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WIRELESS ENGINEER

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WHY ERSIN MULTICORE

the Solder wire with 3 cores of non-corrosive ERSIN FLUX is preferred by the majority of firms manufacturing the best radio and electrical equipment under Government Contracts.

WHY THEY USE CORED SOLDER

Cored solder is in the form of a wire or tube containing one or more cores of flux. Its principal advantages over stick solder and a separate flux are:

(a) it obviates need for separate fluxing (b) if the correct proportion of flux is contained in cored solder wire the correct amount is automatically applied to the joint when the solder wire is melted. This is important in wartime when unskilled labour is employed.

WHY THEY PREFER MULTICORE SOLDER. 3 Cores—Easier Melting

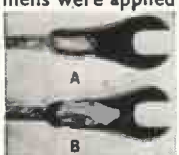
Multicore Solder wire contains 3 cores of flux to ensure flux continuity. In Multicore there is always sufficient proportion of flux to solder. If only two cores were filled with flux, satisfactory joints are obtained. In practice, the care with which Multicore Solder is made means that there are always 3 cores of flux evenly distributed over the cross section of the solder, so making thinner solder walls than single cored solder, thus giving more rapid melting and speeding up soldering.

ERSIN FLUX

For soldering radio and electrical equipment non-corrosive flux should be employed. For this reason either pure resin is specified by Government Departments as the flux to be used, or the flux residue must be pure resin. Resin is a comparatively non-active flux and gives poor results on oxidised, dirty or "difficult" surfaces such as nickel. The flux in the cores of Multicore is "Ersin"—a pure, high-grade resin subjected to chemical process to increase its fluxing action without impairing its non-corrosive and protective properties. The activating agent added by this process is dissipated during the soldering operation and the flux residue is pure resin. Ersin Multicore Solder is approved by A.I.D., G.P.O., and other Ministries where resin cored solder is specified.

PRACTICAL SOLDERING TEST OF FLUXES

The illustration shows the result of a practical test made using nickel-plated spade tags and bare copper braid. The parts were heated in air to 250° C, and to identical specimens were applied ½" lengths of 14 S.W.G. 40/60 solder. To sample A, single cored solder with resin flux was applied. The solder fused only at point of contact without spreading. A dry joint resulted, having poor mechanical strength and high electrical resistance. To sample B, Ersin Multicore Solder was applied, and the solder spread evenly over both nickel and copper surfaces, giving a sound mechanical and electrical joint.



ECONOMY OF USING ERSIN MULTICORE SOLDER

The initial cost of Ersin Multicore Solder per lb. or per cwt. when compared with stick solder is greater. Ordinary solder involves only melting and casting, whereas high chemical skill is required for the manufacture of the Ersin flux and engineering skill for the Multicore Solder incorporating the 3 cores of Ersin Flux. However, for the majority of soldering processes in electrical and radio equipment Multicore Solder will show a considerable saving in cost, both in material and labour time, as compared either with stick solder or single cored solder. Cored solder ensures that the solder and flux are put just where they are required, and by choice of suitable gauge, economy in use of material is obtained. The quick wetting of the Ersin flux as compared with resin flux in single core resin solder ensures that with the correct temperature and reasonably clean surface, immediate alloying will be obtained, and no portions of solder will drop off the job and be wasted. Even an unskilled worker, provided with irons of correct temperature, is able to use every inch of Multicore Solder without waste.

ALLOYS

Soft solders are made in various alloys of tin and lead, the tin content usually being specified first, i.e. 40/60 alloy means an alloy containining 40% tin and 60% lead. The need for conserving tin has led the Government to restrict the proportion of tin in solders of all kinds. Thus, the highest tin content permitted for Government contracts without a special licence is 45/55 alloy. The radio and electrical industry previously used large quantities of 60/40 alloy, and lowering of tin content has meant that the melting point of the solder has risen. The chart below gives approximate melting points and recommended bit temperatures.

ALLOY Tin Lead	Equivalent B.S. Grade	Solidus C.°	Liquidus C.°	Recommended bit Temperature C.°
45/55	M	183°	227°	267°
40/60	C	183°	238°	278°
30/70	D	183°	257°	297°
18.5/81.5	N	187°	277°	317°

VIRGIN METALS — ANTIMONY FREE

The wider use of zinc plated components in radio and electrical equipment has made it advantageous to use solder which is antimony free, and thus Multicore Solder is now made from virgin metals to B.S. Specification 219/1942 but without the antimony content.

IMPORTANCE OF CORRECT GAUGE

Ersin Multicore Solder Wire is made in gauges from 10 S.W.G. (.128"—3.251 m/ms) to 22 S.W.G. (.028"—.711 m/ms). The choice of a suitable gauge for the majority of the soldering undertaken by a manufacturer results in considerable saving. Many firms previously using 14 S.W.G. have found they can save approximately 33⅓%, or even more by using 16 S.W.G. The table gives the approximate lengths per lb. in feet of Ersin Multicore Solder in a representative alloy, 40/60.

S.W.G.	10	13	14	16	18	22
Feet per lb.	23	44.5	58.9	92.1	163.5	481

CORRECT SOLDERING TECHNIQUE

Ersin Multicore Solder Wire should be applied simultaneously with the iron, to the component. By this means maximum efficiency will be obtained from the Ersin flux contained in the 3 cores of the Ersin Multicore Solder Wire. It should only be applied direct to the iron to tin it. The iron should not be used as a means of carrying the solder to the joints. When possible, the solder wire should be applied to the component and the bit placed on top, the solder should not be "pushed in" to the side of the bit.

ERSIN MULTICORE SOLDER WIRE is now restricted to firms on Government Contracts and other essential Home Civil requirements. Firms not yet using Multicore Solder are invited to write for fuller technical information and samples.

MULTICORE SOLDERS LTD., BUSH HOUSE, W.C.2. 'Phone Temple Bar 5583/4

WIRELESS ENGINEER

Editor HUGH S. POCOCK, M.I.E.E. Technical Editor Prof. G. W. O. HOWE, D.Sc., M.I.E.E.

VOL. XX APRIL, 1943 No. 235

Editorial

The Calculation of Aerial Capacitance

ABSTRACT No. 766 on page 141 of the March *Wireless Engineer* deals with "Investigations on Umbrella Aerials" by Rösseler and Vogt published in the German *T.F.T.* of June, 1942. It is stated that "the capacitances were calculated by Howe's 1914/1915 method; the calculated results were on the average 10 per cent. below the measured values—a satisfactory agreement in view of the approximate nature of the calculation." We are in complete agreement with this statement, but wish to make it clear that this observed difference between the calculated and measured values is not entirely accidental. The method is one which necessarily gives a result somewhat less than the measured capacitance, the amount of the discrepancy depending on the arrangement of the wires. It is possible to obtain a closer approximation by going a step further in the assumed distribution of the charge, but at the expense of considerably more calculation. This was explained at the time, but as it is now nearly thirty years ago, it is probably for the majority of our readers wrapped in the mists of radio antiquity. The basis of the method is the assumption that the charge is uniformly distributed over the system of conductors and that the average potential over the system under this condition is equal to the actual uniform potential attained when the same total charge is allowed to take up its natural distribution. This assumption enormously simplifies the calculation of the capacitance of a complicated system of wires such as an umbrella aerial; in fact, we might go further and say that it makes it possible. Before the introduction of this method, the capacitance of such aerials was predetermined by measurements on small scale models. By immersing the model in a conducting liquid and measuring the conductance of the path through the liquid from the aerial to the plate representing the earth, the capacitance could be determined. The method of calculation introduced in 1914 did away with the necessity for such model experiments, and although it does not give the capacitance with 100 per cent. accuracy, one always knows that the actual capacitance will be somewhat greater than that calculated. This follows at once from the well-known fact that the charge naturally distributes itself in such a way that the potential is the same at all points of the conductor and that this distribution gives a minimum of potential energy and therefore of potential difference between the aerial and the earth.

Capacitance of a Straight Wire

As a simple illustration we may consider the capacitance of a straight wire with a length 1000 times its diameter, far removed from any other conducting bodies or dielectrics. If it be assumed to carry a uniformly distributed charge of 1 unit per cm. the average potential V is equal to $2 \log_e \frac{l}{r} - 0.614$

which is 14.59. For the capacitance we therefore have $C = \frac{l}{14.59} = 0.0685\, l$ cm.

As a closer approximation we may assume the wire to be divided into four equal parts, the two inner sections being charged with 1 unit per cm., and the two outer sections with q units per cm. q must be such that all sections have the same average potential. By calculating the mean potential of each section due to its own charge and to the charges on the other sections, and equating the values thus calculated, it is found that $q = 1.10$ units per cm. on the outer sections. The total charge is therefore $1.05\, l$ and as the calculated mean potential is 15.295, the capacitance $C = 0.0688\, l$ cm. This is only 0.5 per cent. greater than the value obtained by the simple assumption of a uniformly distributed charge. By dividing the wire into six equal sections and giving charges of 1, q_1, and q_2 per cm. to the inner, intermediate and outer sections, it is found that for all sections to have the same mean potential, $q_1 = 1.0176$ and $q_2 = 1.141$. This brings out very clearly the increased density near the ends, but on calculating the resulting mean potential and the total charge, the capacitance is still found to be $0.0688\, l$, *i.e.* the same, to a high degree of accuracy, as calculated above for division into four sections. It is thus seen that for cases in which the non-uniformity of distribution is of the same order as in a straight wire, the simplest assumption gives a value differing by less than 1 per cent. from the actual value.

Umbrella Aerials

The German paper from which we quoted above was concerned, however, with umbrella aerials, in which the distribution may be less uniform. In the 1915 paper* to which the authors refer, two umbrella aerials, one simple and the other more elaborate, were calculated. The simple one consisted of a vertical wire 26 metres long with 6 ribs each 30 metres long, all the wire being of the same diameter, viz. 3 mm. With uniform distribution the calculated capacitance was 802 cm., but if the vertical wire be assumed to have a slightly smaller charge per unit length than the ribs, so as to make them both have the same mean potential, the calculated capacitance is only increased to 807 cm., an increase of less than 1 per cent. A more reasonable assumption would be to give the vertical wire and the inner halves of the ribs the same charge, but the outer halves a greater charge per unit length. This would add considerably to the work involved, but would undoubtedly give a slightly greater capacitance, and therefore a closer approximation to the true value.

The other aerial that was calculated had a total height of 70 metres, and five ribs each consisting of four wires situated at the corners of a square of 2 metres side, the length of the ribs being 60 metres. To make the calculation more general it was assumed that the ribs consisted of 10 mm. wire, while the vertical wire had a diameter of 20 mm. It was found that to give the same mean potential on the vertical wire as on the ribs, it was necessary to assume twice the charge per unit length on the former, but as its diameter was twice that of the ribs, this was equivalent to the assumption of a uniform surface distribution of charge. This will apply approximately to all cases in which wires of different sizes are involved ; it is the surface density and not the charge per unit length that should be assumed uniform.

In most cases the simplest application of the method gives a result within 1 or 2 per cent. of the exact value, which is always somewhat greater than the calculated value. Measured values will almost always be even still greater because of the difficulty of excluding extraneous objects which add somewhat to the measured capacitance.

G. W. O. H.

* The Capacity of Aerials of the Umbrella Type. *Electrician*, LXXV, p. 870, also *Wireless World*, Oct. 1915.

Superheterodyne Tuning*

Ganged Control of Signal and Oscillator Tuning

By D. Riach, B.Sc., A.M.I.E.E.

THE most commonly employed method of tuning both the signal and oscillator circuits of a superheterodyne receiver by means of a single control takes the form shown diagrammatically in Fig. 1 (inset). The capacitances of both the signal- and oscillator-frequency tuning condensers, C, which are matched and ganged together, are the same value at any position of the tuning dial. Throughout the tuning range the object is to maintain the tuned frequency of the signal circuit lower than that of the oscillator circuit by a constant difference which is equal to the intermediate frequency. This tracking, as it is called, of the frequencies of these two tuned circuits is accompanied by a certain amount of out-of-tune, the extent of which, however, is not so serious as to detract from the practicability of the scheme.

The algebraic equations for determining the values of the circuit constants to give most accurate tracking and for assessing the tuning errors are somewhat tedious to apply. On this account the writer has worked out the following geometrical method which enables the said information to be obtained more easily and has the further advantage that the effect of adjustment to the circuit constants is immediately shown. The scale on the left of Fig. 1 formed by lines radiating from O on the right is for the measurement of tuning capacitance in terms of frequency, the tuning capacitance being represented by the length of any selected line drawn parallel with OB from the ∞ (infinity) line. For instance, a capacitance range of 540/60 $\mu\mu$F represented on the selected line, would be found by the scale to tune over a frequency range of 1/3; viz. $\sqrt{60}/\sqrt{540}$. The difference of the two lengths representing this capacitance range will represent $C_{max.} - C_{min.}$ and if the capacitance consists of C and C_1 in parallel, then knowing the ratio $C_{max.}/C_{min.}$, the separate values of C_1, $C_{max.}$ and $C_{min.}$ may be set out on the selected line, fixing the position of the line OA as shown.

The actual construction of the scale is purely a matter of applying the relation that the tuning capacitance varies as the inverse of the square of the frequency, viz. $C \propto 1/f^2$. A line (not shown) is drawn parallel to OB, near the left-hand side of the sheet, to represent in length the capacitance for the frequency indicated by line 1. Division points are marked on this line at distances from the lower or ∞ end proportionate to the capacitances to tune to the respective frequencies indicated by the numbers 1.2, 1.4, etc., and intermediate values. The distances to the numbered lines will be in the proportion 6.944, 5.102, 3.906, 3.086, 2.5, 2.066, 1.736, 1.479, 1.276 and 1.111 if the distance from ∞ to 1 is made 10. The relative position of O should be chosen approximately as shown and lines drawn from it to the numbered points. The lines 1 to 3 from O are shaded on each side to a width corresponding to a frequency change of 2 per cent.

The next stage in the development of Fig. 1 is to divide a line so that the lengths from one end to points on it corresponding to the above 1, 1.2, etc., signal frequencies, are proportionate to the capacitances required to tune to the respective oscillator frequencies. The formula for this purpose is $C \propto 1/(f + f_1)^2$ where f is the signal frequency and f_1 the intermediate frequency, which latter is constant. A set of such lines is shown in the lower figure covering a wide range of possible values of the ratio of the minimum to the maximum oscillator frequencies. With a signal range of 500/1500 kc/s (600/200 metres) and an intermediate frequency of 465 kc/s, this ratio is 965 : 1965, which is 1 : 2.036 and requires a tuning capacitance ratio, $LN : LM$, of 4.145 : 1. The particular line is marked dotted and is

* MS. accepted by the Editor, November, 1942.

the one selected for the upper part of Fig. 1, where it is shown thick and pointing obliquely upwards to the right. Ten equal increases of the oscillator frequency are defined by the distances from L of consecutive points which will be found in the proportion 4.145, 3.404, 2.845, 2.413, 2.072, 1.799, 1.577, 1.393, 1.239, 1.11 and one. The set of lines is proportioned so that the part MN of the whole length, LN, is equal. If the lines were made all the same length the lower ones of the set, when applied over the lines emanating from O, in the manner explained later, would take up positions

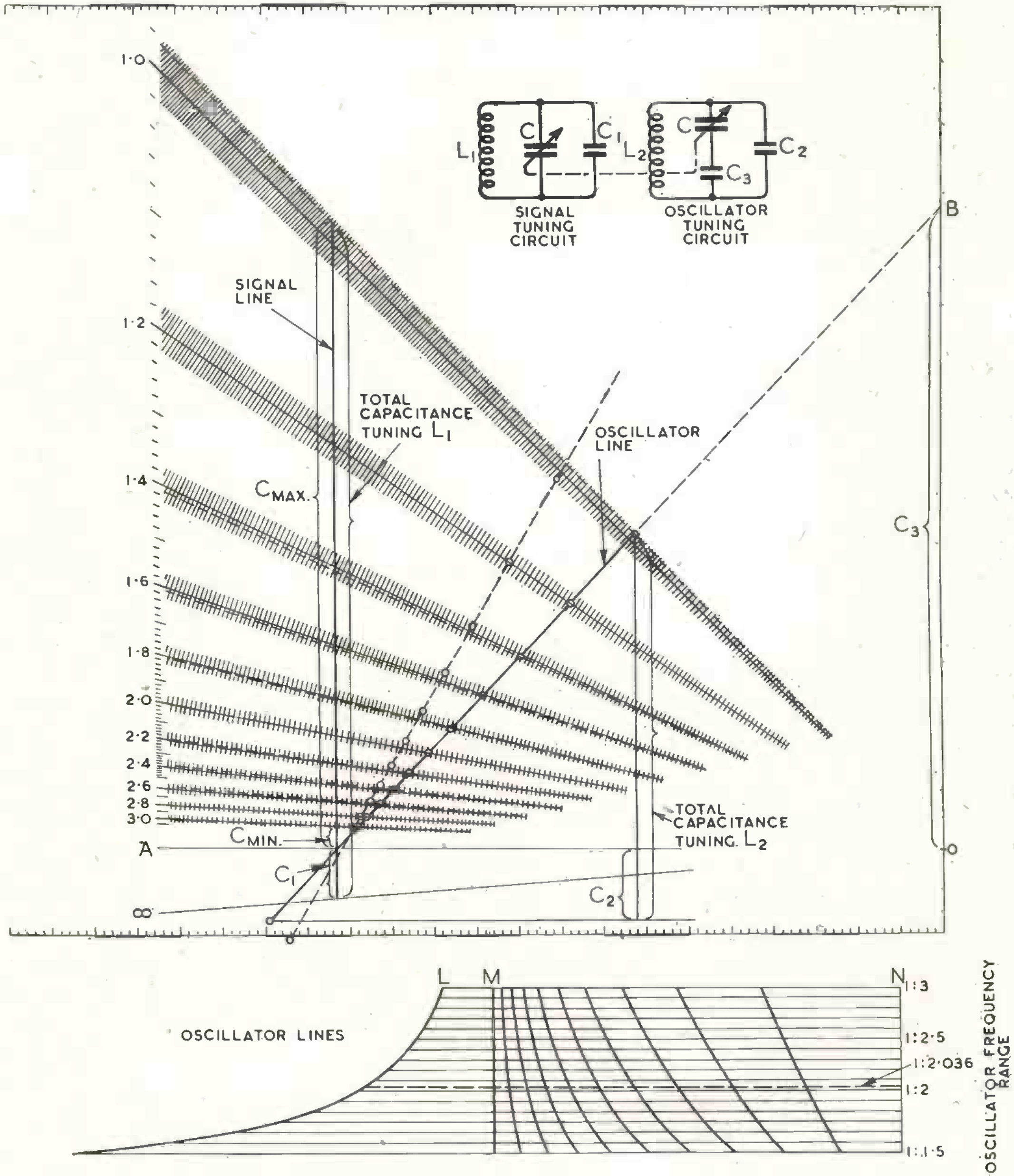


Fig. 1. *Diagram for signal/oscillator frequency tracking.*

which would not allow full advantage to be taken of the accuracy possible in carrying out measurements.

Henceforth, the line representing variation of the oscillator tuning capacitance will be referred to as the oscillator line and the line parallel to *OB* drawn through the point of intersection of the oscillator line and *OA* as the signal line. The signal line is shown drawn thick like the oscillator line. The purpose of the two scale lines at right angles to *OB* is to facilitate the drawing of lines parallel to *OB*.

For what follows, it should be borne in mind that, as the intermediate-frequency amplifier is so much more selective than the signal-frequency amplifier, the operation of tuning the receiver for optimum output, by means of the ganged variable condenser control, results essentially in the oscillator circuit tuning accurately to a frequency higher than the signal by the amount of the intermediate frequency. Whether the signal circuit is simultaneously in tune with the signal will depend on the accuracy of the tracking. The object will, therefore, be not to look for out-of-tune in the oscillator circuit, but in the signal circuit.

The appropriate oscillator line is traced on tracing paper or celluloid and superimposed on the upper part of the figure, being manoeuvred to a position where the points on it are as near as possible to the corresponding lines from *O*, judging their nearness as a percentage frequency difference. For example, a point situated halfway between an edge of the shaded area and the line would indicate a difference of frequency of approximately 1 per cent. and so on. After having found this position of the oscillator line, lines from *O* are drawn through the points *M* and *N* on it corresponding to the maximum and minimum oscillator frequencies to intersect the signal line which is positioned tentatively at this stage. The lengths from ∞ to the points of intersection are proportionate to C_1 in parallel with $C_{min.}$ and $C_{max.}$ respectively and knowing the ratio of $C_{max.}$ to $C_{min.}$ (taken as 25 : 1 for the example shown) the position of the line *OA* is fixed as already explained. The signal line is then moved so that the points of intersection on it of *OA* and the oscillator line coincide.

Having determined the positions of the oscillator and signal lines as above, the relative values of the capacitances C, C_1, C_2 and C_3 which will give the best tracking are found by simply measuring the lengths of the lines named on the figure. C_2, which is the oscillator coil trimming capacitance, is given by the distance parallel to *OB* of the lower end of the oscillator line from *OA*. The inductances L_1 and L_2 are obtained from the formula $L = 2.53 \times 10^4/f^2C$, where L and C are in microhenries and microfarads and f is in kilocycles per second, taking the total tuning capacitances as indicated. It should be noted from the figure that L_1 tunes to a slightly lower frequency than the signal at the lower end of the range and its value must be made accordingly larger.

To prove the values of C_2 and C_3 shown on the figure, the lines representing C and C_3 are parallel. By geometry the distance parallel to them from the intersection of the join of their ends by crossing, to either line joining their ends without crossing, is $CC_3/(C + C_3)$ which therefore is equivalent to C and C_3 in series. For $C_{max.}$ this distance is from *OA* to the top (or lowest frequency) division point of the oscillator line. For $C_{min.}$ it is the corresponding distance from the bottom (or highest frequency) division point and similarly for intermediate values of C. The capacitance C_2, which is in parallel, is added by extending these distances downwards from *OA* till they meet a line parallel to *OA* from the lower end of the oscillator line. From similarity of triangles the figure shows that this is the only value of C_2 in parallel with C and C_3 which varies the oscillator circuit capacitance in the required ratio given by the oscillator line.

If the straight line drawn from *O* through any point of division on the oscillator line intersects the signal line higher than the corresponding point of division on the latter, the signal circuit is tuned to a lower frequency than that of the signal for the particular part of the tuning range. The percentage error is found by dividing the frequency error represented by the distance between the above intersection of the signal line and the corresponding point of division on it, by the signal frequency represented by this point of division and multiplying by 100. It is, however, outside the scope of the article to go further in this direction and consider the effect of the distuning of the signal-frequency circuits on the performance of the receiver. By the shading of the 1 to 3 lines from O to a depth corresponding to a frequency variation of plus and minus 2 per cent. on either side, it is hardly necessary, as described above, unless for greater accuracy, to draw lines from O through the intermediate division points on the oscillator line to see where they intersect the signal line. If a division point is found on the upper edge of the shaded area, the frequency to which the signal circuit is tuned at the particular point of the range is 2 per cent. lower than that of the signal. If it is half-way between the edge and the centre line there is approximately a 1 per cent. tuning error and so on.

It will be seen from the figure that the tracking error will be zero at 2.8, 2 and 1.2,

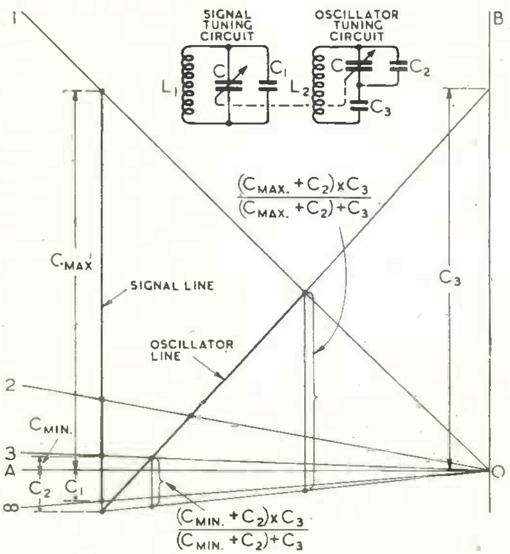


Fig. 2. *Diagram for signal/oscillator frequency tracking (C_2 across C).*

the sign of the error changing in passing through these points of the frequency range. Actually, a somewhat improved result is obtained by arranging the two outer of these points nearer to the ends of the range, but it is shown as above for clarity. For all practical purposes we may say that for the range 3 to 1, if accurate tracking is arranged at 2.9, 2 and 1.1, the best results will be secured. There will, however, be no great loss of tracking accuracy if it is desired to have the two outer points right at the ends of the range.

The oscillator line shown dotted on the figure illustrates the effect of varying the L/C ratio of the oscillator circuit maintaining the tracking error zero at the two outer points, 2.8 and 1.2 of the range. The oscillator inductance, L_2, is too low and the signal circuit, for the range within the two points, is seen to be tuned to too low a frequency.

Instead of connecting the trimming condenser C_2 directly across the inductance, L_2, as in Fig. 1 (inset), which is the more usual practice, it may be connected across the ganged condenser, C. In this case, the respective dimensions on Fig. 1, which give the proportions of the various circuit constants, are shown separately for clarity in Fig. 2, for accurate tracking at 1, 2 and 3.

As a check that the framework of the figure has been constructed correctly, an example should be worked out from it and checked by algebra. The particulars of the formulae can be obtained from the Appendix of W. T. Cocking's "Wireless Servicing Manual."* The following example may be tried :—

Range 500/1500 kc/s ; I.F. 465 kc/s ; Tracking correct at 500, 1000 and 1500 kc/s.

Capacitances in $\mu\mu$F. Inductances in μH.

$C_{max.}$ 500 ; C_2 58.9 L_1 187.6 ; L_2 83.85
$C_{min.}$ 20 ; C_3 554
C_1 40

With a carefully constructed figure, it will be found that results are obtainable which are sufficiently accurate for most design and testing purposes.

* See also M. Wald, *Wireless Engineer*, March, 1940, p. 105, April, 1941, p. 146 ; Editorial, *Wireless Engineer*, April, 1942 ; A. L. Green, *Wireless Engineer*, June, 1942, p. 243 ; Ruby Payne-Scott and A. L. Green, *Wireless Engineer*, July, 1942, p. 290. An extensive bibliography is given in *Wireless Engineer*, June, 1942, p. 250.

I.E.E. Meetings

Sir Edward Appleton, M.A., D.Sc., F.R.S., will lecture on "Radio Exploration of the Ionosphere" at the next meeting of the I.E.E. Wireless Section, on Wednesday, April 7th.

The subject to be discussed at the Informal Meeting of the Wireless Section on Tuesday, April 20th, is "Metal Rectifiers and their Applications to Radio and to Measurements."

Both meetings will begin at 5.30.

Attenuation and Phase-Shift Equalisers*

Semi-Graphical Method of Designing Equalising Networks

By W. Saraga, Ph.D.

SUMMARY.—After a general discussion of equaliser design problems a semi-graphical design method is described.

If the attenuation/frequency curve or the phase-shift/frequency curve of a telephone line, a wave filter, etc., or of a complex communication system, does not possess the shape desired for its most effective operation (e.g. a flat attenuation/frequency curve for a telephone line, a rectangular attenuation characteristic for a filter), this can be corrected by adding to the line, filter, or system in question an equalising network which is so designed that the overall attenuation, or phase shift, of original network and equaliser, as a function of frequency, has the desired shape over a certain specified frequency range. Subtraction of the curve produced by the uncorrected network from the desired overall curve gives the required equaliser curve.

In practical design problems this required curve is usually given as a graph. The designer has provisionally to select a probably suitable type of network and to find such values for its elements that the maximum deviation from the required curve of the curve produced by this network becomes as small as possible. This curve approximation problem is usually solved by trial and error.

However, a trial and error procedure is often undesirable. Therefore a semi-graphical method, "Visual Curve Fitting," has been developed which eliminates trial and error either completely or at least to a very great extent; in the latter case the remaining trial-and-error procedure is of a particularly simple kind.

A description and a mathematical discussion of this method and two design examples are given.

1. Survey of the New Method

(a) Introduction

ATTENUATION and phase-shift equalisers are essential components of any electrical communication system. Ordinarily it is desired that any such system possesses a flat attenuation/frequency curve and that its phase shift increases proportional to the frequency f. If this is not the case, a correcting or equalising network can be inserted into the system with such an attenuation α or phase shift β that the overall frequency characteristic of the system including the equaliser has the desired shape. It is also desirable that the overall attenuation level of the corrected system is as low as possible (see Fig. 1a). Since equalisers, being passive networks, can only increase, never decrease, the overall attenuation, the theoretically lowest possible level of a corrected flat overall curve is equal to the maximum level of the uncorrected curve; in this case the required equaliser characteristic is equal to the difference between the uncorrected curve and its maximum value. Frequently, however, an equaliser which produces a curve of the same shape, but at a higher α level, can be designed with fewer elements than if it has to produce the original difference curve. On the other hand, a higher loss level requires more amplification. Therefore, whereas the required *shape* is usually clearly defined, the required *level* can sometimes only be fixed after some tentative design work for different levels has been carried out showing the relation between level and number of elements.† In addition to the required shape and level some tolerance limits have to be given because, for producing the required curve accurately, networks possessing an uneconomically great number of elements would often be necessary.

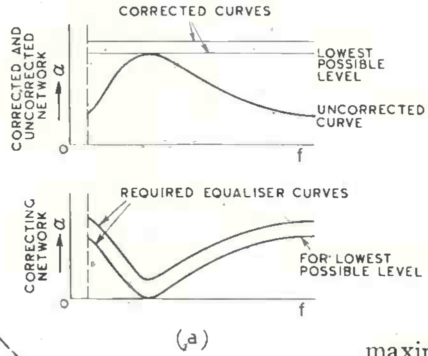
This article does not deal with the more general questions of fixing the level of a required curve and fixing the tolerance limits.

* MS. accepted by the Editor, October, 1942.

† Similar problems arise sometimes in phase shift equaliser design.

It is chiefly concerned with the design problem proper, assuming that a required curve over a specified frequency range is given and a tolerance band enclosing this curve is specified.

Zobel[1] has developed a general design method for correcting networks. To begin with, a "type of network"† probably suitable for the required curve has to be chosen provisionally. Then n points on the required curve are chosen where n is the number of independent‡ network elements or design parameters, and such values for the n elements have to be found that the curve produced by this network goes through the n selected points. Between these points, and between the outer points and the boundaries of the specified frequency range, the designed curve deviates from the required one, and there are at least $(n + 1)$ deviation maxima (see Fig. 1b). The values of these maxima depend on the position of the n selected points, and in order to make the maxima as small as possible a number of consecutive trials have to be made, each trial starting from a different set of points until a satisfactory approximation to the required curve has been obtained.



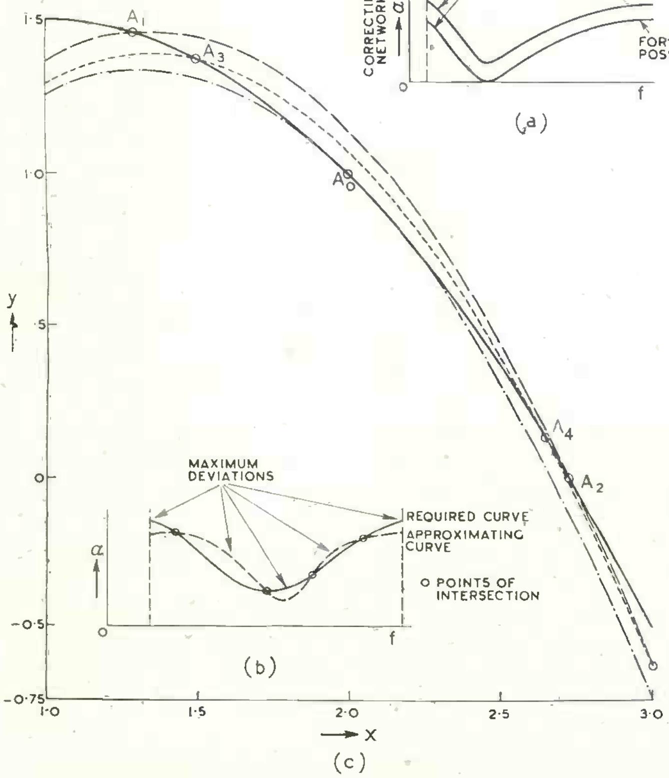


Fig. 1.—*In (c) the full line is the required curve,* $y = 1 + x - \frac{x^2}{2}$, $1 \leq x \leq 3$. *The approximating curves* $y = Px + Qx^2$ *are as follows: broken line, least maximum deviation approximation,* $P = 2.143$, $Q = -0.785$, $|\Delta|_{max} = 0.142$; *dotted line, least mean square deviation approximation,* $P = 2.041$, $Q = -0.751$, $|\Delta|_{max} = 0.21$; *chain-dot line, approximation by Taylor's theorem in* A_0, $P = 2$, $Q = -0.75$, $|\Delta|_{max} = 0.25$, Δ = *deviation.*

(b) *Visual Curve Fitting*

Such a trial-and-error procedure is sometimes very tedious and may lead to a tendency to use more elements than necessary in order to simplify the design work. Therefore other design methods have been developed, but none of them seems to have overcome this difficulty in a general way. It might be thought that only an analytical method—like Cauer's design method for filters—would be capable of eliminating trial and error. It is, however, possible to develop a

[1] This and other numbers refer to the references at the end of the article.

† "Type of network" means here: a network for which only its structure and the character of its elements (resistance, inductance, capacitance), but not their numerical values, are fixed.

‡ In so-called "constant resistance" type equalisers, the actual number of elements is usually twice the number of independent elements because the elements of the shunt arm are determined by those of the series arm and vice-versa; as an example see Fig. 11 (the value of R_o is given).

non-analytical design method which does not proceed by trial and error, i.e. not by a number of separate trials, but by one single, comparatively short, visual curve-fitting process. This method may conveniently be called "visual curve fitting."

If networks with more than two independent elements have to be designed, visual curve fitting requires some auxiliary mechanical, optical or electrical equipment. If such equipment is not available, a certain amount of trial and error has to be carried out, but it is much simpler and, therefore, takes much less time and effort than ordinary trial and error.

Whereas Zobel's design method is based on n single points representing the required curve, visual curve fitting is based on the required curve as a whole and on the specified tolerance band enclosing this curve. Sometimes no tolerance band is specified, but it is desired to approximate the required curve with a given type of network within the smallest possible tolerance range; in such cases this band has to be *roughly* estimated beforehand for use as basis of design.

The required curve as well as the tolerance band are ordinarily given in the attenuation/frequency (or phase-shift/frequency) plane, i.e. the attenuation α is plotted as a function

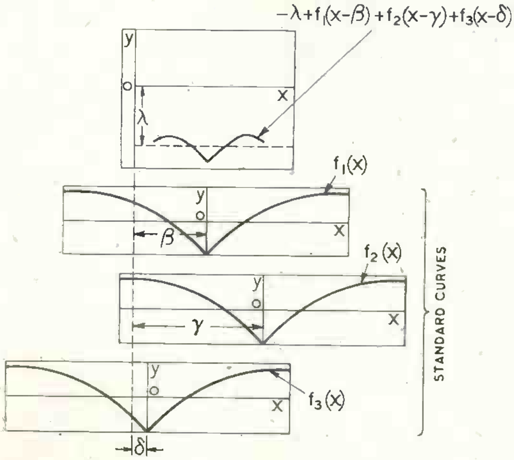


Fig. 2.—*It should be noted that the standard curves have not always an identical shape.*

of the frequency f. Usually, in order to make the visual curve fitting method workable, it is necessary to transfer the tolerance band from this αf plane to another plane,

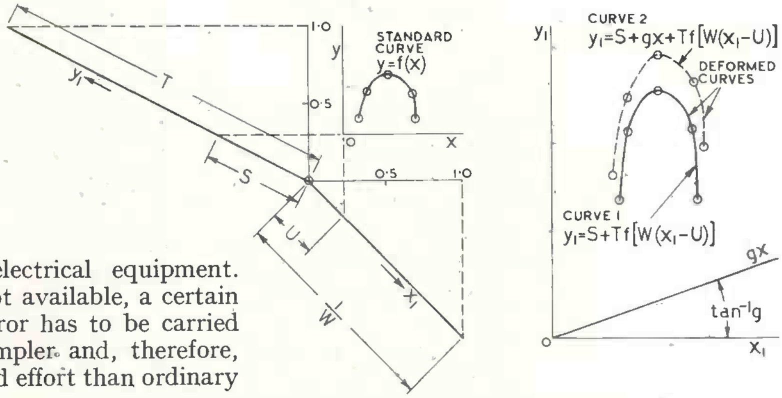


Fig. 3.—*Curve* 1 *is derived from standard curve by shifting and stretching. Curve* 2 *is derived from standard curve by shifting, stretching and shearing.*

a yx plane, where y as well as x are functions of α and f which are more suitable for visual curve fitting.

The design can be carried out in two different ways:—

(*a*) *Summation method:* each type of network is characterised by a set of $(n-1)$ or n "standard curves"; the design consists in shifting these curves along the two co-ordinate axes until the sum of the standard curves (see Fig. 2) fits the required tolerance band in the chosen yx plane. Then the values of the network elements can be derived from the position of the standard curves.

(*b*) *Shaping method:* each type of network is characterised by one single "standard curve"; the design consists in shifting this curve along the two axes and in deforming it according to certain rules (see below) until it fits the tolerance band in the selected yx plane. The values of the network elements can be derived from the position and the deformation of the standard curve. The necessary deformation consists of (see Fig. 3):

(1) Multiplying the ordinate and/or abscissa values of the standard curve by constant factors T and $1/W$ respectively; this may be called "stretching."

(2) Adding a straight line gx to the curve; this may be called "shearing."

In order to eliminate trial and error two conditions have to be fulfilled:

(1) The position of the summation method standard curves or the position and deformation of the shaping method standard curve must be continuously variable;

(2) The result of any such variation, i.e. the curve representing either the sum of the standard curves or the deformed and shifted standard curve, must be observable instantaneously.

For networks with one or two independent elements only, the shaping method requires only shifting, but not deformation, of the standard curve; or even simpler: only a straight line has to be drawn through the tolerance band; in this case, both conditions can easily be fulfilled.

For networks with more than two elements, not only curve shifting but also addition of standard curves or deformation of one standard curve has to be carried out; in order to fulfil conditions (1) and (2) some auxiliary mechanical, optical or electrical equipment is needed (see Section 6). If such equipment is not available, curve fitting by means of addition and deformation constitutes a trial-and-error procedure. However, the operations involved are the simplest possible: addition, reading values of coordinates of points off some sloping scales, plotting of curves; on the other hand in ordinary trial-and-error procedure each separate trial frequently involves lengthy calculations. Moreover, the number of trials necessary is considerably reduced compared with ordinary trial-and-error methods, because the representation of the required band in the yx plane makes it possible to carry out a large part of the fitting work by visual inspection.

2. Curve Approximation

(a) Network Analysis, Synthesis and Design

A discussion of electrical networks is usually divided into "network analysis" and "network synthesis." In network analysis a network is given and it is required to find one or more characteristic network functions, e.g. its attenuation as a function of frequency. In network synthesis one or more network functions are specified and it is required to find the corresponding network. The first step in network synthesis is to find out which type of functions may be prescribed if physically possible networks are to be obtained. It can, for example, be shown that for any physical passive linear network the attenuation α in decibels as function of the frequency f has to be of, or transformable into, the form

$$\alpha = 10 \log_{10} \frac{P_0 + P_2 f^2 + P_4 f^4 + \ldots}{Q_0 + Q_2 f^2 + Q_4 f^4 + \ldots} \quad \ldots \quad (1)$$

where $P_0, P_2, P_4, \ldots, Q_0, Q_2, Q_4, \ldots$, must have such values that for any frequency, $\alpha \geq 0$.

Since, e.g.,

$$\alpha = 10 \log_{10} \frac{1 + 6f^2 + 4f^4}{1 + 5f^2 + 4f^4} \quad \ldots \quad (2)$$

is a particular case of (1), a corresponding physical network can be found. The actual practical synthesis of such a network can be carried out either by means of some general synthesis method (see for instance Gewertz [2]), or by preparing a list of network functions corresponding to given networks and using the list in the opposite direction for finding networks corresponding to given functions.

In practical network design, particularly in equaliser design problems, the required network function, e.g. the attenuation/frequency function, is usually not given as a particular case of equation (1) but as an arbitrary function or a graph and specified for a limited frequency range only. Then the first step is to find an analytical expression for the graph in the form of equation (1). Provided the required α is ≥ 0 over the whole specified frequency range, such an expression can always be found by using a sufficiently large number of coefficients $P_0, P_2 \ldots, Q_0, Q_2 \ldots$—if necessary an infinite number. However, for practical purposes a solution with a great number of coefficients is usually uneconomical because the number of network elements * increases *roughly* in the same proportion as the number of coefficients. In such cases the ideal net-

* The number of inductances and capacitances is usually more important than the number of resistances.

work producing exactly the required curve has to be replaced by an economically possible network producing an approximation to the required curve only. The designer, though using the results of network synthesis, is in practice chiefly concerned with finding the "best possible approximation" (see below), when using a given type of network. Therefore "design method" very often means "approximation method." In this article a new approximation method is developed. As far as network synthesis is concerned, Zobel's results are used and applied.

(b) Types of Approximation

In equaliser design "best approximation" means that the maximum absolute deviation of the designed curve from the required one over the specified frequency range is as small as possible for the chosen type of network. This particular type of approximation is only one of various known types. In order to show their differences some types are shown in Fig. 1c. The required curve has been taken as $y = 1 + x - \frac{1}{2}x^2$ over the range $x_1 \ldots x_2$ with $x_1 = 1$ and $x_2 = 3$. The approximating curve has been chosen to be of the form $y = Px + Qx^2$ and P and Q have to be determined in such a manner that the particular requirements of the type of approximation in question are met. The types of approximation shown are:

(1) *A Taylor series* at the (arbitrary) point A_0. Such an approximation makes the required and the approximating function as well as their differential coefficients (up to as high an order as possible) equal to each other at one selected point A_0. This type of approximation is particularly good at and near the point A_0, but rather bad at the boundaries of the specified range.

(2) *Least mean square deviation.* This type of approximation is defined by

$$M_2 = \sqrt{\frac{1}{x_2 - x_1}\int_{x_1}^{x_2} \Delta^2 dx} \quad \ldots \quad \ldots \quad (3)$$

being a minimum where Δ is the deviation from the required curve for any x. The maximum deviation occurring is greater than that occurring in the following example, where the maximum deviation has been made as small as possible.*

(3) *Least maximum deviation.* This type is called Tschebyscheff-approximation; it is the type required in network design and will be dealt with in this article.

(4) *Zobel-approximation.* This type cannot be represented by one single clearly defined curve. Zobel obtains his approximating curves by choosing n points (for this example $n = 2$) of exact coincidence of required and designed curve. Thus the curves representing type (2) and type (3) which intersect the required curve in two points are also representing Zobel-approximations. The selection of the two points of intersection determines the magnitude of the maximum deviation occurring. Only if A_1 and A_2 have been selected, the least possible maximum deviation is obtained.

(c) Methods of Approximation

The ideal design method would be an analytical method—provided that it does not involve too much work—which derives from the required curve those values of the elements of any chosen type of network which give the least possible maximum deviation. For the design of filter networks such a method has been developed by Cauer [4]. Filter networks can be regarded as a particularly simple case of equaliser networks for which the required αf function is 0 for the pass bands and ∞ for the stop bands. Probably Cauer's method could be extended to cover the general equaliser network where the required αf function may have any shape required. At present, however, such a method seems to be not yet available †.

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* If (3) is replaced by

$$M_{2n} = \sqrt[2n]{\frac{1}{x_2 - x_1}\int_{x_1}^{x_2} \Delta^{2n}\, dx} \quad \ldots \quad \ldots \quad (4)$$

where n is an integer $\gg 1$, M_{2n} is much more determined by the larger than by the smaller values of Δ. For $n \to \infty$ the condition "M_{2n} is a minimum" becomes equivalent to the condition "Δ_{max} is a minimum" which defines the third type of approximation described here (see also M. Wald [3]).

† For the simple network in Fig. 11, which possesses two independent parameters only, an analytical design method can easily be developed; it is described in the appendix.

Zobel's design method, as described in Section 1a, proceeds by trial and error: a tentative network is designed, its attenuation/frequency (or phase-shift/frequency) curve calculated, and compared with the required curve; if the result is not satisfactory, one—or more—new tentative designs, each starting from another set of n points, have to be made. *

In some cases, such trial and error becomes inconvenient and cumbersome. There are cases in which the "tiring" effect of trial and error favours a tendency to use more network elements than necessary, as any approximation can be more easily achieved if more design parameters are available.

Pyrah [5]—and before him in a slightly different form Tucker [6]—has suggested another design method in which a variable equaliser is adjusted until its attenuation/frequency function fits the required curve. The position of the controls of the variable network indicates the required values of the elements.

As it is necessary to adjust the network, to measure the attenuation, to compare it with the required curve, to readjust the network, etc., until a satisfactory approximation is obtained, this method is also a trial-and-error method, but each separate trial takes much less time and is much less tiring than in Zobel's method. However, suitable variable networks and measuring apparatus are not always available, when an equaliser has to be designed.

Curve charts have been frequently suggested for use in equaliser design. The required curve, on a transparent sheet, can be put on top of the curve chart in order to find the best fitting curve of the chart. Usually a great number of chart curves or curve families have to be plotted, in order to prevent large errors, arising from interpolation between the curves.

Pyrah [5] has shown for two particular types of networks that the number of chart curves to be plotted can be considerably reduced by using a logarithmic frequency scale; a variation of one of the parameters can then be obtained by shifting the family of curves, representing various values of the other parameter, along the logarithmic frequency scale. This method does not seem to involve a trial-and-error procedure, but leads directly to the best value of this parameter. The other parameters have to be determined by interpolation between curves of the chart.

From this review of design methods, three facts emerge which are important for the development of a new method.

(1) Network designers are definitely interested in a method involving no, or at least little, trial and error.

(2) Design methods using variable networks require a minimum of trial and error only.

(3) Shifting a chart curve until it fits a required curve does not seem to involve trial and error.

3. Development of the Visual Curve Fitting Method

The basic idea of the new method is to simulate the variable network method without using variable networks. Two such methods have been developed: "summation method" and "shaping method." In the summation method each type of network is represented by one or more "standard curves," the number of curves usually being equal to the number of design parameters minus one. A variation of the element values in the variable network method is represented in the summation method by shifting the curves along both axes. Measuring the network attenuation (and plotting it) is replaced in the summation method by adding the curves (and plotting their sum). This is shown in Fig. 2.*

* For most networks discussed by Zobel the network elements can be derived from the coordinates of the selected n points by solving a set of n linear equations. For some networks the equations are not linear and instead of solving them, Zobel prefers to determine the value of one element—a resistance—beforehand and to apply his usual linear equation method to the remaining $(n - 1)$ elements only. In some cases the resistance value has to be determined by trial and error. If so desired, this *additional* trial and error can easily be removed by applying Zobel's method to all n points and solving the non-linear equations graphically.

* The shifting of standard curves, and their summation, for plotting and design purposes, has recently been described by Truscott [7] in an article which was published after the present article had been written. Truscott deals with circuit problems for which the use of logarithmic scales for frequency and impedance is of particular advantage.

In the shaping method each type of network is represented by a single standard curve. A variation of the network element values is represented in the shaping method by shifting and deforming the standard curve: the deformation consists of "stretching" and "shearing" as described in Section 1(b). As the stretching of a curve cannot be carried out physically without auxiliary equipment, the scales are apparently compressed instead, by rotating them physically (see Fig. 3); shearing, too, can be represented by means of a rotatable scale (see Fig. 3). Measuring the attenuation in the variable network method, and plotting it, is replaced by reading the co-ordinates of the standard curve points off the sloping scales, and replotting it in a normal co-ordinate system.

So far practical design experience seems to indicate that, whenever both methods can be used for the solution of a given problem, the shaping method is somewhat simpler than the summation method.

If networks with only one or two independent parameters have to be designed, a particularly simple design method becomes possible which involves only the shifting of one single standard curve.[7]

To make the summation method and shaping method workable, it is obviously necessary to express the attenuation/frequency (or phase-shift/frequency) function of each type of equaliser network in a form which can be represented either as sum of shiftable standard curves or as one single shifted and deformed (stretched and sheared) standard curve. Moreover, it is necessary that the characteristic frequency functions in this special form possess only *real* and *independent* parameters (this will be explained in detail in the following section).

These mathematical requirements can rarely be met without transferring the αf curve (or the βf curve) from the original αf plane (or βf plane) to another plane, a yx plane, where both y and x are functions of α and f, so chosen as to meet these requirements. For comparison of a curve tentatively designed in the yx plane, with a required αf curve, the required curve too has to be transferred to the yx plane. However, as the deviations in the αf plane, not those in the yx plane, are the criterion according to which any approximation is judged, it is more useful to transfer a "required tolerance band" enclosing the required curve, instead of this curve itself, to the yx plane. If the best possible approximation in the αf plane is required, and therefore no definite tolerance band specified, a probable band can be estimated. Even if this estimate is far out, one transferred tolerance band is usually sufficient for deciding which of several approximating curves is the best. This method of comparison is more practical than transferring each tentative yx curve to the αf plane for comparison with the original required αf curve.

In the new method the network elements are calculated only *after* a satisfactory approximating curve has been found, whereas in conventional trial-and-error procedures each single trial ordinarily involves the calculation of new tentative network elements *before* the calculation of each tentative approximating curve and its comparison with the required curve.

The practical application of the new method results in a great saving of time and effort; it can be carried out by comparatively unskilled people.

4. Mathematical Discussion

(a) Summation Method

The attenuation α in db. of any equaliser section can always be expressed in the form

$$10^{\frac{\alpha}{10}} = \frac{P_0 + P_2 f^2 + P_4 f^4 + \ldots}{Q_0 + Q_2 f^2 + Q_4 f^4 + \ldots} \quad \ldots \quad (5)$$

where the coefficients P_0, P_2, .. , and Q_0, Q_2, .. depend on the network elements and have always such values that $10^{\frac{\alpha}{10}} > +1$.

According to a fundamental theorem of algebra, (5) can always be written in the form

$$10^{\frac{\alpha}{10}} = \rho \frac{(f^2+\beta)(f^2+\gamma)\ldots}{(f^2+\delta)(f^2+\epsilon)\ldots}$$

$$= |\rho| \frac{|f^2+\beta|\,|f^2+\gamma|\ldots}{|f^2+\delta|\,|f^2+\epsilon|\ldots} \quad \ldots \quad (6)$$

or

$$\frac{\alpha}{10} = \log_{10}|\rho| + \log_{10}|f^2+\beta| + \log_{10}|f^2+\gamma| + \ldots - \log_{10}|f^2+\delta| - \log_{10}|f^2+\epsilon| - \ldots \quad \ldots \quad (7)$$

where β^*, γ, δ, ϵ . . . are constants depending on P_0, P_2, P_4 . . . and Q_0, Q_2, Q_4 . . . .

In some cases it is convenient to use $1/f^2$ instead of f^2 as variable. According to equation (7), $\frac{\alpha}{10}$, as a function of f^2, can be represented graphically as sum of a constant $\log_{10}|\rho|$, and a number of curves $\pm \log_{10}|f^2|$, as functions of f^2, shifted by amounts $-\beta$, $-\gamma$, $-\delta$, $-\epsilon$, . . along the f^2 axis, *provided that* $\beta, \gamma, \delta, \epsilon$ . . . *are real*. This is a method for *quick plotting* of α if $\beta, \gamma, \delta, \epsilon$, . . are given.

If it is desired to use this graphical representation for *design* purposes, where the required α is given and $\beta, \gamma, \delta, \epsilon$, . . have to be determined by trying out various values, *it is necessary that* $\beta, \gamma, \delta, \epsilon$, . . *are independent of each other.*

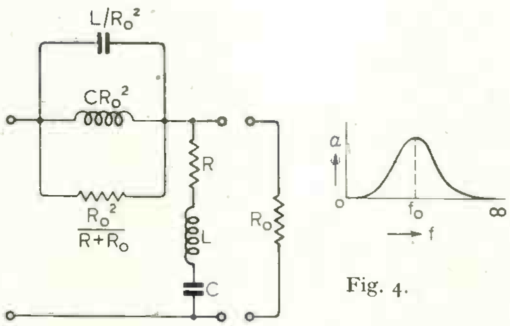

Fig. 4.

Actually, in the αf plane, $\beta, \gamma, \delta, \epsilon$, . . are neither always real nor always independent. Some examples will be discussed in order to show how these difficulties can be overcome by transferring the analysis and the design from the αf plane to a suitable yx plane.

1st example : The ladder type network in Fig. 4 which corresponds to Zobel's networks 5a and 5b has an attenuation

$$\alpha = 10 \log_{10} \frac{\left(1 + \frac{R}{R_0}\right)^2 f^2 + \frac{4\pi^2 L^2}{R_0^2}(f^2 - f_0^2)^2}{\left(\frac{R}{R_0}\right)^2 f^2 + \frac{4\pi^2 L^2}{R_0^2}(f^2 - f_0^2)^2} \quad \ldots \quad (8)$$

where $f_0^2 = \frac{1}{4\pi^2 LC}$. (8) can be written in the desired form of equation (6)

$$\alpha = 10 \log_{10} \rho_1 \frac{(f^2 + \beta_1)(f^2 + \gamma_1)}{(f^2 + \delta_1)(f^2 + \epsilon_1)} \quad (9)$$

Equation (9) has five parameters : ρ_1, β_1, γ_1, δ_1, ϵ_1 ; they cannot be independent because the network has only three independent elements R, L and C. Moreover, as can be derived from (8), β_1, γ_1, δ_1, ϵ_1 are not necessarily real.—From (8) we can obtain

$$y = \frac{f^2}{10^{\frac{\alpha}{10}} - 1} = \frac{R^2}{R_0(R_0 + 2R)} f^2 + \frac{4\pi^2 L^2}{R_0(R_0 + 2R)}(f^2 - f_0^2)^2 \quad \ldots \quad (10)$$

which can be written in the form

$$y = \rho_2(f^2 + \beta_2)(f^2 + \gamma_2) \quad \ldots \quad (11)$$

This is an expression with only three parameters, as desired ; however, two of them, β_2 and γ_2 may still be not real.

This difficulty can be overcome as follows. The meaning of equations (10) and (11) is that a function $\rho_2(f^2 + \beta_2)(f^2 + \gamma_2)$ has to be found, which falls within the tolerance band of the required αf function transferred to the $\left(\frac{f^2}{10^{\frac{\alpha}{10}} - 1}\right)$-$(f^2)$ plane. Now, $\rho_2(f^2 + \beta_2)(f^2 + \gamma_2)$ intersects or touches the f^2 axis if β_2 and γ_2 are real, and does not intersect it if they are complex. Therefore, if the required band intersected the f^2 axis, any *suitable* function $\rho_2(f^2 + \beta_2)(f^2 + \gamma_2)$ would also have to intersect the f^2 axis, as it has to be inside the tolerance band. Unfortunately, $y = \frac{f^2}{10^{\frac{\alpha}{10}} - 1}$ can never intersect the f^2 axis though it may touch it, because it is always ≥ 0, α being always ≥ 0. However, by subtracting from both sides of equation (10) a suitable arbitrary function of f^2, the required band can always be made to intersect the f^2 axis ; the resulting right-hand side of equation (10) (after subtraction of the arbitrary function) must have real roots. If we choose as arbitrary function Kf^2, we obtain from (10)

$$y - Kf^2 = f^2\left(\frac{1}{10^{\frac{\alpha}{10}} - 1} - K\right) = \rho(f^2 - \beta)(f^2 - \gamma) \quad \ldots \quad (12)$$

* This constant β should not be confused with the phase shift β.

with

$$\rho = \frac{(2\pi L)^2}{R_0(R_0 + 2R)};$$

$$\beta, \gamma = f_0^2 + \frac{E}{2\rho} \pm \sqrt{\frac{E^2}{4\rho^2} + \frac{f_0^2 E}{\rho}};$$

$$E = K - \frac{R^2}{R_0(R_0 + 2R)} \qquad \ldots \quad (13)$$

If K has been so chosen that the tolerance band for $(y - Kf^2)$ intersects the f^2 axis, β and γ must be real. From (12) follows

$$\log|y - Kf^2| = \log|\rho| + \log|f^2 - \beta| + \log|f^2 - \gamma| \qquad \ldots \quad (14)$$

which, as a function of f^2, can be represented as sum of a constant $\log|\rho|$, and two curves representing $\log|f^2|$ as functions of f^2 which are shifted by $+\beta$ and $+\gamma$ along the f^2-axis. When values for ρ, β and γ have been found which fit the required tolerance band, L, C, R are obtained by

$$f_0^2 = \sqrt{\beta\gamma}, \qquad T = 4\rho f_0^2,$$

$$S = K - \rho(\beta + \gamma - 2f_0^2),$$

$$R = R_0(S + \sqrt{S + S^2}),$$

$$L = \frac{1}{4\pi f_0}\sqrt{TR_0(R_0 + 2R)},$$

$$C = \frac{1}{4\pi^2 L f_0^2} \qquad \ldots \quad \ldots \quad (15)$$

By introducing a new variable, $x = \log_e f$, equation (12) can be transformed into two other forms which can be represented graphically as the sum of a suitable constant and two standard curves shiftable along the abscissa axis. From (13) follows that β and γ are always positive. Therefore (12) can be written as

If in (12) β or γ were negative, sinh in (16) would have to be replaced by cosh, and $(e^{2x} - 1)$ in (17) by $(e^{2x} + 1)$. Similarly for other types of networks f^2 or $1/f^2$ as variable can be replaced by $x = \log_e f$, and the standard curves $\log|f^2|$ by $\log|\sinh x|$ and $\log \cosh x$, or $\log|e^{2x} - 1|$ and $\log(e^{2x} + 1)$.

Another representation of the αf curve of this network, without introduction of a constant K, can be obtained from equation (10):

$$y = \frac{4\pi^2 L^2}{R_0(R_0 + 2R)} f_0^4 + \left[\frac{R^2}{R_0(R_0 + 2R)} - \frac{8 f_0^2 \pi^2 L^2}{R_0(R_0 + 2R)}\right] f^2 + \frac{4\pi^2 L^2}{R_0(R_0 + 2R)} f^4$$

which is of the form

$$y = \lambda + a\, e^{2 \log_e f} + b\, e^{4 \log_e f}$$

or $\quad y = \lambda + e^{2(x + \phi)} + e^{4(x + \phi)}$

Thus the necessary standard curves are e^{2x} and e^{4x}.

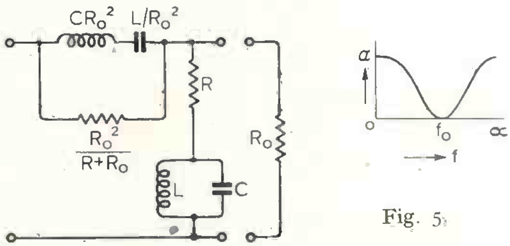

Fig. 5

2nd example: Network shown in Fig. 5. Its attenuation is

$$\alpha = 10 \log_{10} \frac{(R_0 + R)^2\, 4\pi^2 C^2\, (f^2 - f_0^2)^2 + f^2}{R^2\, 4\pi^2 C^2\, (f^2 - f_0^2)^2 + f^2},$$

$$f_0^2 = \frac{1}{4\pi^2 LC}.$$

$$\frac{y - Kf^2}{f^2} = \left(4\rho\sqrt{\beta\gamma}\right)\left[\tfrac{1}{2}\left(\frac{f}{\sqrt{\beta}} - \frac{\sqrt{\beta}}{f}\right)\right]\left[\tfrac{1}{2}\left(\frac{f}{\sqrt{\gamma}} - \frac{\sqrt{\gamma}}{f}\right)\right]$$

or

$$\log\left|\frac{1}{10^{\frac{\alpha}{10}} - 1} - K\right| = \log\rho' + \log|\sinh(x - \tfrac{1}{2}\log_e\beta)| + \log|\sinh(x - \tfrac{1}{2}\log_e\gamma)| \qquad \ldots \quad (16)$$

with $\rho' = 4\rho\sqrt{\beta\gamma}$. Two standard curves $\log|\sinh x|$ are necessary.

Another form for (12) is $(y - Kf^2) = (\rho\beta\gamma)(f^2/\beta - 1)(f^2/\gamma - 1)$

or

$$\log|f^2/(10^{\frac{\alpha}{10}} - 1) - Kf^2| = \log\rho'' + \log|e^{2(x - \log_e\sqrt{\beta})} - 1| + \log|e^{2(x - \log_e\sqrt{\gamma})} - 1| \quad (17)$$

with $\rho'' = \rho\beta\gamma$. Two standard curves $\log|e^{2x} - 1|$ are necessary.

Therefore

$$y = 1/(10^{\frac{a}{10}} - 1) = \frac{S(f^2 - f_0^2)^2 + bf^2}{(f^2 - f_0^2)^2} \quad \ldots \quad \ldots \quad (18)$$

where

$$S = \frac{R^2}{R_0(R_0 + 2R)}, \quad b = \frac{1}{R_0(R_0 + 2R)4\pi^2 C^2},$$

$$f_0^2 = \frac{1}{4\pi^2 LC} \quad \ldots \quad \ldots \quad (18a)$$

(18) can be written in the form of equation (6), viz.

$$y - K = \rho \frac{(f^2 - \beta_1)(f^2 - \gamma_1)}{(f^2 - f_0^2)^2} \quad \ldots \quad (19)$$

where K is so chosen that the required $(y - K)$ band intersects the f^2 axis, in order to make β_1 and γ_1 real. However, equation (19) has four parameters whereas the network has only three independent elements, L, C and R. Therefore, ρ, β_1, γ_1, f_0^2 must be interdependent, and from (18) and (19) follows

$$\beta_1, \gamma_1 = f_0^2 - \frac{b}{2\rho} \pm \sqrt{\frac{b^2}{4\rho^2} - \frac{bf_0^2}{\rho}} \quad \ldots \quad (20a)$$

$$\text{and} \quad \beta_1 \gamma_1 = f_0^4 \quad \ldots \quad \ldots \quad (20b)$$

Thus if equation (19) were used as basis of design and the best positions for three standard curves, $\log|f^2|$, $\log|f^2|$, and $2\log|f^2|$, had to be determined, it would be necessary to keep the shifts of these curves along the f^2-axis, viz., f_0^2, β_1, γ_1, always in the correct relationship given by (20b).

No way has so far been found to eliminate one of the four parameters completely and at the same time to obtain a form suitable for the summation method; but by using $x = \log_e f$ instead of f^2, as variable, the relation between three of the new parameters becomes much simpler. From (19) follows

$$y - K = \rho \frac{\sinh[x - (\frac{1}{2}\log_e \beta_1)] \sinh[x - (\frac{1}{2}\log_e \gamma_1)]}{\sinh^2[x - (\log_e f_0)]} \quad \ldots \quad \ldots \quad (19a)$$

where $\rho = \dfrac{R^2}{R_0(R_0 + 2R)} - K > 0$. The necessary standard curves are

$\log|\sinh x|$, $\log|\sinh x|$,

and $2\log|\sinh x|$. There are four design parameters: ρ, $(\frac{1}{2}\log_e \beta_1)$, $(\frac{1}{2}\log_e \gamma_1)$, $(\log_e f_0)$; but the relation between them is much more convenient for design purposes than (20b): $(\log_e f_0) - (\frac{1}{2}\log_e \beta_1) = (\frac{1}{2}\log_e \gamma_1) - (\log_e f_0)$. When the best values of ρ, β_1, γ_1, f_0 have been found, the corresponding network elements are

$$S = K + \rho, \quad R = R_0(S + \sqrt{S + S^2}),$$

$$C = \frac{\sqrt{1 + K/\rho}}{2\pi R\sqrt{2f_0^2 - \beta_1 - \gamma_1}}, \quad L = \frac{1}{4\pi^2 C f_0^2}$$

It is, of course, possible to use $(e^{2(x - \ldots)} - 1)$ instead of $\sinh(x - \ldots)$ in (19a).

(b) *Design of Reactances by the Summation Method.*

The design of many types of attenuation and phase shift equalisers can be reduced to the design of a purely reactive network producing a required reactance/frequency curve.

According to a general reactance theorem [8] the reactance X of any purely reactive network can be expressed in the form

$$|X| = \left| \rho f^{\pm 1} \frac{(f^2 - f_1^2)(f^2 - f_3^2) \ldots}{(f^2 - f_0^2)(f^2 - f_2^2) \ldots} \right| \quad \ldots \quad \ldots \quad (22)$$

The number of terms in numerator and denominator can never differ by more than one. ρ, f_0^2, f_1^2, ... are always *real* and *independent*; therefore

$$|Xf^{\pm 1}| = \left| \rho \frac{(f^2 - f_1^2) \ldots}{(f^2 - f_0^2) \ldots} \right| \quad \ldots \quad (23)$$

has the ideal form required by the summation method. If desired, f^2 can be replaced by $1/f^2$ or by $x = \log_e f$ as variable. Example: the reactance

$$X = -\rho \frac{f^2 - f_1^2}{f(f^2 - f_0^2)}, \quad \rho > 0,$$

can be represented either by

$$|Xf| = \rho \frac{|f^2 - f_1^2|}{|f^2 - f_0^2|},$$

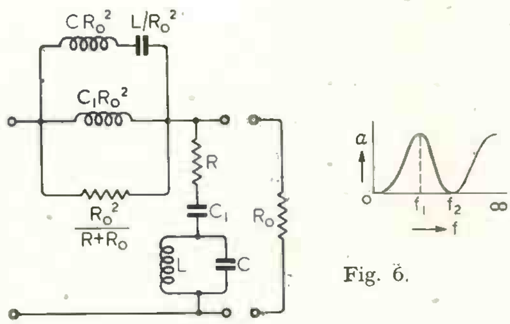

Fig. 6.

or by

$$|Xf| = \left(\frac{\rho f_1^{\,2}}{f_0^{\,2}}\right)\frac{|e^{2(x - \log_e f_1)} - 1|}{|e^{2(x - \log_e f_0)} - 1|}$$

or by

$$|X| = \left|\frac{2\rho f_1}{f_0^{\,2}}\right| \frac{|\sinh\,(x - \log_e f_1)|}{|e^{2(x - \log_e f_0)} - 1|}$$

Phase shift equalisers. Zobel's phase shift equalisers are lattice networks with reactances $\frac{1}{2}X$ in the series arms and reactances $-2R_0^{\,2}/X$ in the lattice arms. Their phase shift β is given by $\tan(\frac{1}{2}\beta) = \frac{1}{2}X/R_0$. Thus the design of a phase shift network is equivalent to the design of a reactance network.

Attenuation equalisers. For the attenuation equalisers shown in Fig. 6 (corresponding to Zobel's network 11) and Fig. 7 it has so far not yet been possible to find equations which at the same time are suitable for the application of the summation method and possess no interdependent parameters. (Zobel experiences a similar difficulty). If, however, the resistance R of their shunt arms is known, the remaining unknown elements of their shunt arms form purely reactive networks, the design of which is always possible by means of the summation method, as shown above.

For the general form of these networks (see Fig. 8)

$$\alpha = 20 \log_{10}\left|1 + \frac{R_0}{R + jX}\right|$$

$$= 10 \log_{10}\left(1 + \frac{R_0(R_0 + 2R)}{R^2 + X^2}\right) \quad \ldots \quad \ldots \quad (24)$$

so that $\alpha_{max} = 20 \log_{10}\left(1 + \dfrac{R_0}{R}\right)$ (if $X = 0$) $\quad \ldots \quad \ldots \quad (25)$

and

$$X^2 = R_0(R_0 + 2R)\left[\frac{1}{10^{\frac{\alpha}{10}} - 1} - \frac{R^2}{R_0(R_0 + 2R)}\right] \quad \ldots \quad \ldots \quad (26)$$

If the αf curve is required to have a maximum α_{max} at a frequency f_m *within* the required frequency range (i.e. not at its boundaries), X has to be made 0 at f_m, and R has to be chosen in accordance with (25), i.e. so that

$$\frac{R_0}{R} = -1 + 10^{1/20(\alpha_{max} \pm \epsilon\Delta)} \quad \ldots \quad (27)$$

where Δ is half the width of the tolerance band and $0 \leq \epsilon \leq 1$.

Usually, required αf curves for which one of the networks in Figs. 6 or 7 is chosen *have* such a maximum and also a narrow tolerance band, so that R can be found by means of (27). Then, by means of (26), a required Xf tolerance band can be derived from the required αf band originally given. After the reactance parameters ρ, f_0^2, f_1^2, f_2^2, ... have been determined by means of the summation method, the network elements can always be found. (For the networks in Figs. 6 and 7 see section 4e).

If R cannot be found in this way, it has to be determined by trial and error. It is necessary to estimate a number of R values and to design the reactance X for each of them. Then the best value of R, i.e. the value allowing the design of X within the narrowest tolerance band, can

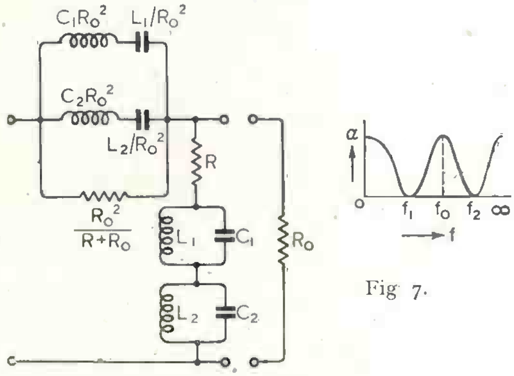

Fig 7.

be found. In order to simplify the trial-and-error procedure we can combine equations (22) and (26) as follows:

$$\left(\frac{1}{10^{\frac{\alpha}{10}} - 1} - d\right)$$

$$= \left[\rho' f^{\pm 1}\frac{(f^2 - f_1^{\,2})(f^2 - f_3^{\,2})\ldots}{(f^2 - f_0^{\,2})(f^2 - f_2^{\,2})\ldots}\right]^2 \quad \ldots \quad \ldots \quad (28)$$

where $\quad d = \dfrac{R^2}{R_0(R_0 + 2R)}$

and $\quad \rho' = \dfrac{\rho}{\sqrt{R_0(R_0 + 2R)}}$

The left-hand side of (28) has to be approximated by its right-hand side for different values of the resistance parameter d until the best d value has been found. The required band for

$$\frac{X^2}{R_0(R_0 + 2R)} = \frac{1}{10^{\frac{\alpha}{10}} - 1} - d$$

can be derived from the required band for $\frac{1}{10^{\frac{\alpha}{10}} - 1}$ simply by shifting the latter band along the ordinate axis (see Fig. 9).

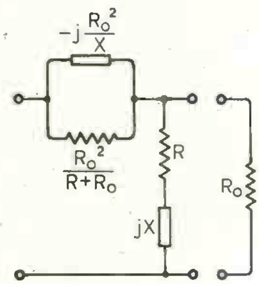

Fig. 8.

The networks in Figs. 4 and 5, discussed in the preceding section, also belong to the type shown in Fig. 8. If their resistance is known, the method described in this section can be applied for their design.

(c) *Application of the Summation Method to Filter Design.*

Cauer's method of filter design is based on the lattice type section with reactances $\frac{1}{2}X_{11}$ and $2X_{21}$ shown in Fig. 10. Though the best values of design parameters can be found analytically, and the results of the analysis have been tabulated, it may be of interest to show the application of the summation method to filter design.

The requirements for an ideal filter are:

pass band: $z_0 = \frac{1}{R_0}\sqrt{-X_{11}X_{21}} = 1$;

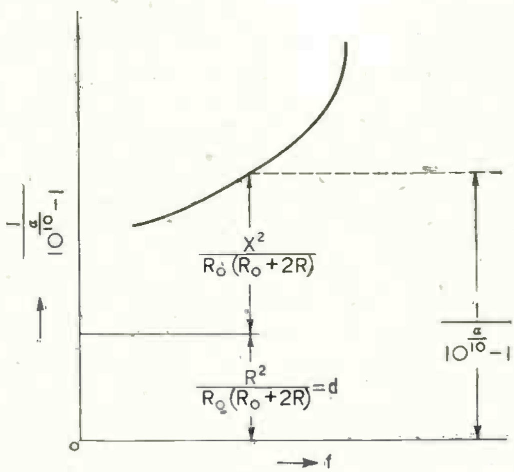

Fig. 9.

stop band: $y_0 = \sqrt{\frac{4X_{21}}{X_{11}}} = 1$.

Example: A low pass filter with cut-off frequency f_1 may have the reactances:

$$X_{11} = -\rho_1 \frac{(f_c^2 - f^2)(f_a^2 - f^2)(f_3^2 - f^2)}{f(f_b^2 - f^2)(f_2^2 - f^2)(f_4^2 - f^2)}$$

and

$$X_{21} = +\rho_2 \frac{f(f_b^2 - f^2)(f_1^2 - f^2)(f_3^2 - f_1^2)}{(f_c^2 - f^2)(f_a^2 - f^2)(f_2^2 - f^2)(f_4^2 - f^2)}$$

Then

$$y_0 = \sqrt{\frac{4\rho_2}{\rho_1}}\ \frac{jf(f_b^2 - f^2)\sqrt{f_1^2 - f^2}}{(f_c^2 - f^2)(f_a^2 - f^2)}$$

and

$$z_0 = \frac{1}{R_0}\sqrt{\rho_1\rho_2}\ \frac{(f_3^2 - f^2)\sqrt{f_1^2 - f^2}}{(f_2^2 - f^2)(f_4^2 - f^2)}$$

where ρ_1 and ρ_2 are positive constants and $f_c, f_b, f_a, f_1, f_2, f_3, f_4$ are resonant and anti-resonant frequencies the relative position of which is shown in Fig. 10a.

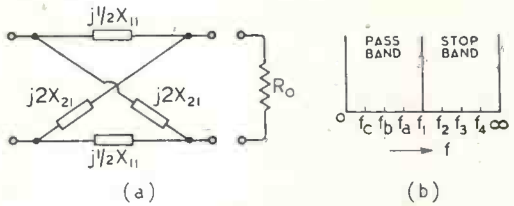

Fig. 10.

All parameters occurring are independent. Thus $\log |y_0|$ and $\log |z_0|$ can easily be represented by the summation method. The required curves are $\log |y_0| = \log 1 = 0$ in the stop band, and $\log |z_0| = \log 1 = 0$ in the pass band. The required tolerance band depends on the maximum loss permitted in the pass band, the minimum loss in the stop band, etc.

(d) *Design of Networks with Two Design Parameters by Means of One Single Standard Curve*

For networks with two design parameters the following design procedures are possible:

(1) Two standard curves are shifted along the x axis until their sum fits the required tolerance band.

(2) One single standard curve is shifted along both the x axis and the y axis until it fits the tolerance band.

(3) A straight line is drawn through the tolerance band.

1st Example: Attenuation Equaliser Network in Fig. 11 (corresponding to Zobel's networks 1a, 1b)

$$\alpha = 10 \log_{10} \frac{P + f^2}{Q + f^2} \qquad (29)$$

where $P = \left(\frac{R_0 + R}{2\pi L}\right)^2$, $Q = \left(\frac{R}{2\pi L}\right)^2$

$$\text{or } y = \frac{1}{10^{\frac{\alpha}{10}} - 1} = \frac{Q}{P - Q} + \frac{f^2}{P - Q} = d + gf^2 \qquad (30)$$

$$\text{with} \quad d = \frac{Q}{P - Q} = \frac{R^2}{R_0(R_0 + 2R)}$$

$$\text{and} \quad g = \frac{1}{P - Q} = \frac{(2\pi L)^2}{R_0(R_0 + 2R)}$$

(30) is the equation of a straight line as a function of f^2. When d and g have been found, R and L are given by

$$R = R_0(d + \sqrt{d + d^2}),$$

$$L = \frac{1}{2\pi} \sqrt{gR_0(R_0 + 2R)}$$

Some alternatives for (29) in which one standard curve occurs are:

(a) $y = d + e^{2[x + (\frac{1}{2} \log_e g)]}$, $x = \log_e f$, with e^{2x} as standard curve; (31)

(b) $\log y = \log g + \log |f^2 + d/g|$, with $\log |f^2|$ as standard curve, as a function of f^2; (32)

(c) $\log (y/f) = \log (2\sqrt{gd})$ $+ \log \cosh (x - \log_e \sqrt{\frac{d}{g}})$, $x = \log_e f$, with $\log \cosh x$ as standard curve. (33)

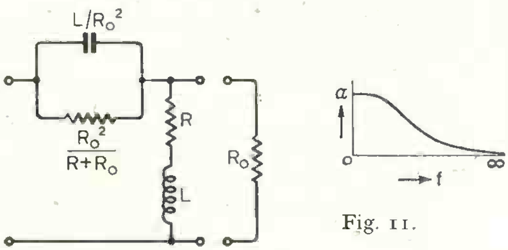

Fig. 11.

Finally, α can be plotted directly according to (29) by applying the summation method:

$$\alpha = \frac{1}{10} \log_{10} |f^2 + P| - \frac{1}{10} \log_{10} |f^2 + Q| \qquad (29a)$$

Though two standard curves $\frac{1}{10} \log_{10} f^2$ are necessary, this representation may be useful, if the level of the required curve, i.e. the level of α, is not quite definite, but it is required to find a compromise between level and width of tolerance band. Then it is very convenient to be able to carry out the design in the αf^2 plane, i.e. a plane in which one variable is α itself. The level becomes a third design parameter.

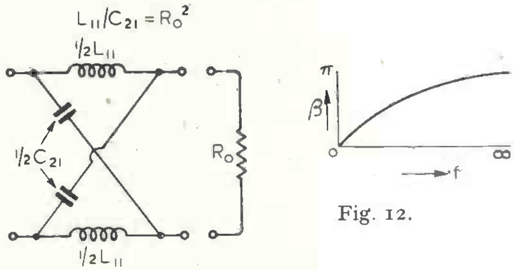

Fig. 12.

2nd Example: Phase Shift Network in Fig. 12 (corresponding to Zobel's network 13).

$$\tan \tfrac{1}{2}\beta = a_1 f \qquad (34)$$

where β is the phase shift and $a_1 = \frac{\pi L_{11}}{R_0}$.

(34) is the equation of a straight line through the origin and a_1 its slope. When a_1 has been found, $L_{11} = \frac{a_1 R_0}{\pi}$.

3rd Example: (a) Network in Fig. 13a which consists of two sections of the network in Fig. 12.

(b) Network in Fig. 13b (corresponding to Zobel's network 14).

$$\tan \tfrac{1}{2}\beta = \frac{a_1 f}{1 - b_2 f^2} \qquad (35)$$

$$\text{where} \quad a_1 = \frac{L_{11}\pi}{R_0} = \frac{(L_{11}' + L_{11}'')\pi}{R_0};$$

$$b_2 = 4\pi^2 L_{11} C_{12} = \frac{\pi^2 L_{11}' L_{11}''}{R_0^2}$$

From (35) follows

$$y = f \cot \tfrac{1}{2}\beta = \frac{1}{a_1} - \frac{b_2}{a_1} f^2 = d - gx \qquad (36)$$

$$\text{with } d = \frac{1}{a_1}, \quad g = \frac{b_2}{a_1}, \quad x = f^2 \qquad (36a)$$

(36) is the equation of a straight line and of the same form as equation (30) so that everything said about the 1st example (the network in Fig. 11) applies to this network

also: in particular the alternatives given there are also applicable here. When d and g have been found, the network elements are given by $a_1 = \frac{1}{d}$, $b_2 = \frac{g}{d}$ and

$$\text{Fig. 13b}: L_{11} = \frac{a_1 R_0}{\pi}, \quad C_{12} = \frac{b_2}{4\pi a_1 R_0} \quad (37)$$

Fig. 13a:

$$L_{11}', L_{11}'' = \frac{R_0}{2\pi}\left(a_1 \pm \sqrt{a_1^2 - 4b_2}\right) \quad (38)$$

From (36) and (37) it follows that any straight line in the yf^2 plane with positive intercept on the y axis and negative slope represents a physically possible network of the type in Fig. 13b. An additional limitation is imposed by (38) on straight

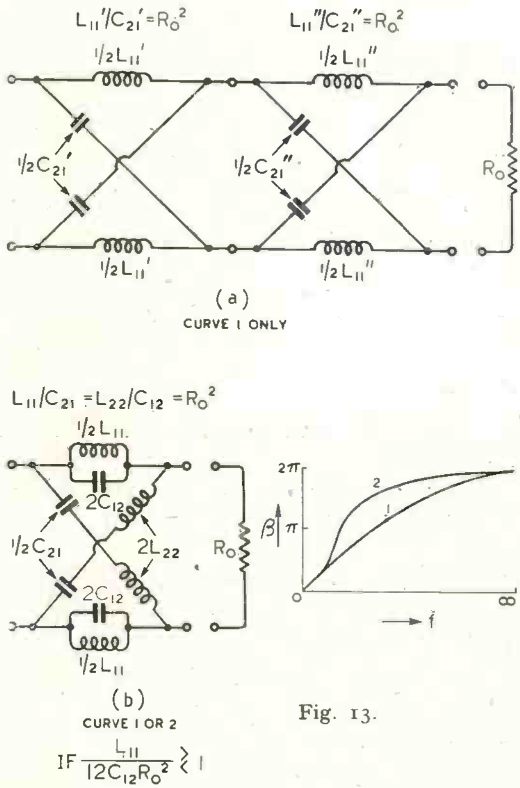

Fig. 13.

lines which represent a physically possible network of the two-section type in Fig. 13a. In order to obtain real expressions in (38) it is necessary that

$$a_1^2 - 4b_2 \geq 0 \quad \ldots \quad \ldots \quad \ldots \quad (39)$$

From (39) and (36a) it follows that

$$dg \leq 1/4 \quad \ldots \quad \ldots \quad \ldots \quad (40)$$

is the condition for the straight line

$$y = d - gx.$$

Condition (40) can be stated graphically: only those straight lines are allowed which,

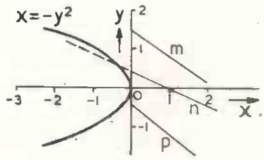

Fig. 14.—*Line m—network in Fig. 13 (b); line n—networks in Figs. 13 (a) or (b); line p—no physical network.*

in addition to having a negative slope and a positive y axis intercept, intersect the parabola $y^2 = -x$ (see Fig. 14), because the ordinate y_0 of the point of intersection of $y = d - gx$ and $y^2 = -x$ is

$$y_0 = \frac{1}{2g} \pm \sqrt{\frac{1}{4g^2} - \frac{d}{g}},$$

Thus y_0 is real only if $\frac{1}{4g^2} \geq \frac{d}{g}$ or if $dg \leq 1/4$. This shows that the condition for the existence of a point of intersection (y_0 real) is identical with the condition for a physically possible two-section network. Therefore only straight lines intersecting the parabola represent a solution for this network.

(*e*) *Shaping Method.*

The principle of the shaping method has been described in section 3 and in Fig. 3. A standard curve $y = f(x)$ representing the chosen type of network is transformed, in the most general case, into

$$y_1 = S + gx_1 + Tf[W(x_1 - U)].$$

S, g, T, W, and U have to be chosen so as to fit the required tolerance band. By visual inspection roughly suitable values for these parameters can be estimated. After adjusting the rotatable scales accordingly, and plotting y_1 as a function of x_1 in a rectangular co-ordinate system we can compare y_1 with the required band and, if necessary, the values of the parameters can be altered until a sufficiently good approximation is obtained.

1st Example: (Network in Fig. 11). As mentioned in section 4d, this network, though having only two parameters, offers a 3-parameter design problem, if the level α_0 of the required αf curve is taken as 3d parameter. From (29) follows

$$\alpha = S + 10 \log_{10}\left(1 - \frac{1}{W(x-U)}\right), \quad x = f^2$$

where $S = -\alpha_0$, $U = -Q$, $W = \dfrac{1}{Q-P}$

Thus the necessary standard curve is $y = 10 \log_{10}\left(1 - \frac{1}{x}\right)$. When S, W, U have been found ($S = \alpha_0$, as small as possible), P and Q are given by $Q = -U$, $P = Q - \frac{1}{W}$.

2nd Example: (Network in Fig. 4). From (10) it follows that

$$\frac{1}{10^{\frac{\alpha}{10}} - 1} = S + T \sinh^2(x-U)\,; \quad x = \log_e f;$$

with $S = \dfrac{R^2}{R_0(R_0+2R)}$, $T = \dfrac{4L}{CR_0(R_0+2R)}$, $U = \log_e f_0$, $f_0^2 = \dfrac{1}{4\pi^2 LC}$. The necessary standard curve is $\sinh^2 x$. When S, T, U have been found, L, C, R are given by

$$R = R_0(S + \sqrt{S + S^2}),$$

$$L = \frac{1}{4\pi f_0}\sqrt{TR_0(R_0+2R)},$$

$$C = \frac{1}{4\pi^2 L f_0^2}.$$

3rd Example: (Network in Fig. 5). From (18) it follows that

$$\left.\begin{aligned} y &= \frac{1}{10^{\frac{\alpha}{10}} - 1} = S + \frac{T}{\sinh^2(x-U)} \\ x &= \log_e f \end{aligned}\right\} \quad (41)$$

with

$$S = \frac{R^2}{R_0(R_0+2R)}, \quad T = \frac{L}{4CR_0(R_0+2R)},$$

$$U = \log_e f_0, \qquad f_0^2 = \frac{1}{4\pi^2 LC}.$$

The necessary standard curve is $y = \dfrac{1}{\sinh^2 x}$.

When S, T, U have been found,

$$R = R_0(S + \sqrt{S + S^2}),$$

$$L = \frac{1}{f_0 \pi}\sqrt{TR_0(R_0+2R)},$$

$$C = \frac{1}{4\pi^2 L f_0^2}.$$

4th Example: (Network in Fig. 6). The reactance of the shunt arm is

$$X = -\frac{C+C_1}{2\pi CC_1}\,\frac{f^2 - f_1^2}{f(f^2 - f_2^2)}, \quad f_1^2 = \frac{1}{4\pi^2 L(C+C_1)}, \quad f_2^2 = \frac{1}{4\pi^2 LC} \quad (42)$$

(a) R is known (see section 4b). From (42) and (26) it can be shown that, with

$$y = \sqrt{f^2\left[\frac{1}{10^{\frac{\alpha}{10}} - 1} - \frac{R^2}{R_0(R_0+2R)}\right]},$$

$$y = S + \frac{T}{x-U},$$

where $x = f^2$, $S = \dfrac{C + C_1}{2\pi CC_1\sqrt{R_0(R_0+2R)}}$, $T = S(f_2^2 - f_1^2)$, $U = f_2^2$. Therefore the necessary standard curve is $y = \frac{1}{x}$. When S, T, U have been found,

$$f_2^2 = U, \quad f_1^2 = U - \frac{T}{S},$$

$$C_1 = \frac{f_2^2}{2\pi f_1^2 S\sqrt{R_0(R_0+2R)}},$$

$$C = \frac{C_1 f_1^2}{f_2^2 - f_1^2}, \qquad L = \frac{1}{4\pi^2 C f_2^2}.$$

(b) R is unknown. From (42) and (26) it can be shown that

$$y = f^2\frac{1}{10^{\frac{\alpha}{10}} - 1} = gx + T_1\left(1 + \frac{1}{W(x-U_1)}\right)^2,$$

$x = f^2$, with $g = \dfrac{R^2}{R_0(R_0+2R)}$, $T_1 = S^2$, $U_1 = U$, $W = \dfrac{\sqrt{T_1}}{T} = \dfrac{1}{f_2^2 - f_1^2}$; S, U see case (a). Thus the necessary standard curve is

$$y = \left(1 + \frac{1}{x}\right)^2.$$

When g, T, W, U_1 have been found,

$$R = R_0(g + \sqrt{g + g^2}), \quad U = U_1,$$

$$S = \sqrt{T_1}, \qquad T = \frac{S}{W}.$$

C_1, C, L see case (a).

5th Example: (Network in Fig. 7). Its reactance is

$$X = -\frac{C_1 + C_2}{2\pi C_1 C_2}\,\frac{f(f^2 - f_0^2)}{(f^2 - f_1^2)(f^2 - f_2^2)} \quad (43)$$

with $f_1^2 = \dfrac{1}{4\pi^2 L_1 C_1}$, $f_2^2 = \dfrac{1}{4\pi^2 L_2 C_2}$,

$$f_0^2 = \frac{L_1 + L_2}{4\pi^2 L_1 L_2 (C_1 + C_2)}.$$

R has to be determined either from a required α_{max} or by trial and error (see section

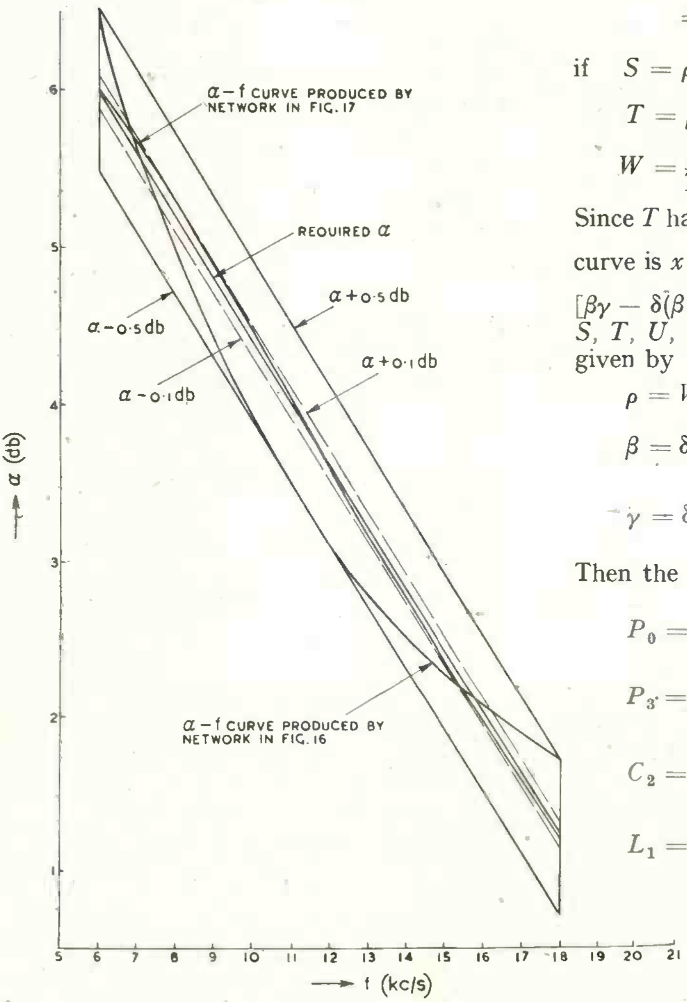

Fig. 15.

4b). L_1, L_2, C_1, C_2 are determined by the shaping method as follows.

From (43) and (26) it follows that

$$\sqrt{\frac{f^2}{\dfrac{1}{10^{\frac{\alpha}{10}} - 1} - \dfrac{R^2}{R_0(R_0 + 2R)}}} = \rho\,\frac{(x-\beta)(x-\gamma)}{x-\delta}$$

with $x = f^2$, $\rho = \dfrac{2\pi C_1 C_2 \sqrt{R_0(R_0 + 2R)}}{(C_1 + C_2)}$,

$$\beta = f_1^2, \quad \gamma = f_2^2, \quad \delta = f_0^2.$$

Now

$$\rho\,\frac{(x-\beta)(x-\gamma)}{x-\delta} = S + T\left[W(x-U) \pm \frac{1}{W(x-U)}\right]$$

if $S = \rho(2\delta - \beta - \gamma)$,

$$T = \rho\sqrt{\pm[\beta\gamma - \delta(\beta+\gamma) + \delta^2]},$$

$$W = \frac{\rho}{T}, \qquad U = \delta.$$

Since T has to be real, the necessary standard curve is $x + \frac{1}{x}$ or $x - \frac{1}{x}$ depending on whether $[\beta\gamma - \delta(\beta+\gamma) + \delta^2] > 0$ or < 0. When S, T, U, W have been found, ρ, β, γ, δ are given by

$$\rho = WT, \qquad \delta = U,$$

$$\beta = \delta - \frac{S}{2\rho} + \frac{1}{\rho}\sqrt{\tfrac{1}{4}S^2 \mp T^2},$$

$$\gamma = \delta - \frac{S}{2\rho} - \frac{1}{\rho}\sqrt{\tfrac{1}{4}S^2 \mp T^2}.$$

Then the network elements are given by

$$P_0 = \frac{1}{4\pi^2\delta}, \quad P_1 = \frac{1}{4\pi^2\beta}, \quad P_2 = \frac{1}{4\pi^2\gamma},$$

$$P_3 = \frac{\rho}{2\pi\sqrt{R_0(R_0 + 2R)}},$$

$$C_2 = \frac{P_3 P_0 (P_1 - P_2)}{P_1 (P_0 - P_2)}, \quad C_1 = \frac{P_3 C_2}{C_2 - P_3},$$

$$L_1 = \frac{P_1}{C_1}, \qquad L_2 = \frac{P_2}{C_2}.$$

5. Design Example.

The design of two attenuation equaliser networks is described in Figs. 15, 16 and 17. The required curve for both networks is shown in Fig. 15. Fig. 16 shows the design of a two parameter network of the type shown in Fig. 11 for a tolerance band with a width of $\pm$ 0.5 db, according to equation (30) (straight line method). In many cases this tolerance band width is too large. Fig. 17 shows the design of a 3-parameter network of the type shown in Fig. 5 by the shaping method according to equation (41) for a tolerance band width of $\pm$ 0.1 db (it is possible but of little practical interest to reduce this width to $\pm$ 0.07 db by using a much larger scale for the graphs in the yx plane). The curves actually produced by the two designed networks are shown in Fig. 15.

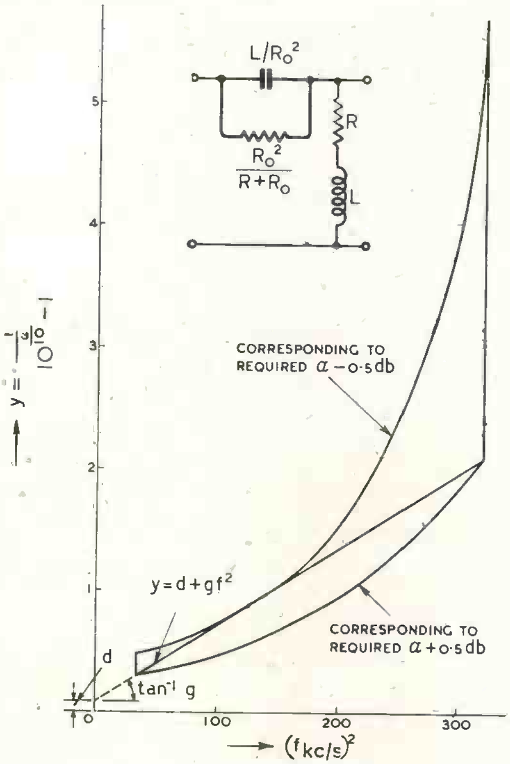

Fig. 16.—*Design of a two-parameter network.* $d = 0.06$, $g = \frac{1}{160} \times 10^{-6}$, $R/R_0 = 0.312$, $L/R_0 = 0.016 \times 10^{-3}\, H/\Omega$.

6. Practical Elimination of Trial and Error

If a two-parameter network is designed by drawing a straight line through the required or estimated tolerance band, or by shifting one single standard curve until it fits the tolerance band, it is not necessary to make a number of separate trials, but the fitting can be done as one single action, under visual observation.

At least from a practical point of view, this is no longer trial and error, but incomparably more convenient. A comparison between the summation method and direct visual fitting can easily be made by designing the network in Fig. 11 according to either equation (29a), or equation (30). It is therefore desirable to extend this method without trial and error to networks with more than two parameters. This can be done by means of some auxiliary equipment, and in particular the shaping method can be used for this purpose.

As described in detail in section 3, in the shaping method a standard curve has to be shifted and deformed until it fits the tolerance band. Shifting can be done without any apparatus and the result of shifting can be seen immediately. However, deformation, viz. stretching and shearing (see Fig. 3) has to be done in two steps: (a) adjusting of rotatable scales and reading the curve point co-ordinates off these scales, (b) replotting these values in a rectangular co-ordinate system. Therefore continuous variation of the stretching and shearing parameters under continuous visual observation—which is essential for practical elimination of trial and error—is not possible if carried out in the manner just described.

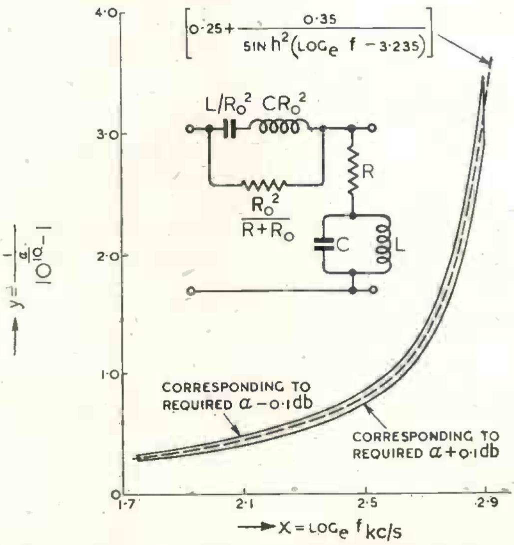

Fig. 17.—*Design of a three-parameter network.* $L/R_0 = 11.98 \times 10^{-6}\, H/\Omega$, $CR_0 = 3.274 \mu F \Omega$, $R/R_0 = 0.809$.

It would become possible if the standard curve were physically stretched and sheared instead of only the scales being varied.

For this purpose various kinds of mechanical, optical, or electrical equipment can be used. Only two examples may be mentioned here.

If two shift and one stretch parameters have to be determined (see e.g. the network in Fig. 5 and equation (41)), a standard curve cut out of hard paper can be used. The required band is plotted on a horizontal mirror surface and the standard curve shifted and *tilted* above the mirror until its projection under vertical observation fits the tolerance band.

Stretching and shifting in two directions can easily be carried out on the screen of a cathode-ray oscillograph. For this purpose the standard curve has to be produced as the trace of a cathode ray beam by means of deviation potentials varying in accordance with the respective standard curve. Then stretching can be carried out by increasing the amplification of the variable deviation potentials, and shifting by adding constant potentials. The cathode ray trace on the screen has to be shifted and stretched until it fits the tolerance band.

In this way it has been found possible to determine up to four design parameters in one single visual fitting operation without trial and error.

Conclusion

After a general discussion of equaliser design problems a semi-graphical design method is developed which either proceeds without trial and error or uses only a particularly simple form of it.

The attenuation/frequency or phase-shift/frequency curves of a great number of correcting four-terminal networks can be transformed in such a manner that they can be represented as the sum of a number of shiftable standard curves or as one single shiftable and deformable (stretchable and shearable) standard curve; the position and deformation of the curves correspond to the values of the network elements. Thus, if the tolerance band of a required attenuation/frequency curve is transformed in this manner, the trial and error necessary to find the best element values for a chosen type of network consists only in determining the best position and/or deformation of some standard curves and is very simple. If only two design parameters have to be determined, trial and error seems to disappear completely; the best position of one single standard curve determines both parameters and can be found by direct visual curve fitting. This visual fitting procedure can be extended to more than two parameters by using some auxiliary mechanical, optical, or electrical equipment.

I wish to thank the Telephone Manufacturing Company, in whose laboratories this investigation has been carried out, for the permission to publish it.

APPENDIX

Example of an Analytical Solution of the Approximation problem.

For the two-parameter network in Fig. 11 the attenuation in nepers is $A = \frac{1}{2}\log_e \frac{P + f^2}{Q + f^2}$ (see equation (29)). Let the required curve be $A = \frac{1}{2}g(f)$ between $f = f_1$, and $f = f_2$, and let us assume that $g(f)$ is of such a shape that maximum deviations occur at 3 frequencies only: somewhere in the interior of the range $f_1 \ldots f_2$, say at f_0, and at the boundaries f_1 and f_2 (this is true, e.g., if $g(f)$ is a straight line, see Fig. 15). Then the best approximation is defined by the following two equations for the deviation

$$\Delta(f) = \tfrac{1}{2}\log_e \frac{P + f^2}{Q + f^2} - \tfrac{1}{2}g(f):$$

$$\Delta(f_1) = \Delta(f_2) \qquad \ldots \quad (44a)$$

$$\Delta(f_1) = -\Delta(f_0) \qquad \ldots \quad (44b)$$

where f_0 is defined by

$$\left(\frac{d(\Delta(f))}{df}\right)_{f=f_0} = 0 \qquad \ldots \quad (45)$$

These are three equations for three unknown magnitudes, P, Q, and f_0. From equation (44a) we obtain

$$\tfrac{1}{2}\left[\log_e \frac{P + f_1^2}{Q + f_1^2} - g(f_1)\right] = \tfrac{1}{2}\left[\log_e \frac{P + f_2^2}{Q + f_2^2} - g(f_2)\right]$$

or $$\frac{(P + f_1^2)(Q + f_2^2)}{(P + f_2^2)(Q + f_1^2)} = e^{[g(f_1) - g(f_2)]} = M$$

or $$P = \frac{Qa - b}{Q - c} \qquad \ldots \quad (46)$$

where $a = \dfrac{Mf_2^2 - f_1^2}{1 - M}$, $\quad b = f_1^2 f_2^2$,

$$c = \frac{Mf_1^2 - f_2^2}{1 - M},\ M = e^{g(f_1) - g(f_2)}$$

From equation (45) we obtain

$$\tfrac{1}{2}\frac{d}{df}\left[\log_e \frac{P + f^2}{Q + f^2} - g(f)\right]_{f=f_0} = 0$$

or $$\frac{2f_0(P - Q)}{(P + f_0^2)(Q + f_0^2)} + g'(f_0) = 0 \qquad \ldots \quad (47)$$

Substituting (46) in (47) we obtain

$$\frac{-Q^2 + Q(a + c) - b}{(Q + f_0^2)[Q(a + f_0^2) - (b + cf_0^2)]} = G$$

where $G = \dfrac{g'(f_0)}{2f_0}$.

This is a quadratic equation for Q in terms of a, b, c, f_0, and G, the solution of which is

$$Q = \tfrac{1}{2}\frac{(a+c) - G[-b + f_0^2(a-c) + f_0^4]}{1 + G(a+f_0^2)} \pm \sqrt{\tfrac{1}{4}\left[\frac{a+c-G[-b+f_0^2(a-c)+f_0^4]}{1+G(a+f_0^2)}\right]^2 + \frac{Gf_0^2(b+cf_0^2)-b}{1+G(a+f_0^2)}} \quad \ldots \quad (48)$$

From equation (44b) we obtain

$$\frac{(P+f_1^2)\,(P+f_0^2)}{(Q+f_1^2)\,(Q+f_0^2)}\,e^{-[g(f_1)+g(f_0)]} = 1 \quad \ldots \quad (49)$$

or, using equation (46)

$$\frac{[Q(a+f_1^2)-(b+cf_1^2)]\quad[Q(a+f_0^2)-(b+cf_0^2)]}{(Q-c)^2\quad(Q+f_1^2)\quad(Q+f_0^2)}\,e^{-[g(f_1)+g(f_0)]} = 1 \quad \ldots \quad (50)$$

Equation (48) allows Q to be plotted as a function of f_0. Using this graph, the left-hand side of equation (50) can be plotted as a function of f_0. The value of f_0 for which this graph is equal to 1 is the required value. Then equation (48) gives the corresponding Q value and equation (46) the corresponding P value. Instead of a purely graphical procedure any of the well-known numerical methods for the solution of equations can be used.

Though we may assume that the solution of these equations can be simplified by previous knowledge of the approximate values f_0, P, and Q, obtained by the "straight line method," described in section 4d, the amount of work necessary is considerable—when compared with the above method—and scarcely worth while as the result is not more accurate than that obtained by the straight line method.

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Correspondence

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

"Optimum Conditions in Class A Amplifiers"

To the Editor, "Wireless Engineer"

SIR,—Professor Howe's Editorial (*Wireless Engineer*, February 1943, p. 53) on Nottingham's paper should prove useful in removing the confusion so often found in the minds of radio engineers over the optimum load for Class A amplification. Some time ago I examined Nottingham's paper (*Proc. I.R.E.* December 1941, p. 620) with a view to incorporating its results in lectures to students, and one or two points arose, which may be of interest to those similarly engaged.

Howe's formula A (Nottingham's 8) can be rearranged as follows:

$$R = \frac{V_{ao} - r_a(2I_o - I_{min}) - \epsilon}{I_o - I_{min}} \quad \ldots \quad (1a)$$

$$= \frac{V_{ao}}{I_o}\left(\frac{I_o}{I_o - I_{min}}\right) - r_a\left(2 + \frac{I_{min}}{I_o - I_{min}}\right) - \frac{\epsilon\, I_{min}}{I_{min}(I_o - I_{min})}$$

Let $R_{DC} = \dfrac{V_{ao}}{I_o}$, the D.C. resistance of the valve

and $r_{min} = \dfrac{\epsilon}{I_{min}}$

then

$$R = R_{DC} - 2r_a + \frac{I_{min}}{I_o - I_{min}}(R_{DC} - r_a - r_{min}) \quad \ldots \quad (1b)$$

$$= R_{DC} - 2r_a + \frac{I_{min}}{I_o - I_{min}}(R_{DC} - r_a) \quad \ldots \quad (1c)$$

when ϵ is zero.

Expression 1c is applicable to the case of a pentode or tetrode valve, and a good approximation to the maker's optimum load is obtained by taking r_a as the slope resistance of the tetrode characteristic curve for $E_g = 0$ at low anode voltages, i.e., before the knee of the curve is reached. R_{DC} is the D.C. resistance of the valve obtained from the recommended D.C. operating conditions and I_{min} can usually be taken as 10 per cent. of I_o.

Nottingham's method of finding the optimum conditions for the third case of fixed anode dissipation and variable D.C. anode voltage proved unduly complicated for teaching purposes and the following analysis was found more suitable.

A.C. power output,

$$P_o = \tfrac{1}{2}(V_{ao} - r_a(2I_o - I_{min}) - \epsilon)(I_o - I_{min}),$$

for maximum power

$$\frac{dP_o}{dI_o} = 0 = (3\,I_{min} - 4\,I_o)r_a - \epsilon - \frac{dV_{ao}}{dI_o}I_{min}.$$

Note that $V_{ao}I_o$ = constant = P_{am}

and $$\frac{dV_{ao}}{dI_o} = -\frac{P_{am}}{I_o^2}$$

$$\therefore \quad \frac{P_{am}}{I_o^2} I_{min} = (4I_o - 3I_{min})r_a + \epsilon \quad \ldots \quad (2)$$

which is identical to Nottingham's equation 17.

Now $\frac{P_{am}}{I_o^2} = R_{DC}$, and by substitution, expression 2 becomes

$$R_{DC} = \left[\frac{4I_o}{I_{min}} - 3\right]r_a + r_{min} \quad \ldots \quad \ldots \quad (3)$$

Replacing R_{DC} in 1b by 3

$$R = \left(\frac{4I_o}{I_{min}} - 5\right)r_a + r_{min}$$
$$+ \frac{I_{min}}{I_Q - I_{min}}\left[\left(\frac{4I_o}{I_{min}} - 3\right)r_a + r_{min} - r_a - r_{min}\right]$$
$$= \left[\frac{4I_o}{I_{min}} - 1\right]r_a + r_{min} \quad \ldots \quad \ldots \quad (4)$$

This form for the optimum load is preferable to Nottingham's expression 16, since it clearly shows the dependence on I_{min} and $\epsilon(r_{min})$.

A point to be stressed is that this third condition of maximum power output found by Nottingham requires a much larger input grid voltage. In the example quoted by him a grid peak input voltage of 104.5 is required for $R = 2r_a$ compared with 334 volts for $R = 24.4\ r_a$, the optimum value, i.e., an increase of 3.2 to 1 compared with a power output increase of 1.77 to 1. Thus increased output efficiency can only be obtained at the cost of a very much increased grid voltage input.

K. R. STURLEY.

To the Editor, "Wireless Engineer"

SIR,—Your Editorial of February, 1943, seems to be a demonstration that the power efficiency of a triode is not related to the conditions for maximum output for a given valve and battery voltage. The results are interpreted for ratios of R to r_a which only hold for the specific example. It would seem, therefore, that a method of approach not involving direct consideration of this ratio is to be preferred, since it could be applied to any form of generator having limiting conditions or characteristics (for example, a pentode valve, where it would be pointless to argue in terms of the ratio of the load resistance to the internal impedance).

The method I propose, is to relate the performance to the magnitude of the negative anode swing with respect to the positive supply busbar. If the anode battery voltage be B, then let this negative swing be $p \cdot B$, where p obviously varies between 0 and 1. Then for equal positive and negative inputs and with choke or transformer coupling of the load resistance R, the total anode swing is $2p \cdot B$, if the amplifier is linear. Let the maximum anode current be I_m, the minimum $q \cdot I_m$ (where q is a small fraction), and the anode D.C. dissipation D; then

$$D = \frac{B \cdot I_m(1 + q)}{2}; \qquad R = \frac{2p \cdot B}{I_m(1 - q)};$$

$$\text{A.C. watts} = w_a = \frac{(2p \cdot B)^2}{8R};$$

for sinusoidal operation.

Therefore the sinusoidal power efficiency

$$= \frac{\text{A.C. watts}}{\text{D.C. watts}} = \frac{p}{2} \cdot \frac{(1 - q)}{(1 + q)} \quad \ldots \quad \ldots \quad (1)$$

Also if, for the sake of example, I_m be limited only by the boundary curve $I = \frac{V_a}{r_a}$ (i.e., the idealised zero grid volts, anode characteristic for a triode), then we can obtain the following two alternative forms for B,

$$B^2 = \frac{2D \cdot r_a}{(1 - p)(1 + q)} \quad \ldots \text{(a), or}$$

$$B^2 = \frac{4 \cdot w_a \cdot r_a}{p(1 - p)(1 - q)} \quad \ldots \text{(b)} \quad \ldots \quad \ldots \quad (2)$$

Thus from (1) we see that any valve under any conditions, gives greater power efficiency as p is increased towards its limiting value of unity, provided q is small or constant, which proves the contention of your Editorial without specific reference to R or r_a, the valve type or the limiting conditions of operation.

From (2a) we can find p maximum for a given D, r_a, B and q for a triode, and we note that there is an irreducible minimum to B, when p is 0, of $\frac{2D \cdot r_a}{(1 + q)}$. If this value of B be called B_0, then a curve $\frac{B}{B_0} = \sqrt{\frac{1}{(1 - p)}}$ could be plotted for all such triode valves, and by calculating B_0, and taking the ratio of the given maximum B, it could be ascertained whether the general run of valves gave p maximum values around 0.8 (for the case when the maximum D.C. dissipation and the maximum supply voltage are used). If so, then there is not much likelihood of any future improvement in the power efficiencies of these valves (for from (1) the maximum sinusoidal efficiency is 50 per cent. when p is unity), and a value of $p = 0.8$ can be generally used, with a corresponding power efficiency.

From (2b) we see that the maximum power output, for a given B, r_a, and q is always obtained for a triode when $p = 0.5$ (assuming linear characteristics, and a zero grid volts boundary).

The same methods can be applied to a pentode or tetrode, when I_m is limited by a constant-current boundary for p values up to about 0.8 of the screen voltage (i.e. up to the "knee" value).

The method can be expanded easily to cover the more general and practical cases where a power-law valve characteristic is assumed for the boundary curve of I_m, and it is hoped to publish this treatment shortly.

Use of the above method avoids possible confusion between the differential impedance of the valve, and the conditions determining a maximum of output. For the differential impedance may be modified at will by appropriate types of feedback, but the boundary conditions which determine the maximum output remain a property of the valve.

Harrow Weald, Middlesex. B. M. HADFIELD.

Simple Quartz-crystal Filters of Variable Bandwidth*

By Geoffrey Builder, Ph.D., and J. E. Benson, B.Sc., B.E.

ABSTRACT.—To obtain, in general purpose communication receivers, a variation in the effective received bandwidth from a few hundred cycles to several kilocycles, it is now common to use a single quartz crystal in the bridge-balanced circuit described by Robinson† (1931). This paper discusses the calculation of the maximum attainable bandwidth and methods of variation of the effective bandwidth in terms of conventional filter theory. It is shown that Robinson's circuit in its balanced condition is indistinguishable from a confluent band-pass filter, but that the latter is not directly realisable physically using a quartz crystal. On the other hand, simple m-derived sections are so realisable and these, together with the bridged-T section described by Mason, provide useful alternative detailed designs.

1. Introduction

GENERAL-PURPOSE communication receivers are now usually designed to provide a variation of selectivity from a bandwidth (measured at 6 db. down) of a few hundred cycles, for the reception of C.W. telegraphy under adverse conditions, to one of several kilocycles, for the reception of telephony. Simple quartz-crystal filter circuits operating at the intermediate frequency are suitable for this purpose and are now widely used. The designer is generally limited to the use of filter-circuit components of a quality and type common in mass-produced receivers, and the quartz-crystal resonator must be suitably designed to provide the required performance with these components.

The filter circuits in general use all seem to have been developed from the selector or "crystal gate" described by J. Robinson† (1931) for the highly-selective tone-corrected type of receiver exemplified by the Stenode Radiostat. Robinson's arrangement is characterised by the use of a bridge circuit, by which the effective parallel capacitance of the crystal resonator is balanced out, to obtain a symmetrical and highly selective resonance curve. A bridged-T single-crystal filter described by Mason (1934) does not seem to have been used in communication receivers, although the arrangement is simple and economical in components. Nor does there appear to have been any use of conventional ladder-type filter structures for the purpose. Nevertheless, m-derived and bridged-T circuits offer some interesting alternatives to Robinson's arrangement, and they have some characteristics that may be advantageous to the designer. We have, therefore, in the following pages, summarised briefly the salient characteristics of these circuits and have discussed their practical application in communication receivers.

2. Circuits

The circuit of Fig. 1 (a) represents, with adequate accuracy for the present discussion, the electrical characteristics of a quartz-crystal resonator mounted between electrodes with a very small or zero air gap; such a mounting is necessary to obtain suitable resonator characteristics. The circuit comprises a series combination L_1C_1 in parallel with a capacitance C_0; the constants L_1C_1 are determined primarily by the design and grinding of the crystal element itself, but C_0 is made up of the effective parallel capacitance due to the crystal itself together with stray capacitances between the electrodes and their electrical connections, and is therefore dependent on the method of mounting the crystal and of connecting it in any specific arrangement. Neglecting dissipation, the frequency-reactance curve of the circuit takes the form shown in Fig. 1 (b), in which the angular resonance fre-

* Reprinted from *A.W.A. Technical Review*, 1941, Vol. 5, No. 3, p. 93.

† Since this was written we have discovered that the bridge-balanced circuit ascribed to Robinson was invented by W. A. Marrison of the Bell Telephone Laboratories. Patent 1,994,658, filed June 7th, 1927, granted March 19th, 1935.

quency ω_r and anti-resonance frequency ω_a are given by :—

$$\omega_r = (L_1C_1)^{-\frac{1}{2}}; \; 2(\omega_a - \omega_r)/\omega_r \doteqdot C_1/C_o \quad (1)$$

If an inductance L, in parallel with a capacitance C, is connected across the crystal, the equivalent circuit is that shown in Fig. 1 (c) and, again neglecting dissipation, has the reactance-frequency characteristic of

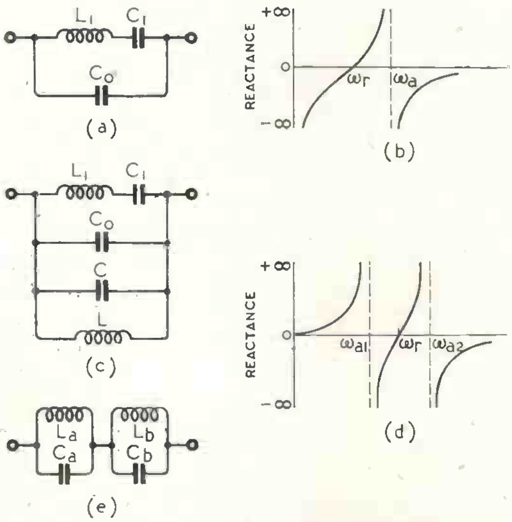

Fig. 1.—*(a) Equivalent electrical circuit of quartz-crystal resonator. (b) Reactance-frequency curve for (a), neglecting dissipation. (c) Equivalent circuit of crystal with inductance L and capacitance C connected across it. (d) Reactance-frequency curve for (c), neglecting dissipation. (e) Circuit equivalent to (c).*

Fig. 1 (d). If $L(C + C_0) = L_1C_1$, the angular frequencies ω_r at resonance, and ω_{a1}, ω_{a2} at anti-resonance, are given by :—

$$\omega_r = (L_1C_1)^{-\frac{1}{2}}; \quad \omega_{a1}\omega_{a2} = \omega_r^2;$$
$$\{(\omega_{a2} - \omega_{a1})/\omega_r\}^2 = C_1/(C + C_o) \quad \ldots \quad (2)$$

The equivalent electrical constants L_1, C_1, C_o may be determined by measuring ω_r, ω_a and using equations (1), or by measuring ω_r, ω_{a1}, ω_{a2} and using equations (2) (Builder, 1940).

The configuration of Fig. 1 (c) is equivalent to that of two parallel tuned circuits L_aC_a, L_bC_b connected in series as in Fig. 1 (e), the constants being related by :—

$$L = L_a + L_b; \quad (C_o + C) = \frac{C_aC_b}{C_a + C_b};$$
$$\omega_{a1} = (L_aC_a)^{-\frac{1}{2}}; \quad \omega_{a2} = (L_bC_b)^{-\frac{1}{2}}. \quad \ldots \quad (3)$$

The bridge circuit of Robinson may take the form shown in Fig. 2 (a). The input is applied to the primary of a transformer of which the secondary is tuned to the mid-frequency of the pass-band. The two ends of the secondary are connected, one through the crystal, and the other through a balancing condenser C_B, to one output terminal (D). The other output terminal (B), usually earthed, is a tap on the tuned secondary of the input transformer. The balancing condenser C_B may be adjusted so that current through it to the load circuit balances out that through the effective parallel capacitance C_o of the crystal. If, as is usual, the tap is at the mid-point of the transformer secondary, this requires $C_B = C_o$. When this balance is effected, the electrical characteristics are indistinguishable from those of the circuit of Fig. 2 (b) (Colebrooke 1932), provided that L_1 and C_1 are the equivalent constants of the crystal resonator and that

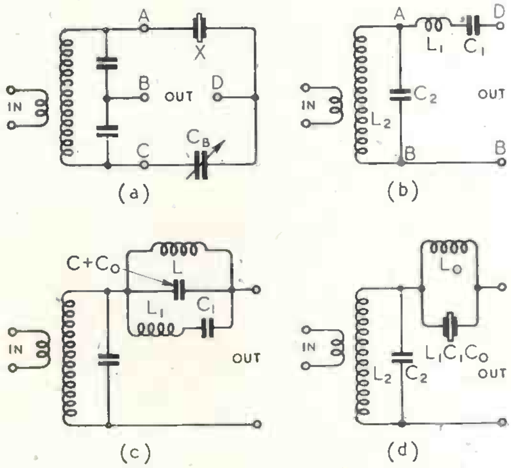

Fig. 2.—*(a) An arrangement of Robinson's bridge-balanced selector circuit. (b) Confluent half-section equivalent to (a). (c) Realisable m-derived half-section. (d) Realisable m-derived half-section that may be regarded as an arrangement in which the effective parallel capacitance C_0 of the crystal is neutralised by an inductance L_0.*

the parallel arm, L_2C_2, has the same impedance as the tuned secondary of the input transformer of Fig. 2 (a), measured between the terminals AB. The circuit of Fig. 2 (b) may be treated as a half-section of a confluent band-pass ladder filter after the manner of Zobel (1923, 1931), and characteristics so calculated may be applied to the circuit of Fig. 2 (a) in the balanced condition.

Although the confluent half-section is not directly realisable physically, using a quartz crystal in the series arm, the corresponding m-derived half-section (Zobel, ibid, or Everitt, 1937) shown in Fig. 2 (c) is so realisable, the series arm consisting, as in the configuration of Fig. 1 (c), of a crystal in parallel with a capacitance C and inductance L. The m-derived section is characterised by two rejector frequencies (i.e. frequencies of infinite attenuation in the dissipationless case) which can be located conveniently with respect to the pass-band by appropriate choice of constants. If $L(C + C_o) = L_1C_1$, the rejector frequencies are given by ω_{a1} and ω_{a2} of equation (2) and are disposed with geometric symmetry about the centre of the pass-band. The spacing between them is determined by the ratio $C_1/(C + C_o)$, as shown in equation (2), and is a maximum when $C = 0$, i.e. when the inductance L becomes equal to L_o, where $L_oC_o = L_1C_1$. When C is very small, or zero, the rejector frequencies are well removed from the pass-band, in the vicinity of which the response is closely similar to that of the confluent section of Fig. 2 (b). When C is zero, the m-derived section has the form shown in Fig. 2 (d) and the inductance L_o may be regarded as neutralising the capacitance C_o, in the same way (Doherty, 1939) that inductance neutralisation of valve inter-electrode capacitances is now commonly effected in high-frequency power amplifiers.

More generally, the m-derived section can be used to advantage to produce high attenuation close to, and on both sides of, the pass-band by suitable choice of the rejector frequencies. In this it has an advantage over the bridge circuit used in its balanced condition, and it also offers an alternative constructional arrangement. A full m-derived π-section may also be used to advantage in some designs and may readily be converted to Mason's bridged-T single-crystal section, but the latter is more economical in components (vide Builder, 1938).

3. Design Considerations

In many communication receivers the crystal selector circuit is arranged to give facilities summarised as follows :—

(a) A maximum received bandwidth (at 6 db. down) equal to or slightly less than that of the receiver with the crystal circuit switched out ; this maximum width is usually about 6 kc/s.

(b) Variation of the bandwidth, by means of a control immediately accessible to the operator, from the maximum to a minimum of a few hundred cycles.

(c) Tuning, by means of a control immediately accessible to the operator, of the position with respect to the pass-band of at least one rejector frequency, to obtain maximum discrimination against any interfering signal.

Any of the circuits described above may be used effectively to obtain these facilities and the design considerations are much the same for all of them. In each case the maximum bandwidth may be calculated most readily by conventional filter formulae using the equivalent forms indicated in the last section.

Maximum Bandwidth.—For the confluent half-section of Fig. 2 (b), the design formulae may, for a narrow-band filter, be written in the simple approximate form

$$C_2 = 1/(R.\Delta\omega), \quad L_1C_1 = L_2C_2 = 1/\omega_r^2$$
$$C_1 = \Delta\omega/(R.\omega_r^2), \quad C_1/C_2 = (\Delta\omega/\omega_r)^2 \quad \ldots \quad (4)$$

where ω_r is the mid-frequency of the pass-band, $\Delta\omega$ is the separation between cut-off frequencies and R is the surge impedance of the filter.

The fractional width $(\Delta\omega/\omega_r)$ of the pass-band is proportional to the square root of the ratio C_1/C_2, so that a wide pass-band requires C_1 to be as large, and C_2 as small, as other design factors permit. By way of example, the width of the pass-band is approximately 10 kc/s at a mid-frequency of 455 kc/s when $C_1 = 0.05$ $\mu\mu$F and $C_2 = 100$ $\mu\mu$F. The bandwidth (at 6 db. down) obtained in practice will differ somewhat from the nominal width of the pass-band ; but practical experience soon shows what allowance must be made on this account, or the actual response curve may be calculated in the usual way, taking into account dissipation and reflection losses.

Dissipation losses may usually be neglected, because losses in the series arm L_1C_1 are negligible owing to the low decrement of the crystal resonator, and because losses in the parallel arm L_2C_2 may be utilised as part of the load terminating the filter as long as they do not exceed the required load. In practice, reflection losses in the region of a cut-off frequency chiefly determine the shape of the response curve in this region. To maintain a flat-topped response, the filter must generally be terminated, at mid-shunt, by a resistance higher than the surge impedance R, and, in mid-series, by a resistance lower than R, the insertion loss being increased in both cases by mismatch at the centre of the pass-band. To utilise the loss in the parallel arm L_2C_2 as the mid-shunt termination, the magnification factor Q_2 of the circuit must be high enough to make $Q_2\omega_rL_2$ equal to, or greater than, the terminating resistance chosen, e.g., in a practical case $Q_2\omega_rL_2 \geqslant 2R$ was required by the choice of a terminating resistance equal to twice the surge resistance.

To obtain a wide band, the effective series capacitance C_1 of the crystal must be made great enough to satisfy equations (4), once the value of C_2 has been fixed. The choice of C_2 depends to some extent on details of design, but any value much less than 100 $\mu\mu$F is likely to introduce production difficulties, while a value approaching 200 $\mu\mu$F is desirable to avoid unwanted effects due to minor changes in capacitance, if the filter is connected directly in a valve amplifier without impedance transformation. In any case, the crystal series capacitance C_1 must be approximately 0.05 $\mu\mu$F at 455 kc/s to obtain bandwidths of 5 to 10 kc/s. Such a high value of C_1 requires some care in the design and mounting of the crystal resonator. The crystal itself having been designed to give the maximum value of C_1 consistent with manufacturing limitations, it must be mounted between electrodes with a very small air gap and, if extreme bandwidth is required, the electrodes should be sputtered directly on to the appropriate faces of the crystal. In practice the sputtered electrodes permit an increase, by a factor of about two, in the effective value of C_1.

Variation of Bandwidth.—Proper variation of the received bandwidth, by variation of the filter cut-off frequencies, would require extensive changes in the value of the filter constants and, in practice, the usual filter design conditions could not be realised for very narrow bands owing to dissipation in the parallel arm L_2C_2. An effective variation of width can, however, be attained readily by

(a) Drastic mismatching, obtained by lowering the resistance terminating the filter at mid-shunt, or

(b) Detuning the parallel arm L_2C_2, which may be regarded as loading the filter at mid-shunt with a reactance.

In either case, when the mismatching process is taken to the extreme to obtain a bandwidth of a few hundred cycles, it is simplest to regard the detuned or loaded parallel arm as a relatively low impedance in series with the crystal. In either case a considerable variation of gain, compared with that at full bandwidth, occurs; with resistive mismatching there is a considerable loss, while detuning of the parallel arm gives an appreciable gain, at a frequency slightly different from ω_r, due to resonance of L_1C_1 and the detuned circuit L_2C_2, all connected in series. The minimum bandwidth attainable is calculable immediately from a consideration of the overall decrement, at the frequency of maximum response, of the circuit comprising L_1C_1 connected in series with the parallel combination L_2C_2 and with the input circuit of the half-section. Where a full π-section is used both parallel arms must be detuned or loaded resistively.

Tuning of Rejector Frequencies.—In any of the arrangements of Fig. 2, the resonance frequency ω_r of the series arm is determined by the crystal itself; but the anti-resonance frequency, or frequencies, may be varied by variation of the effective capacitance across the crystal. In the bridge circuit of Fig. 2(a), variation of the balancing capacitance C_B may be utilised for this purpose, and the rejector frequency is above or below the resonance frequency according as C_B is less than, or greater than, the value required for balance. In the m-derived sections, the value of capacitance across the crystal may be varied directly and the two rejector frequencies move in the same direction, but the frequency-difference between them increases. In any case, a wide variation of the rejector frequencies is obtained and, in practice, this variation is, to a considerable extent, independent of the terminating conditions determining the effective bandwidth, so that it is useful in the rejection of interfering signals.

4. Experimental Results

The aims of the experimental investigation were to confirm the validity of the calculations of maximum bandwidth using equations (4), to confirm the substantial similarity in performance of the bridge and m-derived sections, to confirm the variation in gain due to the effective variation of bandwidth by mismatching, and to compare the minimum effective bandwidth attainable with respect to the value calculated along the lines indicated in the last section.

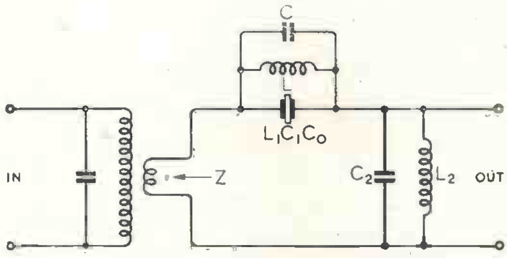

Fig. 3.—*Arrangement of m-derived filter.*

The crystals used were all X-cut bars of a type produced commercially and were approximately 20 mils thick, one-quarter of an inch wide and five-eighths of an inch long. They were entirely free from subsidiary resonances over a range exceeding plus and minus 30 kc/s from the main resonance at 465 kc/s. As produced commercially, these crystals are mounted between flat electrodes and an air gap not greater than 1.0 mil and have an effective series capacitance of approximately 0.017 $\mu\mu$F and an effective series resistance not greater than 2,000 ohms. For the purpose of this investigation a number of these crystals were mounted using gold electrodes sputtered directly on to the major faces, and for each of these crystals the effective series capacitance C_1 was approximately 0.035 $\mu\mu$F, while the series resistance was always considerably less than 2,000 ohms.

The measurements made are exemplified by the curves of Figs. 4 and 5 for an m-derived filter, arranged as shown in Fig. 3. The input circuit comprised a transformer, tuned to the mid-frequency ω_r, of impedance Z, and the constants of the crystal used were :—

$$C_1 = 0.035\ \mu\mu\text{F}$$
$$C_2 = 180\ \mu\mu\text{F}$$
$$C_o + C = 40\ \mu\mu\text{F}$$
$$\omega_r/2\pi = 465\ \text{kc/s}$$
$$L_1C_1 = L_2C_2 = L(C + C_o) = 1/\omega_r^2$$

giving the nominal bandwidth $\Delta\omega$, surge impedance R, and rejector frequencies ω_{a1} and ω_{a2} as follows :—

$$\Delta\omega/2\pi = 6.5\ \text{kc/s},$$
$$R = 136{,}000\ \text{ohms},$$
$$(\omega_{a2} - \omega_{a1})/2\pi = 13.8\ \text{kc/s}.$$

With an effective load of 200,000 ohms across the output circuit, consisting of the loss in coil L_2, the response of the circuit was that shown in curve A of Fig. 4. It is to be noted that the response in the pass-band is substantially flat over a considerable range due to the mismatch termination of 200,000 ohms, and that the rejector frequencies and the bandwidth are in good agreement with the calculated values.

Curves B, C, D of Fig. 4 illustrate variation of effective bandwidth, without change of nominal bandwidth, and without loss of symmetry of response, when the output circuit L_2C_2 was modified by connecting in series in the inductive arm resistances of 100, 1,000 and 10,000 ohms respectively. This method of varying the effective band-

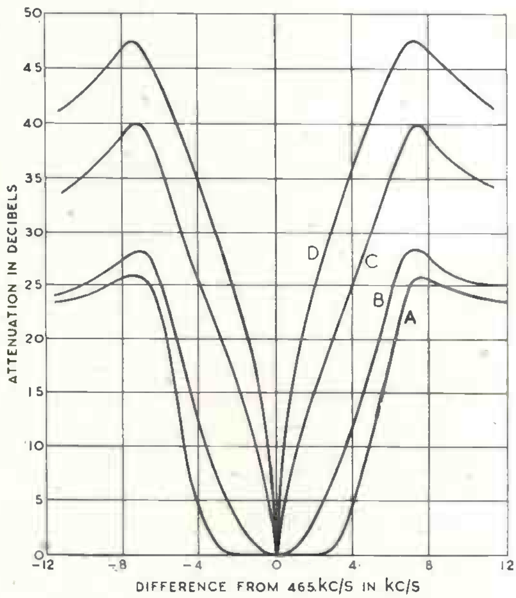

Fig. 4.—*Characteristics of the circuit of Fig. 3, showing variation in effective bandwidth by mismatching.*

width (Oram, 1938) has been used to avoid excessive variation in the transmission loss at the centre of the pass-band; in the present case the loss at mid-band for the conditions of curves B, C, D compared with that of curve A were approximately 1, 3, and 5 db. respectively.

The curves D, E, and F of Fig. 5 illustrate the tuning of the rejector frequencies by variation of the capacitance C

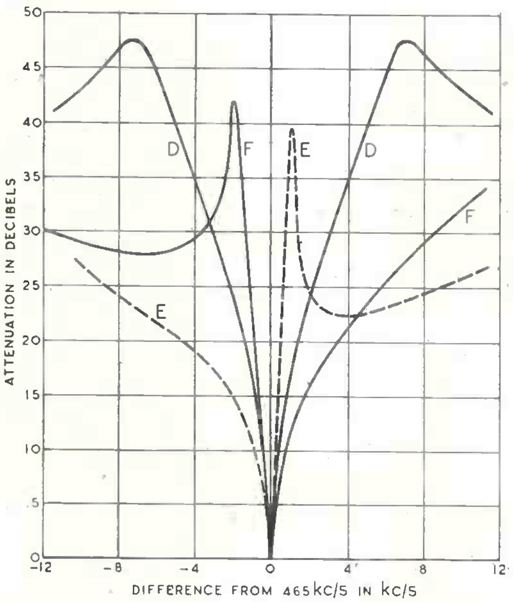

Fig. 5.—*Characteristics of the circuit of Fig. 3, showing tuning of the rejector frequencies for the bandwidth condition D of Fig. 4.*

across the crystal. Curve D is identical with curve D of Fig. 4; curve E was obtained by reducing the value of C from its nominal to its minimum value, both rejector frequencies moving upwards with the lower approaching mid-band frequency; curve F shows the effect of an increase in the value of C. Corresponding curves for the bridge-balanced circuit are well known.

These, and other measurements along the same lines, have confirmed the following conclusions :—

(1) The maximum bandwidth attainable with a given crystal and for given circuit components is substantially the same for Robinson's bridge circuit, for an m-derived filter, and for Mason's bridged-T filter.

(2) This maximum bandwidth may be calculated in any case by the usual filter formulae for confluent or m-derived sections.

(3) The m-derived and bridged-T filters are characterised by having two rejector frequencies, which may be disposed symmetrically about the pass-band and may be spaced in accordance with design requirements by appropriate choice of circuit constants.

(4) The characteristics of Robinson's bridge circuit in the balanced condition are indistinguishable from those of a confluent filter section, characterised by absence of rejector frequencies.

(5) The effective bandwidth may be reduced, without change in the nominal bandwidth and without loss of symmetry of the response curve, by mismatching the filter by (a) resistive mismatch by resistive loading of the parallel arm, (b) reactive mismatch by detuning the parallel arm, or (c) by an effective combination of these attained by adding large values of resistance in series with one or other of the elements of the parallel arm. The gain decreases rapidly with decrease of effective bandwidth for (a), increases rapidly for (b), and decreases somewhat for (c).

(6) The minimum effective bandwidth attainable depends on the effective series resistance of the crystal resonator and the total dissipation resistance in the circuit in series with the crystal. To obtain very narrow bandwidth the filter circuit must be fed from a source Z of low effective series resistance and the output circuit must be reduced to a similar circuit. The minimum effective bandwidth attainable is substantially the same for the various circuit arrangements discussed.

(7) The rejector frequencies in the m-derived and bridged-T filters may be tuned at will by varying the capacitance across the crystal. In Robinson's bridge circuit a rejector frequency is obtained by capacitive unbalance of the bridge, i.e. by variation of C of Fig. 2 (a), or by varying the capacitance across the crystal, and may be tuned to any desired point above or below the pass-band.

(8) When a full m-derived section or a bridged-T section is used, mismatching at both ends of the section is necessary to obtain

any considerable change in effective band-width.

It is therefore to be concluded that Robinson's bridge circuit, the m-derived section, and the bridged-T section are substantially similar in general performance; but that the choice of one or other of them gives some latitude in detailed design.

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In a subsequent issue of the "A.W.A. Technical Review" (Vol. 5, No. 5, 1941) the following note on the history of piezo-electric crystal filters by J. E. Benson was published.

In a recent paper (Builder and Benson 1941) dealing with some aspects of band-pass crystal filters, the authors ascribed the development of the balanced, single-frequency or "gate" filter to J. Robinson, in whose name a description of the Stenode Radiostat, employing this device, was published (Robinson 1931). Since writing this paper the work of earlier investigators has been brought to our notice. In order to clarify the situation, it is proposed in this note to outline briefly the historical development of the crystal filter as far as we have been able to trace it.

The earliest application of piezo-electric crystals to frequency-selective circuits seems to have been achieved by W. G. Cady (1920), whose patent describes the behaviour of piezo-electric elements in the neighbourhood of resonance and their consequent use in the selection and measurement of high frequencies. It appears that the earliest disclosure of the quartz-crystal band-pass filter, having recurrent sections, was due to L. Espenschied whose patent was filed Jan. 3rd, 1927. This was followed by W. A. Marrison's patent (filed June 7th, 1927) for a balanced crystal-gate filter designed for sharp response at a single frequency. In the same year (July 7th, 1927) C. W. Hansell filed a patent for a similar bridge-balanced system, in which the parallel capacitance of the crystal was balanced out by an equal capacitance supplied from the input circuit in opposite phase to the crystal. Single-frequency rejection filters of the T-section type having a piezo-electric element in the shunt arm were described at about the same time by I. F. Byrnes (1927).

Although Robinson had previously filed several patents (1926, 1928) dealing with the application of piezo-electric resonators in high-frequency electric signalling, it was not until July 26th, 1929, that he filed his patent (No. 337,049) for a balanced crystal-filter circuit of the type exemplified in his Stenode Radiostat. As far as the essential principles of the crystal balance circuit are concerned, Robinson's scheme appears to have been essentially similar to those previously described by Marrison and Hansell in 1927.

Acknowledgment

The author desires to thank L. Espenschied, of the Bell Telephone Laboratories, for some valuable comments on the early history of quartz-crystal filters.

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1928. Robinson, J., *British Patent*, 325,232 filed Aug. 10th, 1928; granted Feb. 10th, 1930.
1929. Robinson, J., *British Patent*, 337,049 filed July 26th, 1929; granted Oct. 27th, 1930.
1929. Robinson, J., *British Patent*, 337,050 filed July 26th, 1929; granted Oct. 27th, 1930.
1931. Robinson, J., *Radio News*, Vol. 12, p. 682.
1941. Builder, G. and Benson, J. E., *A.W.A. Tech. Rev.*, Vol. 5, p. 93.

Code for the Use of Valves

With a view to giving guidance to the designers of electronic equipment as to how to obtain the optimum life and performance from valves, some information on their use has been collated by the British Radio Valve Manufacturers' Association, and issued by the British Standards Institution as a War Emergency British Standard Code of Practice. Sections of the Code, which is published as B.S. 1106/1943, deal with valve dimensions, ratings, heater voltage regulation, mounting, heater/cathode insulation, etc. The charge for this four-page leaflet, copies of which may be obtained from the British Standards Institution, 28 Victoria Street, London, S.W.1, is 1s. post free.

Institute of Physics

At the meeting of the Electronics Group of the Institute of Physics on Tuesday, April 6th, Dr. J. H. Fremlin, of Standard Telephones & Cables, will deliver a paper on "Physics and the Static Characteristics of Hard Vacuum Valves." The meeting will be held at 6 o'clock in the Lecture Theatre of the Royal Institution, Albemarle Street, London, W.1.

Wireless Patents

A Summary of Recently Accepted Specifications

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

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

548 842.—Cascade filter arrangement for supplying power to a chain of amplifiers and for preventing undesirable inter-coupling.

The British Thomson-Houston Co. Convention date (U.S.A.), 12th July, 1940.

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

549 267.—Construction of tuning or like control knobs comprising two concentric knobs for giving coarse and fine adjustments.

Marconi's W. T. Co.; G. Payne; and H. H. Lightfoot. Application date, 9th May, 1941.

549 272.—Resistance-coupled broad-band amplifier with substantially uniform negative feed-back over the whole frequency band.

Standard Telephones and Cables (assignees of H. Nyquist). Convention date (U.S.A.), 20th September, 1940.

549 342.—Detector circuit for frequency-modulated signals and comprising linear and square-law frequency-responsive networks.

Hazeltine Corporation (assignees of H. A. Wheeler). Convention date (U.S.A.), 12th August, 1940.

TELEVISION CIRCUITS AND APPARATUS

For Transmission and Reception

549 008.—System for reproducing television signals by a cathode-ray tube with a movable light-modulating screen which is illuminated by a source external to the tube (addition to 543 485).

Ges. Zur. Forderung, &c. Technischen Hochschule (Zurich). Convention date (Switzerland), 11th June, 1940.

549 266.—Construction of a broad-band amplifier with high inherent stability and signal-to-noise ratio, particularly for television signals.

Standard Telephones and Cables; P. K. Chatterjea; and P. M. Brand. Application date, 11th April, 1941.

TRANSMITTING CIRCUITS AND APPARATUS

(*See also under Television*)

549 132.—Two-wire H.F. transmission line fitted with an auxiliary loop of wire for adjusting the characteristic impedance of the line.

E. W. Hayes. Application date, 18th August, 1941.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

549 125.—Valve circuit utilising a secondary-emission electrode for the rapid control of a load which is in series with that electrode and which takes a constant current.

The British Thomson-Houston Co. Convention date (U.S.A.); 13th July, 1940.

549 135.—Construction of a seal or cap for the lead-in wires of a high-powered electron-discharge tube.

The British Thomson-Houston Co.; H. K. Bourne; and E. J. G. Beeson. Application date, 19th August, 1941.

549 275.—Electrode assembly for a short-wave valve comprising a solid support or carrier-frame which also serves to screen a lead-in conductor.

Philips Lamps (communicated by N. V. Philips' Gloeilampenfabrieken). Application date, 14th October, 1941.

549 278.—Means for preventing undesired discharges from the edges of the accelerating anodes or wall-coatings on the glass surface of a C.R. tube.

Philips Lamps (communicated by N. V. Philips' Gloeilampenfabrieken). Application date 19th November, 1941.

SUBSIDIARY APPARATUS AND MATERIALS

549 074.—Timing relays for responding to trains of impulses used, say, for controlling the operation of selective devices in automatic systems of telephony (divided out of 549 047).

Automatic Telephone and Electric Co.; R. Taylor; and G. T. Baker. Application date, 28th April, 1941.

549 092.—Rope-and-pulley operating gear for a ganged variable-permeability tuning device.

Johnson Laboratories, Inc. (assignees of W. A. Schaper). Convention date (U.S.A.), 26th September, 1940.

549 110.—Electrical screening device made by spraying zinc powder over a surface of synthetic resin or other insulating material.

C. F. Lumb. Application date, 21st April, 1941.

549 186.—Frequency-measuring circuit, applicable as a field-intensity meter for directional transmitters or for testing the performance of radio receivers.

Marconi's W. T. Co. (assignees of D. S. Bond). Convention date (U.S.A.), 31st October, 1940.

549 292.—Compact construction of the "stacked" type of fixed condenser designed to allow the plates to be held under constant pressure.

Sangamo Weston. Convention date (U.S.A.), 12th June, 1941.

Waste Paper

It was recently stated by the Controller of Paper at the Ministry of Supply that whereas before the war less than 15 per cent. of the material used in making paper and board was repulped waste, the present waste paper content is 50 per cent.

Abstracts and References

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For the information of new readers it is pointed out that the length of an abstract is generally no indication of the importance of the work concerned. An important paper in English, in a journal likely to be readily accessible, may be dealt with by a square-bracketed addition to the title, while a paper of similar importance in German or Russian may be given a long abstract. In addition to these factors of difficulty of language and accessibility, the nature of the work has, of course, a great influence on the useful length of its abstract.

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PROPAGATION OF WAVES

983. ATTENUATION OF ELECTROMAGNETIC FIELDS IN PIPES SMALLER THAN THE CRITICAL SIZE.—E. G. Linder. (*Proc. I.R.E.*, Dec. 1942, Vol. 30, No. 12, pp. 554–556.)

An extension, mainly mathematical, of the work of Barrow (22 of 1937) and others on the propagation of electromagnetic waves in hollow metal tubes, with particular reference to the frequency region in which the transition from transmission to absorption takes place.

984. ULTRA-HIGH-FREQUENCY TRANSMISSION IN WAVES GUIDES [Brief Outline of the Theory and the Use of Transmission along Hollow Tubes].—L. A. Ware. (*Elec. Engineering*, Dec. 1942, Vol. 61, No. 12, pp. 598–603.)

985. "WAVE GUIDES" [Book Review].—H. R. L. Lamont. (*Wireless Engineer*, Nov. 1942, Vol. 19, No. 230, p. 514.) *See* also 3183 of 1942.

986. FREQUENCY MODULATION [Transmission, Propagation, & Reception].—C. Tibbs. (*Wireless World*, Feb. 1943, Vol. 49, No. 2, pp. 47–51.)

987. CORRECTION TO THE PAPER "ELECTRIC WAVES IN A STRATIFIED MEDIUM: ON THE QUESTION OF PARTIAL REFLECTION AND THE CALCULATION OF THE APPARENT HEIGHT OF IONOSPHERIC LAYERS" [Correction to Errors in Sign (Not affecting Final Equations or Numerical Results)].—K. Rawer. (*Ann. der Physik*, 21st Dec. 1942, Vol. 42, No. 4, pp. 294–296.) *See* 3879 of 1939.

988. THE STOREYS OF THE ATMOSPHERE [and the Need for a Consistent & Systematic Terminology].—H. Flohn & R. Penndorf. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1370: summary, from *Meteorol. Zeitschr.*, No. 1, Vol. 59, 1942, p. 1 onwards.)

According to this plan, based on vertical temperature gradients, prevailing wind directions, intermixing, turbulence, etc., the atmosphere is divided into an outer atmosphere above 800 km ("dissipation-sphere") where the particles can escape from the gravitational and magnetic fields of the earth, and an inner atmosphere up to 800 km. The latter is composed of three spheres, troposphere, stratosphere, and ionosphere, each being divided again into three layers. Thus in the troposphere there are the planetary "boundary" layer up to 1 km (itself subdivided into an "air-ground" layer up to 2 m, a "ground" layer from 2 to 100 m with linear increase of exchange coefficient with height, and an "upper" layer from 100 to 1000 m with constancy or decrease of coefficient with height), the "convection" layer from 1 to 8 km, and the "tropopause" layer from 8 to 12 km (in medium latitudes). Three types are distinguished in the tropopause, the normal type, the rising type, and the sinking type: the tropopause is the seat of the tropopause waves, the maximum interdiurnal or weather-related fluctuations of temperature and pressure, intensive vertical movements, compensation, and in- or out-pumping components of the flow. The stratosphere divides into the "isothermal" layer at 12–35 km (seat of the control waves), the "warm" layer between 35 and 50 km, and the upper "intermixing" layer at 50–80 km. Finally, the ionosphere is divided into three layers also, the E layer at 80–200 km, the F at 200–400 km, and the "atomic" layer between 400 and 800 km. Outside this comes the outer atmosphere or "dissipation-sphere" already mentioned.

989. Temperature and Origin of the D Region of the Ionosphere.—A. Vassy & É. Vassy. (*Comptes Rendus* [Paris], No. 6, Vol. 214, 1942, p. 282 onwards.)

For other recent *Comptes Rendus* Notes *see* 362, 633, & 640 of 1942. On the hypothesis that the luminous night clouds (just above the D layer, which lies at 60–80 km) are composed of ice crystals, Humphreys arrives at 160°K as the temperature at 80 km: this result is rejected because Humphreys' assumption of a constant temperature, independent of height, within the stratosphere is contradicted by other observations, and the writers prefer the Störmer–Vestine hypothesis of cosmic-dust clouds as the explanation of the luminous night clouds, and use it later to support their own argument. Budden, Ratcliffe, & Wilkes, from observations of the heights of the reflecting layer for various zenith distances of the sun, deduced 6 ± 0.5 km as the normal value of the homogeneous height H in the layer, this value decreasing to 4 km after magnetic disturbances, the temperature in the layer changing correspondingly from 205 ± 17°K to 140°K according to the relation $T=mgH/k$. These temperature changes also are rejected as improbable, and the writers conclude that the ionisation of the D layer must depend on the ionising of some component in the atmosphere other than the main constituents, a component to which Chapman's theory does not apply. Extending their earlier suggestion (362 of 1942), they attribute the normal D-layer ionisation to atmospheric sodium and/or other easily ionisable atoms of meteoric or other origin, supporting their belief by citing the presence of cosmic-dust clouds at 80 km (*see* above), the earlier height-determinations of the sodium twilight luminescence, and the observed changes of homogeneous height H at 80 km. The inapplicability of the Chapman theory to the D layer is due to the fact that the distribution of atmospheric sodium (or other dust clouds of cosmic origin) with height is different from that of the normal atmospheric gases. The hypotheses of Martyn and his associates, and of Mitra and his colleagues, attributing the D-layer ionisation to the formation of O^+ and O_2^+, are both rejected, on the ground that the necessary short-wave solar radiation would have been absorbed already by the higher atmosphere. Finally, the Wulf–Deming hypothesis (*see* for example 31 of 1938 and back reference) that the D-layer ions arise from an ionisation of the ozone by ultra-violet sunlight is rejected as qualitatively and quantitatively unsatisfactory, Brewer's investigations being quoted in this connection.

990. The Winter Temperatures of the Aurora over Tromsö.—R. Penndorf. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, pp. 1360–1361: summary, from *Meteorol. Zeitschr.*, No. 12, Vol. 58, 1941, p. 429 onwards.)

Further development of the work dealt with in 3492 of 1942. In Part I the writer discusses the question whether the E-layer temperature shows a yearly variation: mean temperatures for December appear to be −40°C, for March −49°C, the whole auroral layer thus apparently sinking by 800 m in the Spring. But these variations are within the limits of accuracy of the measuring method: no yearly variation of the winter temperature in the E layer over Tromsö could be established. The same result is given by calculations of the mean height of max. ionisation of the E layer over Washington in the months January/April.

Part II deals with the method of determining the temperature, based on spectroscopic investigations of the negative nitrogen bands 4278 and 3914 AU. The writer examines to what extent the temperature calculated from the energy distribution of the vibrational state agrees with the true gaseous temperature: he finds that the temperatures given by Vegard correspond approximately to the true values. In Part III he discusses Harang's contention that the part of an auroral arc which lies in darkness has a height of 100 km, whereas the part lying in the sun-lit region has a height of 130–140 km (*see* 2161 of 1940): he concludes that the supposition that these displacements indicate temperature changes is erroneous: "the lower auroral boundary in the illuminated atmosphere only appears to be higher than that in the dark because the excitation conditions are different in the two cases. Further, it cannot be regarded as a uniform isobaric surface."

In Part IV the wind conditions in the E layer are considered. Meteor observations point to west winds at heights above 80 km. Quantitative considerations actually suggest a pressure drop which would require west winds in the E layer. The velocity of the geostrophic wind for about 60°N works out at 104 m/s: the meteor observations in our latitudes give westerly winds of 20 to 60 m/s. An increase of wind towards North agrees with the writer's views on the temperature distribution, for in our latitudes the temperature contrasts of the E layer are slighter than in high latitudes. The writer's estimate gives an upper limit to the wind velocity and confirms the order of magnitude of the E-layer temperatures from ionospheric measurements in our latitudes, and the Vegard temperatures for the Polar night.

991. The Temperature of the High Stratosphere.—K. Wegener. (*Forschungen u. Fortschritte*, No. 9, Vol. 17, 1941, p. 101 onwards.)

Pressure at 100 km is now estimated at 1/100 mm Hg from the results of auroral observations. The reason for this considerably higher pressure, compared with earlier ideas, may be either (i) a relatively high temperature or (ii) the predominance of radiation-dissociated gases and/or gases of small molecular weight. Against the first possibility are the following facts: extrapolation of exploring-balloon observations in Abisko yields −35°C in summer and −73°C in winter for the boundary of the atmosphere: this last value agrees with the minimum value of upper-atmospheric temperature estimated at −70°C from observations of the temperature on the shadowed side of other planets. Further, Vegard and Tönsberg's spectroscopic researches on the auroral light give a mean temperature of −38.7°C, indicating the absence of any marked temperature-effect on the part of the dissociating radiation; while Regener's measurements show no rise in temperature due to the selective absorption producing, above 21 km, ozone from oxygen. The writer explains the occurrence of dead zones for sound waves, generally attributed to increased temperatures in the 25–50 km region, by the assertion that the waves in question are not ordinary sound waves but pulses of the Riemann type which are propagated more rapidly with decreasing air pressure and are consequently bent back to earth.

The pressure of gases of low molecular weight is supported by Regener's measurements and the Explorer II observations, which gave an increase in the helium content with height: the presence of oxygen also (molecular weight 32) in the upper atmosphere follows from the appearances of the principal line of atomic oxygen in the auroral spectrum. The fact that the observed splitting of the oxygen molecule in the auroral region and the formation of ozone do not bring about the calculated rise in temperature is attributed to erroneous assumptions in the calculation or to the weakness of the dissociating radiation.

992. THE ZENITHAL TWILIGHT BRIGHTNESS AND THE DENSITY OF THE ATMOSPHERE IN THE IONOSPHERE [up to 120–150 km: Recent Results with the Writer's Theory & Technique: Comparison with Density-Distribution Estimates obtained by Other Methods: Slight Gradient of Density above 120 km explained by Turbulence Velocity less than 1 km/s: etc.].—F. Link. (*Meteorol. Zeitschr.*, No. 1, Vol. 59, 1942, p. 7 onwards.) For a summary see *Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1380.

993. THE LIGHT OF THE NIGHT SKY [Survey of Present Knowledge, including Results of Spectrum Analysis].—F. Löhle. (*Zeitschr. f. angew. Meteorol.*, No. 7, Vol. 58, 1941, p. 201 onwards.)

(i) Geographical distribution and variation with time: (ii) connection with magnetic disturbances, earth currents, and sunspot frequency: (iii) contributions to knowledge of the upper-atmospheric layers: (iv) difficulties of orientation in night-flying at great heights.

Knowledge of (i) and (ii) is based exclusively on Rayleigh's long series of observations. Investigations on the brightness distribution lead to the conclusion that the seat of the luminosity lies partly in the earth's atmosphere. By assuming the existence of two, or several, luminous layers, extensive agreement is obtained between observation and calculation. The absolute measurements of brightness have not yet led to any unequivocal results.

994. AEROPLANE MEASUREMENTS OF THE INFRA-RED RE-RADIATION FROM THE ATMOSPHERE [Disagreement with Ångstrom's Formula: at Great Heights the Air radiates More Infra-Red than at Ground Level: Apparent Presence of Additional Infra-Red Source (Not Water Vapour or CO_2) which Increases with Height].—G. Falckenberg & F. Hecht. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1378: summary, from *Meteorol. Zeitschr.*, No. 11, Vol. 58, 1941, p. 415 onwards.)

995. FURTHER SPRING VALUES FOR THE GROUND-LEVEL OZONE AT AROSA.—F. W. P. Götz & R. Penndorf. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, pp. 1370–1371: summary, from *Meteorol. Zeitschr.*, No. 11, Vol. 58, 1941, p. 409 onwards.)

Using the optical method, with additional precautions. No connection between the "upper" ozone, or the general meteorological situation, and the ozone content of the ground-level air could be traced, but the latter appears to be linked with the *local* meteorological conditions by way of the moisture content of the air. The writers point out the unsuitability of wavelengths below 2576 AU for ozone measurements because of the present difficulty in allowing for oxygen absorption.

996. THE EARTH'S ATMOSPHERE AND THE CONSTITUTION OF THE PLANETS [Meteors and the Earth's Upper Atmosphere].—F. L. Whipple. (*Reviews of Modern Phys.*, April/July 1942, Vol. 14, No. 2/3, p. 139.) For a recent paper *see* 1607 of 1941.

997. COSMIC-RAY THEORY.—Rossi & Greisen. (*See* 1303.)

998. HYDROGEN ABUNDANCE AND TURBULENCE IN THE SUN'S ATMOSPHERE [Comprehensive Survey, including the Relation of the Turbulence Theory to the Mechanism of Formation of Faculae, Protuberances, & probably Ultra-Violet Eruptions].—P. ten Bruggencate. (*Physik. Berichte*, 14th July 1942, Vol. 23, No. 14, pp. 1454–1455: summary, from *Forschungen u. Fortschritte*, 1942.)

999. TRAVELLING WAVES ON TRANSMISSION LINES [Use of Laplacian Transformation with Tables of Fourier Integrals reduces Solution of Transmission-Line Problems to Same Status as evaluating Definite Integrals by Tables of Integrals].—E. Weber. (*Elec. Engineering*, June 1942, Vol. 61, No. 6, pp. 302–309.)

1000. PROPAGATION CONSTANT AND CHARACTERISTIC IMPEDANCE OF HIGH-LOSS TRANSMISSION LINES [Graphical and Analytical Method for Their Determination when High Loss results from Series Resistance].—K. Spangenberg. (*Electronics*, Aug. 1942, Vol. 15, No. 8, pp. 57–58.)

1001. ON THE PICKUP OF BALANCED FOUR-WIRE LINES.—C. W. Harrison. (*Proc. I.R.E.*, Nov. 1942, Vol. 30, No. 11, pp. 517–518.)

Author's summary:—"It is demonstrated that for practical purposes the pickup of undesired energy by a balanced four-wire line when compared to the pickup of a balanced two-wire line of the same spacing is so small as to be considered negligible."

1002. ON THE PHYSICAL PHENOMENA OF THE SKIN EFFECT [Flux Displacement] OF A MAGNETIC FIELD PRODUCED BY INDUCTION, AS A RESULT OF THE EDDY CURRENTS IN PLATES [Straightforward Mathematical Treatment, with Curves of Penetration Depths & the Rise in H.F. Resistance].—E. Hameister. (*Zeitschr. f. Fernmeldetech.*, 16th Jan. 1942. Vol. 23, No. 1, pp. 7–9.)

1003. High-Frequency Resistance of Plated Conductors.—Proctor. (*See* 1278.)

1004. The Theory of Optical Polarisability and the Natural Rotating Power [and the Calculation of the Phase Difference between the Electric Field of a Light Wave & the Induced Electric Moment of a Molecule of the Body traversed: Behaviour of the Complex Components of the Hermitic Polarisability Tensor at Very Long and Very Short Wavelengths: etc.].—J. P. Mathieu. (*Physik. Berichte*, 15th June 1942, Vol. 23, No. 12, p. 1253: summary, from *Comptes Rendus* [Paris], No. 9, Vol. 214, 1942, p. 420 onwards.)

1005. Simplified Derivation of the Formulae for the Fraunhofer Diffraction Phenomena [Treatment avoiding Kirchhoff's Use of the Huyghens Principle and based only on the Fourier Integral Theorem].—H. Scheffers. (*Ann. der Physik*, 3rd Dec. 1942, Vol. 42, No. 2/3, pp. 211–215.)

1006. Relation between the Interference Fringes of Two-Wave [Michelson] Interferometers and Multiple-Wave [Pérot-Fabry] Interferometers.—P. M. Duffieux. (*Physik. Berichte*, 15th June 1942, Vol. 23, No. 12, pp. 1251–1252: summary, from *Comptes Rendus* [Paris], No. 6, Vol. 214, 1942, p. 304 onwards.)

ATMOSPHERICS AND ATMOSPHERIC ELECTRICITY

1007. The Separation of Electricity in Clouds [Discussion of Phenomena, with Suggested Theory of Mechanism].—J. A. Chalmers. (*Phil. Mag.*, Jan. 1943, Vol. 34, pp. 63–67.)

1008. Lightning Investigations on Wallenpaupack–Siegfried 220-kV Line of Pennsylvania Power and Light Company.—E. Bell & F. W. Packer. (*Elec. Engineering*, April 1942, Vol. 61, No. 4, Transactions pp. 196–201.)

1009. Lightning Investigation at High Altitudes in Colorado [Determination of Probable Lightning Current at Altitudes from 6000 to 13 500 Feet: Measurement of Corona Current].—L. M. Robertson, W. W. Lewis, & C. M. Foust. (*Elec. Engineering*, April 1942, Vol. 61, No. 4, Transactions pp. 201–208.)

1010. The Application of Wireless Technique to the Study of Lightning Protection [High-Frequency Methods of measuring Earth-Connection Resistances].—V. Fritsch. (*Funktech. Monatshefte*, July 1942, No. 7, pp. 99–104.) *See* also, for example, 707 of March, and 1011, below.

1011. Measurements on Lightning-Conductor Earths: III—High-Frequency Method.—V. Fritsch. (*Arch. f. Tech. Messen*, Oct. 1942, Part 136, V35192-7, Sheets T99–100.) For previous parts *see* 3517 of 1942.

1012. Unusual Behaviour of the Atmospheric-Electrical Potential Drop in Potsdam [at Two Observation Points ("Tower" & "Meadow"), about 100 m apart and with 25 m Difference in Altitude: on Three Occasions (Fine Weather) Potential Drop showed Opposed Behaviour, so that Reduction Factor between the Two Points, generally Constant, was subject to Very Large Fluctuations].—M. Krestan. (*Meteorol. Zeitschr.*, No. 3, Vol. 59, 1942, p. 98 onwards.)

1013. Contribution to Knowledge of Atmospheric Electricity, No. 77: On the Loss of Charge of a Sphere in Stationary and in Moving Air.—E. von Schweidler. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1364.)

1014. A Height Integrator [for Automatic Solution of the Ideal Barometric Height Formula], *and* The "Calculating" Theodolite [for Use with Pilot Balloons].—J. Lugeon: K. Burkhart. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1369: p. 1369.)

PROPERTIES OF CIRCUITS

1015. On the Natural Electromagnetic Oscillations of an Ellipsoidal Cavity.—M. Jouguet. (*Comptes Rendus* [Paris], No. 5, Vol. 214, 1942, p. 214 onwards.)

Pointing out the simplification in the theoretical treatment of the problem of the natural oscillations of a cavity bounded by a surface of rotation, when the cavity is of ellipsoidal form, elongated or flattened.

1016. Oscillatory Circuits for Ultra-High Frequencies [Outline of Calculation & Design of Coaxial & Parallel-Wire Lines as Resonators].—K. Schüssler. (*Zeitschr. f. Fernmeldetech.*, 17th Oct. 1942, Vol. 23, No. 10, pp. 151–154.)

1017. The Properties of Valves as Admittances in the Ultra-Short-Wave Region, and Transit-Time Phenomena.—Hameister. (*See* 1097.)

1018. Reactance Tubes in Frequency-Modulation Applications [Behaviour & Physical Operation].—A. Hund. (*Electronics*, Oct. 1942, Vol. 15, No. 10, pp. 68–71 and 143.)

1019. Frequency-Modulation Detector.—Standard Telephones & Cables. (*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, p. 222.) Brit. Patent No. 546 721.

1020. Distortionless Detection [Discussion before the Wireless Section].—(*Journ. I.E.E.*, Part III, Sept. 1942, Vol. 89, No. 7, pp. 175–176.)

1021. A New Second Detector [using a Triode-Diode with Common Cathode, and having the Advantages of Diode Detection, while presenting a High Impedance to the Preceding Circuit].—F. C. Everett. (*Communications*, Sept. 1942, Vol. 22, No. 9, pp. 10–11 and 34.)

1022. RECEIVER INPUT CIRCUITS: DESIGN CONSIDERATIONS FOR OPTIMUM SIGNAL/NOISE RATIO.—R. E. Burgess. (*Wireless Engineer*, Feb. 1943, Vol. 20, No. 233, pp. 66–76.)

1023. THE CATHODE FOLLOWER: PART II [Performance at Higher Frequencies: Design, Formulae, and Data Sheets].—C. E. Lockhart. (*Electronic Eng:g*, Feb. 1943, Vol. 15, No. 180, pp. 375–382.) Continued from Dec. issue, No. 178, Vol. 15, pp. 287–293, where the analysis for low frequencies, and the relevant data sheets, were given.

1024. DESIGNING A RESISTANCE-LOADED PUSH-PULL INVERTOR [Calculation of Circuit Parameters]. — R. Feinberg. (*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, pp. 206–207.) *See* also 1233, below.

1025. LOW-FREQUENCY CHARACTERISTICS OF THE COUPLING CIRCUITS OF SINGLE AND MULTI-STAGE VIDEO AMPLIFIERS.—Donley & Epstein. (*See* 1169.)

1026. WAVE-FORM CIRCUITS FOR CATHODE-RAY TUBES: PART II [Amplitude and Impedance Method of changing Wave Shape].—H. M. Lewis. (*Electronics*, Aug. 1942, Vol. 15, No. 8, pp. 48–53.) For Part I *see* 3690 of 1942.

1027. OPTIMUM CONDITIONS IN CLASS A AMPLIFIERS. — G.W.O.H: Nottingham. (*Wireless Engineer*, Feb. 1943, Vol. 20, No. 233, pp. 53–55.) Editorial prompted by Nottingham's paper, 1634 of 1942.

1028. TUNING METHOD FOR ELECTRICAL OSCILLATORY CIRCUITS [Phase-Change Method].—Beyerle & Schweimer. (*See* 1189.)

1029. AN OSCILLATOR FOR REMOTE FREQUENCY CONTROL [Single Pentode used as Combined Oscillator and Reactance Tube to vary the Frequency of Generated Oscillations over Two Percent Range through Changes of D.C. Voltage applied to the Tube: Discussion of Design Factors affecting Frequency Stability].—H. C. Lawrence. (*Electronics*, Sept. 1942, Vol. 15, No. 9, pp. 42–43.)

1030. THERMAL-FREQUENCY-DRIFT COMPENSATION [Conditions necessary for Minimising Frequency Drift with Temperature Variation for Various Types of Circuit].—T. R. W. Bushby. (*Proc. I.R.E.*, Dec. 1942, Vol. 30, No. 12, pp. 546–553.) From Amalgamated Wireless, Australasia, Ltd.

1031. RADIO-CIRCUIT STABILITY [Summary of Paper submitted to London Students Section, I.E.E: Types of Thermally-Stable Coils and Condensers: Good Stability of Silvered Mica or Ceramic types].—C. W. Eggleton. (*Elec. Review*, 20th Nov. 1942, Vol. 131, No. 3391, p. 656.)

1032. FREQUENCY STABILITY OF TUNED CIRCUITS [Discussion before the Wireless Section].—(*Journ. I.E.E.*, Part III, Sept. 1942, Vol. 89, No. 7, pp. 173–174.)

1033. STABILISED OSCILLATOR GENERATOR [having Positive Feedback at Oscillating Frequency and Negative Feedback at All Other Frequencies].—F. E. Terman. (*Electronic Eng:g*, Nov. 1942, Vol. 15, No. 177, p. 249.) Brit. Patent No. 541 423, S.T. & C. Ltd.

1034. SQUEGGING OSCILLATORS [Mode of Operation].—E. Hughes. (*Wireless World*, Feb. 1943, Vol. 49, No. 2, pp. 52–53.)

1035. SOME CHARACTERISTICS OF A STABLE NEGATIVE RESISTANCE. — C. Brunetti & L. Greenough. (*Proc. I.R.E.*, Dec. 1942, Vol. 30, No. 12, pp. 542–546.)

The use of positive feedback in a two-stage amplifier makes it possible to obtain an input impedance equal to the impedance in the feedback circuit multiplied by a negative constant. For a resistance–capacitance feedback circuit with a proper choice of circuit constants, the input impedance can be made to approximate closely to a pure negative resistance over any given frequency range. In the apparatus described, high stability is secured by the use of inverse feedback in addition to the positive-feedback loop.

1036. A SELECTIVE CIRCUIT AND FREQUENCY METER USING A TUNING FORK [Tuning Fork and Coil used as a Two-Terminal Sharply-Selective Circuit].—D. G. Tucker. (*Electronic Eng:g*, Aug. 1942, Vol. 15, No. 174, pp. 98–101.)

1037. INTEGRATION IN THE COMPLEX PLANE [a Mathematical Prerequisite for Work on Laplacian Transforms, Fourier Integrals, and Travelling Waves on Transmission Lines].—K. O. Friedrichs. (*Elec. Engineering*, March 1942, Vol. 61, No. 3, pp. 139–143.)

1038. LAPLACIAN TRANSFORM ANALYSIS OF CIRCUITS WITH LINEAR LUMPED PARAMETERS.—J. Millman. (*Elec. Engineering*, April 1942, Vol. 61, No. 4, pp. 197–205.) For criticism *see* Lyon, 726 of March.

1039. THE STEADY-STATE RESPONSE OF CIRCUITS [Part Two of Two-Part Paper].—D. L. Waidelich. (*Communications*, Nov. 1942, Vol. 22, No. 11, pp. 13–15.) Part One was in the issue for October, 1942.

1040. STEADY STATE CURRENTS IN ELECTRICAL NETWORKS [Extension of Operational Circuit Analysis to give the Result in Terms of the Sum Function of a Fourier Series, which is Useful in determining the Wave Form of the Current: Three Methods].—D. L. Waidelich. (*Journ. Applied Phys.*, Nov. 1942, Vol. 13, No. 11, pp. 706–712.)

1041. IMPEDANCE MAGNITUDE AND PHASE SHIFT CURVES FOR SOME COMMON TWO-TERMINAL THREE-ELEMENT LINEAR NETWORKS.—V. L. Edutis. (*Electronics*, Nov. 1942, Vol. 15, No. 11, pp. 76–77.)

1042. Phase Shifting and Amplitude Control Networks [Design Charts for determining the Network required to deliver Proper Power to Each of Several Loads fed by the Same Source: Phase and Amplitude Relations of Coupling Network Easily Determined].—W. S. Duttera. (*Electronics*, Oct. 1942, Vol. 15, No. 10, pp. 53–55.)

1043. "T" to "Pi" Transformations Simplified [Formulae in an Easily Remembered Form, with Examples].—H. Stockman. (*Electronics*, Oct. 1942, Vol. 15, No. 10, pp. 72–73 and 160.)

1044. Symmetrical Electrical Systems: I [Special Method for Determining the Characteristics of Structurally and Electrically Symmetrical Four-Terminal Networks].—E. S. Purington. (*Electronics*, Nov. 1942, Vol. 15, No. 11, pp. 54–57.)

1045. Calculation of Insertion Loss and Phase Change of a 4-Terminal Reactance Network [Simpler and More General Method than Usual, Capable of Great Accuracy, Applicable to Any 4-Terminal Reactance Network operating between Constant Resistance Terminations, provided the Ratio of Image Impedance to Terminating Resistance is the Same at Both Ends].—H. Stanesby, E. R. Broad, & R. L. Corke. (*P.O. Elec. Eng. Journ.*, Oct. 1942, Vol. 35, Part 3, pp. 88–92.) For a further paper, dealing with the use of a special slide rule and with the correction for dissipation, see *ibid.*, Jan. 1943, Vol. 35, Part 4, pp. 111–114.

1046. The Wien Bridge and Some of Its Applications [as a Filter and in Oscillator Circuits].—J. S. Worthington. (*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, pp. 214–216.)

1047. Notes on Band-Pass and Band-Rejection Filters [to Simplify the Calculation of Their Performance, especially when allowing for Dissipation in the Coils].—H. Halubow. (*Electronics*, Aug. 1942, Vol. 15, No. 8, pp. 54–56.)

1048. Notes on R.F. Attenuator Design [Attenuating Networks of Resistive Elements].—R. E. Blakey. (*Journ. of British I.R.E.*, Dec. 1942/Jan. 1943, Vol. 3.)

1049. Unsymmetrical Attenuators [Graphical Method of designing T or Pi Resistance Attenuators, and a Simplified Means of Converting to a Dissymmetrical Network where Impedances of Different Magnitude must be Matched].—P. M. Honnell. (*Electronics*, Aug. 1942, Vol. 15, No. 8, pp. 41–43.)

1050. The Accurate Generation of Sub-Frequencies from a Standard.—E. Newman. (*Electronic Eng:g*, Nov. 1942, Vol. 15, No. 177, pp. 244 and 249.)

1051. The Generation of Groups of Harmonics [by the Use of Rectifiers and by Saturated Iron Inductors].—D. G. Tucker. (*Electronic Eng:g*, Nov. 1942, Vol. 15, No. 177, pp. 232–237.)

1052. Frequency Division without Free Oscillation [System which does Not involve Locked Oscillators].—D. G. Tucker & H. J. Marchant. (*P.O. Elec. Eng. Journ.*, July 1942, Vol. 35, Part 2, pp. 62–64.)

1053. The Half-Wave Voltage-Doubling Rectifier Circuit [Performance Characteristics of Circuit determined by Analysis: Experimental Verification].—D. L. Waidelich & C. H. Gleason. (*Proc. I.R.E.*, Dec. 1942, Vol. 30, No. 12, pp. 535–541.)

1054. Method for A.C. Network Analysis using Resistance Networks.—W. E. Enns. (*Elec. Engineering*, Dec. 1942, Vol. 61, No. 12, Transactions p. 875.)

1055. "A.C. Calculation Charts" [for Reactance Computations: Book Review].—R. Lorenzen. (*Proc. I.R.E.*, Dec. 1942, Vol. 30, No. 12, p. 558.)

1056. Linear Power Scales [Circle Diagrams for Power Relationships in Electrical Machines].—G. F. Freeman. (*Elec. Review*, 12th Feb. 1943, Vol. 132, No. 3403, pp. 210–212.)

TRANSMISSION

1057. A Method of Frequency-Modulating a Quartz-Controlled Transmitter.—I. Koga. (*Zeitschr. f. Fernmeldetech.*, 15th April 1942, Vol. 23, No. 4, pp. 63–64: summary, from *Electrotech. Journ.* [Tokyo], No. 4, 1941.)

By varying the electrical values of the quartz by the use of rectifiers, in parallel, of low electrode capacity, whose internal resistance is modulated. The rectifiers take the form of indirectly heated valves, since the electrode capacities of dry-plate rectifiers, and the filament capacity to earth of directly heated valves, are so large that the quartz would be practically short-circuited. The circuit and some modifications are shown in Fig. 1 a–f, and some modulation curves in Fig. 2.

1058. Grounded-Plate F.M. Amplifier Notes.—D. Phillips: Skene. (*Communications*, Oct. 1942, Vol. 22, No. 10, pp. 49–50.) *See* also 2642 and 3232 of 1942.

1059. Amplitude, Frequency, and Phase Modulation Relations [Comparison between the Three Methods from the Standpoint of Mode of Operation, Modulation Factor, and Frequency Spectrum: Graphical Means of determining Spectrum for F.M. and A.M: Differences between F.M. and P.M.].—A. Hund. (*Electronics*, Sept. 1942, Vol. 15, No. 9, pp. 48–54: Corrections, Oct. 1942, Vol. 15, No. 10, p. 142.)

1060. REACTANCE TUBES IN FREQUENCY-MODULATION APPLICATIONS [Behaviour & Physical Operation].—A. Hund. (*Electronics*, Oct. 1942, Vol. 15, No. 10, pp. 68–71 and 143.)

1061. FREQUENCY MODULATION [Transmission, Propagation, & Reception]. — C. Tibbs. (*Wireless World*, Feb. 1943, Vol. 49, No. 2, pp. 47–51.)

1062. "FREQUENCY MODULATION" [Book Review]. —K. R. Sturley. (*Journ. of Scient. Instr.*, Jan. 1943, Vol. 20, No. 1, p. 20.) For papers by the same writer *see* 2145 of 1942 and 596 of February.

1063. A LOW-POWER AIRCRAFT TRANSMITTER.—W. W. Honnor. (*A.W.A. Tech. Review*, No. 1, Vol. 6, 1942, pp. 37–41.)

A remote-controlled 10-watt transmitter which provides for operation with telephony or tone telegraphy on any one of six crystal-controlled frequencies, three of which are in each of the 3 and 6 Mc/s aircraft bands. The equipment, excluding aerial and 12-volt battery, weighs less than 50 lb.

1064. QUARTZ CRYSTALS IN RADIO.—Baldwin. (*See* 1182.)

1065. COPPER-OXIDE RECTIFIERS IN STANDARD BROADCAST TRANSMITTERS.—Harmon. (*See* 1240.)

1066. THE GENERATION OF SINUSOIDAL ALTERNATING CURRENTS IN THE INFRA-SONIC FREQUENCY BAND.—Müller. (*See* 1355.)

RECEPTION

1067. ULTRA-HIGH-FREQUENCY RADIO-RANGE AND MARKER RECEIVERS FOR AIRCRAFT.—Builder & Downes. (*See* 1120.)

1068. RECEIVER INPUT CONNECTIONS FOR ULTRA-HIGH-FREQUENCY MEASUREMENTS.—Rankin. (*See* 1179.)

1069. REDUCTION OF BAND WIDTH IN FREQUENCY-MODULATION RECEIVERS. — D. A. Bell. (*Wireless Engineer*, Nov. 1942, Vol. 19, No. 230, pp. 497–502.)

A discussion of the possibility of using a high degree of negative feedback of frequency modulation in the i.f. section of a frequency-modulation receiver for the purpose of (a) minimising the necessary i.f. band width and (b) making the detected output independent of amplitude without the use of an amplitude limiter in the i.f. amplifier.

1070. FREQUENCY MODULATION [Transmission, Propagation, & Reception]. — C. Tibbs. (*Wireless World*, Feb. 1943, Vol. 49, No. 2, pp. 47–51.)

1071. "FREQUENCY MODULATION" (Book Review]. —Sturley. (*See* 1062.)

1072. FREQUENCY-MODULATION DETECTOR.—Standard Telephones & Cables. (*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, p. 222.) Brit. Patent No. 546 721.

1073. A NEW SECOND DETECTOR [and Its Advantages].—Everett. (*See* 1021.)

1074. DISTORTIONLESS DETECTION [Discussion before the Wireless Section].—(*Journ. I.E.E.*, Part III, Sept. 1942, Vol. 89, No. 7, pp. 175–176.)

1075. THE NOISE OF RECEIVING APPARATUS AT VERY HIGH FREQUENCIES.—M. J. O. Strutt & A. van der Ziel. (*Philips Tech. Rundschau*, No. 6, Vol. 6, 1941, p. 175 onwards: summary in *Physik. Berichte*, 1st Aug. 1942, Vol. 23, No. 15, pp. 1512–1513.)

The ratio between fluctuation currents and signal currents in the anode circuit of the first valve is taken as a measure of the interference: it can be expressed as a function of the aerial, circuit, and valve properties by the equation $\overline{\delta I_a^2}/I_a^2 = \{4kT\Delta f/(V^2_{ant}/2R_{ant})\}\,.\,\{a/2 + R_r(1/R_{LC} + 1/R_e) + \sqrt{R_r/R_{LC} + 5{\cdot}6R_r/R_e + R_r^2(1/R_{LC} + 1/R_e)^2}\}$, which contains in combination the contributions of the four noise-components, namely the disturbance to the signal in the aerial by other waves from world space (this is represented by the expression given by Bakker, *ibid.*, Vol. 6, p. 129: $\overline{v^2}_{ant} = a\,.\,4kTR_{ant}\Delta f$, where a is a numerical factor 1..10), an interfering voltage v_c in series with the resonance resistance of the oscillatory circuit (represented by $\overline{v_c^2} = 4kTR_{LC}\Delta f$), an interfering current i_k in the anode circuit, arising from spontaneous fluctuations in number and velocity of the electrons from the cathode, and represented by $\overline{i_k^2} = 4kTR_rS^2\Delta f$, where R_r is the equivalent noise resistance, and finally the spontaneous current-fluctuations in the control-grid circuit (represented by $\overline{i_g^2} = 1.43\,.\,4kT_k/R_e\,.\,\Delta f$, where R_e is the reciprocal of the input damping.)

This equation shows that apart from the "cosmic" noise (for which the receiver is not responsible) the ratio between fluctuation current and signal current at "low" frequencies such as those of the broadcast band is determined almost entirely by the ratio R_r/R_{LC}, so that if the circuit resistance R_{LC} is large enough the receiver can amplify even very weak signals practically without trouble from fluctuation noise. With high frequencies the effect of the circuit damping has added to it the electronic input damping, and the ratio between fluctuation and signal currents is then governed by R_r/R_e, the ratio of the equivalent noise resistance to the electronic input resistance of the valve. The damping due to the self-inductance of the cathode lead has no effect on the fluctuation/signal ratio.

The above treatment is then applied to pentodes, whose behaviour is found not to differ in principle from that of triodes. Fluctuations in distribution can be compensated by the insertion of a suitable self-inductance in the screen-grid lead.

1076. RECEIVER INPUT CIRCUITS: DESIGN CONSIDERATIONS FOR OPTIMUM SIGNAL/NOISE RATIO.—R. E. Burgess. (*Wireless Engineer*, Feb. 1943, Vol. 20, No. 233, pp. 66–76.)

1077. RECEIVER INPUT CIRCUITS.—H. Behling. (*Funktech. Monatshefte*, July 1942, No. 7, pp. 89–96.)

The two main points about a receiver are its selectivity and its sensitivity. In a previous paper (*ibid.*, No. 5, 1940) the writer dealt with methods of raising, by comparatively simple means, the selectivity of a receiver, and the present work deals with the design calculations for input circuits with respect to their effect, within certain limits, in increasing the sensitivity. Other workers (for example Fränz, 3126 of 1939) have shown that the optimum signal/noise ratio would be obtained from any aerial by connecting it directly to the grid of the first valve, provided that there were no valve noise: a voltage transformation of the wanted signal is only required because of valve noise. In the case where the over-all amplification of the receiver is so low that even at the output there is no noticeable noise level, a voltage transformation at the input circuit leads to an increase in the receiver amplification. "The following considerations and calculations are thus intended to lead to the obtaining of the greatest possible ratio of the voltage E_g, controlling the grid of the first valve, to the available voltage E_a delivered by the aerial through an internal resistance $\mathfrak{R}_i = \mathrm{R}_i + jX_i$ (Fig. 3). Whether, or to what extent, the highest attainable input value agrees with the highest attainable sensitivity, is shown for a single circuit in Fränz's paper (*loc.cit.*). For band-filter input circuits this question will be calculated in a further work" [presumably that dealt with in 49 of January].

The present paper is divided into two main parts. The first considers input arrangements comprising a single circuit only: its first section deals with the case where the active component R_i of the "generator" internal resistance is not small compared with the resonance resistance of the input circuit, when the reactive component of the generator internal resistance is (a) capacitive, (b) inductive, and (c) zero (for example, a quarter-wave aerial). Its second section (case where the active component R_i is small compared with the resonance resistance of the input circuit) considers only arrangements with capacitive reactive component of the generator internal resistance, since the first section showed that the results for this case could be applied directly to cases where the reactive component was inductive or zero: in fact, eqn. 19 for the optimum value of $a(\doteq E_g/E_a)$ applies in all three cases a, b, and c: it is $|a|_{opt} = \frac{1}{2}\sqrt{R_p/R_i}$, on certain assumptions which are justifiable on most occasions. Here R_p is the parallel resistance into which the losses of the oscillatory circuit are considered as concentrated.

The second part of the paper deals with input arrangements comprising a two-circuit band filter. As before, the first section considers the case when R_i is not small compared with the resonance resistance of (here) the individual circuits of the filter. Eqn. 48 now replaces the simpler eqn. 19 for the optimum value of a: for a given R_i this depends not only on R_p as before but also on k'/d, that is, on the coupling of the band-filter circuits. If $k'/d \gg 1$, eqn. 19 is obtained again, but in practice a compromise between input value and selectivity is chosen: for instance, if $k'/d = 1$, then $|a|_{opt} = 1/\sqrt{2} \times \frac{1}{2}\sqrt{R_p/R_i}$, smaller by the factor $1/\sqrt{2}$ than for the single circuit of eqn. 19. In the case dealt with in the second section, where R_i is small compared with the resonance resistance of the band-filter circuits, the result is as follows:—for an input consisting of a two-circuit band filter, with a generator whose internal resistance is purely reactive, there is (with respect to the coupling of the two filter circuits) an optimum input value (the term "input value" here again represents a, the ratio E_g/E_a) which is attained when these circuits are critically coupled. With respect to the coupling k between the generator and the input circuit, there is no optimum value of a: this ratio increases linearly with k and reaches its maximum for $C_1 = 0$, that is, when the first filter circuit is tuned entirely by C_k. This maximum value is $1/2d$, half as large as with a single circuit as input (eqn. 36).

The paper ends with examples in the design calculations of the 15–60 m arrangements of a broadcast receiver to fulfil certain requirements: a single-circuit input is dealt with first, eqn. 16 for $|a|$ being used: finally a critically coupled band filter is considered as input, table 2 being calculated from eqns. 46/48.

1078. VALVE NOISE, A PHYSICAL LIMIT IN THE DESIGN OF COMMUNICATIONS APPARATUS.—Hameister. (*See* 1104.)

1079. CORRECTION TO DISCUSSION ON "THE DISTRIBUTION OF AMPLITUDE WITH TIME IN FLUCTUATION NOISE."—Landon. (*See* 1103.)

1080. DIODE AS A FREQUENCY-CHANGER.—F. M. Colebrook & G. H. Aston. (*Wireless Engineer*, Jan. 1943, Vol. 20, No. 232, pp. 5–14.)

The behaviour of the diode as a frequency-changer is analysed with particular reference to the so-called "television" type. The dependence of the conversion ratio on circuit conditions and the decrease in conversion ratio involved in the use of harmonics of the diode-current pulses are investigated. Values are found for the effective input resistances at signal and oscillator frequencies and the internal resistance at beat frequency.

1081. DIODE FREQUENCY-CHANGERS.—E. G. James & J. E. Houldin. (*Wireless Engineer*, Jan. 1943, Vol. 20, No. 232, pp. 15–27.)

A general analysis of a two-element frequency-changer is given and the results are applied to diodes having (1) a three-halves power characteristic and (2) an exponential characteristic. The optimum ratio of output power to input power is calculated both for negligible circuit losses and for an input circuit with appreciable loss.

1082. THE SENSITIVITY OF AERIALS TO LOCAL INTERFERENCE.—Cornelius. (*See* 1095.)

1083. VIBRATOR CIRCUIT [to Provide High Tension for Noise-Free Operation of Sensitive Recivers at All Frequencies].—Masteradio.

(*Journ. of Scient. Instr.*, Jan. 1943, Vol. 20, No. 1, p. 18.) For previous work *see* 2094 of 1942.

1084. NOTES ON TRACKING CIRCUITS [including Tracking with Constant Difference of Wavelength].—A. Bloch. (*Wireless Engineer*, Nov. 1942, Vol. 19, No. 230, pp. 508–514.)

1085. SUPERHETERODYNE TRACKING SIMPLIFIED [Chart for determining Series and Shunt Capacities for Tuned Circuits in Single-Control Superheterodyne Receivers].—P. C. Gardiner. (*Electronics*, Nov. 1942, Vol. 15, No. 11, pp. 74–75.)

1086. AN OSCILLATOR FOR REMOTE FREQUENCY CONTROL.—Lawrence. (*See* 1029.)

1087. VALVE REPLACEMENT IN PUSH-PULL STAGES [now that Matched Replacement is Difficult: Ways of obtaining Least Possible Distortion, by Circuit Arrangements].—F. Kunze. (*Physik. Berichte*, 15th June 1942, Vol. 23, No. 12, p. 1268: summary, from *Funkschau*, No. 4, Vol. 15, 1942, p. 49 onwards.)

1088. STANDARDS FOR REPLACEMENT PARTS FOR CIVILIAN RADIO IN WAR TIME [American Standards Association Project].—O. H. Caldwell. (*Industrial Standardisation*, Dec. 1942, Vol. 13, No. 11, pp. 312–313.)

1089. TROPICAL RECEIVER DESIGN [Design of Receiver particularly for Use in India].—M. J. H. Lemmon. (*Journ. I.E.E.*, Part I, July 1942, Vol. 89, No. 19, pp. 321–322.) *See* also 1348 of 1942.

1090. AIRCRAFT RADIO COMPASS AND COMMUNICATION RECEIVER.—Green & Walton. (*See* 1121.)

AERIALS AND AERIAL SYSTEMS

1091. CIRCULAR ANTENNAE [for Ultra-Short Waves].—M. W. Scheldorf & L. M. Leeds. (*Communications*, July 1942, Vol. 22, No. 7, pp. 36–37.)

In this antenna the upper and lower circular elements are each effectively a quarter-wavelength long, being shortened from a physical quarter-wavelength by an adjustable end capacity. It provides simple horizontal polarisation with only two terminals and has uniform horizontal radiating properties. The gain characteristic shows an improvement, due to large spacing, over systems which require a closer spacing for the highest gain. *See* also 762 of March.

1092. LECHER SYSTEMS [including Use as Feeders].—von Lindern & de Vries. (*See* 1178.)

1093. ON RADIATION FROM ANTENNAS.—S. A. Schelkunoff & C. B. Feldman. (*Proc. I.R.E.*, Nov. 1942, Vol. 30, No. 11, pp. 511–516.)

Authors' summary:—"This paper presents some theoretical remarks and experimental data relating to application of transmission-line theory to antennas. It is emphasised that the voltage, current, and the charge are affected by radiation in different ways, a fact which should be considered in any adaptation of line equations to antennas.

"It is shown experimentally and theoretically that in an antenna of length equal to an integral number of half-wavelengths, which is energised at a current antinode, the effect of radiation on the current and the charge (but not on the voltage) can roughly be represented by adding to the resistance of the wires another fairly simple term."

1094. ENERGY AND ENERGY FLOW IN THE ELECTROMAGNETIC FIELD.—Slepian. (*See* 1311.)

1095. THE SENSITIVITY OF AERIALS TO LOCAL INTERFERENCE [Advantages of Frame Aerials over "Indoor" & "Mains" Aerials: Classification of Interference Sources as Electric & Magnetic Radiators, and of Receiving Aerials as Capacitive & Inductive: Action of Various Types of Source on Various Types of Aerial].—P. Cornelius. (*Physik. Berichte*, 15th June 1942, Vol. 23, No. 12, pp. 1264–1265: summary, from *Philips Tech. Rundschau*, No. 10, Vol. 6, 1941, p. 307 onwards.)

1096. SMALL-DIMENSIONAL RADIATING AND RECEIVING SYSTEMS OF HIGH DIRECTIVITY.—Makov. (*See* 1134.)

VALVES AND THERMIONICS

1097. THE PROPERTIES OF VALVES AS ADMITTANCES IN THE ULTRA-SHORT-WAVE REGION, AND TRANSIT-TIME PHENOMENA.—E. Hameister. (*Zeitschr. f. Fernmeldetech.*, 17th Oct. 1942, Vol. 23, No. 10, pp. 145–151.)

Author's summary:—"The properties of valves in the ordinary wave regions are outlined. After a short theoretical treatment of the space-charge region, forming the foundations for the calculation of valve properties, the properties are presented as admittances, from the viewpoint of four-terminal-network theory, for the ultra-short-wave region also; the properties and behaviour in the régime of transit-time phenomena are considered, and the problems of the setting-up of oscillations are discussed. Particular attention is given, all through the paper, to bringing out clearly the methods of calculation."

1098. SOME TECHNOLOGICAL PROBLEMS IN THE DEVELOPMENT OF A NEW SERIES OF TRANSMITTING VALVES [Ultra-Short-Wave Triodes & Pentodes, for 5–7 m Band].—E. G. Dorgelo. (*Philips Tech. Rundschau*, No. 9, Vol. 6, 1941, p. 257 onwards.)

The various considerations which guided the design of this series of valves are discussed: the object in view was to satisfy recent demands for u.s.w. valves of increased output with efficiencies comparable with those of medium-wave valves, and the constructional points necessary to attain this object are described. Molybdenum is used exclusively for the grids and anodes, on account of its high heat-resisting powers. All joints are made either with molybdenum rivets or else by direct welding. Thoriated tungsten is used for the cathodes, with a specific emission of 70 mA/w compared with the 6 mA/w of plain tungsten. A newly developed type of glass is employed for the bulbs, the sealing technique being borrowed from the field of quartz lamp manufacture. This matter of seals, and also the oxide-free leading-in connections of tungsten and molybdenum (which remain vacuum-tight even at working temperatures of 400° C.), are specially mentioned. A zirconium getter in powder form is used, and secondary emission is avoided by zirconising the electrodes.

1099. VACUUM MEASUREMENT IN A MAGNETRON BY MEANS OF THE IONIC CURRENT.—E. Djakoff & A. Rajeff. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1327: summary, from a Bulgarian publication.)

In a magnetron deprived of its magnetic field there is a proportional relation between ionic current on the one hand and gas-pressure and electronic current on the other, as there is in triodes. In a two-slit magnetron the ionic current was measured with one half-anode as positive electrode and the other as negative. With $U_+ = 250$ v and $U_- = 2$ to 4 v, the ionic current reached a maximum, and values of 8.8 and 7.6 were arrived at for the residual-gas constant C_0 for two-slit and four-slit magnetrons respectively. Accuracy of measurement was within 5%.

1100. THE DETERMINATION OF THE RESIDUAL-GAS PRESSURE IN COMMERCIAL VALVES BY GRID-CURRENT MEASUREMENT.—G. Herrmann & O. Krieg. (*Telefunken-Röhre*, No. 21/22, 1941, p. 219 onwards.)

Amplifying valves of various types were connected to a pumping system and filled with carbon monoxide or oxygen at measured pressures. Keeping the anode current constant, measurements were made of the ionic current flowing to the negatively biased grid (in multi-grid valves the grids were connected together in parallel, so as to take all the ions formed into account). A filament-heating voltage raised by 25% above the normal value served to correct for the cooling of the cathode by the gas, and also for a small amount of poisoning of the cathode. Comparison between the results thus obtained and those given by mass-produced valves led to the conclusion that the average residual-gas pressure in the latter was 10^{-5} mm, sinking to 10^{-6} or 10^{-7} mm, or even lower, during service: values which agree with estimates previously made.

1101. DIODE AS A FREQUENCY-CHANGER, *and* DIODE FREQUENCY-CHANGERS.—Colebrook & Aston: James & Houldin. (*See* 1080 & 1081.)

1102. WHAT QUANTITIES REPRESENT THE SUITABILITY OF A VALVE FOR THE AMPLIFICATION OF THE SMALLEST SIGNALS?—W. Engbert: M. J. O. Strutt & A. van der Ziel. (*Physica*, No. 8, Vol. 8, 1941, p. 903 onwards.)

From the Telefunken laboratories. In reply to a remark in Strutt & van der Ziel's paper (*ibid.*, Vol. 8, p. 424) on the definition of a number to represent the noise-resistance of a valve, Engbert maintains that in valve questions in general it is particularly desirable to give valve magnitudes which are not dependent on circuit effects.

1103. CORRECTION TO DISCUSSION ON "THE DISTRIBUTION OF AMPLITUDE WITH TIME IN FLUCTUATION NOISE."—V. D. Landon. (*Proc. I.R.E.*, Nov. 1942, Vol. 30, No. 11, p. 526.) *See* 2129 of 1941 and 389 of February.

1104. VALVE NOISE, A PHYSICAL LIMIT IN THE DESIGN OF COMMUNICATIONS APPARATUS [Survey for the Practical Man, based on Work of Schottky, Spenke, & others].—E. Hameister. (*Zeitschr. f. Fernmeldetech.*, 15th April 1942, Vol. 23, No. 4, pp. 59–62.) Another work referred to in the list at the end is that of Borgnis, 694 of 1942.

1105. DESIGNING A RESISTANCE-LOADED PUSH-PULL INVERTER.—Feinberg. (*See* 1024.)

1106. VALVE REPLACEMENT IN PUSH-PULL STAGES [now that Matched Replacement is Difficult].—Kunze. (*See* 1087.)

1107. A GRAPHICAL METHOD TO FIND THE OPTIMAL OPERATING CONDITIONS OF TRIODES AS CLASS C TELEGRAPH TRANSMITTERS.—J. C. Frommer. (*Proc. I.R.E.*, Nov. 1942, Vol. 30, No. 11, pp. 519–525.)

1108. METHODS OF MEASURING THE SLOPE OF AMPLIFIER VALVES [Review of Methods due to Appleton, Eccles, Schottky, Graffunder, Ballantine, Everitt, & Steimel].—J. Bärisch. (*Zeitschr. f. Fernmeldetech.*, 16th Feb. 1942, Vol. 23, No. 2, pp. 37–41.)

1109. TECHNOLOGICAL POINTS IN THE DESIGN OF RADIO VALVES [including Various Seal-Making Techniques, the Pros & Cons of Metal & Glass Envelopes, and a Description of the Development of the Philips "Key" Valve ("Schlüsselröhre")].—Th. P. Tromp. (*Philips Tech. Rundschau*, No. 11, Vol. 6, 1941, p. 321 onwards.)

1110. IRON-TO-GLASS SEALS [for Vacuum Tubes].—(*Electronics*, Aug. 1942, Vol. 15, No. 8, p. 97.)

1111. SOME ASPECTS OF RADIO VALVE MANUFACTURE [General Description of Process].—T. F. B. Hall & A. H. Howe. (*Electronic Eng.g*, Sept. 1942, Vol. 15, No. 175, pp. 140–146.)

1112. RECENT R.C.A. VALVES: TENTATIVE DATA.—(*Electronic Eng.g*, Nov. 1942, Vol. 15, No. 177, pp. 254–255.)

1113. RECEIVING–TRANSMITTING TUBE MANUALS [Notice of R.C.A. 1942 Handbooks].—(*Communications*, Sept. 1942, Vol. 22, No. 9, p. 39.)

1114. "INTRODUCTION TO VALVES" [Book Review].—F. E. Henderson. (*Elec. Review*, 19th Feb. 1943, Vol. 132, No. 3404, p. 262.)

1115. OPERATION OF A THYRATRON AS A RECTIFIER [Theoretical and Experimental Treatment of Half-Wave Thyratron Rectifier Circuit taking into account Difference between

Firing Potential and Tube Drop during Conduction].—L. A. Ware. (*Proc. I.R.E.*, Nov. 1942, Vol. 30, No. 11, pp. 500–502.)

1116. MEASUREMENT OF THE ANODE-CATHODE VOLTAGE DROP OF AN A.C. DISCHARGE TUBE [Method of measuring Voltage between Anode and Cathode of Alternating-Current Discharge Tube during Interval when Current is Flowing].—R. H. Baulk. (*Electronic Eng:g*, Jan. 1943, Vol. 15, No. 179, pp. 324–326.)

1117. THE MANUFACTURE AND PROPERTIES OF TUNGSTEN AND MOLYBDENUM.—G. A. Percival. (*Electronic Eng:g*, Aug. 1942, Vol. 15, No. 174, pp. 104–108.)

1118. THE USE OF SECONDARY ELECTRON EMISSION TO OBTAIN TRIGGER OR RELAY ACTION.—A. M. Skellett. (*Journ. Applied Phys.*, Aug. 1942, Vol. 13, No. 8, pp. 519–523.) A summary was dealt with in 442 of February.

1119. OBSERVATIONS ON THE SECONDARY-ELECTRON RADIATION FROM NON-CONDUCTORS [by a Method applicable also to Photoelectron Measurements].—K. H. Geyer. (*Ann. der Physik*, 21st Dec. 1942, Vol. 42, No. 4, pp. 241–253.)

Previous measurements of the secondary emission from non-conductors have been obtained almost exclusively with thin films (Bruining & de Boer, 4456 of 1939): by such methods it is impossible to separate the surface properties from the volume properties, particularly as films of material undergo transformations in their structure as their thickness is increased (*see* 1974 of 1942, a previous paper by the present writer). The work now described contributes to our knowledge of the secondary radiation from non-conductors as a volume property of the latter. Also, the usual method of measuring the s.e. coefficient δ has in the past been applied to non-conductors without regard to the fact that in this case the s.e. current which can leave the body is *not* controlled by the p.d. with respect to the surrounding cage (forming the counter-electrode) but by a potential minimum which is formed in front of the non-conductor as a result of the space charge arising from the secondary electrons already emitted. A discontinuity in the angular dependence of the s.e. output from non-conducting bodies, due to this space-charge mechanism, has already been pointed out by Salow (3432 of 1941).

The writer therefore uses an apparatus (Fig. 1) in which the difference between the incident primary electrons and the released secondary electrons is obtained from the displacement current through the non-conducting plate from its backing electrode: dependence on the displacement current makes it necessary to use an oscillograph to observe the instantaneous values. The method used is an improvement of that employed previously by the writer & Heimann (2594 of 1940) and also by Salow, in that instead of velocity-modulation or intensity-modulation of the primary ray (which was found to lead to experimental troubles) it is the collector voltage, the "draining-off" voltage on the dual collector-electrode system Z, M, that is modulated: the method might be termed the "intermittent collector-voltage" method, but because of its fundamental idea it is better called the "intermittent displacement-current" method. It may well be found useful for the measurement of other electron-emission effects from non-conductors, their photoelectric behaviour for example.

The present investigations lead to the following conclusions:—alkaline-earth oxides, Al_2O_3, and colourless glasses have values of δ from 2 to 3: alkali halides rather more, about 3 to 4. Inclusions of foreign-metal atoms produce, in crystalline materials as well as in glasses, an increase of secondary emission up to about twice the original value, provided that they cause no conductivity. The variation with thickness of the s.e. output from thin films shows the presence of maxima, indicating the participation of electrons released by field action from the metal carrier. Thus it is deduced finally that the high output of commercial artificial films (depending often on a definite thickness) is due to co-operation by a non-self-sustaining discharge: it is a surface effect rather than a volume effect.

DIRECTIONAL WIRELESS

1120. ULTRA-HIGH-FREQUENCY RADIO-RANGE AND MARKER RECEIVERS FOR AIRCRAFT.—G. Builder & J. G. Downes. (*A.W.A. Tech. Review*, No. 1, Vol. 6, 1942, pp. 1–15.)

Description of a receiver and associated equipment for use in aircraft to give visual and aural reception of radio-range and marker-beacon signals in the Australian system of u.h.f. (33.3–34.2 and 38.0 Mc/s) navigation aids.

1121. AIRCRAFT RADIO COMPASS AND COMMUNICATION RECEIVER.—A. L. Green & J. G. Walton. (*A.W.A. Tech. Review*, No. 1, Vol. 6, 1942, pp. 17–35.)

A description of a complete aircraft communication receiver with frequency range 275–1700 kc/s, 2.3–3.3 Mc/s, and 6–7 Mc/s, with facilities for both aural and visual indications of homing and direction finding in the first band, and complete remote control of both communication and direction-finding facilities. An explanation of the operation of the radio compass is given in which the loop system is considered as a suppressed-carrier modulator with the carrier re-supplied at a later stage by the fixed aerial. The theory leads to the prediction and achievement of sharper bearings and elimination of overload of the left/right visual indicator.

1122. THEORY OF REVERSED HOMING.—A. L. Green. (*A.W.A. Tech. Review*, No. 1, Vol. 6, 1942, pp. 43–58.)

The radio compass is considered as a navigational aid when flying *away* from a transmitter. The theory indicates the conditions to be observed in order to restrict the errors in flight.

1123. AIRCRAFT COMMUNICATIONS [Description of Bendix Communication Unit No. 508].—C. W. McKee. (*Communications*, Sept. 1942, Vol. 22, No. 9, pp. 12–15 and 30, 34.)

The first of a series of analyses of aircraft communication equipment and components.

1124. Typical Equipment in Aircraft Communications [General Description].—C. W. McKee. (*Communications*, Oct. 1942, Vol. 22, No. 10, pp. 12–13 and 47, 48.)

ACOUSTICS AND AUDIO-FREQUENCIES

1125. Loudspeaker Diaphragms composed Chiefly or Wholly of Mica.—A. Kübel. (*Zeitschr. f. Fernmeldetech.*, 16th Feb. 1942, Vol. 23, No. 2, p. 29.) Telefunken patent, D.R.P. 705 441: especially good for low notes.

1126. Loudspeakers and Frequency-Modulation Distortion.—G. L. Beers & H. Belar. (*Communications*, July 1942, Vol. 22, No. 7, pp. 18 and 20.)

See also 800 of March. Mathematical analysis and measurements indicate the possibility of frequency-modulation distortion in loudspeakers when reproducing a complex wave. To reduce f.m. distortion four methods are suggested: (1) reduction of amplitude by loading with horn, (2) reduction of amplitude by increasing cone diameter, (3) limitation of power input at low frequencies, and (4) use of separate speakers for low and high frequencies.

1127. A Distortion Meter [Frequency Range 30 to 10 000 c/s].—J. E. Hayes. (*Communications*, July 1942, Vol. 22, No. 7, pp. 12–13.) *See* also 798 of March.

1128. Harmonic Distortion in Audio-Frequency Transformers: Part 3 [Influence in Practice of Facts set out in Parts 1 and 2].—N. Partridge. (*Wireless Engineer*, Nov. 1942, Vol. 19, No. 230, pp. 503–507.) For Parts 1 and 2 *see* 3288 of 1942.

1129. "Audio-Transformer Distortion" [Review of 18-page Booklet].—N. Partridge. (*Elec. Review*, 8th Jan. 1943, Vol. 132, No. 3398, p. 58.) *See* also 3288/9 of 1942, and 1128, above.

1130. Radio Loudspeaker Tests Pave Way for Better Reception.—A. N. Goldsmith. (*Industrial Standardisation*, Dec. 1942, Vol. 13, No. 11, pp. 306–308.)

1131. Audio-Frequency Compensating Circuits [to make Sound Reproduction seem More Realistic].—S. Cutler. (*Electronics*, Sept. 1942, Vol. 15, No. 9, pp. 63 and 66..70.)

1132. Sound Reproduction [Summary of Paper read to British Instn. Rad. Eng: Results of Experience as Technical Adviser to Central Council for School Broadcasting: Reliability of Subjective Estimates of Quality].—L. E. C. Hughes. (*Elec. Review*, 11th Dec. 1942, Vol. 131, No. 3394, pp. 749–750.) *See* also 452 of February.

1133. Public Address Systems [History and Development of Public Address Systems traced from First Appearance to the Present Day].—S. Hill. (*Journ. I.E.E.*, Part III, Sept. 1942, Vol. 89, No. 7, pp. 124–134.)

1134. Small-Dimensional Radiating and Receiving Systems of High Directivity.—S. A. Makov. (*Physik. Berichte*, 1st Aug. 1942, Vol. 23, No. 15, p. 1510: summary of paper from the Moscow Acoustic Division.)

"If two systems, each possessing only a weakly directional characteristic, are used together in phase-opposition, the combination has a much stronger directional effect. Such combinations have the advantages of preserving their directional characteristic over a wide range of frequencies and of occupying a comparatively small space. Some examples are given: a straight-line system with radiation at right angles to it, a similar system with parallel beam, and small two- and three-dimensional systems with strongly directional action. The treatment includes also the acoustic problems."

1135. A Frequency-Modulated Control-Track for Movietone Prints [located between Sound and Picture Areas: proposed for Control of Intensity Level of Reproduced Sound].—J. G. Frayne & F. P. Herrnfeld. (*Bell S. Tech. Journ.*, June 1942, Vol. 21, No. 1, p. 75–76: abstract only.) *See* also 454 of February.

1136. A Discussion of Several Factors contributing to Good Recording.—R. A. Lynn. (*R.C.A. Review*, April 1942, Vol. 6, No. 4, pp. 463–472.)

1137. A Study of "Wows" [Equipment to determine "Wow" Content, *i.e.* Effects due to Speed Fluctuations of a Reproducing Turntable: Signal from Magnetic Tone Wheel applied across Tuned Circuits].—H. E. Roys. (*Communications*, July 1942, Vol. 22, No. 7, pp. 42–43.) *Cf.* 4148 of 1937 and 1947 of 1939.

1138. Construction, Mode of Action, and Properties of the Crystal Pick-Up.—P. Beerwald & H. Keller. (*Funktech. Monatshefte*, July 1942, No. 7, pp. 97–99.) *See* 75 of January for a longer paper.

1139. A Piezoelectric Device of Very Great Sensitivity [using a Very Thin Crystal Diaphragm fixed at the Edges and slightly Bulged].—S. Szalay. (*Zeitschr. f. Fernmeldetech.*, 16th Jan. 1942, Vol. 23, No. 1, p. 12.) D.R.P. 704 649.

1140. The Mass Ratio between Membrane and Fluid in the Inner Ear [and the Question of Resonance].—O. F. Ranke. (*Akust. Zeitschr.*, No. 1, Vol. 7, 1942, pp. 1–11.)

1141. Subjective Determination of the Quality of Telephone Systems: II—Statistical Observations on Queries [Requests for Repetition, etc.]: the Technical Procedure.—H. Panzerbieter & A. Rechten. (*Arch. f. Tech. Messen*, Sept. 1942, Part 135, V3719-2, Sheets T90–91.)

1142. THE FUTURE OF TRANSOCEANIC TELEPHONY [Thirty-Third Kelvin Lecture].—O. E. Buckley. (*Journ. I.E.E.*, Part I, Oct. 1942, Vol. 89, No. 22, pp. 454–461 : *Bell S. Tech. Journ.*, June 1942, Vol. 21, No. 1, pp. 1–19.) Summaries have been dealt with in 1995 [*see* also 3282] of 1942.

1143. LOW-FREQUENCY CHARACTERISTICS OF THE COUPLING CIRCUITS OF SINGLE AND MULTI-STAGE VIDEO AMPLIFIERS.—Donley & Epstein. (*See* 1169.)

1144. THE MEASUREMENT OF VOLTAGE PEAKS IN A STUDIO EQUIPMENT [Philips Peak-Indicator for Broadcast Programme Control].—de Fremery & Wenk. (*See* 1297.)

1145. A MECHANICAL OSCILLATOR FOR RECORDING COMPOUNDED LINEAR WAVES [a Teaching Device for Demonstrating Various Wave Phenomena].—A. D. Bulman. (*Journ. of Scient. Instr.*, Jan. 1943, Vol. 20, No. 1, pp. 10–12.)

1146. WAVE ANALYSIS [Part I—General Review: Part II—Analysis of Semi-Periodic Wave-Forms].—Bourne. (*See* 1320.)

1147. 36 and 72 ORDINATE SCHEDULES FOR GENERAL HARMONIC ANALYSIS [to facilitate Calculations: Applicable to Odd and Even Harmonics].—R. P. G. Denman. (*Electronics*, Sept. 1942, Vol. 15, No. 9, pp. 44–47.)

1148. SOUND.—W. Mikelson. (*Gen. Elec. Review*, Dec. 1942, Vol. 15, No. 12, pp. 685–694.) A survey of the technique of the measurement and analysis of sound, and a useful discussion of terminology.

1149. AN ANALYSIS OF AUDIO-FREQUENCY RESPONSE CHARTS [with Charts giving Insertion Loss due to Series and Parallel Reactances].—H. Holubow. (*Communications*, Oct. 1942, Vol. 22, No. 10, pp. 5–7.)

1150. A SELECTIVE CIRCUIT AND FREQUENCY METER USING A TUNING FORK.—Tucker. (*See* 1036.)

1151. THE DETERMINATION OF AN UNKNOWN FREQUENCY FROM A PHOTOGRAPHIC RECORD.—M. Scott. (*Electronic Eng:g*, Nov. 1942, Vol. 15, No. 177, p. 243.)

1152. SOUND INSULATION [Reduction of Noise in Buildings by Planning and Improved Structural Technique].—W. Allen. (*Journ. Roy. Soc. Arts*, 5th Feb. 1943, Vol. 91, No. 4632, pp. 135–147.)

1153. THE NEW D.I.N. NOISEMETER [based on German Acoustics Committee's Recommendations].—H. Kalden. (*Zeitschr. f. Fernmeldetech.*, 16th Jan. 1942, Vol. 23, No. 1, pp. 10–12.) For a correction *see* issue for 16th February, No. 2, p. 32.

1154. THE NOISE REDUCTION IN PRECISION-MADE GEARS [as in Film Cameras, Alarm Clocks, etc.].—R. Berger. (*Akust. Zeitschr.*, No. 1, Vol. 7, 1942, pp. 18–29.)

1155. RECORDING MACHINERY-NOISE CHARACTERISTICS.—Brailsford. (*See* 1356.)

1156. AN ACOUSTIC METHOD FOR DETERMINING THE DYNAMIC COMPRESSIBILITY AND LOSS FACTOR OF ELASTIC MATERIALS.—Meyer & Tamm. (*See* 1359.)

1157. SOME EXPERIMENTS WITH SOUND WAVES GENERATED BY PERIODIC SUCKING IMPULSES [in Powdered Substances].—F. Bruns. (*Akust. Zeitschr.*, No. 1, Vol. 7, 1942, pp. 29–32.)

1158. THE APPLICATION OF SOUND WAVES IN METALLURGY.—Becker. (*See* 1357.)

1159. RECENT RESEARCHES ON THE PHYSICO-CHEMICAL AND SIMILAR EFFECTS PRODUCED BY SUPERSONIC WAVES.—Lovera. (*See* 1358.)

1160. ON THE BIREFRINGENCE PRODUCED IN LIQUIDS BY SUPERSONIC WAVES [Experimental Investigation].—S. Petralia. (*Physik. Berichte*, 15th July 1942, Vol. 23, No. 14, p. 1440.)

PHOTOTELEGRAPHY AND TELEVISION

1161. COLOUR TELEVISION.—Goldmark, Dyer, Piore, & Hollywood. (*Journ. Applied Phys.*, Nov. 1942, Vol. 13, No. 11, pp. 666–677.) Refers to more detailed paper by the same authors—*see* 481 of February. For another condensed version see *Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, pp. 195–200 and 213.

1162. COLOUR TELEVISION DEVELOPMENT [Receiver without Moving Parts].—(*Wireless World*, Feb. 1943, Vol. 49, No. 2, p. 41.) *See* also 830 of March.

1163. COLOUR AND STEREOSCOPIC TELEVISION [Editorial].—(*Electronic Eng:g*, Aug. 1942, Vol. 15, No. 174, pp. 96–97.)

1164. TELEVISION RECEPTION WITH BUILT-IN ANTENNAS FOR HORIZONTALLY AND VERTICALLY POLARISED WAVES.—W. L. Carlson. (*R.C.A. Review*, April 1942, Vol. 6, No. 4, pp. 443–454.)

1165. THEORETICAL CONSIDERATIONS ON A NEW METHOD OF LARGE-SCREEN TELEVISION PROJECTION: PART IV [the Heat Production by Electron Bombardment].—F. Fischer, H. Thiemann, & E. Baldinger. (*Schweizer Arch. f. angew. Wiss. u. Tech.*, Oct. 1942, Vol. 8, No. 10, pp. 299–307.)

In Parts I & II (for all previous parts *see* 3308 of 1942 and back reference) the theory of the "eidophor" was limited to the immediate problem of image production by surface deformation: solid and liquid eidophors were considered, and among other things the necessary ray-current densities were calculated. Part III considered the concomitant effects in liquid eidophors, in order to arrive at the conditions necessary for the production of faultless images: means and ways were given for the realisation of these conditions so that disturbances of the eidophor surface might be reduced to negligible proportions. But no attention was paid to possible effects of electron bombardment on the surface. These might be chemical (polymerisation, etc: footnote "3" gives some references) or thermal. The possibility of the first type occurring can only be decided experimentally, but the thermal problem can be dealt with by calculation. This is done in the present part, and it is concluded that for a current density of 3×10^{-2} A/cm^2 the temperature rises produced in the course of normal working would be somewhere round the harmless value of 0.3°. Larger rises would occur if the line-frequency deflecting unit fell out of action, but even these would not be dangerous to the surface. Provision would have to be made, however, to switch off the ray at once if *both* deflecting systems failed simultaneously.

1166. Television Transmission on High Power [Description of Transmitter].—H. B. Fancher. (*Communications*, July 1942, Vol. 22, No. 7, pp. 20, 30, and 34.) *See* also 826 of March.

1167. Television Studio Lighting [at General Electric Company's Studios, Schenectady].—C. A. Breeding. (*Communications*, July 1942, Vol. 22, No. 7, p. 36: paragraph only.) *See* also 827 of March.

1168. Removable Cathode-Ray-Tube Screens [for Replacement when Worn Out].—J. L. Baird. (*Electronic Eng:g*, Aug. 1942, Vol. 15, No. 174, p. 128.) Brit. Patent No. 544 413.

1169. Low-Frequency Characteristics of the Coupling Circuits of Single and Multi-Stage Video Amplifiers.—H. Donley & D. W. Epstein. (*R.C.A. Review*, April 1942, Vol. 6, No. 4, pp. 416–433.)

The ordinary resistance-coupled amplifier does not faithfully reproduce frequencies below about 200 c/s. The form of compensation usually employed to improve the low-frequency response is a resistance–capacity anode filter circuit. Curves are given for multi-stage as well as single-stage amplifiers showing the effect of changes in the circuit constants on the transient and steady-state characteristics. These facilitate the choice of circuit constants to meet a specific low-frequency requirement. The curves are of value in the design of audio resistance-coupled amplifiers as well as video amplifiers.

1170. A Photocell with Amplification by Secondary Emission.—M. C. Teves. (*Zeitschr. f. Fernmeldetech.*, 16th Jan. 1942, Vol. 23, No. 1, pp. 13–14: summary only.) An Italian summary was dealt with in 2765 of 1941.

1171. Observations on the Secondary-Electron Radiation from Non-Conductors [by a Method applicable also to Photoelectron Measurements].—Geyer. (*See* 1119.)

1172. Semiconductor Photoelements and Rectifiers [Experimental Investigation of Cuprous-Oxide Cells].—C. G. Fink & E. Adler. (*Physik. Berichte*, 15th July 1942, Vol. 23, No. 14, p. 1420: from *Trans. Electrochemical Soc.*, Preprint 1, 1941, Vol. 79.)

By preparing the Cu_2O layers not thermally, as usual, but electrolytically, it was possible to obtain very thin films of exactly controllable thickness and thus to have reproducible specimens. "The rectification and photoelectric e.m.f. of Cu_2O/metal cells is independent of the nature of the carrier metal (not necessarily copper): this is in contradiction to previous observations by other writers. The writers conclude that rectification and photo-effect are not bound up with the presence of an insulating barrier layer. A method of preparing homogeneous oxide films of 10^{-5} to 10^{-2} cm thickness is given. The optical transparency of such oxide films in the near infra-red is measured."

1173. Contribution to Our Knowledge of the Variation of Transparency of Colloidal Iron in a Magnetic Field [and an Inversion analogous to That of the Double Refraction & Dichroism in a Magnetic Field].—G. Frongia & M. Agus. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1327: summary of an Italian paper.)

MEASUREMENTS AND STANDARDS

1174. Hertzian-Wave Spectroscopy with a Magnetron Oscillator [of Stabilised Frequency: Drude's Second Method for measuring Dispersion & Absorption].—Cavallaro. (*See* 1353.)

1175. Anomalous Dispersion of Dipolar Ions [and "a New Method of Measuring the Dielectric Constant & Dielectric Absorption" at Ultra-High Frequencies].—Marcy & Wyman. (*See* 1354.)

1176. Application of Wave-Guides to the Measurement of Dielectric Constants in the Region of Centimetric Waves.—G. Fejér & P. Scherrer. (*Physik. Berichte*, 1st Aug. 1942, Vol. 23, No. 15, pp. 1511–1512.)

Summary of the paper referred to in 3436 of 1941. Wavelengths down to 1 cm are obtained from a magnetron with anode diameter 0.4 mm and length about 4 mm, whose connections form a plane three-wire system tuned by a reflecting disc. A cylindrical wave guide is used to measure the dielectric constant of a solid material in the form of a plate laid on one end-disc of the cylinder. The $H_{o,m}$ wave is specially suitable, owing to its electrical lines of force being parallel to the axis. The formulae for calculating the dielectric constants are given in the actual paper.

1177. The Measurement of Decimetric, Centimetric, and Millimetric Waves [Wavelength Measurement using Resonant Concentric Lines and Resonant Wave Guides].—A. G. Clavier. (*Communications*, Sept. 1942, Vol. 22, No. 9, pp. 5–9 and 19, 35, 36.) *See* also 489 of February.

1178. LECHER SYSTEMS.—C. G. A. von Lindern & G. de Vries. (*Philips Tech. Rundschau*, No. 8, Vol. 6, 1941, p. 241 onwards.)

The behaviour of Lecher systems can be expressed to a certain extent in quasi-stationary terms, although the Lecher system does not represent a quasi-stationary structure. The writers calculate the capacitance, inductance, propagation velocity, and characteristic impedance, and investigate the behaviour to progressive and standing waves: practical conclusions are deduced. Such systems are then considered as feeders for transmitting aerials and as frequency-stabilising elements for oscillatory circuits. Various practical forms are illustrated by photographs.

1179. RECEIVER INPUT CONNECTIONS FOR ULTRA-HIGH-FREQUENCY MEASUREMENTS [Three Methods of obtaining Push-Pull Output Voltage from Signal Generator with Single-Sided Output].—J. A. Rankin. (*R.C.A. Review*, April 1942, Vol. 6, No. 4, pp. 473–481.)

1180. RADIO-FREQUENCY POWER MEASUREMENT [Convenient Bolometric Method].—W. Bacon. (*Electronic Eng:g*, Jan. 1943, Vol. 15, No. 179, pp. 344–345.)

1181. RADIO-CIRCUIT STABILITY.—Eggleton. (*See* 1031.)

1182. QUARTZ CRYSTALS IN RADIO [Source, Nature, Properties, & Applications].—C. F. Baldwin. (*Communications*, Oct. 1942, Vol. 22, No. 10, pp. 20..25 and 37, 40.)

1183. PROPOSED STANDARD CONVENTIONS FOR EXPRESSING THE ELASTIC AND PIEZO-ELECTRIC PROPERTIES OF RIGHT AND LEFT QUARTZ.—W. G. Cady & K. S. Van Dyke. (*Proc. I.R.E.*, Nov. 1942, Vol. 30, No. 11, pp. 495–499.) *Cf.* 500/2 of 1941.

1184. ON THE INFLUENCE OF INACCURACIES IN THE FORM AND THICKNESS OF A CIRCULAR RING ON THE FREQUENCY OF ITS PLANE FLEXURAL VIBRATIONS. — K. Federhofer. (*Physik. Berichte*, 15th July 1942, Vol. 23, No. 14, p. 1386.)

1185. THE TECHNIQUE OF FREQUENCY MEASUREMENT, AND ITS APPLICATION TO TELECOMMUNICATIONS [Improvement in Frequency Standards briefly Outlined: Methods used to obtain Accurate Measurements in terms of Improved Standards.]—J. E. Thwaites & F. J. M. Laver. (*Journ. I.E.E.*, Part III, Sept. 1942, Vol. 89, No. 7, pp. 139–165.)

1186. THE ACCURATE GENERATION OF SUB-FREQUENCIES FROM A STANDARD.—E. Newman. (*Electronic Eng:g*, Nov. 1942, Vol. 15, No. 177, pp. 244 and 249.)

1187. NEW REFERENCE-FREQUENCY EQUIPMENT [provides Frequencies, accurate to 2 Parts in 10^7, of 100, 10, 1.0 and 0.1 kc/s, with Apparatus to detect Errors and to operate Alarm if Error exceeds Limit].—V. J. Weber. (*Bell Lab. Record*, Nov. 1942, Vol. 21, No. 3, pp. 73–76.)

1188. A SELECTIVE CIRCUIT AND FREQUENCY METER USING A TUNING FORK.—Tucker. (*See* 1036.)

1189. TUNING METHOD FOR ELECTRICAL OSCILLATORY CIRCUITS.—K. Beyerle & K. P. Schweimer. (*Arch. f. Tech. Messen*, Sept. 1942, Part 135, V373-4, Sheet T92.)

The common way of tuning a circuit to a given frequency is to alter its inductance or capacitance until an indicator shows the maximum or minimum value of current or voltage. But the change in value of the impedance is particularly small just in the neighbourhood of its maximum or minimum, so that the optimum tuning point (especially for highly damped circuits such as are encountered in l.f. technique) is often poorly marked and hard to recognise. The change of phase angle, on the other hand, is especially large as the zero point is passed through, and therefore provides a sharper indication of tuning. A tuning device (circuit Fig. 5) based on this principle, and embodying a bridge circuit and a cathode-ray oscilloscope with its associated circuits, is shown in Fig. 6: it was designed for the frequency range 200–1000 c/s. For circuits of factor of merit around 100 the accuracy of adjustment is within about 0.5‰: for deviations of this order the straight-line trace becomes an elongated ellipse or two adjacent lines. Besides its primary use in tuning a parallel circuit (a similar arrangement is possible for series circuits) the device is valuable for adjusting a number of equal circuits so that the phase characteristic in each is the same, and also for determining the temperature characteristics of coils and condensers.

1190. THE MAGIC EYE AS A RESONANCE INDICATOR.—J. M. A. Lenihan. (*Electronic Eng:g*, Sept. 1942, Vol. 15, No. 175, p. 164.)

1191. THE WIEN BRIDGE AND SOME OF ITS APPLICATIONS [as a Filter and in Oscillator Circuits].—J. S. Worthington. (*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, pp. 214–216.)

1192. A CARRIER-FREQUENCY HETERODYNE OSCILLATOR [having Frequency Range 50 c/s to 170 kc/s: developed on Similar Lines to Earlier Ryall–Sullivan Oscillator].—K. W. Bourne. (*P.O. Elec. Eng. Journ.*, July 1942, Vol. 35, Part 2, pp. 65–68.)

1193. FORMULAS FOR THE INDUCTANCE OF RECTANGULAR TUBULAR CONDUCTORS [using G. Stein's $H(x)$ Function, a Table of which is Included].—T. J. Higgins. (*Journ. Applied Phys.*, Nov. 1942, Vol. 13, No. 11, pp. 712–715.) *Cf.* 3657 of 1942.

1194. FORMULAS FOR THE MAGNETIC-FIELD STRENGTH NEAR A CYLINDRICAL COIL.—H. B. Dwight. (*Elec. Engineering*, June 1942, Vol. 61, No. 6, Transactions pp. 327–333.)

1195. STANDARDS FOR REPLACEMENT PARTS FOR CIVILIAN RADIO IN WAR TIME [American Standards Association Project].—O. H. Caldwell. (*Industrial Standardisation*, Dec. 1942, Vol. 13, No. 11, pp. 312–313.)

1196. Mica Capacitor Standard, First American War Standard on Radio.—H. P. Westman. (*Industrial Standardisation*, Dec. 1942, Vol. 13, No. 11, pp. 297–299.)

"The object was to prepare a purchase specification which could be approved by all branches of the (American) Services, and which would permit interchangeability in manufacture and in service."

1197. Fixed Condensers [New British Standard Specification for Fixed Capacitors].—(*Wireless World*, Feb. 1943, Vol. 49, No. 2, p. 53.)

1198. Standard Resistance Values [Manufacturers and Government Decision].—(*Elec. Review*, 29th Jan. 1943, Vol. 132, No. 3401, p. 146.)

1199. Silver Alloys as Resistance Materials: I & II [Tin-Free Ag–Mn and "NBW" Alloys: Advantages in resisting Corrosion, and in Compensation Circuits to give Independence of Temperature].—A. Schulze. (*Arch. f. Tech. Messen*, Sept. & Oct. 1942, Parts 135 & 136, Z931-7 & 8, Sheets T97–98 & 110). For previous work *see* 2134 of 1942.

1200. A Simple Method of Measuring the Specific Resistance of the Material of a Circular Disc [based on Formula involving Resistances between Two Points on Periphery & Two Points on Surface].—B. J. Filippowitsch. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1321: summary of Russian paper.)

1201. The Measurement of the Loss Coefficients of Magnetic Dust Core Materials.—V. G. Welsby. (*P.O. Elec. Eng. Journ.*, July 1942, Vol. 35, Part 2, pp. 46–49.)

A method is described by which the losses in a toroidal dust-cored coil may be measured and those associated with the winding eliminated. Splitting up the core losses into three components, the hysteresis and eddy-current coefficients can be evaluated with possible error of $\pm$ 5 per cent, but the error in the residual coefficients may reach $\pm$ 10 per cent. These coefficients are of considerable importance in the development of core materials.

1202. Methods of Measuring the Slope of Amplifier Valves.—Bärisch. (*See* 1108.)

1203. An Improved Inter-Electrode-Capacitance Meter.—(*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, pp. 212–213: summary, from *R.C.A. Review*, April 1942.)

1204. The Q Meter and Its Theory [Theory of Direct-Reading Instruments including Corrections to increase Accuracy of Measurments].—V. V. L. Rao. (*Proc. I.R.E.*, Nov. 1942, Vol. 30, No. 11, pp. 502–505.) *See* also 112 of January.

1205. A Distortion Meter.—Hayes. (*See* 1127.)

1206. A Versatile Valve-Voltmeter.—O. Limann. (*Physik. Berichte*, 15th June 1942, Vol. 23, No. 12, p. 1244.)

From *Funkschau*, No. 4, Vol. 15, 1942, p. 55 onwards. A triode is connected as a d.c. amplifier, and supplied with various auxiliary units for different applications: a resistance–capacitance filter for d.c. measurements, a similar filter preceded by a Sirutor rectifier for a.c. measurements, and a diode pick-up unit for high frequencies. The whole is mains-fed, and in spite of its simplicity is notably unaffected by mains fluctuations.

1207. A Multi-Range Direct-Current and Voltmeter Circuit [making Use of a "Multi-Stage" Shunt and Multiplier Circuit to Reduce the Number of Shunt and Series Resistors Required].—G. E. Roth. (*Journ. of Scient. Instr.*, Jan. 1943, Vol. 20, No. 1, pp. 12–15.)

1208. Rectifier Instruments [Characteristics, Possibilities, & Limitations].—F. R. Axworthy. (*Elec. Review*, 18th Dec. 1942, Vol. 131, No. 3395, pp. 791–793.) "A general appreciation of the characteristics of rectifier instruments will enable users to obtain the maximum benefit from them and may in some cases explain apparently anomalous readings."

1209. Design of Long-Scale Indicating Instruments.—A. J. Corson, R. M. Rowell, & S. C. Hoare. (*Elec. Engineering*, June 1942, Vol. 61, No. 6, Transactions pp. 318–327.)

SUBSIDIARY APPARATUS AND MATERIALS

1210. Applications of Cathode-Ray Tubes [in Ultra-High-Frequency Technique: Survey].—Dudley. (*Electronics*, Oct. 1942, Vol. 15, No. 10, pp. 49–52 and 154, 155.)

1211. The Cathode-Ray Tube used Stroboscopically [Beam Quenched except at Regular Intervals].—Bocking. (*Electronic Eng:g*, Aug. 1942, Vol. 15, No. 174, pp. 102–103.) *Cf.* 436 of February.

1212. An Auxiliary Circuit for Cathode-Ray Photography [of Transient Phenomena].—Roberts. (*Electronics*, Sept. 1942, Vol. 15, No. 9, pp. 59–60 and 144.)

1213. A Simple Electronic Switch [for Viewing Two Phenomena Simultaneously on a Single-Beam C.R. Tube].—Russell. (*Electronic Eng:g*, Dec. 1942, Vol. 15, No. 178, pp. 284–285.)

1214. Wave-Form Circuits for Cathode-Ray Tubes: Part II.—Lewis. (*See* 1026.)

1215. Calculation of Axially Symmetrical Fields [in particular, the Potential Distribution of a Simple Electron Lens: Comparison with Plot made in an Electrolytic Plotting Tank].—Bertram. (*Journ. Applied Phys.*, Aug. 1942, Vol. 13, No. 8, pp. 496–502.) For a previous paper *see* 451 of 1941.

1216. Electrostatic Electron Lenses with a Minimum of Spherical Aberration: Errata [Correction of Errors in "Note Added in Proof"].—Plass. (*Journ. Applied Phys.*, Aug. 1942, Vol. 13, No. 8, p. 524.) *See* 2774 of 1942.

1217. Electron Optics [Lecture before the Electronics Groups, Institute of Physics].—Gabor. (*Electronic Eng:g*, Dec. 1942 and Jan. & Feb. 1943, Vol. 15, Nos. 178/180, pp. 295–299, 328–331 and 337, & 372–274.) Part of this paper was dealt with in 526 of February.

1218. A Simplified Electron Microscope.—Bachman & Ramo. (*Phys. Review*, 1st/15th Nov. 1942, Vol. 62, No. 9/10, p. 494.)

1219. The Measurement of Depth in Electron-Microscopic Objects [Long Account of Writer's Technique, using the Stereoscopic Method and also the "Oblique Bedusting" Method (Silver-Vapour Particles incident at an Angle to the Carrier Foil: a Shadow-Effect Method): the Necessary Precise Determination of the Electron-Microscopic Magnification].—Müller. (*Kolloid-Zeitschr.*, No. 1, Vol. 99, 1942, p. 6 onwards.) This is the work mentioned by Gotthardt, 3082 of 1942. For a long summary see *Physik. Berichte*, 1st Aug. 1942, Vol. 23, No. 15, pp. 1476–1477.

1220. Observations on the Secondary-Electron Radiation from Non-Conductors [by a Method applicable also to Photoelectron Measurements].—Geyer. (*See* 1119.)

1221. Action of Red Light during the Excitation of Phosphorescent Substances [Experimental Investigation].—Bertolino. (*Physik. Berichte*, 1st Aug. 1942, Vol. 23, No. 15, pp. 1504–1505: summary of Italian paper.)

1222. On the Destruction of the Phosphorescence of ZnS–Cu Phosphors.—Becker & Schaper. (*Ann. der Physik*, 21st Dec. 1942, Vol. 42, No. 4, pp. 297–336.)

The present investigation, based on extensive and important new experimental results, shows "that the findings hitherto set out in the literature require correction in many ways and are also susceptible to extension," and that the influences which more or less destroy luminescence may lead to new properties of the phosphor which can be considered to be positive results of the disintegrating action. While the latter is to be considered a desirable phenomenon for particular purposes, on the other hand a way is shown in which, in certain cases, it can be eliminated: a result for which up to now no possibility seemed to exist. For previous work on the subject, some of it carried out in the same laboratories (Philipp Lenard Institute, Heidelberg University) *see* for instance Goos, 2714 of 1940 [for later work, 2468 of 1941], Bandow, 2068 of 1942, and Streck, 1147 & 3295 of 1939.

Practically the whole of the long section A deals with the disintegrating action of alpha rays. Subsection III investigates at length Goos's discovery that the long-wave extinguishing effect is greatly increased by radioactive disintegration of ZnS phosphors: "the relation between the height of the extinction and the intensity of the exciting and quenching light is thoroughly examined. The rise in the extinguishing action is especially steep in the region of low quenching intensities, the more so the further the phosphor-disintegrating action has progressed. Thus the extinguishing action is particularly well adapted to the detection and measurement of very small infra-red intensities." The quenching action sets in without appreciable inertia, stretches widely into the infra-red band, and in strongly disintegrated phosphors can effect a complete extinction of an existing excitation. It affects only the photoluminescence: alpha-luminescence is not appreciably quenched by infra-red irradiation of the phosphor.

Subsection Av deals with the temperature dependence of the alpha-red disintegration: "the disintegration of a phosphor by alpha rays can be completely prevented by a temperature rise to several hundred degrees." Subsection Avi deals with beta and gamma irradiation: even high values produced no signs of disintegration.

Finally, the much shorter Section B compares the alpha-ray results with the disintegration produced by light: the behaviours are so different that the internal mechanisms cannot be the same. Disintegration by light requires a definite waveband and the cooperation of moisture: alpha-ray disintegration is at least approximately independent of moisture. And in every case disintegration by light was found to *decrease* the quenching effect instead of increasing it. On the other hand, disintegration produced by pressure (in a hydraulic press) increases the quenching effect, like alpha-ray disintegration: the action is considerably greater than can be explained merely as due to the resulting decrease in excitability. But it is incorrect to deduce that alpha-ray disintegration is a pressure effect due to the kinetic energy of the ray atoms, and the alpha-ray disintegration gives a better ratio of increase of quenching effect to decrease of phosphorescence, and is thus more favourable to infra-red investigations.

1223. Spectrum Analysis of the Fluorescence of Crystalline Platinocyanides.—Genard & others. (*Physik. Berichte*, 15th July 1942, Vol. 23, No. 14, p. 1435: summary of Belgian paper.)

1224. Preserving the D.C. Level in Oscillograph Amplifiers.—Russell. (*Electronic Eng:g*, Sept. 1942, Vol. 15, No. 175, p. 173.) Vibrator used to convert direct voltage into square-topped wave with any alternating component superimposed on the flat tops. This can be handled by most oscillograph amplifiers.

1225. Precise Voltage Stabiliser [Summary of Paper on Minimisation of Mains Fluctuations by a Stabilising Metal-Filament Lamp Bridge coupled to Thermionic Amplifier].—Glynne. (*Elec. Review*, 20th Nov. 1942, Vol. 131, No. 3391, pp. 655–656.) *See* also 519 of February.

1226. VOLTAGE-REGULATING TRANSFORMERS [Review of Applications of Saturation].—Lambert. (*Electronic Eng:g*, Feb. 1943, Vol. 15, No. 180, pp. 384–387.)

1227. A NEW INSTRUMENT FOR RECORDING TRANSIENT PHENOMENA [employing Magnetic-Tape Recording].—Begun. (*Elec. Engineering*, April 1942, Vol. 61, No. 4, Transactions pp. 175–177.)

1228. NOTES ON HIGH-VACUUM TECHNIQUE.—Burrows. (*Journ. of Scient. Instr.*, Feb. 1943, Vol. 20, No. 2, pp. 21–28.)

1229. THE CURE OF BUMPING IN MERCURY–GLASS DIFFUSION PUMPS [by Passage of a Few Hundred C.C. of Sulphuretted Hydrogen through Pump before Use].—Adam & Balson. (*Journ. Chemical Soc.* [London], Sept. 1941, p. 620.)

1230. IRON-TO-GLASS SEALS [for Vacuum Tubes].—(*Electronics*, Aug. 1942, Vol. 15, No. 8, p. 97.)

1231. PERMEABILITY OF METALS TO GASES [Diffusion depends Not Only on Geometrical Conditions in Metal Lattice but also on a "Chemical" Affinity between Gas & Metal Atoms: etc.].—Fast. (*Philips Tech. Rundschau*, No. 12, Vol. 6, 1941, p. 369 onwards.) For summary see *Physik. Berichte*, 15th July 1942, Vol. 23, No. 14, p. 1387.

1232. AN "ANODE-CHANNEL" SOURCE OF BEAMS OF POSITIVE CARRIERS [New Design, with over 8 mA Output].—Schmitthenner. (*Ann. der Physik*, 21st Dec. 1942, Vol. 42, No. 4, pp. 273–286.)

1233. STATIC CONVERSION OF DIRECT CURRENT TO ALTERNATING CURRENT WITH GRID-CONTROLLED MERCURY-ARC MUTATORS.—Feinberg. (*Journ. I.E.E.*, Part I, Oct. 1942, Vol. 89, No. 22, pp. 462–470.)

1234. A NEW METHOD OF DETERMINING THE TEMPERATURE OF A HIGH-PRESSURE DISCHARGE.—Elenbaas. (*Physica*, No. 1, Vol. 9, 1942, p. 53 onwards.) For a summary see *Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1324.

1235. MEASUREMENT OF THE ANODE–CATHODE VOLTAGE DROP OF AN A.C. DISCHARGE TUBE.—Baulk. (*See* 1116.)

1236. THE CALCULATION OF THE TOWNSEND COEFFICIENT OF IONISATION.—Zaitzev. (*Physik. Berichte*, 15th June 1942, Vol. 23, No. 12, p. 1240.)

From *Journ. of Exp. & Theoret. Phys.* [in Russian], No. 4, Vol. 9, p. 469 onwards. Calculation of the number of atoms per cm path ionised by electrons, for two different velocity distributions. Comparison with measurements on a glow discharge in neon shows that a Maxwellian distribution yields too high values for these coefficients.

1237. ON THE DISCHARGE OF POSITIVE POINTS [Investigation of Mechanism of Point Discharges & Electrical Wind: Discovery of a Hysteresis Effect in Wind from Positive (Not from Negative) Charge].—Yadoff & Yadoff. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1324: summary, from *Comptes Rendus* [Paris], No. 4, Vol. 214, 1942, p. 158 onwards.)

1238. OPERATION OF A THYRATRON AS A RECTIFIER.—Ware. (*See* 1115.)

1239. SEMICONDUCTOR PHOTOELEMENTS AND RECTIFIERS [and Some New Conclusions regarding Cuprous-Oxide Cells].—Fink & Adler. (*See* 1172.)

1240. COPPER-OXIDE RECTIFIERS IN STANDARD BROADCAST TRANSMITTERS.—Harmon. (*Proc. I.R.E.*, Dec. 1942, Vol. 30, No. 12, pp. 534–535.)

Improvements in the processing of copper used in copper-oxide rectifiers make practical the use of this type of rectifier in broadcast transmitters. Forced-draught cooling greatly increases the output allowable.

1241. METAL RECTIFIERS [Discussion of I.E.E. Paper].—Williams & Thompson. (*Journ. I.E.E.*, Part I, Aug. 1942, Vol. 89, No. 20, pp. 357–362.) *See* 248 of 1942.

1242. CONTACT NON-LINEARITY, WITH REFERENCE TO THE METAL RECTIFIER AND THE CARBORUNDUM CERAMIC NON-LINEAR RESISTOR [Experimental Analysis designed to correlate and formulate the Behaviour of the Various Structures possessing Contact Non-Linearity].—Fairweather. (*Journ. I.E.E.*, Part I, Nov. 1942, Vol. 89, No. 23, pp. 499–518.)

1243. THE HALF-WAVE VOLTAGE-DOUBLING RECTIFIER CIRCUIT.—Waidelich & Gleason. (*See* 1053.)

1244. THE ELECTROTECHNICAL FOUNDATIONS OF THE CONTACT CONVERTER: Part II.—Koppelmann. (*Elektrot. u. Masch:bau*, 24th April 1942, Vol. 60, No. 17/18, pp. 189–194.) For this "K-Converter" *see* 1780 of 1942.

1245. VIBRATOR CIRCUIT.—Masteradio. (*See* 1083.)

1246. LOW-CAPACITANCE A.C. POWER SUPPLIES [Need for Economy of Material, particularly Aluminium: Design of Power Filters with Minimum of Capacitance].—Mountjoy & Finnigan. (*R.C.A. Review*, April 1942, Vol. 6, No. 4, pp. 455–462.)

1247. SOME PECULIARITIES OF THE LEAD ACCUMULATOR WORKING AT LOW TEMPERATURES.—Briner & Yalda. (*Physik. Berichte*, 15th July 1942, Vol. 23, No. 14, p. 1423.) Summary of a report on some Geneva University researches.

1248. FREQUENCY DIVISION WITHOUT FREE OSCILLATION.—Tucker & Marchant. (*See* 1052.)

1249. The Dielectric Constants of Higher Alcohols at the Solidifying Point [and the Reason for Their High Values].—Frosch. (*Ann. der Physik*, 21st Dec. 1942, Vol. 42, No. 4, pp. 254–272.)

Experimental investigation of recently solidified n-$C_{12}H_{24}OH$ and n-$C_{16}H_{33}OH$. "It is the object of this work to explain the high capacitance values, possessed by a condenser with a dielectric which has just solidified, by a stratification (layers of dielectric already solidified and of dielectric still in the liquid state), and to confirm this explanation by measurements. The question why such a stratification should occur is discussed [section va, pp. 263–264]. By this explanation it is no longer necessary to attribute to the solid material just after solidification a high para-electric polarisation." The stratification hypothesis can reasonably be applied to explain also Damköhler's high dielectric constant for HBr at its transition point of 89° absolute, Wintsch's results with ice, and other experimental findings. An added footnote calls attention to Stachowiak's recent measurement (2317 of 1941) of dielectric constants up to 17 000 in biological substances, in the wave-range 400–10 000 m: here again it is not a matter of high para-electric polarisation but a structural effect (stratification of cell elements).

1250. Dielectric Relaxation as a Chemical Rate Process.—Kauzmann. (*Reviews of Modern Phys.*, Jan. 1942, Vol. 14, No. 1, pp. 12–44.)

This comprehensive survey is in two parts. Part I is theoretical and deals with the derivation of the differential equation for the relaxation of the dielectric polarisation. It then gives the solution of this equation for static and oscillating fields. The physical nature of the energy losses in dielectrics is discussed and the factors determining the transition probabilities are given. Part II is a comparison of theory with experiment, including the interpretation of observed data on relaxation rates in terms of molecular processes.

1251. Modern Materials in Telecommunications: Part VIIa—Non-Conductors: Insulation Resistance and Electrical Strength [the Relationship between Structure and Electrical Properties in Unidirectional Fields].—Radley, Speight, Richards, & Walker. (*P.O. Elec. Eng. Journ.*, Oct. 1942, Vol. 35, Part 3, pp. 84–87.) For previous parts *see* 2500 of 1942.

1252. Electrical Insulating Materials from Bentonite Clays [Dehydrating Process yielding Films with Dielectric Strength of 90 kV/mm: Elastic & Non-Hygroscopic].—Wolkowa. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1346: summary of Russian paper.)

1253. "Neuzeitliche Verwendung feinkeramischer Werkstoffe in der Technik" [Recent Application of Fine-Ceramic Materials: Book Review].—Assoc. of German Electrotech. Porcelain Manufacturers. (*Zeitschr. f. Fernmeldetech.*, 17th Oct. 1942, Vol. 23, No. 10, p. 160.)

1254. Experimental Investigations on the Fine Structure of the Röntgen Absorption Edges [including Results for Rutile, Anatas, & Brookite Modifications of TiO_2].—Saur. (*Ann. der Physik*, 3rd Dec. 1942, Vol. 42, No. 2/3, pp. 223–240.)

1255. Insulator Glazes [Effect of Glazing on the Insulating Properties (especially in Relation to Humidity) and on the Mechanical Strength of Porcelain Insulators: Microphotographs of Surfaces].—Rosenthal. (*Elec. Review*, 27th Nov. 1942, Vol. 131, No. 3392, pp. 675–678.) From Bullers, Ltd.

1256. Fibrous Glass Insulation [Properties & Applications].—Robertson. (*Elec. Review*, 19th Feb. 1943, Vol. 132, No. 3404, pp. 247–248.) For other recent papers on the subject *see* 228/9 of 1942 and back references.

1257. Insulating Materials for Electrical Conductors ["Surprising" Improvement of Polyvinyl Chloride by mixing with Silicates (Quartz, Glass, or Mica)].—I.G. Farbenindustrie A.G. (*Zeitschr. f. Fernmeldetech.*, 16th Jan. 1942, Vol. 23, No. 1, p. 12.) D.R.P. 704 361.

1258. More Mica from the Western Hemisphere [Survey of Mica Supply for United States Industry].—(*Elec. Engineering*, Dec. 1942, Vol. 61, No. 12, pp. 607–609.) For other recent references to the mica situation *see* 2818 & 3369 of 1942.

1259. Mica Capacitor Standard, First American War Standard on Radio, *and* British Standard Specification for Fixed Condensers.—Westman: British Standards. (*See* 1196 & 1197.)

1260. Insulating Wood [Experimental Development of Improved Qualities for Electrical Purposes].—Jervis. (*Elec. Review*, 6th Nov. 1942, Vol. 131, No. 3389, pp. 581–583.) For a previous article *see* issue of 27th March, 1942.

1261. The Use of Wood, dried in a High-Frequency (particularly an Ultra-High-Frequency) Field, for Electrical Insulation.—Siemens-Schuckert. (*Zeitschr. f. Fernmeldetech.*, 16th Feb. 1942, Vol. 23, No. 2, p. 28.) D.R.P. 704 667.

1262. Paper Dielectrics containing Chlorinated Impregnants [Deterioration under Certain Conditions due to Chemical Decomposition].—McLean & others. (*Bell S. Tech. Journ.*, June 1942, Vol. 21, No. 1, p. 77: abstract only.) For another summary *see* 847 of 1942.

1263. Physical Chemistry of Resin Solutions: Part I [Anomalous Solubility of Shellac & Other Resins in Organic Solvents]: Parts II & III.—Palit. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1305: p. 1305: pp. 1305–1306: summaries of Indian papers.)

1264. Infra-Red Heating [Report of Lecture: Application of Infra-Red Lamps to Drying or Softening of Synthetic Resin Materials].—Rowland. (*Elec. Review*, 20th Nov. 1942, Vol. 131, No. 3391, p. 655.) *See* also 665 of February.

1265. The Pre-Heating of Moulding Materials [to Shorten the Compression Process: Use of Temperatures above 100°C by Uniform Heating, particularly by Radiation].—Brandenburger. (*Physik. Berichte*, 1st July 1942, Vol. 23, No. 13, p. 1345.)

1266. "Physics in the U.S.S.R." [Review of Series of Ten Articles, including Study of Insulation].—Pilot Press, Ltd. (*See* 1309.)

1267. Conducting Silver Films on Refractory Materials [Non-Inflammable Base for Silver-Oxide Paint used to produce Silver Films on Reduction by Heating].—King. (*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, p. 216.)

1268. Electro-Plating [Application to Non-Conducting Materials].—(*Electrician*, 29th Jan. 1943, Vol. 130, No. 3374, p. 122.)

1269. The Properties of Materials [Their Classified Description as an Aid to Their Selection for Particular Applications].—Owen. (*BEAMA Journ.*, Jan. 1943, Vol. 50, No. 67, pp. 6–10.)

1270. The Measurement of the Loss Coefficients of Magnetic Dust Core Materials.—Welsby. (*See* 1201.)

1271. Electrical Conductivity and Transition Point of Fe_3O_4 [and the Reason of Its High Conductivity compared with Fe_2O_3 and Co_3O_4].—Verwey. (*Physik. Berichte*, 15th July 1942, Vol. 23, No. 14, p. 1420: summary of a Philips' Company paper.)

1272. The Magnetic Structure of Iron Crystals.—Elmore. (*Phys. Review*, 1st/15th Nov. 1942, Vol. 62, No. 9/10, pp. 486–493.)

1273. The Magnetic Properties and Uses of Iron Alloys [Properties and Uses of Iron Alloys for Communication Engineering].—Haigh. (*Journ. I.E.E.*, Part I, Oct. 1942, Vol. 89, No. 22, pp. 473–475.)

1274. On the Physical Phenomena of the Skin Effect of a Magnetic Field produced by Induction, as a Result of the Eddy Currents in Plates.—Hameister. (*See* 1002.)

1275. Magnet Windings of Tubular Conductors with Liquid Cooling [and Their Design for Minimum Volume & Exciting Input].—Plechl. (*Elektrot. u. Masch:bau*, 3rd July 1942, Vol. 60, No. 27/28, pp. 289–293.)

1276. Permanent Magnets [Recommendation of Hysteresis Loop rather than B/H Curve as Basis of Design].—Hamilton-Adams. (*Elec. Review*, 13th Nov. 1942, Vol. 131, No. 3390, p. 622.)

1277. Electrical Conductivity of Graphite Films.—Acheson, Ltd. (*Electronic Eng:g*, Aug. 1942, Vol. 15, No. 174, p. 124.) For application to electrical screening *see* 1530 of 1942.

1278. High-Frequency Resistance of Plated Conductors [Mathematical Analysis with Examples, including the Use of Copper-Plated Iron Conductors for Coaxial Cables].—Proctor. (*Wireless Engineer*, Feb. 1943, Vol. 20, No. 233, pp. 56–65.) *Cf.* Klein & others, 910/11 of March.

1279. Silver Alloys as Resistance Materials.—Schulze. (*See* 1199.)

1280. Electrical-Resistance Properties of Dilute Binary Alloys of Copper, Silver, and Gold [including a Silver-Manganese Alloy very suitable for the Measurement of Pressure].—Linde. (*Zeitschr. f. Instr:kunde*, Oct. 1941, Vol. 61, No. 10. pp. 353–355: summary only.)

1281. The Relay in Communications [Topical Review of Many Types of Relay in Effective Use To-day].—Peters. (*Communications*, Nov. 1942, Vol. 22, No. 11, pp. 10–12.)

1282. On the Response Times of Telecommunication Relays [and Electrical, Mechanical & Thermal Methods of Delay: Compressed Survey].—Schüssler. (*Zeitschr. f. Fernmeldetech.*, 16th Jan. 1942, Vol. 23, No. 1, pp. 9–10.)

1283. Anti-Vibration Material [New Textile Material for Use in Shock Absorption].—(*Electrician*, 12th Feb. 1943, Vol. 130, No. 3376, p. 164.)

1284. The "Tuchel" Contact ["Wiping", Self-Cleaning Contact Design for Small or Large, Single or Multiple Plugs & Sockets, Cable Connectors, etc: replacing Banana Plugs, Knife-Blade Contacts, & Other Devices for Communications or Power Purposes].—Dewald. (*Zeitschr. f. Fernmeldetech.*, 15th April 1942, Vol. 23, No. 4, pp. 55–59.)

The dagger-shaped plug is cleaned and gripped as it forces its way along successive sections of three (or more) pairs of edges formed by slitting an eccentric nest of tubes spot-welded together along a line diametrically opposite to the slits, and by division into sections by slots at intervals along the axial length of the tubes.

1285. The Manufacture and Properties of Tungsten and Molybdenum.—Percival. (*Electronic Eng:g*, Aug. 1942, Vol. 15, No. 174, pp. 104–108.)

1286. Beryllium-Copper: a High-Strength Alloy for Electrical and Instrument Springs.—Hunt. (*Electronic Eng:g*, Dec. 1942, Vol. 15, No. 178, pp. 276–279.)

1287. Beryllium–Copper and Its Applications [also Its Production, Properties, and Availability].—Crossley. (*Journ. of Scient. Instr.*, Jan. 1943, Vol. 20, No. 1, pp. 7–9.)

1288. Diamond Dies for the High-Speed Drawing of Copper Wire.—Padowicz. (*Bell S. Tech. Journ.*, June 1942, Vol. 21, No. 1, pp. 20–36.)

1289. The Soldering of Aluminium [Advantages of Hard and Soft Methods].—Einerl : Neurath. (*Electronic Eng:g*, Feb. 1943, Vol. 15, No. 180, p. 388.)

1290. Solder Alloys [Tin Conservation & Soldering-Iron Temperature : Editorial].—(*Elec. Review*, 18th Dec. 1942, Vol. 131, No. 3395, p. 771.)

1291. Wartime Soldering [Cored Solder & Emergency Alloys : Correct Method of Use of Cored Solders].—Arbib. (*Elec. Review*, 18th Dec. 1942, Vol. 131, No. 3395, pp. 777–779.)

1292. The Mechanical Design of German Army Wireless Components [Description of the More Unusual Designs].—Hull. (*Electronic Eng:g*, Nov. 1942, Vol. 15, No. 177, pp. 238–241.)

STATIONS, DESIGN AND OPERATION

1293. The Star Spangled Network [Wired Radio in American Army Camps].—Frank & Phillips. (*Communications*, July 1942, Vol. 22, No. 7, pp. 5–10.)

As contrasted with earlier wired systems that used low frequencies, reception is obtained with any radio set plugged into the power line and tuned to the appropriate broadcast frequency. Details are given of one transmitting system and of the method of coupling to 4000-volt power lines.

1294. High-Frequency Resistance of Plated Conductors.—Proctor. (*See* 1278.)

1295. Post-War Planning in Radio Communications [Discussion before the Wireless Section].—(*Journ. I.E.E.*, Part III, Sept. 1942, Vol. 89, No. 7, pp. 168–173.)

1296. Australia—New Empire Radio Network Plan [Proposals, accepted by Australian Cabinet, for a Network linking All Parts of Empire].—(*Elec. Review*, 22nd Jan. 1943, Vol. 132, No. 3400, p. 127.)

1297. The Measurement of Voltage Peaks in a Studio Equipment [and the Philips Peak-Indicator for Broadcasting Transmission Control].—de Fremery & Wenk. (*Philips Tech. Rundschau*, No. 1, Vol. 7, 1942, p. 20 onwards.)

The indicator consists of two push-pull amplifier stages followed by a rectifier and an output stage. The anode current from the latter is read off the -45 to $+45$ db scale of a line-of-light instrument, zero corresponding to an input alternating voltage of 1.55 V_{eff}, which for broadcasting transmissions over lines with a 200-ohms impedance serves as the standard level (12 mw). A resistance–capacitance time circuit makes the peaks reach their final deflection within 5 ms, whilst recovery takes place with a speed of at most 20 db/s : this renders observation more simple and less tiring. The grid-leak of the output valve takes the form of the non-linear resistance of a dry-plate rectifier (kept accurately at a constant temperature because of the temperature-dependence of its characteristic) to provide a linear scale.

1298. Subjective Determination of the Quality of Telephone Systems.—Panzerbieter & Rechten. (*See* 1141.)

1299. Aircraft Communications [Description of Bendix Communication Unit No. 508].—McKee. (*Communications*, Sept. 1942, Vol. 22, No. 9, pp. 12–15, 30, 34.) The first of a series of analyses of aircraft communication equipment and components.

1300. Typical Equipment in Aircraft Communications [General Description].—McKee. (*Communications*, Oct. 1942, Vol. 22, No. 10, pp. 12–13 and 47, 48.)

1301. Nazi Aircraft Radio [Equipment in Ju. 88, He. 111H, Me. 109 : also Forced Landing Sets].—Jupe. (*Electronics*, Nov. 1942, Vol. 15, No. 11, pp. 58–59, 167–169.) *See* also 3015 & 3016 of 1942 and 598 of February.

GENERAL PHYSICAL ARTICLES

1302. Energy Flow in Electric Systems : the *Vi* Energy-Flow Postulate.—Slepian. (*Elec. Engineering*, Dec. 1942, Vol. 61, No. 12, Transactions pp. 835–840.)

Author's summary :—" The conditions which a valid postulated electric-energy flow must satisfy are given and are stated to be insufficient for its unique determination. The commonly used *Vi* energy-flow postulate is shown by examples to be not generally valid, but by adding a simple term it can be made equally valid with other valid energy-flow postulates. Various examples are given of the application of this corrected energy-flow postulate. On power systems the engineer commonly limits his use of the uncorrected *Vi* postulate to applications where the correcting term should have a negligible net effect. Various examples of such use are discussed." *Cf.* 1311, below.

1303. Cosmic-Ray Theory.—Rossi & Greisen. (*Reviews of Modern Phys.*, Oct. 1941, Vol. 13, No. 4, pp. 240–309.)

The interaction of cosmic rays with matter gives rise to a great variety of secondary effects. The present article gives a very comprehensive survey of the more theoretical aspects of the subject. Part I deals with collision processes, radiation processes, the production of pairs and scattering. Part II deals with the theory of multiplicative showers.

1304. A New Table of Values of the General Physical Constants [as of August, 1941].—Birge. (*Reviews of Modern Phys.*, Oct. 1941, Vol. 13, No. 4, pp. 233–239.)

A discrepancy between the value of the electronic charge *e* obtained from oil-drop work, and that obtained by a later method involving wavelengths of X-rays determined by ruled gratings, made previously accepted values of the general physical constants wrong. It has been shown definitely that the value assumed for the viscosity of air by Millikan in his oil-drop work was seriously in error. Numerous recent determinations of the viscosity of air, together with new oil-drop work, have now brought the two values of e into satisfactory agreement. The present grating value appears, however, to be much more accurate than the oil-drop value, and is therefore adopted in the present tables. The quantity obtained in this method is primarily Avogadro's number. This now becomes a fundamental constant and e becomes a derived constant.

The earlier value of e was roughly 4.77×10^{-10} e.s.u.; the new value is about 4.80×10^{-10} e.s.u. This large change gives rise to a whole new set of discrepancies, affecting chiefly the radiation constants. *Cf.* Laby, 610 of March.

1305. ON THE THEORY OF THE METALLIC CONDUCTION OF ELECTRICITY, *and* THE ELECTRICAL AND THERMAL PROPERTIES OF METALS IN A MAGNETIC FIELD.—Sauter: Kohler. (*Ann. der Physik*, 3rd Dec. 1942, Vol. 42, No. 2/3, pp. 110–141: pp. 142–164.)

1306. A DEVELOPMENT OF LONDON'S THEORY OF SUPERCONDUCTIVITY.—von Laue. (*Ann. der Physik*, 3rd Dec. 1942, Vol. 42, No. 2/3, pp. 65–83.) Followed by other papers on superconductivity (Justi: Steiner & Gerschlauer).

1307. THE THEORY OF UNITS.—Bedford. (*Journ. of British I.R.E.*, Dec. 1942/Jan. 1943, Vol. 3, pp. 82–103.)

Author's summary:—" In the formulation of electromagnetic theory, there are four stages at which arbitrary constants are conveniently introduced for the purpose of defining units. It is customary to assign immediately the value unity to certain of these constants, after which they are lost sight of and have often at a later stage to be painfully resuscitated. In the present treatment a sketch of electromagnetic theory is given in which the suppression of these fundamental constants (k_1, k_2, k_3, k_4) does not occur. Maxwell's theory is shown to lead to a certain relationship between the k's involving the velocity of light. Subject to this and to one other restriction, the assignment of k-values is an arbitrary matter, the process of formulating a unit system being one of two degrees of freedom.

" A table is constructed showing the assignment of k-values corresponding to the various known unit systems. This table can, amongst other things, be used to derive the numerical relationships between the unit quantities of the various systems. One necessary example of this is given, but it is pointed out that the (advocated) use of a generalised system of units (k's unspecified) eliminates once and for all the tiresome and inelegant process of unit changing. Moreover, this generalised system of units is free of the dimensional inconsistencies which characterise the known systems, at least in their more usual modes of expression.

" Examples are given of the direct use of the generalised system of units; a particular problem which would normally present the most tiresome process of unit changing is solved straight out in generalised units from which the numerical answers are written down at once in practical units.

" Serious inconsistencies of method nomenclature in connection with the practical system of units are brought to light and proposals for rationalisation are considered." For a short summary see *Elec. Review*, 6th Nov. 1942, p. 592.

1308. ON THE DIMENSIONS OF PHYSICAL QUANTITIES [Editorial Discussion of Recent Papers in *Phil. Mag.* and *Proc. Phys. Soc.*].—G.W.O.H. (*Wireless Engineer*, Jan. 1943, Vol. 20, No. 232, pp. 1–3.) *See* for example 132 & 239 of January.

1309. " PHYSICS IN THE U.S.S.R." [Review of Series of Ten Articles by Various Authors, published by the Pilot Press Ltd: Study of Insulation: Electrification in the U.S.S.R.].—(*Elec. Review*, 20th Nov. 1942, Vol. 131, No. 3391, p. 646.)

1310. " INTRODUCTION TO MODERN PHYSICS " [Third Edition: Book Review].—Richtmyer & Kennard. (*Journ. Applied Phys.*, Nov. 1942, Vol. 13, No. 11, p. 687.)

MISCELLANEOUS

1311. ENERGY AND ENERGY FLOW IN THE ELECTROMAGNETIC FIELD [a New Derivation of Poynting's Vector, which also yields Other Energy-Flow Vectors].—Slepian. (*Journ. Applied Phys.*, Aug. 1942, Vol. 13, No. 8, pp. 512–518.) A summary was referred to in 3259 of 1942. *Cf.* 1302, above.

1312. STEADY STATE CURRENTS IN ELECTRICAL NETWORKS [Extension of Operational Circuit Analysis].—Waidelich. (*See* 1039 & 1040.)

1313. THE OPERATIONAL THEORY OF LONGITUDINAL IMPACT [by Analogy with Theory of Electrical Transmission Lines].—Pipes. (*Journ. Applied Phys.*, Aug. 1942, Vol. 13, No. 8, pp. 503–511.)

1314. THE ELECTRO-MECHANICAL ANALOGY IN OSCILLATION THEORY.—Manley. (*Journ. Roy. Aeron. Soc.*, Jan. 1943, Vol. 47, No. 385, pp. 22–26.)

1315. HEAVISIDE'S DIRECT OPERATIONAL CALCULUS [Application to the Solution of Engineering Problems].—Russell. (*Elec. Engineering*, Feb. 1942, Vol. 61, No. 2, pp. 84–88.)

1316. INVERSE FUNCTIONS OF COMPLEX QUANTITIES [Explicit Formulae for Such Functions].—Dwight. (*Elec. Engineering*, Dec. 1942, Vol. 61, No. 12, Transactions pp. 850–853.)

1317. THE TERM " VECTOR " [Letter indicating Ambiguity in the Present Use of the Term].—Feinberg. (*Electrician*, 12th Feb. 1943, Vol. 130, No. 3376, p. 172.)

1318. INTEGRATION IN THE COMPLEX PLANE [a Mathematical Prerequisite for Work on Laplacian Transforms, Fourier Integrals, and Travelling Waves on Transmission Lines]. — Friedrichs. (*Elec. Engineering*, March 1942, Vol. 61, No. 3, pp. 139–143.)

1319. ANALYSIS OF SYSTEMS WITH KNOWN TRANSMISSION–FREQUENCY CHARACTERISTICS BY FOURIER INTEGRALS.—Sullivan. (*Elec. Engineering*, May 1942, Vol. 61, No. 5, pp. 248–256.)

1320. WAVE ANALYSIS [Part I—General Review: Part II—Analysis of Semi-Periodic Wave-Forms].—Bourne. (*Electronic Eng:g*, Sept. & Dec. 1942, Vol. 15, Nos. 175 & 178, pp. 149–151 & 280–282.)

1321. 36 AND 72 ORDINATE SCHEDULES FOR GENERAL HARMONIC ANALYSIS [to facilitate Calculations: Applicable to Odd and Even Harmonics].—Denman. (*Electronics*, Sept. 1942, Vol. 15, No. 9, pp. 44–47.)

1322. SIX-FIGURE TRIGONOMETRICAL TABLES [Reference to Recently Published Tables]. —(*Engineer*, 30th Oct. 1942, Vol. 174, No. 4529, p. 368.)

1323. PRODUCTION TESTING. — Summers. (*Gen. Elec. Review*, Dec. 1942, Vol. 15, No. 12, pp. 701–703.)

A description of the application of cathode-ray tubes and reflecting galvanometers to the testing of mass-produced electrical products. *See* also 3668 of 1942 and back reference.

1324. AN EXTENSION OF NOMOGRAPHY.—Hansel. (*Phil. Mag.*, Jan. 1943, Vol. 34, No. 228, pp. 1–26.)

The author considers that the possible scope of nomographic methods is not generally appreciated. The generalised application of these methods includes the numerical solution of differential equations, the reduction of observations, and the determination of laws. Nomographic calculation should find a place in elementary mathematical training and nomographic paper with printed scales should be available cheaply.

1325. GOVERNMENT AND SCIENCE IN GREAT BRITAIN [Organisation of Scientific Effort for War Purposes].—Cripps. (*Nature*, 6th Feb. 1943, Vol. 151, pp. 152–153.)

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1334. LABORATORY SLANG [Correspondence criticising the Use of Undefined Unfamiliar Terms]. —Ipfeed: Puckle. (*Electronic Eng:g*, Oct. 1942, Vol. 15, No. 176, p. 217.)

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1337. "RADIO REFERENCE DATA" [Review of Booklet].—Standard Telephones & Cables. (*Electronic Eng:g*, Sept. 1942, Vol. 15, No. 175, p. 174.)

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1339. "THE TELEPHONE HANDBOOK" [Book Review].—Poole. (*Elec. Review*, 1st Jan. 1943, Vol. 132, No. 3397, p. 6.)

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1349. The Application of Carrier Systems to Submarine Cables: Part I [General Analysis of Problem of providing the Maximum Number of Carrier Circuits]: Part II [Design of Terminal Equipment].—Halsey. (*P.O. Elec. Eng. Journ.*, Oct. 1942, Vol. 35, Part 3, pp. 79–83: Jan. 1943, Vol. 35, Part 4, pp. 121–125.)

1350. Infra-Red Heating.—Rowland. (*See* 1264.)

1351. Disintegration of Phosphors by Alpha Rays with Increase in the Long-Wave Quenching Action: and Its Application to the Detection and Measurement of Infra-Red Radiation.—Becker & Schaper. (In paper dealt with in 1222, above.)

1352. Radiation Instruments using Geiger-Müller Tubes [for measuring Intensity of Radiation].—Weisz. (*Electronics*, Oct. 1942, Vol. 15, No. 10, pp. 44–48 and 118.) For other recent work by the same writer *see* 2224, 2585, & 2920 of 1942.

1353. The Operation of Proportional Counters [Use of Geiger Counters in which Magnitude of Pulse observed on Collecting Electrode is Proportional to Size of Initial Ionising Event].—Korff. (*Reviews of Modern Phys.*, Jan. 1942, Vol. 14, No. 1, pp. 1–11.)

1354. Hertzian Wave Spectroscopy with a Magnetron Oscillator [Physico-Chemical & Technical Importance of Dispersion & Absorption Measurements at High & Ultra-High Frequencies: Equipment using Magnetron Oscillator of Stabilised Frequency in Wave Range 0.70–1.80 m, and Drude's Second Method for measuring Dispersion & Absorption: Results with Alcohols].—Cavallaro. (*Physik. Berichte*, 1st Aug. 1942, Vol. 23, No. 15, p. 1485.)

1355. Anomalous Dispersion of Dipolar Ions [and "a New Method of Measuring the Dielectric Constant & Dielectric Absorption" at Ultra-High Frequencies, using C.R. Oscillograph to determine Phase-Displacement of Test Field in Two Similar Cells, One containing Water & the Other the Liquid under Test].—Marcy & Wyman. (*Physik. Berichte*, 1st Aug. 1942, Vol. 23, No. 15, p. 1486: summary of paper in *Journ. Am. Chem. Soc.*, Vol. 63.)

1356. The Production of Sinusoidal Alternating Currents in the Infra-Sonic Frequency Band [and Their Penetration into the Ground: Use in Radio Prospecting].—Müller. (*Physik. Berichte*, 1st Oct. 1942, Vol. 23, No. 19, p. 1811: summary, from *Zeitschr. f. Geophys.*, No. 5/6, Vol. 17, 1942, p. 181 onwards.)

The penetration depth of a current of frequency ν is approximately $s = k\sqrt{1/\nu}$; depths are calculated for frequencies between 1000 and 0.1 c/s for grounds of various conductivities, and it is seen that it is often necessary to use frequencies below 10 c/s if large depths are required. Ordinary generators are not suitable for such low frequencies, especially since pure sine waves are required. The writer describes two suitable circuits, both containing only condensers and ohmic resistances. The range of the oscillation period is from several hours to 1/100 000 second. The calculation of the oscillatory circuits is given and the conditions for the setting-up of oscillations are discussed. By the variation of these conditions, selective effects may be produced; for example, the percentage of silicic-acid content in a gold seam was thus determined. The circuits are suitable for all purposes where a small output is sufficient.

1357. Recording Machinery-Noise Characteristics [Apparatus & Technique, including Suggestions for Subsequent Laboratory Analysis by Oscillographic Means].—Brailsford. (*Electronics*, Nov. 1942, Vol. 15, No. 11, pp. 46–51.)

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