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Colin Hinson
In the village of Blunham, Bedfordshire.

THE QUADRANT AERIAL: AN OMNI-DIRECTIONAL WIDE-BAND HORIZONTAL AERIAL FOR SHORT WAVES*

By N. WELLS, M.Sc.†

(The paper was first received 3rd September, 1943, and in revised form 25th February, 1944.)

SUMMARY

As a preliminary to the problem of designing a simple wide-band short-wave aerial for omni-directional requirements, the paper briefly discusses the comparative merits of vertical and horizontal polarization when low suspension is involved. The stages leading to the development of a novel form of aerial based on known principles are outlined, and the antecedent and subsequent experimental tests are described. The factors limiting band-width are examined, and this in turn leads to an examination of the factors determining the form and dimensions of an aerial under working conditions. The economical grouping of the aerial is. considered, and some particulars are given of a check on the arrangement finally adopted. Ideal polar diagrams are illustrated, and attention is devoted to the modifying effect of physical conditions on the vertical diagrams. The paper concludes with an analysis of the quasi omni-directional properties of the horizontal dipole. The mathematical treatment of the diagrams is given in the appendices, and, for consistency, a short appendix outlines the derivation of the formula for transfer loss.

(1) INTRODUCTION

In the field of radio communication it is probable that the advantage of horizontal polarization applies principally to reception, for it seems to be generally agreed that a horizontally supported short-wave aerial has greater immunity from extraneous pick-up than its vertical fellow, apart from that species of atmospheric disturbance caused by rain, to which it is the more susceptible of the two. In any event, if choice is limited to comparatively low aerials employed for reception there can be little doubt that horizontal suspension is the better proposition, for not only is the signal/noise ratio superior in many cases, but when low aerials are involved the vertical polar diagram is less influenced by ground conditions for horizontally than for vertically polarized waves. For high aerials the case is different. It was considerations such as these that led the author to decide upon horizontal polarization for dealing with the aerials of a scheme that entailed' omni-directional reception over a wide band of frequencies, and that also entailed a multiplicity of comparatively low aerials. Later the design was applied to transmission, largely because of the practical advantage that it was independent of an earth system, though obviously the greater freedom from ground absorption applies also to low transmitting aerials.

As is well known, the radiation from a simple straight aerial in its equatorial plane is omni-directional, the diagram being in fact a circle; on the other hand, both because of polarization and because of space-phasing along the path of the conductor, the diagram is directional in a plane containing the aerial. These notes are concerned with the horizontal aerial. The problem here discussed is two-fold, and amounts to asking what form of horizontal aerial will (*a*) yield a substantially omni-directional diagram in the horizontal plane at some given frequency, and (*b*) will not depart from that diagram by more than a predetermined amount when the frequency is changed over a reasonably wide band.

* Wireless Section paper.

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Consideration will show that the problem as a whole can be sub-divided into five main Sections under the following headings:

Form of aerial.

Limitation of frequency band imposed by radiation diagram.

Limitation of frequency band imposed by variations of aerial terminal impedance.

Optimum value of transfer impedance.

Grouping and dimensions.

Three other Sections, dealing respectively with calculated diagrams, stacked aerials, and signals from an ordinary dipole, complete the paper.

(2) CHOICE OF QUADRANT AERIAL

As is well known, there are various types of horizontally disposed aerials giving an omni-directional pattern, perhaps the most familiar being the "Turnstile" due to G. H. Brown, and the ellipsoidal television variant due to N. E. Lindenblad, but, so far as the author is aware, none of these is designed for a sufficiently wide band-width to cover an octave, or more, in frequency. A study of this requirement led the author to the opinion that the inherent flatness, as regards change of frequency, of the ordinary figure-of-eight diagram due to a simple dipole, in combination with some expedient by means of which two such diagrams were superimposed in space quadrature, might best supply the solution. Preliminary analysis confirmed this opinion, and it was decided to carry out a series of experiments to ascertain .to within what limits such a scheme was applicable, and also how best to implement it. The experiments were conducted with a rotating transmitting system and a fixed receiver some little distance away, and were primarily made on what might be termed the quadrant aerial, i.e. a right-angled V-aerial with its limbs fed in anti-phase at the apex, though experiments were also conducted on the balanced centre-fed U-shaped aerial, and on a cruciform aerial in which the two component aerials were, respectively, detuned positively and negatively in order to obtain a rotating field. As a result of these tests it was established that the quadrant aerial gave a substantially omni-directional diagram when its limbs were of the order of $\frac{1}{2}\lambda$ long, and that it was the only -one of those tried out for which the change of shape of diagram with change of frequency was sufficiently slow: since this type of aerial is simple in construction, and also comparatively easily erected, it was selected as fulfilling the object in view.

The frequencies employed during the tests were approximately 60 Mc/s and 100 Mc/s, the latter frequency at a later date for corroborating some of the earlier results. The transmitter consisted of the usual three stages—drive, buffer, and amplifier—developing 25 watts at the lower frequency and IS watts at the higher, with a ,stability of the order of 1 : 500. A rotating framework carried the transmitter at almost ground level, the aerials being supported at 3.8 m, a purely fortuitous and convenient height for the existing framework. The aerials under test were fed via a 500-ohm twin feeder, while a reference dipole, which replaced the aerials before and after each series of readings, was fed via a 70-ohm h.f. twin cable. Some difficulty was experienced in selecting a suitable receiving site, but eventually one was found at a distance of 50 m giving consistent results, inasmuch as the figure-of-eight diagram from the datum dipole was repeatable, symmetrical and comparatively free from distortion. The receiver aerial was supported at a height of 5.5 m, and was eventually similar to the datum dipole, being split and 0.475) overall: it was also connected via a 70-ohm h.f. cable to a screened linear rectifier, the latter being self-contained and supported at about I m above ground. Owing to prevailing conditions it was decided to maintain a steady frequency and to conduct the actual tests by cutting back the length of the transmitting aerial members from oversize to undersize a few centimetres at a time. In practice this appeared to be quite satisfactory, and had something to recommend it, apart from the main reason for its adoption. As a source of reference the signal strength from the dipole was measured at the beginning and end of each series of readings. Preliminary tests were carried out to ensure linearity and to check for possible sources of error.

(3) FREQUENCY BAND AND RADIATION DIAGRAM

Experimental tests with the quadrant aerial have indicated that when the arms are each physically #A long the radiation diagram is very much as shown in Fig. 1(a); this diagram may be regarded

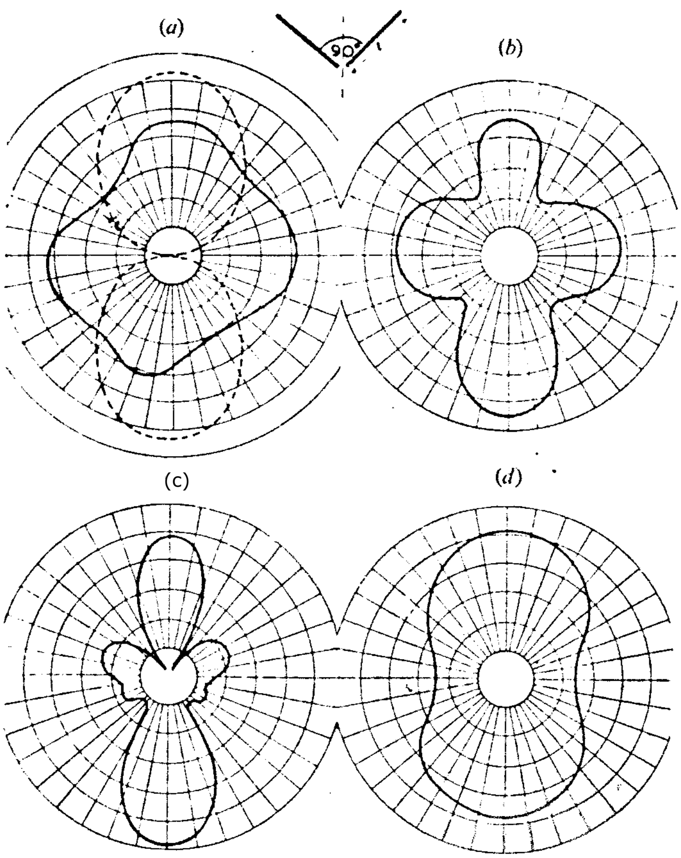


Fig. 1.—Quadrant aerial. Polar diagrams.
(a) 1 0.5A (b) I = 0.65A
(c) I — 0·75λ (d) l = 0·25A

as satisfactory in respect of its omni-directional property, and the corresponding frequency will therefore be taken as a datum value. The dotted-line curve superimposed on Fig. 1(a) is the diagram of the reference dipole, which has been inserted for the purpose of comparison: in all cases the input power is maintained constant. As the frequency increases above datum value there will be an alteration in current distribution, the effect of which is seen in the changing pattern of the diagram. If it is assumed that results are no longer satisfactory when any of the diagram minima fall below 50% of the average strength of Fig. 1(a), then the experiments have shown that this limit is reached when the equivalent physical arm length increases to 0·65λ, corresponding to an increase in frequency of 30% over the datum frequency; the relevant diagram is shown in Fig. 1(b). A further increase in frequency to, say, a total of 50%, corresponds to an equivalent physical arm length of 0·75λ, and yields the diagram of Fig. 1*(c)*, which, obviously, is quite unsatisfactory. With a decrease in frequency the change in diagram is not so critical, but as decrease in frequency implies decrease, of equivalent arm length and, hence, decrease in radiation resistance, there is an increase in ohmic losses which reduces the overall efficiency of the aerial and imposes a limit; generally an equivalent arm length of 0.25A, corresponding to a drop in frequency of 50 %, marks a reasonable limit at the low-frequency end of the band, the corresponding experimental diagram for this arm length being shown on Fig. 1*(d)*.

Summing up, the limit in band-width imposed by a drop in signal strength to 50% of the datum average is 130/50, say 2·5/1·0. The appropriate rule for dimensions would be to make the arm length either 0.25 of the longest wave, or as giving the same answer when the ratio is as above, 0.65 of the shortest wavelength, but, as will be seen in the next Section, a narrower limit is preferable.

The above conclusions are based on experimental results, and it should be noted that the arm lengths are in measured physical units and cannot accord with the actual electric wave distribution. As regards this latter point an inspection of the reaction-resistance curves of Fig. 3A reveals that the natural frequency of the experimental cage aerial is around 8.75 Mc/s, equivalent to a wavelength of 34.29 m, while the physical length of the aerial is 15.25 m; at zero retardation this length would yield a natural wavelength of 30.5 m, so that the actual retardation works out at (34.29 — 30-5)/34.29 = 11-5%; this is appreciable and should be taken into account when comparing experimental results with the calculated curves. These latter are discussed in Section 6, though meanwhile it may be of interest to note that the polar curve for arms of 0·65λ physical length as given in Fig. 1(b) is more nearly comparable with the calculated distribution for 0.75A, while the curve of Fig. 1(c) for arm lengths of 0·75λ is tending towards a calculated distribution of 1·0λ.

(4) LIMITATION IMPOSED BY TERMINAL RESPONSE

Coming to the limitation of frequency band imposed by the permissible variation of terminal impedance, it is well known that the greater the distributed capacitance of an aerial the less its surge impedance, and hence the less the variation of impedance over a given frequency band; in other words we are at first concerned with the problem of designing a high-capacitance cage aerial. In order to facilitate the solution of this problem groups of capacitance curves were calculated for cylindrical cages of various radii and formed of various numbers of wires, each group being for a given height above earth; one such group is shown on Fig. 2A for the height of 20 m, while Fig. 2B exhibits the corresponding values of surge impedance. The formula employed was experimentally verified, and found to be correct within the limits of experimental error; in fact the harmony between calculated and measured values was good. The argument is that the capacitance per cm length of cage is

$$\frac{\quad}{2 \log_e (A/B)} \text{ e.s.u.}$$

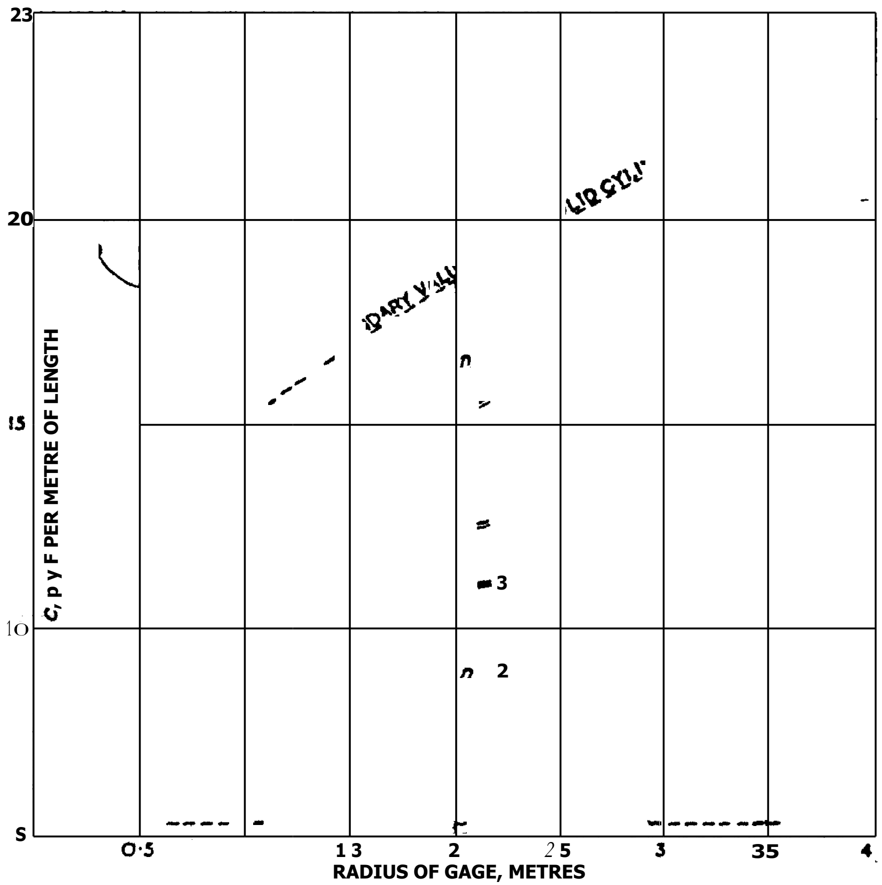


Fig. 2A.—Capacitance of horizontal cage aerial at height of 20 metres above earth.

$$C = \frac{24 \cdot 1\, n}{[n \log_{10}(2h/r) - \log_{10}(nr/R)]}$$

where C = μμF per metre, R = radius of enveloping cylinder (metres), r = radius of conductor (metres), h = height to centre of cage (metres).

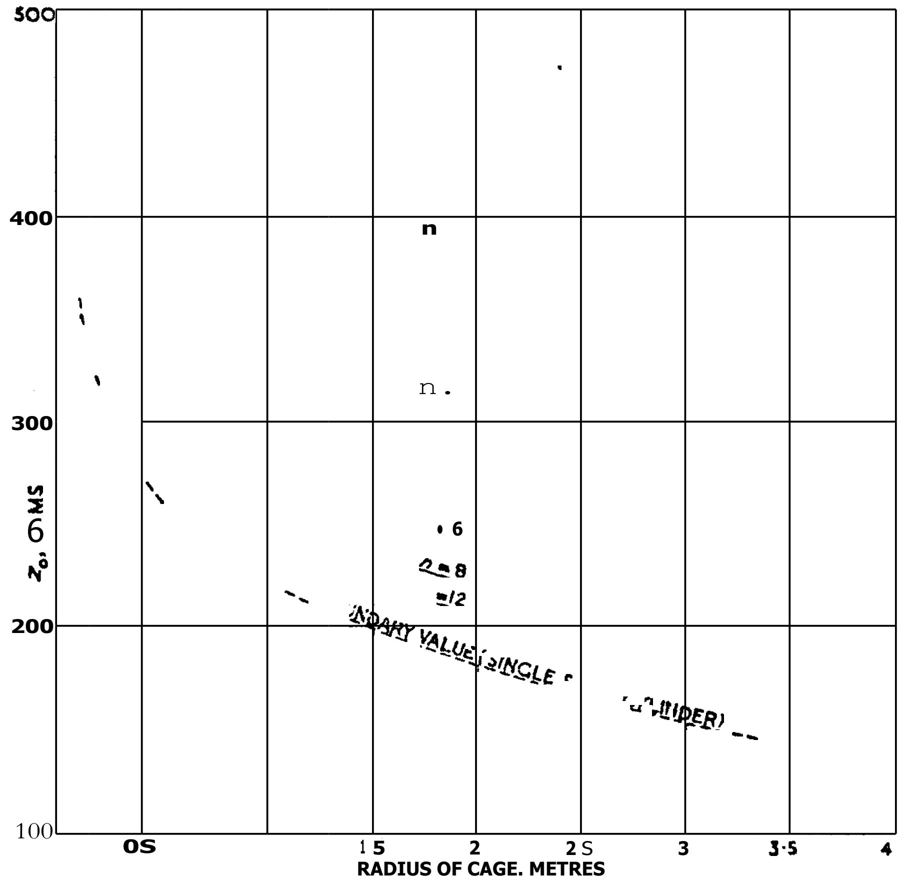


Fig. 2B.—Surge impedance of horizontal cage aerial at height of 20 metres above earth.

n = number of 1·0 mm (radius) conductors (No. 14 S.W.G.)

Note—Z_0 for single conductor at height of 20 metres = 635 Ω.

where n is the number of symmetrically disposed conductors comprising the cage, B the geo-mean distance between the n conductors, and A the geo-mean distance between the conductors of the cage and those forming its image. It can be shown that B . r. R^{n-1}), where r is the radius of the conductor, and R that of the cage, and if we assume that A is equal to twice the height h of the cage above earth the expression becomes

$$C = \frac{n}{2L}\left[\log_e \frac{2h}{(n\,.\,r\,.\,R^{n})}\right]^{-1} \text{e.s.u.}$$

Actually the formula given in Fig. 2 has been transposed into a slightly more convenient form for the purpose of calculation; also it is for $\mu\mu$F per metre of cage. Various considerations, including construction, weight and windage, lead to the conclusion that a 4-wire combination is the most practical form of high-capacitance cage: there are, of course, special cases when 6 wires or even 8 might be desirable. To take an example, a 4-wire cage with 1.0-m sides (0·707-m radius) yields a capacitance of about 10.5 $\mu\mu$F per metre run; this corresponds to a surge impedance of 320 ohms. If the inner ends of the cages are each given a slow taper, the progressive rate of change of capacitance between the arms will tend to be equalized.

Having decided upon the size and type of cage, the next step is the measurement of the variations of impedance at the terminals of an actual quadrant aerial; for this experiment twin square-form 4-wire cages having 1-m sides were employed, built of No. 14 wires measuring 15.25 m long, as depicted in Fig. 3A, the height above ground being about 15.0 m; connection was made to a high-frequency bridge via 16·5 m of 500-ohm twin feeder. Having measured the variations of impedance looking into the feeder, the variations at the aerial terminals can be computed by means of the circle diagram or some such artifice. In the present instance the resultant curves are shown in Fig. 3A, while Fig. 3a gives the results for a single-wire quadrant aerial of similar arm length; this latter set is added simply for the sake of comparison. Clearly, within the limits of efficient radiation the minimum variation of impedance will be around peak value, i.e. on either side of, and including, the natural frequency of the aerial. From an examination of Fig. 3A it appears that the natural frequency of the aerial under test is about 8.75 Mc/s, and if we take the octave between 5.75 and 11.5 Mc/s the variation in modulus impedance is seen to be a rise from a value of 320 ohms to 1 000 ohms, and subsequently a fall to 335 ohms; since experience has shown that the variation is entirely satisfactory this band-width will be adopted as standard for all ordinary requirements. Before accepting the limits of 5.75-11.5 Mc/s it is important to check that the equivalent arm length does not exceed 0·65λ at the higher frequency, as discussed in the previous Section: 11.5 Mc/s corresponds to 26.09 m, and

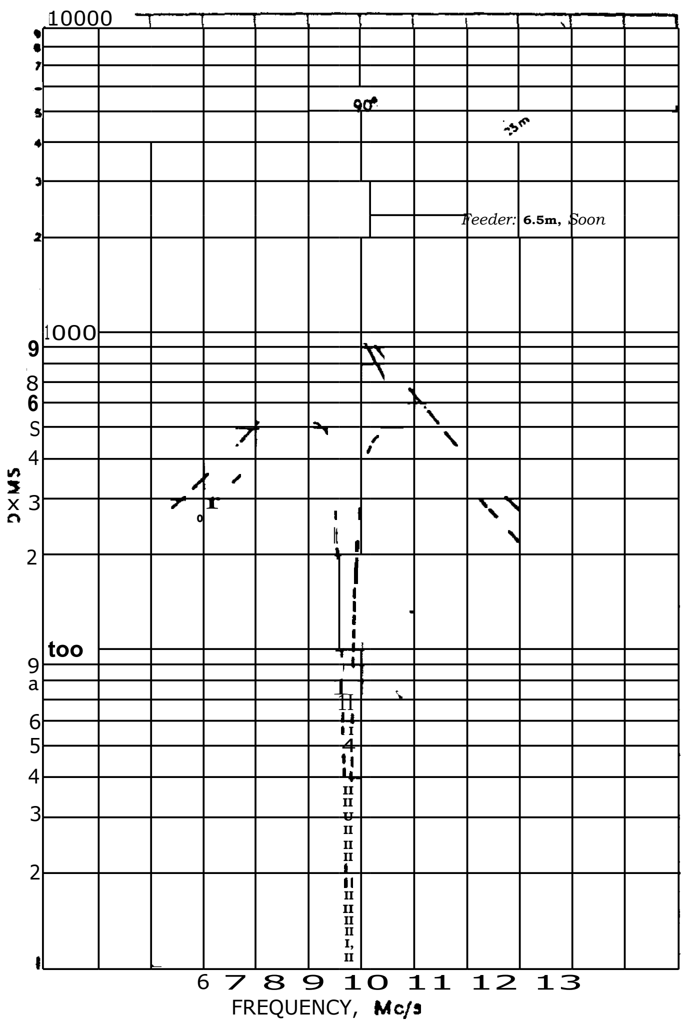


Fig. 3A

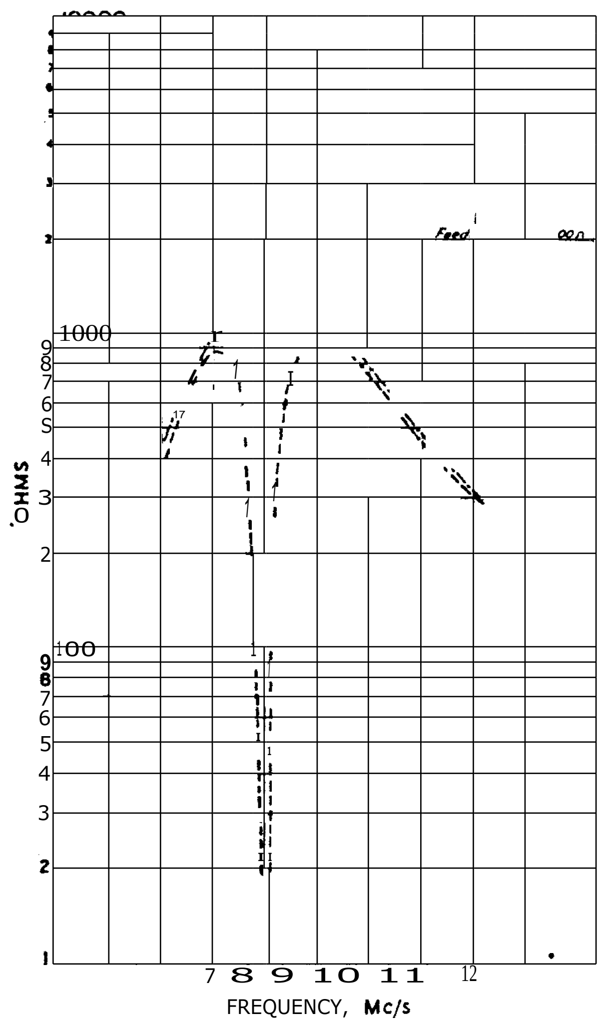


Fig. 3B

the arm length is 15.25 m, hence the equivalent in this case is 0.50 and is well within the prescribed limit.

We have now arrived at an overall working limitation of a 2 : 1 frequency band. Special requirements can be dealt with either by a narrower band-width, e.g. 6.5-11.0 Mc/s, or by enlarging the diameter of the aerial cage. On the other hand the octave may be safely exceeded for many classes of service where requirements are not stringent. Calculated polar diagrams covering the octave band are shown in Figs. 6(a), 6(b) and (6c).

Two other problems remain: the choice of an optimum value of fixed impedance to match the varying aerial terminal impedance over the prescribed frequency band; and the determination of the number, dimensions and grouping of quadrant aerials to cover a given total frequency band.

(5) TRANSFER IMPEDANCE

When aerial and receiver are connected via a length of twin balanced feeder the impedance of the latter must be chosen to give a good average match with the varying impedance across the aerial terminals. When the connection between aerial and receiver is via a length of concentric cable there will be a transformer at the junction of aerial and feeder; here it is the input impedance of this transformer that must be chosen to give the best possible match. Taking the resistance and reactance va over the given frequency band, from the curves of Fig. 3A can easily calculate the transfer of energy between the termi of the aerial and a known resistance across them (see Appe 11.5) and compare the result with the maximum possible tra fer; further, if this is done for a series of resistance values it be possible to determine an optimum value. The plot of s a series for the cage aerial under consideration is indicated Fig. 4, from which it will be seen that between the limits 5.75 and 11·5 Mc/s a value of 500 ohms exhibits a fairly unifo curve showing an average loss of slightly under 1·0 db, wl for the narrower frequency limits of 6.5 and 11·0 Mc/s value of 600 ohms would seem to be the best choice. Clea the actual figure is not critical.

When transformers are employed to couple aerial and fee the added loss due to this cause must be taken into accou under present war conditions the figure should be reckoned about 2.5 db when large quantities of transformers are concerne though under normal conditions of production it should be lo

Touching on the aspect of losses, it is perhaps well to recoll that for reception the degree of matching between aerial a feeder has no bearing upon feeder losses, but determines, in effe the coupling between the two, i.e. the amount of energy tran ferred from one to the other. The losses in the feeder a

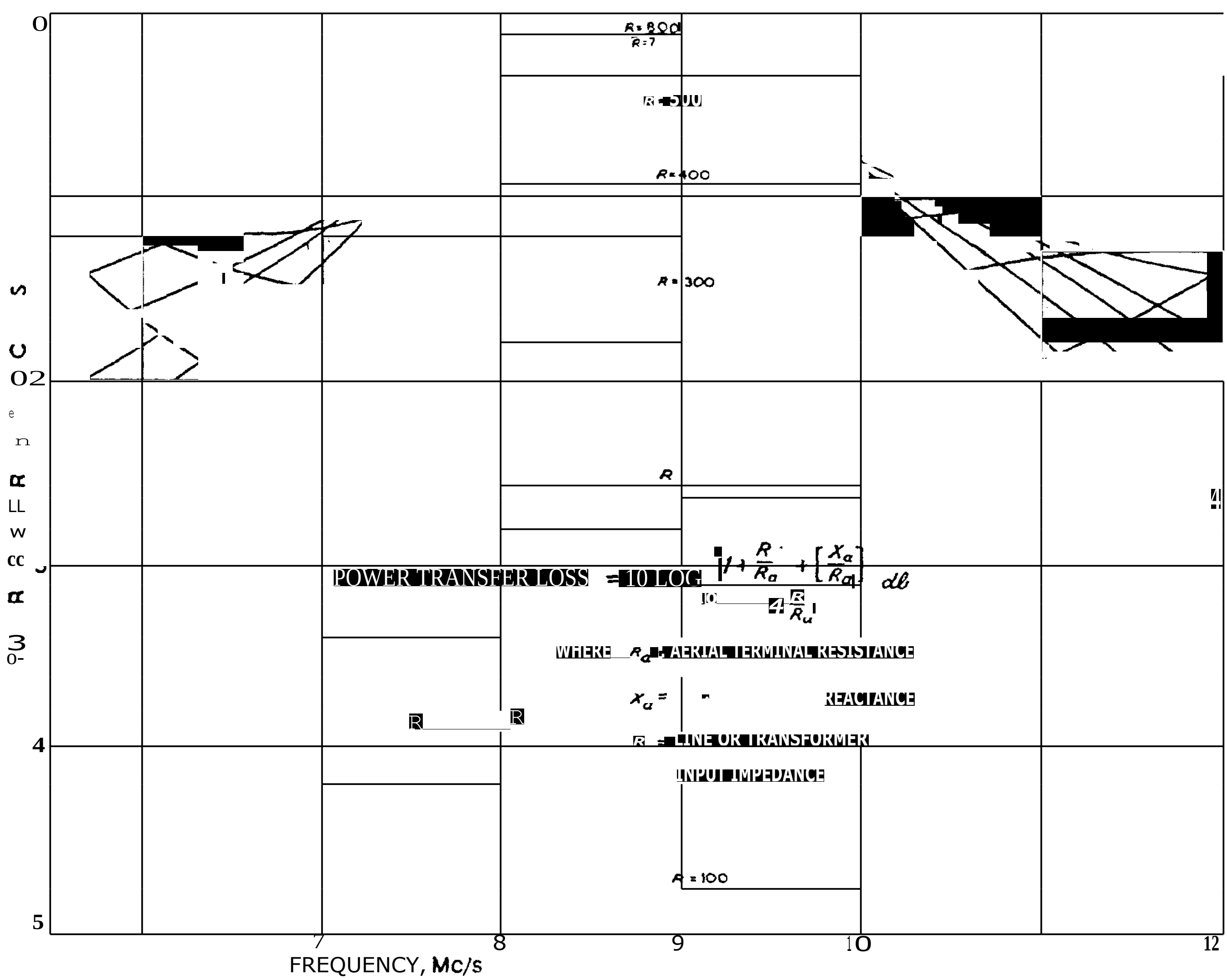


Fig. 4

entirely governed by correct or incorrect termination, at the receiver end in reception and at the aerial end in transmission.

(6) DIMENSIONS AND GROUPING

(6.1) Dimensions

If it is assumed that at mid-frequency the distribution along the arms of the quadrant aerial is equal to $\frac{1}{2}\lambda$, it is a simple matter to assess the mid-frequency for any given band, and to calculate the physical length on the basis of the known retardation; in the cage quadrant this is of the order of 10%, so that the physical length in metres becomes

$$0{\cdot}90 \times 0{\cdot}50\lambda_0 = 0{\cdot}45\lambda_0$$

But since the band is usually given in terms of its limits, the arm dimension may sometimes be more conveniently worked out on the basis of either the minimum or the maximum wavelength: for instance, in the case of the octave band there is a variation of 33A around mid-frequency; hence if the distribution at mid-frequency is assumed to be equal to $\frac{1}{2}\lambda_0$, the corresponding distribution at, say, λ_{min} will be $\frac{2}{3}\lambda_{min}$, and the physical length will be

$$0{\cdot}90 >: \quad = 0.60\lambda_{min}$$

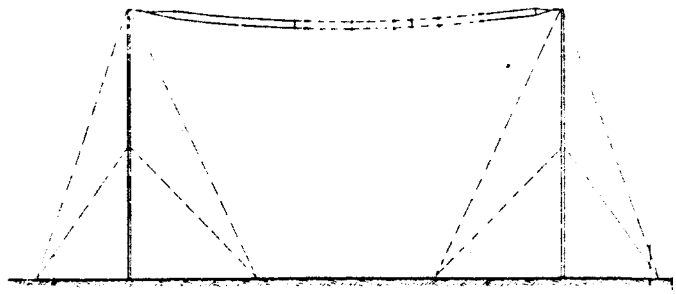
The factor $0{\cdot}45\lambda_0$ might be termed the dimension factor, and no apology is offered for reminding readers that λ_0 is not the mean wavelength but the wavelength corresponding to mean frequency. Other retardations will call for other factors.

It might be argued that the mid-frequency of 9·625 Mc/s, deduced from the limiting frequencies of 5.75 Mc/s and 115 Mc/s, does not quite coincide with the natural frequency of 8.75 Mc/s deduced from Fig. 3A, but this slight discrepancy is simply because the round figures of 5.75 Mc/s and 11.5 Mc/s were deliberately chosen, rather than more precise values.

(6.2) Grouping

A single quadrant aerial requires three poles to support it, and, in common with all aerials when distortion is to be avoided, it should be erected as far as practicable from other aerials, but since the use of more than one aerial is often desirable the problem of grouping presents itself; for instance reception may be required over a band-width of 3.0-20.0 Mc/s, say by three quadrants individually covering 3.0-6.0 Mc/s, 5·5–11·0 Mc/s and 10·0–20·0 Mc/s; or again there may be a requirement for 2.0-20.0 Mc/s covered by individual quadrants of 2·0–4·0 Mc/s, 3.5-7.0 Mc/s, 6.0-12.0 Mc/s and 10·0–20·0 Mc/s. It is clear that in these and similar cases the form of grouping presents some difficulties, but ultimately the problem has been overcome by evolving the four-square mast design indicated in Fig. 5B, a design which appears to be both economically and technically sound. In principle the idea is that the proximity of similar aerials can produce only symmetrical distortion, whatever the phasing and coupling; also, provided the latter is weak enough, as it can readily be in practice, the energy lost by absorption as ohmic losses is negligible. To check the actual effect a square rotating framework was set up with diagonals of twice the transmitted wavelength, a quadrant transmitting aerial being erected in one corner, and another *quadrant* aerial with limbs originally $1{\cdot}4\lambda$ long being erected in the opposite corner; this latter quadrant had its terminals shunted by means of a 600-ohm resistance. Tests were then carried out during which a series of polar radiation diagrams was plotted as the second quadrant aerial was cut down in stages from 1.4λ to $0{\cdot}15\lambda$.

The outcome of these tests showed that neither the distortion nor the absorption was serious, and that as the second quadrant arms were reduced, distortion increased slightly at first and then decreased slowly, the difference being negligible as the second



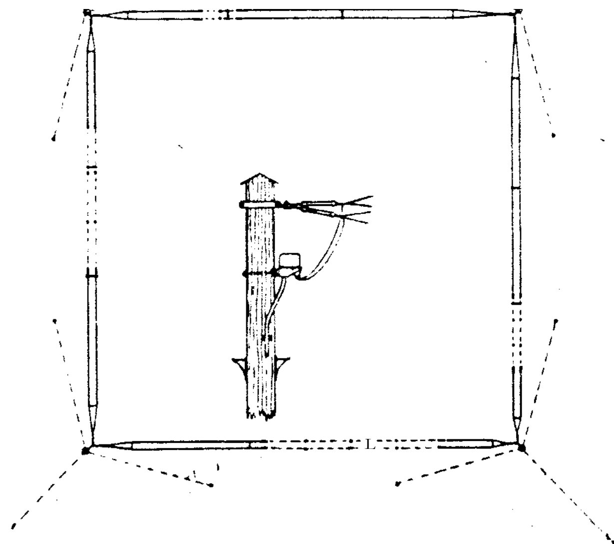


Fig. 5A.—Group of 4 quadrant aerials.

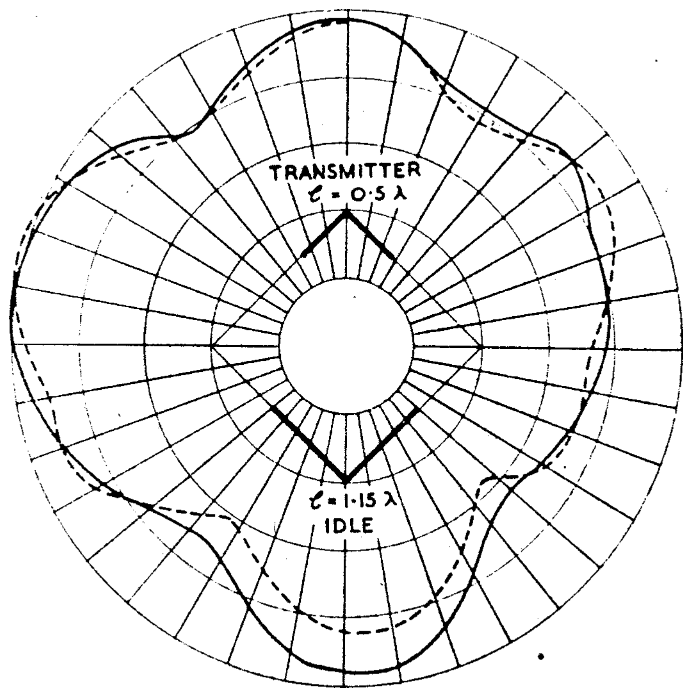


Fig. 5B.—Quadrant aerial polar diagrams.

quadrant approached and receded beyond the half wavelength. The greatest recorded departure from datum occurred when the second quadrant arm lengths were reduced to 1.15A, which does not appear to correspond to any particular distribution: the curve is shown in dotted line superimposed on the datum curve, in full line, in Fig. 5B, and it will be apparent that the difference is sufficiently small to justify the adoption of this form of grouping. The author is aware that these tests, carried out under war conditions, have a limited scope; nevertheless they appear to be reliable for most operational requirements.

(7) CALCULATED HORIZONTAL AND VERTICAL POLAR DIAGRAMS

Altogether six horizontal diagrams have been calculated by the method given in Appendix 11.1; they are shown in Figs. 6(*a*)–6(f), and cover arm lengths of $\frac{1}{4}\lambda$, $\frac{1}{2}\lambda$ and $\frac{3}{4}\lambda$; also +A, IA and 1A. In the calculations it has been assumed that the retardation is nil, whereas, owing to the comparative shortness of arms diverging at 90', the end-to-end change in impedance is rapid, and there is appreciable retardation on these aerials. This fact will not materially alter the shape of the diagrams for given distributions, but, as already mentioned, should be taken into account when assessing the actual distribution on an aerial of given physical dimensions. As can be deduced from Fig. 3a, the retardation with single conductors amounts to some 18.5%, whereas with the cage conductors depicted in Fig. 3A the change of impedance is not so rapid because of the taper towards the terminals, and retardation is reduced to 11·5%.

It will be appreciated that the above diagrams are calculated on the basis of free-space aerials, though it is considered that they afford a fairly accurate picture of the relative diagrams of radiation at various inclinations to the horizontal, corresponding, that is, to aerials at various heights above earth. In order to form a more objective picture of such radiation, five vertical polar diagrams have also been calculated for the mid-vertical plane of a quadrant aerial with +A distribution, these arc shown in Figs. *7(a)* to *7(e)* inclusive, and cover the respective cases of "free space" [Fig. *7(a)]*. and heights of 0·25λ, 0.50A, 1·0λ and 2.0A [Figs. *7(b)* to 7(e)] above a perfect earth. The relevant calculations are given in Appendix 11.2, but it must be apparent that such diagrams cannot differ appreciably from those appropriate to a horizontal dipole; in fact the difference between Fig. *7(a)*

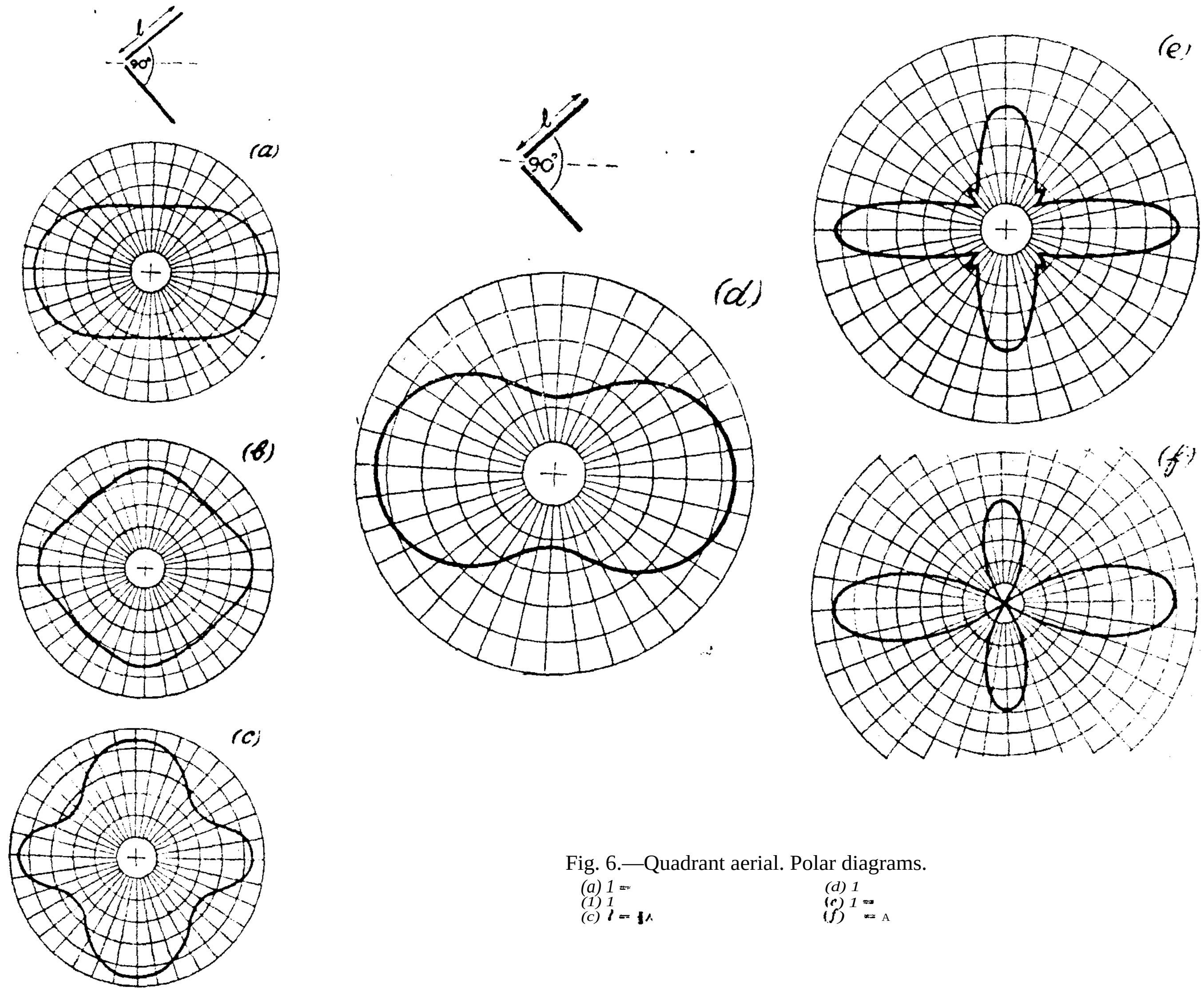


Fig. 6.—Quadrant aerial. Polar diagrams.
(*a*) 1 =
(1) 1
(*c*) *l* = λ
(*d*) 1
(*e*) 1 =
(*f*) = A

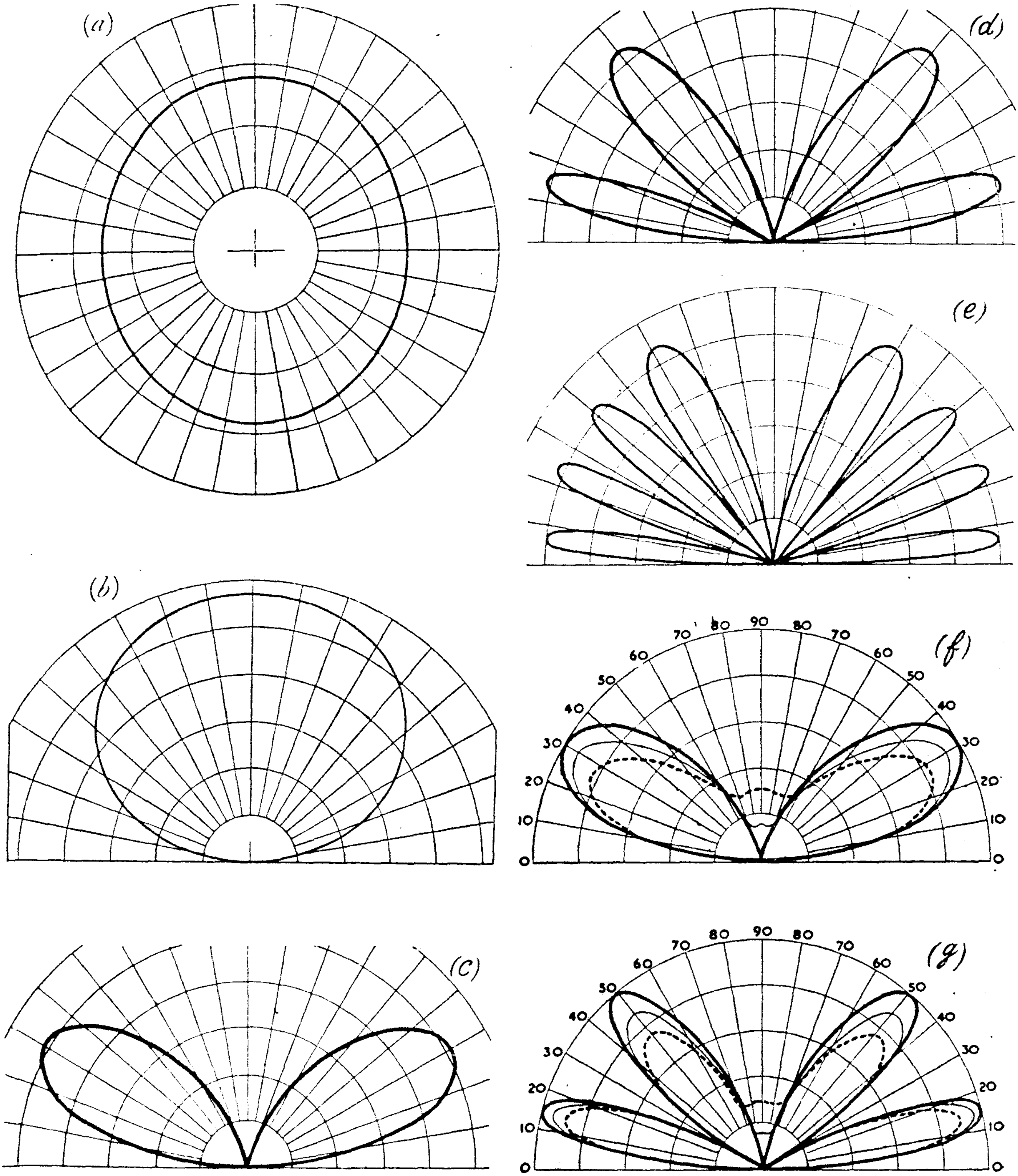


Fig. 7.---Quadrant aerials with $0 \cdot 5\lambda$ arms. *(a)* to *(e)* are vertical polar diagrams; *(f)* and *(g)* are for a horizontal dipole.

(a) Free space (b) h $0 \cdot 25\lambda$, (c) (d) (e) $h = 2 \cdot 0\lambda$ (f) $h = 0 \cdot 5\lambda$ (g) $h = X$

—— Perfect earth (o = x ; x = Good earth (o 10 e.m.u.; x 25 e.s.u.) ——— Poor earth ($\sigma = 10^{-14}$ e.m.u.; r. = 5 e.s.u.)

and a circle is a measure of the difference between these same diagrams and those appropriate to the equatorial plane of a simple horizontal dipole.

As regards the vertical polar diagrams of quadrant aerials with other than the $\frac{1}{2}\lambda$ distribution, careful calculations have shown that between $0 \cdot 25\lambda$ and 0.625A the change in the pattern is scarcely discernible, being perhaps 5% for $\frac{1}{4}\lambda$ arms and negligible for $\frac{5}{8}\lambda$ arms; thus it may be concluded that Figs. *7(a)* to *7(e),* and also Figs. 8(b) to 8(f) give *a* reasonably good picture of the distribution common to the octave frequency band.

Before concluding this Section we must consider the effect of physical earth: this is a problem that presents mathematical difficulties, but fortunately it is also one that can be resolved by deduction from the results worked out for the horizontal dipole. It has been seen that under similar conditions the two vertical polar diagrams are closely akin; also it is established that in the case of horizontal polarization the diagram for the physical earth approximates in shape to that for ideal earth, w absorption being seen in a progressive reduction of the diagra radii with increase of inclination. For these reasons we a justified in deducing the quadrant diagram by comparing it w the appropriate diagram of the horizontal dipole, this latter, c course, being comparatively easily worked out. For instanc the diagrams shown in Figs. 7(*f*) and *7(g)* for two heights horizontal dipole, each for two conditions of physical earth, ma reasonably be employed as proportionately applicable to the tw corresponding quadrant diagrams of Figs. *7(b)* and *7(c).*

(8) STACKED QUADRANTS

For spot-wave transmission, and to a less extent for spot-wave reception, the quadrant aerials may be stacked as indicated i Fig. 8(*a*), with the consequent advantage of greater control ove the vertical polar diagram. Figs. 8(h), 8(c) and 8(d) illustrat the vertical polar diagrams of twin stacked $\frac{1}{2}\lambda$ quadrants separate(

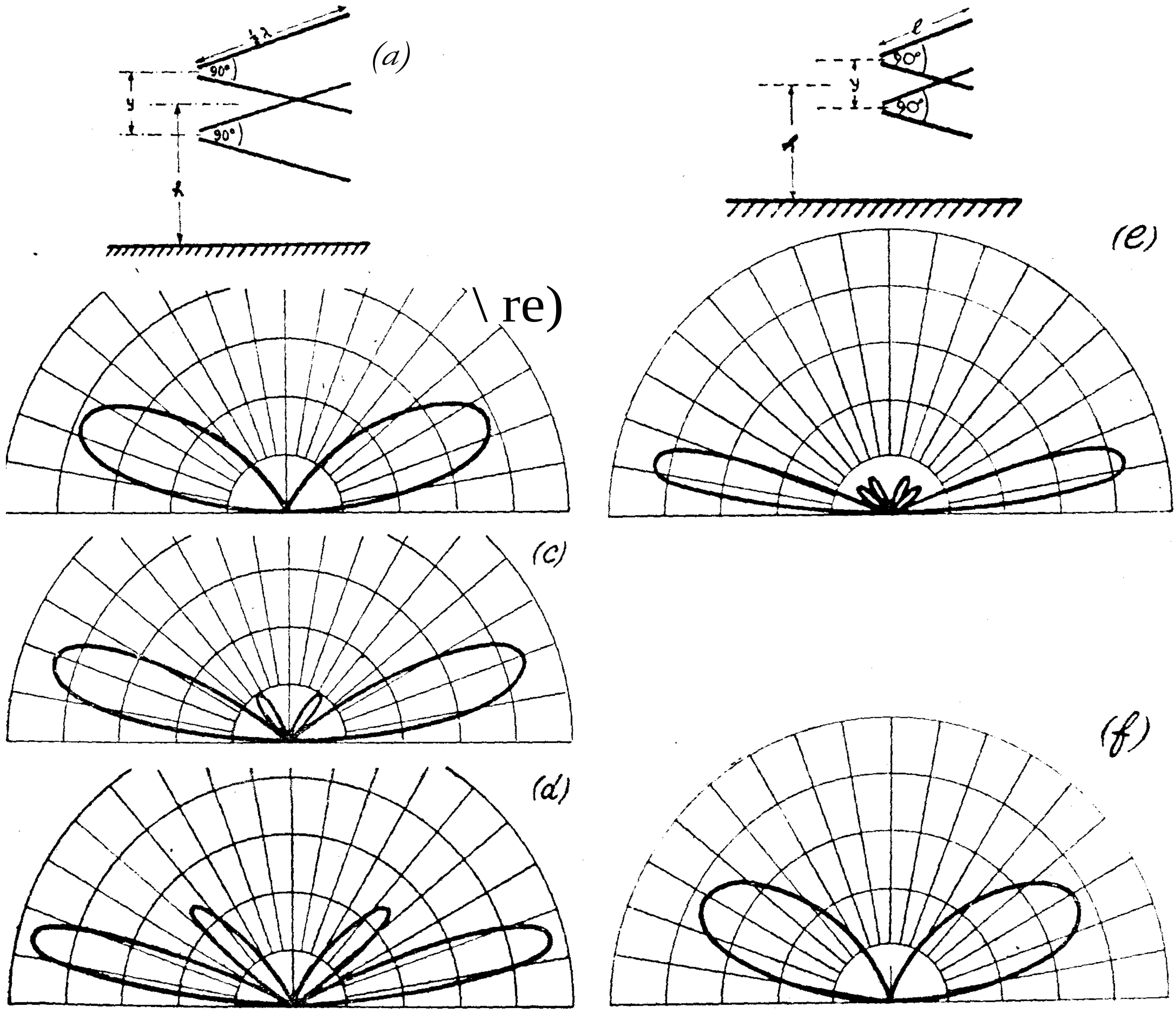


Fig. 8.—Stacked quadrant aerials. Vertical polar diagrams.

(b) *h* 0·5λ; overall height = 0·75λ; y = 0·5λ   *(r)* *h* *y*
(*c*) h = 0·75λ; overall height = *y* 0·5). 0·5). arms   *h* *y*
h = 1·0λ; overall height = 1·25λ; y 0.5λ.

0.50A and at the three heights above perfect earth indicated. The calculations are given in Appendix 11.3.

If the design of the feeders relative to power transmitted is such that standing waves are permissible, and also if re-adjustment of transmitter coupling with change of frequency is not an obstacle, it is possible to retain the major advantage of stacking while covering an octave (or smaller) frequency band on transmitting aerials; in this event the feed takes the form of a centre tap, mid-way between the tiers, on the interconnecting feeder. As an example we may take the case of an aerial covering the wave-band for which the limits are 20.0 mλ and 40·0 mλ; if on the shortest wavelength the tiers are designed to be $\frac{2}{3}\lambda$ apart, and at a height of 1.0A from mid-way between them to ground, it will be apparent that on the longest wavelength such dimensions become equivalent to $\frac{1}{3}\lambda$ between tiers, and $\frac{1}{2}\lambda$ height from ground. As regards arm lengths, if on 40 mλ the arms correspond to a distribution of 3A, then on 80 m the arms will correspond to a distribution of The two vertical polar diagrams are given in Figs. 8(a) and 81]).

(9) NOTE ON SIGNALS FROM A HORIZONTAL DIPOLE

In conclusion, the following note on the quasi-omni-directional property of the ordinary horizontal dipole may be of interest.

A moment's consideration of Fig. 9(a) reveals that in a horizontal dipole there is radiation through 360° of azimuth for all inclined rays, though at and near azimuth angles of $\psi = 0°$ and tit = 180', radiation may be small. Further, although a horizontal dipole is regarded as emitting horizontally polarized waves, it is obvious by symmetry from Fig. 9(a) that the field is only completely horizontally polarized in the equatorial plane, i.e. when ψ At any other orientation there is a component of vertical polarization the value of which increases with A, while for any one value of Δ it is also obvious that the relative proportion between vertical and horizontal components increases with decrease of towards the axis, reaching a maximum when $\psi = 0$ and *0* coincides with A. If in Fig. 91a) the length TA is taken as proportional to the total field at any given orientation, we can deduce the corresponding values of TC and AC, and thus obtain the proportional values of vertical and horizontal components: it only remains to calculate the precise value of TA in order to deduce the precise values of TC and AC, and this is done in Appendix (11.4).

As a typical example we may take the case of a horizontal $\frac{1}{2}\lambda$ dipole (I = 0.5A in the Figure) erected $\frac{1}{2}\lambda$ above earth, at which height the inclination of maximum radiation is 30'. Adopting this value for ι throughout, and taking distance as 1.0 km and loop current as 1.0.A, Table 1 shows the figures (in mV/m) for five values of azimuth orientation.

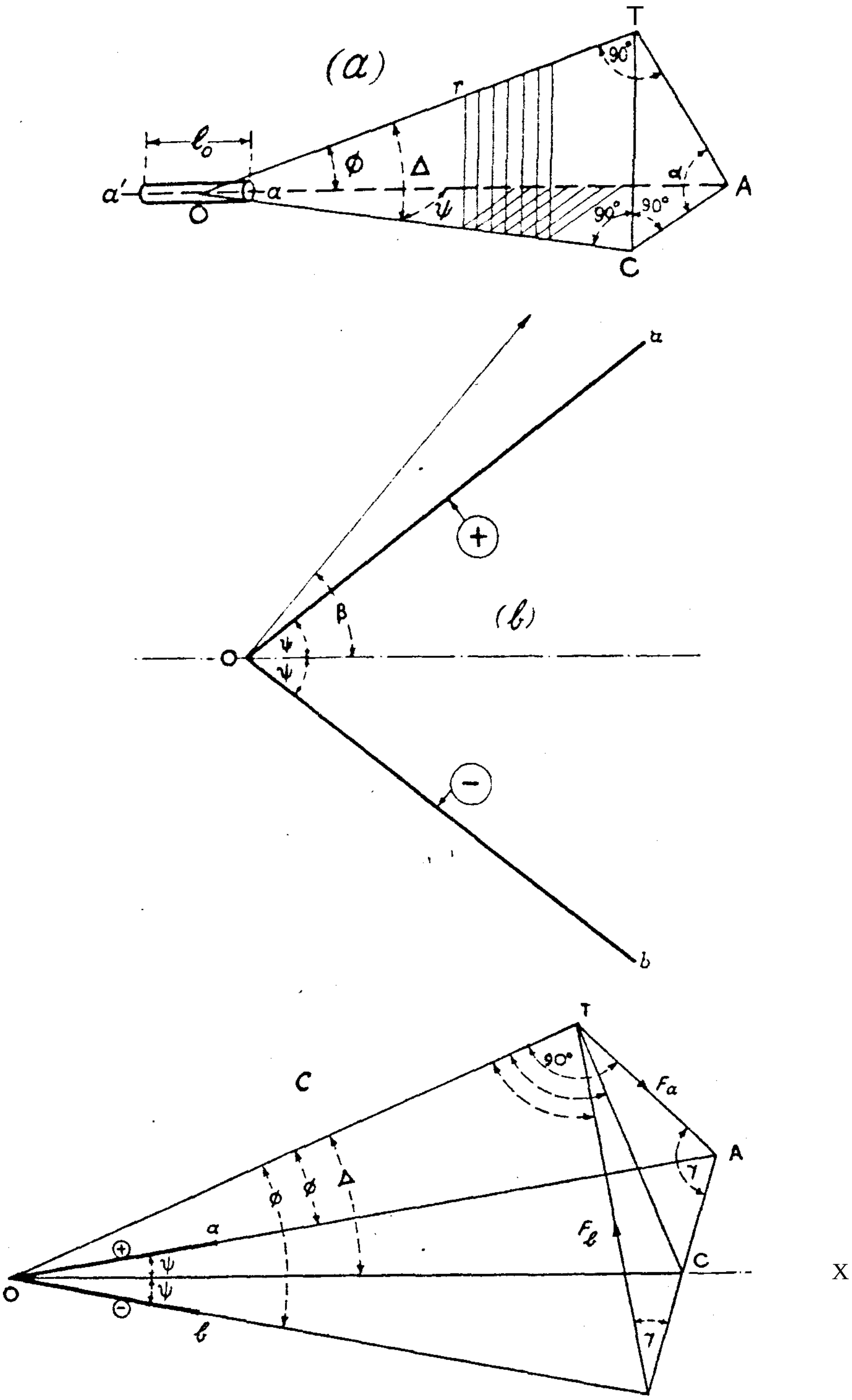


Fig. 9

(a) Single conductor. *(b)* and (c) Quadrant aerials.

In connection with long-distance communication, when effective rays are those at an inclination, if conditions are such that there is substantially no divergence from the great circle joining two otherwise non-related transmitting and receiving sites, the transmitting dipole being by chance in line with the receiving site, then the above figures explain why elliptically

Table 1. $\Delta = 30°$

ψ	Total strength (TA)	Vertical component (TC)	Horizontal component (AC)
90°	120	0	120
20°	59.5	41.8	42.3
10°	52.5	42.9	30
5°	51.0	43.4	27
0°	50.0	43.3	25

polarized signals (in the plane of the great circle) of moderate strength are transmitted towards the receiver: obviously when both dipoles chance to be in line and conditions fortuitously favour the particular elliptical polarization, received signals will be quite strong.

(10) ACKNOWLEDGMENTS

The author wishes to acknowledge his indebtedness to Messrs. R. W. Meadows and M. P. Quinlivan for their care in conducting the early experimental work, and to Mr. H. C. Cafferata for his patience in working out the calculations.

(11) APPENDICES

(11.1) **Calculation of Field Strength in the Plane of a Horizontal Quadrant Aerial**

List of symbols.

I_0 = loop current, in amperes.

r = distance from the source, in kilometres.

l_0 = angular length of conductor ($l_0 = 2\pi l/\lambda$, where l is in metres).

λ = wavelength, in metres.

h_0 = angular height of aerial above ground ($h_0 = 2\pi h/\lambda$ when h is in metres).

$\mathscr{E}$ = field strength in mV/m at distances large compared with λ.

Consider Fig. 9(a) as representing a length of conductor at 0 in free space; the general expression for the field strength at T a distance r from 0 due to standing waves on such a conductor is known to be

$$\mathscr{E} = \frac{30I_0}{r\sin\phi}\{\cos(l_0\cos\phi) - \cos l_0 + j[\sin(l_0\cos\phi) - \sin l_0\cos\phi]\} \quad (1)$$

where ϕ is the inclination of OT to the length of the conductor.

Next consider the horizontal V-aerial represented in Fig. 9(b), with an angle 2ψ at its apex, which becomes the quadrant aerial when $\psi = 45°$. If $\mathscr{E}_a$ and $\mathscr{E}_b$ are the field strengths in the horizontal plane at a distance r from the apex, and at an azimuth β to the mid-axis, $\mathscr{E}_a$ and $\mathscr{E}_b$ are given by expressions of the form in (1) where ϕ has the values ϕ_a and ϕ_b respectively. As the arms Oa and Ob are fed in anti-phase, the resultant field $\mathscr{E}_r$ will be given by the vector difference of $\mathscr{E}_a$ and $\mathscr{E}_b$, so that the modulus $|\mathscr{E}_r|$ is given by

$$|\mathscr{E}_r| = \sqrt{|\mathscr{E}_a|^2 + |\mathscr{E}_b|^2 - 2|\mathscr{E}_a||\mathscr{E}_b|\cos(\alpha_a - \alpha_b)} \quad (2)$$

where the phase angle

$$\alpha_a = \arctan\left[\frac{\sin(l_0\cos\phi_a) - \sin l_0\cos\phi_a}{\cos(l_0\cos\phi_a) - \cos l_0}\right] \quad (3)$$

the phase angle α_b being similar with respect to ϕ_b.

Taking as a particular case the quadrant aerial with arms each $\frac{1}{4}\lambda$ long, i.e. $l_0 = \pi/2$, and a loop current of 1·0 amp, the field at a distance of 1 km has been calculated for various values of β, the results being given in Table 2.

Table 2

Horizontal Polar-Diagram Values for Quadrant Aerial with $\frac{1}{4}\lambda$ Arms

β	$\|\mathscr{E}_a\|$	$\|\mathscr{E}_b\|$	α_a	α_b	$\alpha_a - \alpha_b$	$\mathscr{E}_r$
0°	37.5	37.5	+ 75.3°	− 104·8°	+ 180.1°	75.0
10°	30·84	43.5	+ 60·3°	− 86.3°	+ 146.6°	71·7
20°	29.2	47.4	+ 48·7°	− 64.4°	+ 113.3°	61.5
30°	15.15	50.1	+ 41.3°	− 40·0°	+ 81.3°	50.1
40°	5.265	51.0	+ 36·4°	− 13.5°	+ 49.9°	47.94
50°	5.265	51.0	− 146·6°	+ 13.5°	− 157.1°	56.1
60°	15.15	50.1	− 138·7°	+ 40.0°	− 178·7°	65·7
70°	29.2	47.4	− 131.3°	+ 64.4°	− 195.7°	69.3
80°	30.84	43.5	− 119.7°	+ 86.3°	− 206.0°	72.6
90°	37.5	37.5	− 104.75°	+ 104·8°	− 209·55°	72·6

When $l_0 = \frac{1}{2}\lambda$ the expressions for the moduli of the component field strengths $|\mathscr{E}_a|$ and $|\mathscr{E}_b|$ simplify considerably, becoming

$$|\mathscr{E}_a| = \frac{60I_0}{r}\frac{\cos(\frac{1}{2}\pi\cos\phi_a)}{\sin\phi_a} \quad . \quad . \quad (4)$$

with a phase angle $\alpha_a = \frac{1}{2}\pi\cos\phi_a$

while the expression for conductor b is similar.

(11.2) **The Vertical Polar Diagram of the Quadrant Aerial**

In Fig. 9(c), OX is the mid-axis of the V-aerial with its origin at O, and T is a point in the vertical plane through OX distant r from O at an elevation Δ. The field at T due to the limb Oa acts along TA, where TA is at right angles to OT, and A lies in the horizontal plane along Oa produced. Similarly, the field due to the limb Ob acts along TB, being reversed in sense with respect to T because the limbs are fed in anti-phase, and by symmetry we have $|\mathscr{E}_a| = |\mathscr{E}_b|$.

It is obvious by symmetry that the horizontal line AB is at right angles to OX and that its mid-point lies on OX. As OT is perpendicular to both TA and TB, it is also perpendicular to the line TC in the plane TAB, i.e. the angle OTC is a right angle. We can see from the Figure that as $|\mathscr{E}_a|$ and $|\mathscr{E}_b|$ are equal, the resultant field at T is parallel to BA and equal to $2\mathscr{E}_a\cos\gamma$, being wholly horizontal and transverse to the mid-vertical plane.

$$\text{In the Figure}\cos\gamma = \frac{BC}{BT} = \frac{OB\sin\psi}{OB\sin\phi} = \frac{\sin\psi}{\sin\phi}$$

As we wish to express the resultant in terms of the angle of elevation Δ rather than ϕ, we note that

$$\cos\Delta = \frac{OT}{OC} = \frac{OB\cos\phi}{OB\cos\psi} = \frac{\cos\phi}{\cos\psi}$$

so that $\cos\phi = \cos\Delta\,.\cos\psi$

and $\sin\phi = \sqrt{(1 - \cos^2\Delta\cos^2\psi)}$

$$\text{Thus}\quad\cos\gamma = \frac{\sin\psi}{\sqrt{(1 - \cos^2\Delta\cos^2\psi)}}$$

and using this value of $\cos\gamma$ in the value $2\mathscr{E}_a\cos\gamma$ of the field at T, and our values of $\sin\phi$ and $\cos\phi$ in the value for $\mathscr{E}_a$ given by (1) we derive for the field at T

$$\frac{60I_0}{r}\cdot\frac{\sin\psi}{1 - \cos^2\Delta\cos^2\psi}[\cos(l_0\cos\Delta\cos\psi) - \cos l_0 + j(\sin(l_0\cos\Delta\cos\psi) - \cos\Delta\cos\psi\sin l_0)] \quad (5)$$

When l_0 is an integral number of half wavelengths, i.e. $l_0 =$ *where n is a positive integer,* the scalar value of the field becomes

$$\frac{120 I_0}{r} \cdot \frac{\sin}{(1 - \cos\text{-} a \cos' \psi)} \sin [(\tfrac{1}{2} n\pi)(1 + \cos A \cos \psi)] \quad (6)$$

When a is unity equation (6) reduces to

$$\mathscr{E} = \frac{120 I_0}{r} \cdot \frac{\cos (\tfrac{1}{2}\pi \cos A \cos \psi)}{(1 - \cos^2 A \cos^2 \psi)} \sin \psi \quad (7)$$

For the quadrant aerial $_b$ is 45°, so that for a quadrant aerial with ?.A arms

$$\frac{120/_0}{r} \cdot \frac{\cos [(\pi/2 \ /2) \cos A]}{\sqrt{2}(1 - \cos^2 A)} \quad (8)$$

For a $\tfrac{1}{4}\lambda$ quadrant aerial above perfect earth the introduction of the height factor 2 sin (h_0 sin A) changes the expression into the form

$$\frac{240 I_0}{r} \cos \frac{[(\pi/2 \sqrt{2}) \cos A]}{\sqrt{2}(1 - 4 \cos^2 A)} \sin (h_0 \sin A) \quad (9)$$

(11.3) Calculation of the Vertical Polar Diagram of the Stacked Quadrant Aerial

The free-space polar diagram for the mid-vertical plane of a quadrant aerial with $\tfrac{1}{4}\lambda$ arms is given in scalar form by equation (8). Call this value $\mathscr{E}'$, then for two stacked quadrant aerials separated by a distance of y_0 radians the field at an elevation A with reference to the field that would be produced by a single quadrant aerial placed mid-way between them may be written

$$'[e^{j(y_0/2)\sin A} \pm e^{j(y_0/2)\sin} \quad 2F' \cos (\tfrac{1}{2} y_0 \sin A) \quad (10)$$

When the mid-plane of the stacked aerials is at a height h_0 above a perfectly conducting earth, the field if with reference to the field *if,* that would be produced by the stacked aerials if their mid-plane were placed at ground level and the earth were removed, may be written

$$\mathscr{E} = \mathscr{E}_s (e^{jh_0 \sin} - e^{jh_0 \sin} \quad 2j\mathscr{E}_s \sin (h_0 \sin A)$$

The factor) merely represents a phase angle of in-, so that in scalar form we may write the field $\mathscr{E}$, using (10) above, as

$$\mathscr{E} = 4\mathscr{E}' \cos (\tfrac{1}{2} y_0 \sin A) \sin (h_0 \sin A) \quad (11)$$

(11.4) The Horizontal and Vertical Component Field Strengths Due to Horizontal Dipole above Earth

For a IA dipole in free space the scalar field strength (in mV/m) is known to be

$$\frac{60/ \cos iv \cos \phi}{r \sin 0}$$

where ϕ is the angle of direction relative to the conductor axis [as depicted in Fig. 9(a)), r is in kilometres, and I_0 is in amperes. It is apparent from this equation that the locus of $\mathscr{E}$ is a solid figure of revolution about the axis, with a circular diagram in the (vertical) equatorial plane. When the dipole is above perfect earth the latter is regarded as a horizontal plane with a perfect reflecting surface; hence the force represented by the lower half of the solid figure will be reflected skywards, and the resulting figure will no longer be symmetrical about the conductor axis. The actual effect of reflection is indicated as regards vertical polar diagrams by the height factor 2 sin (h_0 sin A), where A is the inclination in the vertical plane. For the purpose of analysis it is obviously more convenient to determine the locus of X by means of the inclination A and the azimuth angle ψ, and in Appendix 11.2 expressions for replacing ϕ by A and ψ are given; adopting these expressions, and assuming unity values for r and 4, the above equation becomes

$$\mathscr{E} = 60 \frac{\cos (\pi \cos \Delta \cos 0)}{\sqrt{(1 - \cos^2 A \cos'}} \cdot 2 \sin (h_o \sin A)$$

In the example discussed in the text the angular height of the dipole is π radians, this being the optimum height for the given inclination of 30', so that the height factor has its maximum value of 2. We have already noted that when 0 = 0' and 180' the angle 0 coincides with A.

This is the total force, and the present requirement is to find expressions for its horizontal and vertical component forces. Consider Fig. 9(a), in which a′a is a horizontal conductor with mid-point at 0, and A lies in the horizontal plane along a′a produced: TA is at right angles to OT, and TC is normal to the horizontal plane, indicated by OCA, while points T, C and A lie in a vertical plane. If TA is proportional to $\mathscr{E}$ it is clear from the Figure that CA and CT are proportional to the required component forces; thus the problem is to find expressions for CA and CT, which we shall do in terms of TA and the angle a between TA and AC. The angles 0, A and 0 have already been defined and are as indicated.

It is seen that $\quad TC = OT \sin A = TA \dfrac{\sin \phi \sin A}{\cos \phi}$

Adopting the values for sin ϕ and cos ϕ already found in Appendix 11.2 we have

$$TC = TA \frac{\cos A \cos b \sin A}{\sqrt{(1 - \cos^2 A \cos^t 0)}}$$

and, since TC = TA sin a, we are in position to express the required vertical component in terms of $\mathscr{E}$ as given in equation (12), and sin a as given above. Thus

Vertical force = X sin a

where $\quad \arcsin \dfrac{\cos A \cos 0 \sin A}{\sqrt{(1 - \cos^2 A \cos^2 0)}}$

Having found a we may now write

Horizontal force = $\mathscr{E}$ cos a

(11.5) **Transfer Loss Due to Mismatching**

The circuit schematic is shown in Fig. 4, from which it is apparent that

$$\text{Current through R} = \frac{}{\sqrt{[(R_a \quad R)^2}}$$

$$\text{Power absorbed by R} = \frac{E^2 R}{(R_a \quad R)^2 \quad X_a^2}$$

Optimum power is drawn from the aerial when $R = R_a$, and X_a is neutralized. Hence

$$\text{Optimum power absorbed} = \frac{E^2}{4R_a}$$

$$\text{Loss} = 10 \log \frac{\text{Optimum power}}{\text{Actual power}}$$

$$10 \log \frac{(R_a \ R)^2 \quad X_a^2}{4R_a R}$$

Chapter 3

FEEDERS AND MATCHING

LIST OF CONTENTS

LIST OF ILLUSTRATIONS

Introduction

I. The source of the radio-frequency currents fed to the various types of antenna described in Chapter 2 was referred to as a *generator.* In practice, of course, the *generator* is actually the output stage of the radio transmitter and, although in some circumstances, (usually in portable or mobile transmitters used with vertical antennae) it is possible to connect the antenna directly to the transmitter, more frequently the antenna is necessarily separated from the transmitter for the following reasons:—

(1) Some antennae are erected at heights well above ground level.

(2) The building housing the transmitter, if in close proximity to the antenna, may act as an obstruction and effect the polar diagram.

(3) In communication centres where direction of transmission and reception and also variation of frequency are essential to maintain consistent communications, different types of antenna may necessarily be required for use at various times of day and night by the same transmitter. These antennae must he adequately spaced and therefore must he located at some distance from the transmitter or receiver.

2. The connecting link between the transmitter or receiver and the distant antennae is the r.f. feeder or transmission line, the sole purpose of which is to carry r.f. power from one point to another as efficiently as possible to ensure that the r.f. output from the transmitter is fed to the antenna terminals with minimum loss. In reception, of course, the function of the line is to transfer the minute current produced in the antenna by the travelling waves propagated by a distant transmitter to the receiver input terminals, again with minimum loss.

3. Losses in low frequency (50-60 Hz) power lines may be minimized by adequate insulation and reduction of line resistance by selection of conductor size sufficient for the load. At radio frequencies, in addition to the resistance of conductors and dielectric losses, each conductor having appreciable length comparable with the wave length in use, will radiate power and become an antenna. Unless special care is taken to minimize this radiation the power loss may considerably exceed the resistance losses.

4. Power loss in resistance is inherent to some extent, but loss by radiation is generally avoidable by using two conductors, so arranged that the electromagnetic field of one conductor is completely balanced by an equal and opposing field in the other. The resultant space field then reduces almost to zero and radiation is largely prevented. This is achieved in practice by placing two parallel conductors a very small fraction of a wave length apart, for the frequencies involved, and insulating one from the other. The feeder may be either balanced or unbalanced with respect to earth (para. 51).

5. The configuration of the two conductors may vary in different types of feeder, examples being the open-wire, coaxial, screened pair and single wire with earth return. The open-wire and coaxial feeders are types in common use and are considered in more detail in this chapter.

Open-wire feeders

6. The open-wire transmission line or feeder consists of two identical parallel wires spaced a very small fraction of a wave-length apart and usually supported about ten feet above ground at intervals between 70 and 100 feet on insulated poles (fig. la). In Chapter 2 on the subject of travelling waves, the effect of a voltage applied to one end of a wire on the resultant travelling wave was discussed. If one end of an open-wire transmission line is connected to a generator balanced with respect to earth, the voltage applied to one wire is equal to the voltage applied to the other wire but opposite to it in phase. The travelling waves on the two wires are also equal but opposite in phase (fig. 2a). If the wires arc very close together, the electromagnetic waves are in anti-phase throughout the length of the line and in all directions around it, thus reducing radiation from the line virtually to zero. For complete absence of radiation the distance between the wires should actually be zero. but spacing of a small fraction of a wave length of the frequency in use results in almost complete cancellation.

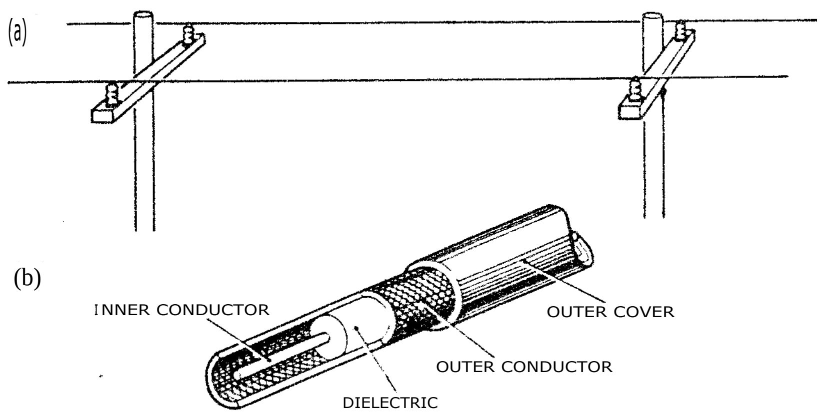


Fig. 1 Common types of feeder

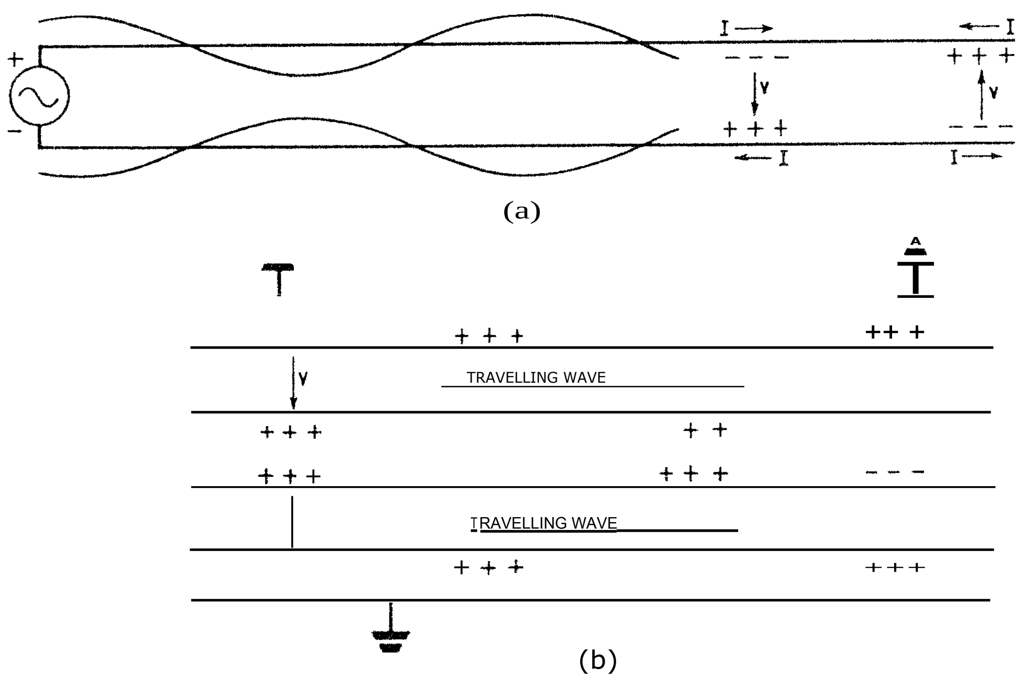


Fig. 2 Travelling waves on a feeder

Coaxial feeders

7. The coaxial transmission line (fig. 1b) consists of an inner wire conductor concentrically encased within a cylindrical outer conductor with the spacing maintained by the use of insulating washers at frequent intervals or by completely filling the space between the conductors with a low-loss insulating material termed the dielectric. The outer conductor is usually constructed of metal braiding for mechanical flexibility and is weather-proofed by a plastic outer covering.

8. The coaxial line functions in a similar way to the open-wire feeder with the travelling waves on the two conductors being equal but opposite in phase. In addition, due to skin effect, the wave on the outer conductor travels along its inner surface, the outer surface acting as a screen to radiation from feeder conductors. Because the wave does not penetrate to the outer surface of the coaxial line, the outer conductor may be earthed.

Characteristic impedance

9. Each of the conductors of a transmission line has a characteristic impedance, the value of which is affected by the proximity of the other conductor, but it is more convenient to consider the characteristic impedance of the line itself. The characteristic impedance of a transmission line or feeder is defined as the ratio of the voltage between the conductors at any point to the current flowing in either conductor at that same point, when no reflection occurs at the far end of the line.

10. With an open-wire line, the voltage between the wires at any point on the line is twice the voltage in each wire, since the generator is balanced with respect to earth. The current has the same magnitude (though directionally opposed) in either wire, provided that the line itself is balanced i.e. provided the total impedance (due to capacitance etc.) between wire and earth is the same for each conductor.

11. The characteristic impedance 4 of a pair of parallel wires separated by air is given by the formula,

$$4 = 276 \log_{10} \left(\frac{2S}{d} \right)$$

where S is the centre-to-centre spacing of the wires and d is the diameter of either wire.

Note . . .

The dimensions S and d may be measured in any convenient unit provided the same unit is used for both. This formula should only be used where d is small in terms of S.

The values of the characteristic impedance for various spacings of different sizes of hard drawn copper wire and the relationship between frequency and attenuation for sizes of wire commonly used in open-wire feeders are given in fig. 17 and 18 respectively.

12. The characteristic impedance of a coaxial feeder depends on the diameters of the inner conductor and inner surface of the outer conductor together with the dielectric constant of the material between the conductors. The characteristic impedance can be found from the following formula when the dielectric constant is known:

$$4 = \frac{138}{\sqrt{K}} \log_{10}$$

where D is the inside diameter of the outer conductor
d is the external diameter of the inner conductor
K is the dielectric constant of the dielectric.

Losses

13. For efficient operation, a feeder must deliver all the energy fed in at the transmitter end to the load or antenna at the other end. The energy actually delivered depends on the correct matching of the line to the source and load and the requirement that the line shall be balanced (para. 51). If incorrectly matched, some of the energy is reflected back from the load to the transmitter and if the line is not balanced, unequal currents flow in two conductors, the electromagnetic waves radiated from the conductors do not cancel out and energy is lost through radiation.

14. Even if the feeder is correctly matched and balanced, some energy is lost during its passage along the line for the following reasons:—

(1) The normal resistance of the conductors causes energy dissipation by heat (I^2R or ohmic loss).

(2) Energy dissipation by heat in the dielectric material and insulators, particularly in coaxial cables (dielectric loss).

(3) Leakage of energy due to the capacity between the line and earth and adjacent earthed objects such as open-wire supports.

(4) Radiation loss of energy due to residual radiation effects, mainly in unscreened lines.

Generally the total losses due to such effects are similar in both open-wire and coaxial feeders of the same length. Open-wire feeders however, are usually cheaper and also introduce less ohmic loss in long lines than coaxial cable and for those reasons are normally used in ground installations where they can also be mounted on poles beyond the reach of accidental contact.

15. At the higher frequencies, where wavelengths are short, the distance required between the conductors of an open-wire feeder to avoid radiation loss becomes too short to be practicable and some form of screened feeder must be substituted. This may consist of parallel-wire feeders enclosed by an earthed screen but coaxial cable is more commonly used.

Matching

16. When a feeder is terminated by a load anntenna) with an impedance equal to the characteristic impedance of the line, the energy travelling along the line from the transmitter is completely absorbed by the load and none is reflected back to the transmitter. if the load impedance does not match that of the line some of the energy is fed back along the line in the form of a reflected wave and the combination of the advancing and reflected waves is a standing wave. The presence of standing waves in a feeder, therefore, indicates that the line is incorrectly matched to the load and that the transfer of energy is inefficient.

Standing Waves

17. The presence of standing waves in an unscreened feeder can readily be detected by comparison of the magnitude of the r.f. current at various points along the line (fig. 3). A relative measure of the r.f. current can be obtained by connecting a hot-wire or thermocouple type of ▪ mmeter to the ends of a short straight wire mounted on a long insulated handle and by placing the wire close to one of the feeder wires. If the feeder is correctly matched to its load, the only wave along the line is the advancing travelling wave and the meter should give similar readings at any point along the line provided that the test wire is held at the same distance from the feeder wire for each reading. The presence of standing waves will give rise to r.f. current variations which will show as maximum and minimum values as the meter is moved along the feeder. The amount of variation indicates the extent of the mis-match and is usually expressed as a ratio of the maximum and minimum currents indicated by the meter.

i.e. standing wave ratio (s. $\frac{I\ (max)}{I\ (min)}$

Note . . .

Where the feeder and load are properly matched there is no standing wave: hence $I_{(max)} = I_{(min)}$ and s.w.r. = I : 1

18. If there is a slight mis-match the reflected wave will be small compared with the incident wave and the resultant s.w.r. will be small.

e.g. if $I_{(max)}$ 1.0A
and $I_{(min)}$ = 0·8A
s.w.r. = 1.25: I

Hence, as the degree of mis-match increases, so the standing wave ratio becomes greater. The extreme condition occurs when the load end of the feeder is either short or open-circuited, when I (min) becomes very small and the standing wave ratio very large (para. 23).

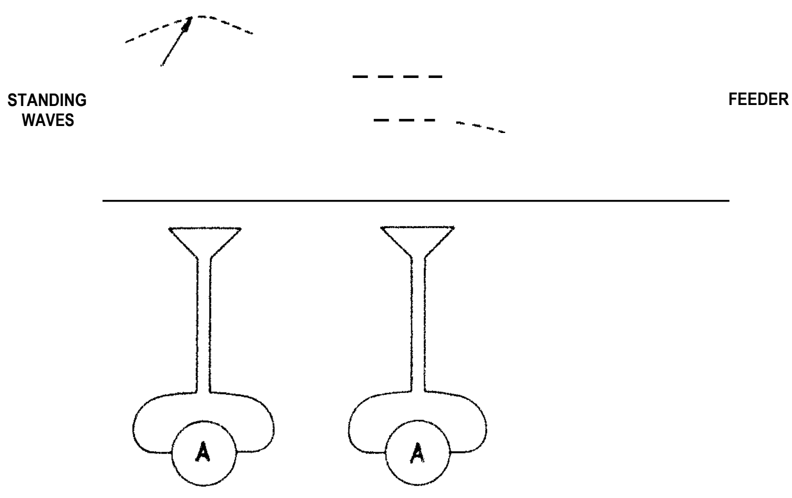


Fig. 3 Measuring standing waves

Methods of matching

19. Where possible a feeder is selected so that its characteristic impedance is comparable to the impedance, at its operating frequency, of the antenna it is required to feed, matching being facilitated by the use of broad-band antennae with reasonably level values of impedance at their feed points over the frequency ranges for which they are designed. Where the impossibility of direct matching of transmission line to antenna feed point is imposed by system design considerations, an impedance matching device between the feeder and antenna becomes necessary. This also applies in the case of matching an unbalanced feeder to a balanced antenna (para. 51) or vice versa. Matching devices may be grouped as follows:

(1) Conventional r.f. transformers which may also include other components such as capacitors. The increased usage of wide-band antenna systems and the development of air-transportable radio installations involve the extensive use of ferrite-cored wide-band transformers both for impedance matching and conversion from balance to unbalance (balun) functions.

(2) Matching stubs, which consist of a short length of feeder terminated in either an open or short circuit and connected to the main feeder at a selected point.

(3) A short length of feeder of suitable dimensions connected between main feeder and load, which functions in a mariner similar to a conventional transformer. The two conductors comprising the short length of feeder may be either parallel (Q-bars) or divergent ('V-match).

(4) The Windum antenna method (para. 45) using a single-wire feeder. Owing to feeder radiation this method is of limited application.

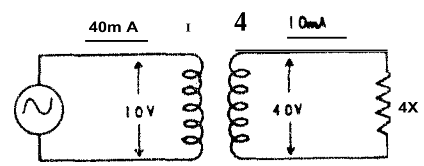
Matching Units

20. The r.f. transformer type of matching unit works on the same principle as its counterpart in general electronic circuitry. The basic example of this principle is presented by an ideal transformer in which the coupling between primary and secondary windings is 100 per cent, ignoring impedance effects. If the secondary winding is wound with four times the number of turns as the primary and a 4,000 ohm resistor is connected across the secondary (fig. 4a), with 10 volts applied to the primary, the voltage measured across the secondary will be 40 and the current in the resistor, 10 mA. The power dissipated in the resistor will be 400 mW, but as this power must have its source in the primary, the current flowing in the primary circuit must be 40 mA. With an applied voltage of 10 and a current of 40 mA, the resistance of the primary circuit must be $\frac{10}{0.04}$ = 250 ohms, or one sixteenth of the value of the secondary circuit resistance. Such a transformer could be used to match a 4,000 ohm load to a 250 ohm feeder or conversely a 250 ohm load to a 4,000 ohm feeder by reversing the primary and secondary functions of the windings. An auto-transformer could be used in the same way.

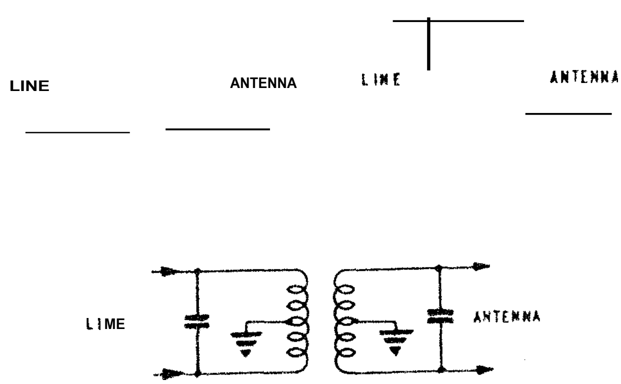
21. The simple explanation of para. 20 serves to illustrate the principle of the transformer hut at radio frequencies the effects of capacitance between and self-inductance of the windings introduce factors which complicate the design of a practical r.f. transformer. This type of matching unit is particularly suitable for the h.f. spectrum of frequencies and can be designed for use with the broad band type of antennae currently preferred. Such transformers are usually designed for the frequency coverage of the associated antenna and may not be suitable for use for other frequencies. Capacitors may be introduced into the matching circuitry (fig. 4(b)) to neutralize the self-inductance of the transformer windings.

Impedance of a terminated line

22. Before describing the method of stub-matching, (para. 37) it is necessary to consider the effective impedance at various points on a feeder where standing waves exist. This effective impedance is similar in nature to the antenna impedance of a dipole (Chap. 2).



(a)



(b)

Fig. 4 Matching units

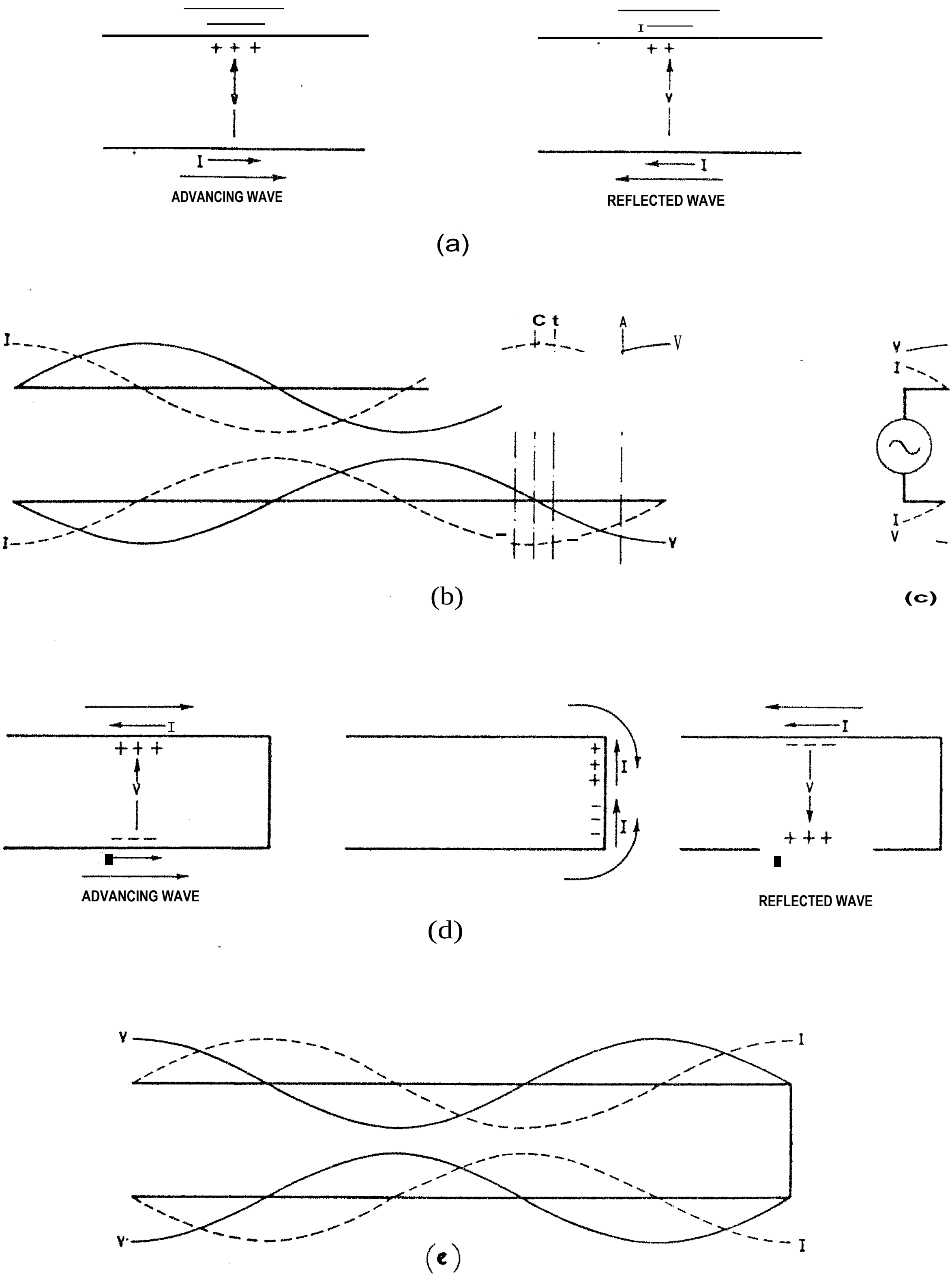


Fig. 5 Open and short-circuited lines

Open and short circuited lines

23. Discussion in the previous paragraphs has been concerned with the behaviour of feeders terminated in a load at the end distant from the generator. Fig. 5(a) represents a voltage pulse and associated current on a feeder whose far end is open-circuited. As with the terminated feeder, the travelling pulse on the feeder consists of a pulse on each wire which travel together but are opposite in phase. On reaching the far (open-circuited) end, the pulse on each wire is reflected back with no change in voltage but a reversal in direction of the current. It may similarly be said for the complete feeder, that the voltage (between the wires) is unchanged on reflection and that the current is reversed. A continuous wave fed down a line may be considered to be composed of a large number of adjacent pulses and will be reflected on reaching the far end. The advancing and reflected waves together produce voltage and current standing waves on the feeder (fig. 5(b)) comparable to those produced on a dipole antenna. The curve V in fig. 5(b) indicates the maximum instantaneous voltage at any point along the line and the voltage rise and fall together. The current (curve I) behaves in a similar manner but the voltage and current reach their maxima a quarter of a cycle apart or, in other words, are 90' out of phase.

24. At the point marked A in fig. 5(b), the voltage is approaching a maximum value and the current a minimum whilst the instantaneous voltage lags the instantaneous current by 90'. At the actual open-circuit, the voltage reaches a maximum (anti-node), there is no current flow (node), the energy cannot be absorbed or dissipated and the reflection process follows. The reflected waves are similar in magnitude and phase variation to the incident waves and could be considered as being produced by a second generation connected at the open-circuit end of the feeder (fig. 5(c)), with the short section of line from A to the open-circuit as the load across its terminals. The standing waves produced are the same as those depicted in fig. 5 (b). The voltage across the hypothetical generator load is high, the current neglible and the voltage lags the current. The load therefore behaves as an impedance with a reactive component which is large and capacitive and with no resistive component.

25. Moving to the left of A in fig. 5(b) to point B, the voltage has reduced to a low value and the current is large whilst the reactance still capacitive has reduced to a low value. At C which is exactly a quarter-wavelength from the open circuit end of the feeder, the current is at its maximum, the voltage has reduced to zero and therefore the impedance has also reduced to zero, virtually a short circuit. It will be seen that by selecting a section of open circuited line less than a quarter wave long, anv required value of capacitive reactance can be obtained by adjusting the length accordingly. At the point D in fig. 5(b), the voltage is in antiphase to the voltage at B but the current is still in-phase with the current at B. Therefore the voltage now leads the current and the reactance becomes inductive.

26. The case of the feeder terminated in a short-circuit at the far end is now considered. In common with the open-circuited line, a pulse travels along each wire, the pulses being in anti-phase (fie. 5(d)). On reaching the far end of the line, each puke travels through the short circuit and returns along the other wire. Considering the pulse pr the feeder as a whole, it can be said that the pulse is reflected at the short-circuited end with a change in direction of voltage but no change in direction of current, which is opposite to the conditions at the end of the open-circuited line. If the feeder is fed with a continuous wave, the incident and reflected waves together form a standing wave as illustrated in fig. 5(e). Throughout the first quarter-wavelength from the short-circuit, the voltage leads the current by 90° and by selecting the appropriate length of feeder, less than a quarter-wave long, it is possible to obtain any desired value of inductive reactance.

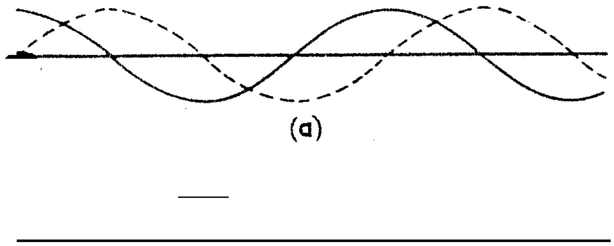
27. The discussion may be summarized as follows:—

(1) By taking a suitable section of feeder less than a quarter-wavelength long, it is possible to obtain any value of reactance by suitably adjusting the length of the section and leaving the end either open-circuited for capacitive or short-circuited for inductive reactance.

(2) The reactances obtained in this way are pure reactances, without any resistive component, since voltage and current differ in phase by 90°.

Representation of standing waves

28. The normal method of depicting the standing waves on a short or open-circuited line is shown in fig. 6(a), with half of each curve shown above and half below the line. At those points where the voltage curve (full line) crosses the line, the voltage is always zero in both conductors. At all points of the half wave length between two adjacent zero points, the voltages on the conductors rise and fall together, i.e. at any instant they are both positive, both zero or both negative and their maximum values are reached at the same instant although the value of the maximum may vary at different points. It may therefore be said that the voltages on that half-wavelength of line are all of the same phase. In the next half-wavelength of the line, the voltages also rise and fall in phase with one another but are in antiphase, with the voltages on the first half-wavelength. As mentioned in para. 25, the voltages at points A, B and C in fig. 5(b) are in phase with one another but in antiphase with the voltage at D. Following the voltage curve from left to right in fig. 6(a) the phase is constant until the curve intersects the zero line and then the phase changes abruptly by



(b)

Fig. 6 Representation of standing waves

180° and remains constant until the next zero intersection when it again changes by 180° and so on. The same changes occur in the current standing wave shown by the broken line in fig. 6(a).

29. When using a meter to detect standing waves on a feeder, the meter readings will give no indication of the phase relationships and the curves obtained will take the form shown in fig. 6(b). Such curves provide all the information normally required in practice.

30. The type of representation given in fig. 5 and fig. 6(a) can be used only for the exactly-in-phase and exactly-in-antiphase relationships, which are shown above and below a line. When a feeder is terminated by a resistance or impedance other than a short or open circuit, the phase varies continuously from point to point in such a manner that representation by the form of curve shown in fig. 6(b) only is possible.

Lines with other terminations

31. If a feeder is terminated by a resistance of a value less than its characteristic impedance, the wave travelling along each conductor passes through the load to the other conductor similar to the short circuited feeder, **but** in this case part of the waves' energy is absorbed by the load and only the unabsorbed portion is reflected down the line, The voltage of the reflected portion of the wave is reversed in phase but the current phase remains unchanged, whilst the combination of the advancing and reflected waves results in a standing wave. The instantaneous voltages and currents on the feeder at intervals of one-eighth of a cycle are shown in fig. 7(a). The curves obtained when the r.f. voltage and current are measured with a meter at various points along the feeder are reproduced in fig. 7(b). Because the reflected wave is lower in amplitude than the advancing wave, the r.f. voltage and current do not fall completely to zero at any point, as compared with a short-circuited feeder where reflection is complete.

32. In addition to the instantaneous voltages and currents at one-eighth cycle intervals, fig. 7(a) also enables comparisons to be made at points A, B, C and D on the feeder of the voltage and current phase and amplitude variations at the beginning, one-eighth and one-quarter cycle intervals at those points. It will be seen that the rates of variation of phase and amplitude are different for voltage and current and that the phase relationship between voltage and current also varies from point to point. At points an exact number of quarter-wavelengths from the toad, the voltage and current are in phase and the impedance is purely resistive, but at intermediate points the voltage and current differ in phase, the impedance at these points having both resistive and reactive components. Both the resistive and reactive components vary from point to point along the feeder.

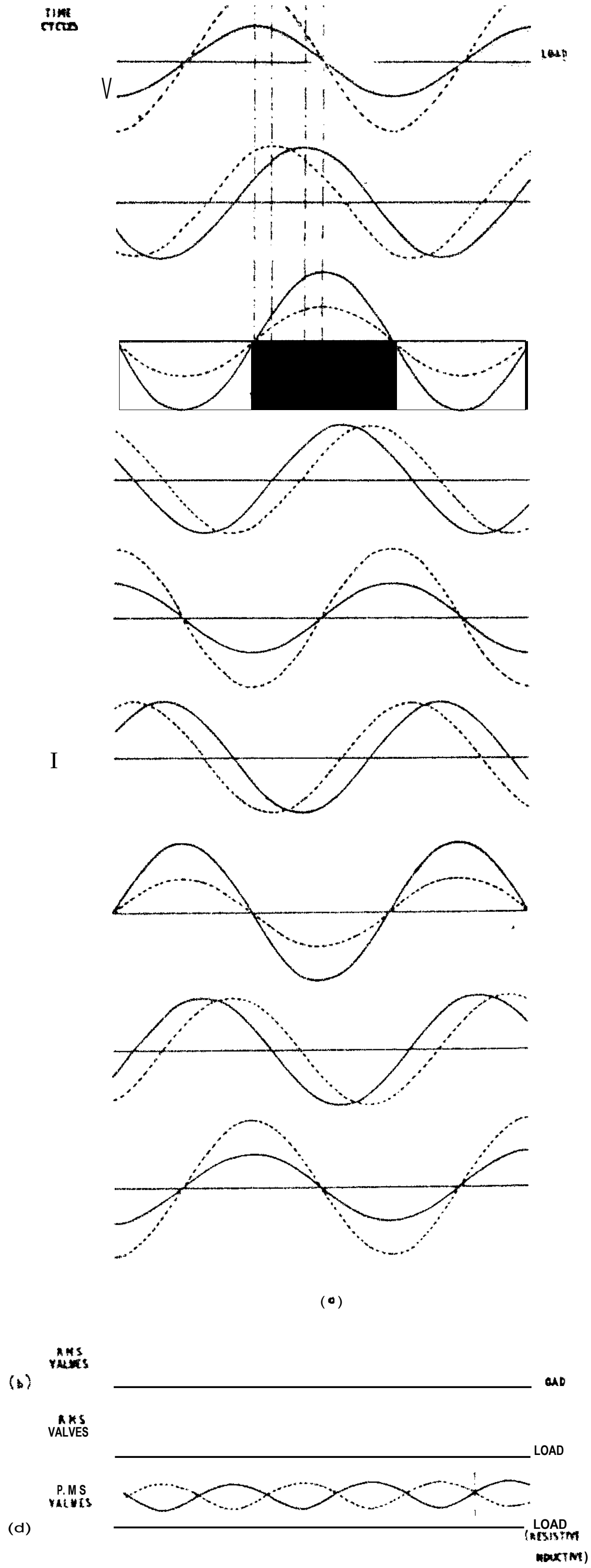


Fig. 7 Examples of standing waves

33. When the feeder is terminated in a resistance greater than its characteristic impedance, the advancing wave meets conditions similar to those in an open-circuited feeder with the result that the wave is partially absorbed in the resistance and the remaining part is reflected back along the line. As with the open-circuited feeder, the current phase is reversed in the reflected portion of the wave whilst the voltage phase remains unchanged. The resultant standing wave is illustrated in fig. 7(c). It will be noted that the curves differ from those of fig. 7(b) by a half-wavelength.

34. If the load terminating a feeder is not a pure resistance but an impedance consisting of both resistive and reactive components, a portion of the wave is still reflected, but compared with the purely resistive load, the phase changes occuring on reflection are more complex. In this case, both voltage and current are subject to a phase change and the change is no longer a phase reversal, but a change through an angle of less than 180^0. The standing wave produced is of the form shown in fig. 7(d) which follows the pattern of fig. 7(c) but with the curves displaced by a fraction of a quarter-wavelength.

35. At the point marked E in fig. 7(d), the voltage and current are in phase and the impedance at this point is therefore a pure resistance. The standing waves to the left of point E are exactly the same as the waves that would be produced if the line was terminated at point E by a purely resistive load of appropriate value. The line may be considered as actually terminated at E by a resistive load, this load consisting of a short length of line terminated by an impedance.

36. The waves depicted in fig. 7 all have a standing wave ratio of 2 : 1. Other values of terminating impedance could be chosen to give a different standing-wave ratio. For any particular ratio, there are two possible values of purely resistive load (para. 31, 33) and numerous possible loads with varying combinations of resistive and reactive components. For all dissimilar loads producing a given standing-wave ratio, the standing waves are identical except for their location along the feeder (para. 30. 31). If one load is replaced by another load which gives the same ratio, the standing waves move bodily along the feeder to another position but otherwise remain unchanged.

Stub Matching

37. A feeder terminated in a load with an impedance which differs from the characteristic impedance, and is neither short or open-circuited, has an effective impedance of a value that varies at different points along the feeder for both the reactive and resistive components of the impedance (para. 32). There are, however, always two points in any half-wavelength of line at which the resistive component becomes equal to the characteristic impedance of the line. At such a point, if the reactive component could be cancelled out, the impedance of the line at the point would be purely resistive and equal to the characteristic impedance of the line. The part of the line from the generator to that point would then be terminated in a resistive load equal to its characteristic impedance, i.e. correctly matched, and with no standing wave condition.

38. A capacitive reactance can be cancelled by connecting an inductive reactance of equal value in parallel with it or vice versa, the two reactances forming a tuned circuit. A convenient reactance for cancelling out the unwanted reactive component (para. 37) is provided by the length of line, under a quarter-wavelength long, described in para. 27. A length of line used in this way is called a matching stub. Typical examples of the current distribution on a line feeding a dipole and using a matching stub are given in fig. 8(a).

39. Apart from the usual feeder losses (para. 14) all the energy fed into a stub-matched feeder is delivered to the load. The total energy in the circuit may be regarded as consisting of two parts, that which is continuously fed from the generator to the load and a constant amount of energy which is oscillating to and from the resistance-free resonant circuit formed by the stub and that portion of feeder on which standing waves exist. The energy in the resonant circuit is provided by the generator during the first few cycles of applied voltage when equilibrium conditions are being established.

40. It is unnecessary to measure the impedance at points on a feeder to determine the length and location of the required stub, this information being obtained from the standing-wave ratio using the curves given in fig. 8(b). The impedance is capacitive in the quarter-wavelength on the load side of a current antinode and inductive on the generator side of a current antinode. Either a short-circuited or an open-circuited stub can be used, the short-circuited stub being connected on the load side of a current antinode and the open-circuited stub on the generator side of the current antinode.

41. The stub and portion of the feeder on which standing waves exist do not radiate to any extent as the waves on the individual conductors are in antiphase. In order to minimize residual radiation and ohmic losses which can be appreciable at current maxima, the selected stub should be connected as closely as possible to the load.

42. When using a short-circuited stub, the distance Y in fig. 8(b) should be measured to the centre of the shorting bar. This point may be earthed if protection against lightning discharge is required. When an open-circuited stub is used, the open end should be adequately insulated because of the high voltage at this point. If the power in use is greater than 5 kW, the voltage may be high enough to produce corona discharges from the conductor open ends and insulators with corona rings should be fitted to prevent this. The ring capacitance is appreciable and shortening of the stub length will be necessary to compensate.

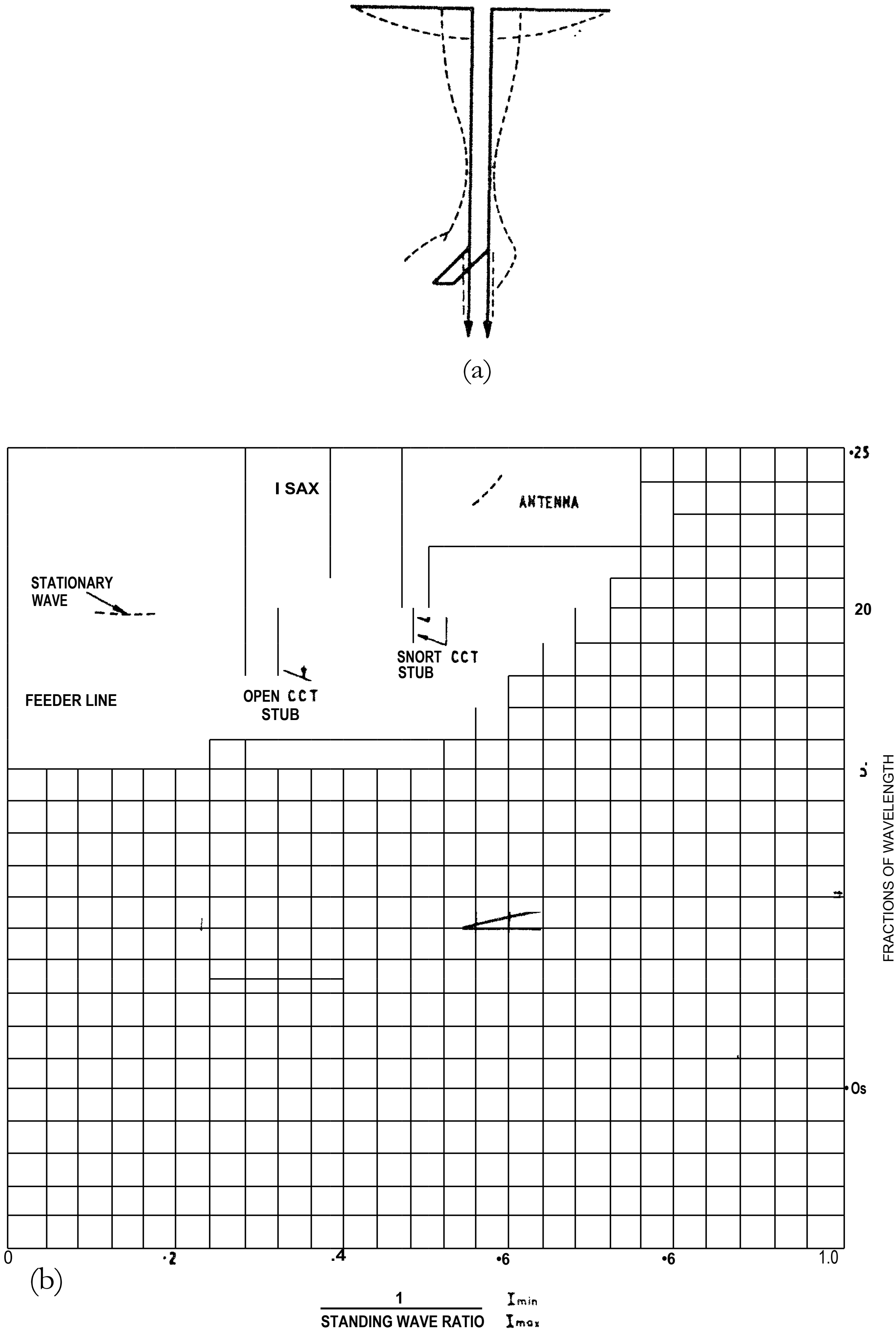


Fig. 8 Matching stubs

43. The dimensions given in fig. 8(b) may not provide a perfect match owing to stray capacitance etc., which cannot be allowed for in the calculation. The stub should first be fitted temporarily and adjusted until no standing wave occurs. This may not be possible to achieve in practice and stub adjustment should then be directed to producing a standing-wave ratio as close to unity as possible.

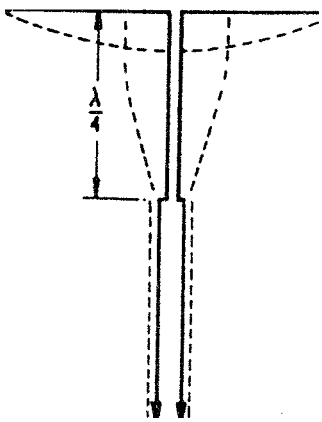


Fig. 9 Quarter-wave transformer (Q-bars)

Q—bars

44. If the load at the end of a feeder is purely resistive, the impedance one quarter-wavelength from the load is also purely resistive. Where RI is the value of the load, R2 is the impedance one quarter-wavelength back and Z_0 is the characteristc impedance of the feeder, the three quantities are related by the expression 4^2 = RI R2. A feeder whose characteristic impedance is RI can therefore be matched to a resistive load R2 by using a quarter wavelength of line whose characteristic impedance is 4 = √RI R2 as illustrated in fig. 9. A quarter-wavelength used in this way is called a quarter-wave transformer. It may be constructed from the appropriate length of open or coaxial line or, at the higher frequencies from two ¼ inch or inch diameter tubes a quarter-wavelength long, known as Q—bars.

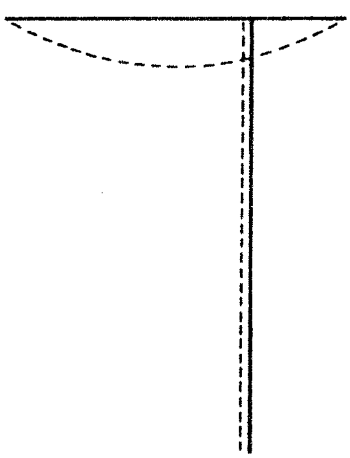


Fig. 10 Windom antenna

Windom antenna

45. The impedance of a half-wave dipole varies from about 73 ohms at the centre to about 2,500 ohms at the ends. On either side of the centre there are points where the impedance is from 500 to 700 ohms, suitable for matching a single-wire feeder so that the feeder is free from current standing waves and the feeder radiation is low compared to the dipole radiation. A dipole fed in this manner is called a Windom antenna (fig. 10). The exact distance of the feed point from the centre must be established empirically but is approximately 0.66 of a wavelength. Owing to the feeder radiation, the system is confined to installations such as low power mobile equipment.

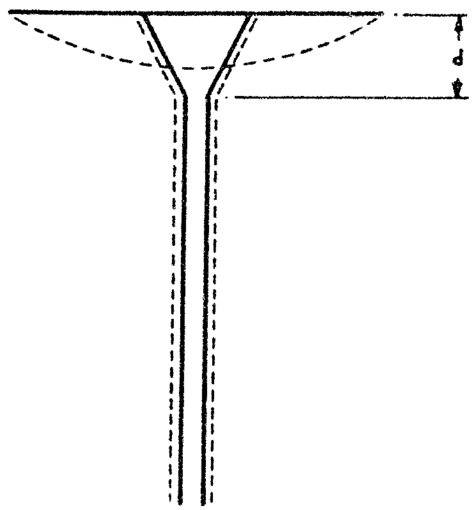


Fig. 11 Y-matched dipole

Y or delta matching

46. If a second feed-wire were connected to a Windom antenna at the corresponding point on the other side of the centre, a balanced system would result with the travelling waves in both wires equal in amplitude. The connection point for each wire would differ slightly from that for a single wire, as the characteristic impedance of each wire is affected by the other. The distance between the two wires at the connection points would be about one eighth-wavelength which is too great for the radiation from the wires to cancel completely. In practice, therefore, an ordinary feeder is used and opened out at the antenna (fig. I I) the correct value for the dimensions "d" being dependent upon the operating frequency of the antenna. As an example. for a half-wave dipole delta-fed by a feeder of 600 ohms characteristic impedance, distance "d" (feet) $= \frac{148}{(MHz)}$ and the distance in feet between connection points is $\frac{123}{(MHz)}$, at the fundamental resonant frequency of the antenna. A dipole fed in this way is said to be a **Y** or delta-matched dipole.

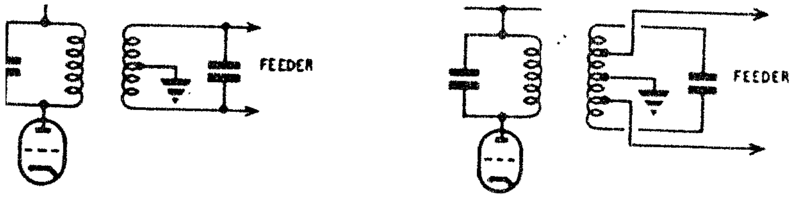


Fig. 12 Matching transmitter to feeder

Other requirements

47. It is not only necessary to match the feeder to the transmitting antenna (load) but the matching of the transmitter output stage (generator) to the feeder is of equal importance, otherwise the feeder cannot accept all the available energy from the transmitter and some is reflected back into the output stage with possible disastrous effects to the equipment. Such matching is usually accomplished by antenna couplers or transformers, examples of the circuitry being given in fig. 12.

48. It is general practice for the transmitter output to be fed through a coupling network into a coaxial r.f. connection and by coaxial cable to an antenna coupler or tuning unit to which the feeder is connected. In those cases where the antenna is already fed by a coaxial line it is possible for the feeder to be directly connected to the transmitter coaxial output socket. More generally, it is necessary to interpose an intermediate coupling or transforming device, because the line, although of equal impedance, may be a balanced twin feeder. The antenna coupler can **be** built into the transmitter, but may be designed as a separate unit and incorporated in a junction box on the wall of the transmitter building or other convenient location.

49. The antenna coupler or tuning unit may be required to incorporate any or all of the following important functions:—

(1) Tuning the antenna or feeder to resonance.

(2) Transforming from one impedance to another, that is the transmitter output to the feeder.

(3) Linking an unbalanced output to a balanced line or vice versa.

The subject of balance is discussed in later paragraphs.

50. Matching is equally desirable between a receiving antenna and its feeder by means of either a matching unit or stub and also between the feeder and receiver which is usually provided by the input circuitry of the receiver.

Balance

51. In addition to being correctly matched, a **feeder** system must also be balanced or, in other words equal currents should flow in both conductors. An unbalanced feeder will radiate (para. 13).

Open- wire feeder

52. An open-wire feeder system is balanced if the impedance to earth is the same for both conductors and if the effective generator and load impedance are also balanced to earth. For the ordinary type of pole-supported feeder, the impedance to earth of both wires is the same unless the line is so run that one wire is appreciably closer to an earthed object than the other. The balancing of the *generator* is obtained by suitable design of the coupling circuitry between the transmitter and feeder, examples of which are given in fig. 12. Types of antenna which are symmetrically disposed relative to earth, such as the horizontal dipole and rhombic, automatically present a balanced load to the feeder. Other types of antenna, such as the Marconi and other vertical antennas. which are not symmetrically disposed, present an unbalanced load and the matching unit must be designed not only to match the impedances, but also to correct for the unbalance to balance condition. Such a unit is shown in fig. 13.

53. Unbalance will occur in an open-wire system if any one of the three constituent parts (generator, feeder, load) is unbalanced.

(1) Unbalance in the generator will result in unequal voltages and currents being fed into the feeder wires.

(2) If a path between one wire and earth is inadvertently provided, the currents in the feeder wires will be unequal.

(3) An unbalanced load will terminate each feeder wire in a different resistance, resulting in unequal currents in the wires, reflection in only one wire or differing reflections in both.

The possibility of producing a balanced feeder system by counteracting the unbalance or one part of the system by a compensating unbalance in another part of the system is not a practical proposition.

Effective earth

54. A balanced load is conventionally represented symbolically by a centre-tapped resistor with the centre-tap connected to earth. In practice the load may not be physically connected to earth, but any point at which the voltage is always zero may he regarded as being effectively an earthed point. A point on a feeder where the voltage standing wave is at zero is a typical example. This does not

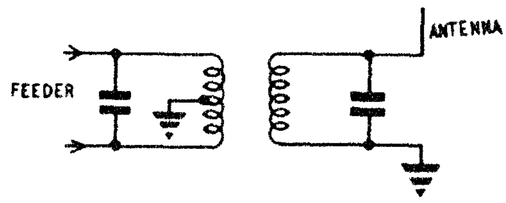


Fig. 13 Matching balanced feeder to unbalanced load.

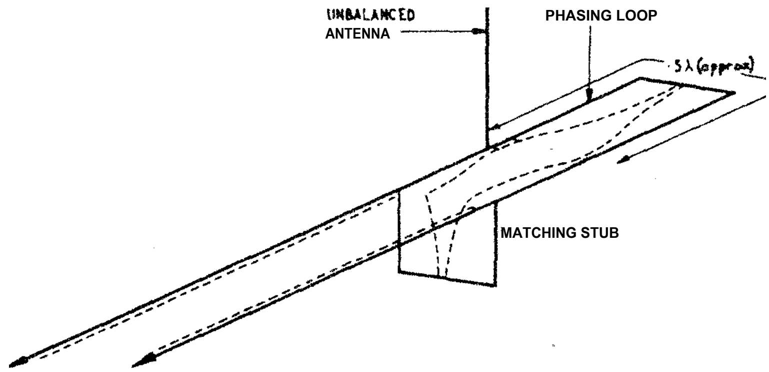


Fig. 14 Phasing loop

necessarily mean that such a point can be directly connected to earth as the connection may radiate, introduce capacitance, or otherwise disturb the conditions responsible for the standing waves. A physical connection to earth must be made somewhere in the circuit, however, in order to provide a reference potential, as shown in fig. 12.

Phasing loop

55. Apart from the use of a matching unit, a balanced open-wire feeder can be connected to an unbalanced load, such as a vertical antenna, by using a phasing loop. The loop consists of about a half-wavelength of wire folded back on itself to prevent radiation and connected as shown in fig. 14. The voltages and currents at points on a wire a half-wavelength apart are in antiphase and, in the absence of the antenna, the half-wavelength of wire bent into a loop would provide a balanced load. Connecting the antenna to one wire of the feeder introduces a lack of balance which can be rectified by suitable adjustment of the length of the loop. The adjustment required is found by measuring the standing waves on both wires of the main feeder and noting any difference between the distances of the current maxima from the antenna. Using the point of current maximum on the antenna side of the feeder as datum, the length of the loop is adjusted until the current maximum on the other wire is exactly opposite. The effective earth for the vertical antenna is a point midway between the two wires at the position where the antenna is connected. A matching stub may be required in addition to the phasing loop and should. be fitted after the loop has been adjusted.

Coaxial cables

56. The coaxial feeder, initially described in para. 7 and 8, functions in a similar manner to the open-wire feeder with one travelling wave moving along the inner conductor and a travelling wave of opposite phase moving along the inner conductor and a travelling wave of opposite phase moving along the inner surface of the outer conductor where the current is confined at radio frequencies by skin effect. The outer surface of the outer conductor acts as a screen to any radiation and when earthed, all radiation is confined within the envelope of the cable. For connection to the generator and the load the coaxial cable must be cut to the length required, an example being illustrated in fig. 15(a), where the wave travelling along the inner surface of the outer conductor will find its way to earth at the outer surface. Obviously the generator and load connected to the feeder must be unbalanced as in each case one terminal will be earthed on connection if not already earthed.

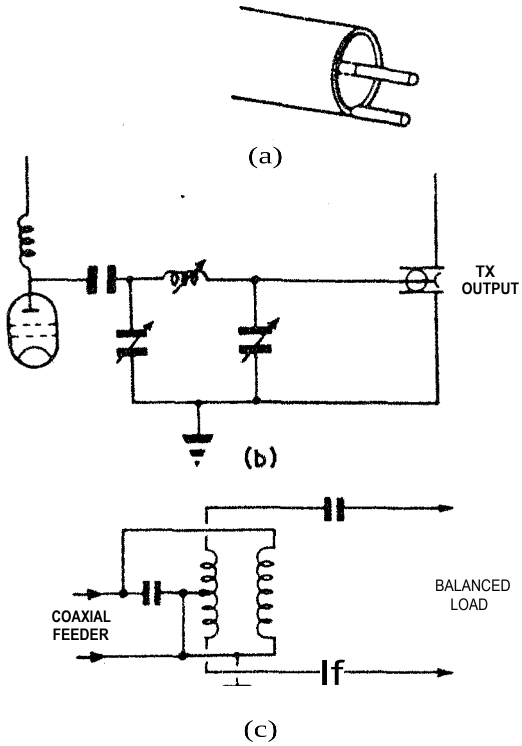


Fig. 15 Matching coaxial feeders

57. As both generator and load connected to a coaxial line must be unbalanced, this type of feeder is said to be inherently unbalanced. An unbalanced *generator* for working into a coaxial feeder is normally provided by a suitably designed coupling between the output stage of the transmitter and the feeder, a typical example being illustrated in fig. 15(b). The unbalanced *load* for connection to the coaxial feeder can be either an unbalanced antenna, such as the Marconi which if impedance matched, can be directly connected, or a balanced antenna with a suitably designed matching unit interposed between the antenna and feeder. An example of such a matching unit is given in fig. 15(c).

Balance to unbalance devices (Baluns)

58. If a coaxial feeder is connected directly to a balanced antenna such as the half-wave dipole, the transmitting current will flow in the screen of the feeder in addition to the antenna current and there will be radiation from the outer conductor which will upset the pattern of the polar diagram and reduce the antenna efficiency. The balance can be restored by the use of a balancing device known as a balun at the connection between the feeder and antenna.

59. One type of balun, in its simplest form, consists of a metal cylinder or sleeve, a quarter-wavelength long which is connected to the outer conductor at a point one quarter-wavelength of the feeder (fig. 16(a)). It is only effective over a narrow band of frequencies on either side of the frequency for which the sleeve is a quarter-wavelength. The sleeve effectively separates the inner and outer surface of the outer conductor and reduces the currents on the outside of the feeder to a negligible value, enabling the antenna to operate in a well-balanced condition. The outer surface of the balun becomes a continuation of the feeder screen, whilst the inner surface of the sleeve and connector end with the enclosed outer surface of the feeder outer conductor together form a short-circuited coaxial stub.

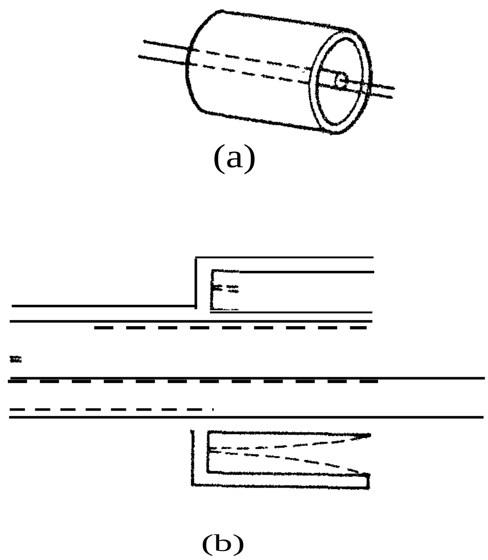


Fig. 16 **Balun**

60. During the initial cycles of applied voltage the travelling wave passes from the outer to inner surfaces of the feeder outer conductor and sets up a similar wave of opposite phase, travelling in the same direction, on the inner surface of the sleeve which forms the other conductor of the stub. These waves together produce a standing wave as illustrated in fig. 16(b). This standing wave consists of wave-fronts surging to and fro in the stub as in a resonant circuit, its initial energy being provided by the generator during the first few cycles whilst the equilibrium condition is being established. Once the standing wave has formed, the current is zero at the open end of the stub and no current flows into the stub from the feeder (except a low value compensating ohmic loss) or from the stub to earth. The feeder can therefore he connected to a balanced load. The voltage at the open end of the stub is of the same amplitude and phase as the voltage of the travelling wave at the end of the outer conductor of the coaxial feeder.

61. The type of balun previously described is also known as a simple choke type and by the addition of an extra choke, the balun can be made capable of operating over a broader frequency band. The transformer type of balun mentioned in para. 19(1) is also an example of a balance to unbalance matching device for broad band operation.

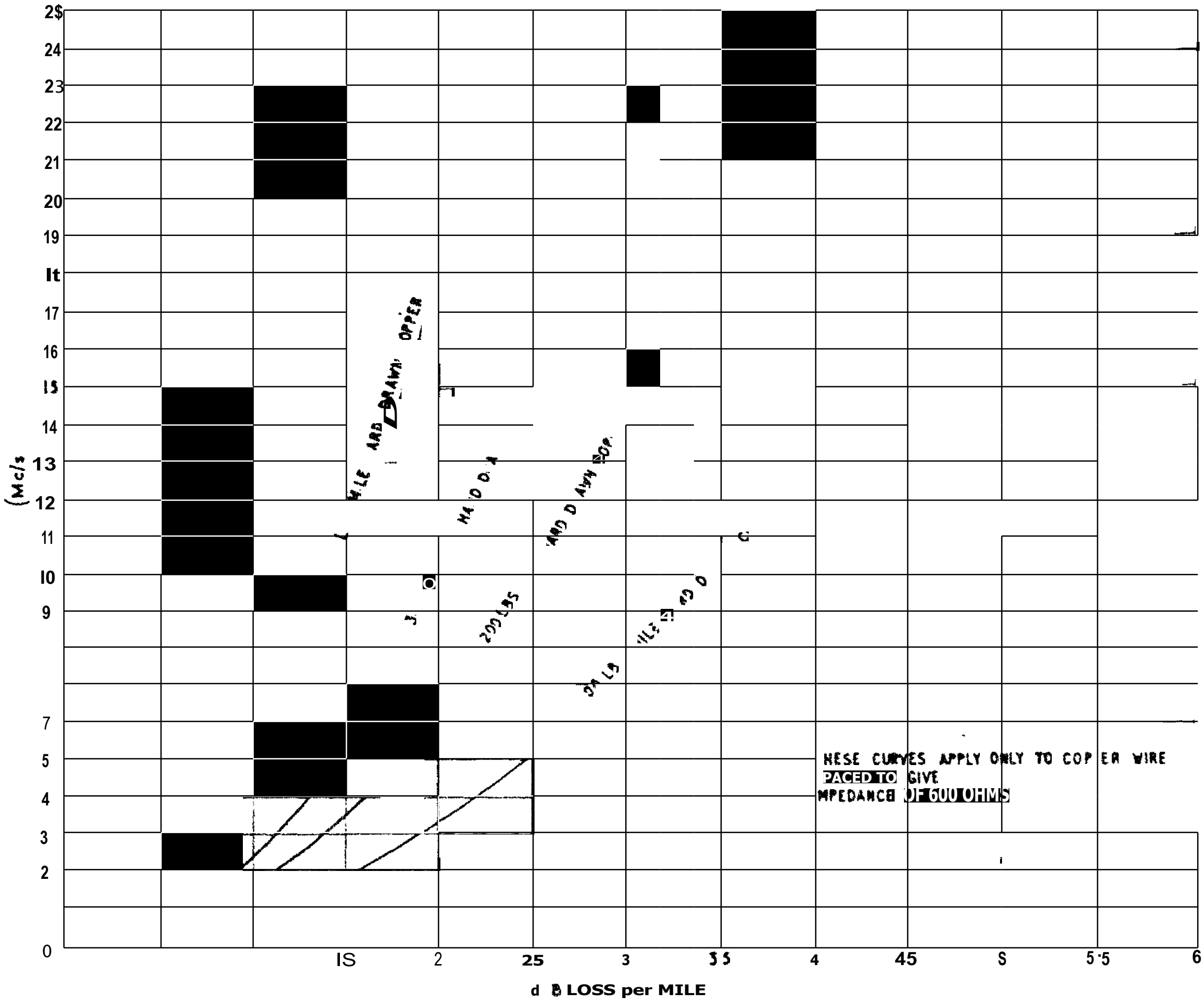


Fig. 17 Effect of frequency on open wire feeder attenuation

Feeder characteristics

62. Fig. 17 shows the effect of frequency on the attenuation of open-wire feeders, spaced to give the standard characteristic impedance of 600 ohms. The graph covers four commonly used weights of hard drawn copper wire viz. 600, 300, 200 and 100 lb. per mile.

63. Fig. 18(a) shows the relationship between impedance and wire spacing for 600 lb. per mile (6 s.w.g.) and 100 lb. per mile (14 s.w.g.) hard drawn copper wire.

64. Fig. 18(b) shows the variation of nominal attenuation with frequency for four commonly used coaxial cables viz. uniradio 57, 63, 67 (this is the replacement for uniradio 4) and 132 (mineral filled).

65. Fig. 18(c) shows the variation of power rating in air with frequency for the same four coaxial feeder cables referred to in para. **4.**

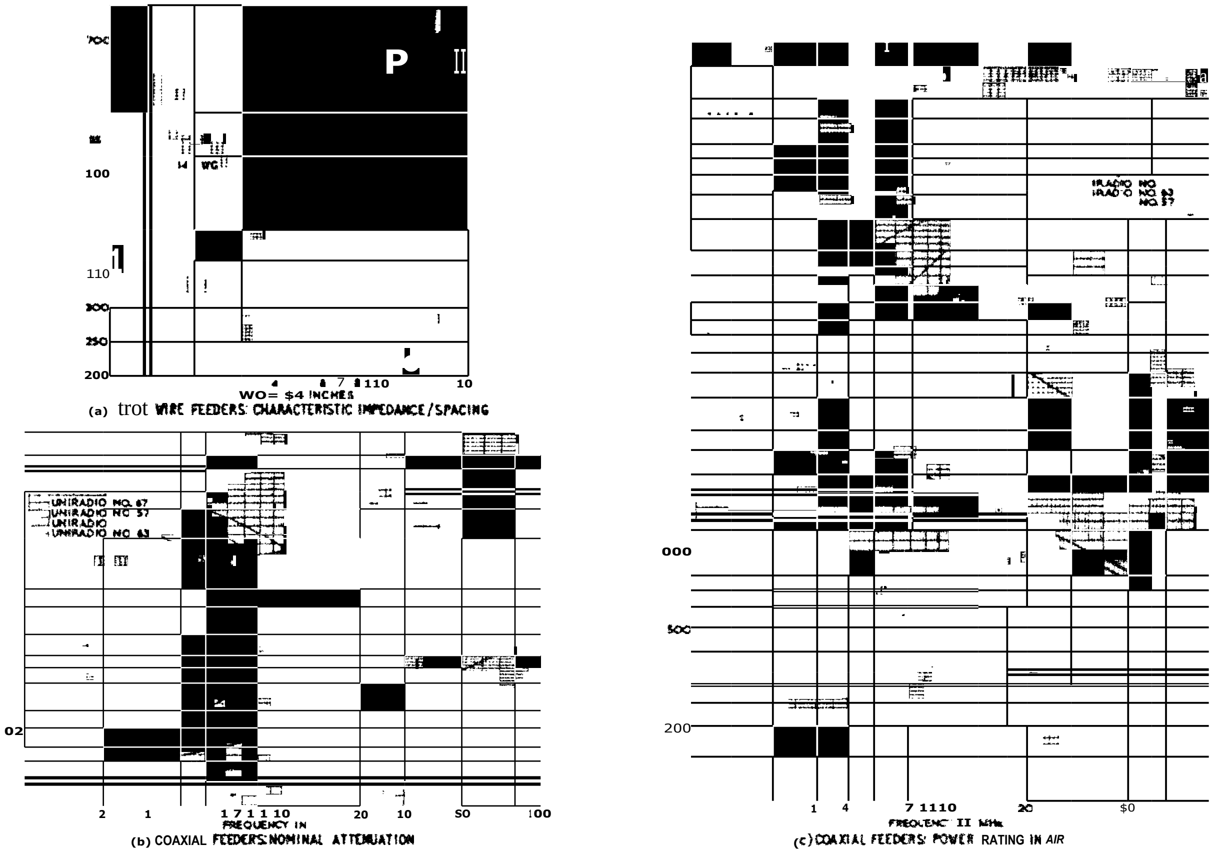


Fig. 18 HF and MF antenna feeder characteristics