohms, at resonance at 46 Mc/s. when the aerial is mounted on

A New Industrial X-Ray Unit

THE practice of using X-rays as a method of inspection in industry is rapidly being adopted, and numerous new applications are constantly being found, with the result that X-ray equipment is now being designed for specific types of work. An example of this is the new M.100 Industrial X-ray Unit developed by Messrs. Philips Lamps, Ltd., Shaftesbury Avenue, London, W.C.2.

The equipment consists of an H.T. Transformer, a shockproof and ray-proof X-ray tube and a portable control table. The shockproof X-ray tube is provided with forced air cooling and connexion to the H.T. Transformer is effected by means of two shockproof H.T. cables. The H.T. transformer is accommodated on the base of a mobile trolley and the shockproof X-ray tube is mounted on a vertical column, provided with universal movements so that the X-ray tube can be angulated in virtually any direction. For ease in transportation, the control table is mounted on an angle iron frame over the H.T. transformer, and can be removed and placed if desired, on a bench in an adjoining room. Alternatively, the frame which is also removable can be used as a stand. Connexion to the H.T. transformer from the control table is made via a multi-core cable.

This type of equipment is capable of delivering up to 100 kVp, and the controls are of a very simple nature. With this apparatus it is possible to examine up to $\frac{5}{8}$ in. steel and 4 in. of aluminium by the radiographic method. It is eminently suitable for the inspection of electrical assemblies, thin gauge spot welding and plastic materials, etc. The unit design of the equipment readily lends itself for incorporation in a conveyor belt system for the continuous visual inspection of such commodities as sparking plugs and moulded assemblies.

A Circular Aerial (*continued*)

a 4-inch diameter steel pole. With poles of larger diameter the radiation resistance decreases because of out-of-phase currents induced in the surface of the pole.

Since the aerial is appreciably smaller than an ordinary dipole, some loss of signal strength is to be expected as compared to the latter. However, it turns out that this loss is only one decibel as compared to a vertical dipole (which also has a uniform horizontal pattern). The aerials can be stacked vertically to increase the field strength, and it has been found that optimum gain is obtained when the spacing between units is about one wavelength. The gain in decibels over a vertical half-wave aerial, as a function of number of aerials or "bays," is shown in Fig. 6. It can be seen that doubling the number of elements results in approximately 3 db. gain. This is to be expected in view of the fact that the mutual impedance between aerial units or bays has been determined experimentally to be very low, when the spacing is one wavelength, hence the bays act almost independently of one another.

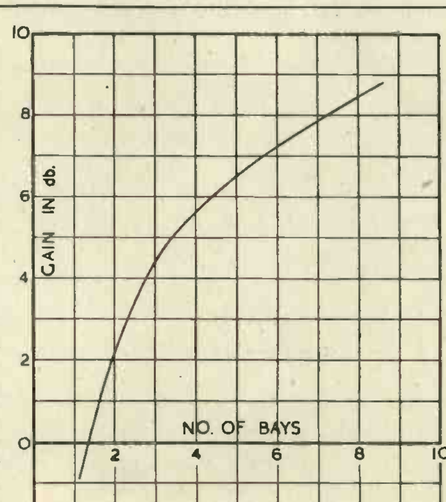


Fig. 6. Gain of circular antenna over a vertical half-wave dipole. Bay spacing is 1 wavelength.

REFERENCES.

[1] A. Alford and A. G. Kandoian, "Ultrahigh-Frequency Loop Aerials" *A.I.E.E. Trans.* Supplement, 1940.
[2] P. S. Carter, "Simple Television Aerials," *RCA Rev.* October, 1939.

Straight Line Rotating Plate Condensers with Large Angle of Rotation

IN high grade special appliances for ultra short wave engineering special rotating plate condensers are desirable for selecting from a large number of u.h.f. channels. They should fulfil the following requirements:

(a) small space
(b) high accuracy of tuning
(c) easy adjustment
(d) large angle of rotation
(e) linear relation between capacitance and angle of rotation.

How these requirements may best be fulfilled is described in a recent publication of E. Leider and O. Zinke.* In the following an abstract is given of its principal contents.

With the usual shape of stator plates covering one quadrant and having a circular face neither a rotor with regularly stepped circular quadrants nor one in which the circular faces were replaced by spirals gave a straight capacitance characteristic. An improvement is obtainable by cutting the stator face as well as the rotor face in a spiral.

Fig. 1 and 2 show the usual type of a rotating plate condenser if instead of the customary angle of 180° an angle of 270° is chosen for the rotor while the stator angle is 90°. In Fig. 1 the rotor face has three circular steps, while the stator face is also circular. In Fig. 2 the rotor faces are formed by spirals. As is shown in Fig. 3 the capacitance is by no means proportional to the angle of rotation.

Calling C = capacitance
e = the dielectric constant
F = the active area
n = the sum of stator and rotor plates diminished by 1
d = distance of plates.

we have

$$C = \frac{eFn}{d}$$

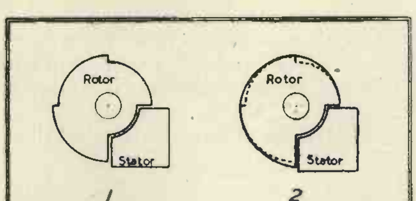


Fig. 1 (left). Condenser with stepped circular faces of rotor, stator face circular.
Fig. 2 (right). Condenser with spiral faces of rotor, stator face circular.

Stator

Angle of rotation		0°	30°	60°	90°
$K\frac{\alpha}{\alpha_{st}}$		0	0.48	0.96	1.44
$r_0^2 + K\frac{\alpha}{\alpha_{st}}$	(cm²)	1.0	1.48	1.96	2.44
r_1	(cm)	1.0	1.22	1.40	1.56

Rotor

Angle of rotation		0°	30°	60°	90°
$K\frac{\alpha}{\alpha_{st}}$		0	0.48	0.96	1.44
$R_0^2 + K\frac{\alpha}{\alpha_{st}}$	(cm²)	2.44	2.92	3.40	3.88
R	(cm)	1.56	1.71	1.84	1.97
Angle of rotation			120°	150°	180°
$K\frac{\alpha}{\alpha_{st}}$			1.92	2.40	2.88
$R_0^2 + K\frac{\alpha}{\alpha_{st}}$	(cm²)		4.36	4.84	5.32
R	(cm)		2.09	2.20	2.31
Angle of rotation			210°	240°	270°
$K\frac{\alpha}{\alpha_{st}}$			3.36	3.84	4.32
$R_0^2 + K\frac{\alpha}{\alpha_{st}}$	(cm²)		5.80	6.28	6.76
R	(cm)		2.41	2.51	2.60

As we require a linear relation between capacitance C and angle of rotation α we get

$$F = \frac{K}{2}\alpha \text{ and therefore } dF = \frac{K}{2}\,d\alpha$$

The constant $K/2$ is then given by the relation

$$\frac{K}{2} = \frac{C_{max}\ d}{en\ \alpha_{max}}$$

and has the dimension of an area.

Calling the variable rotor radius R and the variable stator radius r and referring to Fig. 4 we see that the area F for a certain angle of rotation α is given by the relation

$$F = \int_0^\alpha \frac{R^2}{2}\,d\alpha - \int_0^\alpha \frac{r^2}{2}\,d\alpha = \tfrac{1}{2}\int_0^\alpha (R^2 - r^2)\,d\alpha$$

therefore

$$\frac{dF}{d\alpha} = \tfrac{1}{2}(R^2 - r^2) = K/2$$

or for a given angle α

$$R^2 = r^2 + K$$

Similar relations are found for the second and third quadrant.

In order to get a smooth transition when passing through $\alpha = 90°$ or—more general—through α_{st}, *i.e.*, the angle of the stator plate, it is essential that r at angle α_{st} equals R_0. The

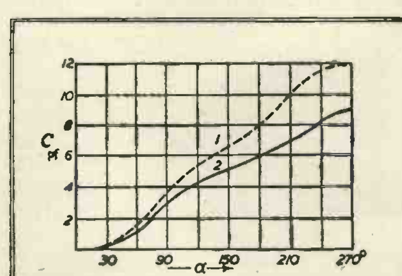


Fig. 3. Characteristics with circular stator face.
(1) Rotor face stepped circular.
(2) Rotor face stepped spiral.

* *Elektrotechnische Zeits.* 63, pp. 433-436, Sept. 24, 1942

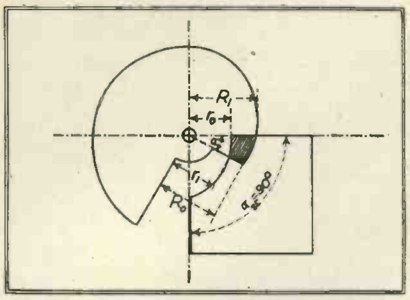


Fig. 4. Area F determined by α, R and r.

actual curve for the stator face may be chosen at random, but it is preferable to use a linear slope for the square of the radii as shown in Fig. 5 which represents the connexion between the face curves of rotor and stator for three quadrants corresponding to a total angle of rotation of 270° (with $a_{st} = 90°$). If for the stator angle 30° is chosen it is even possible to design a condenser with a total angle of rotation of 330°. In Fig. 5 the index 2 indicates that a lies between a_{st} and $2a_{st}$; the index 3 that a lies between $2a_{st}$ and $3a_{st}$.

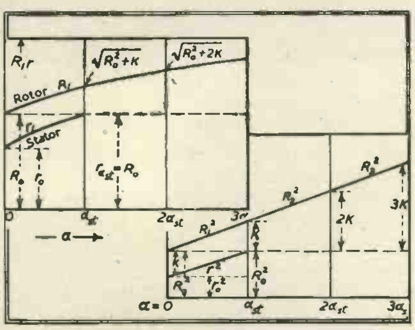


Fig. 5. (right) Connexion between the face curves of stator and rotor.

Fig. 6. (left) Actual face curves for condenser with total angle of rotation of 270°.

Fig. 6 shows the actual face curves for stator and rotor for a total angle of rotation of 270°.

With the linear slope mentioned above the radii are given by the following expressions

$$r = \sqrt{r_o^2 + K\frac{a}{a_{st}}}$$

$$R = \sqrt{R_o^2 + K\frac{a}{a_{st}}}$$

with other words the radii increase with $\sqrt{a}$.

For a condenser of $C_{max} = 10$pF having one stator and two rotor plates and $r_o = 1$ cm. the table on page 434 give the radii of stator and rotor for a maximum angle of rotation of 270°.

R. NEUMANN.

March Meetings

Institution of Electrical Engineers

Ordinary Meetings.

On March 25 at 5.30 p.m. at the Institution, Sir Frank Gill, K.C.M.G., O.B.E., will give an address on "Engineering Economics." This will be a joint meeting with the Institutions of Civil and Mechanical Engineers.

Wireless Section.

At a meeting to be held on March 3 at the Institution at 5.30 p.m., a paper will be read on "Amplifying and Recording Technique in Electrobiology" by G. Parr and W. Grey Walter, M.A. The paper will be followed by a demonstration of the electrical potentials produced by the human brain.

On March 16, also at 5.30 p.m. at the Institution, an informal meeting will be held and a discussion on "Factory Testing of Radio Equipment" will be opened by F. L. Hogg.

Institute of Physics.

The next meeting will be held on March 17 at 2.30 p.m. at the Royal Institution, Albemarle Street, London, W.1. This will be a joint meeting with the London and South-Eastern Counties section of the Institute of Chemistry. The address will be given by E. D. Eyles, B.Sc., A.Inst.P., of Messrs. Kodak, Ltd., on "High Speed Kinematography."

British Kinematograph Society.

At a meeting to be held on March 17 at the Gaumont British Theatre, Wardour Street, W1, at 6 p.m., a paper will be read by Dr. N. Fleming on "Acoustics and the Motion Picture."

Notice is hereby given that a special general meeting of active members of the Society will be held at the small theatre, Film House, Wardour Street, W.1, at 5.45 p.m., to consider whether an election of officers and committee shall, as required by the Constitution, be held this year.

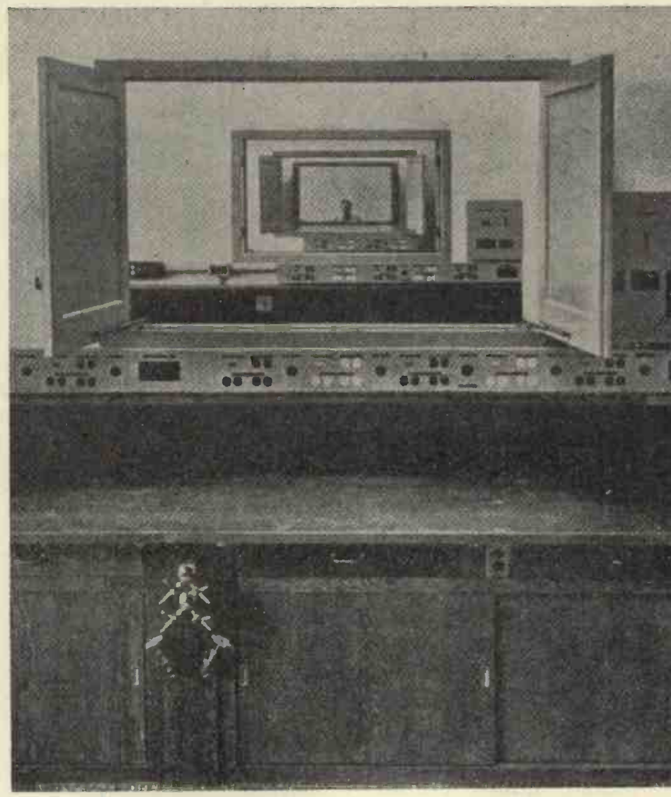
Brit. I.R.E.

The next meeting will be held on March 26, at the Institution of Structural Engineers, 11 Upper Belgrave Street, London, S.W.1, when the address will be given by L. C. Pocock, on "The Functions and Properties of Acousto-Electric Transducers."

Do realise that a poor view is taken of the "heavy-hammer-and-a-light-heart" method of working. If things stick or get tight suddenly, there's an obstruction somewhere to be cleared. As in dentistry, courtship and sardine-tin opening, persuasion is better than force.

(From a de Havilland Aircraft Co. Advt.)

NOTES FROM THE INDUSTRY



In the new R.C.A. research laboratories the optical laboratory bays have hatches in the walls to permit of long focus set-ups for television testing. The benches are fitted with numerous electrical outlets as shown and in addition are piped for air gas, water H. and O.

B.S.I.

British Standard Electrical Glossary

At the outbreak of war a revision of the B.S. Glossary of Terms used in Electrical Engineering had already commenced. The progress has necessarily been slow, but it has now reached a stage where publication in sections can begin. Under present conditions the main portion of the work will be issued in 8 parts.

Terms relating to Telecommunication, which were given in Sections 9 and 10 of the 1936 Edition of the Glossary will be issued separately in due course as a revision of B.S. 204.

Each part will be published at 2s. Part 1 is now ready, the others will follow at short intervals. Orders for other parts may now be placed.

Copies may be obtained from the Publications Department, British Standards Institution, 28 Victoria Street, London, S.W.1.

Inserted Tip Drills for Economy in High Speed Steel

The idea of using drills with inserted tips is not new, but in view of the urgent need for the utmost economy in high speed steel, it is suggested that the principle might with advantage be applied more extensively at the present time.

It is true that the drills have certain limitations. They cannot, for instance, be used for drilling from the solid, although they are perfectly satisfactory for opening out holes produced by a pilot drill, or by piercing.

One type of drill has two flutes and the other four. The fluted stem and tapered shank in each instance are manufactured from carbon steel, and the inserted tips are made from 18 per cent. tungsten high speed steel.

The 2-flute drill is provided with a tube for carrying cutting fluid to the cutting edge.

The saving in high speed steel will be appreciated from the following figures:—

The 2-flute drill is 2 inches in diameter, and the overall length is 21 inches. The total weight is 20 lb., of which only 6 oz. is high speed steel. The 4-flute drill is 25 inches long overall, and 2.70 inches in diameter. 1 lb. of high speed steel is used for the tip and 46 lb. of carbon steel for the body and shank of the tool.

Production and Engineering Bulletin, Vol. 2, *No.* 4, 1943.

Conference on X-Ray Analysis

The Institute of Physics is arranging a second conference to take place in Cambridge on April 9 and 10. The provisional programme includes a lecture on "Future Developments in X-Ray Crystallography" by Prof. J. D. Bernal and discussions on "Quantative Treatment of Powder Photographs," "The Fine Structure of X-Ray Diffraction" and "Line Broadening."

Further particulars may be obtained from the Secretary of the Institute of Physics, The University, Reading.

Catalogues Received

Resistances and Resistance Networks

Messrs. Muirhead's publication C.102A gives full information on the range of resistance boxes, slide wires and attenuators manufactured by them. Their "Munit" construction, in which the various components are assembled in metal boxes with drilled flanges on two sides, enables apparatus such as Wheatstone bridges, special attenuators, etc., to be made up in semi-permanent form from stock components.

Precision slide wires of constant inductance can be supplied in circular form with resistances from 0.5-10 Ω.

There is also an attenuator available wound to 75 ohms impedance for use with H.F. cables at frequencies up to several Mc/s. Muirhead & Co., Elmers End, Beckenham.

Potentiometers and Stud Switches

Messrs. Painton & Co. (Kingsthorpe, Northampton) can supply stud switches of instrument quality with beryllium copper brushes (if required). It has been found that the use of these brushes lowers the contact resistance appreciably and that the "noise" on rotating the switch is reduced.

The type CV.25S potentiometer is of massive construction with a substantial metal shielding case. Resistances from 100 to 50,000 ohms. Dissipation 25 watts. Vitreous enamelled and other types of fixed resistances are also manufactured.

Further Letter from Dr. J. R. Baker

SIR,—A friend has pointed out to me that one sentence in my letter in the January number of ELECTRONIC ENGINEERING implies that Prof. J. D. Bernal introduced contradictory passages into his book, *The Social Function of Science,* from unworthy motives. I wish to retract this implication as to Prof. Bernal's motives. The retraction is particularly necessary because the discussion has been closed and Prof. Bernal is thus prevented from answering in your columns.—Yours faithfully,

JOHN R. BAKER.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ABSTRACTS OF ELECTRONIC LITERATURE

TELEVISION

Analysis, Synthesis, and Evaluation of the Transient Response of Television Equipment

(A. V. Bedford and G. L. Fredendall)

The sharpness of detail in a television picture is directly dependent upon the capability of the transmitter for the transmission of abrupt changes in picture half tone. A suitable test signal is a square wave of sufficiently long period.

Rules are deduced for the evaluation of the subjective sharpness to be expected in transmitted pictures and may be applied when the square wave response of the transmitting apparatus is known. Rapid chart methods have been devised for (1) the analysis of a square wave output into sine-wave amplitude and phase response and (2) the synthesis of a square wave response from a given set of amplitude and phase characteristics. Analysis furnishes an immediate solution to the familiar but troublesome problem of finding the sine-wave characteristics of television apparatus. The four aspects of the application of square waves to television, *i.e.*, measurement, analysis, synthesis and evaluation are presented as a basis for a unified and complete technique.

The authors hope that this paper will be a contribution to the general problem of working out electrical specifications for television transmitters and other television apparatus, giving information regarding the steepness of rise and the amplitude of overswing of the square wave response.

—*Proc. I.R.E.*, Vol. 30, No. 10 (1942), page 440.

A Portable High Frequency Square Wave Oscillograph for Television

(R. D. Kell, A. V. Bedford and H.W. Kozanowski)

A portable high frequency oscillograph for television is described by which a square wave (100-kilocycle) response may be viewed as a dotted wave and readily recorded as a series of readings. The dots are spaced at 1/30—(or 1/20) microsecond intervals. No electrical connexion is required between the oscillograph and the square wave generator other than that established through the apparatus under test since the synchronous sweep and timing dots are derived from the square wave response of the apparatus. Circuit diagrams of the square wave generator and square wave oscillograph are given.

—*Ibid.*, page 458.

THERMIONIC DEVICES

A Diffraction Adapter for the Electron Microscope

(J. Hillier, R. F. Baker and V. K. Zworykin)

An adapter has been developed which allows a conventional electron microscope to be used interchangeably as an electron diffraction camera or an electron microscope. The adapter comprises a unit which takes the place of the projection lens unit of the microscope, and includes a newly designed microscope projection lens, a specimen holder and a focusing lens.

To transfer the instrument from a microscope to a diffraction camera (or vice versa) it is necessary only to transfer the specimen from the regular object chamber to the adapter. Diffraction patterns may be obtained by either reflexion or transmission. As a result of the excellent reproducibility of voltages and currents from the regulated power supplies used in the electron microscope, the diffraction camera holds its calibration to within 0.1 per cent. over long periods. Using a calibration determined by measurements of gold patterns, lattice spacings of a number of common materials, were determined and found to agree with X-ray values to within 0.5 per cent.

—*Jour. App. Phys. Vol.* 13, *No.* 9 (1942), *page* 571.

Electronic Counter for Rapid Impulses

(B. Wellman and K. Roeder)

In a circuit for the biological study of nerve potentials this thyraton circuit scales down the incoming pulses so that 600 impulses per second can be counted.

—*Electronics*, Vol. 15, No. 10 (1942), page 75.

An Electronic Circuit for Studying Hunting

(M. J. Deherno and R. T. Basnett)

A method for the determination of the power angle oscillations of a synchronous motor during hunting is described. Mirrors, arranged 360 electrical degrees apart on the motor coupling, reflect a beam of light on to a photo-electric cell which is so connected as to discharge a condenser every time it receives a light flash. The condenser cycle is observed with an oscilloscope with a specially arranged non-repeating sweep such that the envelope of the resultant oscillogram indicates the hunting characteristics.

—*El. Engg.* December, 1942, page 603.*

Operation of a Thyratron as a Rectifier

(L. A. Ware)

The half-wave thyratron rectifier circuit is treated theoretically taking into account the difference between the firing potential and the tube drop during conduction. Four loads are considered ranging from a pure resistance to a pure inductance, the impedance angles being 0, 59.15, 85.6 and 90°. The first three of these are checked oscillographically and good correspondences are obtained between (1) calculated average current and measured current, and (2) oscillographic waveshape of current and calculated waveshape. It is also noted that errors in current calculation due to erroneous values of Ef (firing potential) are higher for loads of higher impedance angle.

—*Proc. I.R.E.*, Vol. 30, No. 11 (1942), page 500.

INDUSTRY

An Instrument for Measuring Surface Roughness

(C. K. Gravley)

A tracer instrument for the measurement of surface roughness is described, and details of the main parts are given. These are a pickup using a bimorph element of piezoelectric Rochelle salt, a three-stage amplifier with a gain of approximately 100,000 and a direct inking oscillograph. The calibration and checking of the instrument are described and examples are given of its use.

—*Electronics*, November, 1942, page 70.*

Magnetostriction made Visible

(S. C. Leonard)

It is suggested that the audio-sound of iron-core apparatus is due to magnetostriction. An optical method of measuring the effect of a magnetic field on the length of a specimen in the direction of the applied field is described. Elongation or contraction of from 0.5 to 80 × 10^{-6} ins. can be observed and a photoelectric-recording type fluxmeter records magnetostriction against flux-density directly. No exact correlation between audio-sound and magnetostriction is derived. Other uses of the instrument, *e.g.*, for tensile strength and temperature coefficient measurements are mentioned

—*G. E. Rev.*, November, 1942, page 637.*

* Supplied by the courtesy of Metropolitan-Vickers Electrical Co. Ltd., Trafford Park, Manchester.

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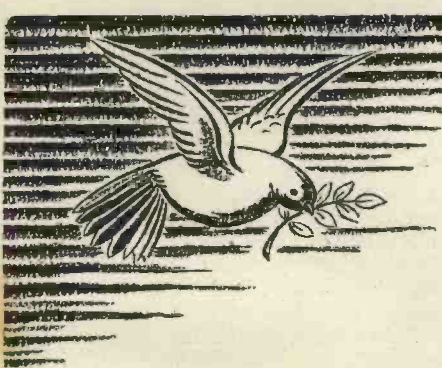

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PATENTS RECORD

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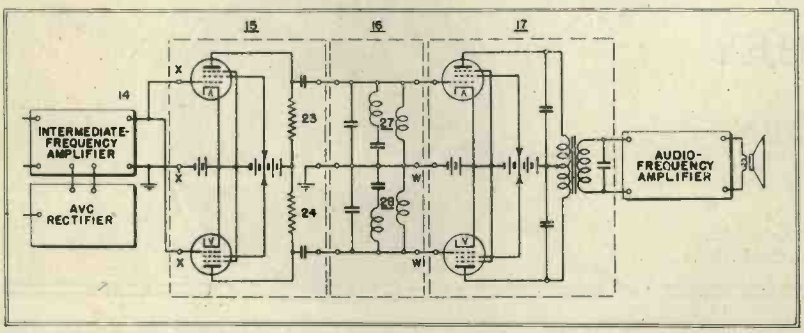
CIRCUITS

Carrier-Signal Frequency-Detector System

The ideal detector for frequency modulation is one which not only faithfully reproduces the true form of the f.m. components, but also is unresponsive to amplitude fluctuations of the carrier signal.

This invention is to provide an improved frequency detector system which, while of general application, is especially suitable for use in a frequency-modulated carrier-signal receiver.



The output of the i.f. amplifier 14 is applied to an additional stage of i.f. amplification 15 which may be biased to operate to limit to a predetermined amplitude level carrier signals translated by this amplification stage. The carrier signals developed across load impedances 23 and 24 at the nominal frequency of the carrier signal are of opposing phase, but of equal intensity and are, therefore, balanced voltages with respect to earth. These are applied to a frequency selective stage 16 consisting of a pair of impedance networks 27 and 28.

The output is applied to a rectifier stage 17 which includes a pair of pentode valves biased near to cutoff to operate as anode current rectifiers for detecting amplitude variations. Modulation components are applied to the input of the a.f. amplifier.

—Hazeltine Corp.. (Assignees of H. A. Wheeler). Patent No. 549,342.

Beam Deflecting Circuit for C.R. Tubes

Means for reducing the flyback in magnetic scanning.

In the invention means are provided for applying alternating potentials developed across the secondary winding of the output transformer and deflecting coils upon the screen grid of a pentode.

The mutual conductance of the amplifier varies in the same sense as the screen grid potential. Thus during the interval of trace the anode to cathode resistance of the pentode decreases by a relatively small amount, but during the retrace interval a relatively high negative potential peak is impressed on the screen grid through a condenser. The anode to cathode resistance of the pentode arises to a high value. This means that the current in the primary winding of the output transformer drops to a low value. This value is theoretically zero, though in reality, due to the capacity of the transformer windings, the current drops back quickly to the value it had at the beginning of the trace period.

Consequently, the rate of change of current in the transformer is accelerated appreciably and the anode to cathode resistance of the valve exerts a reduced damping effect upon the deflecting coils.

—The British Thomson-Houston Co., Ltd. Patent No. 548,463.

Attenuation Equaliser

Transmission circuits often require the use of variable attenuation equalisers which may be regulated to compensate for attenuation changes caused by changes in the temperature or other weather conditions. Under some circumstances it is desirable that the equaliser be continuously adjustable to provide a family of similar loss characteristics all of which pass through a common point. To make the equaliser more useful it is sometimes required that the family of curves be extended into the region of transmission gain.

According to the invention the network comprises a number of sections connected in parallel at their input ends and coupled at their output ends by means of a potentiometer to a high impedance load. The potentiometer may be of the condenser type and the output load the input circuit of a valve. The equaliser sections are preferably of the constant resistance type and one or more of the sections may be designed to provide a voltage gain over part of the useful frequency range. The individual sections may be so designed that their loss characteristics over the useful frequency range are substantially straight lines having different slopes, some positive and some negative, but all having the same loss at some reference frequency. Under these circumstances by an adjustment of the potentiometer there may be selected any one of an infinite family of loss characteristics all of which are linear and in effect pivot about a fixed point.

—Standard Telephones and Cables, Ltd. (Assignees of W. R. Lundry). Patent No. 549,926.

TELEVISION

Electron Camera

To correct for irregularities in the photoelectric surface of a cathode in a dissector employed as a line scanning apparatus.

This is effected by an arrangement in which a distorted line image from a continuously moving film is formed on a photosensitive cathode surface by means of a lens combination which is slightly cylindrical about a horizontal axis. The effect of the cylindrical lens is to spread the line image out into a broad band covering a large number of granules of the photosensitive surface.

A vertical slit is provided in the dissector to scan the electron stream from the cathode instead of the square aperture usually used, in order to pick up simultaneously all electrons emitted from an elemental transverse strip of the band. The effect of any irregularly sensitised spot of the photoelectric cathode is thus minimised by being combined with the effects from a large number of other spots, the image current depending on the average effect.

—Standard Telephones and Cables, Ltd. (Assignees of H. E. Ives). Patent No. 549,890.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The Encephalophone—*Continued from page 420.*

Because of the high value of the time constant of the amplifier it takes a considerable time (in the order of magnitude of one minute) before normal conditions are established after the electrodes had been handled or the instrument had been switched on. During this period the pitch of the telephone tone changes and finally becomes steady. Only then the sensitivity of the instrument reaches its normal value. With the two-stage amplifier alone the waving of a charged insulator (fountain pen) at a distance of about 10 cm from the unearthed electrode transforms the steady tone into a trill, and with the pre-amplifier in operation (normal sensitivity) the same effect is obtained at a distance of about 60 cm.

Disturbances from 50 c/s. alternating currents are heard as a kind of roughness of the telephone tone, which makes the observation of small alterations in pitch difficult. Such disturbances must therefore be avoided as far as possible by working at some distance from leads carrying a.c. and by having such leads shielded by earthed metal tubes. Otherwise the instrument is remarkably steady in its operation and corrections in the setting of the different controls are hardly necessary once they have been properly done.

The experimental model of the Encephalophone described above has been tried out in Edinburgh Royal Infirmary for a period of a month. It gives a readily perceivable and often very striking indication of the brain potentials.

Some of the finer details of wave forms are, of course, beyond the analytical capacity of the listener's auditory mechanism, such details are, however, so far of little diagnostic value. It seems therefore likely that the apparatus will provide an adequate method for clinical purposes. The most important feature of the EEG is the frequency of the waves, and this is directly heard on the encephalophone; it can with experience be estimated with accuracy sufficient for most practical purposes.

Finally, the following future development of the instrument may be suggested. In its present form the audio method does not permit a very accurate measurement to be made of the size of the activity, but this can be improved by fitting the instrument with an artificial source of very small potential changes which can be made to produce the same changes in pitch as the EEG, and measured by means of a potentiometer. Also it is often important to observe the EEG potentials between more than one pair of electrodes simultaneously, in order to study phase relations between the activity in different parts of the brain. In the standard methods of examination this is accomplished by using a number (usually three) of "channels" simultaneously. In the audio method the simultaneous observation of two regions could be carried out by using two independent channels and leading the outputs of these to the two ears of the observer. If the activities of the two regions were entirely different the observer would probably find it difficult to analyse the two sounds together. However, in normal subjects the two hemispheres of the brain are always very similar, showing waves of the same frequency, size, wave-form, and phase. A departure from identity in any of these features provides evidence suggestive of dysfunction, so that if the two audio channels were used symmetrically on both hemispheres an experienced observer would readily detect any asymmetry of electrical activity.

ACKNOWLEDGMENTS.—We thank the University of Edinburgh and Mr. Norman M. Dott for the opportunity of doing this work, and the latter for advice and encouragement.

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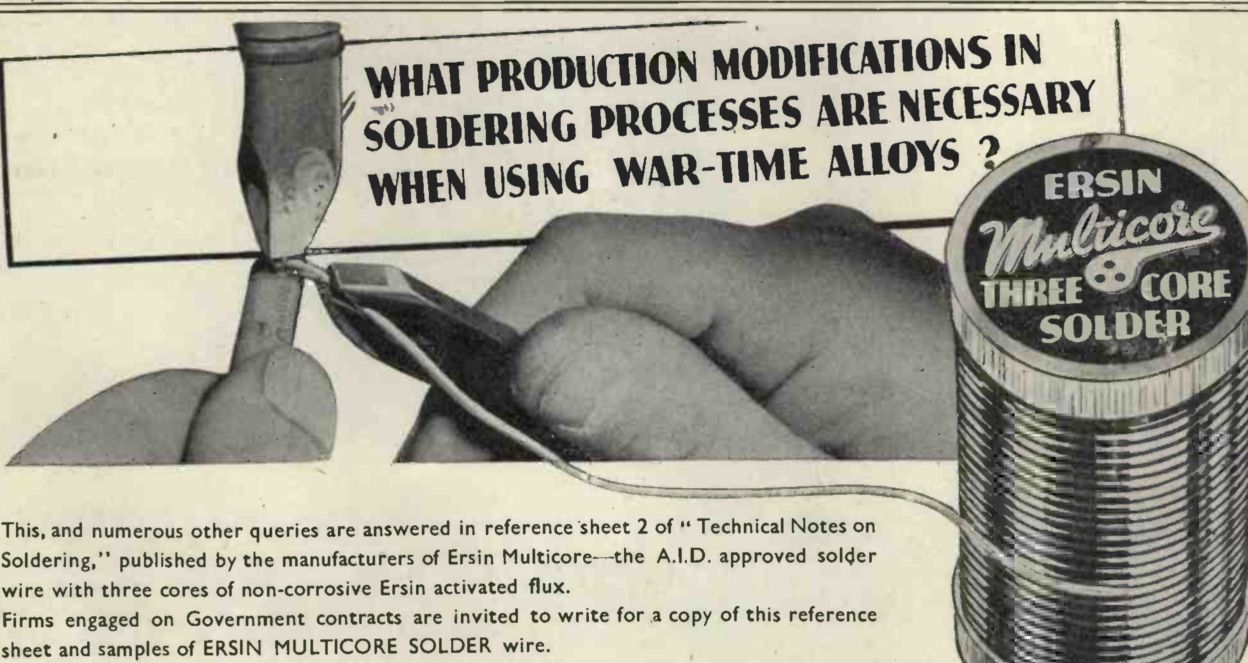

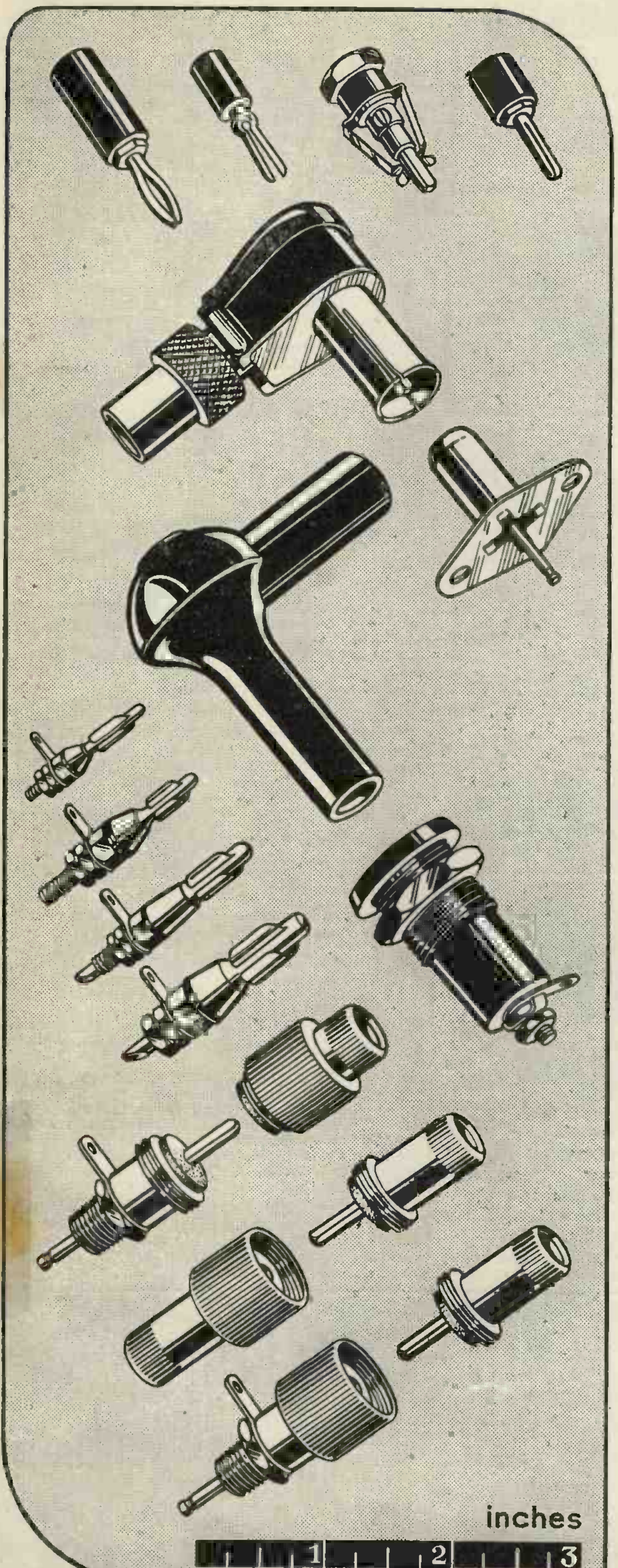
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Classified Announcements

From trimmers

to transmitters

Practically every component the short-wave worker is likely to want is in stock at Webb's, and where equipment of a specialised nature is required for official work Webb's have facilities for constructing it. In fact, if it is anything to do with short wave radio . . . see Webb's about it.

Typical Webb's Services

SPECIAL PRODUCTS DEPT. This Department offers very useful facilities for the construction of specialised equipment for transmission and reception. One of its recent productions—a 75 watt 'phone and C.W. Transmitter is illustrated above.

SHORT WAVE COMPONENTS. We are still able to offer practically any component required by the Short-wave enthusiast at our Retail Sales Dept., 14, Soho Street. If you are unable to call, our Mail Order Dept. provides a very efficient and careful service.

SERVICE OF COMMUNICATION RECEIVERS. A restricted service can now be given on all types of Communication Receivers but please do not despatch instruments without first communicating with us.

WEBB'S

Wartime Service

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