WIRELESS ENGINEER

Vol. 32 MARCH 1955 No. 3

Magnetic Demodulation

CONSIDERABLE work has been done on this subject in recent years at the Brunswick Technische Hochschule by Professor L. Pungs and Dr. G. Meinshausen. The latter has recently published* a very interesting article in which he discusses the close relationship between magnetic demodulation and demodulation by means of a diode. The basis of the relationship can be seen by interchanging the axes of the ordinary magnetization curve and plotting the flux horizontally and the magnetizing current vertically as in Fig. 1. If a steady current i_0 is applied, so that the operating point is at the knee of the curve, the characteristic, neglecting hysteresis, is very similar to that of a diode over a wide range. Fig. 2 shows an iron core with a winding B through which a steady current i_0 is maintained, and a winding A through which the generator maintains a high-frequency current i_h. The choking coil D reduces any alternating current in B to a negligible value.

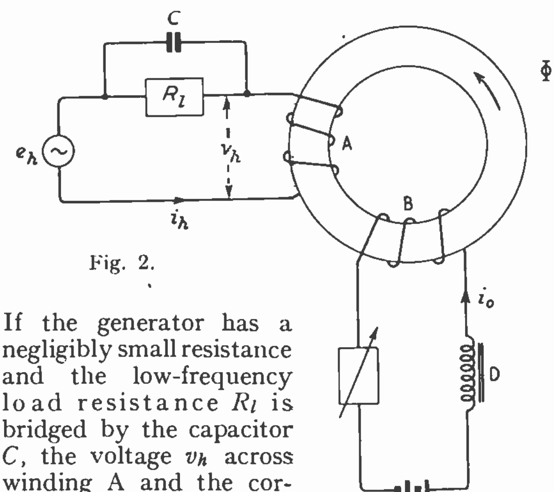


Fig. 2.

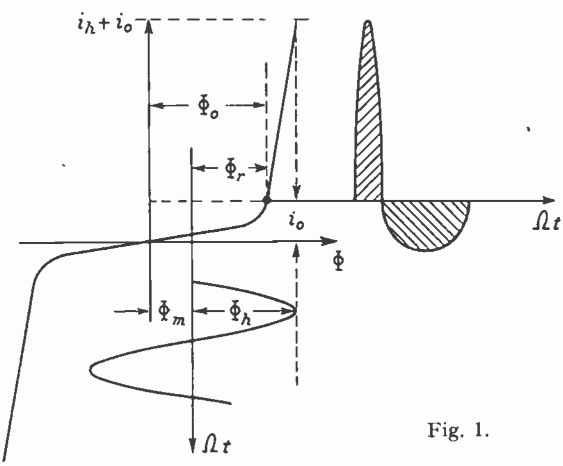


Fig. 1.

If the generator has a negligibly small resistance and the low-frequency load resistance R_l is bridged by the capacitor C, the voltage v_h across winding A and the corresponding flux Φ_h will both be sinusoidal, but the current i_h will be as shown on the right of Fig. 1. The integral $\int i_h dt = 0$, hence the two shaded areas are equal. The resultant flux $\Phi = \Phi_h \sin \Omega t + \Phi_m$, where Ω is the angular high frequency; the corresponding total current linking the core (assuming the coils to be of single turns) is $i_h + i_0$. The steady flux upon which the alternating flux is superposed is reduced from Φ_0 to $\Phi_0 - \Phi_r = \Phi_m$, which will vary with the amplitude of Φ_h and therefore follow the modulation and induce a corresponding low-frequency e.m.f.

* Pungs, *E.N.T.*, Vol. 15, 1938, p. 161. Pungs and Meinshausen, *Frequenz*, Vol. 7, 1953, p. 153. Meinshausen, Hochschule Dissertation, 1953. Meinshausen, *Archiv der Elektrischen Übertragung*, Vol. 8, Sept. 1954, p. 373.

The suggested equivalent circuits of the diode and magnetic demodulators are shown in Figs. 3(a) and (b). In Fig. 3(a) the diode is assumed to have infinite resistance in one direction and a constant resistance R_i in the other. R_a is the low-frequency load resistance; the capacitor C_l prevents the low-frequency voltage being short-circuited by the generator, which is assumed to have negligible resistance. In Fig. 3(b) R_l constitutes the low-frequency load resistance; the steady bias magnetizing current is represented by i_0. Just as the diode can be represented by two resistances in series, which function alternately, depending on the direction of the current, so the magnetizing circuit can be represented by two inductances L and L_e in parallel, where L_e is the very high inductance over the unsaturated portion of the characteristic and L a much smaller inductance, which comes into parallel with L_e and reduces the resultant inductance to the correct value L_0 over the saturated part of the cycle. The current i through R_l will then follow the curve on the right of Fig. 1.

The author sets out the equivalent formulae for the two types of demodulator in parallel columns, for example, in Fig. 3(a)

$$\frac{de_h}{dt} = \frac{i}{C_l} + \frac{d}{dt}(Ri_1),$$

whereas in Fig. 3(b) $e_h = iR_l + \frac{d}{dt}(Li_1)$ and so, assuming that the mathematical formulae for the diode modulator are known, those for the magnetic case can be written down by comparison.

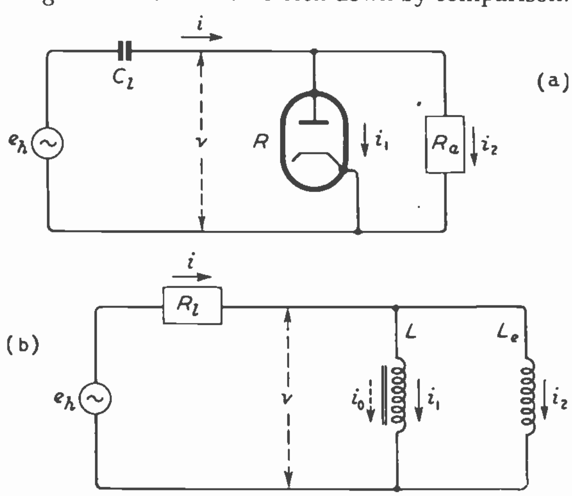


Fig. 3.

The development is simplified by assuming the straight-line magnetization curve shown in Fig. 4, for which the author determines the relation between the mean value Φ_m of the flux and the current i_r for a constant value of the flux amplitude Φ_h. In Fig. 4 this is shown by point A, and the dotted curve shows how the position of A varies with change of the current i_r.

We do not propose to go into the subject more fully; we have merely given an indication of the interesting parallel between the diode and the magnetic demodulator. The original paper goes into the matter very fully; measurements are described which were made by winding a third coil on the iron core and connecting it to a ballistic galvanometer and to a cathode-ray oscilloscope. In addition to the assumption of straight lines, calculations are also made on the assumption that the characteristic curves are exponential.

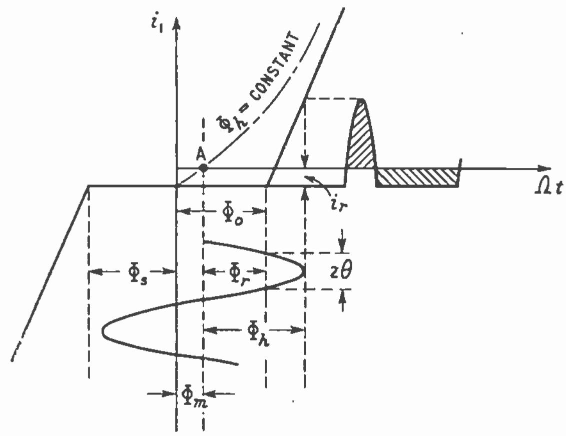


Fig. 4.

There are many difficulties to overcome before the magnetic demodulator can be regarded as a serious competitor with the diode. One of the disadvantages is that to obtain distortionless demodulation with 100% depth of modulation, it is necessary to re-adjust the biasing current i_0 whenever the carrier voltage changes, whereas no such re-adjustment is necessary with a properly adjusted diode. Another disadvantage is that it is only of practical importance if the output is fed into a load of very low resistance—the author says "approaching short-circuit"—but under these conditions it is more efficient than the diode. The tests were carried out at low carrier frequencies; at higher frequencies the losses would necessitate the use of ferrites which have a lower maximum permeability and a less sharply pronounced knee in the magnetization curve. Their future depends on the development of suitable ferrites and the author concludes by confessing that one cannot yet predict whether the magnetic demodulator will be superior to the diode, although it can be shown that an ideal magnetic demodulator could have advantages over an ideal diode demodulator.

G. W. O. H.

WAVEGUIDES AND WAVEGUIDE JUNCTIONS

Geometrical Analysis of their Properties

By J. M. C. Dukes, M.A., D.I.C., A.M.I.E.E.

SUMMARY.—It has been shown elsewhere[7–9] that the properties of waveguides and waveguide junctions may be expressed in terms of certain theorems in non-Euclidean geometry. In the present instance the discussion is limited to those aspects of the topic which do not require concepts more advanced than those of simple plane geometry. A method of measurement and analysis is described, with examples of its use in typical situations. It is shown that the method has considerable advantage when measurements must be made through an unavoidably mismatched junction.

LIST OF SYMBOLS

u = complex coefficient defining amplitude and phase of wave travelling towards load.
v = complex coefficient of wave travelling back from load.
$\Gamma = v/u$ = reflection coefficient (complex).
$|\Gamma|$ = modulus (amplitude) of Γ.
$\arg \Gamma$ = argument (phase) of Γ.
r = voltage standing-wave ratio.
u' = complex coefficient defining amplitude and phase of wave travelling towards junction.
v' = complex coefficient of wave travelling back from junction.
$\Gamma' = v'/u'$ = image of Γ by the junction.
S_{11} = scattering coefficient, input terminals of junction.
S_{22} = scattering coefficient, output terminals of junction.
S_{12} = scattering coefficient, transmission through junction.
l = length of transmission line between junction and load.
L = insertion loss of the junction in decibels.

1. Introduction

THIS article is concerned with the problem of measuring the properties of a microwave component through a junction, when this junction, for unavoidable reasons, provides rather a poor match between the measuring equipment and the waveguide system of which the component forms a part, and consequently any measured results thus obtained are not directly representative of the properties of the component.

If the parameters of the junction are known then the measured results may be manipulated so as to eliminate the effect of the junction. Such a calculation, however, is tedious if performed algebraically, as the properties of the junction cannot be specified properly in terms of less than three independent parameters, each of which is complex.

Now it is a well-established property of passive two-terminal-pair networks that the relationship between the input and output conditions is that known mathematically as a bilinear transformation[1]. In the case of poly-terminal-pair networks, which will not be dealt with here, a bilinear relationship exists between any two terminal-pairs, but the complex constants are then a function also of the conditions existing at the other terminal-pairs.

By exploiting the geometrical properties of the bilinear transformation[2] it is accordingly possible to express the input–output relationship of a junction in a geometrical manner. That is to say, if we have a set of points on a diagram such as the Smith chart, this set of points will be transformed into another set of points situated elsewhere on the chart, and the relationship between the two sets, which may be determined geometrically, specifies the properties of the junction. In the special case where the set of points lie on the circumference of a circle, then the transformed set of points will also lie on the circumference of a circle. In this case the geometrical relationship is particularly easy to identify.

The method is essentially practical in character and is of the greatest value in the laboratory. Not only is the tedium of such calculations reduced, but also the liability to numerical error. Moreover, the method permits the visual averaging of sets of readings of questionable accuracy, and anomalous effects, such as multi-moding, are shown up as a destruction of the geometry of the system.

The geometric approach has been developed by a number of writers[3–6] but the most significant contribution of recent years is the introduction by G. A. Deschamps[7–11] of certain concepts, due principally to Cayley (1821-95), Klein (1849-1925) and Poincaré (1854-1912), in the field of non-Euclidean projective geometry. In order that this article may have the widest possible appeal, only a few topics will be treated, and these topics will be expounded in terms of elementary Euclidean geometry in the plane It is hoped, however, to publish a subsequent article

MS accepted by the Editor, March 1954

describing other aspects and applications of the technique[15].

It should be stressed that the method is applicable to any type of transmission line or medium where propagation takes place in one single mode.

2. Background[1]

In Fig. 1 is shown in diagrammatic form the experimental set-up for the measurement of the properties of a length of waveguide (or transmission line) 'bc'. The designation 'fixed termination' implies that the electrical properties of the termination, when measured at the reference plane 'c', are invariant of the length l of the line 'bc'.

At this stage the following simplifying assumptions will be made:

(i) The junction at 'b' between the line 'ab' and 'bc' is perfectly matched and lossless.

(ii) The losses in the line 'ab' are negligibly small over a range of several wavelengths.

(iii) For convenience, the reference plane of the s.w.r. detector will be made coincident with the junction at 'b'.

The state of the field in the line 'abc' may be described by means of two parameters u and v, where u is a complex quantity defining the amplitude and phase of the incident wave at a given point along the line, and v is a complex quantity at the same point of the wave reflected by the fixed termination. The intensity of the field at this point, as measured by the probe, will be proportional to $| u + v |$.

If it is assumed that the line 'ab' is virtually lossless then it is not difficult to show that $| u + v |^2$ varies sinusoidally along the line in the following manner:—

$$|u+v|^2 = |u|^2 + |v|^2 + 2.|u|.|v|.\cos 4\pi x/\lambda_g \quad (1)$$

where x is the distance along the line from some suitable reference plane and λ_g is the wavelength in the guide. Hence to describe the state of the field in the line, the only measurements required are:—

(i) The value of $| u + v |_{max}$

(ii) The value of $| u + v |_{min}$

(iii) The distance x of a minimum from the reference plane 'b', (Fig. 1.)

(iv) The spacing between successive minima.

From the first two measurements we obtain the voltage standing-wave ratio r:—

$$r = \frac{| u + v |_{max}}{| u + v |_{min}} = \frac{| u | + | v |}{| u | - | v |} \quad \ldots \quad (2)$$

From a knowledge of the spacing of the minima we obtain the wavelength λ_g in the line 'ab'. Assuming we know the wavelength in the line 'bc' and the position x of one minimum then we can determine the conditions at the reference plane 'c', and hence the properties of the component under test.

If, however, we do not yet know the wavelength in the line 'bc' we determine it by varying the length of the line 'bc'. Thus if a change Δl in the length of the line 'bc' produces a shift in the position of the minimum as measured in the line 'ab' by an amount equal to say half a wavelength, then we know that the wavelength in the line 'bc' is $2\Delta l$.

A convenient parameter to work with is the reflection coefficient Γ defined as:—

$$\Gamma = v/u \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

The reflection coefficient therefore describes, by means of one complex number only, the state of the wave at any given point along the line 'ab'. From equations (2) and (3) it is easy to obtain the relationship between the reflection coefficient Γ and the standing-wave ratio r, namely:—

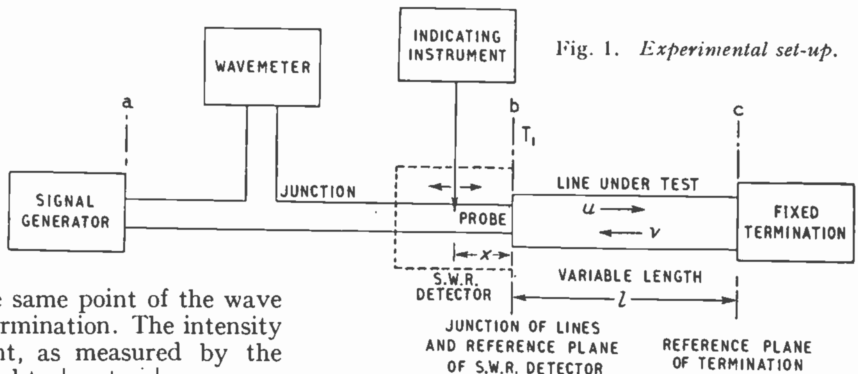


Fig. 1. *Experimental set-up.*

$$| \Gamma | = \frac{r - 1}{r + 1} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

It is important at this stage to see clearly how Γ, when measured at the reference plane 'b', varies when the length l of the line 'bc' is varied. This is illustrated in Fig. 2. The point Γ_0 corresponds to zero length of line 'bc', and hence represents the reflection coefficient of the fixed termination. As l is increased the point Γ traces a spiral curve, the rate of change of $| \Gamma |$ with respect to l being a function of the attenuation of the line 'bc'.

In most cases of practical interest the loss in the line 'bc' is so small that there is very little change in $| \Gamma |$ over a range of one half wavelength (i.e., one complete rotation of Γ in the Γ plane). Hence by making two sets of measurements with two very different lengths of line l_1 and l_2 we obtain, for all practical purposes, two isolated circles as in Fig. 3. In addition to the data which may have been obtained by the means already discussed, we may also evaluate the attenuation in the line 'bc'. Thus the attenuation α of the length $l_2 - l_1$ is given by :—

$$\alpha = 10 \log_{10} (\mid \Gamma_1 \mid / \mid \Gamma_2 \mid) \text{ db} \quad .. \quad (5)$$

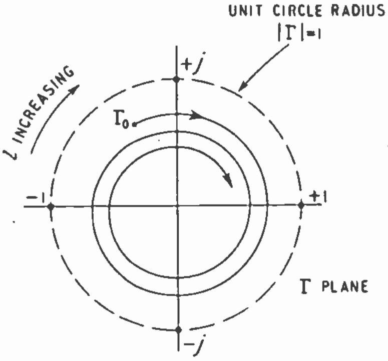


Fig. 2. *Variation of reflection coefficient with variation in length of terminated line.*

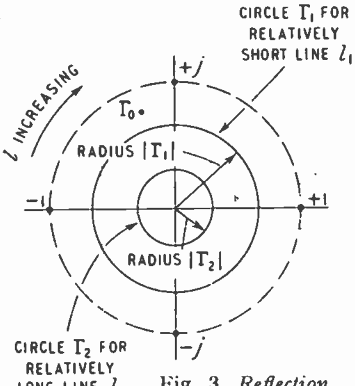


Fig. 3. *Reflection coefficient circles.*

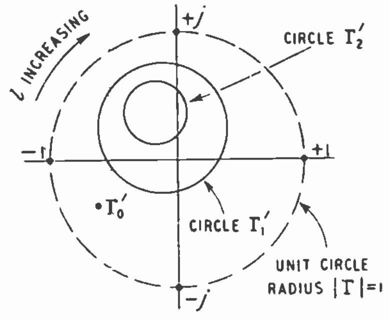


Fig. 4. *Image reflection coefficient circles.*

3.2. *Junction as a Bilinear Transformation*

In Fig. 6 is shown an enlarged sketch of the junction between the two lines. On the right-hand side we have a reflection coefficient Γ and on the left-hand side we have a reflection coefficient Γ' both measured with respect to the common reference plane of the junction. Now, provided that the junction is linear and passive,

Hence the attenuation per unit length in decibels

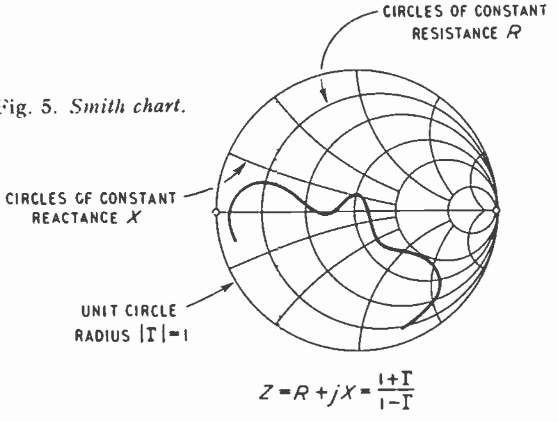
$$= \{10 \log_{10} (\mid \Gamma_1 \mid / \mid \Gamma_2 \mid)\} \div (l_2 - l_1) \quad (6)$$

(These relationships follow from the fact that the incident wave u decreases exponentially in amplitude as it travels towards the fixed termination, while the reflected wave v decreases exponentially in amplitude as it travels *away* from the fixed termination.)

In practice it is convenient to choose as a fixed termination either an open circuit or a short circuit. In the first instance Γ_0 in Fig. 2 moves to $+ 1$ and in the second instance to $- 1$.

In conclusion, it should be pointed out that the Γ plane is in fact the plane of the Smith Chart[1] and as such may be more familiar to many readers. The way in which the lines of constant resistance and reactance are superimposed upon the Γ plane is illustrated in Fig. 5.

Fig. 5. *Smith chart.*



3. Determination of the Properties of a Line Measured Through a Mismatched Junction

3.1. *Experimental Set-Up*

The experimental set-up is as described in Section 2 and illustrated in Fig. 1. However, we now remove the restriction that the junction between the lines 'ab' and 'bc' be matched, with the result that Fig. 3 is modified in the general manner indicated in Fig. 4. It will be noted that although the two circles have suffered both displacement and contraction in size, they still remain circles. This property, which is very important, and certain others, are discussed below.

it can be shown that Γ and Γ' must always be related by an equation of the general form

$$\Gamma' = \frac{a \, \Gamma + b}{c \, \Gamma + d} \quad .. \quad .. \quad .. \quad (7)$$

where a, b, c, d, are complex quantities. However, since we can divide through by either c or d there are only three essentially *independent* parameters, each complex.

The more important properties of a bilinear transformation[2] are:—

(1) The mapping is conformal. In other words, a *small* figure in the Γ plane is preserved in shape in the Γ' plane, although the scale may be changed and the figure rotated.

(2) If two lines cross at a given angle in the Γ plane they will cross at the same angle in the Γ' plane.

(3) Any circle, whether *large* or *small* in the Γ plane is mapped by the transformation into a circle in the Γ' plane and vice versa.

(4) The result of two successive bilinear transformations is itself a bilinear transformation.

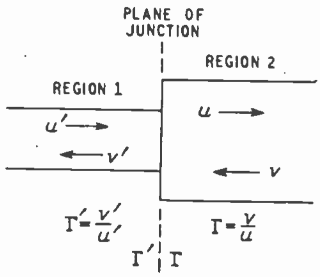
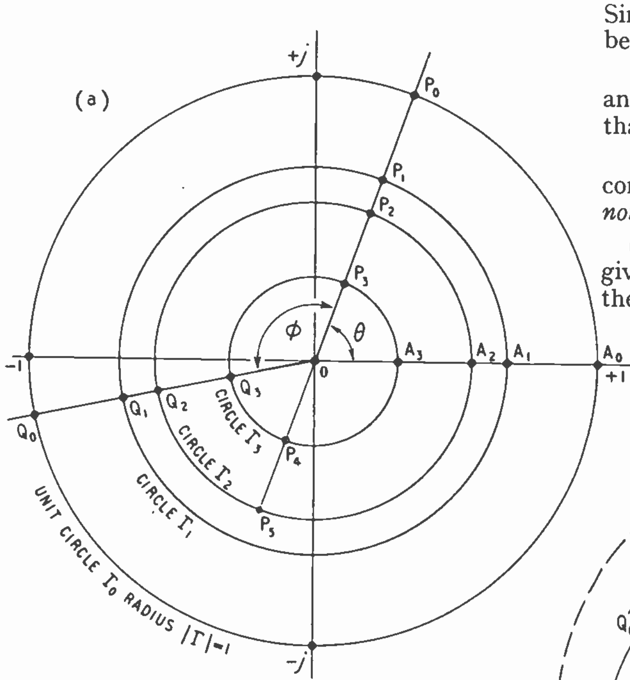


Fig. 6. *Junction relationships.*

In order to illustrate these principles more clearly, and in greater detail, Figs. 3 and 4 have been redrawn in Fig. 7 (a) and (b) to a larger scale and in modified form. An extra circle Γ_3' has been included corresponding to an even longer length of line, l_3. Both diagrams are drawn to the same scale, and, therefore, are each enclosed within the same unit circle radius $|\Gamma| = 1$ (also the edge of the Smith chart).



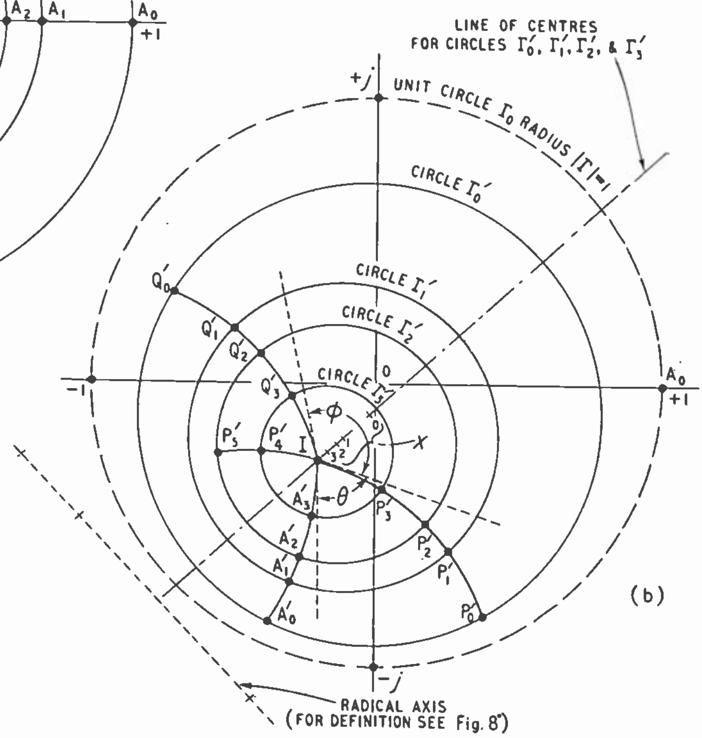


Fig. 7. (a) *Enlarged diagram of reflection coefficient circles;* (b) *enlarged diagram of image reflection coefficient circles.*

The following properties, can be deduced from these diagrams, based on the above theorems. They will not be proved but merely stated:—

(1) The set of circles Γ_0', Γ_1', Γ_2', Γ_3' are the 'image' by the junction of the set of circles Γ_0, Γ_1, Γ_2, Γ_3. (The effect of a junction is analogous to that of a lens in an optical system).

(2) These circles have a common line of centres. The centre of the circle of zero radius is the point I, known as the *Iconocentre*. This point is in fact the image by the junction of the point O, the centre of the unit circle, or Smith Chart. *Hence it represents the value of Γ' as measured when the fixed termination comprises a power-absorbing load whose impedance is matched to the characteristic impedance of the line 'bc'.*

(3) The circular arcs IA_0', IP_0', IQ_0' are the image by the junction of the radii OA_0, OP_0, OQ_0. Since the mapping is conformal these arcs must be orthogonal to the circles Γ_0', Γ_1', Γ_2', Γ_3'.

(4) These arcs intersect at the Iconocentre I, and the angular relationship at I is the same as that between the radii OA_0, OP_0 and OQ_0.

(5) A set of such circles Γ_0', Γ_1', Γ_2', etc., constitutes a family. (Note that the circle Γ_0 is *not* a member of this family.)

(6) If any two member circles of a family are given, then the position of the Iconocentre I is thereby completely defined.

(7) Conversely if one circle and the Iconocentre are given, then the remaining members of the family are thereby defined.

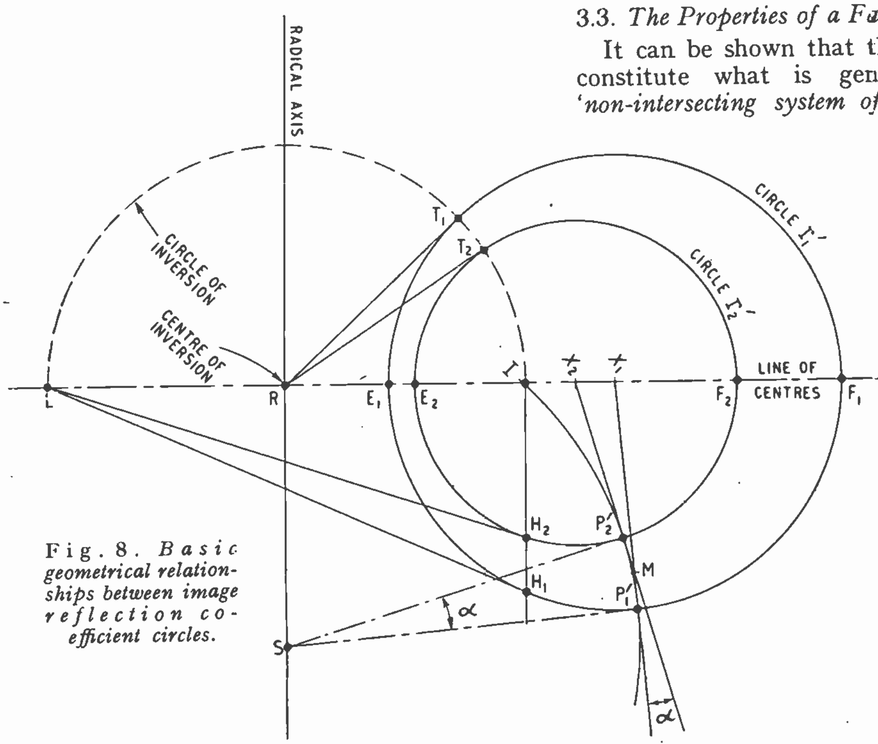


Fig. 8. *Basic geometrical relationships between image reflection coefficient circles.*

From the above considerations it should be evident:—

(1) That to determine the properties of the line we require exactly the same data, whether the junction is matched or not. (Given that the junction is matched only one point per circle is required to define the circle. In practical situations it is not usually known in advance whether the junction is matched or not.)

In other words we must define two circles, and this requires a minimum of six separate measurements. (In practice it is preferable to have four points per circle, spaced approximately $\lambda/8$ apart; i.e., a total of eight separate measurements.)

(2) That it should be possible to deduce from this data the properties of the junction. This follows from the fact that two circles define the whole family, and the relationship between the members of the family (relative to the boundary of the Smith chart) therefore defines the four complex quantities *a*, *b*, *c*, and *d* in equation (7).

It will now be shown that the task of determining the properties of either the line or the junction from the circle diagram of Fig. 7 (b) is not difficult.

First, however, it is necessary to demonstrate certain elementary relationships between the member circles of a family.

3.3. *The Properties of a Family*

It can be shown that the family of Γ' circles constitute what is generally known as a '*non-intersecting system of coaxial* circles*'. As the properties of such systems are adequately described in a number of standard textbooks on plane geometry[12] only the more important relationships will be considered here.

In Fig. 8 are shown two member circles Γ_1' and Γ_2', centres X_1 and X_2, of a family. In addition to the Iconocentre (otherwise known as the *first limiting point*) the system also possesses:

(1) a common *centre of inversion* R, such that

$$RI^2 = RE_1 . RF_1 = RE_2 . RF_2 = \text{etc.} \quad \quad (8)$$

(2) a *radical axis* RS, which is the line through R perpendicular to the common line of centres XI.

(3) a *second limiting point* L such that LR $=$ RI.

(4) a *circle of inversion;* i.e., the circle centre R passing through the points I and L.

From equation (8), which in fact specifies the system completely, a number of useful geometrical relationships may be derived. For example:

Theorem I. The tangents to the two circles from a common point on the radical axis are equal in length. Thus

$$RT_1 = RT_2, \text{ and similarly } SP_1' = SP_2'.$$

Theorem II. From the above it follows that the locus of a point P_1', for a given point S, as the circle Γ_1' is varied, must be a circle orthogonal to Γ_1' passing through I. Hence the radical axis RS is also the locus of the centres of a system of circles each orthogonal to all the member circles of the Γ' family of circles.

Theorem III. Since $SP_1' = SP_2'$, it follows that $P_1'M = P_2'M$.

* Sometimes the term 'coaxal' is used in standard textbooks on plane geometry. However, the implication is the same, namely that all the circles have the same *radical axis* (but *not* the same centres).

Theorem IV. If a perpendicular is drawn through I so as to intersect the circles Γ_1' and Γ_2' at H_1 and H_2, then the tangents at H_1 and H_2 will intersect at the second limiting point L.

The above theorems are simple to prove and may all be derived from equation (8). The reader who wishes to gain familiarity with the method is recommended to attempt the proof of these theorems himself. As a further exercise it will be found helpful to run through the simple problems given in the Appendix, based on the application of the above theorems.

3.4. *Geometrical Identification of the Junction Parameters*

The bilinear transformation of Γ into Γ' has been defined in Section 3.2 by means of equation (7). This equation may conveniently be rewritten

$$\Gamma' = \frac{A\,\Gamma + B}{C\,\Gamma + 1} \qquad \ldots \quad \ldots \quad \ldots \qquad (9)$$

thus eliminating one superfluous constant.

Before attempting a geometrical interpretation of the complex constants A, B and C, it is convenient to relate these constants to some recognized system of parameters defining the properties of a junction.

One such system of parameters would be the input, output, and transfer impedances of the junction[6]. Since equation (9) is expressed as a function of the reflection coefficient rather than of voltages or currents this is not very convenient.

A more useful system of parameters is the set of scattering coefficients.[1] These are defined as follows (see Fig. 6).

$$v' = S_{11}\,u' + S_{12}\,v \qquad \ldots \quad \ldots \qquad (10)$$

$$u = S_{22}\,v + S_{21}\,u' \qquad \ldots \quad \ldots \qquad (11)$$

where by the principles of reciprocity of passive networks $S_{12} = S_{21}$.

In order to attach a physical significance to the scattering coefficients, let the line to the right of the junction be terminated in a matched load, in which case:

$$v = 0,\ S_{11} = v'/u' = \Gamma', \text{ and}$$
$$S_{12} = S_{21} = u/u'$$

Furthermore if we interchange the positions of the generator and the matched load:

$$u' = 0, S_{22} = u/v = 1/\Gamma, \text{ and}$$
$$S_{12} = S_{21} = v'/v$$

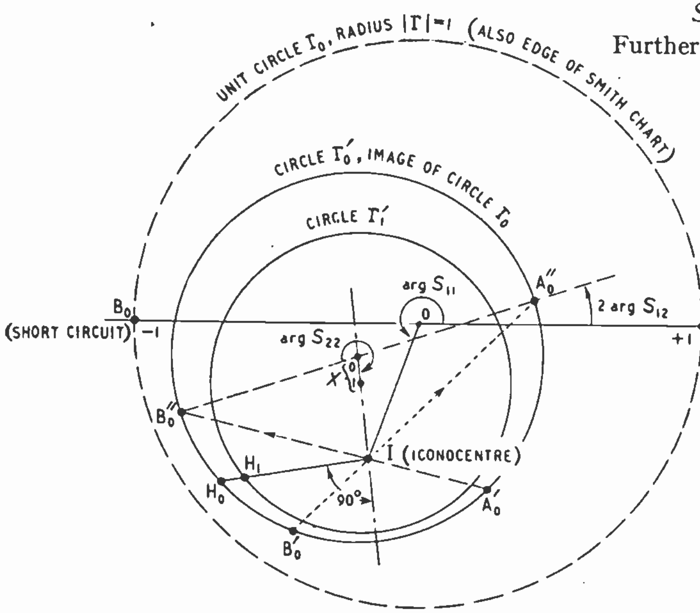


Fig. 9. *Graphical determination of the scattering coefficients and insertion loss of the junction.* 0 *is the centre of the Smith Chart; X_0 and X_1 are the centres of circles Γ_0' and Γ_1'; r_0 and r_1 are the radii of circles Γ_0' and Γ_1'; A_0' and B_0' are the images of A_0 and B_0. The reference line A_0'' B_0'' may be constructed from a knowledge of either A_0' or B_0' whichever is more convenient. The scattering coefficients are given by:*
$|S_{11}| = OI$; $\arg S_{11} = (A_0O, OI)$;
$|S_{12}| = IH_0/\sqrt{r_0}$; $\arg S_{12} = \frac{1}{2}(A_0O, X_0A_0'')$;
$|S_{22}| = IX_0/r_0$; $\arg S_{22} = (B_0''X_0, X_0I)$.
Junction insertion loss = $20 \log_{10} IH_0 - 10 \log_{10} r_0$ db.
Loss in line length $(l_1 - l_0)$ = $20 \log_{10} IH_0/IH_1 - 10 \log_{10} r_0/r_1$ db.

(Note that the inversion of Γ in the expression for S_{22} is quite consistent, since v is now the incident wave, and u is the reflected wave.)

The physical significance of the scattering coefficients can be further clarified by deriving expressions for the insertion loss of the junction. This may be defined as the ratio of the power supplied by the generator to that delivered by the junction to the line plus load expressed in decibels. In other words:

Insertion loss, $L = 10 \log_{10} (\,|u'|^2/|u|^2)$ db

Using equations (10) and (11) this can be reduced to:

$$L = -\,20 \log_{10} |\,S_{12}/(1 - \Gamma S_{22})\,| \text{ db} \qquad (12a)$$

from which it is evident that in the general case the loss is a function of the reflection coefficient of the circuit connected at the output terminals of the junction.

In practice, however, the insertion loss is frequently defined on the implicit assumption that the junction is inserted in a line terminated with a matched load. In this case $\Gamma = 0$, and equation (12a) reduces to:

$$L = -\,20 \log_{10} |S_{12}| \text{ db} \qquad \ldots \quad \ldots \qquad (12b)$$

This may be further modified to *exclude* the loss by reflection thus:

$$L = -10 \log_{10}\{(|v'|^2 + |u|^2)/|u'|^2\} \text{ db}$$
$$= -10 \log_{10}(|S_{11}|^2 + |S_{12}|^2) \text{ db} \quad (12c)$$

(The positive sign in this last result will not seem so surprising if it is remembered that the coefficients S_{11}, S_{12}, S_{22} are always less than unity.)

Of these three equations, equation (12b) is not only the simplest, but also by far the most useful. It has the additional merit of further clarifying the role of the coefficient S_{12}.

From equations (10) and (11) and the definitions $\Gamma' = v'/u'$ and $\Gamma = v/u$ it follows directly that:—

$$\Gamma' = \frac{S_{11} + (S_{12}^2 - S_{11}S_{22})\Gamma}{1 - S_{22}\Gamma} \quad .. \quad (13)$$

It will be recognized that the form of this equation is the same as that of equation (9). Not only does this justify the previous assumption of a bilinear transformation but it provides us with the desired expressions for the complex constants A, B and C.

The correct identification of the quantities S_{11}, S_{12}, and S_{22} with the geometrical properties of the circle diagram can be achieved by the application of the principles of simple co-ordinate geometry. The detailed derivation of the relationships is rather tedious however, and since it has been given in full elsewhere[8,11] it will not be repeated here. On the other hand the final result, illustrated in Fig. 9, is elegant in its simplicity.

3.5. *Summarized Routine*

In practice, therefore, the technique involves the following steps:—

(1) Make a series of measurements with two lengths of line resulting in two circles Γ_0' and Γ_1'. The first length of line corresponding to the circle Γ_0' should therefore be chosen so that the loss in the line is negligible. (If this is not possible the desired circle can be constructed relatively simply from two other circles. Space precludes a description of this construction.)

(2) From either of the two circles calculate the wavelength in the line 'bc'.

(3) Determine the position of the Iconocentre by means of a measurement with a matched load, or alternatively by means of the construction outlined in Problem II of the Appendix, if no matched load is available.

(4) From a knowledge of the wavelength in the line locate the point A_0' (or alternatively B_0') in Fig. 9, on the circle Γ_1'. Draw in the line $A_0'B_0''$ through I; project B_0'' through X_0 to A_0''; join X_0 to I and I to O. Project I to H_0 at right angles to X_0I. Measure the distances OI, IH_0, X_0I and the radius r_0 of the circle Γ_0' and hence obtain the values of $|S_{11}|$, $|S_{12}|$ and $|S_{22}|$ as indicated in the table of Fig. 9. By measuring the appropriate angles obtain the values of arg S_{11}, arg S_{12} and arg S_{22}.

(5) If the circle Γ_1' intercepts the line IH_0 at H_1, then the value of $|S_{12}|$ for the junction *plus the line* is given by

$$IH_1/(\text{radius of circle } \Gamma_1')^{\frac{1}{2}}$$

Hence, the insertion loss of junction

$$= 20 \log_{10} |S_{12}|_0 \text{ db} \quad .. \quad .. \quad (14)$$

and line loss for length $(l_1 - l_0)$

$$= 20 \log_{10} |S_{12}|_0 - 20 \log_{10} |S_{12}|_1 \text{ db} \quad .. \quad .. \quad .. \quad (15)$$

Where the additional subscripts $_0$ and $_1$ indicate the appropriate line lengths, l_0 and l_1.

At first sight the method of computation may seem a little confusing and somewhat involved. With a little practice, however, the plotting of the circles, the geometrical construction, and the final extraction of the required quantities can be accomplished very rapidly with the expenditure of very little mental effort.

4. Conclusions

It has been shown that the input–output relationship of a two-terminal-pair junction may be represented by a bilinear transformation. Furthermore, if a suitable measuring technique is used, such as that indicated, it is possible to exploit the geometrical properties of a bilinear transformation, and thereby analyse geometrically both the properties of the junction and also of the transmission line connected to the output terminals of the junction.

The method, however, can be extended to cover situations of considerably greater complexity than those discussed here. For example, the method may be used to measure the properties of one junction through another junction, or alternatively to compute the combined effect of a number of junctions in cascade[11,14]. It may also be adapted to the case where there is a change in frequency, provided the frequency convertor is reasonably linear in operation[4].

This article, however, is only an introduction to the subject. A number of other and more elegant geometrical constructions are possible[8,11] and in certain cases constructional work can be eliminated by the use of a special 'hyperbolic protractor'[9]. G. A. Deschamps has also suggested another form of transmission-line chart, known as the Projective Chart[9], which in many cases has considerable advantages over the Smith Chart.

Acknowledgments

The author wishes to thank Standard Telecommunication Laboratories for permission to publish this article. Special acknowledgment is also due to G. A. Deschamps and M. Arditi, both of Federal Telecommunication Laboratories, N.J., for their instruction in the technique, and helpful discussion of difficult points.

APPENDIX

Some Elementary Problems

The following problems are based on the four theorems of Section 3.3 and the geometrical definition of the junction scattering coefficients in Section 3.4. The relevant figures are Figs. 7(a) and (b), 8 and 9.

Problem I. Given the two circles Γ_1' and Γ_2' find the Radical Axis. (*Answer:* draw the line of centres X_1X_2. Draw any radius X_1P_1'. Draw a second radius X_2P_2' such that the two radii intersect at a point M, and such that $P_1'M = P_2'M$. Draw the tangents at P_1' and P_2', intersecting at S. The perpendicular SR to the line of centres is the Radical Axis. *Alternative:* describe a circle of any radius and any centre. Let this circle intersect circle Γ_1' at J_1 and K_1, and circle Γ_2' at J_2 and K_2. The chords J_1K_1 and J_2K_2 intersect at a point on the Radical Axis.)

Problem II. Given the circles Γ_1' and Γ_2' find the Iconocentre. (*Answer:* proceed as in Problem I and hence obtain the point R. Draw tangents RT_1 and RT_2 to the circles Γ_1' and Γ_2'. Then the circle centre R through T_1 and T_2 will intersect X_1X_2 at I. *Alternative:* having found S, simply describe the circle centre S through points P_1' and P_2' intersecting X_1X_2 at I.)

Problem III. Given the circles Γ_1' and Γ_2' find the circle orthogonal to both Γ_1' and Γ_2' passing through a given point P_1' on Γ_1'. (*Answer:* proceed as in Problem I and hence obtain the point S. The circle, centre S, passing through P_1' is the desired circle.)

Problem IV. Given the circle Γ_1' and the Iconocentre I, and an isolated point P_2' find the circle Γ_2' through the point P_2'. (*Answer:* draw in the line IH_1 perpendicular to the line of centres X_1I. Draw the tangent at H_1 intersecting X_1I at L. Bisect LI and hence obtain the radical axis through R. Find a point S on the radical axis such that S is equidistant from both I and P_2'. Draw in SP_2'. The line perpendicular to $P_2'S$ through P_2' will intersect the line X_1I at X_2 and X_2 is the centre of the required circle through P_2'.)

Other and somewhat more elegant constructions may be used instead of the above, and some of these are described elsewhere[7,8,10]. The constructions indicated in the answers to Problems I to IV have, however, the merit that they follow in a very direct manner from the fundamental relationship of equation (8).

Three problems with a rather more practical flavour will now be given. Suppose that we have made a series of measurements on a line through a mismatched junction, by the method outlined in Section 3.1. Furthermore we have obtained as a result a diagram identical to Fig. 7 except that we have not troubled to delineate the circle Γ_3'.

Problem V. Given the circles Γ_1' and Γ_2' in Fig. 7(b) it is required to determine the value of Γ' that will correspond to a matched load termination of the line. (*Answer:* if the line is terminated by a matched load then the corresponding value of Γ' will be given by the point I. Proceed, therefore, as in Problem II and hence find the point I.)

Problem VI. Given the circles Γ_1' and Γ_2' in Fig. 7(b), find the electrical angle between the points P_1' and Q_1'. (*Answer:* proceed as in Problem III and hence obtain the circular arcs $P_1'I$ and $Q_1'I$. Measure the angle between the tangents at I to these two arcs. The electrical angle is one half the geometrical angle ϕ.)

Problem VII. Given the circles Γ_0' and Γ_1' in Fig. 7(b), together with the position of the point A_0' (image of the point A_0), determine the scattering coefficients of the junction, the insertion loss of the junction and the loss per unit length in the line 'bc'. (*Answer:* proceed as in Problem II, utilizing the circles Γ_0' and Γ_1', and hence find the Iconocentre I. Carry out the construction illustrated in Fig. 9 based on the circle Γ_0'. Obtain thereby the values of the scattering coefficients of the junction as given by the table in Fig. 9, and also the insertion loss of the junction, and the loss in the line 'bc'. If this construction is performed on the circle diagram of Fig. 7(b) it will be found that the numerical results are:—

$$|S_{11}| = 0{\cdot}31 \qquad \text{Arg } S_{11} = +234°$$
$$|S_{12}| = 0{\cdot}87 \qquad \text{Arg } S_{12} = +130°$$
$$|S_{22}| = 0{\cdot}27 \qquad \text{Arg } S_{22} = +217°$$

Insertion loss of junction $= 1{\cdot}2$ db

Loss in length of line $(l_1 - l_0) = 1{\cdot}6$ db.

REFERENCES

[1] C. G. Montgomery, R. H. Dicke, E. M. Purcell: "Principles of Microwave Circuits", M.I.T. Radiation Laboratory Series, Vol. 8, McGraw-Hill, 1948

[2] E. G. Phillips, "Functions of a Complex Variable—with Applications", University Mathematical Texts, Oliver and Boyd, 1947.

[3] H. A. Wheeler, "Geometrical Relations in Circle Diagrams of Transmission-Line Impedance", Wheeler Monograph No. 4, July 1948, Wheeler Laboratories Inc., Great Neck, N.Y.

[4] H. A. Wheeler, D. Dettinger, "Measuring the Efficiency of a Superheterodyne Convertor by the Input Impedance Circle Diagram", Monograph No. 9, March 1949, *ibid.*

[5] W. A. Alter, "Q Circles—a Means of Analysis of Resonant Microwave Systems", *Proc. Inst. Radio Engrs*, April 1947 and May 1947, Vol. 35, Nos. 4, 5, pp. 355–361, 478–484.

[6] S. Fedida, "Impedance Transformation by Four-Terminal Networks", *Wireless Engineer*, Aug.–Sept. 1954, Vol. 31, Nos. 8, 9, pp. 204–214, pp. 239–249.

[7] G. A. Deschamps, "Geometric Viewpoints in the Representation of Waveguides and Waveguide Junctions", Proc. of Symposium on Modern Network Synthesis, New York, pp. 277–295, 30th September 1952.

[8] G. A. Deschamps, "Determination of Reflection Coefficients and Insertion Loss of a Waveguide Junction", *Journal Applied Physics*, August 1953, Vol. 24, No. 8, pp. 1046–1050.

[9] G. A. Deschamps, "New Chart for the Solution of Transmission-Line and Polarization Problems", *Electrical Communication*, September 1953, Vol. 30, No. 3, pp. 247–254. Also *Trans. Inst. Radio Engrs*, Prof. Group on Microwave Theory and Techniques, March 1953, Vol. 1, pp. 5–13.

[10] G. A. Deschamps. "Geometrical Representation of the Polarisation of a Plane Electromagnetic Wave", *Proc. Inst. Radio Engrs*, May 1951, Vol. 39, pp. 540–544.

[11] G. E. Storer, L. S. Sheingold, "A Simple Graphical Analysis of a Two-Port Waveguide Junction". *Proc. Inst. Radio Engrs*, August 1953, Vol. 41, No. 8, pp. 1004–1013.

[12] L. P. Benny, "Plane Geometry", Blackie & Son, 1922.

[13] D. Hilbert, S. Cohn-Vossen, "Geometry and The Imagination", Chelsea Publishing Co., N.Y., 1952 (Translation of "Anschauliche Geometrie"), pp. 242–268."

[14] M. Arditi, "Experimental Determination of the Properties of Microstrip Components", Convention Record of the I.R.E., 1953 National Convention, N.Y., Part 10, Microwaves.

[15] J. M. C. Dukes, "Experimental Determination of the Properties of an Unknown Termination when measured through a Mismatched Junction", *Wireless Engineer* (to be published later).

DIFFERENTIAL-AMPLIFIER DESIGN

By A. M. Andrew, B.Sc.

(*Departments of Physiology and Psychological Medicine, University of Glasgow*)

SUMMARY.—The theoretical possibility exists of a differential amplifier which will give good rejection of in-phase input voltages without requiring a balancing adjustment or the use of selected valves or other components. The practical design of such an amplifier is difficult. The result can be achieved by employing in-phase negative feedback, but may not be achieved if the feedback voltages affect the amplifier in a way which is not exactly equivalent to summation with the input voltages. When in-phase negative feedback is introduced by including a resistance or a pentode valve in series with the cathodes of a push-pull stage, the feedback voltage affects the amplifier in a way which is not equivalent to summation with the input voltages, and good in-phase rejection without a balancing adjustment is not possible by this means. Two types of amplifier stage are described which make possible good in-phase rejection without a balancing adjustment. Because of practical disadvantages these are unlikely to come into general use, but they illustrate important general principles.

DIFFERENTIAL amplifiers are frequently required for biological and other purposes. A true differential amplifier will give zero output when its two input terminals undergo identical voltage variations; that is to say, when it receives an in-phase input. Practical amplifiers will generally give some output under these conditions, and it is convenient to speak of the 'rejection ratio' or 'discrimination factor' of the amplifier, which may be defined as the ratio between the in-phase input voltage and the anti-phase input voltage required to give the same output. [See opening section of reference (2)]. It is assumed that the output is a single variable, as it is when the amplifier gives a single output voltage, or when it drives a pen recorder or an oscillograph. The general push-pull amplifier, giving a double output, is considered later.

In some applications, one of the most important of which is electro-encephalography, the amplifier is required to respond to a small anti-phase signal which may occur together with a much larger in-phase interfering voltage. It is then important to have a high rejection ratio, and in electro-encephalography a value of at least 10,000 should be aimed at.

Need for Balancing Control

Amplifiers which are required to have a high rejection ratio generally incorporate some form of balancing control which can theoretically be adjusted to make the in-phase rejection perfect. An exception is the amplifier described by Johnson[1], but Parnum[2] has shown that Johnson's amplifier cannot be relied upon to give good in-phase rejection when built with valves and components of commercial tolerance.

From Parnum's results it might be supposed that it is physically impossible to build an amplifier which can be relied upon to have a high rejection ratio without using selected components and without a balancing adjustment. On the other hand, Offner[3] has shown that, in theory at least, an amplifier can be constructed, using components of commercial tolerance, to have a high rejection ratio without the need for a balancing control.

An amplifier with an intrinsically high rejection ratio, not depending on critical selection or adjustment of components, would be very useful, for it would not require readjustment when valves were replaced, nor would it require periodical adjustment to allow for aging of valves and other components. However, although such an amplifier is theoretically possible, it is not easy to devise a satisfactory practical circuit.

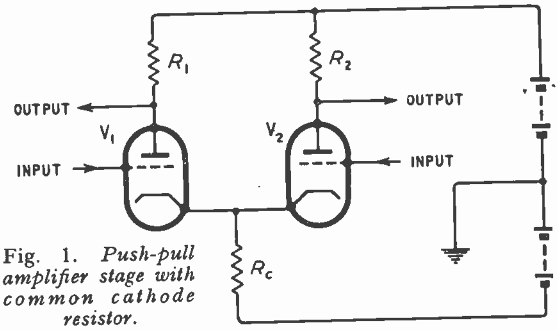


Fig. 1. *Push-pull amplifier stage with common cathode resistor.*

Parnum considers amplifiers whose stages are of the form shown in Fig. 1, in which R_c may be either a resistor as shown, or a pentode valve having extremely high differential resistance. He shows that even when, effectively, $R_c = \infty$, a balancing control is needed if high rejection ratio is to be ensured with components of commercial tolerance. Offner, however, indicates that his results can be applied to the circuit of Fig. 1, and, if his results are applied as he suggests, a conclusion is reached which is at variance with Parnum's conclusions, for it appears that when $R_c = \infty$, the circuit should give perfect in-phase rejection even though $R_1 \neq R_2$ and the valves are dissimilar.

The discrepancy between the findings of

MS accepted by the Editor, February 1954

Parnum and those of Offner will be removed by showing that Offner's general conclusions cannot strictly be applied to the circuit of Fig. 1. Consideration of the reason for the apparent discrepancy has suggested two ways in which the circuit of Fig. 1 can be modified to make Offner's results approximately applicable, and these will be described. For either of the modified circuits the in-phase rejection is very much less dependent on valve characteristics and component values than for the circuit of Fig. 1.

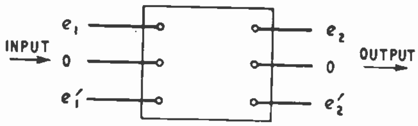


Fig. 2. *General push-pull amplifier.*

General Push-Pull Amplifier

Offner defines two sets of four gain factors for a push-pull amplifier or amplifier stage. The specification of values for either set gives, if linearity is assumed, a complete description of the behaviour of the amplifier. In general, the gain factors will be complex numbers.

Let e_1 and e_1' be the input voltages, and e_2 and e_2' the output voltages of a push-pull amplifier (Fig. 2). Then the first set of gain factors consists of m, m', γ and γ', defined as follows:—

$$m = e_2/e_1 \quad \text{for} \quad e_1' = 0$$
$$m' = e_2'/e_1' \quad \text{for} \quad e_1 = 0$$
$$\gamma = -e_2'/e_1 \quad \text{for} \quad e_1' = 0$$
$$\gamma' = -e_2/e_1' \quad \text{for} \quad e_1 = 0$$

Of the other set, only three are required for the present purpose. These are the anti-phase, in-phase and inversion gains, defined as follows:—

Differential or anti-phase gain (for $e_1' = -e_1$)

$$= (e_2 - e_2')/(e_1 - e_1')$$
$$= G_0 \text{ (Offner's notation)}$$
$$= M \text{ (Parnum's notation)}$$

In-phase gain for ($e_1' = e_1$)

$$= (e_2 + e_2')/(e_1 + e_1')$$
$$= G_c \text{ (Offner's notation)}$$
$$= \overline{M} \text{ (Parnum's notation)}$$

Inversion gain for ($e_1' = e_1$)

$$= (e_2 - e_2')/\tfrac{1}{2}(e_1 + e_1')$$
$$= \frac{\text{anti-phase output}}{\text{in-phase input}} \left(\begin{array}{c}\text{for zero anti-}\\ \text{phase input}\end{array}\right)$$
$$= G_i \text{ (Offner's notation)}$$
$$= K \text{ (Parnum's notation)}$$

$$K = G_i = \frac{\mu_1 R_1 r_{a2} - \mu_2 R_2 r_{a1} + (\mu_1 - \mu_2)\{R_1 R_2 + (R_1 + R_2) R_c\}}{(r_a + R)\{r_a + R + 2R_c(\mu + 1)\}} \quad \ldots \quad \ldots \quad (2)$$

Parnum defines 'transmission factor' H as follows:—

$$H = M/K = G_0/G_i$$

Modern biological amplifiers are generally push-pull throughout, and the indicating device generally responds only to the anti-phase component of the output voltage. In that case the rejection ratio of the amplifier is equal to its transmission factor H.

Amplifiers which do not include feedback loops involving more than one stage will first be considered. In such amplifiers, when fitted with balancing controls, high overall transmission factor is often achieved by intentionally unbalancing the input stage to compensate for lack of balance in later stages. If this is done, high overall transmission factor may be achieved although the individual stages have low transmission factors.

In an amplifier which does not include a balancing control, however, it is not possible to compensate for imperfect balance in one stage by unbalancing another. To ensure high overall transmission factor the first stage must be designed to have a high transmission factor and, unless the first stage has negligible in-phase gain, the second and perhaps later stages must have high transmission factors also. An amplifier without a balancing control and which can be relied upon to have a high overall transmission factor (without feedback loops involving more than one stage) can be built if, and only if, an amplifier stage having high transmission factor can be built without a balancing control.

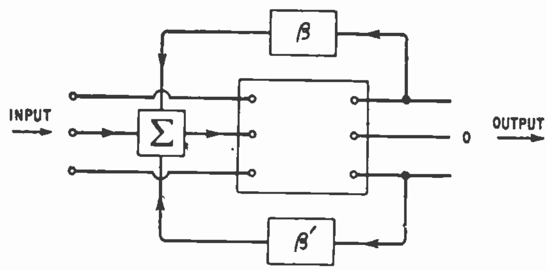


Fig. 3. *General push-pull amplifier with in-phase negative feedback.*

Stage with Common Cathode Resistor

For the push-pull stage of Fig. 1, Parnum obtains the following expressions for the anti-phase and inversion gains:—

$$M = G_0 = \mu R/(r_a + R) \quad \ldots \quad \ldots \quad (1)$$

and for the inversion gain

where μ_1 and r_{a1} are the amplification factor and anode resistance of V_1; μ_2 and r_{a2} are constants of V_2. The expressions have been simplified by replacing R_1 and R_2 by their mean value R, μ_1 and μ_2 by their mean value μ, and r_{a1} and r_{a2} by their mean value r_a in places where the values are not critical.

Hence the transmission factor is given by

$$H = \frac{\mu R\{r_a + R + 2R_c(\mu + 1)\}}{\mu_1 R_1 r_{a2} - \mu_2 R_2 r_{a1} + (\mu_1 - \mu_2)\{R_1 R_2 + (R_1 + R_2) R_c\}} \quad \ldots \quad (3)$$

and if $R_c = \infty$,

$$H = \mu(\mu + 1)/\Delta\mu \approx \mu^2/\Delta\mu \quad \ldots \quad (4)$$

where $\Delta\mu = \mu_1 - \mu_2$

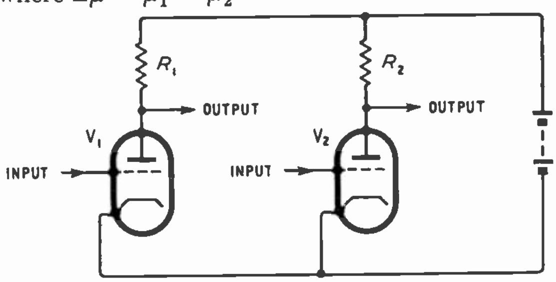


Fig. 4. *Push-pull amplifier stage, as in Fig. 1, but without common cathode resistor.*

If a pentode valve is substituted for R_c, its differential resistance is very high, and the value of H is approximately $\mu^2/\Delta\mu$. However, as pointed out by Parnum, this value will often not be as high as is required and a balancing control must be included to compensate for the difference between μ_1 and μ_2.

Offner has shown that if negative feedback is applied to an amplifier as shown in Fig. 3, the expression for the inversion gain with feedback is

$$G_i' = G_i/[1 + \beta(m - \gamma') + \beta'(m' - \gamma)] \quad (5)$$

where G_i, m, m', γ and γ' are for the amplifier without feedback.

The common cathode resistor R_c in Fig. 1 introduces in-phase negative feedback. Without R_c the circuit becomes that of Fig. 4, which consists of two unconnected single-sided amplifier stages. The gain factors m and m' are the respective gains of these stages and $\gamma = \gamma' = 0$.

If now the resistor R_c is inserted, as in Fig. 1, the feedback fractions are given by

$$\beta = R_c/R_1$$
$$\beta' = R_c/R_2$$

Substituting for β, β', γ and γ' in equation (5) gives for the inversion gain

$$G_i' = G_i/[1 + mR_c/R_1 + m'R_c/R_2] \quad \ldots \quad (6)$$

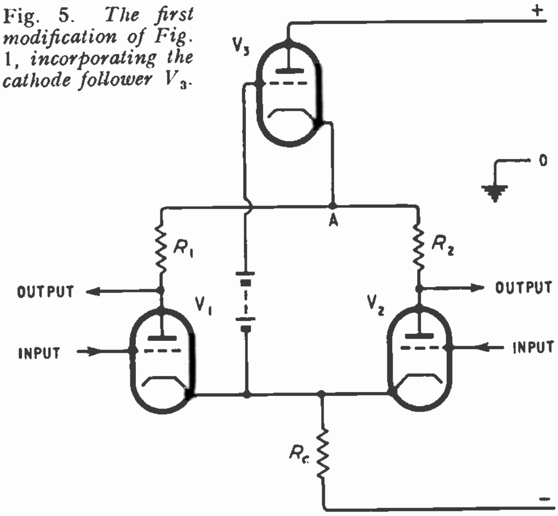
According to equation (6), when $R_c = \infty$, $G_i' = 0$ and hence the transmission factor $H = \infty$, even though $\mu_1 \neq \mu_2$. This result differs from that of Parnum [equation (4)].

The explanation of the discrepancy is that Offner's results are not applicable to the circuit of Fig. 1. The variations in cathode potential due to the inclusion of R_c produce a change in the grid-to-cathode voltage of each valve and also a change in the anode-to-cathode voltage. When R_c is regarded as introducing in-phase negative feedback, only its effect on the grid-to-cathode voltage is being taken into consideration. The application of Offner's theory therefore produces an incorrect result.

First Modified Circuit

Consideration of this explanation of the discrepancy suggests the circuit of Fig. 5 as a circuit to which Offner's results are applicable in the ideal case where the cathode follower V_3 has unity gain ($M = 1$) and the output voltages are taken relative to point A instead of relative to earth. It can be verified by straightforward circuit analysis that the inversion gain is zero when $M = 1$ and $R_c = \infty$. If an in-phase voltage e is applied to the grids, the resulting change in the anode current of V_1, for $M = 1$, is given by

Fig. 5. *The first modification of Fig. 1, incorporating the cathode follower V_3.*



$$i_{a1} = \frac{e\mu_1(r_{a2} + R_2)}{(r_{a1} + R_1)(r_{a2} + R_2) + R_c\{\mu_1(r_{a2} + R_2) + \mu_2(r_{a1} + R_1)\}} \quad \ldots \quad (7)$$

$= 0$ when $R_c = \infty$.

By symmetry, $i_{a2} = 0$ when $R_c = \infty$.

Hence if the output voltages are measured relative to point A, there is no output for an in-phase input. Hence when $M = 1$ and $R_c = \infty$ the transmission factor is infinite.

The values of anti-phase gain, inversion gain and transmission factor are unaffected if the output voltages are taken relative to earth instead of relative to the point A, for a change in reference point does not affect the anti-phase component of output voltage.

In the general case where the gain of the cathode follower V_3 is M, the transmission factor is given by

$$H = \frac{\mu R \{r_a + R + 2R_c(\mu + 1 - M)\}}{\mu_1 R_1 r_{a_2} - \mu_2 R_2 r_{a_1} + (\mu_1 - \mu_2)[R_1 R_2 + R_c(1 - M)(R_1 + R_2)]} \quad \ldots \quad (8)$$

$\approx \mu^2/\Delta\mu(1 - M)$ when $R_c = \infty$.

By comparison with equation (4) it is seen that the effect of the cathode follower is to multiply the transmission factor by $1/(1 - M)$ in the case where $R_c = \infty$. In practice M may be very close to unity, especially when R_c is effectively infinite, for then the cathode load of V_3 includes a constant-current device. Hence a considerable improvement in transmission factor is possible by the use of the circuit of Fig. 5.

Second Modified Circuit

Another circuit to which Offner's results are approximately applicable is shown in Fig. 6(a). In this circuit the variations in cathode potential due to R_c affect the anode-to-cathode voltages, but have little effect on the anode currents because of the high anode resistance of the pentode valves. When R_c is replaced by a pentode valve, a high transmission factor is obtained without the need for a balancing adjustment, with valves and components of commercial tolerance.

The high transmission factor of this circuit could also have been predicted from Parnum's results, for equation (4) shows that a high transmission factor can be obtained when μ is very large even when $\Delta\mu/\mu$ is also appreciable. By using pentodes with fixed screen-grid voltage, an extremely high value of μ is obtained.

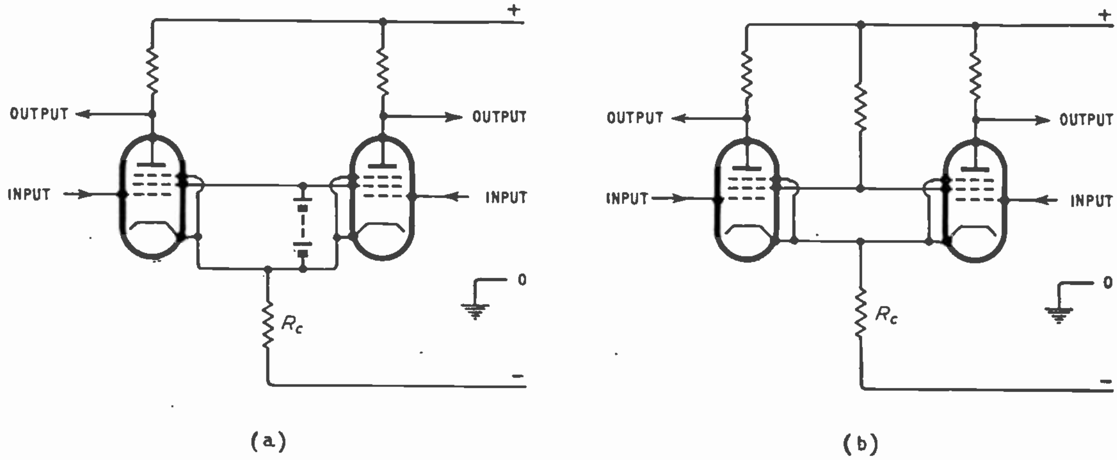


Fig. 6. (a) *The second modification of Fig. 1, incorporating pentode valves with a battery between the cathodes and the screen-grids.* (b) *Another stage incorporating pentode valves, which generally gives a lower transmission factor than circuit* (a).

The circuit of Fig. 6(b) does not in general give a transmission factor as high as that of the circuit of Fig. 6(a). Offner's results cannot be applied to the circuit of Fig. 6(b) because the variations in cathode potential affect the screen-to-cathode voltage and hence the anode currents.

Experimental Results

Tests were made with the three amplifier stages shown in Fig. 7, in which (a), (b) and (c) are practical versions of Figs. 1, 6(b) and 6(a) respectively. The stages were connected as shown in Fig. 8. The two-stage amplifier and oscilloscope by themselves had a measured rejection ratio of 30. For measurement of rejection ratios, the height of the oscilloscope trace was used as a measure of the output of the

TABLE 1

Values of rejection ratio obtained with the amplifier stages of Figs. 7(a), (b) and (c) in the arrangement shown in Fig. 8.

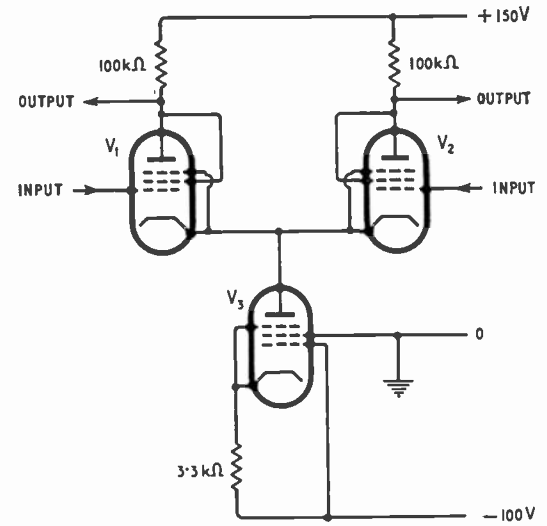
Valve pair	1	2	3	4	5
Ratio for Fig. 7(a)	630	630	220	1,500	320
Ratio for Fig. 7(b)	450	1,500	320	45,000	1,050
Ratio for Fig. 7(c)	14,000	40,000	30,000	40,000	30,000

amplifier, while known in-phase and anti-phase inputs were applied in turn. The rejection ratio is the ratio of in-phase to anti-phase inputs giving the same output. Throughout the tests the same EF37A valve was used as V_3. Five random pairs of EF37A valves were inserted in turn as V_1 and V_2. The measured rejection ratios were as shown in Table 1. The superiority of the circuit of Fig. 7(c), which is a practical version of that of Fig. 6(a), is evident.

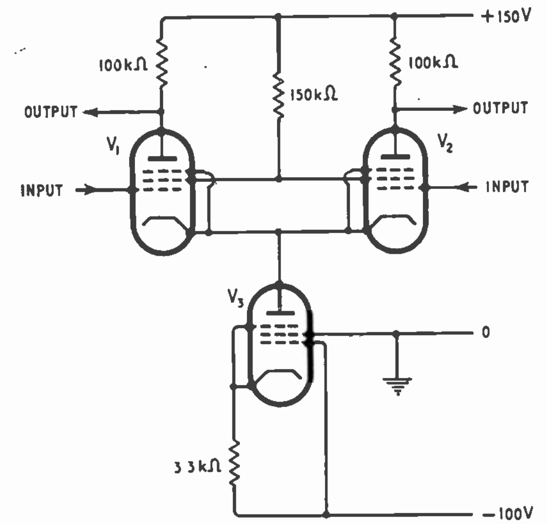
The values of rejection ratio shown in Table 1 were only obtained when an isolated, well-insulated heater supply was provided for V_1 and V_2. If this is not done, the effective value of R_c is reduced owing to the heater-cathode leakage in these valves. The need for an isolated heater supply is a serious practical difficulty in the obtaining of high rejection ratios without a balancing control, by methods which depend on an extremely high value for R_c.

Amplifiers having In-Phase Feedback over more than One Stage

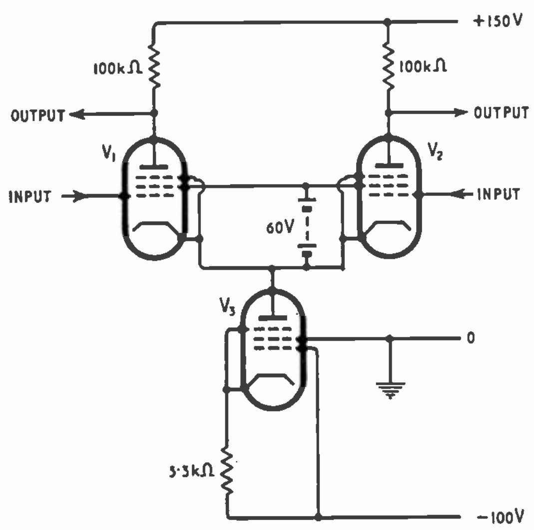
In-phase feedback can be applied over a number of stages in either of two ways. The feedback voltage may be applied either to a grid circuit or to a cathode circuit. Offner gives examples of each type. In his example in which the feedback voltage is introduced to the grids (his Fig. 7), it is necessary that either the source of signals or the power supply to the amplifier should 'float'. Neither of these requirements is likely to be met in practice, especially in multichannel equipment where it is desirable to operate several channels from the same power supplies in the interests of economy and simplicity.



(a)



(b)



(c)

Fig. 7. *Experimental amplifier stages. In each case the heaters of V_1 and V_2 were supplied from an isolated, well-insulated battery. Resistors were of 10% tolerance. In all cases the outputs were taken off via the usual coupling capacitors and grid leaks, but these components are not shown in the circuits.*

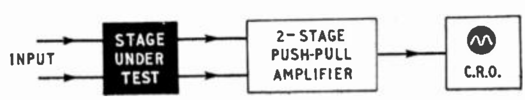


Fig. 8. *Experimental arrangement for tests on the amplifier stages of Fig. 7.*

Where the feedback voltage is applied to cathodes, Offner's theoretical treatment is not applicable, for the same reason as it was inapplicable to the circuit of Fig. 1, unless the stage to which the feedback is applied incorporates a cathode follower, as in Fig. 5, or uses pentodes and a battery as in Fig. 6(a). When either of these precautions is taken, Offner's theory becomes applicable, and a high rejection ratio can be obtained without a balancing adjustment. The cathode circuit need not be so highly insulated from earth as when the feedback depends on R_c only, so an isolated heater supply is not essential.

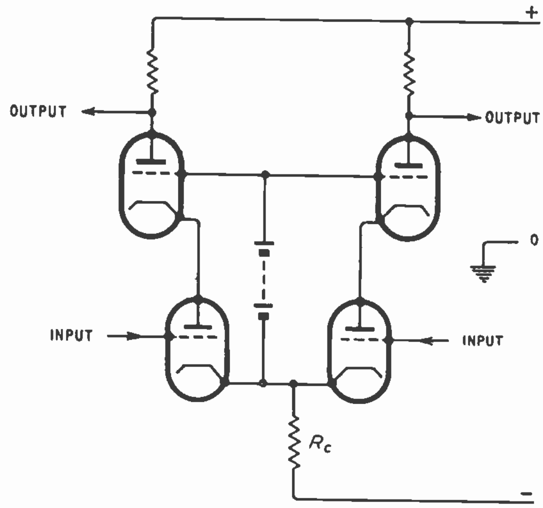
Discussion

It should not be inferred from Parnum's results that a differential amplifier which will give good in-phase rejection without a balancing adjustment when built with components of commercial tolerance is a physical impossibility. Offner's results show that such an amplifier can be built by utilizing in-phase negative feedback. However, the design of a practical amplifier having in-phase rejection good enough for use in electro-encephalography, without a balancing control, requires care.

The two types of amplifier stage described in this note provide means of achieving the result, but they are subject to practical disadvantages, since both require 'floating' batteries and, except when in-phase feedback from a later stage is applied to the cathodes, they require isolated heater supplies if full advantage is to be taken of their properties. In addition, the second modified circuit [Fig. 6(a)] has the disadvantage that it uses pentodes, and will therefore be more noisy than a stage using triodes.

It is possible to devise further modifications of the first modified circuit (Fig. 5) which do not require the floating battery. Fig. 9 shows one such modification. This circuit, however, has the disadvantage that with it the value of M achieved will not be so close to unity as with the circuit of Fig. 5.

A further possibility, which eliminates the noise due to the pentodes in the second modified



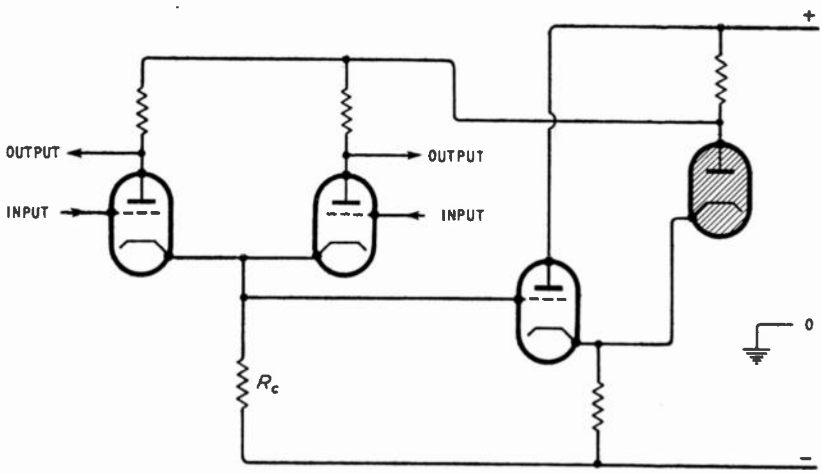
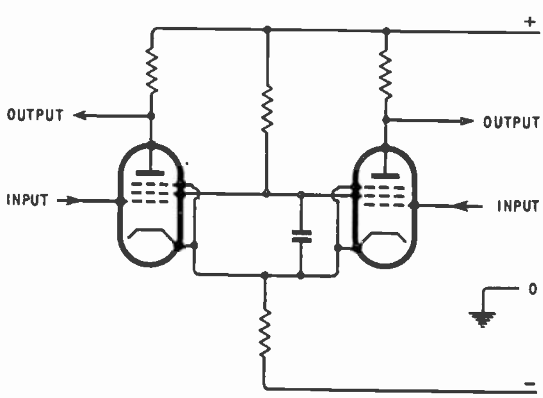


Fig. 9 (left). *Stage similar to that of Fig.* 5, *without a floating battery. The cathode-follower gain is likely to be lower than in Fig.* 5.

Fig. 10 (above). *Stage similar to that of Fig.* 6(a), *using pairs of triodes in cascode instead of pentode valves.*

Fig. 11 (below). *Stage similar to that of Fig.* 6(a), *using a capacitor instead of the floating battery. This type of stage is used by Offner.*



circuit [Fig. 6(a)] is to replace the pentodes with pairs of triodes connected so as to form a cascode amplifier, as in Fig. 10. A floating battery is still required, but now no current is drawn from it. The cascode pair resembles a pentode in having very high anode impedance.

Offner[4] has pointed out that instead of using a battery as in Fig. 6(a), the screen-grids and cathodes can be connected by a capacitor as shown in Fig. 11, and the advantages of Fig. 6(a) will be obtained at frequencies which are high enough for the capacitor to have negligible impedance. The effective value of R_c, for the purposes of the theoretical treatment, consists of the cathode resistor and the screen-grid resistor in parallel. In this way, good in-phase rejection is obtained, except at very low frequencies, without using floating batteries. The rejection may be further improved by applying in-phase negative feedback from a later stage, to the cathodes.

Because of their practical disadvantages it seems unlikely that the modified circuits will come into general use in the forms in which they have been described. Differential amplifiers will probably continue to be built either with balancing controls, or in the manner due to Offner, described in the preceding paragraph. The modified circuits are thought to be worth describing because of the general principles which they illustrate.

REFERENCES

[1] D. L. Johnson, "Electro-Encephalograph Amplifier", *Wireless Engineer*, 1947, Vol. 24, p. 231.
[2] D. H. Parnum, "Transmission Factor of Differential Amplifiers", *Wireless Engineer*, 1950, Vol. 27, p. 125.
[3] F. F. Offner, "Balanced Amplifiers", *Proc. Inst. Radio Engrs*, 1947, Vol. 35, p. 306.
[4] Franklin F. Offner, Offner Electronics Inc., Chicago. Personal communication.

H.F. DIRECTION FINDING

Observations on a Transmitter of Adjustable Beam-Direction

By C. G. McCue, M.Sc.*

(Communication from D.S.I.R. Radio Research Station, Slough)

SUMMARY.—Some night-time d.f. observations were made at the Radio Research Organization's stations at Slough and Winkfield on h.f. radio signals from Sterling, Virginia, U.S.A., during February and March 1953. The purpose of these measurements was to test whether the observed bearings depended upon the direction of the transmitting aerial beam. No significant correlation of this nature was disclosed.

1. Introduction

A SERIES of night-time direction-finding observations was carried out at the Radio Research Organization's stations at Slough and Winkfield between 17th February and 6th March 1953 on 13·7 Mc/s continuous-wave and pulse radio signals from Sterling, Virginia, U.S.A. The transmitting system was a Yagi aerial of adjustable orientation, with a beam-width in the horizontal plane of $\pm 34°$ between 6-db points. In the experiments to be described the maximum deviation of the beam axis from the bearing line to Slough was 30 degrees. The transmissions provided an opportunity to investigate whether the bearing observations were affected by the orientation of the transmitting aerial.

The continuous-wave signals were sent with a power of 400 watts according to the programme outlined in Table 1. This programme was repeated every 20 minutes from 1800 hours U.T. till 2300 hours U.T. Bearings were taken at Slough on a U-Adcock cathode-ray direction finder on 17th, 24th, 25th and 27th February 1953. Visual observations were made as rapidly as possible using only amplitude maxima: this gave an average of 11 readings in each 2·5-minute period.

* Scientific Officer of the Australian Department of Supply, until recently working at the Radio Research Station, Slough.

MS accepted by the Editor, March 1954

TABLE 1

Transmitting Aerial Orientation (E of N)	Period of Operation (U.T.)
30°	1800 – 1802½
40°	1802½ – 1805
*50°	1805 – 1807½
60°	1807½ – 1810
70°	1810 – 1812½
80°	1812½ – 1815
Transmitter off	1815 – 1820

* Pointing almost directly towards Slough and Winkfield.

The pulse signals were sent with a peak power of 300 kW. The pulse duration was 40 μs and the pulse repetition rate 25 per second. The transmitting programme was the same as that for the c.w. signals. These pulse observations were taken visually, as for the c.w., on the U-Adcock apparatus at Slough on 3rd–6th March 1953. An average of 10 readings was obtained each 2·5-minute period. Additional observations were made photographically on 6th March using a long-base loop-aerial phase comparison d.f. equipment at Winkfield about 10 kilometres south-west of Slough. This also yielded an average of 10 readings each 2·5-minute period. During these pulse experiments, one particular echo, tentatively identified as the third-order F-reflection, was most frequently observed. A few measurements were made also on the second- and fourth-order reflections.

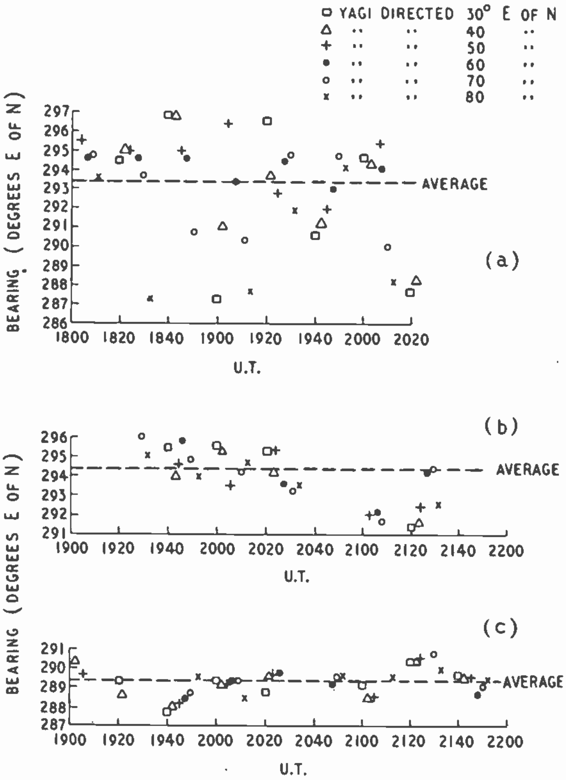


Fig. 1. *Mean bearings for* 2·5-*minute periods:* (a) *on* 25*th February* 1953 *at Slough. Transmitter Sterling bearing* 288·3°, 13·7 *Mc/s, CW;* (b) *Slough and* (c) *Winkfield, on* 6*th March* 1953. *Transmitter Sterling bearing* 288·3°, 13·7 *Mc/s, pulses.*

2. Results

For the purpose of analysing the results of these observations, the d.f. readings were grouped in the same two-and-a-half-minute periods in which they had been taken. The mean values and the variance of the bearings were calculated for each of these periods. Examples of these means are shown in Fig. 1 for (a) Slough observations on continuous waves on 25th February 1953, (b) Slough on pulses on 6th March 1953, and (c) Winkfield on pulses on 6th March 1953.

The variance computed by grouping all individual readings made on a given orientation of the transmitting aerial into one large sampling distribution can be analysed into the two components, the 'between-period' variance and the 'within-period' variance. In all cases, the between-period variance was more than 10 times the within-period variance which, for the number of degrees of freedom involved, is highly significant of difference. This prevented the application of the *t*-test to a comparison of the means using variances computed in the described manner.

Accordingly, the method of analysis adopted was to apply the *t*-test to a comparison of the means of the results obtained for the different aerial orientations, computed using the two-and-a-half-minute period means as the population values. The results of the analysis are presented in Table 2.

TABLE 2

Slough C.R.D.F.: Continuous Waves

Transmitting Aerial Orientation (E of N)	Mean Bearing Degrees E of N	Variance of Population (Deg.)²	Number of Period Means Used
30°	292·8	10·4	11
40°	292·6	6·7	10
50°	293·0	10·9	13
60°	292·4	6·0	13
70°	291·9	7·5	12
80°	291·2	7·2	9

Slough C.R.D.F.: Pulses using Third-Order Echo

Transmitting Aerial Orientation (E of N)	Mean Bearing Degrees E of N	Variance of Population (Deg.)²	Number of Period Means Used
30°	294·5	1·57	20
40°	294·6	3·13	22
50°	294·5	3·75	24
60°	294·0	4·49	22
70°	293·9	4·20	24
80°	294·3	1·90	19

Winkfield: Pulses using Third-Order Echo

Transmitting Aerial Orientation (E of N)	Mean Bearing Degrees E of N	Variance of Population (Deg.)²	Number of Period Means Used
30°	289·2	0·67	7
40°	289·3	0·69	8
50°	289·3	0·62	7
60°	289·1	0·25	5
70°	289·5	0·68	5
80°	289·4	0·25	6

This statistical analysis for the c.w. and the third-order pulse echo revealed no significant differences among the bearings for the different transmitting aerial orientations above the 90% t-test level. A similar analysis of the smaller amount of data for the second-order echo also revealed no significant bearing differences for the various aerial orientations.

It will be noted from Table 2 that the mean bearings and their variances observed at Slough are significantly different for the c.w. and pulse transmissions respectively; the c.w. bearings are lower than the pulse bearings and the c.w. variances are larger than the pulse variances. In considering these differences it must be remembered that the pulse and c.w. transmissions were not made on the same days so that changes in ionospheric and instrumental conditions may be at least partially responsible. It is now known that there is a pronounced resonance effect in the Adcock aerial system near the operating frequency; errors in the observed bearings are, therefore, large and particularly sensitive to frequency changes, changes in angle of elevation and variations in the electrical characteristics of the aerial system. Unfortunately, no adequate measurements of instrumental error were made at the time of the tests so that it cannot be said with certainty that the observed difference between pulse and c.w. mean bearings is, in fact, instrumental in origin. It is not easy to explain the observed difference between the variances of the 2½-minute-means for pulse and c.w. conditions. The smaller variance obtained with pulses might be a result of the elimination of errors due to wave interference between the various propagation modes present in the c.w. case. But this is unlikely as wave interference errors change rapidly in periods of seconds[1] so that the contribution to the variance of the 2½-minute-means would be small compared with the variance of each constituent observation and in any case, as already noted, the latter variance was itself small compared with the period-to-period variance. Nor does it seem likely that the reduction of variance in the pulse case resulted from reduced effects of lateral deviation, since presumably all of the energy, even in the c.w. case, arrived at low angles of elevation. It is possible that a small part of the difference may result from a change in observational error brought about by a change in effective signal/interference ratio, as the pulse signals were stronger than the c.w. and, therefore, easier to observe.

3. Conclusions

This particular series of night-time d.f. observations made on radio signals from Sterling, Virginia, U.S.A., during February and March 1953, was undertaken to test whether the observed bearings depended on the direction of the transmitting aerial beam. No dependence of this nature was disclosed by the experiments but it would be unwise to generalize from the results of these limited tests.

Acknowledgments

Thanks are due to the National Bureau of Standards, U.S.A., for providing the transmissions.

The author acknowledges the assistance of his colleagues who helped in the taking of the bearing observations.

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research.

REFERENCE

[1] W. C. Bain, " On the Rapidity of Fluctuations in Continuous-Wave Bearings". *Proc. Instn elect. Engrs*, Pt. B, 1955, Vol. 102 (to be published).

PREMIUMS FOR TECHNICAL WRITING

The Radio Industry Council has announced the award of six further premiums of 25 guineas each for technical writing. They are to:—

H. S. Jewitt, for his articles "Wideband I.F. Amplifiers" and "Feedback I.F. Amplifiers for Television", which appeared in *Wireless World*, February and December 1954.

A. E. Maine, for his articles "High-Speed Magnetic Amplifiers" and "Three-Phase High-Speed Magnetic Amplifiers", which appeared in *Electronic Engineering*, May and December 1954.

J. F. Field and D. H. Towns, for their article "A Torquemeter for Testing Gas Turbine Components", which appeared in *Electronic Engineering*, November and December 1954.

W. R. Cass and R. M. Hadfield, for their article "Dip-Soldered Chassis Production", which appeared in *Wireless World*, November 1954.

The remaining two premiums have been awarded to the authors of three articles which appeared in *Electronic Engineering*, July, August and September 1954.

J. M. M. Pinkerton and E. J. Kaye, "Leo Lyons' Electronic Office",

E. H. Lenaerts, "Leo Operation and Maintenance",

E. J. Kaye and G. R. Gibbs, "Leo—a Checking Device for Punched Data Tapes".

The awards are to be made at a luncheon to be held by the Public Relations Committee of the Radio Industry Council on 10th March 1955.

I.E.E.

His Royal Highness the Duke of Edinburgh, K.G., P.C., K.T., G.B.E., F.R.S., has been elected to Honorary Membership of the Institution of Electrical Engineers.

WAVEGUIDE PHASE CHANGER

New Type with Linear Characteristic

By R. E. Collin, B.Sc., Ph.D.

(Dept. of Electrical Engineering, Imperial College of Science and Technology, London)

SUMMARY.—The properties and uses of the main types of waveguide phase changers are discussed in a qualitative way. This is followed by a theoretical discussion of a new type of rectangular waveguide phase changer with a linear characteristic. The design and performance of a working model are described and a comparison with other types of phase changers is made.

1. Introduction

WAVEGUIDE phase changers play an important role in measurements at microwave frequencies, and are also used when a variable electric length of waveguide line is required. During the last decade several phase changers operating on different principles have been developed. The merits of any particular type depend to a large degree on the purpose for which it is to be used. The various phase changers that have been used are:—

(1) Rotary type[1,2,3].
(2) Squeeze-section type[4].
(3) Transverse moving-dielectric type[5,6,7,8].
(4) Trombone type[9].

Before comparing the performance of the various types it is necessary to consider some of the desirable properties in a phase changer and the various applications. Among the desirable features are:—

(1) Ample mechanical movement to enable accurate setting and reading.
(2) A simple calibration law.
(3) A standing-wave ratio near unity over an adequate band of frequencies.
(4) Ease of construction and reasonable size.
(5) Suitability for use at high field intensities without flash over.

All the above features are not usually found in any one type. However, depending on the use to which the phase changer is to be put, this may or may not be a disadvantage. Some of the more important uses are given below:—

(1) Measurement of the phase of aerial-radiation patterns and diffraction fields.
(2) Adjustment of the load conditions for a microwave oscillator.
(3) Determination of magnetron and klystron frequency-pulling figures.
(4) Aerial scanning systems.
(5) Phase modulation of a carrier wave.

The above applications may be broadly classed into two divisions. The first is that of phase measurement where a simple calibration law and accuracy of setting and reading are of primary importance. The second division is where a line stretcher is required, but the exact phase change produced is only of secondary importance.

The squeeze-section type of phase changer is perhaps the simplest to construct but it is, at best, a rather inferior instrument for the following reasons. A very small mechanical movement is required for a phase change of 360° and hence the accuracy of setting and reading is small. The calibration curve is a complicated function of frequency and setting. The slot may radiate unless it is narrow and properly centred. The physical length is excessive; i.e., approximately 25 in. for a 360° phase change, when operating at a wavelength of 3 cm.

The rotary phase changer is the most difficult to construct but it has a simple calibration law that is independent of frequency. The change in phase of the transmitted wave is equal to twice the angle through which the half-wave plate is rotated. Because of this frequency independence this type is suitable for use in aerial scanning systems. Furthermore, rotary motion is by far the simplest to produce, and several rotary phase changers have been used ganged together in aerials[1].

The rotary phase changer is the only feasible type for providing phase modulation of a carrier wave. The rotating half-wave plate is usually spun at a suitable number of revolutions per second by means of an induction motor built into the phase changer. For this application small departures of the phase shift from the theoretical law give rise to harmonics which can be filtered out and hence are only of secondary importance. The rotary phase changer has the great advantage that continuous rotation of the half-wave plate produces a continuous change in phase of the transmitted wave and this feature makes it superior to all other types for use in automatic phase plotters.

For the determination of oscillator frequency-pulling figures it is essential that the standing-wave ratio of the phase changer be nearly unity over as wide a band as possible while the calibration law is only of minor importance. For example, a standing-wave ratio of 1·03 can

MS accepted by the Editor, April 1954

produce an error of the order of 7·5% in the pulling figure. The reflected waves from the fixed discontinuity and the phase changer may add in phase or out of phase and hence the oscillator is subjected to a standing-wave ratio varying in phase through 180°, and which may vary in magnitude from 1·45 to 1·55.

For accurate phase measurements it is desirable to have a phase changer that is capable of measuring phase to within 1° or less. The rotary phase changer does not appear to have been developed to this degree of perfection. Slight misalignment of the quarter-wave plates and multiple reflections between the various discontinuities cause the phase change to depart from its theoretical value,[10]. The bandwidth is largely limited by the dispersion of the quarter-wave plates and, for large bandwidths, it is necessary to use some form of achromatic quarter-wave plates. To minimize the effect of the multiple reflections, resistive mode suppressors are required and this introduces some amplitude modulation of the transmitted wave. These mode suppressors, together with the two sections tapering from circular waveguide to rectangular waveguide, make the overall length rather large. The standing-wave ratio of the phase changer is not of major importance since, in the usual arrangement for phase measurement, the transmitted wave only is of importance and a small standing-wave ratio in the input line to the phase changer is of no consequence.

The transverse moving-dielectric-strip phase changer is not ideally suited for accurate phase measurements, since its calibration law is a rather complicated function of setting and frequency and the overall mechanical movement for 360° phase change is limited to about one-third of the waveguide width. Thus at 3 cm it becomes necessary to be able to set the dielectric strip to within 0·003 cm to obtain a reading accuracy of 1°.

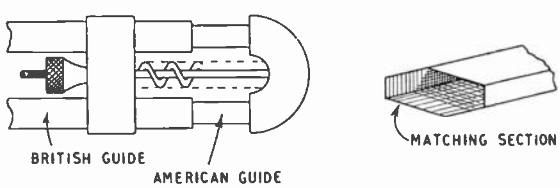


Fig. 1. *Trombone-type phase changer.*

The trombone phase changer is more suitable for phase measurements since it is of simple construction and has a linear law; i.e., the phase shift varies directly with the mechanical displacement. This type has been quite widely used but there does not seem to be any published information on its behaviour. At 3 cm, a simple type of trombone phase changer may be constructed from standard 1-in. × ½-in. British waveguide and 0·9-in × 0·4-in. American waveguide, as illustrated in Fig. 1. The dimensions of the American guide are such that it makes a sliding fit in the British guide. By cutting away the top and one side of the ends of the American guide for a quarter-wave length a quarter-wave transformer is obtained that will match the American guide to the British guide[11]. For a 360° phase change the total mechanical movement is half the guide wavelength so that the mechanical movement must be measured by a calibrated dial gauge reading to 0·001 cm. Measurements made on such a phase changer showed that the departure from a linear law was not greater than ± 0·60° over a 360° range. The variation was probably due largely to small variations in the waveguide dimensions.

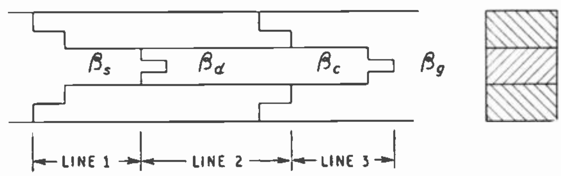


Fig. 2. *Waveguide section with central dielectric strip displaced longitudinally with respect to the side dielectric strips.*

2. Rectangular Waveguide Phase Changer with a Linear Characteristic

During some work on dielectric discontinuities in waveguides it was noticed that the sum of the propagation constants of an empty and a completely-filled guide was less than the sum of the propagation constants for a guide with a dielectric strip in the centre and a guide with dielectric strips along the sides. It was realized that this principle could be used to produce a phase shift by displacing the central dielectric strip longitudinally with respect to the two side strips which are kept fixed.

2.1 *Theory*

With reference to Fig. 2, it is found that $\beta_c + \beta_s > \beta_d + \beta_g$. For a 3-cm waveguide and polystyrene with a dielectric constant of 2·54 this difference is of the order of 0·40 radian per cm. Let this difference be called α; i.e., $(\beta_c + \beta_s) - (\beta_d + \beta_g) = \alpha$. When the central dielectric strip is displaced longitudinally by an amount x cm the effect is to lengthen sections 1 and 3 and to shorten section 2 and the empty waveguide length all by this same amount x. A phase shift of αx radians is produced and is, therefore, directly proportional to the displacement. Furthermore if the ends of the dielectric strips can be matched at a given frequency, they are matched for all settings of the phase changer at this frequency. For values of dielectric constant greater than 2·54 the phase shift produced per cm of movement is greater. Typical values that can be obtained are shown in Fig. 3. It is, however, essential to use a dielectric constant less than the value which allows the H_{03} mode to propagate in the completely-filled guide. The H_{02} mode is not excited because of the symmetry of the structure. Also from Fig. 3 it can be seen that a maximum phase shift is produced when the thickness of the centre strip is between a quarter and a third of the guide width a.

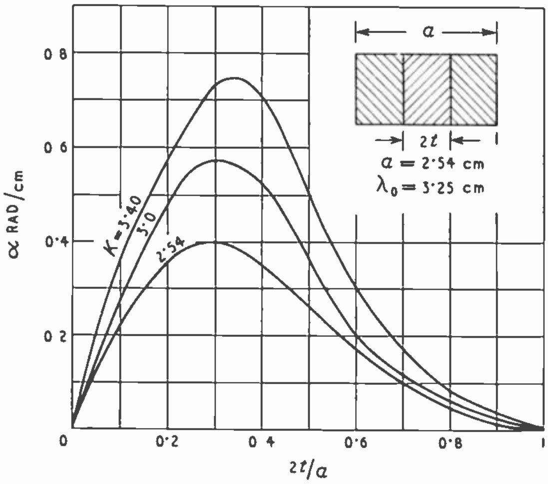


Fig. 3. *Variation of α with dielectric strip thickness.*

The dielectric strips can be matched by quarter-wave steps. If a standing-wave ratio of less than 1·05 is to be maintained over the desired band, then the maximum value of the reflection coefficient for the whole structure must not exceed 0·025. Since there are four discontinuities and the reflected waves from these are almost certain to add in phase for some particular setting of the phase changer, each discontinuity must be matched to give a reflection coefficient not greater than 0·025/4 or 0·006 over the required band. This is quite a severe matching condition and a two- or three-section transformer is required if a large bandwidth is to be obtained.

It is shown in Appendix 1 that β_c and β_s are given to within a few per cent by the following expressions.

$$\beta_c^2 = 2k_0^2\left[\frac{Kt}{a}+\frac{d}{a}+\frac{K-1}{2\pi}\sin\left(\frac{2\pi t}{a}\right)\right]-\frac{\pi^2}{a^2} \quad (1)$$

$$\beta_s^2 = 2k_0^2\left[\frac{t}{a}+\frac{Kd}{a}-\frac{K-1}{2\pi}\sin\left(\frac{2\pi t}{a}\right)\right]-\frac{\pi^2}{a^2} \quad (2)$$

where $2t$ is the thickness of the centre dielectric strip, $k_0^2 = \omega^2\mu_0\epsilon_0$, K is the relative permittivity of the dielectric and $2d = a-2t$. Although these expressions are not suitable for calculating the phase shift produced per cm of displacement they are of value in analysing the general properties of the phase changer. The phase shift produced is a second-order quantity and a small error in β_c or β_s produces a much larger error in the calculated phase shift per cm. When the above two equations are added it is seen that

$$\beta_c^2 + \beta_s^2 = (K+1)\,k_0^2 - 2\,(\pi/a)^2 \quad .. \quad (3)$$

an expression that is independent of the thickness of the dielectric strips. To maximize the phase shift it suffices to maximize $(\beta_c + \beta_s)$ subject to the condition that $\beta_c^2 + \beta_s^2 =$ constant. By elementary calculus it is easily shown that the maximum phase shift is obtained if $\beta_c = \beta_s$. Substituting this result in equations (1) and (2) and equating them gives

$$\sin\left(\frac{2\pi t}{a}\right) + \frac{2\pi t}{a} = \frac{\pi}{2} \quad .. \quad .. \quad .. \quad (4)$$

The solution to the above equation is $2t/a = 0{\cdot}265$ and it appears that the only parameter that fixes the maximum phase shift is the ratio of the thickness of the centre dielectric strip to the waveguide width. An exact calculation shows that the above result is accurate for dielectric constants less than about 2·65. As the dielectric constant is increased up to 3·50 this optimum ratio approaches 0·33. (Fig. 3.)

Movement of the central dielectric strip can be obtained by cutting a longitudinal slot in the centre of the broad face of the guide and cementing a short dielectric tongue to the central dielectric strip, this tongue protruding up through the slot and being fixed to some mechanical drive. Although such a slot does not cut any current flow lines if it is narrow and centred in the broad face of the guide, it may still produce some undesirable effects in the linear characteristics of the phase changer. The phase shift produced is a second-order quantity and a small change of the guide wavelength in any one section produces a considerably larger percentage change in the phase shift per cm. For this reason the slot must be positioned so that the central dielectric strip moves in the slotted waveguide section over its total range of movement.

2.2. *Errors*

If the four discontinuities are perfectly matched then the linear characteristic of the phase changer can be destroyed only by variation in the dielectric constant throughout the strips, non-parallelism of the strips or by the effect of the slot. The first of these can be quite safely ignored, since present manufacturing technique ensures that the dielectric constant of a given sample remains essentially uniform throughout the sample.

The variation in any of the propagation factors caused by changes in the dimensions of the centre or side strips can be determined by differentiating the appropriate expression for the propagation constants with respect to the transverse dimensions of the strips. As an example, consider a variation of $\pm$ 0·01 cm in the thickness of the centre dielectric strip. From equation (1) it is seen that

$$\delta\beta_c = \frac{(K-1)\,k_0^2}{\beta_c a}\left(1 + \cos\frac{2\pi t}{a}\right)\delta t \quad .. \quad (5)$$

For $K = 2{\cdot}56$, $k_0^2 = 3{\cdot}75$, $\beta_c = 2{\cdot}26$, $\beta_d = 2{\cdot}76$, $2t/a = 0{\cdot}265$, $\alpha = 0{\cdot}40$ rad/cm, and $2\delta t = -0{\cdot}01$ cm, the change in β_c is less than 0·01 and hence the change in α is of the order of $2\frac{1}{2}\%$. If this change in thickness occurs in the region where the centre strip is moving between the two side strips then the effect on the linear characteristic is very small since the decrease in β_c is almost completely neutralized by the decrease in β_d. The actual change in β_d with gap thickness $\delta_s = 0{\cdot}005$ cm as illustrated in Fig. 4 is found to be

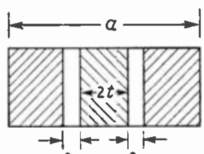


Fig. 4. *Gaps Δs between the dielectrics are illustrated here.*

$$\delta\beta_d = -\frac{(K-1)}{\beta_d a}\,k_0^2\left[1 + \cos\frac{2\pi t}{a}\right]\delta_s = -0{\cdot}008 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

by an application of the Ritz method similar to that used to obtain equations (1) and (2). The net residual change is of the order of $\frac{1}{2}\%$ in the phase shift produced per cm. By careful machining, the tolerance on the transverse dimensions can be kept to within $\pm$ 0·003 cm, so that this error can be made negligible.

The perturbation in the guide wavelength due to a covered slot can be calculated by considering propagation along the transverse axis of the waveguide [12]. At the cut-off frequency the problem reduces to that of an equivalent parallel-plate transmission line with a change in height. Consider a slotted waveguide as in Fig. 5. The equivalent circuit consists of two lengths of transmission line with characteristic admittance inversely proportional to the guide height and a capacitive susceptance at the junction. For the usual thickness of guide wall the junction susceptance can be neglected. The transverse wave number k_x is found from the condition that the admittance at the centre of the guide must be zero. Neglecting the junction susceptance this condition is

$$j\left(-\cot k_x\,l_1 + \frac{b}{b+y}\tan k_x\,l_2\right) = 0 \quad .. \quad (7)$$

or $\tan k_x l_2 = \dfrac{b+y}{b}\cot k_x l_1$. If $l_2 \ll l_1$ this can be approximated by $k_x l_1 \tan k_x l_1 = \dfrac{l_1}{l_2}\dfrac{b+y}{b}$. The perturbed propagation constant β_g is obtained from the eigenvalue equation $\beta_g^2 = k_0^2 - k_x^2$. Similar techniques may be used when the guide is partially filled with a dielectric. In this case, the transverse wave numbers are different in the two sections but are related through the condition that the propagation constant along the z axis must be the same over the guide cross section.

Multiple reflections between the four discontinuities will also affect the linear law of the phase changer but this is only a second-order effect, being proportional to the square of the reflection coefficient, and can be neglected. (See Appendix 2.)

Slight asymmetry in the dielectric strips results in the excitation of the H_{02} mode which can propagate in the completely-filled guide, and this affects the linear law of the phase changer. However, careful construction can ensure that the amplitude of this unwanted mode does not become large enough to cause any significant departure from linearity.

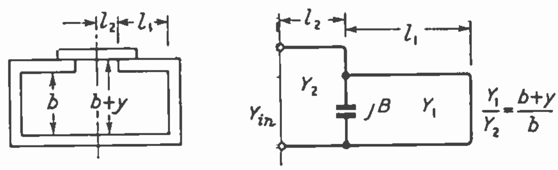


Fig. 5. *Slotted waveguide* (a) *and its equivalent circuit* (b).

2.3. *Working Model*

One model was constructed using polystyrene strips in 0·9-in. by 0·4-in. guide. The central dielectric strip was fixed to a brass slide which covered the slot on top of the guide by a polystyrene key. The guide had a $\frac{3}{32}$-in. slot centred longitudinally along the guide. The central dielectric strip moved in the slotted section for the first half of the range and in the unslotted section for the last half of the range. The phase shift produced per cm of movement was found to be about 5% less when the dielectric strip moved in the slotted section than when it moved in the unslotted section. A theoretical calculation of the change in the guide wavelength due to the covered slot indicated that this effect was caused by the slot. Consequently a slightly modified model was built. This latter model had a $\frac{1}{16}$-in. slot located at one end of the guide such that the dielectric strip moved in the slotted section for the whole range of movement. The effect of the slot can be reduced by cutting a longitudinal groove in the broad face of the guide and making the brass slide with a tongue which projects down into this groove. Making the groove half the thickness of the guide wall effectively reduces the thickness of the slot to one half of its former value. However, so long as the dielectric strip moves in the slotted section over its total range of movement the slot does not alter the linear characteristic of the phase changer. The two larger discontinuities were matched by three-step binomial transformers while the two smaller discontinuities were matched by two-step binomial transformers. The design of the quarter-wave matching transformers is discussed in Appendix 3.

A simple mechanical drive is a rack and pinion arrangement with the rack bolted to the brass slide. A simplified drawing of a suitable design is given in Fig. 6. The dimensions of the quarter-wave steps are given in Table 1.

This model gives quite a good overall perform-

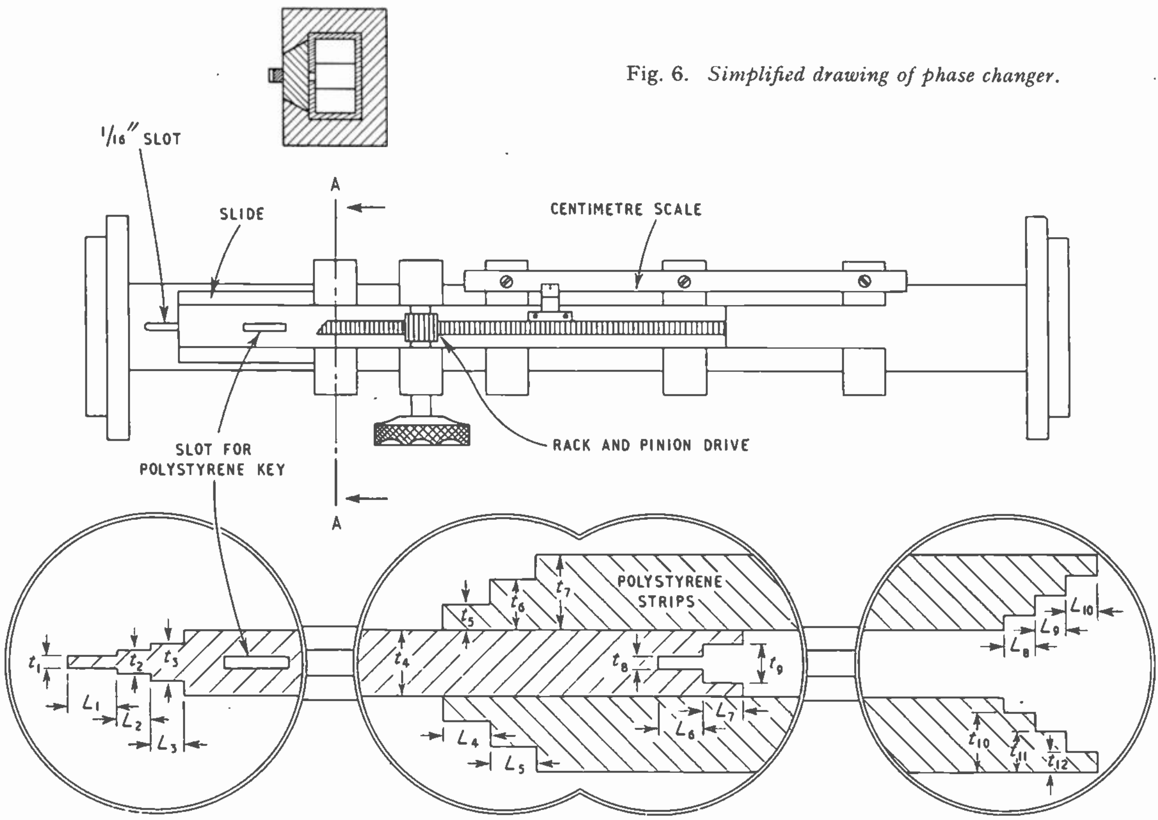


Fig. 6. *Simplified drawing of phase changer.*

ance. The worst standing-wave ratio of the phase changer and the two choke couplings is plotted against wavelength in Fig. 7. Although a perfect match is never obtained the performance remains quite constant for wavelengths from 3·00 to 3·50 cm. Measurements made on several choke couplings showed that they produced a standing-wave ratio of the order of 1·01 per coupling throughout the band. In view of this, it is reasonable to assume that the dielectric strips were matched to produce a standing-wave ratio of less than 1·03 in the vicinity of the design wavelength. No doubt the slot had some effect on the matching of the centre strip. It is thought that a better match might be obtained by making the slide with a projecting tongue as explained earlier and taking more care with the matching

TABLE 1

Length of steps in inches	Thickness of steps in inches
$L_1 = 0{\cdot}425$	$t_1 = 0{\cdot}020$
$L_2 = 0{\cdot}350$	$t_2 = 0{\cdot}088$
$L_3 = 0{\cdot}288$	$t_3 = 0{\cdot}195$
$L_4 = 0{\cdot}257$	$t_4 = 0{\cdot}250$
$L_5 = 0{\cdot}234$	$t_5 = 0{\cdot}025$
$L_6 = 0{\cdot}236$	$t_6 = 0{\cdot}122$
$L_7 = 0{\cdot}261$	$t_7 = 0{\cdot}324$
$L_8 = 0{\cdot}293$	$t_8 = 0{\cdot}063$
$L_9 = 0{\cdot}353$	$t_9 = 0{\cdot}189$
$L_{10} = 0{\cdot}426$	$t_{10} = 0{\cdot}293$
	$t_{11} = 0{\cdot}216$
	$t_{12} = 0{\cdot}126$

Dimensions for quarter-wave steps. Notation as in Fig. 6. All tolerances $\pm$ 0·001 in. Design wavelength 3·25 cm. 0·9 in. $\times$ 0·4 in. rectangular waveguide.

of the dielectric strips. All the standing-wave ratios were obtained from the width of the Weissfloch curves which were plotted for every setting of the phase changer [13]. At each frequency eleven curves were plotted corresponding to one centimetre increments in the phase changer setting.

If the phase changer and the two choke couplings are replaced by an equivalent circuit consisting of two lengths of transmission line and an ideal transformer, then it is readily seen that the displacement of the mean centre of the Weissfloch curve of $\beta_g (l + d)$ against $\beta_g l$ with phase changer setting gives the change in phase of the wave transmitted past the output coupling. The position of the movable short circuit is a distance l from the arbitrary reference plane in the output guide, while d is the distance of a field minimum from some convenient reference plane in the input guide. At 3 cm using a square-wave modulated klystron, a high-gain selective amplifier, and measuring the position of a field minimum with a dial gauge reading to 0·011 cm it is possible to measure the phase of the transmitted wave to better than 0·2°. It was found that the maximum departure from linearity of the phase changer was less than 1° over the whole band. This departure from linearity was probably due to a combination of several effects such as small variations in the waveguide dimensions, multiple reflections between the ends of the dielectric strips, slight non-parallelism of the dielectric strips, and excitation of the H_{02} mode. The mean phase shift per cm is given in Table 2. This mean value was determined so that the sum of the errors over the total range was zero; i.e., $\sum_{n=0}^{10} \beta_g (nd_0 - d) = 0$, where $\beta_g d_0$ is the mean phase shift per cm and $\beta_g d$ is the measured total phase change in the transmitted wave with reference to the initial setting of the phase changer $n = 0$. The mean phase shift produced was a few per cent below the theoretical value because of the effect of the slot.

TABLE 2

λ (cm)	3·00	3·15	3·25	3·35	3·50	3·65
α $\left(\frac{\text{rad}}{\text{cm}}\right)$	0·417	0·397	0·378	0·375	0·371	0·368

2.4. *Comparison with other Types*

The above model has an overall length of 14½ in. and covers a range of 215°. If the dielectric strips are made 2·5 in. longer a range of 360° can be covered and the overall length becomes 20 in. The use of material with a dielectric constant of about 3 would reduce this overall length considerably. With a vernier scale reading to 0·01 cm the phase changer can be set accurately to better than 0·25°. This type of phase changer is physically smaller than the rotary type and of considerably simpler construction. There are only four discontinuities to match as compared with as many as eight or more for the rotary phase changer. Further, there is no evidence that the rotary phase changer is any more accurate. One disadvantage is the variation of α with wavelength. However, this variation is small and consequently not serious. Because of the linear law this type of phase changer is more suitable for phase measurements than the transverse moving-dielectric strip type. Its performance is comparable to that of the trombone type and avoids the 180° reversal of the waveguide.

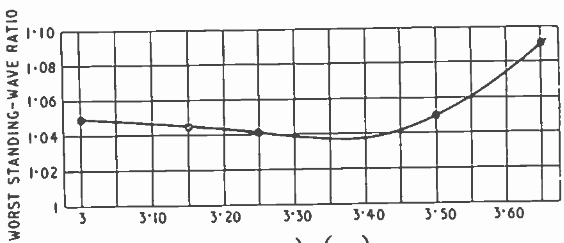


Fig. 7. *Standing-wave ratio of phase changer and two choke couplings.*

3. Conclusion

A new type of rectangular waveguide phase changer has been developed. It is of simple construction and has a linear law, the phase change in the transmitted wave being proportional to the mechanical displacement. Its performance is comparable to that of the trombone phase changer, the departure from linearity being less than one degree for wavelengths between 3·00 and 3·50 cm. The largest standing-wave ratio of the phase changer for any setting is less than 1·05 throughout this band.

APPENDIX 1.

Approximations to the propagation constants of an inhomogeneously filled guide.

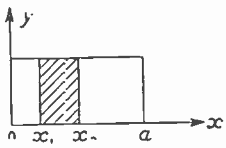


Fig. 8. *Rectangular guide partly-filled with lossless dielectric.*

For a rectangular guide partially filled, as in Fig. 8, and supporting H_{0n} modes an approximation to the propagation constant, β, can be obtained by minimizing the following energy integral,[14].

$$\beta^2 = \frac{\int_0^a \left[K(x)\, k_0^2\, E_y^2 - \left(\frac{\delta E_y}{\delta x}\right)^2 \right] dx}{\int_0^a E_y^2\, dx}.$$

where $K(x)$ is the dielectric constant of the filling and is considered as a function of x and $k_0^2 = \omega^2\mu_0\epsilon_0$. If E_y is approximated by the series,

$$\sqrt{\frac{2}{a}}\sum_{n=1}^{N} a_n \sin(n\pi x/a),$$

then

$$\beta^2 = \left[\sum_{n,s=1}^{N} a_n a_s\, T_{ns} - \sum_{n=1}^{N} (n\pi/a)^2\, a_n^2\right] \Big/ \sum_{n=1}^{N} a_n^2$$

where

$$T_{ns} = \frac{2k_0^2}{a}\int_0^a K(x)\sin(n\pi x/a)\sin(s\pi x/a)\,dx.$$

When the partial derivatives with respect to a_n are set equal to zero a set of N homogeneous equations is obtained. For a non-trivial solution, the determinant of this set of equations must vanish and this leads to N roots of β^2, which are approximations to the first N propagation constants of the partially-filled guide. The smallest of these is the propagation constant of the H_{01} mode. Equations (1) and (2) in Section 2.1 were obtained by using a single term approximation to E_y. Thus for the central dielectric strip

$$\beta_c^2 = (a_1^2 T_{11} - \pi^2 a_1^2/a^2) \div a_1^2 = T_{11} - \pi^2/a^2,$$

where

$$T_{11} = \frac{4k_0^2}{a}\left[\int_0^{\frac{a}{2}-t}\sin^2(\pi x/a)\,dx + K\int_{\frac{a}{2}-t}^{a}\sin^2(\pi x/a)\,dx\right]$$

$$= k_0^2\left[\frac{Kt}{a} + \frac{a-t}{a} + \frac{K-1}{2\pi}\sin\left(\frac{2\pi t}{a}\right)\right]$$

APPENDIX 2.

Effect of multiple reflections on phase of transmitted wave.

Consider n arbitrary spaced similar discontinuities as in Fig. 9. Let each discontinuity have a reflection coefficient ρ, where ρ is small compared with unity. Since ρ is small it is sufficient to consider only second-

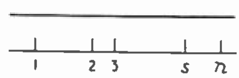


Fig. 9. *Arbitrarily-spaced discontinuities in a guide.*

order reflections between the elements. The voltage transmission coefficient of each element is $t = (1 - \rho^2)^{1/2}$. Let a wave of unit amplitude and zero phase be incident on element 1. The principal wave transmitted past element n is $t^n e^{-j\theta}$, where θ is the phase angle of the wave. Incident on element s is a principal wave of magnitude t^{s-1}. This wave is reflected from element s and reflected back again by all the elements preceding s. Thus element s will give rise to $s-1$ second-order waves whose amplitudes does not exceed $\rho^2 t^n$ after being transmitted past element n. The worst possible case is when all these second-order waves from all the elements add in phase and are $\pm 90°$ out of phase with the principal wave. The magnitude of all the second-order waves will not exceed

$$t^n\rho^2\sum_{s=2}^{n}(s-1) = \tfrac{1}{2}n(n-1)\,\rho^2 t^n$$

Therefore, the maximum change in the phase of the total transmitted wave is $\pm\tan^{-1}\left[\frac{n(n-1)\rho^2}{2}\right]$. When $n = 6$ and $\rho = 0{\cdot}01$, this change in phase is of the order of $\pm 0{\cdot}09°$ which is negligible.

APPENDIX 3.

Design of quarter-wave matching transformers

The results of some theoretical and experimental work done on quarter-wave dielectric matching sections has shown that, for the symmetrical H-plane dielectric step which excites only odd order H_{0n} modes, the reactance at the step discontinuity is negligible so far as the design of a quarter-wave matching section is concerned[15]. The step illustrated in Fig. 10 may be considered as an abrupt change from the impedance of the empty guide to

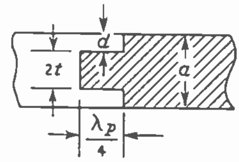


Fig. 10. *Dielectric step in matching transformer.*

that of the partially-filled guide. Further, the impedance of any section may be assumed proportional to the guide wavelength in that section. For the centrally-positioned dielectric strip the wavelength in the partially-filled guide is obtained from a simultaneous solution of the following two equations,[12,15]

$$\beta_p^2 = (2\pi/\lambda_p)^2 = K\,k_0^2 - l_1^2$$
$$= k_0^2 - k_1^2 \quad k_1\cot l_1 t = l_1\tan k_1 d$$

where l_1 and k_1 are the transverse wave numbers in the dielectric-filled section and empty section respectively, λ_p is the guide wavelength in the partially-filled guide and t and d are the dimensions given in Fig. 10. The smallest roots of k_1 and l_1 are appropriate for the H_{01} mode while the higher roots give the propagation factors for the higher-order H_{0n} modes. When the role of dielectric and air is interchanged (i.e., dielectric strips along the sides of the guide) it is necessary to interchange l_1 and k_1 in the second equation given above. Each dielectric strip is now of thickness d. These equations may be used to calculate the constant α for the phase changer.

For a maximally-flat transformer, the impedances of the sections are arranged according to the coefficients of the binomial expansion and each section is a quarter-wave long at the design wavelength, l. It is also possible to design the matching transformers so that the reflection coefficient vanishes at as many different frequencies as there are quarter-wave sections,[7,16]. This design procedure will not be considered here.

REFERENCES

[1] Southworth, "Principles and Applications of Waveguide Transmission", Chap. 9 (D. Van Nostrand).

[2] A. G. Fox, "An Adjustable Waveguide Phase Changer", *Proc. Inst. Radio Engrs*, 1947, Vol. 35, pp. 1489–1498.

[3] A. G. Fox, "An Adjustable Waveguide Phase Changer", *Bell Lab. Rec.*, 1948, Vol. 26, pp. 245–250.

[4] J. J. Brady, M. D. Pearson, S. Peoples, "Squeeze-section Phase Shifter for Microwave Measurements", *Rev. sci. Instrumen.*, Nov. 1952, Vol. 23, pp. 601–604.

[5] G. E. Bacon, J. C. Duckworth, "Some Applications of the Principle of Variation of Wavelength in Waveguides by the Internal Movement of Dielectric Sections", *J. Instn elect. Engrs*, Pt. IIIA, 1946, Vol. 93.

[6] S. Sonoda, T. Makimoto, S. Yokouchi, "On the Design of the Dielectric Slab in Waveguide Phase Shifters", *Mem. Sci. Industrial Research*, Osaka Univ., 1951, Vol. 8.

[7] G. J. Halford, "A Wide-band Waveguide Phase Shifter", *Proc. Instn elect. Engrs*, Pt. III, 1953, Vol. 100.

[8] G. L. Ragan, "Microwave Transmission Circuits", Radiation Laboratory Series, Vol. 9 (McGraw-Hill), p. 514.

[9] G. L. Ragan, "Microwave Transmission Circuits", *Ibid.*, p. 481.

[10] A. J. Simmons, "Errors in a Microwave Rotary Phase Shifter", *Proc. Inst. Radio Engrs*, July 1952, Vol. 40, p. 869.

[11] L. G. H. Huxley, "A Survey of the Principles and Practice of Waveguides". (Cambridge Univ. Press) 1947.

[12] Montgomery, Dicke, Purcell, "Principles of Microwave Circuits". Radiation Laboratory Series, Vol. 8 (McGraw-Hill), Section 6.16.

[13] Weissfloch, "*Hochfreq. u. electro*". 1942, Vol. 60, p. 67.

[14] G. Chambers, "Compilation of the Propagation Constants of an Inhomogeneously Filled Waveguide, *Brit. J. Appl Phys.*, Jan. 1952, Vol. 3.

[15] R. E. Collin: University of London. Ph.D. Thesis, 1953.

[16] R. E. Collin and J. Brown, "Transmission-Line Matching System", *Wireless Engineer*, Feb. 1954, p. 31.

CORRESPONDENCE

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

Pulse Response of Signal Rectifiers

SIR,—In his article in your January issue, Mr. M. V. Callendar makes a significant contribution and also draws attention to many complex problems still unresolved. It may, therefore, be of interest to note that partial answers to some of these were given in an article on "The Detector", published in your journal as long ago as September 1932, which has probably been forgotten because of its unusual approach to the problems.

Mr. Callendar, referring, in the second paragraph of his introduction, to the design of receivers for picture or pulse-modulated transmissions, observes "the time constant of the rectifier is usually taken as that of the rectifier output (load) circuit" but goes on to explain that some experimental observations do not accord with this. Later, on page 7, he writes "The problem of the pulse response in the general case, where C_dR_d is not negligible, awaits the attention of those versed in the application of operational methods to non-linear circuits. The only point which is certain is that the increase in overall time constant as C_dR_d is increased does not correspond to that given by a linear amplifier stage with time constant C_dR_d". In the summary at the end of my (1932) article I concluded "The time constant of the (equivalent) filter stage may be taken as roughly the geometric mean of the time constants of the detector circuit when the rectifier is conducting and non-conducting". It may now be added that the effective internal resistance of the input tuned circuit may often be replaced by an appropriate increase in the rectifier resistance when conducting.

The degree to which the observed response, quoting Mr. Callendar, "to a positive going r.f. step modulation" differs from "the response to a negative step" is directly given by the change of slope along the appropriate response line such as those of Figs. 8(a), (b) and (c) of my article.

In complex non-linear phenomena such as these it is not often possible to formulate both precise and general relations, but the approximate significance of component elements may be derived from simple analyses such as those in my article. These results may be used to supplement the analyses such as those of Wheeler and Terman developed by Mr. Callendar where these encounter their peculiar limitations.

W. B. LEWIS

Atomic Energy of Canada, Ltd.,
Chalk River, Ontario,
Canada.
1st February 1955.

Analogy of Ferrite in Waveguide

SIR,—Gyroscopes, behaving classically, have been used as models in the qualitative investigation of gyromagnetism in ferrites. This seems a justifiable aid to thinking. In a pioneer paper by Hogan in the *Bell System Technical Journal* of January 1952, a gyroscope suspended above its centre of gravity is suggested. As the major contribution to the magnetization of a ferromagnetic material is believed to be supplied by spinning electrons, and as, in the cases considered, these are supposed to find themselves in approximately uniform magnetic fields of changing direction, it would appear that the gyroscopes representing them should be suspended not above, but rather at their centres of gravity.

Such a model is used in the following attempt to provide a picture of asymmetric attenuation due to a ferrite rod suitably placed in a waveguide, across the axis of which is applied a steady magnetic field.

In Fig. 1 the magnetic field pattern of an H_{01} wave in a rectangular guide is shown. It will be seen that, as the pattern proceeds along the guide, the field at a point on the dotted line AB will rotate. The direction of rotation depends upon the direction of travel of the wave, being clockwise if the wave travels from right to left and counter-clockwise if it travels from left to right.

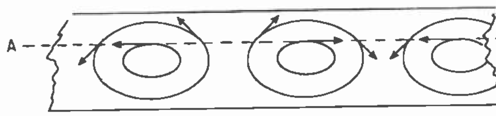


Fig. 1.

If a steady field is applied at right-angles to the plane of the pattern of Fig. 1. it will combine with the rotating field to produce a conically-oscillating field as indicated in Fig. 2. The disc at the point P in the figure represents a spinning electron in a ferrite rod assumed to be placed along the line AB in Fig. 1. Due to its magnetic moment the electron will try to take up such a position that its axis is in line with the resultant field. The axis of torque will be perpendicular to the component of field lying in the plane of spin, and will itself lie in that plane. Due to its rotation the electron will precess about an axis in the plane of spin perpendicular to the torque axis. At some frequency of rotation, which in this context we will designate as resonant, the precession will be at the same speed as the rotating field and in such a phase-relationship to it that a large component of the axis of magnetization of the electron will lie along the rotating H-vector in the guide. The sense of the precession will depend upon the direction of this vector, and so, as we have seen, upon the direction of travel of energy in the guide. For one direction the electron flux will reinforce that of the travelling wave, for the other direction it will weaken it. In the former case the energy will be constrained to flow largely through the lossy ferrite and will be severely attenuated; in the latter it will tend to by-pass the ferrite. We might be granted licence to say that *at the wave frequency* the ferrite behaves paramagnetically to waves travelling in one direction, and diamagnetically to waves travelling in the other.

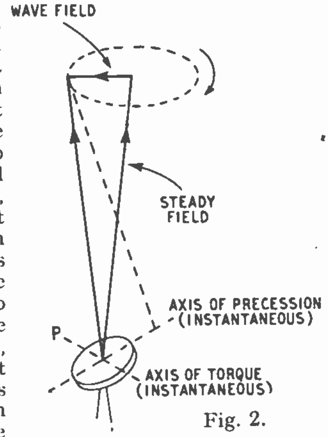


Fig. 2.

So far the only test to which this picture has been put has been a prophecy by its help of the direction of the steady field required for high attenuation. The picture indicated the correct one of the two possible answers, which was comforting rather than convincing.

The author's thanks are due to his colleague, Mr. M. F. McKenna, for having set up the apparatus and demonstrated the effect. He is about to investigate the field pattern near the ferrite as a test of the validity of the model.

G. F. NICHOLSON

Royal Naval College,
Greenwich, London, S.E.10.
28th January 1955.

NEW BOOKS

English-German Technical and Engineering Dictionary

By Dr. LOUIS DE FRIES of Iowa State College. Pp. 997 +xv. McGraw-Hill Publishing Co., Ltd., 95 Farringdon Street, London, E.C.4. Price £7.

This is a companion volume to the already published German-English Dictionary. It contains 130,000 terms used in all branches of engineering and technology, including hundreds of new terms which have evolved in recent years in radar, television, electronics, rocket and jet propulsion, nuclear engineering, etc. It has been a great undertaking and involved the editing of over 14,000 pages of manuscript. A list is given of 76 collaborators to whom the author expresses his indebtedness, and a list of 85 reference works. The last ten pages are devoted to abbreviations, the meanings of which are given in both languages. Presumably this volume will not have such a large appeal in this country as its companion volume, since people are more likely to want to know the meaning of German terms than to convert English into German. Where the spelling usage differs in this country and America, the latter is adopted; for example, one looks in vain for 'caesium' or 'centre' but finds 'cesium' and 'center'. Similarly 'beveled' and 'traveler' look strange to an English reader, but these are minor details and are unavoidable. The same cannot be said of "Ohms' law", which is a definite error, but we feel sure that there are very few such errors, for the book gives every appearance of having been prepared with great care.

This is undoubtedly a valuable addition to the available technical dictionaries.

G. W. O. H.

Theory and Design of Electron Beams (2nd Edition)

By J. R. PIERCE. Pp. 222 + xiv. D. Van Nostrand Co., Inc., 250 Fourth Avenue, New York, U.S.A. Price $4.50.

Applied Electronics (2nd Edition)

By TRUMAN S. GRAY. Pp. 881 + xxviii. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 72s.

Amplitude-Frequency Characteristics of Ladder Networks

By E. GREEN, M.Sc. Pp. 156 + iii. Marconi's Wireless Telegraph Co., Ltd., Marconi House, Chelmsford, Essex. Price 25s.

Antenna Diagrams

INTERNATIONAL RADIO CONSULTATIVE COMMITTEE (C.C.I.R.). Pp. 52. The Publications Department, International Telecommunication Union, Palais Wilson, Geneva, Switzerland. Price S.Fr. 18.45 (S.Fr. 17.15 to I.T.U. members).

This is a loose-leaf book with a page 11⅝ in. by 8 in. containing formulae and power-distribution diagrams for vertical, horizontal dipole curtain and rhombic aerials as well as of aerials for tropical broadcasting.

Electronic Measuring Instruments

By E. H. W. BANNER, M.Sc., M.I.E.E., M.I.Mech.E., F.Inst.P., M.Const.E. Pp. 395 + xiv. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 45s.

MEETINGS

I.E.E.

9th March. "Artificial Reverberation", by P. E. Axon, O.B.E., M.Sc., Ph.D., C. L. S. Gilford, M.Sc. and D. E. L. Shorter, B.Sc.(Eng.).

21st March. "Materials for Valves", discussion to be opened by R. O. Jenkins, Ph.D.

5th April. "High Speed Electronic-Analogue Computing Techniques", by D. M. MacKay, B.Sc., Ph.D.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30.

Brit.I.R.E.

30th March. Discussion on "The Maintainability of Service Equipment". Principal Speakers:—Capt. (L) A. J. B. Naish, R.N., M.A., Maj. R. B. Brenchley, Wing Comdr. G. C. Godfrey and G. W. A. Dummer, M.B.E. Meeting will commence at 6.30 at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for January 1955

Date 1955 January	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	−0·7	−4	NM
2	−0·7	−3	NM
3	−0·6	−21 *	NM
4	−0·8	−1	+77·5
5	−0·8	−1	+79·2
6	−0·7	0	NT
7	−0·7	−1	+78·4
8	−0·7	0	NM
9	−0·7	+1	NM
10	−0·7	+1	+79·4
11	−0·7	+1	+77·6
12	−0·8	+2	+76·6
13	−0·6	+1	+75·9
14	−0·6	+2	+75·4
15	−0·6	0	NM
16	NT	+2	NT
17	NM	+2	+73·0
18	−0·6	+3	+71·6
19	NT	+2	NT
20	−0.5	+2	+70·5
21	−0·6	+2	+69·0
22	−0·6	+2	NM
23	−0·6	+2	NM
24	−0·5	+2	+67·5
25	−0·5	+2	+67·0
26	−0·6	+1	+65·9
27	−0·5	+1	+65·0
28	−0·5	+2	+64·1
29	−0·5	+3	NM
30	−0·5	+3	NM
31	−0·5	+2	+61·5

The values are based on astronomical data available on 1st February 1955.

The transmitter employed for the 60-kc/s signal is sometimes required for another service.

NM = Not Measured.
NT = No Transmission.
* = B.B.C. report faulty adjustment of frequency between 0900 and 1200 G.M.T.