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AUTOMATIC FREQUENCY CORRECTION

As this paper is primarily concerned with the principles involved in the operation of A.F.C. circuits, the circuits themselves will not be treated in great detail; in fact such circuits as are shown are intended to assist those unfamiliar with the technique involved in correlating theory with practice. In the first part of the paper a basic circuit is described; the second part presents graphical methods whereby the designer may quickly and easily estimate the performance of any given circuits or alternatively determine the circuit characteristics for any required performance; in the third part, a rigorous analytical investigation of the problem confirms the correctness of the graphical methods employed. It will be appreciated that the methods and analysis are not confined to any particular form of circuit, but may be applied quite generally.

THE object of automatic frequency correction (A.F.C.) in a receiver is to compensate automatically for any errors or mal-adjustment which tend to result in the listener hearing a distorted programme. For instance in a superheterodyne receiver an intermediate frequency is produced which is equal to the difference between the incoming signal and the local oscillator frequency and it is at this frequency that the greater part of both the receiver gain and selectivity is produced. The selectivity requirements are generally such that should the actual intermediate frequency given by the difference between the incoming signal and the local oscillator frequencies differ by more than ± 1.0 kc. from the designed or predetermined intermediate frequency, distortion effects due to sideband cutting, etc., will be noticeable. Since the average listener usually does not tune a receiver with this degree of accuracy, it is thus the function of the A.F.C. circuit to ensure that the actual I.F. does not depart greatly from the designed I.F. in spite of mal-adjustment of the manual tuning control or subsequent frequency drift either at the transmitter or in the local oscillator. In passing, attention may be drawn to the fact that automatic frequency correction systems are sometimes incorrectly termed automatic tuning control (A.T.C.) systems; the latter term should preferably be restricted to systems in which the act of tuning is automatic rather than manual (or to systems in which "searching" is eliminated).

As applied to a superheterodyne receiver in which, for instance, the oscillator frequency is higher than the signal frequency, the A.F.C. may operate from the intermediate frequency amplifier back on to the local oscillator through (A) a discriminator which provides an output dependent on the magnitude and sense of the deviation of the received I.F. signal from the designed I.F., and (B) a control device operating from the output of the discriminator on the local oscillator circuit so as to alter the frequency of this circuit in such a manner that the original deviation from the designed I.F. is considerably reduced.

In a well-known form of discriminator, energy is fed from the intermediate frequency amplifier (Fig. 1A) to a pair of tuned circuits, P, Q, one tuned above and the other an equal amount below the designed intermediate frequency. The radio frequency outputs of these circuits are rectified and the resultant direct current outputs combined differentially. The individual response curves are shown in

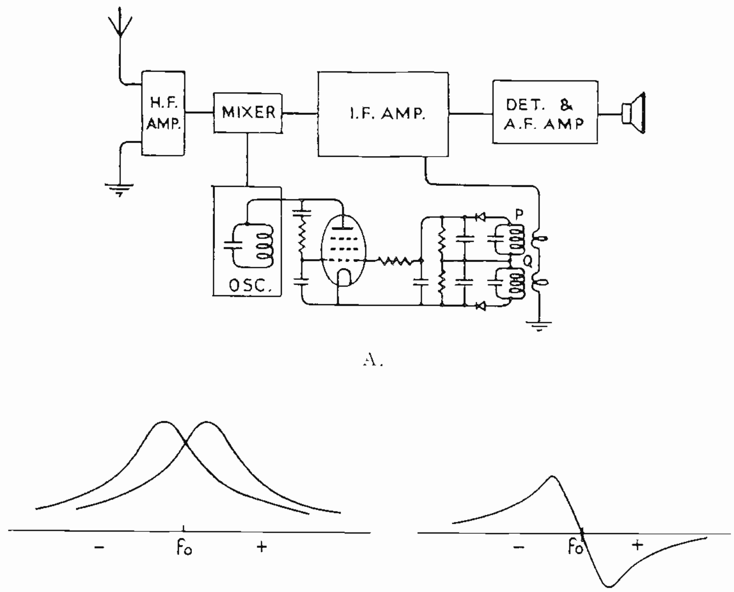


FIG. I.

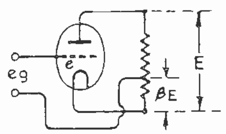
Fig. 1 (B) in relation to the designed intermediate frequency f_o, while Fig. 1 (C) shows the characteristic obtained by combining the previous ones differentially. One thus obtains a voltage output which varies in magnitude and sign with the departure of the actual intermediate frequency from the designed intermediate frequency. This voltage is now employed to effect a reactance change in the oscillator tuned circuit so as to alter the oscillator frequency in such a way as to cause the actual intermediate frequency to approach more closely the designed intermediate frequency. There are several ways of doing this, that generally employed being to use a control valve whose anode-cathode space is in parallel with the oscillator tuned circuit and whose grid is fed with voltage at the oscillator frequency which is substantially in phase quadrature with that across the tuned circuit. The anode-cathode space of this control valve then appears as a reactance to the tuned circuit, in parallel with the tuned circuit reactances themselves. The value of this reactance may be varied by variation of the mutual conductance of the valve (by means of bias derived from the discriminator circuit), thus varying the oscillator frequency.

The peaks of the discriminator curve may in a broadcast receiver conveniently be set at ± 1 kc. With the A.F.C. system in operation, on tuning towards a station the receiver " jumps " unto tune when the dial pointer indicates say 5 kc. off tune, and on tuning through the station, tune is still held until a point is reached, say, where the pointer indicates 7 kc. off tune, where tune is no longer held and the receiver "jumps out." With strong stations it frequently happens that jump out occurs with the pointer indicating as much as 20 kc. off tune, i.e., the strong station is held in tune and adjacent channels are skipped until at last the strong station is released. To avoid this, use is made of a "limited characteristic" control device which, as may be seen from the curves of Fig. 8, prevents these excessive values of jump-out and permits reception of weak stations adjacent to powerful ones.

The curves shown in the following notes are all of the limited characteristic type.

The discriminator will, therefore, be a device which produces a voltage or current output dependent upon the magnitude and sense of an initial frequency error F_1 say, due to mistuning, while the control device will produce a frequency change or correction f dependent on the voltage or current output of the discriminator. As a result, after the circuits have stabilised there will be a (final) frequency error F_0 such that

$$F_0 = F_1 - f.$$



Exact or over-compensation cannot of course be attained since the existence of some deviation is essential to the operation of the circuits and the maintenance of the correction; the system is, in fact, one which is operated essentially in a back-coupled state (i.e., with feedback), the feedback being in terms of frequency deviation or errors.

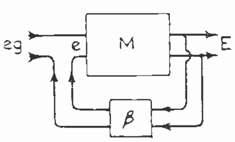


FIG. 2.

Before proceeding further and for the sake of clarity consider a valve operating under conditions of feedback (Fig. 2). Let a voltage e_g be applied to the input terminals of a valve such that without feedback a voltage E_1 is produced at the output, then $E_1 = M\,e_g$ where M is the magnification of the stage. Now let a fraction β of the output be fed back to the input, then the net voltage between grid and cathode is now $e = e_g + \beta E$ (the output voltage now being E), then $E = Me = M\,(e_g + \beta E)$ or $E = \dfrac{Me_g}{1 - \beta M}$.

The usual sign convention has been adopted for β, so that for β negative a reduction of output occurs for a fixed input. Now if the magnification of the circuit be unity we have

$$E = \frac{e_g}{1 - \beta}$$

or

$$e_g = E\,(1 - \beta) = E - \beta E.$$

In the case of frequency correction we may also apply the feedback principle (Fig. 3) for if F_1 be the initial or "input" error, F_0 the final or "output" error and f the correction, then as we have shown

$$F_0 = F_1 - f.$$

Now f is some fraction γ of the output frequency F_0

so that $\quad F_0 = F_1 - \gamma\,F_0 \qquad$ i.e. $\quad F_0 = \dfrac{F_1}{1 + \gamma} \qquad$ or $\qquad F_1 = F_0\,(1 + \gamma)$

which is of the same form as that obtained for valve feedback.

As stated above the correction f is obtained in practice as a result of the combined action of a discriminator whose characteristic may be expressed as volts against kc. error and a control device having a characteristic relating kc. correction and input (discriminator) volts. For feedback purposes the overall characteristic expressing kc. correction against kc. error is of importance and this may easily be obtained from the individual characteristics of the two units. It represents the relation $f = \gamma F_o$ and is not linear. This implies that the coefficient of feedback γ is not constant since γ is the ratio of kc. correction to kc. error or more correctly the differential of the kc. correction — kc. error curve. A typical set of curves is shown in Fig. 4 where A represents a normal discriminator curve, and B is a control curve. The curve C is obtained by combining curves A and B and expresses the relation $f = \gamma F_o$. Differentiating this we obtain curve D, from which it will be seen that γ may be positive, negative or zero. In this case, however, the criterion

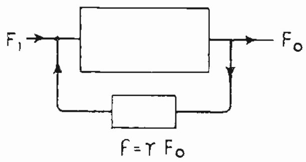


FIG. 3.

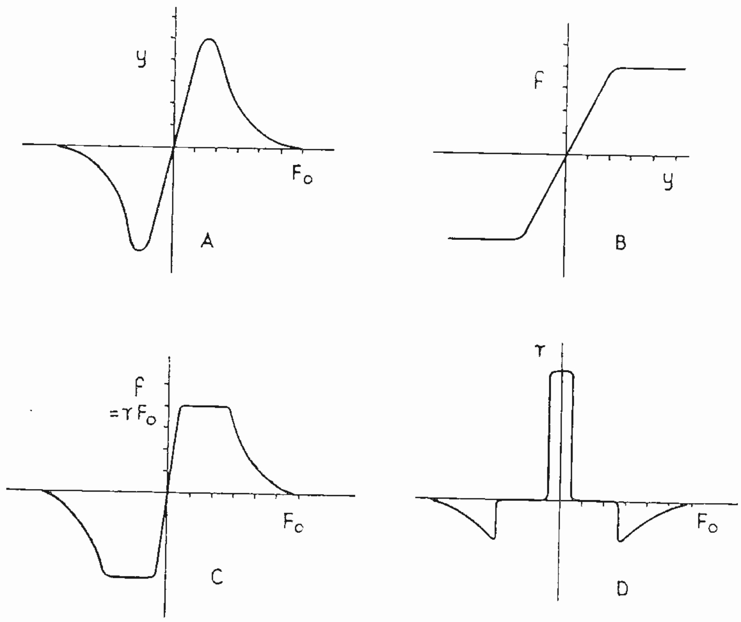


FIG. 4.

for the sign of the feedback is not the sign of γ but the slope of the γF_o — F_o curve (Fig. 4 (c)). If the slope of this curve is less than 135° (measured with respect to the positive F_o axis) the feedback is positive, if more than 135° feedback is negative, except for 180° when feedback is zero. This will be appreciated from a consideration of later curves.

In Fig. 5, for instance, which is in effect Fig. 4 (c) skewed 45° (so that the slope of the curve relative to 180° is now taken), it will be readily seen that, starting at any point on the curve in the regions AR or PC, a small decrease of F_I will result in a small increase of F_o, the slope of the curve being negative in these regions and hence indicative of positive feedback ; whilst on any other portion of the curve the slope is positive, a decrease of one quantity being associated with a decrease of the other and negative feedback exists.

Since γ is not a constant it is simpler, however, to use the curve of Fig. 4 (c) representing kc. correction against kc. error, i.e., f against F_o.

Now $\qquad F_o = F_I - f$

or $\qquad F_I = F_o + f = F_o + \gamma F_o$

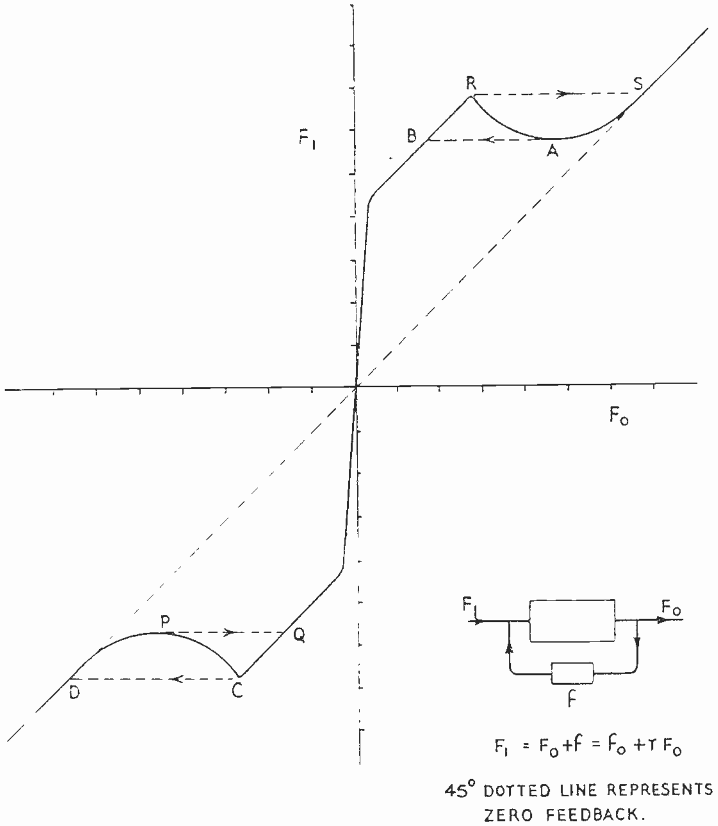


FIG. 5.

i.e., the input or initial error is equal to the output or final error plus the correction at that particular value of final error. It will be observed that in common with all back coupling problems, it is preferable to determine the input conditions from known output conditions rather than to use the reverse, but more usual, process. The relationship between F_I and F_o is shown graphically in Fig. 5.

It will be observed that without feedback the relationship between F_I and F_o is a straight line through the origin making an angle of 45° with the axes.

From this curve it is now possible to deduce the output conditions for given input conditions. For instance, suppose F_I is large and positive, then $F_o = F_I$, there being no feedback under these conditions. Now let F_I be gradually reduced. A set square, an edge of which is parallel to the F_o axis, can be moved up and down over the diagram to represent changes of F_I. Up to the point where feedback commences, $F_o = F_I$, but on further reduction of F_I, feedback comes into operation and F_o reduces at a greater rate. A horizontal line through the value of F_I chosen

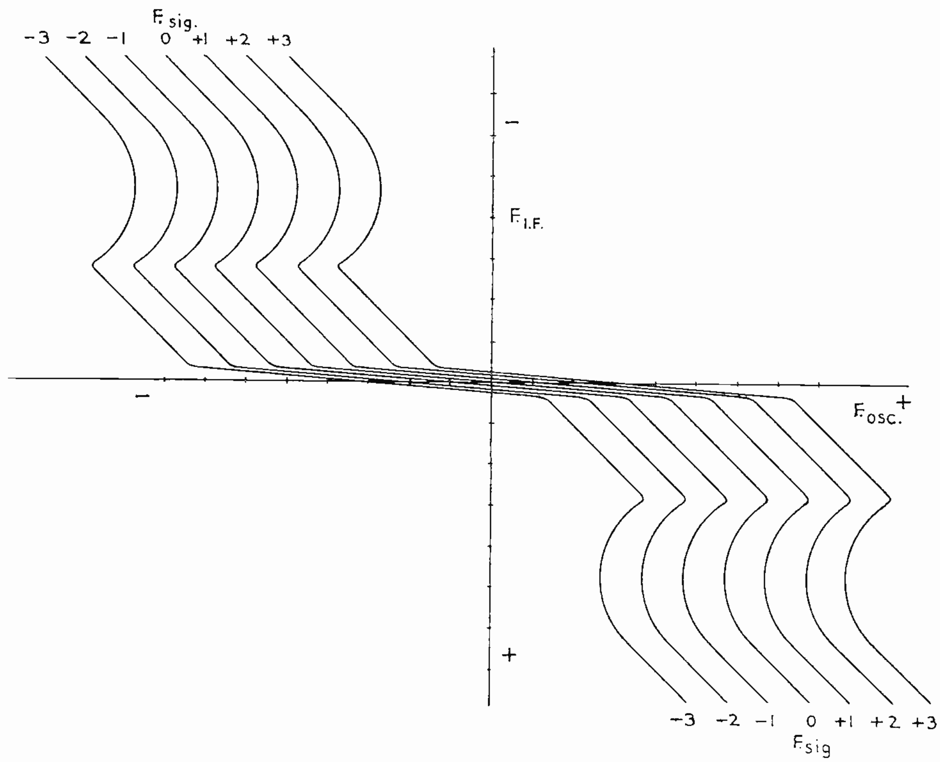


FIG. 6.

will, as F_I is reduced, become tangential to the characteristic at the point A and cut it in the straight portion of lesser slope at B. This position corresponds to a jump in F_o towards zero. Further reduction of F_I results in F_o reducing at the same rate as F_I at first, and then as the portion of characteristic of greater slope is entered, F_o has nearly reached zero and the reduction is proportional to the reduction of F_I. This continues until $F_I = F_o = 0$. If now F_I increases in the negative direction F_o also increases (but very slowly in comparison) in the negative direction until the less steep portion of the charcteristic is reached, when F_o increases at the same rate as F_I. At the end of this portion of lesser slope F_o increases suddenly in a jump to a value $F_o = F_I$ from C to D in the diagram. If F_I is initially large or negative the curve is covered in the reverse direction and the jumps occur from P to Q and then from R. to S. It is shown below that the points at which these jumps occur must also be the points at which jump-in or jump-out occurs in an actual circuit.

So far the discussion has been quite general and the results obtained may equally as well be applied to a valve circuit (working under the conditions of unity overall magnification) in which non-linear back-coupling is employed by replacing F_o by E, the output voltage, F_I by e_g the applied voltage, and f by βE, the voltage fed back. Note here also that when β is non-linear it is in general simpler from the point of view of the graphical construction to use not β but βE since the Bartels* construction apparently breaks down when $\beta = 0$.

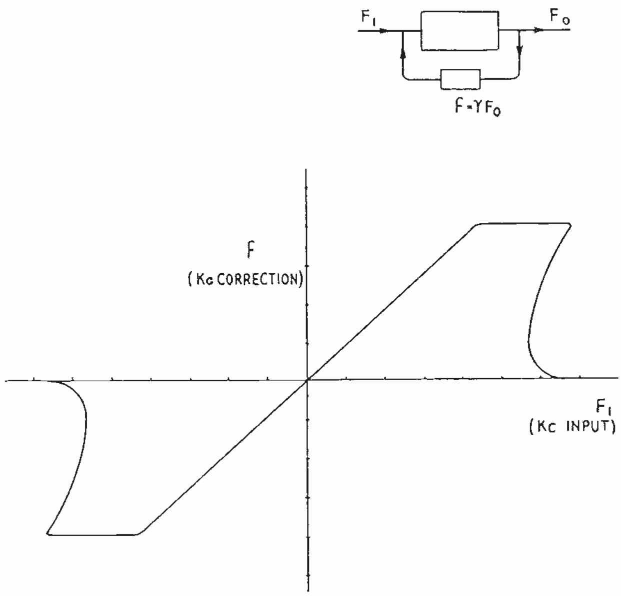


FIG. 7.

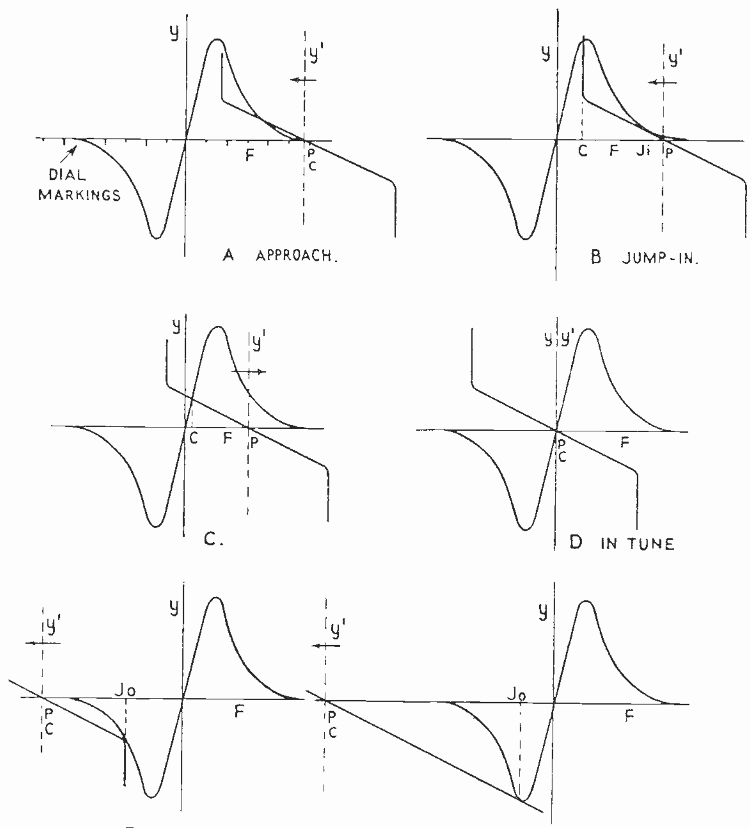
Now F_I is the departure of the initial I.F. from the predetermined or designed I.F., and this I.F. is in turn given by the difference between the local oscillator frequency and the signal frequency, both of which are capable of variation. Hence F_I may be the result of a variation of oscillator or signal or both. If, therefore, we put F_o equal to deviation in I.F. ($F_{I.F.}$), F_I equal to deviation in oscillator frequency (F_{osc}) from the true tune frequency, we may draw a family of $F_{I.F.} - F_{osc}$ curves with F_{sig} as parameter (similar to the $I_a - E_a$ characteristics of a valve with e_g at parameter) for the backcoupled (frequency corrected) condition.

These curves will be of similar shape to the $F_I = F_o + f$ characteristic. For this purpose, however, it is convenient to represent F_{IF} ($= F_o$) as ordinate and F_{osc} ($= F_I$) as abscissa (Fig. 6).

We may also represent the relationship in another way by putting frequency fed back (f) against input frequency (or initial tuning error F_I). Such a curve is

* H. Bartels: Linear and Non-Linear Backcoupling. E.N.T. 11, H.9. (1934).

shown in Fig. 7. As before (Fig. 5) as F_1 is reduced a jump occurs in f from a low to a high value (at the point corresponding to where the gradient of the curve is infinite in this case). The value of F_1 is of course the same whether the construction of Fig. 5 or Fig. 7 is adopted. It will also be observed in Fig. 7 that the slope of the curve through the origin tends to approach 45° as the degree of feed-back increases.



E JUMP-OUT. (LIMITED CONTROL CHARACTERISTIC.)

F JUMP-OUT (NON-LIMITED CONTROL CHARACTERISTIC)

y Discriminator output axis.
y^1 Control input axis (after rotation).
J_1 Jump-in frequency.
J_0 Jump-out frequency.
p Actual pointer position.
c Equivalent pointer position (or frequency to which circuits adjust themselves).

Arrow on y' axis indicates direction of tuning.

FIG. 8.

It will be observed that both the curves Fig. 5 and Fig. 7 possess the property of being what may be termed re-entrant, and but for this property the phenomena of jump-in and jump-out would not occur. A and P represent the points of jump-in and C and R the points of jump-out. The performance of a receiver employing A.F.C. circuits is completely determined from either of these curves in the manner indicated (together with the relation $F_1 = F_0 + f$), but the curves are not in a form suitable for designs purposes in so far as they represent the combined action of two separate units, and the effect of alteration of one unit alone demands that a new combined curve be constructed. A method will now be described that obviates this difficulty. No originality is claimed since it has been proposed before,* but the method seems to have been neglected in recent literature. Briefly, the method consists of using the voltage (or current) output-frequency characteristic of the discriminator together with the frequency-voltage (or current) characteristic of the control device in the following manner. The control characteristic (preferably drawn on transparent paper) is rotated or inverted and placed over the discriminator characteristic so that the frequency axes are coincident (the control characteristic being, of course, in the correct sense in relation to the discriminator characteristic). This characteristic may now be slid over the discriminator characteristic keeping the frequency axes coincident. The point of intersection of the control characteristic with the frequency scale of the discriminator characteristic represents the frequency error that would exist as a result of initial mal-adjustment of the circuits without frequency correction, while the point or points of intersection of the control characteristic with the discriminator characteristic when projected on to the discriminator frequency scale give the frequency error existing after the circuits have adjusted themselves due to the frequency correction. The significance of more than one point of intersection is dealt with below. When the characteristics are tangential a point of jump-in or jump-out is indicated, the initial frequency setting and the corrected frequency before and after jump-in (or jump-out) being indicated by the curves. We may express this in another way as follows. The frequency error axis of the discriminator curve may be regarded as the dial or tuning scale of the receiver and suitable dial markings made thereon. These will, of course, vary, depending upon the particular station represented by the discriminator curve. The now vertical voltage (or current) axis of the rotated control curve may be regarded as the pointer which moves over the dial. The projection of the appropriate intersection of control and discriminator characteristics then gives both the frequency error after the circuits have adjusted themselves and also the equivalent position of the pointer, i.e., the position to which the pointer would have to be set in the absence of frequency correction to obtain the same accuracy of tuning. Fig. 8 illustrates the procedure.

This method, originally devised in connection with the application of A.F.C. to a transmitter, is of the utmost practical value to the receiver designer. One may set out a family of discriminator characteristics taken at different input frequencies and levels (the curves thus being "overall" characteristics, taking into account the selectivity of preceding circuits, oscillator efficiency and modification due to A.V.C., etc.), and then estimate the shape of a control characteristic which will provide the best compromise in the range to be considered. It is then necessary to design control devices or circuits that will fulfil this requirement. Further, by extending the horizontal discriminator axis, drawing in curves representing adjacent channel stations, adding the ordinates to form a single curve and then slid'ng the control curve along the axis, useful information may be obtained, though it must be remembered that the effect of the time constant of the A.V.C. circuits is not taken into account.

* Kusonose and Ishikawa : Proc. I.R.E. 20, p. 310.

Considering this construction in the light of the feedback analysis above in which it is fundamental that $F_1 = F_0 + f$ or the initial frequency setting equals the final frequency plus the correction, let the relationship between discriminator output (y) and frequency error (x) be represented by $y = f(x)$ and the relationship between correction (z) and control device input (y) by $z = F(y)$. Now if the system originally assumed to be in correct tune (so that there is no output from the discriminator) is detuned slightly, the error after it has attained equilibrium may be some value x kc. (Fig. 9); the voltage (or current) output from the discriminator will, therefore, be $y = f(x)$ and this voltage is producing a correction z given by $z = F(y)$. The amount of detuning which would be apparent without feedback would, therefore, be $x + z$ corresponding to F_1 in the feedback equation. We thus have that $F_1 = x + z$ = setting or initial error, $f = z$ = correction and $F_0 = x$ = final error. This is, of course, the same thing as sliding the rotated (and if necessary inverted) control curve over the discriminator curve in the manner described above. A rigorous proof of the validity of this curve sliding technique is given later, and it is sufficient here to say that experimental checks under a variety of conditions confirm this.

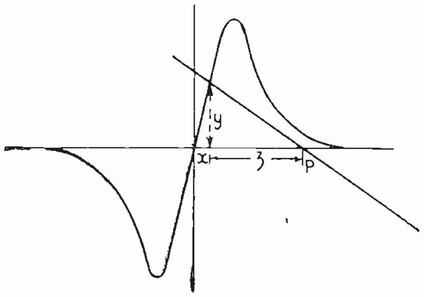


FIG. 9.

The experimental checks were taken in the following manner. An intermediate frequency of 80 kc. was used together with a normal discriminator. To avoid possibilities of error in measurement due to the high internal impedance of the discriminator output source, the current output rather than the voltage output of this circuit was employed, a milliammeter being inserted across the extreme ends of the discriminator diode load resistances. The discriminator characteristic then is expressed in terms of current output against frequency deviation or error (input).

The control device consisted of what was termed a capacity meter, an instrument which is essentially a milliammeter with scale removed in which the pointer is replaced by a light metal vane which moves between two suitably positioned fixed plates so as to form a small variable condenser. The construction of the instrument, which was made to our requirements by Messrs. Elliott Bros. (London) Ltd., will be apparent from the photograph of Fig. 10, showing case removed. To the right are seen the two fixed spaced plates insulated from the remainder of the instrument, whilst the moving vane is in its zero position half in mesh with the fixed plates. The moving vane itself is a silvered mica sheet held by means of a tongue on the pointer arm, which is in electrical connection with the side of the operating winding

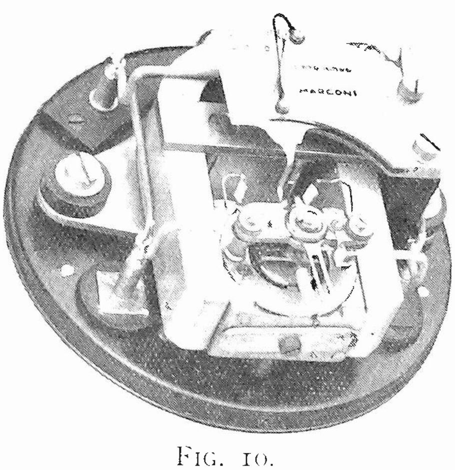


FIG. 10.

normally earthed ; a suitable scale is marked on the vane and an edgewise metal strip attached to the fixed plates facilitates reading the same. This scale is of assistance for zero setting and subsidiary calibration purposes, the main calibration being obtained by relating either capacity or frequency to current input measured externally. The capacity change available was $\pm$ 6 $\mu\mu f$. For the purposes of the experiments, the calibration was obtained by observing oscillator frequency change against current input to the capacity meter which was effectively in parallel with the oscillator tuning condenser.

The operating winding of the capacity meter was inserted in series with the discriminator output meter and could be shorted out if required. A small high quality trimmer condenser with slow motion dial attached was used across the oscillator tuned circuit, and although calibrated against frequency, it was found more convenient to express both discriminator or control characteristics against dial reading rather than frequency deviation. The results obtained checked against those predicted by the curve sliding technique to within a few cycles. The slight errors observed could well be accounted for by meter friction.

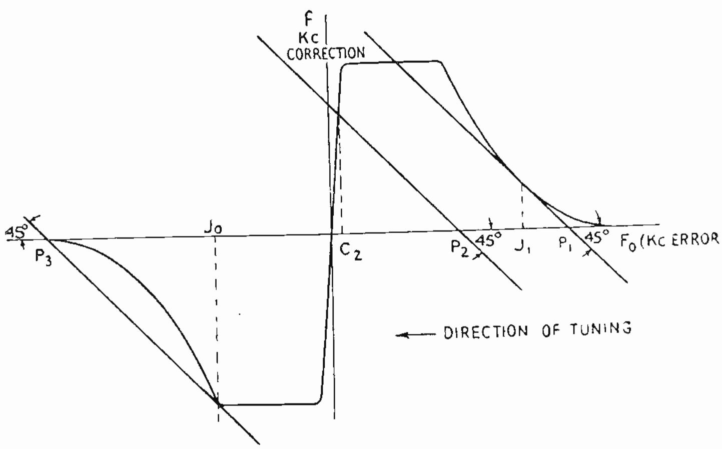
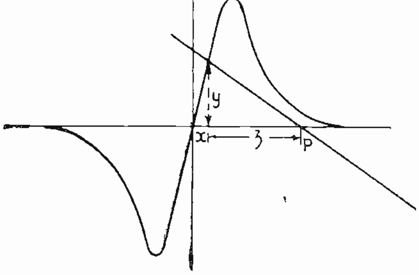


FIG. 11.

As a result of the analytical investigation, not only is the correctness of the curve sliding method confirmed but a new method of greater value analytically is introduced. This method, although more fundamental than the curve sliding method described, is in our opinion not of such practical value, inasmuch as the rôles played by the discriminator and control device individually are not made apparent. If one plots kc. correction against kc. error, i.e., $z = F\ (y) = F\ (f(x)) = \phi\ (x)$ a 45° line slid along the kc. error axis will determine the circuit performance, for the intersection of the line with the kc. error axis determines the dial setting or initial (tuning) error, while the appropriate intersection of the line with the curve yields both the amount of correction which is taking place and the final frequency error. Examples are shown in Fig. 11. It will be observed that Fig. 7 is similar to this but skewed 45°.

The analytical investigation now follows, and in a concluding note additional experimental evidence is adduced in confirmation of its correctness.

Considering this construction in the light of the feedback analysis above in which it is fundamental that $F_1 = F_0 + f$ or the initial frequency setting equals the final frequency plus the correction, let the relationship between discriminator output (y) and frequency error (x) be represented by $y = f(x)$ and the relationship between correction (z) and control device input (y) by $z = F(y)$. Now if the system originally assumed to be in correct tune (so that there is no output from the discriminator) is detuned slightly, the error after it has attained equilibrium may be some value x kc. (Fig. 9); the voltage (or current) output from the discriminator will, therefore, be $y = f(x)$ and this voltage is producing a correction z given by $z = F(y)$. The amount of detuning which would be apparent without feedback would, therefore, be $x + z$ corresponding to F_1 in the feedback equation. We thus have that $F_1 = x + z$ = setting or initial error, $f = z$ = correction and $F_0 = x$ = final error. This is, of course, the same thing as sliding the rotated (and if necessary inverted) control curve over the discriminator curve in the manner described above. A rigorous proof of the validity of this curve sliding technique is given later, and it is sufficient here to say that experimental checks under a variety of conditions confirm this.



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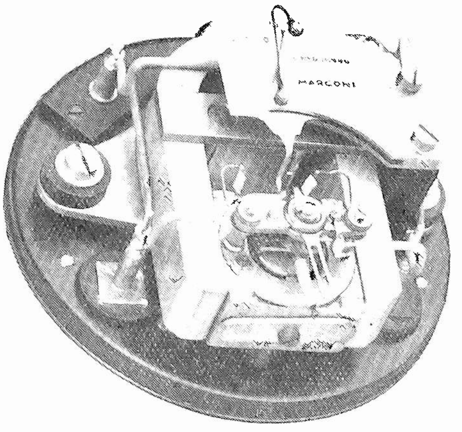

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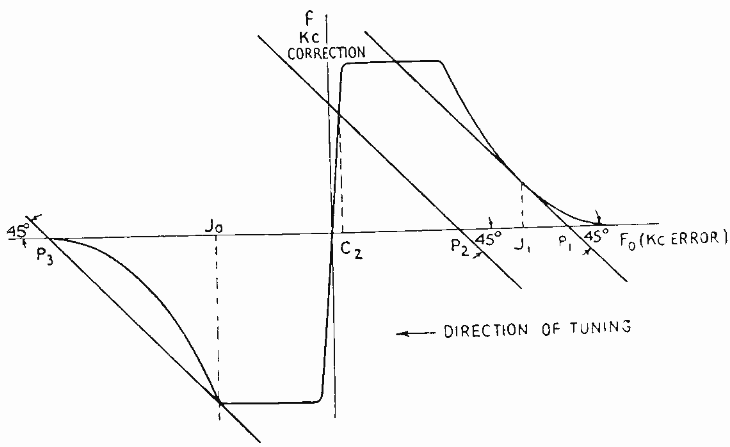

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The analytical investigation now follows, and in a concluding note additional experimental evidence is adduced in confirmation of its correctness.

Analysis of Automatic Frequency Correction.

The essential elements of the device are :—

1. A circuit operated from a special I.F. network such that it develops a D.C. voltage when the injected I.F. is off the tune point of the network. This voltage may be regarded as a function of the I.F.
2. A circuit by which this voltage may be made to control the frequency of the heterodyne oscillator and hence the I.F. It is assumed that the I.F. is produced by heterodyning a fixed incoming signal frequency so that changes in heterodyne frequency represent actual changes in I.F.

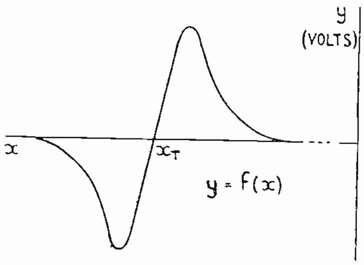


FIG. 12.

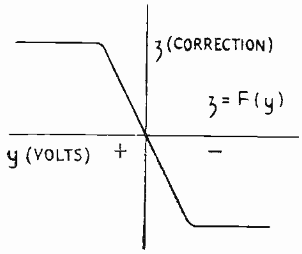


FIG. 13.

The two circuits thus interact, and the success of the device depends upon the attainment of a state of stable equilibrium in which the I.F. is changed from an initial value to some final value which produces the necessary voltage required to maintain the change in frequency involved.

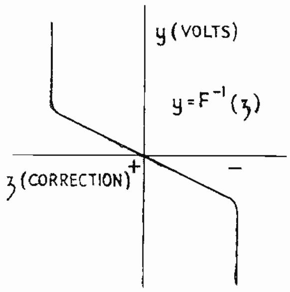


FIG. 14.

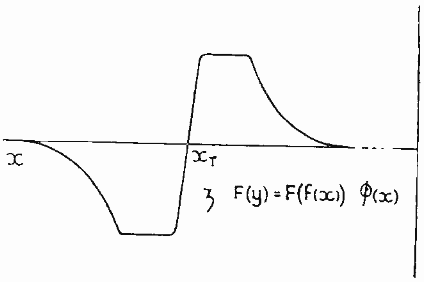


FIG. 15.

In order to retain the same curve shape as that hitherto used, the signs of the axes have in some cases been reversed. Further, frequency x has been taken as actual frequency, not frequency deviation from true tune.

In analysing the process we consider first that the two circuits are separate. If the I.F. is held at some particular value x and the voltage developed is not fed back to the heterodyne oscillator, this voltage will take up a value y which can be represented as some function of x. Say $y = f(x)$. In Fig. 12 a typical curve for

$f(x)$ is shown diagrammatically. The curve is usually adjusted so that $y = 0$ when $x = x_T$ the tune point of the I.F. circuit, and is symmetrical about the point x_T, y changing sign as we pass through x_T.

Similarly, if we have a fixed voltage y and feed it by the control circuit on to a similar but separate heterodyne oscillator, it will produce a change of frequency z which is a function of y, say $z = F(y)$. In Fig. 12 a typical curve is shown. z is usually zero when y is zero.

Now from these two functions we can derive two new functions. Firstly we can invert the curve in Fig. 13 and draw y as a function of z, which may be written $y = F^{-1}(z)$. This is shown in Fig. 14.

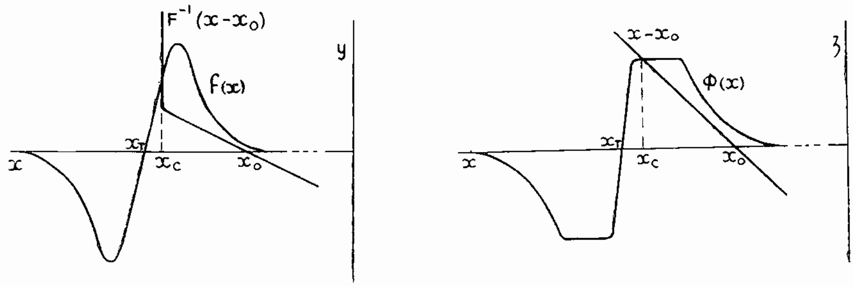


FIG. 16.

Secondly we may combine the two graphs and by writing $z = F(y) = F(f(x)) = \phi(x)$ we can plot a graph of z as a function of x, where it must be remembered that to get z, x is operated on by f first and then by F. We get a graph as in Fig. 15.

Now when the circuits are made to interact we may assume that the heterodyne oscillator is fixed initially so that the I.F. has some value x_0. If a position of equilibrium is reached when $x = x_c$ $z = z_c$ and $y = y_c$ then we must have

$$z_c = x_c - x_0$$

$$y_c = f(x_c)$$

$$z_c = F(y_c)$$

$$\therefore \quad y_c = F^{-1}(z_c) = F^{-1}(x_c - x_0)$$

$$\text{and} \quad z_c = x_c - x_0 = \phi(x_c).$$

The values of x_c and y_c are therefore given by the point where the curves $y = f(x)$ and $y = F^{-1}(x - x_0)$ cut when plotted together on the y x axes; or by the intersection of the curves $z = x - x_0$ and $z = \phi(x)$ when plotted together on the z x axes as in Fig. 16.

The curve $F^{-1}(x - x_0)$ is obtained by sliding the curve $F^{-1}(z)$ along until it passes through x_0; while the curve $z = x - x_0$ is simply the straight line of slope 45° through the point x_0.

From the point of view of determining graphically the position of x_c the two methods are equivalent. The former one in which the $F^{-1}(z)$ curve is moved along is the more convenient in practice, and the sliding along of the curve is equivalent to adjusting the initial value x_0 of the I.F. by adjusting the initial value of the heterodyne oscillator. The second method is very useful analytically.

In sliding the curve along it is obvious that there are several types of intersection of the curves. These are shown in Fig. 17.

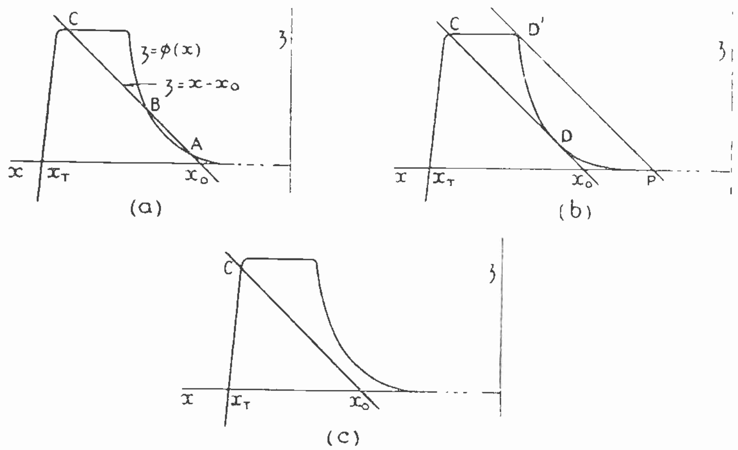


FIG. 17.

The curves are drawn for z against x, although it is obvious that similar diagrams can be drawn with y against x. In (*a*) we have three possible points A, B, C. In (*b*) tuning towards the station, A and B have come to the limiting position D, or tuning away from the station, B and C have come to the limiting position D′. In (*c*) C alone is possible.

Now it is known in practice that if we have case (*a*) and start from x_0 we come to stable equilibrium at A. B is essentially unstable and as we alter x_0, i.e., slide the curve, we come to case (*b*) where D is a neutral point beyond which the smallest increase of x_0 takes us immediatlely over to C (jump-in) as in case (*c*). Tuning away from the station, D′ is the neutral point beyond which any further mistuning results in jump-out to P. This point is generally on the axis, outside the range of the characteristic.

We seek to find some means of determining theoretically how the I.F. changes from x_0 to the stable points A or C, and to show that they actually are stable points.

First of all a graphical method suggests itself. Suppose we start at the point x_0. If the circuits respond instantaneously, this will mean that we develop a voltage y_0 given by

$$y_0 = f\,(x_0)$$

But these volts y_0 will immediately produce a change of frequency z_0 given by

$$z_0 = F\,(y_0) = \phi\,(x_0)$$

so that the I.F. takes up a new value x_1 given by $x_1 = x_0 + z_0 = x_0 + \phi\,(x_0)$. But now the voltage will change to y_1 and z will change to z_1 giving a new I.F. x_2 where

$$y_1 = f\,(x_1), \quad z_1 = F\,(y_1) \text{ and } x_2 = x_0 + z_1$$

and so on.

The I.F. jumps to a series of values x_0, x_1, x_2, etc. Now we can write

$$x_1 = x_0 + \phi\,(x_0)$$
$$x_2 = x_0 + \phi\,(x_1)$$
$$\cdots\cdots\cdots$$
$$x_n = x_0 + \phi\,(x_{n-1})$$

so that if this step-by-step process does give a convergent series, x_n must approach a value x_c such that x_n when n is very large $= x_c$ while $\phi\,(x_{n-1}) = \phi\,(x_n) = \phi\,(x_c)$.

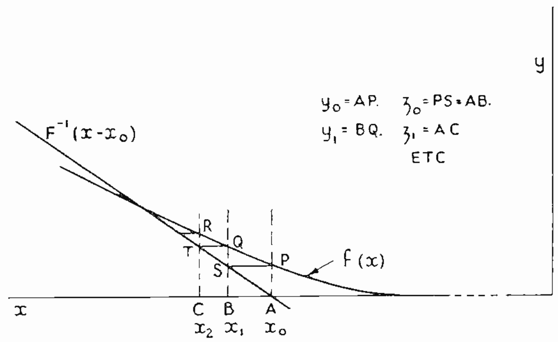


FIG. 18.

Thus $\qquad x_c = x_0 + \phi\,(x_c)$

which agrees with the condition given above. Thus if the process does lead to a convergent result it will lead to the correct convergent, and moreover if we want to find the convergent we do not have to go through the labour of calculating the

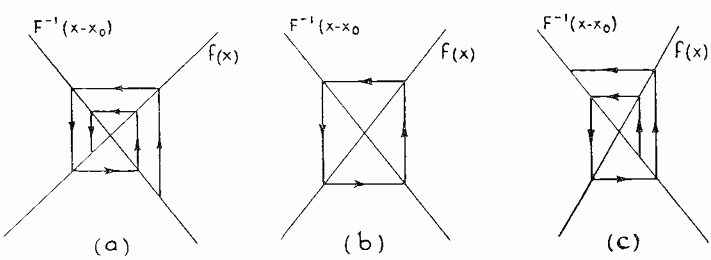


FIG. 19.

successive convergents, since we see that the final result must be the same as that given by the direct graphical method.

To see whether our assumption that the step-by-step process does converge is justified we can represent the process graphically (Fig. 18). We can use either the y, x or z, x curves. Using the $y\ x$ curves let A be the point (x_0 .o), then in the figure AP $= y_0$. We draw PS across to cut the control curve (inverted) in S. Then since BS $= y_0$, AB $= z_0$ and B is therefore x_1. We now have B Q $= y_1 =$ CT and AC $= z_1$ and C is the point x_2 and so on, i.e., starting at x_o we draw up to the curve $f\,(x)$ then across to the curve $F^{-1}\,(x - x_0)$ then up to $f\,(x)$ again and so on.

Now this process shows that for a point of the type A (Fig. 17) where the $f(x)$ curve is less steep than the $F^{-1}(x - x_0)$ curve we always converge on to the point, while if we take up the process in the neighbourhood of a point of the type B, we move away from it.

But for a point of the type C we jump from side to side and we get nearer and nearer or further and further away according as the slope of $f(x)$ is less steep or more steep than that of $F^{-1}(x - x_0)$ (where we take the numerical value of the slope).

This can be seen from Fig. 19 where (*a*) represents the convergent case where $f(x)$ is less steep than $F^{-1}(x - x_0)$ (*b*) is the intermediate case where the slopes are equal and (*c*) is the divergent case where $f(x)$ is steeper than $F^{-1}(x - x_0)$.

The results for types A and B agree with practice, while the result for C is at variance, since in practice type C is a stable condition whatever the slope and is only approached in an oscillatory way when the slope of the $f(x)$ curve relative to the $F^{-1}(x - x_0)$ curve exceeds a certain value.

It is obvious that the graphical step-by-step method breaks down because we have ignored the time-factor. The frequency and voltage values cannot alter in a series of instantaneous finite jumps, and although for type A the graphical method gives a rough idea of how the final equilibrium is obtained, the picture for type C where the jumps may increase indefinitely in value is completely wrong.

To analyse the process we must therefore write down the differential equations taking into account the time constants of the circuits, and we can then study by solving the equations graphically exactly how the frequency does change from x_0 to x_c and can study in detail the nature of the equilibrium for types A, B, C and D, i.e., assuming we are at one of these points and disturb the conditions slightly we can tell whether we will return to the point again or diverge from it.

Suppose that at any time t the voltage which is being fed back to the oscillator control circuit is y and that the frequency of the oscillator is such that with the given station frequency the resultant I.F. is x.

If x_0 is the I.F. corresponding to the frequency at which the oscillator regarded by itself is set, the change in frequency z is given by

$$z = x - x_0 \qquad \ldots \qquad (1)$$

If the frequency could be held at x the voltage y would alter and approach exponentially a final value of $f(x)$, the exponential time factor being of the form $exp(-k_1 t)$ where k_1 depends on the time constant of the circuit involved.

Actually therefore the rate of change of y with respect to time can be represented by the equation

$$\frac{dy}{dt} = k_1(-y + f(x)) \qquad \ldots \qquad (2)$$

since y would be given by $f(x)(1 - e^{-k_1 t})$ if $f(x)$ were to remain constant. This equation replaces the relation $y = f(x)$ of the "step-by-step" graphical method.

Similarly, if y were to remain fixed the change of frequency produced by the control circuit would approach exponentially a final value of F (y) and we may write

$$\frac{dz}{dt} = k_2 \left(- z + \mathrm{F}(y)\right) \quad \ldots \quad (3)$$

instead of $z = \mathrm{F}(y)$ by the graphical method.

From (1) and (3) we get

$$\frac{dx}{dt} = k_2 \left(- (x - x_0) + \mathrm{F}(y)\right) \quad \ldots \quad (4)$$

Thus instead of the "step-by-step" process we actually move along a $(y\ x)$ curve obtained by eliminating the time t from the solutions of (2) and (4).

We now define a critical point $(x_c\ y_c)$ as one for which $\frac{dx}{dt}$ and $\frac{dy}{dt}$ are simultaneously zero.

Thus from (2) and (4) we have

$$y_c = f(x_c) \text{ and } x_c - x_0 = \mathrm{F}(y_c)$$

$$\therefore \quad x_c - x_0 = \mathrm{F}(f(x_c)) = \phi(x_c)$$

$$\therefore \quad x_c = x_0 + \phi(x_c) \quad \ldots \quad (5)$$

These critical points are therefore the same as those found graphically. We need therefore to study the values of x and y in the region of these critical points.

In order to study the way in which (x, y) gets in the region of a critical point we would have to be able to solve (2) and (4) for any form of $f(x)$ and F (y) for any given initial conditions. This is in general an impossible task analytically,* but we may suppose that in course of time $(x\ y)$ gets into the region of $(x_c\ y_c)$ and that at some moment which we may take as the origin of time $x = x_s$ and $y = y_s$ where the point $x_s\ y_s$ is in the region of $(x_c\ y_c)$.

Then we may replace the functions $f(x)$ and F (y) by the tangents to these functions at the point $(x_c\ y_c)$, in fact we can define the region of $(x_c\ y_c)$ as those points for which we can consider this approximation to be valid. $(x_s\ y_s)$ can be any point within this region.

We therefore replace $f(x)$ by

$$f(x) = (x - x_c) \tan\theta_1 + y_c \quad \ldots \quad (6)$$

where θ_1 is the slope of the tangent at $(x_c\ y_c)$
and we replace F (y) by

$$\mathrm{F}(y) = (y - y_c) \tan\theta_2 + (x_c - x_0) \quad \ldots \quad (7)$$

where θ_2 is the slope of the control curve at $(x_c\ y_c)$ when y is taken as abscissæ.

Thus	when $x = x_c$	$f(x) = y_c$
and	when $y = y_c$	$\mathrm{F}(y) = x_c - x_0$

as they should do.

* The actual way by which $(x\ y)$ gets into the region $(x_c\ y_c)$ can be studied by solving the equations for any given values of k_1 and k_2 by approximate graphical means.

Putting these values in (2) and (4) gives

$$\frac{dy}{dt} = k_1 (-y + (x - x_c) \tan \theta_1 + y_c) \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

$$\frac{dx}{dt} = k_2 (-(x - x_0) + (y - y_c) \tan \theta_2 + x_c - x_0) \quad \ldots \quad \ldots \quad (9)$$

We now change the origin to $x_c\ y_c$ by writing

$$x^{\mathrm{I}} = x - x_c \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

$$y^{\mathrm{I}} = y - y_c \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

so that $\dfrac{dx^{\mathrm{I}}}{dt} = \dfrac{dx}{dt}$ and $\dfrac{dy^{\mathrm{I}}}{dt} = \dfrac{dy}{dt}$

so that (8) and (9) become

$$\frac{dy^{\mathrm{I}}}{dt} = k_1 (-y^{\mathrm{I}} + x^{\mathrm{I}} \tan \theta_1) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

$$\frac{dx^{\mathrm{I}}}{dt} = k_2 (-x^{\mathrm{I}} + y^{\mathrm{I}} \tan \theta_2) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

Differentiating (13) we get

$$\frac{d^2x^{\mathrm{I}}}{dt^2} = k_2 \left(-\frac{dx^{\mathrm{I}}}{dt} + \tan \theta_2 \frac{dy^{\mathrm{I}}}{dt}\right)$$

and substituting for $\dfrac{dy^{\mathrm{I}}}{dt}$ from (12)

$$\frac{d^2x^{\mathrm{I}}}{dt^2} = -k_2 \cdot \frac{dx^{\mathrm{I}}}{dt} + k_2 \tan \theta_2 (-k_1 y^{\mathrm{I}} + k_1 x^{\mathrm{I}} \tan \theta_1)$$

$$= -k_2 \cdot \frac{dx^{\mathrm{I}}}{dt} - k_1 k_2 \tan \theta_2 \cdot y^{\mathrm{I}} + k_1 k_2 \tan \theta_1 \tan \theta_2 \cdot x^{\mathrm{I}}$$

and putting in the value of $k_2 \tan \theta_2 \cdot y^{\mathrm{I}}$ from (13) we have finally

$$\frac{d^2x^{\mathrm{I}}}{dt^2} + (k_1 + k_2) \cdot \frac{dx^{\mathrm{I}}}{dt} + k_1 k_2 (1 - \tan \theta_1 \tan \theta_2)\, x^{\mathrm{I}} = 0.$$

Similarly

$$\frac{d^2y^{\mathrm{I}}}{dt^2} + (k_1 + k_2) \cdot \frac{dy^{\mathrm{I}}}{dt} + k_1 k_2 (1 - \tan \theta_1 \tan \theta_2)\, y^{\mathrm{I}} = 0.$$

Thus both equations can be summarised by

$$(D^2 + 2aD + b)\, \psi = 0 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

where ψ is x^{I} or y^{I}

$$\text{and } a = \frac{(k_1 + k_2)}{2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

$$\text{and } b = k_1 k_2 (1 - \tan \theta_1 \tan \theta_2) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

From (14) we have for the operator D

$$D = -a \pm \sqrt{a^2 - b}$$

$$= -a \pm c$$

where $$c = \sqrt{a^2 - b} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

Thus the solution of (14) is

$$= e^{-at} (Ae^{ct} + Be^{-ct}) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (18)$$

where the values of A and B for x^1 and y^1 are found by putting $x^1 = x_s - x_c$ and $y^1 = y_s - y_c$ when $t = 0$ in the expressions for x^1 and y^1 from (18) and in the values of $\frac{dx^1}{dt}$ and $\frac{dy^1}{dt}$ in (12) and (13).

We do not, however, need to consider the actual values of A and B, since we wish to show whether as time progresses ψ approaches zero or diverges from it for any starting point $(x_s\ y_s)$ we may like to choose within the region of $(x_c\ y_c)$.

From the form of (18) we have the following general results.

Remembering from (15) that a is essentially positive since k_1 and k_2 are positive, we have

Case (1).

When c is real and less than a, then ψ is a decaying non-oscillatory function. As $t \rightarrow \infty$, $\psi \rightarrow 0$.

Case (2).

When c is real and greater than a, then ψ is a divergent function unless A happens to be zero.

Case (3).

When c is real and equal to a, then either $\psi \rightarrow 0$ or some constant value. Due to practical considerations this case actually passes over to a divergent condition.

Case (4).

When c is zero, ψ is a decaying function with critical damping.

Case (5).

When c is imaginary then ψ is a decaying oscillatory function. As $t \rightarrow \infty$, $\psi \rightarrow 0$.

These cases will correspond to different conditions for the value of $\tan\theta_1 \tan\theta_2$. The possibilities are :—

(a) $\tan\theta_1 \tan\theta_2$ is positive and < 1

Then from (16) b is positive and $< k_1 k_2$

also since $k_1 k_2 \leqslant \left(\frac{k_1 + k_2}{2}\right)^2$

by virtue of the relation $\left(\frac{k_1 + k_2}{2}\right)^2 - k_1 k_2 = \left(\frac{k_1 - k_2}{2}\right)^2$

$\therefore \qquad b < a^2$

Hence $c = \sqrt{a^2 - b}$ is real and less than a. We therefore have Case (1).

(b) $\tan\theta_1 \tan\theta_2$ is positive and > 1

$\therefore \qquad b$ is negative and c is real and greater than a.

We therefore have Case (2).

(c) $\tan\theta_1 \tan\theta_2$ is positive and $= 1$ so that $b = 0$, $c = a$ and we have Case (3).
(d) $\tan\theta_1 \tan\theta_2$ is negative so that b is positive.

Three cases arise

(d.1) $\quad b < a^2$ corresponding to $|\tan\theta_1 \tan\theta_2| < \frac{1}{k_1k_2}\left(\frac{k_1 - k_2}{2}\right)^2$

then c is real and less than a, i.e. we have Case (1).

(d.2) $\quad b = a^2$ corresponding to $|\tan\theta_1 \tan\theta_2| = \frac{1}{k_1k_2}\left(\frac{k_1 - k_2}{2}\right)^2$

then c is zero and we have Case (4).

(d.3) $\quad b > a^2$ corresponding to $\tan\theta_1 \tan\theta_2 > \frac{1}{k_1k_2}\left(\frac{k_1 - k_2}{2}\right)^2$

then c is imaginary and we have Case (5).

Now for our values of $F(y)$ and $f(x)$ in the region $(x_c\ y_c)$ the function

$$\phi(x) = F(f(x)) \text{ of the graphical method becomes}$$
$$\phi(x) = [(x - x_c)\tan\theta_1 + y_c - y_c]\tan\theta_2 + x_c - x_0$$
$$= (x - x_c)\tan\theta_1 \tan\theta_2 + x_c - x_0$$

which gives $\phi(x) = x_c - x_0$ as it should do when $x = x_c$ from (5). Thus $\tan\theta_1 \tan\theta_2$ is the slope of the $\phi(x)$ curve in the region $(x_c\ y_c)$.

Referring to Fig. 17, for a point of intersection such as

A	$\tan\theta_1 \tan\theta_2$ is $+ ve.$ and < 1
B	is $+ ve.$ and > 1
D,D1	is $+ ve.$ and $= 1$
C	is $- ve.$

We may conveniently summarise these results in tabular form as shown below.

	$\tan\theta_1 \tan\theta_2$	Case	Intersection Type	Mode of Operation
a	$+ve < 1$	1	A	Decaying, non-oscillatory. Stable.
b	$+ve > 1$	2	B	Divergent. Unstable.
c	$+ve = 1$	3	D,D1	Divergent in practice. Jump point.
d.1	$-ve < \frac{1}{k_1k_2}\left(\frac{k_1 - k_2}{2}\right)^2$	1	C	Decaying, non-oscillatory. Stable.
d.2	$-ve = \frac{1}{k_1k_2}\left(\frac{k_1 - k_2}{2}\right)^2$	4	C	Decaying, critical damping. Stable.
d.3	$-ve > \frac{1}{k_1k_2}\left(\frac{k_1 - k_2}{2}\right)^2$	5	C	Decaying, oscillatory. Stable.

The analysis thus bears out the graphical method, showing that a type A intersection is essentially stable and non-oscillatory, while a type B is essentially unstable, while as suggested, the introduction of the time factor renders a type C stable and actually non-oscillatory for slopes below a certain value, and stable though oscillatory for any slope greater than this.

The possibility in the divergent case of the constant A being equal to zero is essentially a case of unstable equilibrium, and corresponds to initial conditions which just happen to bring it to rest on the critical point B. Any subsequent disturbance, however small, will cause divergence. It is analogous to a rod swinging about a horizontal axis through one end, which is given a blow which just happens to bring it to rest in a position vertically above the axis. Any subsequent displacement will cause it to fall down. In practice receiver noise would be sufficient to make this point impossible of attainment. It may be emphasised again that the analysis deals with the subsequent history of events once $(x\ y)$ has approached within the region (as defined above) of a critical point $(x_c\ y_c)$. Given this condition, which on physical principles must in general arise the analysis shows under what conditions $(x\ y)$ proceeds to converge on $(x_c\ y_c)$.

The particular case for $\tan\theta_1 \tan\theta_2 = 1$, corresponding to points of the D type, which is of considerable practical interest and which has not so far been dealt with in detail, results in $b = o$, $c = a$, so that $\psi = A + Be^{-2at}$. When $t = \infty$, $\psi = A$. We thus in general converge to the point (A_x, A_y) instead of (o, o) where the values of A_x and A_y depend upon the values of x_s and y_s chosen as the starting point in the region of $(x_c\ y_c)$. When, however, $\tan\theta_1 \tan\theta_2 = 1$ it implies that the slope of $\phi\ (x)$ is 45°, i.e., parallel to the 45° line through x_o. But since we have assumed that the lines cut in $x_c\ y_c$, they must coincide and we have not a critical point but a critical line. The analysis shows in fact that the values of A_x and A_y are such that the convergent point is always on this 45° critical line.

Equilibrium may thus be attained at any point in this line; movement to a portion of characteristic of lesser slope will result in a return to this line, whereas movement to a portion of greater slope results in movement away from the line and equilibrium is attained at some point of the C type. In practice points of the D type are essentially unstable on account of receiver noise and a jump occurs either to a point of the C type (when tuning towards a station) or to some point on the extreme skirt of the characteristic or more generally outside the range of the characteristic altogether (when tuning away from a station).

The oscillatory state for points of C type, dealt with above, has been observed during the course of experiments. When using the very steep curve resulting from the use of a quartz crystal discriminator, the oscillations persisted. This persistence may have resulted from the use of the capacity meter, which, of course, possesses a certain inertia, or it may imply that where the oscillations are lightly damped, circuit noise may be sufficient to maintain them. The most probable cause, however, was "ringing" due to the quartz itself. In any case it would appear that there is a practical limit beyond which it is inadvisable to increase the goodness of the circuits employed.

O. E. KEALL.
G. MILLINGTON.

COMPARISON OF PARALLEL AND SERIES COUPLING CIRCUITS FOR TRANSMITTERS

This article, the first part of which appeared in THE MARCONI REVIEW, No. 68, *p.* 31, *deals with the load circuit of transmitters using inductive coupling. The article is concluded in the present issue.*

Comparison of Series and Parallel Circuit.

AFTER this preliminary survey of the properties of the parallel circuit we can proceed with the comparison of the series and parallel arrangements.

It is when working with loads between 100 and 500 ohms that the decision of which type of circuit to use will present difficulties. The choice will mainly be guided by the relative ease and cost of construction of the coils and condensers required. Of course, it is theoretically possible to use a tapped circuit as shown in Fig. 1 (C) or (D) to make the damping of the circuit any desired value, but (C) may not be practically convenient and (D) involves the use of more copper in the coils and larger condensers.

We shall confine our comparison to the arrangements of Fig. 1 (A) and (B) on the following basis :—

At a fixed frequency, but varying turns in coupling coil. It is assumed that the position and size of the coupling coil has been fixed, but that it consists of a number of turns that can be combined in series and (or) parallel as required. Then the following relations hold.

Series Circuit.

Assuming all turns of coupling coil are paralleled—

Inductance of equivalent single turn $= \mathrm{L}$

Reactance ,, ,, $= \mathrm{X}$

Resistance ,, ,, $= r$

EMF induced in ,, ,, $= \mathrm{E}.$

Load resistance $= \mathrm{R}$

Load Current $\mathrm{I} = \dfrac{\mathrm{E}}{\mathrm{R}}$

Power $= \dfrac{\mathrm{E}^2}{\mathrm{R}} = \mathrm{P_S}$

Energy capacity of inductance or condenser $= \frac{1}{2}\,\mathrm{LI}^2 = \frac{1}{2}\,\mathrm{L}\,\dfrac{\mathrm{E}^2}{\mathrm{R_S}^2}$

Loss in coupling coil $= r\mathrm{I}^2 = \dfrac{r\mathrm{E}^2}{\mathrm{R}^2}$

Now coupling coil of " n " turns in series.

Inductance of n turns $= \mathrm{L}n^2$

Reactance ,, $= \mathrm{X}n^2$

Resistance ,, $= rn^2$

EMF $= \mathrm{E}n$

Load current $\dfrac{\mathrm{E}n}{\mathrm{R}} = n\mathrm{I}$

Power $= \dfrac{\mathrm{E}^2n^2}{\mathrm{R}} = n^2.\mathrm{P_S}$

Energy capacity of inductance or condenser $= \frac{1}{2}\,\mathrm{L}n^2.n^2\mathrm{I}^2 = n^4\,(\frac{1}{2}\,\mathrm{LI}^2)$

Loss in coupling coil $= r.n^2.n^2\mathrm{I}^2 = n^4(r\mathrm{I}^2).$

It will be noticed that on the basis of a fixed EMF in the single turn the energy capacity (volume) of the condenser, and the losses in the coupling coil vary as the square of the power in the load.

Parallel Circuit.

The constants of the coupling coil will vary in exactly the same way as for the series case, but as turns are increased the power output for a fixed position of coil decreases as shown below.

Assume $R_P >> X_P$ or X_L in the range considered. (It will be seen later that the same rule holds good over the whole practical working range down to $R_P = X_P$ and X_L).

Assume all turns of coupling coil paralleled as before.

Inductance of equivalent single turn	$= L$
Reactance ,, ,,	$= X$
Resistance ,, ,,	$= r$
EMF induced in ,, ,,	$= E$
Load resistance	$= R_P$
Equivalent series load resistance	$= \frac{X^2}{R_P}$

Power to load $= \frac{E^2 R_P}{X^2} = P_P$

Energy capacity of inductance or condenser $\frac{1}{2} LI^2 = \frac{1}{2} L \frac{E^2 R_P^2}{X^4}$

Losses in coupling coil $= r \frac{E^2 R_P^2}{X^4}$

Now if coupling coil is charged to " n " turns in series,

Inductance of " n " turns	$= Ln^2$
Reactance ,,	$= Xn^2$
Resistance	$= rn^2$
EMF	$= En$

Equivalent series load resistance $= \frac{(Xn^2)^2}{R_P}$

Power to load $= \frac{E^2 n^2 R_P}{(Xn^2)^2} = \frac{E^2 R_P}{X^2 n^2} = \frac{P_P}{n^2}$

Energy capacity of inductance or condenser—

$$= \frac{1}{2} Ln^2 \cdot \frac{E^2 n^2 R_P^2}{X^4 n^8} = \frac{1}{n^4} \cdot \frac{1}{2} L \frac{E^2 R_P^2}{X^4}$$

Losses in coupling coil $= \frac{1}{n^4} \quad r \frac{E^2 R_P^2}{X^4}$

Large Change in Frequency.

It is assumed (1) that the primary circuit has a constant capacity (more or less) and that the change of frequency is obtained by varying the arrangement of turns in the coil maintaining all the conductors in use. (2) That the HT volts on the valve are constant, and hence the volts across the circuit are constant.

Hence to increase the frequency " m " times we must decrease the turns $\frac{1}{m}$

times. The inductance of the coil is reduced to $\frac{1}{m^2}$, and the reactance $\frac{1}{m}$ times.

$\therefore$ current per turn increases m times.

Hence the ampere turns of the primary and the flux through the coupling coil are unchanged. The EMF induced in one turn of the coupling coil has increased "m" times.

Series Circuit.

To produce the same loading we require the same EMF. Hence turns can be reduced in ratio $\frac{1}{m}$ and paralleled so that inductance is reduced in ratio $\frac{1}{m^2}$ and reactance in the ratio $\frac{1}{m}$.

In the case of the series circuit this means that the same condenser value should be required. The volts across it, however, will be reduced in ratio $\frac{1}{m}$.

Parallel Circuit.

If we reduce the coupling coil as in the series case keeping the same EMF and retain the same value of condenser its reactance will have been reduced in the ratio $\frac{1}{m}$.

In the range where $X_P << R_P$ this decreases the equivalent series resistance in ratio $\frac{1}{m^2}$. [Equation (1)]

The same EMF will produce m^2 times more power. Much heavier loading will be produced for the same position of coupling coil. Even when X_P is comparable with R_P the same tendency will persist though not in so marked a form.

If we retain the original coupling coil the EMF will be increased "m" times, and the reactance of coupling coil and required condenser will be increased "m" times. This will increase the effective series resistance m^2 times. The power $\left(\frac{E^2}{R}\right)$ will therefore be unchanged. This will be approximately true provided X_L and X_P are less than R_P.

The required condenser capacity will be decreased $\frac{1}{m^2}$ times, and the current in the coupling coil $\frac{1}{m}$ times. Losses in the coupling coil decrease $\frac{1}{m^2}$ assuming a constant resistance, but $\frac{1}{\sqrt{m^3}}$ assuming resistance increases as $\sqrt{m}$.

Small change in frequency such as would be covered on the output circuit by the range of the tuning condenser, but for which the primary inductance must be adjusted.

Series Circuit.

Assume frequency increased "m" times then the EMF in the coupling coil is increased "m" times. Therefore power output increases "m^2" times for same position of coil.

The tuning condenser has to be decreased in ratio $\frac{1}{m^2}$ (reactance increased m times) so that voltage across condenser for same power in load has increased "m" times.

Parallel Circuit. $X_P << R_P$.

As for the series case the EMF in the coupling coil has increased "*m*" times; the reactance of the tuning condenser has increased "*m*" times. The equivalent series resistance has increased m^2 times from equation (1). Hence the power delivered $\left(\frac{E^2}{R}\right)$ will remain unchanged. Even when X_P is comparable with R_P this tendency will remain.

These results show the general trend of behaviour, and we can now study a few cases of the parallel circuit where X_L is comparable with R_P. The conditions will be those studied under 3, i.e.:

Value of X_P *adjusted to give maximum current in the load* R_P. Fig. 6 (A), (B) and (C).

(A) $X_L = R_P$. Fig. 6 (A).

Then we found $X_P = X_L$ and the current in load $= \frac{E}{R_P} \cdot \frac{BD}{GD} = \frac{E}{R_P}$ exactly the same as would be produced by the series circuit. The current in the coupling coil is $\sqrt{2}$ times the current in the load, whereas in the series circuit the coil current would be equal to the load current. Hence the coil heating is doubled in the parallel circuit.

The condenser required has the same capacity and the same voltage across it in both cases.

This case of $X_L = R_P$ is a convenient meeting point for the comparison of series and parallel circuits.

(C) $X_L = \frac{1}{2} R_P$. Fig. 6 (C).

This may be produced from $X_L = R_P$ in two ways.

(i) By decrease of coupling coil turns in ratio $\frac{1}{\sqrt{2}}$

(ii) By decrease of frequency to $\frac{1}{2}$ original value.

(i) *Decrease of coupling coil turns* $\frac{1}{\sqrt{2}}$.

Then EMF is reduced $\frac{1}{\sqrt{2}}$.

Referring back to Fig. 6 (C) and the previous discussion.

Resultant impedance of circuit $= GD_2 = \frac{R_P}{2} \cdot \frac{1}{\sqrt{5}} = \frac{R_P}{2\sqrt{5}}$

and has been reduced in ratio $\frac{1}{\sqrt{10}}$

$\therefore$ compared with parallel circuit when $X_L = X_C = R$.

Coil current has increased in ratio $\frac{\sqrt{10}}{\sqrt{2}} = \sqrt{5}$

$\therefore$ Current per unit has increased $\sqrt{\frac{5}{2}}$ times.

$\therefore$ Coil heating has increased $\frac{5}{2}$ *times* compared with parallel case ($X_L = X_C = R$)
or 5 times compared with series case.

Power in Load.

Volts across load $= \frac{BD_2}{GD_2} \cdot \frac{E}{\sqrt{2}} = \sqrt{2}\ E$

$\therefore$ Power in load is doubled.

Energy in Condenser.

Condenser has been doubled in value and volts across it have increased $\sqrt{2}$ times.

$\therefore$ Energy has increased 4 times.

Equivalent Series Resistance of Circuit $= FD_2$.

For this we have :—

$$\frac{FD_2}{BC} = \frac{GF}{GB} \text{ or } FD_2 = 2\ GF$$

$$GF = GB - BF$$

$$BF = \frac{2}{\sqrt{5}} BD_2 = \frac{2}{\sqrt{5}} \cdot \frac{2}{\sqrt{5}} GB = \frac{4}{5} GB.$$

$$\therefore GF = GB - \frac{4}{5} GB = \frac{1}{5} GB = \frac{1}{10} R_P$$

$$\therefore \text{Series Resistance : } FD_2 = 2\ GF = \frac{1}{5} R_P$$

Series Reactance $GF = \frac{1}{10} R_P$.

Compare these with the results we should have obtained assuming the same laws held good as when $X_P << R_P$.

$$n = \frac{1}{\sqrt{2}}$$

Power increases as $\frac{1}{n^2} = 2$ times.

Condenser energy increases as $\frac{1}{n^4} = 4$ times.

Coil heating increases as $\frac{1}{n^4} = 4$ times.

This last item is the only one giving a different figure, and indicates the merging into the general case.

To obtain the same increase of power from the series circuit we should have to increase the coil turns, and therefore the EMF by $\sqrt{2}$. The coil inductance is doubled, and the tuning condenser halved. Volts across tuning condenser are increased $2\sqrt{2}$ times.

$\therefore$ Energy in condenser is increased $\frac{1}{2}(2\sqrt{2})^2$ times $= 4$ times.

Coupling coil losses are also increased 4 times.

(ii) *Decrease of frequency to $\frac{1}{2}$ original value.*

It is assumed that primary turns have been readjusted. Then the EMF in this case has been reduced to $\frac{1}{2}$ E instead of $\frac{1}{\sqrt{2}}$ E as in (i). Hence the power to the load would be unchanged by the change in frequency. As in (i)

$$\text{Effective series resistance} = \frac{1}{5} R_P$$

$$\text{,, \quad ,, \quad reactance} = \frac{1}{10} R_P$$

$$\text{,, \quad ,, \quad impedance} = \frac{1}{2\sqrt{5}} R_P$$

$$\text{Current in coil} = \frac{\text{EMF}}{\text{impedance}} = \frac{E}{2} \cdot \frac{2\sqrt{5}}{R_P} = \frac{E\sqrt{5}}{R_P}$$

$\therefore$ current in coil has increased $\sqrt{\frac{5}{2}}$ times.

$\therefore$ coil losses have increased $\frac{5}{2}$ times

as compared with parallel circuit when $X_L = X_C = R$ (or 5 times compared with the series circuit $X_L = X_C = R$).

Condenser reactance is halved and frequency halved

$\therefore$ capacity is X by 4. Volts on condenser are unchanged.

$\therefore$ Energy capacity has increased 4 *times*.

As previously stated in the series circuit to maintain the same leading it would be necessary to double the turns in the inductance, and coupling coil losses are increased 4 times and energy in condenser is increased 4 times.

(D) $X_L = \frac{1}{k} R_P$.

As before this may be produced from $X_L = R_P$ in two ways. (i) By decrease of coupling coil turns in ratio $\frac{1}{\sqrt{k}}$. (ii) By decrease of frequency to $\frac{1}{k}$ of original value.

Only (i) will be considered in detail and for purposes of reference we shall use Fig. 6 (c).

In this we shall have:

$$BC = R_P,\ BG = X_L = X_C = \frac{1}{k} R_P$$

$$\text{EMF in circuit} = \frac{1}{\sqrt{k}} E$$

Resultant impedance in circuit $= GD_2$ and we have $\frac{GD_2}{GB} = \frac{GB}{GC}$

$$GC = R_P\sqrt{\left(1 + \frac{1}{k^2}\right)} = R_P \frac{\sqrt{k^2+1}}{k}$$

$$\therefore \text{Resultant impedance} = GD_2 = \frac{GB^2}{GC} = \frac{R_P^2}{k^2} \times \frac{k}{R_P\sqrt{k^2+1}}$$

$$= \frac{R_P}{k\sqrt{k^2+1}}$$

Series Resistance $= FD_2$.

$$\frac{FD_2}{GD_2} = \frac{BC}{GC}$$

$$\therefore FD_2 = \frac{GD_2 \cdot BC}{GC} = \frac{R_P \cdot R_P \cdot}{k\sqrt{k^2+1}} \quad \frac{k}{R_P\sqrt{k^2+1}}$$

$\therefore$ Series Resistance $= FD_2 = \dfrac{R_P}{k^2 + 1}$

Series Reactance $= GF$.

$$\frac{GF}{FD_2} = \frac{GB}{BC}$$

Series reactance $= GF = \dfrac{FD_2 \cdot GB}{BC} = \dfrac{R_P}{k(k^2+1)}$

Power in Load.

$$\text{Volts across load} = \frac{BD_2}{GD_2} \cdot \frac{E}{\sqrt{k}} = \frac{BC}{BG} \cdot \frac{E}{\sqrt{k}} = \sqrt{k}.\ E$$

$$\therefore \text{Power in load } \frac{(\sqrt{k} \cdot E)^2}{R_P} = k\ \frac{E^2}{R_P}$$

and has increased "k" times compared with parallel circuit where $X_L = X_P = R_P$.

Losses in Coupling Coil.

$$\text{Current in coupling coil} = \frac{\text{EMF}}{\text{impedance}} = \frac{E}{\sqrt{k}} \cdot \frac{k\sqrt{k^2+1}}{R_P}$$

$$= \sqrt{k(k^2+1)}\ \frac{E}{R_P} \qquad \text{Instead of } \frac{\sqrt{2}\,E}{R_P}$$

Turns in coil have decreased $\dfrac{1}{\sqrt{k}}$

$\therefore$ Current per unit turn has increased

$$\frac{\sqrt{k(k^2+1)}}{\sqrt{k} \cdot \sqrt{2}} = \sqrt{\frac{k^2+1}{2}}$$

$\therefore$ Coil losses have increased in ratio $\dfrac{k^2+1}{2}$ compared with parallel circuit when $X_L = X_P = R_P$ or (k^2+1) times that for the series circuit.

Energy in Condenser.

Condenser has increased "k" times and volts across it have increased $\sqrt{k}$ times.

$\therefore$ Energy capacity has increased k^2 times.

Summary.

For a given load resistance R_P the choice of series or parallel circuit depends on the relative ease of constructing the coupling coil and condenser to give the required loading, For values of R_P below $100\,\Omega$ the series circuit is usually chosen and for R_P above $500\,\Omega$ the parallel circuit. For the range in between each case must be decided on its merits.

With the series circuit working into a high resistance and heavy loading the turns on the coupling circuit may be so large that a stationary wave distribution may occur.

For the parallel circuit, on the other hand, the inductance of the coupling coil becomes smaller for heavy loading, and the inductance of the connecting leads might become a limitation.

It is pointed out that for the parallel circuit to function reasonably X_L must not be greater than R and should preferably be considerably less. The arrangement where $X_L = X_C = R$ forms a convenient meeting point for comparing the properties of series and parallel circuits.

The following tables help in giving a picture of the results.

(1) Comparison at a fixed frequency and fixed position of coupling coil.

$X_L = X_C = R$ gives same loading for both series and parallel circuit. The required changes to increase the load current $\sqrt{2}$ and also $\sqrt{k}$ times in both cases are shown in the table.

	$X_L = X_C = R$		To multiply power in load by 2.		To multiply power in load by "k."	
	Series Circuit.	Parallel Circuit.	Series Circuit.	Parallel Circuit.	Series Circuit.	Parallel Circuit.
Resistance of load	R	R	R	R	R	R
Reactance of coupling coil or tuning condenser ..	X = R	X = R	2X	$\frac{1}{2}$X	kX	$\frac{1}{k}$X
Turns in coupling coil	T	T	$\sqrt{2}$.T	$\frac{1}{\sqrt{2}}$.T	$\sqrt{k}$.T	$\frac{1}{\sqrt{k}}$.T
E.M.F. in coupling coil	E	E	$\sqrt{2}$.E	$\frac{1}{\sqrt{2}}$.E	$\sqrt{k}$.E	$\frac{1}{\sqrt{k}}$.E
Current in load ..	I = $\frac{E}{R}$	I = $\frac{E}{R}$	$\sqrt{2}$.I	$\sqrt{2}$.I	$\sqrt{k}$.I	$\sqrt{k}$.I
Power in load ..	P = $\frac{E^2}{R}$	P = $\frac{E^2}{R}$	2.P	2.P	k.P	k.P
Capacity of tuning condenser ..	C	C	$\frac{1}{2}$.C	2.C	$\frac{1}{k}$.C	k.C
Volts across tuning condenser ..	V = E	V = E	$2\sqrt{2}$.V	$\sqrt{2}$.V	$k\sqrt{k}$.V	$\sqrt{k}$.V
Energy in tuning condenser ..	$\frac{1}{2}$CV²	$\frac{1}{2}$CV²	4($\frac{1}{2}$CV²)	[illegible]($\frac{1}{2}$CV²)	k^2($\frac{1}{2}$CV²)	[illegible]($\frac{1}{2}$CV²)
Resistance of coupling coil ..	r	r	$2r$	$\frac{1}{2}r$	$k.r$	$\frac{1}{k}.r$
Current in coupling coil ..	I	$\sqrt{2}$.I	$\sqrt{2}$.I	$\sqrt{10}$.I	$\sqrt{k}$.I	[illegible] I
Losses in coupling coil	rI²	$2r$I²	$4r$I²	$5r$I²	k^2rI²	$(k^2+1)r$I²
Resultant series R of circuit ..	R	$\frac{1}{2}$R	R	$\frac{1}{5}$R	R	$\frac{1}{k^2+1}$R
Resultant series X of circuit ..	0	$\frac{1}{2}$X	0	$\frac{1}{10}$R	0	$\frac{k}{k^2+1}$R
Resultant series impedance circuit	R	$\frac{1}{\sqrt{2}}$R	R	$\frac{1}{2\sqrt{5}}$R	R	$\frac{1}{k\sqrt{k^2+1}}$.R

(2) To give a fixed degree of loading at a fixed position of coil when frequency is decreased "m" times, commencing where $X_L = X_C = R$.

	λ $X_L = X_C = R$		m Frequency decreased "m" times.	
	Series Circuit.	Parallel Circuit.	Series Circuit.	Parallel Circuit.
Resistance of load	R	R	R	R
Reactance of coupling coil or tuning condenser ..	$X=R$	$X=R$	mX	$\frac{1}{m}X$
Turns in coupling coil ..	T	T	mT	T
EMF in coupling coil ..	E	E	E	$\frac{1}{m}E$
Current in load	$I=\frac{E}{R}$	$I=\frac{E}{R}$	I	I
Power in load	$P=\frac{E^2}{R}$	$P=\frac{E^2}{R}$	P	P
Capacity of tuning condenser	C	C	C	$m^2 C$
Volts across tuning condenser	V	V	mV	V
Energy in tuning condenser ..	$\frac{1}{2}CV^2$	$\frac{1}{2}CV^2$	$m^2(\frac{1}{2}CV^2)$	$m^2(\frac{1}{2}CV^2)$
Resistance of coupling coil ..	r	r	m^2r	r
Current in coupling coil ..	I	$2I$	I	$\sqrt{(m^2+1)}I$
Losses in coupling coil ..	rI^2	$2rI^2$	$m^2(rI^2)$	$(m^2+1)(rI^2)$
Resultant Series R of circuit	R	$\frac{1}{2}R$	R	$\frac{1}{m^2+1}R$
Resultant Series X of circuit	O	$\frac{1}{2}X$	O	$\frac{1}{m(m^2+1)}R$
Resultant Series Z of circuit..	R	$\frac{1}{2}R$	R	$\frac{1}{m\sqrt{m^2+1}}R$

Variation of loading over range of output tuning condenser decreasing frequency "m" times.

	Series Circuit.	Parallel Circuit.
Loading for fixed coil position	$\div\, m^2$	constt.
Condenser volts for fixed loading	$\div\, m$	constt.

These tables bring out the following points :—

(1) To increase the loading for the series circuit the number of turns in the coupling coil must be increased, while for the parallel circuit the turns must be decreased.

(2) Tune point on the parallel circuit as found by varying the condenser to give maximum current in the load does not give a circuit of zero reactance.

(3) *Tuning Condenser.*

For a given loading the energy capacity ($\frac{1}{2}$ CV^2) of the tuning condenser at a given wavelength is the same for the parallel or series circuit. For the parallel circuit the volts across the condenser are constant throughout the waverange, but the capacity required increases as the square of the wavelength. Such a condenser can conveniently be made by switching units in parallel.

For the series circuit the required capacity is constant throughout the waverange, but the volts across it are proportional to the wavelength. This is not so convenient to arrange. Also with a fixed coupling coil and a variable condenser of the parallel vane type the volts across the condenser will increase as the capacity is reduced, and the spacing must be sufficient to stand the voltage at minimum working capacity.

In either case the energy capacity of the condenser increases as the square of the power in the load.

(4) *Coupling Coil.*

With the parallel circuit a fixed loading is obtained over a wide range of wavelengths with a fixed coupling coil. With the series circuit the coupling coil turns must be increased in proportion to the wavelength. The losses in the coupling coil are not widely different in the two cases, though in the area where $X_L = X_C = R$ the series circuit has the advantage. In either case the losses in the coupling coil are proportional to the square of the power in the load.

For medium resistance loads the series circuit is likely to require all turns in series, while the parallel circuit will need parallel turns. It is difficult to arrange that equal EMF's will be induced in each turn. With series turns, however, the EMF's will add O.K. With parallel turns the unequal EMF's give rise to circulating currents and the resultant EMF will take up a mean value.

Change of load resistance.

Series Circuit.

If load R increases " n " times, then to retain same loading EMF must increase $\sqrt{n}$ times.

$\therefore$ Turns in coil increase $\sqrt{n}$ times.

Current in each turn decreases $\frac{1}{\sqrt{n}}$

$\therefore$ Coil losses are unchanged.

Inductance of coil increases n times.

Capacity of condenser decreases $\frac{1}{n}$ times.

Energy capacity of condenser is unchanged.

Parallel Circuit.

If load resistance R increases n times, then to retain same loading :

Increase turns in coil $\sqrt{n}$ times.

,, inductance n times.

Effective series R increases n times.

EMF is increased $\sqrt{n}$ times.

$\therefore$ Power to load is unchanged.

Capacity of condenser decreases $\frac{1}{n}$

Energy capacity of condenser is unchanged.

Appendix No. 1 to Comparison of Parallel and Series Coupling Circuits for Transmitters.

NOTE.—The tables given on page 29 start from the case when $X_L = X_C = R$ and show what alterations are required to increase the loading " k " times. If two cases are to be compared in neither of which $X_L = X_C = R$ they can be referred to this case. Let k_1 and k_2 be the respective increases in loading, and $\frac{k_2}{k_1} = k$ the relative increase in loading. Then it is clear that for the series circuit the original table will still hold good throughout, while for the parallel circuit it holds strictly down to the item " Resistance of Coupling Coil." For the current in the coupling coil the factor will be $\sqrt{\frac{k_2 (k_2^2 + 1)}{k_1 (k_1^2 + 1)}}$ and so on.

Similar rules can be formulated for the table on page 30.

Variation of Load Resistance, Coupling Coil Losses, and Energy Capacity of Condenser at Fixed Loading.

Series Circuit.

If load resistance is increased " n " times

EMF in coupling coil must be increased $\sqrt{n}$ times,

i.e., coupling coil turns ,, ,, $\sqrt{n}$ times,

,, ,, inductance ,, n times.

Capacity of condenser decreases $\frac{1}{n}$ times.

Current in load and condenser decreases $\frac{1}{\sqrt{n}}$ times.

$\therefore$ Volts on condenser increase $\sqrt{n}$ times.

Energy capacity of condenser $\frac{1}{2}CV^2$ is unchanged.

Resistance of coupling coil increases " n " times.

$\therefore$ Losses in coupling coil rI^2 are unchanged.

The same relation can be shown to hold for the parallel circuit provided $R_P >> X_P$.

Change of Loading, etc., when R_P is changed to nR_P, X_L and X_C remaining unchanged.

For the series circuit the various quantities are easily calculated. In particular the power $\frac{E^2}{nR_P}$ has been reduced in ratio $\frac{1}{n}$.

For the parallel circuit provided $R_P >> X_L$ or X_C we shall have effective series resistance $\frac{X_L^2}{nR_P}$.

$$\text{Power} = \frac{E^2\, n\, R_P}{X_L^2} \text{ and has increased " } n \text{ " times.}$$

For the parallel circuit in other cases

(A) $X_L = X_C = R_P$.

This case is shown in Fig. 7 (A), the dashed letters referring to the load resistance $n\ R_P$.

This is very similar to the case treated on page 27. There $X_L = \frac{1}{k} R_P$ whereas here $X_L = R_P = \frac{1}{n}(n\ R_P)$. We must therefore substitute nR_P for R_P in the various circuit constants there derived.

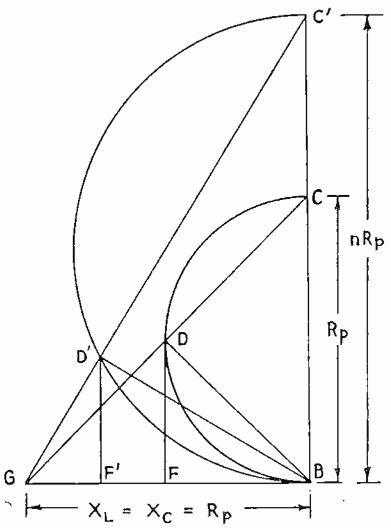


FIG. 7.

Resultant impedance of circuit GD^1

$$= \frac{n\ R_P}{n\sqrt{n^2+1}} = \frac{R_P}{\sqrt{n^2+1}}$$

Series resistance $F^1D^1 = \frac{n\ R_P}{n^2+1}$

Series reactance $GF^1 = \frac{n\ R_P}{n\ (n^2+1)} = \frac{R_P}{n^2+1}$

Current in coil $= \frac{\text{EMF}}{\text{impedance}} = \frac{E\sqrt{n^2+1}}{R_P}$

Power in load

$$= \frac{n\ R_P}{n^2+1} \times \left(\frac{E\sqrt{n^2+1}}{R_P}\right)^2 = n\frac{E^2}{R_P} = \frac{n^2E^2}{n\ R_P}$$

Coupling coil losses $= n^2+1\quad r\ \frac{E^2}{R_P{}^2}$

Volts across condenser $= nE$.

(B) $X_L = X_C = \frac{1}{k} R_P = \frac{1}{nk}(n\ R_P)$;

so that in this case we have from the results on pages 27 and 28

Resultant impedance of circuit $GD^1 = \frac{nR_P}{nk\sqrt{n^2k^2+1}} = \frac{R_P}{k\sqrt{n^2k^2+1}}$

Series resistance $F^1D^1 = \frac{nR_P}{n^2k^2+1}$

Series reactance $= \frac{nR_P}{nk\ (n^2k^2+1)} = \frac{R_P}{k\ (n^2k^2+1)}$

Current in coil $= \frac{E.k\sqrt{n^2k^2+1}}{R_P}$

Power in load $= \frac{n\ R_P}{n^2k^2+1} \times \left(\frac{Ek\sqrt{n^2k^2+1}}{R_P}\right)^2 = \frac{nk^2E^2}{R_P}$

Coupling coil losses $= r\ \frac{E^2k^2\ (n^2k^2+1)}{R_P{}^2}$

Volts across condenser $= nkE$.

These results are summarised in the following table. The outstanding points are that for all ratios of $X_L \gtrless R_P$ the change from R_P to nR_P increases the power in the load "n" times, the voltage on the condenser "n" times and the energy capacity of the condenser n^2 times.

Parallel Circuit :—Load Resistance Varied.

	$X_L=X_C=R_P=R$		$X_L=X_C=\frac{1}{k}R_P$	
Resistance of load	R_P	nR_P	R_P	nR_P
EMF in coupling coil ..	E	E	E	E
Series impedance load ..	$\frac{R_P}{\sqrt{2}}$	$\frac{R_P}{\sqrt{n^2+1}}$	$\frac{R_P}{k\sqrt{k^2+1}}$	$\frac{R_P}{k\ n^2k^2+1}$
Series resistance	$\frac{R_P}{2}$	$\frac{nR_P}{n^2+1}$	$\frac{R_P}{k^2+1}$	$\frac{nR_P}{n^2k^2+1}$
Series reactance	$\frac{R_P}{2}$	$\frac{R_P}{n^2+1}$	$\frac{R_P}{k(k^2+1)}$	$\frac{R_P}{k(n^2k^2+1)}$
Current in coil	$\frac{E\sqrt{2}}{R_P}=\sqrt{2}.I$	$\frac{E\sqrt{n^2+1}}{R_P}=\sqrt{(n^2+1)}I$	$\frac{Ek\sqrt{k^2+1}}{R_P}$	$\frac{Ek\sqrt{n^2k^2+1}}{R_P}$
Power in load	$\frac{E^2}{R_P}=P$	$\frac{n^2E^2}{nR_P}=nP$	$\frac{k^2E^2}{R_P}=P^1$	$\frac{nk^2E^2}{R_P}=nP^1$
Current in load	$\frac{E}{R_P}=I$	$\frac{E}{R_P}=I$	$\frac{kE}{R_P}=I^1$	$\frac{kE}{R_P}=I^1$
Volts across load	E	nE	kE	nkE
Energy in condenser ..	$\frac{1}{2}CE^2$	$n^2(\frac{1}{2}CE^2)$	$\frac{1}{2}C(kE)^2$	$n^2[\frac{1}{2}C(kE)^2]$
Losses in coil	$2\ rI^2$	$(n^2+1)rI^2$	$(k^2+1)rI^{1\,2}$	$(n^2k^2+1)rI^{1\,2}$

Other Methods of Increasing Loading.

Instead of altering the turns to increase the loading we can use the circuits shown in Fig. 8. Fig. 8 (A) is a transformer arrangement for the parallel circuit, while (B) is a type using a reactance transformer, the constants of which are easier

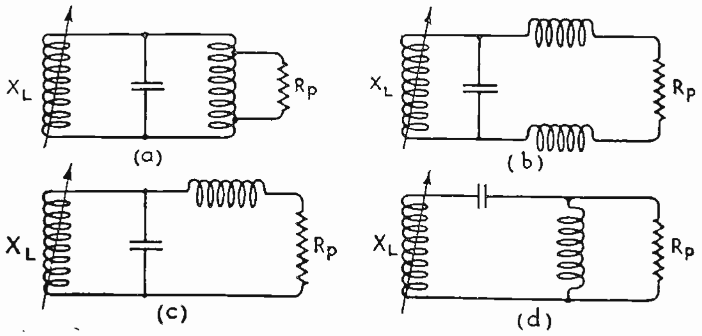


FIG. 8.

to calculate and should be as good as (A) in practice. (C) is the same as (B) in an unsymmetrical form. (D) shows a reactance transformer arrangement for the series circuit.

Fig. 9 shows 8 (B) and its equivalent. It will be seen that the effective load resistance across the coupling coil can be increased to nR_P provided

$$X_1 = \frac{n}{\sqrt{n-1}} R_P$$

and

$$X_1^1 = \sqrt{n-1}\, R_P$$

It may be noted that X_1 the reactance of the extra condenser required becomes larger as "n" increases, so that the extra capacity required becomes smaller. We

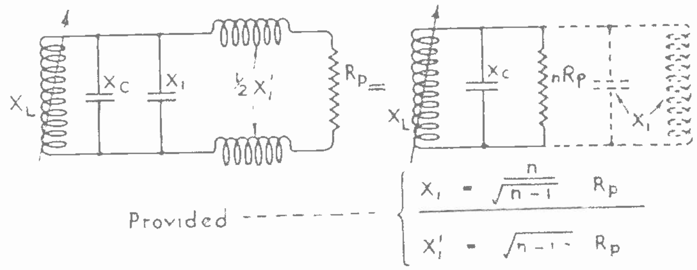


FIG. 9.

have already seen that increasing the load resistance from R_P to "nR_P" other things being unchanged will increase the power in the load "n" times and the volts across the condenser "n" times.

Optimum Ratio of Primary and Secondary Capacity.

It will have been noticed that to increase the power "n" times for a given position of coupling coil we have to increase the energy capacity $\frac{1}{2}CV^2$ of the condenser "n^2" times. This suggests that there should be a point beyond which it will be more economical to increase both primary and secondary energy capacity in proportion; and that this will occur when they become equal. This is proved later, but it may be noted that it will not usually come into operation on short wave sets. The minimum primary capacity is fixed by the valve capacity, and normally the secondary capacity is less than this.

Relative Capacities in Primary and Secondary to give minimum total energy capacity.

Assume fixed dimensions and positions of primary and secondary.

C_1	Primary condenser.	
L_1	Primary inductance.	
C_2	Secondary condenser.	} Series Circuit.
L_2	Secondary inductance.	
R	Load resistance.	
k =	Coefficient of coupling.	
M =	$k\sqrt{L_1 L_2}$ = mutual inductance.	
V_1 =	Volts across primary.	

V_2 = Volts across secondary.
I_1 = Primary current.
I_2 = Secondary current.

Then $L_1C_1 = L_2C_2 = \dfrac{1}{\omega^2}$

and $M = k\sqrt{L_1L_2} = \dfrac{k}{\sqrt{C_1\,C_2}.\omega^2}$

$I_1 = C_1\,\omega V_1$

Secondary EMF, $E_2 = M\,\omega I_1$

$$= \frac{k\omega\,C_1\,\omega V_1}{\sqrt{C_1C_2}\,\omega^2} = \frac{k\,C_1\,V_1}{\sqrt{C_1\,C_2}}$$

Secondary current $I_2 = C_2\,\omega V_2$

Power delivered to secondary load $= E_2I_2 = \dfrac{k.\,C_1V_1.\,C_2V_2\omega}{\sqrt{C_1C_2}}$

$$= k\omega\,\sqrt{(C_1V_1^2)\,(C_2V_2^2)}$$

If the total energy capacity of primary and secondary are fixed

$$C_1V_1^2 + C_2V_2^2 = \text{constant}.$$

Then the power delivered to the load will be a maximum when

$$C_1V_1^2 = C_2V_2^2$$

which was to be proved.

It could also be shown that in this case maximum efficiency would result if the amount of copper in primary and secondary were also equal.

E. GREEN

A NOTE ON THE PHOTOGRAPHIC ENGRAVING OF SCALES ON CATHODE RAY TUBES

In the majority of cases where Cathode Ray Tubes are used for quantitative measurements, the need arises for some type of scale to which the movements of the cathode ray spot may be referred.

A process has been evolved whereby such scales can be photographed on the end of the tube, and a brief description of this is given below.

WHEN first Cathode Ray Tubes began to be used for such measurements as modulation percentage, waveform analysis, and the like, it was essential, of course, to have means available whereby the deflection of the spot could be measured.

In the first type of Cathode Ray Equipment manufactured by the Marconi Company the measuring device took the form of two horizontal arms with their adjacent edges parallel and mounted in front of the screen of the tube in such a manner that they could be moved vertically. The device is shown in Fig. 1, from which it will be seen that the top arm could be moved alone, or the two arms could be moved together by rotating the upper or lower of the two concentric thumbscrews respectively. The movement of the lower arm moved a scale along a fixed pointer, so that the position of the figure traced by the spot was indicated on the scale, which was engraved in centimetres. The upper arm also carried a pointer which moved in a slot in the moving scale, and indicated the distance apart of the two arms, thus enabling the height or change of height of any figure to be measured.

Such a device suffered from several disadvantages. It took an appreciable time to adjust and measurement was very prone to parallax errors.

An obvious improvement was to mount in front of the tube a sectionally scaled celluloid or other transparent screen, the markings of which were adapted to the particular use to which the tube was being put. This type of measuring screen was quicker to read, but still suffered from the defect of parallax errors, especially since the screen was plane, whereas the end of the cathode ray tube was convex.

The necessity for some more accurate screen was emphasised when a new type of cathode ray direction finder was developed by the Marconi Company, in which the bearing was given by the orientation, on the tube end, of a linear cathode ray spot trace. The problem, simply stated, was to engrave an accurate compass scale on the convex end of the tube.

Two alternatives presented themselves. Either the scale could be etched chemically on the end of the tube or photographic methods could be employed. The first method is obviously not preferable both on account of the fact that the process is a lengthy one and also that a scale, so etched, could not be removed from the tube. The method given below of originally fixing the position that the scale should occupy on the tube is one which is not entirely free from errors and until the tube has been rechecked after the scale is on, it cannot be definitely stated whether the first scale is correct or not. If incorrect and if the scale had been etched, the tube would have to be discarded and for this reason the possibility of etching the scale had to be abandoned.

Various methods of producing photographic images on glass are of course available, but the majority of these were unsuitable owing to the fact that the end of the tube is not plane. Thus the employment of transfer methods, using, for example, some form of stripping film where a paper base is coated with a thin tough celluloid-like film carrying the emulsion which is stripped from the base after

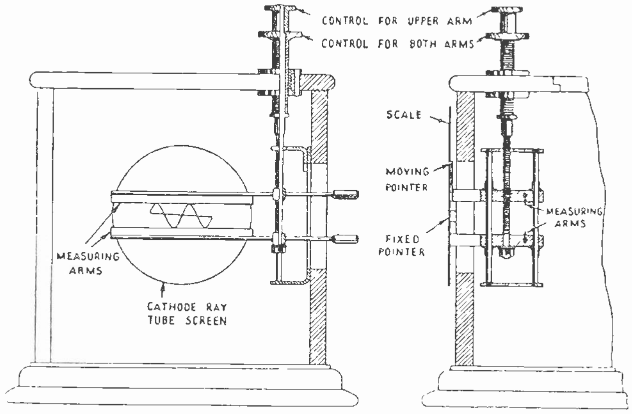


FIG. 1.

exposure, development, and fixing, and mounted on the glass surface, are impracticable owing to the appreciable distortion of the image which would result and the difficulty of obtaining good contact between film and glass over the whole of the convex surface.

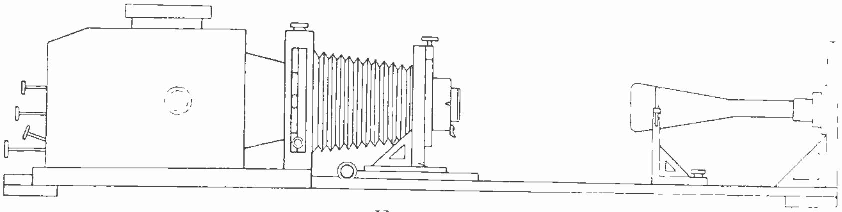


FIG. 2.

Modifications of the normal gum-bichromate process were also tried. The end of the tube was first sensitised by a solution of albumen or gum arabic and ammonium or potassium bichromate in water and then exposed to a negative image in the enlarger to be described later. After exposure the image was rubbed with photo-litho ink and those portions of the image which had not been acted on by light washed away with water, the exposed portions remaining insoluble and holding the ink, thus producing the required image. The objections to this process were

(A) It was extremely difficult to obtain a sufficiently good substratum for the albumen or gum solution.

(B) The exposure needed was very long; some 30 minutes for the particular scale tried.

(C) The permanency of the image was not great and the density of the black lines not good.

Several firms were consulted with a view to obtaining help on the problem and finally it was decided to employ a modification of the wet collodion process. The change of technique involved by the fact that the surface to be coated was convex and not plane introduced certain difficulties, and a brief description of the final process is given below.

A specially adapted type of enlarger was first built, shown in Fig. 2. The Cathode Ray Tube is held firmly in a socket and its end supported on a semicircular wooden support, to which the tube is held by a spring band. The adjustment of the position of the image on the tube end is effected by mounting the photographic image negative of the scale in a specially constructed holder which can be moved either angularly or horizontally. Vertical movement of the image is effected by raising or lowering the lens by a small slide in front of the enlarger. A diagram illustrating the actual scale is shown in Fig. 3.

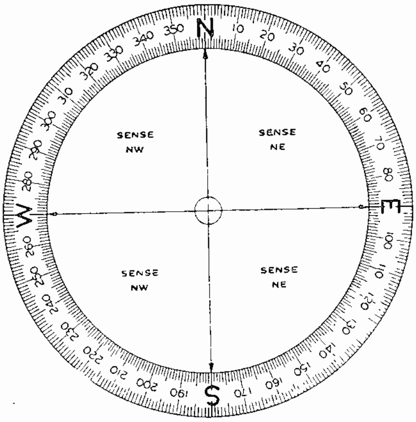


FIG. 3.

The actual procedure is as follows. Each Cathode Ray Tube is first mounted in the direction finder and has marked on it five points in Indian ink. These points correspond to the rest or central position of the spot and the extremities of lines indicating North, South, East and West traces. The tube is then removed from the direction finder and placed in the enlarger and the image of the scale adjusted so that these five marks (actually, of course, defining two lines intersecting at right angles) coincide with the scale.

The adjustments made, the tube is removed from the enlarger and the end thoroughly cleaned, first in nitric acid or, if necessary, to remove grease, in sodium or potassium hydroxide solution, and then in hot water. When it is seen by inspection that no greasy appearance is left, an alkaline albumen substratum is applied bv pouring the solution over the end of the tube, which is kept rotating horizontally and dried rapidly. The tube is then collodionised, again by pouring the collodion over the end of the tube, which is slowly rotated meanwhile. The excess collodion is allowed to drain off and when the coating is just tacky it is sensitised. Various methods of applying the collodion were tried. The orthodox method of flowing the collodion over the tube end could not, of course, be employed, as the end of the tube is convex and not plane. An experimental whirler was made up which held the tube vertically and rotated it slowly while the collodion was poured on the centre, and the application of the collodion by immersion was also tried. It was found, however, that the best results were obtained by rotating the tube almost horizontally but with the end pointing slightly downward and pouring the collodion over the end. A certain amount of dexterity is necessary to obtain a good even coating, but once this has been achieved no trouble should be found.

The collodionised end of the tube is now sensitised by immersing it in a silver bath. The bath is made up according to the well-known wet collodion formula. The normal silver bath consists of 35 grains of silver nitrate per ounce of water acidified by a solution of 1 part nitric acid, 1 part acetic acid to 4 parts water. It was found that better results were obtained by increasing the silver nitrate until a concentration of approximately 40 grains/oz. was reached. The bath was placed in a hemispherical bowl and kept up to strength by addition of silver nitrate from time to time, the strength of the solution being checked by an argentometer. The temperature of the bath must not be allowed to fall too low and the bowl can be kept in warm water with beneficial results. The end of the tube is immersed, by one continuous sweep, into the solution, and is left until all the streaky appearance has disappeared. The initial immersion determines whether the sensitisation is to be effective or not. Any irregularity in immersion will lead to silver marks, and a considerable amount of practice is necessary at this stage before good results can be obtained.

When the tube end has been sensitised, it is now placed in its socket in the enlarger and exposed. The exposure time varies naturally according to light source, negative, etc., but in the case of the tubes intended for direction finding work is of the order of twelve seconds.

Development, fixing and intensification follow. The usual iron developer consisting of an acid solution of ferrous sulphate in water is used, and the image after development is washed and fixed with a potassium cyanide solution.

The normal method of separate intensification, whilst possessing certain advantages, has been discarded for the method wherein intensification is effected by a lead nitrate solution, made up according to the formula :—

Lead Nitrate	1 oz.
Potassium Ferricyanide ..	1 oz.
Acetic Acid	1 oz.
Water	20 oz.

and the image, after being bleached by this, is finally blackened by a solution of sodium sulphide in water.

As regards the solutions employed in the process, it undoubtedly pays to use fresh solutions for each tube. Especially does this apply to the iron developer and sulphide blackening solutions. The silver bath can, of course, be kept the same for many tubes and in fact improves with use.

This method of intensification is both quicker and cheaper than the copper-iodine process and with black and white images produces excellent results.

After the final intensification the image is well washed and dried, and is then varnished. Several types of varnish were tried. Any type of cellulose varnish, whilst giving perfect protection and a very neat appearance was ruled out on account of the fact that it tended to lift the fragile image from the collodion base and to break up the lines. If a coating of albumen or other substance unaffected by the cellulose base was first applied to the unvarnished image and the cellulose varnish then flowed over, this trouble to a certain extent disappeared and this method was employed before it was found that a normal hard shellac negative varnish gave practically as good results and was much quicker in its application. The varnish used consisted of 3 parts of shellac to 30 parts methylated spirit, the resultant solution being decanted before use. In application the end of the tube was first heated to about 80° C. and the varnish flowed over the heated end. As it dries a somewhat opaque coating results, which becomes completely transparent on warming again to about the same temperature. An alternative to a spirit varnish can be made by dissolving borax in boiling water and adding white shellac. The solution is allowed to cool and is decanted. This type of water varnish can be flowed over the image after the final washing. Varnishing is rendered necessary by the fact that the tube is likely to be subjected to repeated handling, and as the image resides wholly on the outer surface of the collodion it is quite fragile and would quickly rub away were it not protected in some way. In addition to protecting the image the varnish considerably enhances its appearance.

Scales photographed in this way serve admirably the purpose for which they are intended. The scale is fully opaque on a transparent background and parallax errors are reduced to a minimum. In addition, any slight irregularity in the end of the tube is automatically compensated for by the manner in which the image of the scale is projected into the tube.

The process is obviously not limited to this type of scale, any markings required being readily transferred to the end of the tube. In addition, coloured scales can be easily obtained by suitably toning the silver image.

L. E. Q. Walker.

PATENT ABSTRACTS

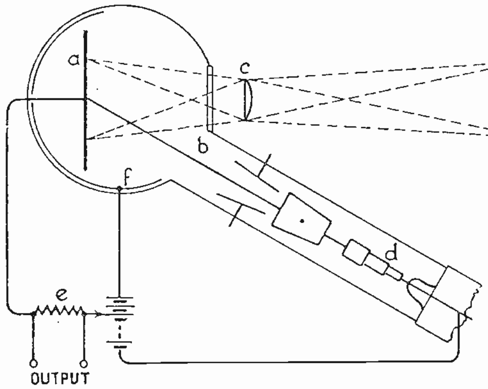
Under this heading abstracts are given of a selection from the most recent inventions originating with the Marconi Co. These abstracts stress the practical application of the devices described.

TELEVISION TRANSMITTING APPARATUS

Application date, December 3rd, 1935. **No. 465,963.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and G. B. Banks.

This invention discloses a form of electron television transmitter tube wherein use is made of the phenomenon that secondary emission from a metallic surface bombarded by electrons is greatly increased by subjecting the surface to light.



PATENT NO. 465,963.

In the type of television transmitter tube disclosed a homogeneous metallic plate (*a*) of low work function and maintained at a positive potential with respect to the cathode of the cathode ray tube gun is scanned by a cathode ray beam (*b*) in a normal fashion at the same time having projected upon it an image from an optical system (*c*). In a circuit connecting the plate (*a*) with the cathode of the cathode ray gun (*d*) there will therefore at any time be two opposing currents: one caused by the primary electrons from the cathode ray beam attracted to the plate and the other by the secondary electrons emitted from the plate. Now the first of these currents will be constant in magnitude for any picture element whilst the second will vary, by virtue of the phenomenon mentioned above, as the light or shade in the particular picture element being scanned.

Across a resistance therefore included in this circuit such as is shown at (*e*) there will be voltages set up corresponding to the light and shade in the picture being scanned.

In use the plate (*a*) is maintained a few volts negative to a metal deposit on the inside of the bulb which facilitates disposal of the secondary electrons.

Preferably the arrangement is such that when the plate (*a*) is in darkness the two opposing currents just cancel each other so that no picture voltage is produced across the resistance.

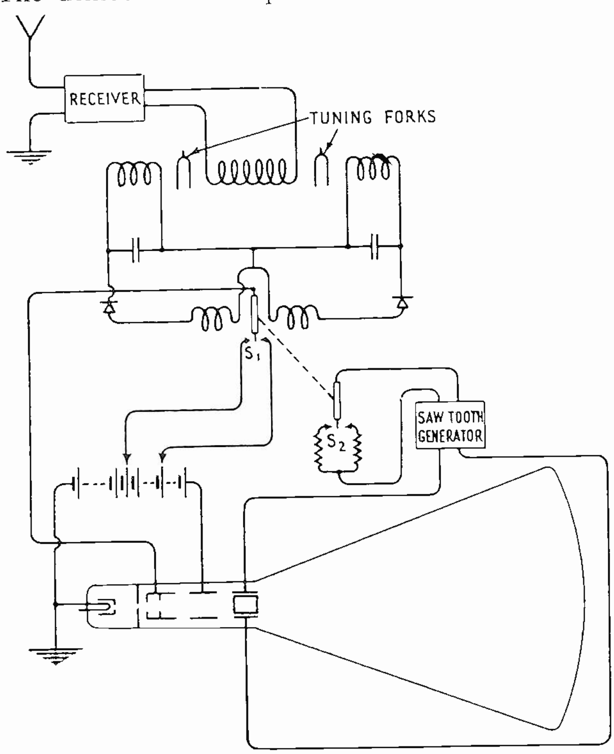
TELEVISION RECEIVERS

Application date, December 3rd, 1935. **No. 466,756.**

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

Automatic changeover of a television receiver line frequency and cathode ray spot size to correspond with changes of transmission is described in this specification.

The circuit, in Fig. 1, shows an arrangement for operating on one of two predetermined line frequencies, but the device is not limited to a choice of two frequencies. The demodulated output from the television receiver is applied to two auxiliary resonant circuits tuned to the predetermined line frequencies. The selectivity of these circuits is increased by using tuning forks or some mechanical resonant device. Each tuned circuit is connected to a detector and the outputs from these are combined differentially to operate relay switches S_1 and S_2. Switch S_1 changes the size of the spot on the cathode ray tube by varying the voltage on the second anode and switch S_2 changes the resistance controlling the frequency of the saw tooth generator giving the receiver line scan.



PATENT No. 466,756.

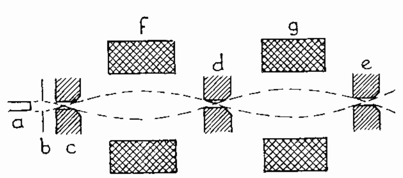
It is possible to replace the mechanical relay by valves the bias for which is derived from the differential detector voltages. Control of line frequency is obtained by variation of valve anode resistance in parallel with the resistance frequency control of the saw tooth generator. Variation of spot size is achieved by change of bias on a valve the anode of which is joined to the second anode of the cathode ray tube which is connected to H.T. supply through a resistance.

CATHODE RAY TUBES

Application date, February 3rd, 1936. **No. 469,897.**

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

This invention utilises the phenomenon of secondary emission to increase the brightness of the fluorescent spot produced by the cathode ray stream in a cathode ray tube.



PATENT No. 469,897.

Referring to the drawing, a cathode modulating electrode and first anode are mounted as shown at *a*, *b* and *c*, in a cathode ray tube. The first anode (*c*) is coated as regards the internal surface of its aperture with some material of low work function such as caesium so that the electrons reaching it will give rise to secondary electrons which fall on a second anode (d) also arranged to emit secondary electrons. Successive apertures such as (e) may be provided, each anode being maintained at a suitably positive potential with reference to the preceding anode.

Focusing of the beam is accomplished first by the configuration of a, b and c, and after passing (c) by suitable electron lens arrangements, either electromagnetic or electrostatic in action, such as are shown at f and g. The track of the cathode ray stream is shown diagrammatically in the figure by the dotted lines, and it will be seen that matters are so arranged that the beam is brought to a focus at or near the aperture of each anode.

Amplification of the original primary cathode ray stream may be obtained up to 5^n where n is the number of anodes at which secondary emission takes place.

HIGH FREQUENCY CIRCUIT ARRANGEMENTS

Application date, April 29th, 1936. **No. 474,382.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and N. M. Rust and E. F. Goodenough.

The use of band pass filters of the normal type is restricted because of the variable phase shifts produced over the frequency pass band. They may not, for

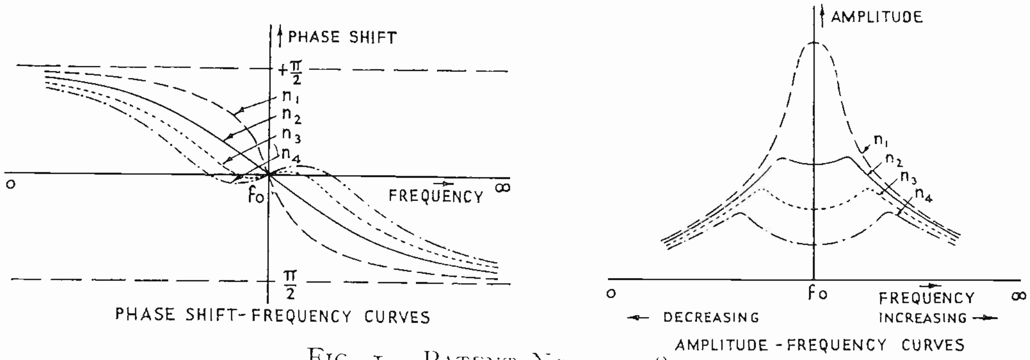


Fig. 1. Patent No. 474,382.

example, be used in reaction circuits since this phase change produces frequency discrimination which will seriously modify the overall amplitude characteristic of the pass region.

A special type of band pass filter, in which the phase shifts may be controlled or almost entirely eliminated over the pass band, is disclosed in this specification. Two particular methods are given; the first involves the use of a parallel resonant circuit shunted by a series circuit tuned to the same frequency. By varying the ratio $\frac{\text{series circuit resonant admittance}}{\text{parallel circuit resonant admittance}}$ (this is designated by the letter "n") various degrees of phase shift and amplitude characteristics may be obtained. Typical curves of phase shift and amplitude are shown in Fig. 1 for different values of n and it will be noted that for a particular value of $n=n_3$ the phase shift over most of the pass band is almost zero. Increase of n above the value n_3 produces reversed phase shift which can be used to compensate for the phase shift produced in any other part of the circuit.

The second method uses two parallel resonant circuits in series and tuned to the limit frequencies of the pass band. The two circuits are designed to have vectors of equal amplitude and phase angles of $\pm\tan^{-1}K$ at the mid frequency f_0 $(\sqrt{f_1 f_2})$. The phase shift and amplitude curves obtained by this method are identical with those given in Fig. 1 if $K^2=n$. Modifications of these two circuits, which may be designed to yield the same results, are discussed.

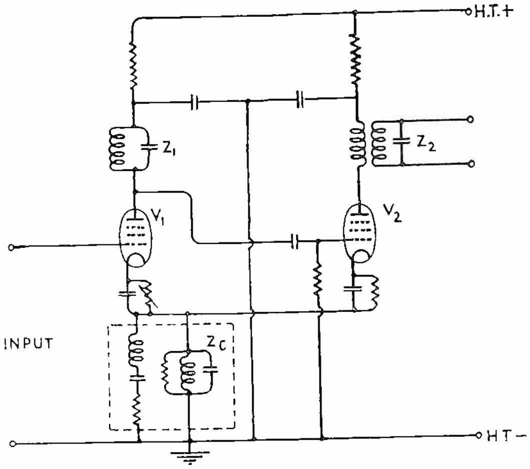


FIG. 2. PATENT NO. 474,382.

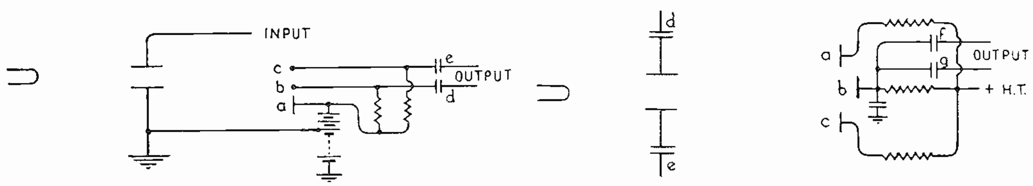
These filters may be used in all circuits where negligible or controlled phase shift is required over a given pass band, such as for television and especially for circuits in which reaction is employed. One form of reaction circuit is illustrated in Fig. 2. The cathode impedance Z_c consists of the special band pass filter. The ratio n is governed by the form of the first anode coupling impedance Z_1. If Z_1 is aperiodic, Z_c must be designed for zero phase shift over the pass band, but where Z_1 is a tuned circuit, Z_c must be designed to give a phase shift curve such as that for n_4 in Fig. 1. Compensation for the phase shifts produced by Z_1 then occurs. The highest gain is obtained from such a reaction arrangement when Z_1 is a tuned circuit.

TELEVISION RECEIVERS

Application date, April 29th, 1936. **No. 474,386.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and D. L. Plaistowe.

This specification discloses means for separating synchronising signals from picture signals in television receiving systems.



PATENT NO. 474,386.

A cathode ray tube having three anodes is employed, these three anodes being co-planar and perpendicular to the normal direction of the cathode ray stream. In the first application one anode consists of a rectangular plate and the other two anodes are small. They are represented diagrammatically in the first figure as a, b and c.

The input signal consisting of television and synchronising signals combined are applied to one of a pair of deflector plates, the other plate being maintained at a constant positive potential. If the synchronising signals are supposed of higher amplitude than the picture signals, matters may be so arranged that the picture signal only causes the ray to be deflected over (*a*) whereas line synchronising signals, being of superior amplitude, may cause the ray to be deflected to (*b*) and thence send a separated synchronising signal through a condenser (*d*) to the output. Similarly frame synchronising signals being superior in amplitude, both to the line synchronising signals and to the picture signals, may cause the ray to be deflected to (*c*) and thus cause an output frame synchronising signal to be sent out through the condenser (*e*).

In this way frame and synchronising signals may be separated from each other and from the television signals and may be applied to their respective time bases.

In a second application which deals with cases where the synchronising signals correspond to zero carrier a cathode ray tube is provided with three anodes, *a*, *b* and *c*.

Modulated unrectified signals are applied in push-pull to two deflector plates through condensers (*d*) and (*e*). During normal picture modulation the ray sweeps between (*a*) and (*c*), but during synchronising periods it remains on (*b*) and a mixed synchronising signal is sent to line and separated for the two time bases by appropriate selector circuits.

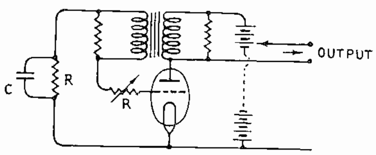
A third modification operates on a similar system to the above but synchronising signals are produced by the positive potentials assumed by the two outside anodes during the intervals when no neutralising negative charges are incident on them from the cathode ray beam, i.e., during the synchronising periods.

ELECTRICAL OSCILLATION GENERATORS

Application date, February 3rd, 1936. **No. 469,898.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and S. W. H. W. Falloon.

For measuring the height of the various layers of the ionosphere, it is usual practice nowadays to employ a cathode ray receiver of the type that shows the direct signal and echoes on a linear time screen and send from what is technically known as a pulse transmitter. A pulse transmitter consists of a low power transmitter sending a regular series of very short pulse signals. It is essential that the duration of the pulses should be only a fraction of a second, usually $\frac{1}{5,000}$th or less, as if the signal lasts too long, the echo signals are not separated sufficiently for purposes of analysis. Hitherto it has been usual to make periodic pulse signals locked from the A.C. 50 cycle main. It is, however, an advantage to be able to build a transmitter which can be set up in a place not provided with mains supply, and the present invention describes a method of producing a transmitter delivering a series of pulses with a constant repetition frequency independent of the A.C. mains. The oscillator circuit is shown in Fig. 1, from which it can be seen that the anode circuit is back coupled



PATENT NO. 469,898.

to the grid through a transformer, and the grid contains a CR circuit which determines the repetition frequency, and a resistance, R, which controls the duration of pulse (usually about 300 per sec.).

The operation of the circuit is as follows. When H.T. is switched on, anode current increases, but due to back coupling, the E.M.F. induced in the grid circuit causes grid current to flow and this charges up the condenser C, the effect being to back off the grid and stop the anode current. The charge then leaks away and the cycle is repeated. Thus it can be seen to be a type of squegger oscillator.

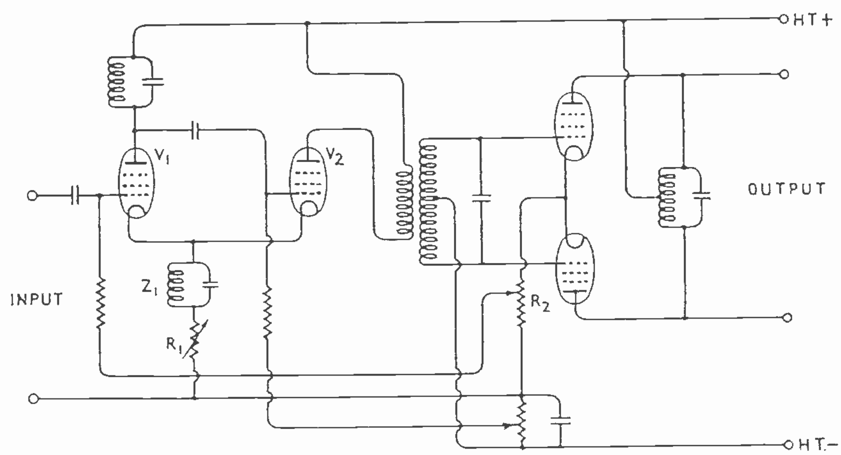
Such a pulse transmitter is then used to modulate a key of a low power wireless transmitter by connecting the output to the grid input circuit of the modulator of the transmitter.

ELECTRON DISCHARGE DEVICE CIRCUITS

Application date, February 8th, 1936. **No. 470,085.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and N. M. Rust and E. F. Goodenough.

The object of this invention is the cancellation of distortion produced by non-linearity of the I_a E_g characteristic curve of a valve. The most important distortion term in the power series representing the I_a E_g relationship is generally the square law term (a_2 E_g^2), which produces second harmonic distortion and this may be



PATENT NO. 470,085.

cancelled by antiphase feedback of the correct amplitude of second harmonic. This second harmonic is derived from the common cathode circuit of two valves operating in push-pull and having I_a E_g characteristic curves of the same form as the distorting valve. A second harmonic voltage is thus produced across the common cathode resistance and this is fed back to the input of the first valve in such a way that cancellation of the second harmonic distortion produced by this valve occurs. This method of compensation is particularly applicable to amplifiers with controlled reaction and it is found that a high degree of reaction may be obtained with low distortion, resulting in high gain, high selectivity and low cross modulation.

The principle may be applied to audio and radio frequency amplifiers and to oscillators for controlling the output voltage with low distortion and improved frequency stability.

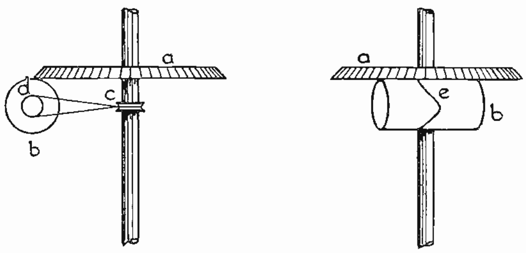
A typical R.F. reaction circuit with this form of compensation is shown in Fig. 1. Valves V_1 and V_2 have a common cathode impedance Z_1 and a reaction control resistance R_1. The output from V_2 is fed to two push-pull valves each having a characteristic I_a E_g curve producing second harmonic distortion. Owing to the push-pull connection this second harmonic distortion cancels in the output circuit. It is, however, present in the common cathode resistance R_2 and is fed back to the valve V_1 so as to cancel the second harmonic distortion normally produced by the first two valves.

DIRECTION FINDING APPARATUS

Application date, June 25th, 1936. **No. 477,346.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and R. J. Kemp.

This specification relates to automatic correction of direction finder readings for certain forms of errors.



PATENT NO. 477,346.

In connection with the moving scale of the installation on which the bearings are obtained, a correction chart in accordance with the quadrantal and other errors of the installation is prepared; this chart being so arranged that it acts as the pointer or index of the moving scale and rotates with the scale in such a way that the reading of the scale with reference to the pointer gives a corrected reading.

The diagram illustrates the action of the apparatus. The rotating scale (*a*), marked in degrees, rotates also through a gear (*cd*) a cylinder (*b*) on which a line (*e*) is engraved which acts as the index to the scale on (*a*). The disposition of (*e*) on (*b*) and the gearing is so chosen that the readings on (*a*) with reference to the index (*e*) are automatically corrected.

ULTRA SHORT WAVE VALVE APPARATUS

Application date, June 26th, 1936. **No. 477,362.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and N. E. Davis, E. Green and A. W. Hall.

This patent describes methods of construction of ultra short wave transmitting sets (below 10 metres) for medium and large powers. With the growth of communication systems employing ultra high frequencies such as may be used for television systems, the efficiency of a transmitting system depends more upon the mechanical design of the circuit than upon any fundamental change in circuit connection, i.e., ultra short wave transmitters employ conventional circuits consisting usually of a constant frequency drive, a series of frequency doubling circuits, and a series of power amplifiers at the signal frequency neutralised in the usual way. Unless special precautions are taken, however, in the design of these signal frequency

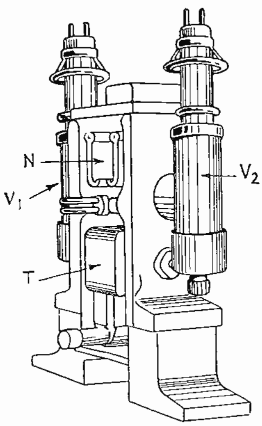


FIG. 1.

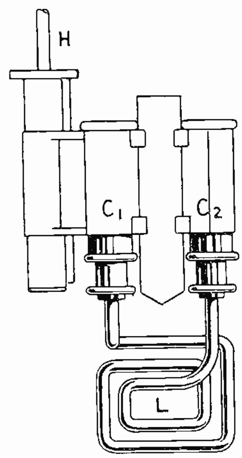


FIG. 2.

PATENT NO. 477,362.

amplifiers, such as the elimination where possible of insulating material in high frequency fields, the elimination of connecting leads between circuits, and design to assure uniform current distribution in the components, considerable loss will be experienced and in consequence low efficiencies. The present invention describes a method of mechanical design which enables high efficiencies to be obtained with large power amplifying systems employing fluid-cooled valves.

Fig. 1 shows an isometric projection of the output stage of a large water-cooled ultra short wave transmitter in which V_1 and V_2 are the valves and valve jackets bolted to a hollow metal box in which the cooling water also circulates. Between the inner faces of the two cooling boxes is a sliding box-like structure constituting the anode tuning condenser T, and above this are joined hinged variable plates for neutralisation purposes N. Bolted directly to the face of these box structures is the tuning inductance consisting usually of a one or two turn " pancake " coil through which the cooling water flows, the ends of the coil being fitted to sliding tubular condensers capable of variation. In certain variations of this invention the centre of the coil is broken by a tubular condenser in order that the anodes of the two valves may be separately excited from the D.C. supply.

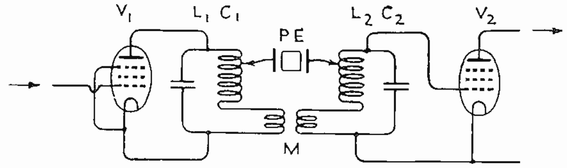
The tubular condensers described in the patent are of special importance as the current distribution is very uniform, and Fig. 2 shows such a coil and condenser assembly in which L is a " pancake " coil, C_1 and C_2 variable tubular condensers, controlled from the handle H.

PIEZO-ELECTRIC CRYSTAL CIRCUIT ARRANGEMENTS

Application date, April 29th, 1936. **No. 474,387.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and N. M. Rust.

This patent specification refers to the control of the response characteristics of a crystal gate coupled circuit. If we consider the case of two parallel tuned circuits coupled through the intermediary of the piezo-electric crystal, it is found that although the latter may be regarded as a series resonant circuit at the natural frequency of the crystal, owing to the carry-through capacity of the crystal and its associated holder, the admittance curve of the system, and in consequence the output volts frequency curve, is not symmetrical, but asymmetrical.



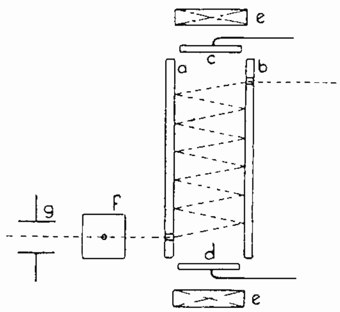
PATENT NO. 474,387.

This is because the electrical equivalent of a piezo-electric crystal is to be thought of as a series resonant circuit having very high "Q" value shunted by a capacity which is large compared with the series resonant capacity of the crystal itself. In consequence the dip on the side of the resonance curve may be quite pronounced. Now it may be desirable in certain cases to smooth out this dip or even accentuate it to form a crevasse in order that we can obtain both an acceptance of frequencies at the resonance frequency of the crystal and a rejection of nearby frequencies. In the present invention the control of the shape of the curve, and, in particular, the shape of the crevasse formed beside the resonance peak is accomplished by providing a variable magnetic coupling between the two circuits, and by providing tapped points so that the circuit crystal can be matched to the circuits It is found that careful adjustment of the tapping points and the mutual will enable one to balance out the carry-through susceptance of the crystal or to accentuate the crevasse as desired.

CATHODE RAY TUBES

Application date, June 25th, 1936. **No. 477,345**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and L. M. Myers



PATENT NO. 477,345.

This specification deals with the production of relatively great electron beam currents in cathode ray tubes of simple construction.

A cathode ray tube consisting of the usual gun which projects the cathode ray beam in the direction shown in the diagram, has a pair of secondary emitter electrodes, *a*, *b*, provided on its envelope. The beam passes through an aperture in (*b*), is brought to a focus on (*a*) and causes secondary emission therefrom. The electrons emitted from (*a*) now pass back to (*b*), where they give rise to secondary electrons which impact on (*a*), and so on, until the multiplied electron beam finally passes through an aperture in (*a*) and out to the normal deflection plate system, *f* and *g*.

A suitable D.C. potential is supplied between two plates *c* and *d* at right angles to *a* and *b* in order to cause the staggered electron path shown in the diagram. Electrons emitted from one plate are brought to a focus on the other by means of a magnetic field set up in a coil (*e*) which produces a field substantially normal to the plates *a* and *b*. If a very high frequency electrostatic field is set up between *a* and *b* and if the frequency of this field is adjusted suitably electron reflections with multiplication at each reflection will take place between *a* and *b*.

In this way a considerably magnified cathode ray beam can be produced.

Preferably space charge grids may be placed in front of each secondary emitter plate.

PIEZO-ELECTRIC CRYSTALS

Application date, June 25th, 1936. **No. 477,344.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and O. E. Keall and E. O. Smith.

This patent describes an unconventional form of crystal holder. Instead of positioning the crystal between solid contacts or held free by cord suspension, it is held between flat surface contacts made by a solid backplate to which are fixed pads of steel wool, the plated faces of the crystal being held between these pads of steel wool and lateral movement being prevented by means of a mica bridge. This virtually holds the crystal in free space between the solid contacts.

The patent describes various methods of making up this crystal holder, and, in particular, one supported in an evacuated envelope.