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G. Marconi

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THE MARCONI REVIEW

No. 73. *April-June, 1939.*

Editor: L. E. Q. WALKER, A.R.C.S.

THE MARCONITRACK AIRCRAFT RADIO BEACON SYSTEM

The following article describes the development of an improved type of ultra short wave Radio Beacon System for assisting the approach of aircraft to an aerodrome during QBI conditions.

THE procedure adopted by the pilot of an aircraft when approaching an aerodrome in conditions of very bad visibility and with the assistance of a landing beacon and the apparatus such a system involves are nowadays widely known, but for the sake of those who are unfamiliar with the subject, a brief outline of the apparatus and the method in which it is used is not out of place.

A beacon situated at the end of a suitable landing runway and termed " The Main Beacon " transmits rhythmically a series of dots and dashes, the aerial arrangements being so disposed that on a given desired bearing line from the beacon, i.e., in line with the landing runway, they merge into a continuous unbroken signal. At any position within range of the main beacon and not on this bearing line or " Equisignal Track," a predominance of dots or dashes is heard.

On this equisignal track are situated two further beacons of low power, radiating in a mainly vertical direction and known as " Marker Beacons." Of these, one termed the " Boundary Marker Beacon," is situated on the aerodrome boundary, while the other, known as the " Approach Marker Beacon," is located several miles from the aerodrome. Each marker beacon transmits its own characteristic signal.

A pilot wishing to land on the aerodrome and having obtained permission from the airport control officer to use the beacon, flies by normal navigational means to a point in the approximate neighbourhood of the equisignal track but well outside the approach marker beacon. He then turns right or left according to whether he hears a predominance of dots or dashes in his earphones in conjunction with the indications given on his visual indicator forming part of the aircraft beacon receiver. Having reached the equisignal track as indicated by the fact that he now hears a continuous signal and his visual indicator needle is central and stationary, he maintains his course along the track by constant correction whenever he hears any tendency for dots or dashes to predominate and sees his visual indicator needle begin to flicker one way or the other. He continues in this way, flying at a safe altitude, until he reaches the approach marker beacon as indicated by a characteristic warbling note in the earphones. From previous information he then knows his exact distance along the track from the aerodrome boundary and proceeds to lose height so that by the time he arrives over, and receives signals from, the boundary marker beacon, he is in a position preparatory to land ahead. For the final flattening out and touch down a vertical visibility of some 60 feet at least is necessary. Claims that aircraft may be brought right down to the ground in conditions of zero visibility by means of a "glide path indicator" forming part of the aircraft receiver have now largely been abandoned. This is due, among other causes, to the fact that the glide path indicator depends for its action on the shape of the vertical polar diagram of the radiation from the main beacon, which shape, being liable to alter with changing ground conductivity, is largely influenced by changes in the ground surface moisture content. An added difficulty is the fact that the optimum glide path of various types of aircraft differs widely, and the pilots of many types would be confronted with an almost impossible task in attempting to follow an arbitrarily imposed one.

Methods of Providing the Equisignal Track.

Having briefly recapitulated the basic principles of such landing systems, it is of interest to investigate the methods by which an equisignal track may be provided in space and the effect of such methods on running economy, mutual interference between adjacent beacons, site errors, and their navigational aspect from the pilot's point of view.

The conventional aerial system for producing an equisignal zone on ultra short wavelengths consists of a vertical rod dipole centrally disposed between two reflector rods, the array thus comprised being situated at right angles to the line of approach.

The reflectors are keyed in the dot-dash interlock system, the resultant radiation diagram being two overlapping ellipses causing the formation of two equisignal paths 180 degrees opposed, where the diagrams, when changing over in the keying rhythm, overlap.

In the Marconitrack system, on the other hand, the equisignal path is formed by the overlapping of two beams projected from two arrays of the well-known Marconi-Franklin Series Phase type.

The advantages of this method are that :—

(1) Improved signal discrimination and a resultant sharpening of the equisignal track are obtained.

(2) Due to the forward concentration of radiated energy, mutual interference with neighbouring beacon systems is minimised.

(3) A saving is also effected in the radiated energy by its projection in a mainly forward direction with a resultant economy in input power for a given length of equisignal track.

(4) The 180 degrees track ambiguity is eliminated.

(5) The track width is more easily controllable.

(6) The beam elevation can be controlled.

(7) Serious site errors due to lateral reflection effects are minimised.

Steering Accuracy.

With regard to the most important question of steering accuracy, we shall show why it is simpler for a pilot to follow a sharp, narrow track than a comparatively broad one.

In Fig. 1 a track having a divergency of 5 degrees is shown. The line ABC represents the course of an aircraft travelling along one edge of the equisignal zone and commencing to deviate therefrom at the point B through the arc of a circle of radius R and cutting the other side of the equisignal track at an angle θ. To

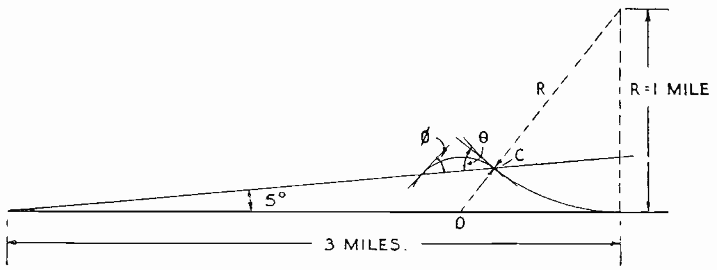


FIG. 1.

return to the track, the aircraft must now make a turn in the opposite direction through the arc of a circle, let us say, of radius OC. Owing to the magnitude of the angle θ this arc is seen to represent a sharp change in direction necessitating a steep turn and furthermore, unless correction is anticipated, the aircraft will be seen to re-enter the track also at a considerable ϕ.

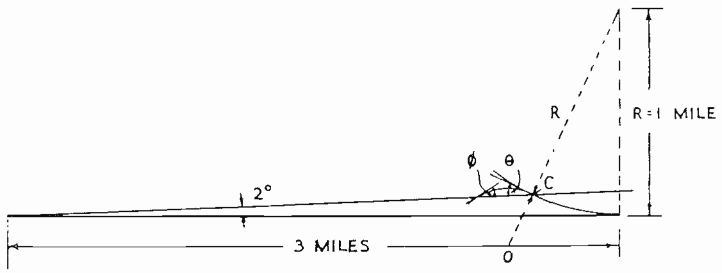


FIG. 2.

It is true that the resultant liability to wide amplitude "weaving" is in normal practice substantially reduced by co-operation with another entirely different directive system, i.e., a compass or directional gyro. Resort to this alternative aid, however, does not eliminate the case in favour of a sharper track, but rather accentuates its need.

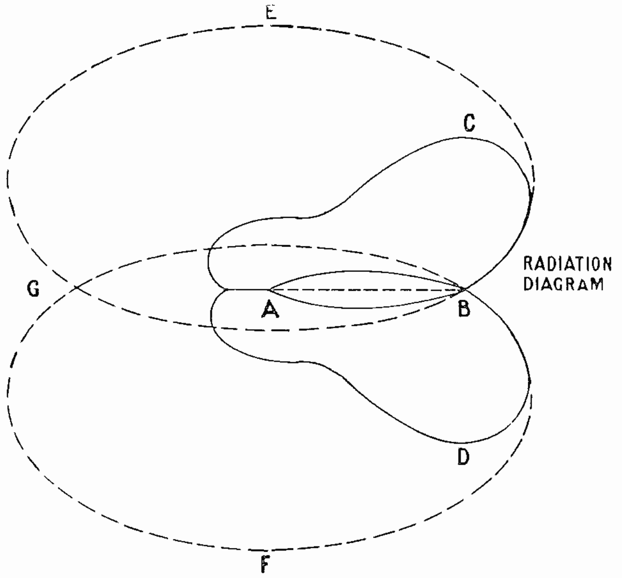
Fig. 2 shows a track 2 degrees in width and it can be seen that although the aircraft commences to deviate from the track on a turning circle of the same radius, the amount of course change required to re-enter the track on the same turning radius OC is much reduced by reason of the much smaller angle θ necessitating only a slight medium turn and the angle of re-entry into the track ϕ is also very appreciably smaller.

It is clear, therefore, that the sharper course reduces undesirable weaving whilst facilitating the following of the precise track with the minimum of error. This, of course, is particularly desirable when lateral deviations are not permissible; for instance, on crossing the aerodrome boundary in a gap between obstructions.

Mutual Interference between Adjacent Beacons.

With regard to the second point, Fig. 3 shows the considerable saving in interference area between the diagrams produced by the overlapping ellipses and the overlapping beams for an equal length of equi-signal track. The very substantial reduction in interference area enables stations using a common frequency to be more closely spaced, thereby economising in frequencies and avoiding resort to unnecessary complications in wavechanging in flight.

As a direct result of the improved radiation diagram given by the beam system, the transmitter itself can be rated down from the medium power of 500/800 watts to the feeder, to a relatively small power of about 100 watts or less.



A.B. = APPROVED TRACK.
A.B.C } = TWO OVERLAPPING
A.B.D } BEAMS.
E.F.G.B. = TWO OVERLAPPING ZONES PRODUCED BY A VERTICAL AERIAL WITH TWO REFLECTORS ALTERNATELY USED

FIG. 3.

Site Errors Minimised.

It will be realised that the ideal approach track to an aerodrome does not always provide along its path a very suitable site for the location of an ultra short wave directional system, which, fundamentally, must be of a high order of accuracy. It is clear that whereas the radiation diagram provided by two overlapping ellipses develops powerful fields at right angles to the approach line, producing the maximum of reflection errors from objects such as metal hangars, etc., laterally disposed to the approach line, in the case of the beam system, radiation is at a minimum laterally. Furthermore, the sharper discrimination round about the equisignal track provides greater safeguards against errors caused by small inequalities of signal intensity on either side of the track.

In regard to the site problem, it will, of course, be apparent that the adjustment of the elements of a composite array provides a certain measure of flexibility in compensating for inequalities, a means of correction not so readily available with the more simple forms of radiator.

Difficulties Overcome.

While still in the experimental stage and in order to test out some of the points

The system was found to suffer from two major defects, i.e., instability of the main forward track and a numoer of false courses at the side and back caused by the overlapping of subsidiary lobes.

Upon examination, each of the beams was found to have a radiation diagram as shown in Fig. 5, the two overlapping beams forming a plurality of false courses as shown in Fig. 5A.

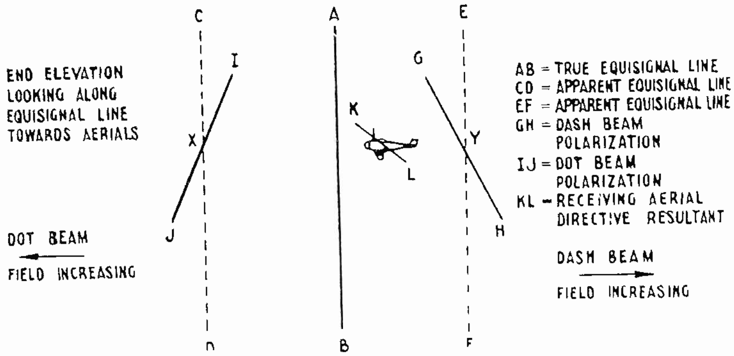


Fig. 6.

Attention was firstly directed to the instability noticed in the forward course, an effect which was found to vary with different types of aircraft with which flight trials were made.

The effect was a shift of course—through as much as 15 degrees—when the equisignal line was approached from different directions. The course was projected from West to East, and when approached from North to South the track would be found to lie to the south of the proper line, and, vice versa, when approached from the south, it would be found to lie to the north. On turning the head of the aircraft into the apparent track, a further shift resulted. Track widths of unequal value when flying laterally in opposite directions was another observed phenomenon.

Upon investigation, the cause of the instability was found to be the existence of a horizontally polarised component in the receiving system of the aircraft. It is well known that spurious polarisation from the intended plane of polarisation may be brought about by reflection and other factors affecting the wave propagation from the radiating source.

In the case under consideration a certain amount of horizontal component was found to be introduced due to the attenuation along the arrays from front to back, causing an out-of-balance radiation from the horizontal components of the series-phase aerials. This gave an apparent forward tilt to the wave front and, with the aerial layout shown in Fig. 4, an apparent lateral tilt on either side of the equisignal line.

The aircraft is fitted with a vertical mast aerial and when the mass of the aircraft is grounded, the vertical aerial functions as such. When existing as a free object in space, however, the aircraft, especially if of the composite type of construction, appears to behave like a bent dipole, thus introducing a horizontal component to the receiving system.

The point of equal coupling of the receiving system to the lines of field strength of the two overlapping beams will thus vary with the machine's orientation. The effect is shown in Fig. 6, from which it will be clear that the aeroplane characterised by the receiving system KL will have to fly beyond the true equisignal line AB in the more powerful field towards the point X in order to compensate for the reduced coupling its aerial has to the wave front it is approaching.

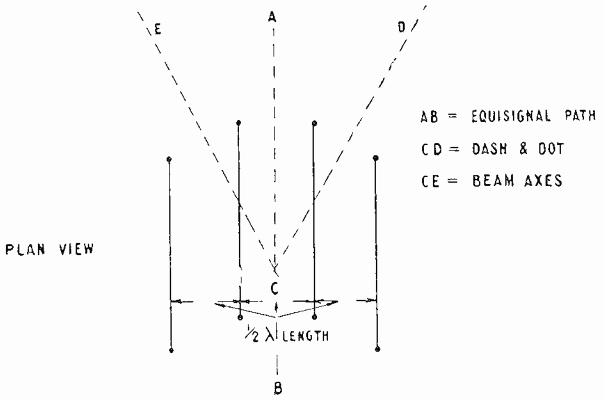


FIG. 7.

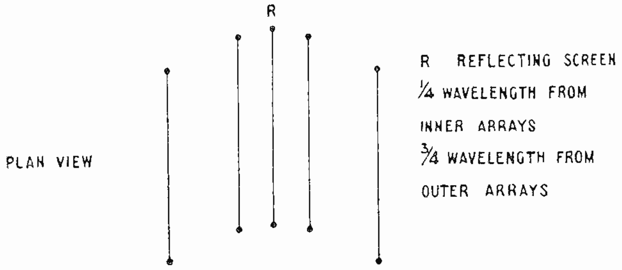


FIG. 8.

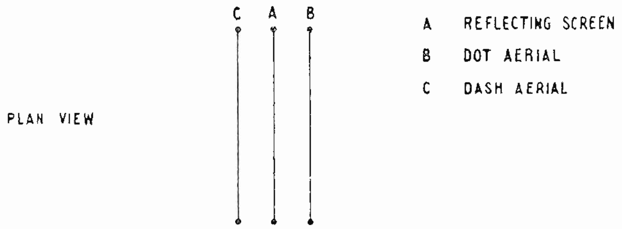


FIG. 9.

Errors introduced by the existence of a horizontal component may take other forms. Such, for example, as that caused by a transitory change of signal on steeply banked turns. Course splitting, too, resulting in multiple equisignals has been noticed on ground tests on many occasions when the horizontal component is present.

A convenient solution was found by re-aligning the arrays so that they were parallel to the equisignal line, the required overlap of the beams being adjusted by stepping back the two outer arrays. Fig. 7 shows this arrangement. The effect of paralleling the arrays is to eliminate completely the apparent lateral tilt along the line of equisignal. The slight apparent forward tilt remaining is equal and in the same sense in the case of both dot and dash wavefronts and therefore the effect on the receiving aerial from each beam is the same. It will be seen that the horizontal component is not eliminated but is so disposed that it cannot detrimentally affect the steering accuracy along the equisignal track. Many flight trials have shown that the track marked out by radiation from the aerial system shown in Fig. 7 is sharply defined, stable and reliable.

Having advanced to this point, the problem of the composite back radiation diagram (Fig. 5A) was next considered.

As the forward to back radiation is in the ratio of about 9 : 1 there is a sharp fall in intensity after passing over the aerials from front to back, and hence, if a manually biased receiver is used, this ratio can be put to advantage so as to eliminate the back radiation almost completely. Although it is easy to demonstrate this effect,

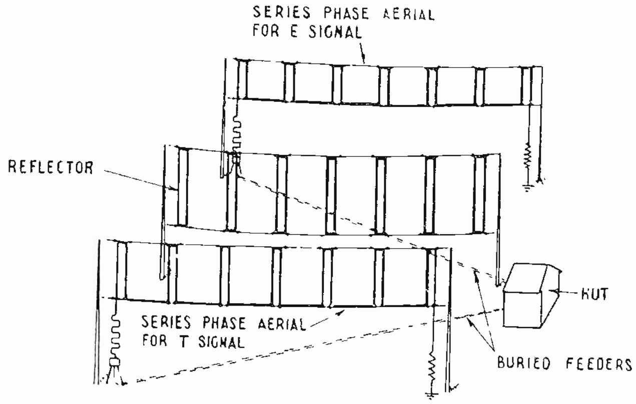


FIG. 10.

which gives the impression of a single forward track with very sharp beam formation when entering the forward area after a circuit, it is not a solution of practical value since other considerations require receiver sensitivity to be controlled by automatic means (delayed A.V.C.).

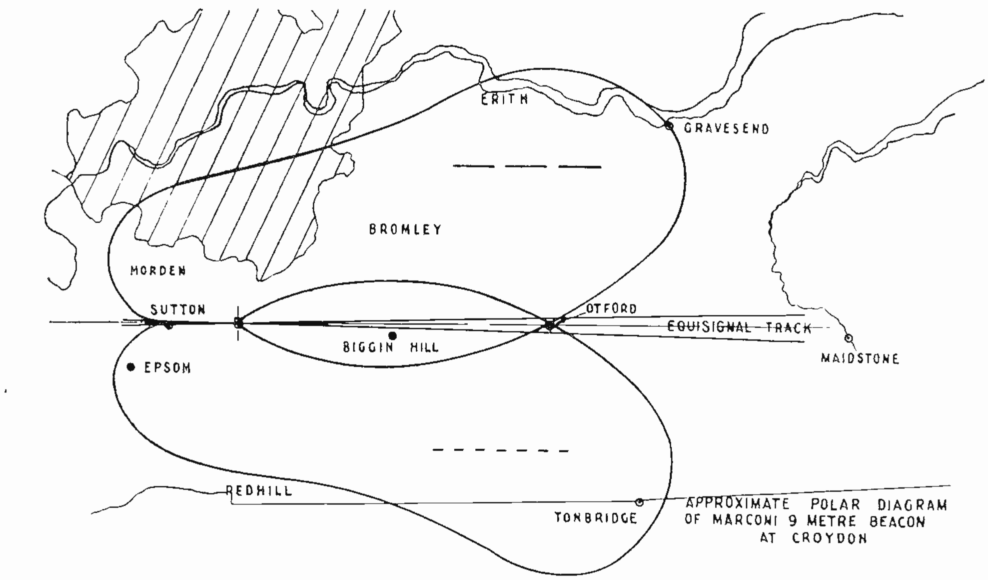


FIG. 11.

It was therefore decided to examine the possibility of reflecting back the side radiation from each of the aerials, and from the results of preliminary exploration with single half-wavelength rods, a central reflecting screen was installed as shown in Fig. 8. It became apparent that the central reflector was not only having the desired effect in reflecting back side radiations and hence preventing the overlapping of lobes of opposite sign which had led to the formation of false equisignal courses, but this new addition to the arrays was tending to sharpen the main beams and to reduce the overlap between them.

It was therefore found possible to dispense with the two outer arrays altogether, thus simplifying the feeder system, bringing the geometrical centres of the beam radiation nearer together and conserving the space needed on any given site. Fig. 9 shows this arrangement.

Fig. 10 shows the resulting radiation diagram in approximate outline and Fig. 11 shows a sketch of the aerial and reflector system showing feeder arrangements, etc.

From the results given above it is evident that the objectives aimed at, viz., high quality discrimination, reduction of interference area and of power required for a given length of path, have been realised in the " Marconitrack " system, and an important step forward in the technique of ultra short wave directive systems has been achieved.

J. M. FURNIVAL.

RETIREMENT OF MR. H. M. DOWSETT

Mr. H. M. Dowsett, Principal of the Marconi School of Wireless Communication, and Editor of THE MARCONI REVIEW since its inception in 1928, has retired from the Marconi Company on reaching the age limit after 40 years' service with the Company.

Mr. Dowsett has been succeeded by Mr. L. E. Q. Walker as Editor of THE MARCONI REVIEW.

THE REDUCTION OF DIODE DETECTOR DISTORTION BY POSITIVE BIAS

This note shows that positive bias may be used in a diode detector to eliminate coupling circuit distortion. For satisfactory operation the bias should be controlled by the signal input and a method of achieving this is described.

A CAPACITANCE-RESISTANCE coupling is necessary to transfer the audio frequency voltages developed across a diode detector load resistance to the grid of the first A.F. amplifier. The grid leak resistance across the amplifier grid circuit affects the performance of the detector by limiting the maximum modulation percentage which can be accepted without distortion. It can be proved for a linear diode detector that the modulation percentage at which distortion occurs is

$$K\,\% = \left[1 - \eta_D \cdot \frac{R_1}{R_1 + R_2}\right] \times 100 \qquad \text{(1)}$$

where R_1 = diode load resistance.

R_2 = A.F. amplifier grid leak resistance.

$\eta_D = \dfrac{I_m R_1}{\hat{E}}$ the diode detection efficiency,

and I_m = mean D.C. current through R_1 for a given carrier voltage $\hat{E}$ sin wt (1 + K sin pt)

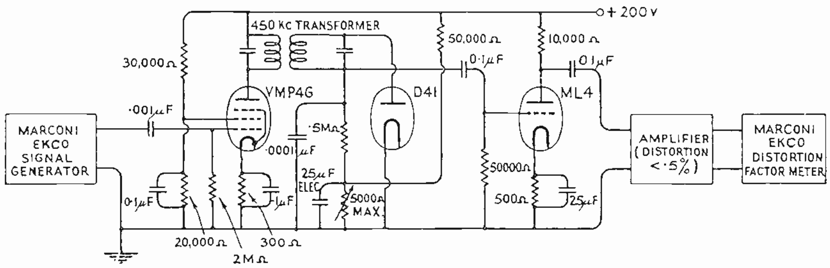


FIG. 1.

Expression (1) indicates the desirability of making $R_2 >> R_1$. It is not, however, possible always to obtain this relationship, particularly when the detector supplies a power valve directly, for the value of R_2 is limited by the danger of developing "softness" in the power valve.

Further theoretical investigation shows that the above expression is modified by applying positive D.C. bias to the diode anode to

$$K\,\% = \frac{\hat{E} + E_b}{\hat{E}}\left[1 - \eta_D \cdot \frac{R_1}{R_1 + R_2}\right] \times 100 \qquad \text{(2)}$$

where E_b = positive applied bias.

It appears possible from expression (2) that the modulation percentage maximum may be increased for a given value of R_2, R_1 and $\hat{E}$ by applying positive bias to the diode.

The value of K is the modulation percentage at which distortion begins and not the value at which it becomes important. In practice the permissible maximum is always higher than K.

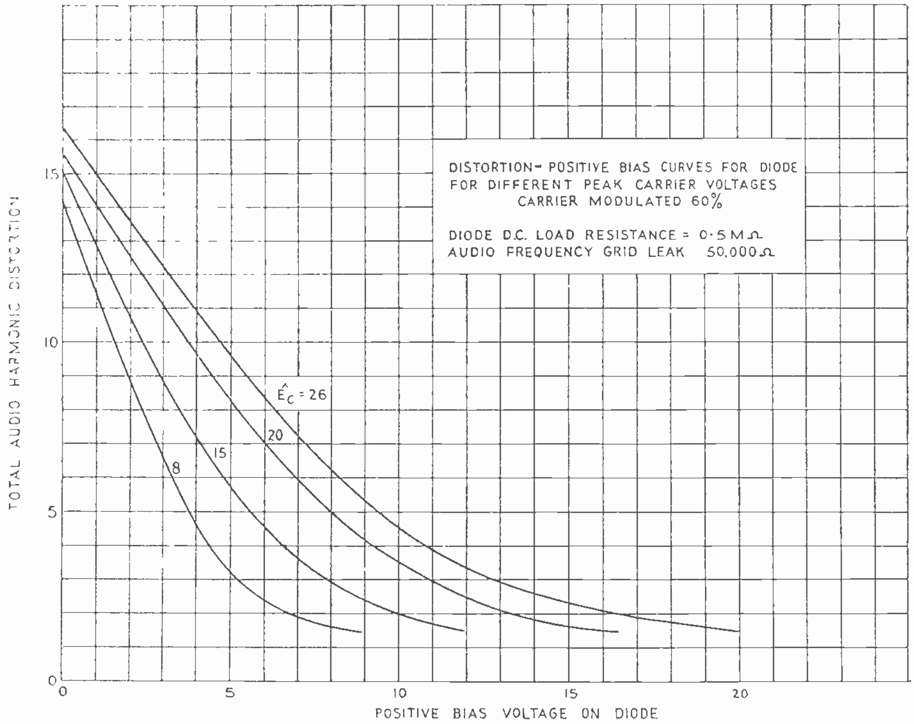


FIG. 2.

To check the theory, apparatus, a diagram of which is given in Fig. 1, was constructed. Distortion in the apparatus other than the diode was reduced to its lowest value and a low ratio of $\frac{R_2}{R_1}\left(\frac{1}{10}\right)$ was chosen in order to exaggerate the distortion. Positive bias was obtained by returning the diode load resistance to a potentiometer system connected across the H.T. supply. Curves of distortion against positive bias are plotted in Fig. 2 for different values of carrier peak voltage $\hat{E}$ and a fixed modulation percentage of 60%. It is quite clear from the curves that distortion can be very considerably reduced by applying positive bias of value approximately $\frac{2}{3}\hat{E}$.

There is a disadvantage to the application of positive bias to a diode detector ; it increases the damping of the tuned circuit supplying the detector. The damping is not constant but increases as the carrier voltage decreases. Experiments showed that it becomes appreciable when the carrier peak voltage is less than half the positive bias.

The effect of this increased damping with reducing carrier voltage is to give suppression of weak signals and noise suppression can be obtained on a receiver by applying positive bias to the diode detector.

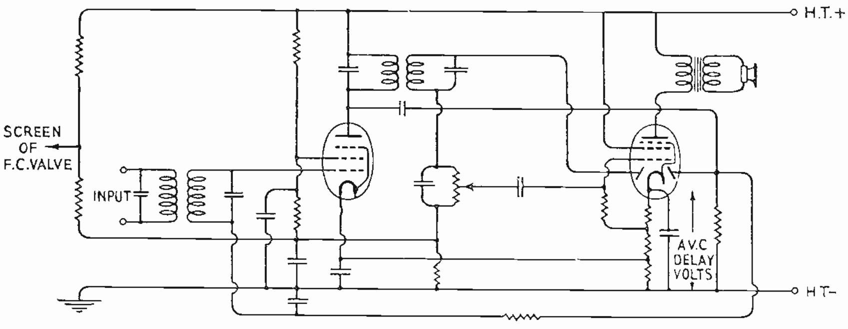


FIG. 3.

A possible method of preventing excessive suppression is shown in Fig. 3. A double diode pentode valve acts as a combined detector, automatic gain control and output valve. Resistance in the cathode circuit provides self bias for the pentode and delay voltage (about 10 volts) for the A.G.C. diode. The delay voltage is neutralised for the detector valve by returning the load resistance to a resistance (R_1) connected in series with the potentiometers supplying the screens of the R.F. valves. The current in the lower limb of these potentiometers varies from approximately 6 to 10 mA (two controlled R.F. stages are assumed) for no A.G.C. to full A.G.C. conditions. By adjusting the resistance R_1 so that the voltage across it exceeds the delay voltage of 10 volts, any degree of noise suppression may be obtained. At the same time, the application of A.G.C. causes the voltage across R_1 to rise so that increasing positive bias is applied to the detector as the carrier voltage increases. If the initial bias on the R.F. controlled stages is included in the delay voltage, the change of positive bias may be further increased by about 1.5 volts.

RADIATION RESISTANCE OF A HORIZONTAL DIPOLE ABOVE EARTH

The calculation of the radiation resistance of a vertical antenna above earth was completed many years ago by van der Pol. His method was to evaluate the field strength at a distant point, and then to integrate the Poynting vector over the surface of a sphere. As far as the writer is aware, however, the analysis for a horizontal dipole above earth has never been carried to a similar conclusion. An expression for it was stated by McPherson in "Electrical Communications," April, 1937, but no attempt was made to evaluate the integral involved, though an approximate value was indicated.

In this article, the analysis is completed, and an expression for the radiation resistance is given in a form suitable for practical calculation.

CONSIDER a radiating element at A (Fig. 1) in which the direction of the current is perpendicular to the plane of the paper and from above. Let A′ be the image of A in the earth DE, and FE the direction of the magnetic field at a point E on the earth.

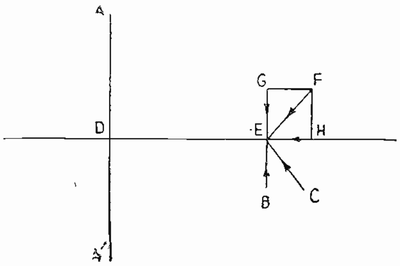


FIG. 1.

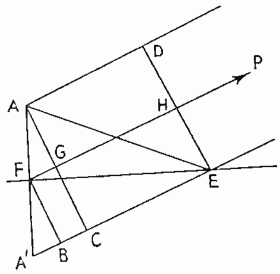


FIG. 2.

The variation of the magnetic field at E will produce currents in the earth, and, assuming it is perfectly conducting, these currents will be such that :—

1. There is no component of magnetic induction perpendicular to the earth.
2. There is no horizontal component of the magnetic field in the earth.

If GE represents the vertical component of field strength, then an equal and opposite field strength BE must be produced by the induced currents. This secures that there is no component of magnetic induction perpendicular to the earth. Other surface currents (perpendicular to the plane of the paper, in Fig. 1) will flow to produce just under the surface of the earth a field strength EH equal and opposite to the horizontal component HE of the field strength produced by the radiating element at A, thus making the total horizontal field strength in the earth zero. These surface currents will produce at the surface of the earth a field strength equal to HE. Therefore the total effect of the earth is to produce a field strength CE. This field strength would be produced by a radiating element at A′, in which the direction of the current is opposite to the direction of current in A. Thus, as far as the magnetic field is concerned, the effect of the earth can be replaced by the effect of a radiating element at A′, the image of A, in which the direction of current is reversed. Considerations similar to the above show that the effect also holds for the electric field, so that, in the following analysis, the earth will be replaced by a radiating element at A′.

Let P (Fig. 2) be a point far removed from A and its image A′ in the earth FE. It is convenient to measure phases relative to F. The phase of A is advanced by an amount corresponding to the distance FG and the phase of A′ retarded by an amount corresponding to a distance AE — FH = A′E — BE = A′B = FG.

If the field strength produced at P by A is δf, then that produced at P by A′ is $-\delta f$, since the current in A′ is reversed.

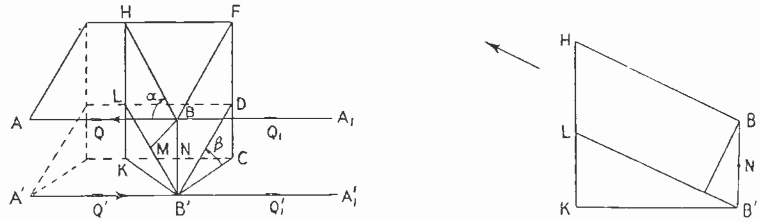


FIG. 3.

Taking phases relative to F, the total field strength produced at P is

$$\delta F = \delta f e^{+jqFG} - \delta f e^{-jqFG} = 2\, j \delta f \sin (qFG)$$

where $q = \frac{2\pi}{\lambda}$ and λ is the wavelength of the oscillations at A.

The factor j means that the field strength is advanced in phase relative to F by 90°. As this does not affect the radiated power it will be ignored in what follows.

In Fig. 3, let ABA_1 be a dipole and $A'B'A'_1$ its image in the earth. BH is the direction of a distant point P from the dipole system.

In the figure, let $\widehat{ABH} = \alpha$ and $\widehat{CB'D} = \beta$.

Taking phases relative to N, the mid-point of BB′, the field strength produced at a point P by an element dx at Q, and its image at Q′ is, when PN is large, so that P subtends the same angle at all points of AA_1 and $A'A'_1$,

$$\delta f_1 = \frac{30}{r} q\, I_x\, dx \sin \alpha\, e^{jqx \cos \alpha}\, 2 \sin q \left(\tfrac{1}{2} MB'\right)$$

where PN $= r$, I_x is the current in the element dx and QB = Q′B′ $= x$. The field strength produced by elements at Q_1 and Q'_1 where $Q_1B = Q'_1B' = -x$ is,

$$\delta f_2 = \frac{30}{r} q\, I_x\, dx \sin \alpha\, e^{-jqx \cos \alpha}\, 2 \sin q \left(\tfrac{1}{2} MB'\right).$$

Assuming a sinusoidal distribution of current,

$$I_x = I_o \sin q\, (l + a - x)$$

where I_o is the value of current at an antinode, l the half length of the dipole, = AB, and $I_o \sin qa$ the magnitude of the current at A, the total field strength at P is,

$$\delta F = \delta f_1 + \delta f_2 = \frac{30}{r} q\, I_o \sin \alpha\; 2 \sin \left(\tfrac{1}{2} qMB'\right)\, 2 \cos (qx \cos \alpha) \sin q\, (l + a - x)\, dx.$$

If H is the height of the dipole above ground, $BB' = 2H$.

$$MB' = BB' \sin \widehat{MBB'} = 2H \sin \widehat{LB'K} = 2H.\frac{LK}{LB'} = \frac{\frac{2H.DC}{DB'}}{\sin\alpha} = 2H \sin\alpha \sin\beta.$$

$$F = \int_{x=o}^{x=l} \delta F = \frac{30}{r} \cdot q\, I_o \sin\alpha\, 2 \sin (q\, H \sin\alpha \sin\beta) \int_o^l 2\cos (qx \cos\alpha) \sin q\, (l + a - x)\, dx.$$

$$= \frac{120}{r} \cdot \frac{I_o}{\sin\alpha} \sin (qH \sin\alpha \sin\beta) \Big[(\cos qa \cos (ql \cos\alpha) - \cos q\, (l + a) - \cos\alpha \sin qa \sin (ql \cos\alpha) \Big]$$

The power P is given by

$$P = \frac{1}{4\pi 30} \int_o^\pi \int_o^\pi F^2 r^2 \sin\alpha\, .\, d\alpha\, .\, d\beta.$$

and the radiation resistance by

$$R = \frac{P}{I_o^2}$$

Putting $qH = \phi$, $ql = \theta$ and $qa = \psi$.

$$R = \frac{120}{\pi} \int_o^\pi \int_o^\pi \frac{\sin^2 (\phi \sin\alpha \sin\beta)}{\sin^2\alpha} \Big[\cos\psi \cos (\theta\cos\alpha) - \cos (\theta + \psi) - \cos\alpha \sin\psi \sin (\theta \cos\alpha) \Big]^2 \sin\alpha . d\alpha . d\beta.$$

Now $\sin^2 (\phi \sin\alpha \sin\beta) = \frac{1}{2} [1 - \cos (2\phi \sin\alpha \sin\beta)]$

and $\int_o^\pi \cos (K \sin\beta)\, d\beta = \pi J_o (K)$ where $J_o (K)$ is Bessel's function of the first kind and order zero, and K is a constant.

Putting $\cos\alpha = z$, and carrying out the integration with respect to β,

$$R = 120 \int_o^1 \frac{[1 - J_o (2\phi \sqrt{1 - z^2})]}{1 - z^2} [\cos\psi \cos\theta z - \cos (\theta + \psi) - z \sin\psi \sin\theta z]^2\, dz. \qquad (1)$$

By replacing $J_o (2\phi\sqrt{1 - z^2})$ by its expansion

$$1 - \frac{(\phi\sqrt{1 - z^2})^2}{(1!)^2} + \frac{(\phi\sqrt{1 - z^2})^4}{(2!)^2} \dots$$

it is seen that the following integrals will occur.

(a) $\int_o^1 (1 - z^2)^n \cos^2 \theta z dz$ (b) $\int_o^1 (1 - z^2)^n \cos \theta z dz.$

(c) $\int_o^1 (1 - z^2)^n\, dz.$ (d) $\int_o^1 (1 - z^2)^n z^2 \sin^2 \theta z dz.$

(e) $\int_o^1 (1 - z^2)^n z \sin \theta z dz.$ (f) $\int_o^1 (1 - z^2)^n z \sin \theta z \cos \theta z dz.$

Let $f\,(n) = \int_o^1 (1 - z^2)^n \cos \theta z dz.$

$$\frac{df(n)}{d\theta} = \int_o^1 - (1 - z^2)^n z \sin \theta z dz.$$

$$= \left[\frac{(1 - z^2)^{n+1}}{2(n+1)} \sin \theta z\right] - \frac{\theta}{2(n+1)} \int_0^1 \cos \theta z \, (1 - z^2)^{n+1} \, dz.$$

$$= - \frac{\theta}{2(n+1)} f(n+1), \text{ and putting } n \text{ for } n+1$$

$$f(n) = - 2n \frac{1}{\theta} \frac{d}{d\theta} f(n-1) = -2n(-2n-2) \left[\frac{1}{\theta} \frac{d}{d\theta}\right]^2 f(n-2)$$

$$= (-1)^n \; 2^n \; n! \left[\frac{1}{\theta} \frac{d}{d\theta}\right]^n f(0)$$

$$f(0) = \int_0^1 \cos \theta z dz = \frac{\sin \theta}{\theta} = - \frac{1}{\theta} \frac{d}{d\theta} \cos \theta$$

Therefore, $$f(n) = (-1)^{n-1} 2^n \; n! \left[\frac{1}{\theta} \frac{d}{d\theta}\right]^{n+1} \cos \theta \quad \ldots \quad (2)$$

Let $$F(n) = \int_0^1 (1-z^2)^n \, dz = \left[z(1-z^2)^n\right]_0^1 - \int_0^1 z(1-z^2)^{n-1} (-2z) n \, dz$$

$$= 2n \int_0^1 z^2(1-z^2)^{n-1} \, dz = - 2n \int_0^1 (1-z^2-1) \, (1-z^2)^n \, dz.$$

$$= - 2n \, F(n) + 2n \, F(n-1)$$

whence, $$F(n) = \frac{2n}{2n+1} F(n-1) = \frac{2n}{2n+1} \cdot \frac{2(n-1)}{2n-1} F(n-2)$$

$$= \frac{2n.2n-2 \; \ldots\ldots\ldots 2}{2n+1.2n-1 \ldots\ldots 3} \int_0^1 (1-z^2)^0 \, dz.$$

$$= \frac{2n. \; 2n. \; 2n-2. \; 2n-2 \ldots\ldots\ldots 2. \; 2.}{2n+1. \; 2n. \; 2n-1. \; 2n-2 \ldots 3. \; 2.}$$

$$= \frac{2^{2n} \, (n!)^2}{2n+1} \quad \ldots \quad (3)$$

Now $$\int_0^1 (1-z^2)^n \cos^2 \theta z \, dz = \tfrac{1}{2} \int_0^1 (1-z^2)^n \, dz + \tfrac{1}{2} \int_0^1 (1-z^2)^n \cos 2\theta z \, dz \ldots \quad (4)$$

and $$\int_0^1 (1-z^2)^n z^2 \sin^2 \theta z \, dz = -\tfrac{1}{2} \int_0^1 (1-z^2)^n (1-z^2-1)(1-\cos 2\theta z) \, dz$$

$$= \tfrac{1}{2} \int_0^1 (1-z^2)^n dz - \tfrac{1}{2} \int_0^1 (1-z^2)^{n+1} dz + \tfrac{1}{2} \int_0^1 (1-z^2)^{n+1} \cos 2\theta z dz - \tfrac{1}{2} \int_0^1 (1-z^2)^n \cos 2\theta z \, dz \quad (5)$$

Also $$\int_0^1 (1-z^2) z \sin \theta z \, dz = \left[\frac{-1}{2(n+1)} \cdot (1-z^2)^{n+1} \sin \theta z\right]_0^1$$

$$+ \frac{\theta}{2(n+1)} \int_0^1 (1-z^2)^{n+1} \cos \theta z \, dz = \frac{\theta}{2(n+1)} \int_0^1 (1-z^2)^{n+1} \cos \theta z \, dz \quad (6)$$

and $$\int_0^1 (1-z^2)^n z \sin \theta z \cos \theta z = \tfrac{1}{2} \int_0^1 (1-z^2)^n z \sin 2\theta z \, dz = \frac{\theta}{2(n+1)} \int_0^1 (1-z^2)^{n+1} \cos 2\theta z dz \quad (7)$$

in Fig. 8. It became apparent that the central reflector was not only having the desired effect in reflecting back side radiations and hence preventing the overlapping of lobes of opposite sign which had led to the formation of false equisignal courses, but this new addition to the arrays was tending to sharpen the main beams and to reduce the overlap between them.

It was therefore found possible to dispense with the two outer arrays altogether, thus simplifying the feeder system, bringing the geometrical centres of the beam radiation nearer together and conserving the space needed on any given site. Fig. 9 shows this arrangement.

Fig. 10 shows the resulting radiation diagram in approximate outline and Fig. 11 shows a sketch of the aerial and reflector system showing feeder arrangements, etc.

From the results given above it is evident that the objectives aimed at, viz., high quality discrimination, reduction of interference area and of power required for a given length of path, have been realised in the "Marconitrack" system, and an important step forward in the technique of ultra short wave directive systems has been achieved.

J. M. FURNIVAL.

RETIREMENT OF MR. H. M. DOWSETT

Mr. H. M. Dowsett, Principal of the Marconi School of Wireless Communication, and Editor of THE MARCONI REVIEW since its inception in 1928, has retired from the Marconi Company on reaching the age limit after 40 years' service with the Company.

Mr. Dowsett has been succeeded by Mr. L. E. Q. Walker as Editor of THE MARCONI REVIEW.

THE REDUCTION OF DIODE DETECTOR DISTORTION BY POSITIVE BIAS

This note shows that positive bias may be used in a diode detector to eliminate coupling circuit distortion. For satisfactory operation the bias should be controlled by the signal input and a method of achieving this is described.

A CAPACITANCE-RESISTANCE coupling is necessary to transfer the audio frequency voltages developed across a diode detector load resistance to the grid of the first A.F. amplifier. The grid leak resistance across the amplifier grid circuit affects the performance of the detector by limiting the maximum modulation percentage which can be accepted without distortion. It can be proved for a linear diode detector that the modulation percentage at which distortion occurs is

$$K\,\% = \left[1 - \eta_D \cdot \frac{R_1}{R_1 + R_2}\right] \times 100 \quad \ldots \quad (1)$$

where R_1 = diode load resistance.

R_2 = A.F. amplifier grid leak resistance.

$\eta_D = \dfrac{I_m R_1}{\hat{E}}$ the diode detection efficiency,

and I_m = mean D.C. current through R_1 for a given carrier voltage $\hat{E} \sin wt\,(1 + K \sin pt)$

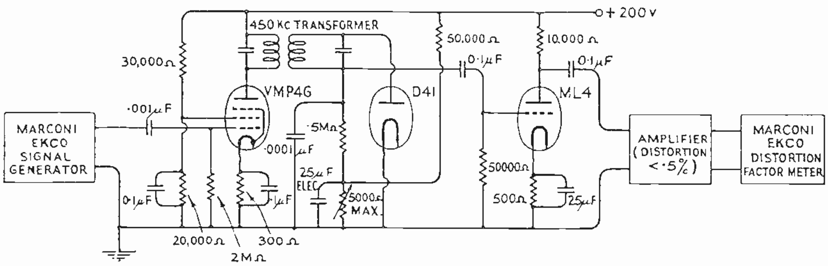


FIG. 1.

Expression (1) indicates the desirability of making $R_2 >> R_1$. It is not, however, possible always to obtain this relationship, particularly when the detector supplies a power valve directly, for the value of R_2 is limited by the danger of developing "softness" in the power valve.

Further theoretical investigation shows that the above expression is modified by applying positive D.C. bias to the diode anode to

$$K\,\% = \frac{\hat{E} + E_b}{\hat{E}}\left[1 - \eta_D \cdot \frac{R_1}{R_1 + R_2}\right] \times 100 \quad \ldots \quad (2)$$

where E_b = positive applied bias.

It appears possible from expression (2) that the modulation percentage maximum may be increased for a given value of R_2, R_1 and $\hat{E}$ by applying positive bias to the diode.

The value of K is the modulation percentage at which distortion begins and not the value at which it becomes important. In practice the permissible maximum is always higher than K.

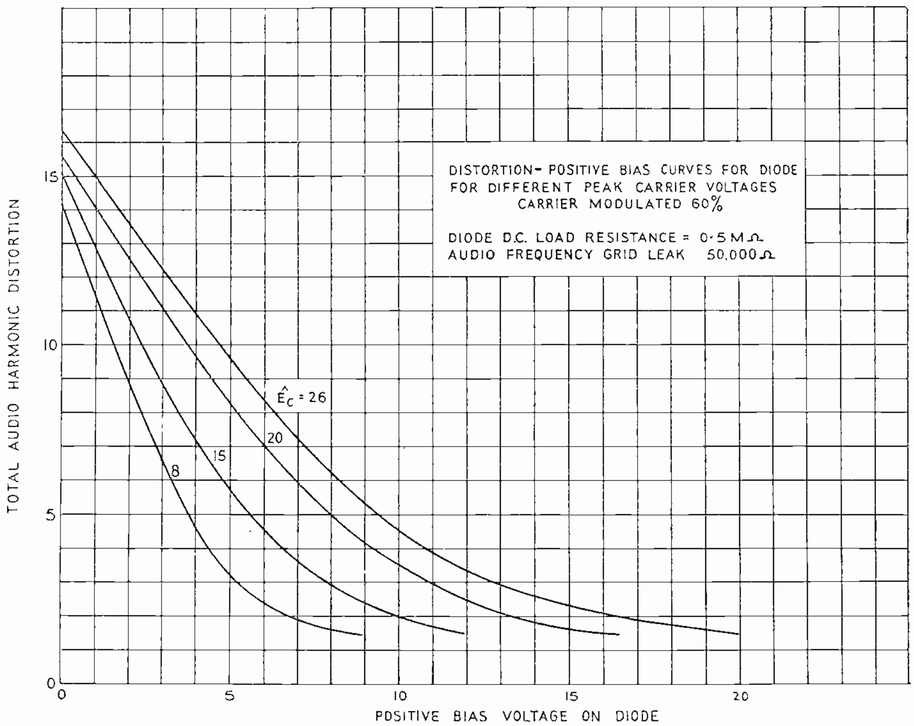


FIG. 2.

To check the theory, apparatus, a diagram of which is given in Fig. 1, was constructed. Distortion in the apparatus other than the diode was reduced to its lowest value and a low ratio of $\frac{R_2}{R_1}\left(\frac{1}{10}\right)$ was chosen in order to exaggerate the distortion. Positive bias was obtained by returning the diode load resistance to a potentiometer system connected across the H.T. supply. Curves of distortion against positive bias are plotted in Fig. 2 for different values of carrier peak voltage $\hat{E}$ and a fixed modulation percentage of 60%. It is quite clear from the curves that distortion can be very considerably reduced by applying positive bias of value approximately $\frac{2}{3}\hat{E}$.

There is a disadvantage to the application of positive bias to a diode detector ; it increases the damping of the tuned circuit supplying the detector. The damping is not constant but increases as the carrier voltage decreases. Experiments showed that it becomes appreciable when the carrier peak voltage is less than half the positive bias.

The effect of this increased damping with reducing carrier voltage is to give suppression of weak signals and noise suppression can be obtained on a receiver by applying positive bias to the diode detector.

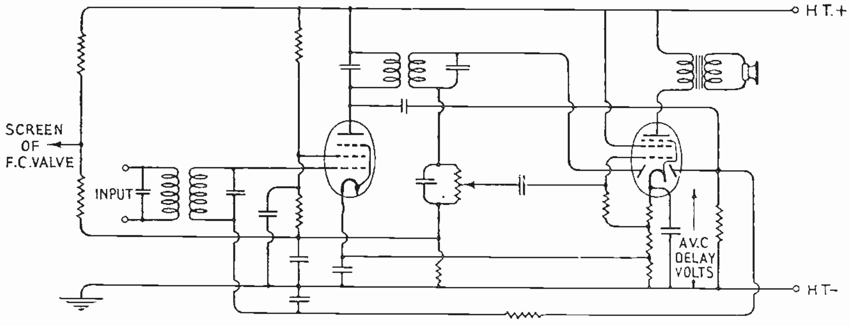


FIG. 3.

A possible method of preventing excessive suppression is shown in Fig. 3. A double diode pentode valve acts as a combined detector, automatic gain control and output valve. Resistance in the cathode circuit provides self bias for the pentode and delay voltage (about 10 volts) for the A.G.C. diode. The delay voltage is neutralised for the detector valve by returning the load resistance to a resistance (R_1) connected in series with the potentiometers supplying the screens of the R.F. valves. The current in the lower limb of these potentiometers varies from approximately 6 to 10 mA (two controlled R.F. stages are assumed) for no A.G.C. to full A.G.C. conditions. By adjusting the resistance R_1 so that the voltage across it exceeds the delay voltage of 10 volts, any degree of noise suppression may be obtained. At the same time, the application of A.G.C. causes the voltage across R_1 to rise so that increasing positive bias is applied to the detector as the carrier voltage increases. If the initial bias on the R.F. controlled stages is included in the delay voltage, the change of positive bias may be further increased by about 1.5 volts.

RADIATION RESISTANCE OF A HORIZONTAL DIPOLE ABOVE EARTH

The calculation of the radiation resistance of a vertical antenna above earth was completed many years ago by van der Pol. His method was to evaluate the field strength at a distant point, and then to integrate the Poynting vector over the surface of a sphere. As far as the writer is aware, however, the analysis for a horizontal dipole above earth has never been carried to a similar conclusion. An expression for it was stated by McPherson in "Electrical Communications," April, 1937, but no attempt was made to evaluate the integral involved, though an approximate value was indicated.

In this article, the analysis is completed, and an expression for the radiation resistance is given in a form suitable for practical calculation.

CONSIDER a radiating element at A (Fig. 1) in which the direction of the current is perpendicular to the plane of the paper and from above. Let A′ be the image of A in the earth DE, and FE the direction of the magnetic field at a point E on the earth.

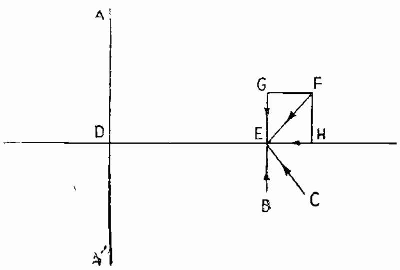


FIG. 1.

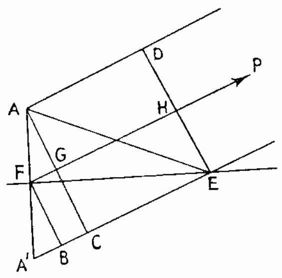


FIG. 2.

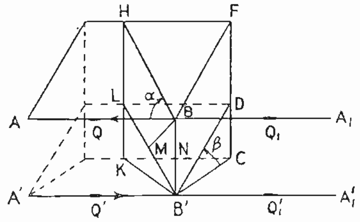
The variation of the magnetic field at E will produce currents in the earth, and, assuming it is perfectly conducting, these currents will be such that :—

1. There is no component of magnetic induction perpendicular to the earth.
2. There is no horizontal component of the magnetic field in the earth.

If GE represents the vertical component of field strength, then an equal and opposite field strength BE must be produced by the induced currents. This secures that there is no component of magnetic induction perpendicular to the earth. Other surface currents (perpendicular to the plane of the paper, in Fig. 1) will flow to produce just under the surface of the earth a field strength EH equal and opposite to the horizontal component HE of the field strength produced by the radiating element at A, thus making the total horizontal field strength in the earth zero. These surface currents will produce at the surface of the earth a field strength equal to HE. Therefore the total effect of the earth is to produce a field strength CE. This field strength would be produced by a radiating element at A′, in which the direction of the current is opposite to the direction of current in A. Thus, as far as the magnetic field is concerned, the effect of the earth can be replaced by the effect of a radiating element at A′, the image of A, in which the direction of current is reversed. Considerations similar to the above show that the effect also holds for the electric field, so that, in the following analysis, the earth will be replaced by a radiating element at A′.

Let P (Fig. 2) be a point far removed from A and its image A′ in the earth FE. It is convenient to measure phases relative to F. The phase of A is advanced by an amount corresponding to the distance FG and the phase of A′ retarded by an amount corresponding to a distance AE — FH = A′E — BE = A′B = FG.

If the field strength produced at P by A is δf, then that produced at P by A′ is $-\delta f$, since the current in A′ is reversed.



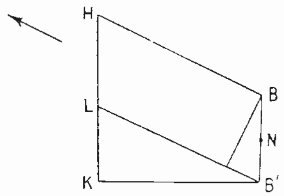


FIG. 3.

Taking phases relative to F, the total field strength produced at P is

$$\delta F = \delta f e^{+jq\mathrm{FG}} - \delta f e^{-jq\mathrm{FG}} = 2\, j \delta f \sin (q\mathrm{FG})$$

where $q = \dfrac{2\pi}{\lambda}$ and λ is the wavelength of the oscillations at A.

The factor j means that the field strength is advanced in phase relative to F by 90°. As this does not affect the radiated power it will be ignored in what follows.

In Fig. 3, let $\mathrm{ABA_1}$ be a dipole and $\mathrm{A'B'A'_1}$ its image in the earth. BH is the direction of a distant point P from the dipole system.

In the figure, let $\widehat{\mathrm{ABH}} = \alpha$ and $\widehat{\mathrm{CB'D}} = \beta$.

Taking phases relative to N, the mid-point of BB′, the field strength produced at a point P by an element dx at Q, and its image at Q′ is, when PN is large, so that P subtends the same angle at all points of $\mathrm{AA_1}$ and $\mathrm{A'A'_1}$,

$$\delta f_1 = \frac{30}{r}\, q\, \mathrm{I}_x\, dx \sin \alpha\, e^{jqx \cos \alpha}\, 2 \sin q\, (\tfrac{1}{2}\,\mathrm{MB'})$$

where $\mathrm{PN} = r$, I_x is the current in the element dx and $\mathrm{QB} = \mathrm{Q'B'} = x$. The field strength produced by elements at $\mathrm{Q_1}$ and $\mathrm{Q'_1}$ where $\mathrm{Q_1B} = \mathrm{Q'_1\,B'} = -x$ is,

$$\delta f_2 = \frac{30}{r}\, q\, \mathrm{I}_x\, dx \sin \alpha\, e^{-jqx \cos \alpha}\, 2 \sin q\, (\tfrac{1}{2}\,\mathrm{MB'}).$$

Assuming a sinusoidal distribution of current,

$$\mathrm{I}_x = \mathrm{I}_o \sin q\, (l + a - x)$$

where I_o is the value of current at an antinode, l the half length of the dipole, = AB, and $\mathrm{I}_o \sin qa$ the magnitude of the current at A, the total field strength at P is,

$$\delta \mathrm{F} = \delta f_1 + \delta f_2 = \frac{30}{r}\, q\, \mathrm{I}_o \sin \alpha\; 2 \sin (\tfrac{1}{2}\, q\mathrm{MB'})\; 2 \cos (qx \cos \alpha)\, \sin q\, (l + a - x)\, dx.$$

If H is the height of the dipole above ground, $BB' = 2H$.

$$MB' = BB' \sin \widehat{MBB'} = 2H \sin \widehat{LB'K} = 2H.\frac{LK}{LB'} = \frac{\dfrac{2H.DC}{DB'}}{\sin\alpha} = 2H \sin\alpha \sin\beta.$$

$$F = \int_{x=0}^{x=l} \delta F = \frac{30}{r} \cdot q\, I_o \sin\alpha\, 2 \sin(q H \sin\alpha \sin\beta) \int_0^l 2\cos(qx\cos\alpha) \sin q(l + a - x)\, dx.$$

$$= \frac{120}{r} \cdot \frac{I_o}{\sin\alpha} \sin(qH\sin\alpha\sin\beta) \Big[(\cos qa \cos(ql\cos\alpha) - \cos q(l+a) - \cos\alpha \sin qa \sin(ql\cos\alpha) \Big]$$

The power P is given by

$$P = \frac{1}{4\pi\, 30} \int_0^\pi \int_0^\pi F^2 r^2 \sin\alpha \, . \, d\alpha \, . \, d\beta.$$

and the radiation resistance by

$$R = \frac{P}{I_o^2}$$

Putting $qH = \phi$, $ql = \theta$ and $qa = \psi$.

$$R = \frac{120}{\pi} \int_0^\pi \int_0^\pi \frac{\sin^2(\phi\sin\alpha\sin\beta)}{\sin^2\alpha} \Big[\cos\psi\cos(\theta\cos\alpha) - \cos(\theta+\psi) - \cos\alpha\sin\psi\sin(\theta\cos\alpha) \Big]^2 \sin\alpha\, d\alpha\, d\beta.$$

Now $\sin^2(\phi \sin\alpha \sin\beta) = \frac{1}{2}[1 - \cos(2\phi \sin\alpha \sin\beta)]$

and $\int_0^\pi \cos(K \sin\beta)\, d\beta = \pi J_o(K)$ where $J_o(K)$ is Bessel's function of the first kind and order zero, and K is a constant.

Putting $\cos\alpha = z$, and carrying out the integration with respect to β,

$$R = 120 \int_0^1 \frac{1 - J_o(2\phi\sqrt{1-z^2})}{1-z^2} \Big[\cos\psi\cos\theta z - \cos(\theta+\psi) - z\sin\psi\sin\theta z \Big]^2 dz \qquad (1)$$

By replacing $J_o(2\phi\sqrt{1-z^2})$ by its expansion

$$1 - \frac{(\phi\sqrt{1-z^2})^2}{(1!)^2} + \frac{(\phi\sqrt{1-z^2})^4}{(2!)^2} - \ldots.$$

it is seen that the following integrals will occur.

(a) $\int_0^1 (1-z^2)^n \cos^2\theta z\, dz$ (b) $\int_0^1 (1-z^2)^n \cos\theta z\, dz$

(c) $\int_0^1 (1-z^2)^n\, dz$. (d) $\int_0^1 (1-z^2)^n z^2 \sin^2\theta z\, dz$.

(e) $\int_0^1 (1-z^2)^n z \sin\theta z\, dz$. (f) $\int_0^1 (1-z^2)^n z \sin\theta z \cos\theta z\, dz$

Let $f(n) = \int_0^1 (1-z^2)^n \cos\theta z\, dz$.

$$\frac{df(n)}{d\theta} = \int_0^1 - (1-z^2)^n z \sin\theta z\, dz.$$

$$= \left[\frac{(1 - z^2)^{n+1}}{2\,(n + 1)} \sin\,\theta z\right] - \frac{\theta}{2\,(n + 1)} \int_0^1 \cos\,\theta z\,(1 - z^2)^{n+1}\,dz.$$

$$= -\frac{\theta}{2\,(n + 1)}\,f\,(n + 1), \text{ and putting } n \text{ for } n + 1$$

$$f(n) \;=\; -\,2n\,\frac{1}{\theta}\,\frac{d}{d\theta} f\,(n-1) \;=\; -2n\,(-2n-2)\left[\frac{1}{\theta}\,\frac{d}{d\theta}\right]^2 f(n-2)$$

$$= (-1)^n\;\;2^n\;\;n!\;\left[\frac{1}{\theta}\,\frac{d}{d\theta}\right]^n f(o)$$

$$f(o) \;=\; \int_0^1 \cos\,\theta z dz \;=\; \frac{\sin\,\theta}{\theta} \;=\; -\,\frac{1}{\theta}\,\frac{d}{d\theta} \cos\,\theta$$

Therefore, $$f(n) \;=\; (-1)^{n-1}\,2^n\;\,n!\left[\frac{1}{\theta}\,\frac{d}{d\theta}\right]^{n+1} \cos\;\theta \quad \ldots \quad \ldots \quad \ldots \quad \ldots \tag{2}$$

Let $$F(n) \;=\; \int_0^1 (1-z^2)^n\,dz = \Big[z(1-z^2)^n\Big]_0^1 - \int_0^1 z(1-z^2)^{n-1}\,(-2z)n\;dz$$

$$= 2n \int_0^1 z^2(1-z^2)^{n-1}\;dz \;=\; -\;2n \int_0^1 (1-z^2-1)\;(1-z^2)^n\;dz.$$

$$= -\;2n\;F(n) \;+\; 2n\;F\;(n-1)$$

whence, $$F(n) \;=\; \frac{2n}{2n+1}\;F(n-1) \;=\; \frac{2n}{2n+1}\;\cdot\;\frac{2(n-1)}{2n-1}\;F(n-2)$$

$$= \frac{2n.2n-2\;\;\ldots\ldots\ldots 2}{2n+1.2n-1\ldots\ldots 3}\int_0^1 (1-z^2)^0\;\;dz.$$

$$= \frac{2n.\;2n.\;2n-2.\;2n-2.\ldots\ldots\ldots 2.\;2.}{2n+1.\;\;2n.\;\;2n-1.\;\;2n-2\ldots.3.\;2.}$$

$$= \frac{2^{2n}\;(n!)^2}{2n+1} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \tag{3}$$

Now $$\int_0^1 (1-z^2)^n \cos^2\theta z\,dz \;=\; \tfrac{1}{2}\int_0^1 (1-z^2)^n\,dz + \tfrac{1}{2}\int_0^1 (1-z^2)^n \cos\,2\theta z\,dz\,.\,. \tag{4}$$

and $$\int_0^1 (1-z^2)^n\,z^2 \sin^2\theta z\,dz \;=\; -\tfrac{1}{2}\int_0^1 (1-z^2)^n\,(1-z^2-1)\;(1-\cos\,2\theta z)\;dz$$

$$= \tfrac{1}{2}\int_0^1 (1-z^2)^n dz - \tfrac{1}{2}\int_0^1 (1-z^2)^{n+1} dz + \tfrac{1}{2}\int_0^1 (1-z^2)^{n+1}\cos\,2\theta z dz - \tfrac{1}{2}\int_0^1 (1-z^2)^n \cos\,2\theta z\,dz \tag{5}$$

Also $$\int_0^1 (1-z^2)z\sin\,\theta z\,dz = \left[\frac{-1}{2(n+1)}\;\cdot\;(1-z^2)^{n+1}\sin\,\theta z\right]_0^1$$

$$+\;\frac{\theta}{2(n+1)}\int_0^1 (1-z^2)^{n+1}\cos\,\theta z\,dz = \frac{\theta}{2(n+1)}\int_0^1 (1-z^2)^{n+1}\cos\,\theta z\,dz \tag{6}$$

and $$\int_0^1 (1-z^2)^n z\sin\,\theta z\cos\,\theta z = \tfrac{1}{2}\int_0^1 (1-z^2)^n z\sin\,2\theta z\,dz = \frac{\theta}{2(n+1)}\int_0^1 (1-z^2)^{n+1}\cos 2\theta z dz \tag{7}$$

On replacing $J_0(2\phi\sqrt{1-z^2})$ in Eq. (1) by its expansion we have

$$R = 120\sum_1^\infty \int_0^1 \frac{(-1)^{n-1}\phi^{2n}(1-z^2)^{n-1}}{(n!)^2}$$

$$\left[\begin{array}{l}\cos^2\psi\cos^2\theta z + \cos^2(\theta+\psi) + z^2\sin^2\psi\sin^2\theta z - 2\cos\psi\cos(\theta+\psi)\cos\theta z \\ + 2\sin\psi\cos(\theta+\psi)z\sin\theta z - 2\sin\psi\cos\psi\, z\sin\theta z\cos\theta z\end{array}\right] dz$$

This reduces by Eqs. 4, 5, 6 and 7 to

$$R = 120\sum_1^\infty \int_0^1 \frac{(-1)^{n-1}\phi^{2n}(1-z^2)^{n-1}}{(n!)^2}$$

$$\Big\{\left[\tfrac{1}{2}+\cos^2(\theta+\psi)\right] + \left[(1-z^2)\tfrac{1}{2}\sin^2\psi\right] - \left[\cos\theta z.\ 2\cos\psi\cos(\theta+\psi)\right]$$

$$+ \left[\cos 2\theta z.\ \tfrac{1}{2}\cos 2\psi\right] + \left[(1-z^2)\cos\theta z\sin\psi\cos(\theta+\psi)\frac{\theta}{n}\right]$$

$$- \left[(1-z^2)\cos 2\theta z.\ \tfrac{1}{2}\sin^2\psi\right] - \left[(1-z^2)\cos 2\theta z\sin 2\psi\ .\ \frac{\theta}{2n}\right]\Big\}\,dz \qquad (8)$$

The first term reduces, by Eq. 3, to

$$60\,(1+2\cos^2(\theta+\psi))\sum_1^\infty \frac{\phi^{2n}\,2^{2n-2}\,(n-1)!\,(-1)^{n-1}}{(2n-1)!\,(n!)}$$

$$= 60\left[(1+2\cos^2(\theta+\psi))\right]\sum_1^\infty \frac{(2\phi)^{2n}(-1)^{n-1}}{2n\cdot 2n!}$$

Now $$\frac{1-\cos x}{x} = \frac{x}{2!} - \frac{x^3}{4!} + \ldots\ldots$$

Therefore $$\int_0^{2\phi}\frac{1-\cos x}{x}dx = \frac{(2\phi)^2}{2\cdot 2!} - \frac{(2\phi)^4}{4\cdot 4!} + \ldots\ldots = \sum_1^\infty \frac{(2\phi)^{2n}(-1)^{n-1}}{2n\cdot 2n!}$$

Now $$\int_0^x \frac{1-\cos x}{x}dx = \log x + .5772 - Ci(x) = S_1(x),$$ where $Ci(x)$ and $S_1(x)$ are tabulated functions (Jahnke–Emde, Funktionentafeln, p. 7[illegible])

Thus, the first term of Eq. 8, is

$$60\,(1+2\cos^2(\theta+\psi))\,S_1(2\phi)$$

The second term is $-60\sin^2\psi\sum_1^\infty \frac{\phi^{2n}(-1)^{n-1}}{(n!)^2}\cdot\frac{2^{2n}(n!)^2}{(2n+1)!}$ by Eq. (3)

$$= -\frac{60\sin^2\psi}{2\phi}\left[2\phi - \frac{(2\phi)^3}{3!} + \frac{(2\phi)^5}{5!} - \ldots\ldots - 2\phi\right]$$

$$= -60\sin^2\psi\left[\frac{\sin 2\phi - 2\phi}{2\phi}\right]$$

The third term is $-240\cos\psi\cos(\theta+\psi)\sum_1^\infty\int_0^1 \frac{(-1)^{n-1}\phi^{2n}}{(n!)^2}(1-z^2)^{n-1}\cos\theta z\,dz,$

$$= 240\cos\psi\cos(\theta+\psi)\sum_1^\infty \frac{\phi^{2n}\,2^{n-1}}{n\cdot n!}\left[\frac{1}{\theta}\frac{d}{d\theta}\right]^n\cos\theta \text{ by Eq. (2)}$$

Let $\theta^2 = 4\,w$. $\cos\theta = \cos 2\sqrt{w}$

$$\frac{2}{\theta}\frac{d}{d\theta} = \frac{2}{\theta}\frac{dw}{d\theta}\cdot\frac{d}{dw} = \frac{d}{dw}$$

The third term becomes,

$$120\cos\psi\cos(\theta+\psi)\sum_{1}^{\infty}\frac{\phi^{2n}}{n\,n!}\left(\frac{d}{dw}\right)^{n}\cos 2\sqrt{w}$$

Now Taylor's theorem may be written,

$$\frac{f(h+x)-f(x)}{x} = \sum_{1}^{\infty}\frac{x^{r-1}}{r!}\left(\frac{d}{dh}\right)^{r} f(h)$$

Therefore $\quad \sum_{1}^{\infty}\frac{x^{r}}{r\,\lfloor r}\left(\frac{d}{dh}\right)^{r} f(h) = \int_{0}^{x}\frac{f(h+x)-f(h)}{x}\,dx$

Put $\quad x = \phi^2,\ h = w,\ f(x) = \cos 2\sqrt{x}$

Then $\quad \sum_{1}^{\infty}\frac{\phi^{2n}}{n\,n!}\left(\frac{d}{dw}\right)^{n}\cos 2\sqrt{w} = \int_{0}^{\phi^2}\frac{\cos 2\sqrt{\phi^2+w}-\cos 2\sqrt{w}}{\phi^2}\,d\phi^2$

$$= \int_{1}^{\sqrt{1+\frac{4\phi^2}{\theta^2}}}\left(\cos\theta t-\cos\theta\right)\left(\frac{1}{t+1}+\frac{1}{t-1}\right)dt \text{ where } w+\phi^2 = wt^2$$

$$= \int_{2}^{\left(\sqrt{1+\frac{4\phi^2}{\theta^2}}+1\right)}\frac{\cos\theta\,(t_1-1)-\cos\theta}{t_1}\,dt_1 + \int_{0}^{\left(\sqrt{1+\frac{4\phi^2}{\theta^2}}-1\right)}\frac{\cos\theta\,(t_2+1)-\cos\theta}{t_2}\,dt_2$$

with $\quad t+1 = t_1,\ t-1 = t_2$

$$= \cos\theta\,[S_1(2\theta) - S_1(\sqrt{\theta^2+4\phi^2}+\theta) - S_1(\sqrt{\theta^2+4\phi^2}-\theta)]$$
$$+ \sin\theta\,[Si(\sqrt{\theta^2+4\phi^2}+\theta) - Si(\sqrt{\theta^2+4\phi^2}-\theta) - Si(2\theta)]$$

where $\quad Si\,\theta = \int_{0}^{\theta}\frac{\sin x}{x}\,dx$

Calling this expression $N(\theta)$, the third term can be written,

$$120\cos\psi\cos(\theta+\psi)\,N(\theta) \quad \ldots \quad (9)$$

The fourth term is $60\cos 2\psi\cdot\sum_{1}^{\infty}\int_{0}^{1}\frac{(-1)^{n-1}\phi^{2n}(1-z^2)^{n-1}}{(n!)^2}\cos 2\theta z\,dz$

By comparison with the preceding term this is seen to be

$$-30\cos 2\psi\,N(2\theta)$$

The fifth term is $120\sin\psi\cos(\theta+\psi)\cdot\theta\sum_{1}^{\infty}\int_{0}^{1}\frac{(-1)^{n-1}\phi^{2n}}{n\,(n!)^2}(1-z^2)^{n}\cos\theta z\,dz$

$$= 120\sin\psi\cos(\theta+\psi)\cdot\theta\sum_{1}^{\infty}\frac{\phi^{2n}\,2^{n}}{n\,n!}\left(\frac{1}{\theta}\frac{d}{d\theta}\right)^{n+1}\cos\theta \text{ by Eq. (2)}$$

By comparison with the expression for the third term and Eq. (9) this is seen to be

$$120\sin\psi\cos(\theta+\psi)\cdot\theta\,\frac{1}{\theta}\frac{d}{d\theta}\,[N(\theta)] = 120\sin\psi\cos(\theta+\psi)\,\frac{d}{d\theta}\,[N(\theta)]$$

$$= 120\sin\psi\cos(\theta+\psi)\begin{bmatrix}\cos\theta\,[Si(\sqrt{\theta^2+4\phi^2}+\theta) - Si(\sqrt{\theta^2+4\phi^2}-\theta) - Si\,2\theta] \\ -\sin\theta\,[S_1(2\theta) - S_1(\sqrt{\theta^2+4\phi^2}+\theta) - S_1(\sqrt{\theta^2+4\phi^2}-\theta)]\end{bmatrix}$$

The sixth term is $60 \sin^2\psi \sum_{1}^{\infty} \frac{\phi^{2n}\, 2^n}{n!} \left[\frac{1}{2\theta} \frac{d}{d(2\theta)}\right]^{n+1} \cos 2\theta$

$$= 30 \sin^2\psi \sum_{1}^{\infty} \frac{\phi^{2n}}{n!} \left(\frac{d}{dw}\right)^{n+1} \cos 2\sqrt{w} \text{ where } w = \theta^2$$

By Taylor's theorem, $f(x + h^2) - f(x) = \sum_{1}^{\infty} \frac{h^{2n}}{n!} \left(\frac{d}{dx}\right)^n f(x)$

therefore, $\sum_{1}^{\infty} \frac{h^{2n}}{n!} \left(\frac{d}{dx}\right)^{n+1} f(x) = \frac{d}{dx} [f(x + h^2) - f(x)]$

Put $f(x) = \cos 2\sqrt{x}$, $h = \phi$, $x = w$,

Then the sixth term is $30 \sin^2\psi \frac{d}{dw} [\cos 2\sqrt{w + \phi^2} - \cos 2\sqrt{w}]$

$$= 30 \sin^2\psi \left[\frac{\sin 2\theta}{\theta} - \frac{\sin 2\sqrt{\theta^2 + \phi^2}}{\sqrt{\theta^2+\phi^2}}\right] \text{ since } w = \theta^2$$

The last term is $\frac{-60}{2} \sin 2\psi \cdot 2\theta \sum_{1}^{\infty} \int_0^1 \frac{(-1)^{n-1} (1 - z^2)^n \phi^{2n} \cos 2\theta z}{n. (n!)^2} dz$

By comparison with the expression for the fifth term, this is seen to be

$$-\tfrac{1}{4}\, 120 \sin 2\psi \begin{bmatrix} \cos 2\theta\, [S_1 (\sqrt{4\phi^2 + 4\phi^2} + 2\theta) - Si (\sqrt{4\phi^2 + 4\phi^2} - 2\theta) - Si (4\theta)] \\ - \sin 2\theta\, [S_1 (4\theta) - S_1 (\sqrt{4\theta^2 + 4\phi^2} + 2\theta) - S_1 (\sqrt{4\theta^2 + 4\phi^2} - 2\theta)] \end{bmatrix}$$

Collecting the terms it is found that,

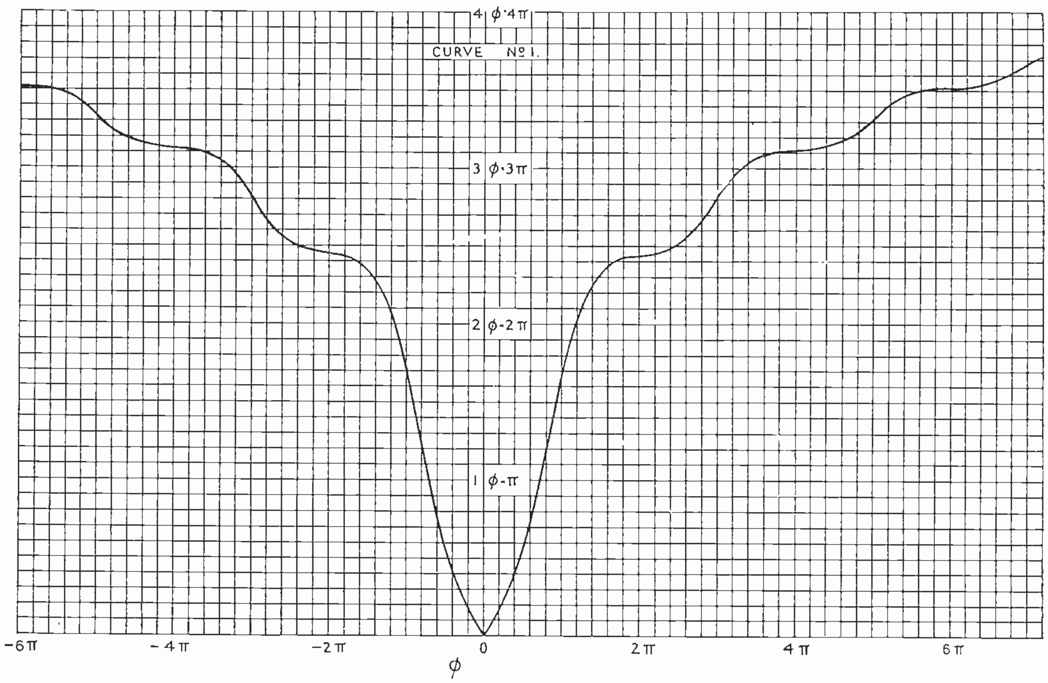
$$R = 30 \left\{ \begin{array}{l} 2 S_1 (2\phi) [1+2\cos^2 (\theta+\psi)] + 2 \sin^2\psi \left[\frac{\sin 2\phi - 2\phi}{2\phi} + \frac{\sin 2\theta}{2\theta} - \frac{\sin 2\sqrt{\theta^2+\phi^2}}{2\sqrt{\theta^2+\phi^2}}\right] \\ + 4 \cos^2 (\theta+\psi) [S_1 (2\theta) - S_1 (\sqrt{\theta^2 + 4\phi^2} + \theta) - S_1 (\sqrt{\theta^2 + 4\phi^2} - \theta)] \\ + 2 \sin 2 (\theta + \psi) [Si (\sqrt{\theta^2 + 4\phi^2} + \theta) - Si (\sqrt{\theta^2 + 4\phi^2} - \theta) - Si (2\theta)] \\ - \cos 2 (\theta + \psi) [S_1 (4\theta) - S_1\, 2 (\sqrt{\theta^2 + \phi^2} + \theta) - S_1\, 2 (\sqrt{\theta^2 + \phi^2} - \theta)] \\ - \sin 2 (\theta + \psi) [Si\, 2 (\sqrt{\theta^2 + \phi^2} + \theta) - Si\, 2 (\sqrt{\theta^2 + \phi^2} - \theta) - Si (4\theta)] \end{array} \right\}$$

This probably represents the simplest form into which the expression for the radiation resistance can be put. It is clearly much simpler when $\psi = 0$ and $\theta = \frac{\pi}{2}$ or π, which are the most commonly occurring cases.

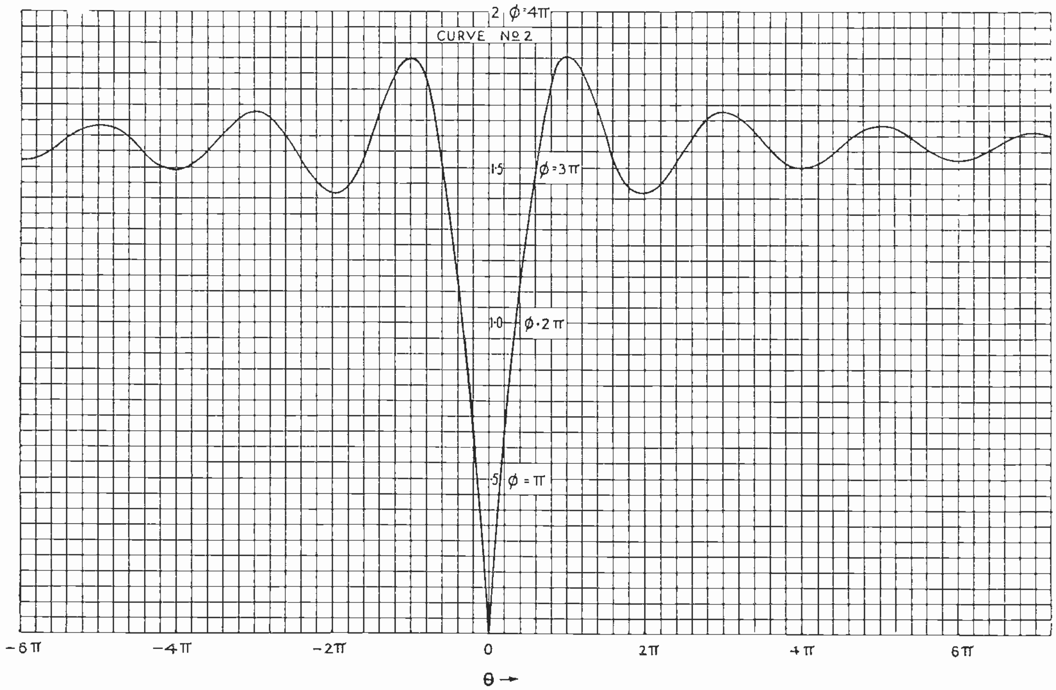
On letting ϕ tend to infinity, that is, removing the dipole from the earth, the radiation resistance becomes

$$R = 30 \left\{ \begin{array}{l} 4 \cos^2 (\theta + \psi) S_1(2\theta) - 2 \sin 2 (\theta + \psi) Si (2\theta) - \cos 2 (\theta + \psi) S_1 (4\theta) \\ + \sin 2 (\theta + \psi) Si (4\theta) + 2 \sin^2\psi \left[\frac{\sin 2\theta - 2\theta}{2\theta}\right] \end{array} \right\}$$

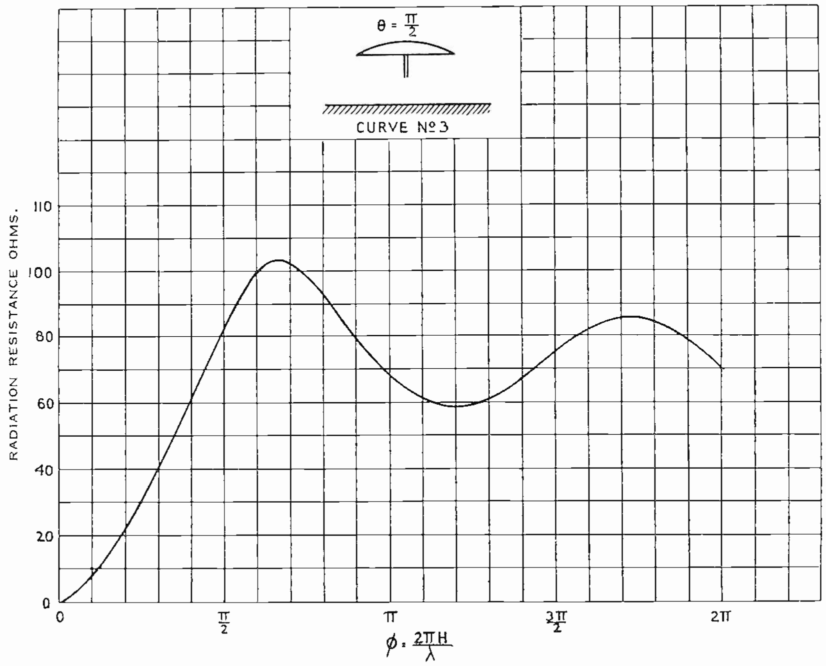
which is the well known expression for the radiation resistance of a dipole in free space.



Curve for determining the value of $[S_1(\sqrt{\theta^2+4\phi^2}+\theta) + S_1(\sqrt{\theta^2+4\phi^2}-\theta) - S_1(2\theta)]$



Curve for determining the value of $[Si(2\theta) + Si(\sqrt{\theta^2+4\phi^2}-\theta) - Si(\sqrt{\theta^2+4\phi^2}+\theta)]$


θ = π/2
CURVE Nº3
RADIATION RESISTANCE OHMS.
110
100
80
60
40
20
0
0
π/2
π
3π/2
2π
φ = 2πH/λ

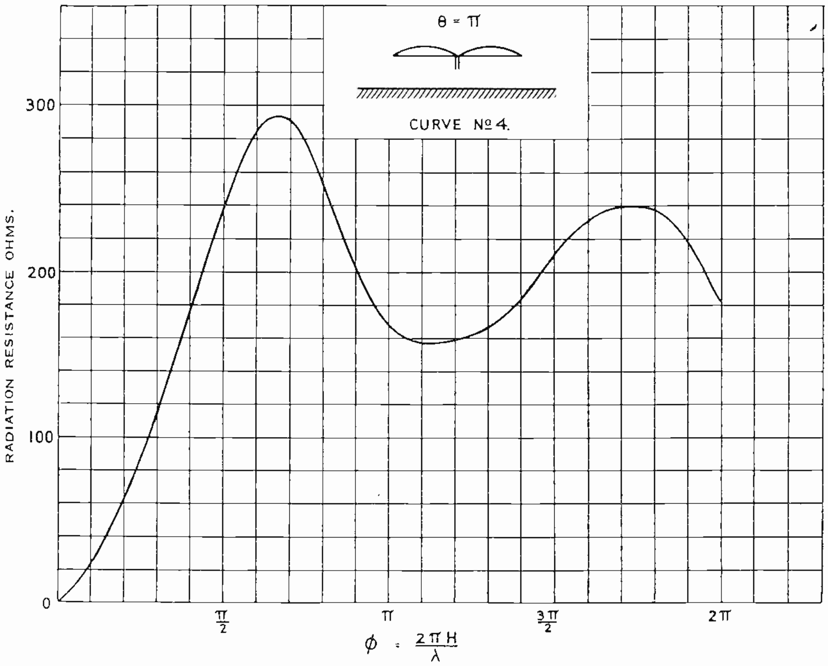

θ = π
CURVE Nº4.
RADIATION RESISTANCE OHMS.
300
200
100
0
π/2
π
3π/2
2π
φ = 2πH/λ

TABLE I.

$$S_1 x = \log_e x + .5772157 - Ci\,x$$

x	S_1 (x)	x	S_1 (x)	x	S_1 (x)	x	S_1 (x)	x	S_1 (x)
0.0	0.00000	5.0	2.37669	10.0	2.92527	15.0	3.23899	20.0	3.52853
0.1	0.00249	5.1	2.38994	10.1	2.94327	15.1	3.25090	20.1	3.53173
0.2	0.00998	5.2	2.40113	10.2	2.96050	15.2	3.26308	20.2	3.53535
0.3	0.02241	5.3	2.41044	10.3	2.97688	15.3	3.27552	20.3	3.53946
0.4	0.03973	5.4	2.41801	10.4	2.99234	15.4	3.28814	20.4	3.54402
0.5	0.06185	5.5	2.42402	10.5	3.00688	15.5	3.30087	20.5	3.54905
0.6	0.08866	5.6	2.42866	10.6	3.02045	15.6	3.31363	20.6	3.55456
0.7	0.12002	5.7	2.43210	10.7	3.03300	15.7	3.32641	20.7	3.56049
0.8	0.15579	5.8	2.43452	10.8	3.04457	15.8	3.33911	20.8	3.56687
0.9	0.19578	5.9	2.43610	10.9	3.05514	15.9	3.35167	20.9	3.57368
1.0	0.23981	6.0	2.43704	11.0	3.06467	16.0	3.36401	21.0	3.58085
1.1	0.28766	6.1	2.43749	11.1	3.07323	16.1	3.37612	21.1	3.58840
1.2	0.33908	6.2	2.43764	11.2	3.08083	16.2	3.38790	21.2	3.59629
1.3	0.39384	6.3	2.43766	11.3	3.08749	16.3	3.39932	21.3	3.60446
1.4	0.45168	6.4	2.43770	11.4	3.09322	16.4	3.41032	21.4	3.61288
1.5	0.51233	6.5	2.43792	11.5	3.09814	16.5	3.42088	21.5	3.62155
1.6	0.57549	6.6	2.43847	11.6	3.10225	16.6	3.43096	21.6	3.63037
1.7	0.64088	6.7	2.43947	11.7	3.10561	16.7	3.44050	21.7	3.63935
1.8	0.70820	6.8	2.44106	11.8	3.10828	16.8	3.44947	21.8	3.64842
1.9	0.77713	6.9	2.44335	11.9	3.11038	16.9	3.45788	21.9	3.65751
2.0	0.84739	7.0	2.44643	12.0	3.11190	17.0	3.46568	22.0	3.66662
2.1	0.91865	7.1	2.45040	12.1	3.11301	17.1	3.47288	22.1	3.67568
2.2	0.99060	7.2	2.45534	12.2	3.11370	17.2	3.47945	22.2	3.68465
2.3	1.06295	7.3	2.46130	12.3	3.11412	17.3	3.48543	22.3	3.69348
2.4	1.13540	7.4	2.46834	12.4	3.11429	17.4	3.49077	22.4	3.70216
2.5	1.20764	7.5	2.47649	12.5	3.11436	17.5	3.49553	22.5	3.71059
2.6	1.27939	7.6	2.48577	12.6	3.11437	17.6	3.49969	22.6	3.71879
2.7	1.35038	7.7	2.49619	12.7	3.11438	17.7	3.50330	22.7	3.72670
2.8	1.42035	7.8	2.50775	12.8	3.11453	17.8	3.50639	22.8	3.73427
2.9	1.48903	7.9	2.52044	12.9	3.11484	17.9	3.50895	22.9	3.74153
3.0	1.55620	8.0	2.53423	13.0	3.11540	18.0	3.51107	23.0	3.74838
3.1	1.62163	8.1	2.54906	13.1	3.11628	18.1	3.51276	23.1	3.75483
3.2	1.68511	8.2	2.56491	13.2	3.11754	18.2	3.51404	23.2	3.76089
3.3	1.74646	8.3	2.56171	13.3	3.11924	18.3	3.51500	23.3	3.76651
3.4	1.80552	8.4	2.59938	13.4	3.12142	18.4	3.51568	23.4	3.77170
3.5	1.86211	8.5	2.61786	13.5	3.12414	18.5	3.51610	23.5	3.77644
3.6	1.91613	8.6	2.63704	13.6	3.12745	18.6	3.51633	23.6	3.78072
3.7	1.96745	8.7	2.65686	13.7	3.13134	18.7	3.51645	23.7	3.78459
3.8	2.01600	8.8	2.67721	13.8	3.13587	18.8	3.51648	23.8	3.78801
3.9	2.06170	8.9	2.69799	13.9	3.14104	18.9	3.51648	23.9	3.79101
4.0	2.10449	9.0	2.71909	14.0	3.14688	19.0	3.51660	24.0	3.79360
4.1	2.14438	9.1	2.74042	14.1	3.15338	19.1	3.51661	24.1	3.79582
4.2	2.18131	9.2	2.76186	14.2	3.16054	19.2	3.51685	24.2	3.79767
4.3	2.21535	9.3	2.78332	14.3	3.16835	19.3	3.51727	24.3	3.79917
4.4	2.24648	9.4	2.80468	14.4	3.17677	19.4	3.51790	24.4	3.80036
4.5	2.27479	9.5	2.82583	14.5	3.18583	19.5	3.51879	24.5	3.80129
4.6	2.30033	9.6	2.84669	14.6	3.19545	19.6	3.52002	24.6	3.80197
4.7	2.32317	9.7	2.86713	14.7	3.20564	19.7	3.52156	24.7	3.80243
4.8	2.34344	9.8	2.88712	14.8	3.21630	19.8	3.52348	24.8	3.80271
4.9	2.36124	9.9	2.90651	14.9	2.22746	19.9	3.52578	24.9	3.80288
								25.0	3.80295

TABLE 2.

$$Si\,(x) = \int_{o}^{x} \frac{\sin u}{u}\, du$$

x	*Si* (x)	x	*Si* (x)	x	*Si* (x)	x	*Si* (x)	x	*Si* (x)
0.0	0.00000	5.0	1.54993	10.0	1.65835	15.0	1.61819	20.0	1.54824
0.1	0.09994	5.1	1.53125	10.1	1.65253	15.1	1.62226	20.1	1.55289
0.2	0.19956	5.2	1.51367	10.2	1.64600	15.2	1.62575	20.2	1.55767
0.3	0.29850	5.3	1.49732	10.3	1.63883	15.3	1.62865	20.3	1.56253
0.4	0.39646	5.4	1.48230	10.4	1.63112	15.4	1.63093	20.4	1.56743
0.5	0.49311	5.5	1.46872	10.5	1.62294	15.5	1.63258	20.5	1.57232
0.6	0.58813	5.6	1.45667	10.6	1.61439	15.6	1.63359	20.6	1.57714
0.7	0.68122	5.7	1.44620	10.7	1.60556	15.7	1.63396	20.7	1.58186
0.8	0.77210	5.8	1.43736	10.8	1.59654	15.8	1.63370	20.8	1.58641
0.9	0.86047	5.9	1.43018	10.9	1.58743	15.9	1.63280	20.9	1.59077
1.0	0.94608	6.0	1.42469	11.0	1.57831	16.0	1.63130	21.0	1.59489
1.1	1.02869	6.1	1.42087	11.1	1.56927	16.1	1.62291	21.1	1.59873
1.2	1.10805	6.2	1.41871	11.2	1.56042	16.2	1.62657	21.2	1.60225
1.3	1.18396	6.3	1.41817	11.3	1.55182	16.3	1.62339	21.3	1.60543
1.4	1.25623	6.4	1.41922	11.4	1.54356	16.4	1.61973	21.4	1.60823
1.5	1.32468	6.5	1.42179	11.5	1.53571	16.5	1.61563	21.5	1.61063
1.6	1.38918	6.6	1.42582	11.6	1.52835	16.6	1.61112	21.6	1.61261
1.7	1.44959	6.7	1.43121	11.7	1.52155	16.7	1.60627	21.7	1.61415
1.8	1.50582	6.8	1.43787	11.8	1.51535	16.8	1.60111	21.8	1.61525
1.9	1.55778	6.9	1.44570	11.9	1.50981	16.9	1.59572	21.9	1.61590
2.0	1.60541	7.0	1.45460	12.0	1.50497	17.0	1.59014	22.0	1.61608
2.1	1.64870	7.1	1.46443	12.1	1.50088	17.1	1.58443	22.1	1.61582
2.2	1.68763	7.2	1.47509	12.2	1.49755	17.2	1.57865	22.2	1.61510
2.3	1.72221	7.3	1.48644	12.3	1.49501	17.3	1.57285	22.3	1.61395
2.4	1.75249	7.4	1.49834	12.4	1.49327	17.4	1.56711	22.4	1.61238
2.5	1.77852	7.5	1.51068	12.5	1.49234	17.5	1.56146	22.5	1.61041
2.6	1.80039	7.6	1.52331	12.6	1.49221	17.6	1.55598	22.6	1.60806
2.7	1.81821	7.7	1.53611	12.7	1.49287	17.7	1.55070	22.7	1.60536
2.8	1.83210	7.8	1.54894	12.8	1.49430	17.8	1.54568	22.8	1.60234
2.9	1.84219	7.9	1.56167	12.9	1.49647	17.9	1.54097	22.9	1.59902
3.0	1.84865	8.0	1.57419	13.0	1.49936	18.0	1.53661	23.0	1.59546
3.1	1.85166	8.1	1.58637	13.1	1.50292	18.1	1.53264	23.1	1.59168
3.2	1.85140	8.2	1.59810	13.2	1.50711	18.2	1.52909	23.2	1.58772
3.3	1.84808	8.3	1.60928	13.3	1.51188	18.3	1.52600	23.3	1.58363
3.4	1.84191	8.4	1.61981	13.4	1.51716	18.4	1.52339	23.4	1.57945
3.5	1.83313	8.5	1.62960	13.5	1.52291	18.5	1.52128	23.5	1.57521
3.6	1.82195	8.6	1.63857	13.6	1.52905	18.6	1.51969	23.6	1.57097
3.7	1.80862	8.7	1.64665	13.7	1.53352	18.7	1.51863	23.7	1.56676
3.8	1.79339	8.8	1.65379	13.8	1.54225	18.8	1.51810	23.8	1.56262
3.9	1.77650	8.9	1.65993	13.9	1.54917	18.9	1.51810	23.9	1.55860
4.0	1.75820	9.0	1.66504	14.0	1.55621	19.0	1.51863	24.0	1.55474
4.1	1.73874	9.1	1.66908	14.1	1.56330	19.1	1.51967	24.1	1.55107
4.2	1.71837	9.2	1.67205	14.2	1.57036	19.2	1.52122	24.2	1.54762
4.3	1.69732	9.3	1.67393	14.3	1.57733	19.3	1.52324	24.3	1.54444
4.4	1.67583	9.4	1.67473	14.4	1.58414	19.4	1.52572	24.4	1.54154
4.5	1.65414	9.5	1.67446	14.5	1.59072	19.5	1.52863	24.5	1.53897
4.6	1.63246	9.6	1.67316	14.6	1.59702	19.6	1.53192	24.6	1.53672
4.7	1.61101	9.7	1.67084	14.7	1.60296	19.7	1.53357	24.7	1.53484
4.8	1.58998	9.8	1.66757	14.8	1.60851	19.8	1.53954	24.8	1.53333
4.9	1.56956	9.9	1.66338	14.9	1.61360	19.9	1.54378	24.9	1.53221
								25.0	1.53148

The chief modification to this formula, due to the presence of the earth, is to replace the functions $S_1\,(2\theta)$ and $Si\,(2\theta)$ by

$$[S_1\,(2\theta) - S_1\,(\sqrt{\theta^2+4\phi^2}+\theta) - S_1\,(\sqrt{\theta^2+4\phi^2}-\theta)] \text{ and}$$
$$[Si\,(2\theta) + Si\,(\sqrt{\theta^2+4\phi^2}-\theta) - Si\,(\sqrt{\theta^2+4\phi^2}+\theta)] \text{ respectively.}$$

These expressions are the eyesore of the formula, but a simple and speedy graphical method for evaluating them exists.

In curves Nos. 1 and 2, the electrical length of the dipole θ is the abscissa. The ordinate is marked in two scales, one the electrical height ϕ and the other the magnitude of the required expression.

Place one point of a pair of compasses or dividers on the point $(\theta, 0)$ and the other on the point $(0, \phi)$, and with this radius and centre $(\theta, 0)$ mark two points on the θ axis. At these points erect two ordinates of the curve. The sum of these ordinates on the *magnitude* scale gives $S_1\,(\sqrt{\theta^2+4\phi^2}+\theta) + S_1\,(\sqrt{\theta^2+4\phi^2}-\theta)$ in the case of curve No. 1, while their difference on the *magnitude* scale gives $Si\,(\sqrt{\theta^2+4\phi^2}+\theta) - Si\,(\sqrt{\theta^2+4\phi^2}-\theta)$ in the case of curve No. 2. In this second case, the sign of the difference is to be taken as positive if the higher ordinate is to the right of the origin, but negative if it is to the left. In both cases the values of $S_1\,(2\theta)$ and $Si\,(2\theta)$ can be read off from their corresponding curves.

Curves Nos. 3 and 4 represent the radiation resistance plotted against electrical height ϕ, for the cases of a half wave and whole wave dipole. It is seen that for heights greater than half a wavelength the resistance fluctuates about the free space value, and soon settles down to the values of 73 and 199 ohms respectively.

Tables 1 and 2 (see P. O. Pedersen, " Radiation from a Vertical Antenna over Flat Perfectly Conducting Earth ") give respectively the values of $S_1 x$ and $Si\,x$ to six figures for values of x from 0 to 25 in steps of .1. The function $S_1 x = \log x + .5772 - Ci x$ is found to be more useful in problems concerning aerials than the more usually tabulated function $Ci x$ as it does not become infinite for small values of x.

L. Lewin.

THE MARCONI SPLIT GAP HORN ARRESTER

The following note describes a type of discharge gap developed by the Marconi Company which has been found of great use in transmitter installations.

IN the employment of horn discharge gaps, for protecting radio frequency equipment from the effects of lightning, a very common experience is that the steep potential gradient of a lightning discharge heavily ionises the region in the vicinity of the gap, which then becomes conductive to the comparatively low radio frequency potential of the equipment to be protected. This condition is highly undesirable, if not actually dangerous, because of the tendency to continued after-discharge on the part of the radio equipment, which becomes virtually shorted to earth.

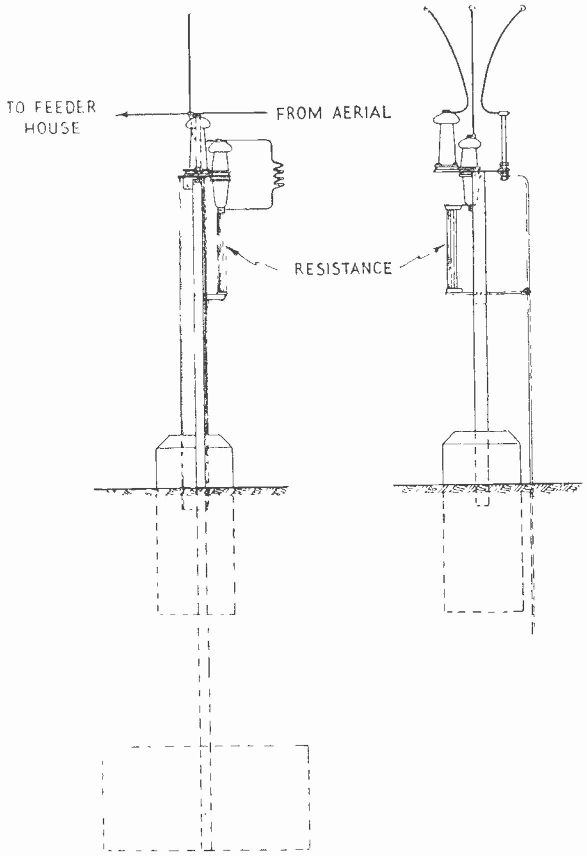


FIG. 1.

In order to quench the after-discharge and to minimise its interference with programmes or routine working, a resistance is inserted in the earth lead of the gap; this resistance generally takes the form of a carbon rod specially manufactured to carry the very heavy transient current of a lightning discharge, during which the impedance of the rod is effectively nil, while offering a high impedance to the subsequent after-discharge maintained by the radio equipment. Unfortunately it frequently happens that, however well made, the resistance rod fails under the repeated shock of numerous discharges, so that either the resistance is broken down or becomes disintegrated; in either case the gap can no longer function as a safety device for the equipment concerned.

In the design of the Marconi Split Gap Arrester, however, this defect is largely if not completely overcome. As its name implies, the main gap is built up of a series of two gaps, the outer electrodes being respectively connected to the live conductor and directly to earth, while the central electrode is connected to earth via the resistance previously mentioned and also via a choke coil. This latter effectively blocks the path of the steep gradient associated with lightning, while offering but little impedance to the after-discharge of radio frequency; thus while the original lightning shock bridges the main gap and avoids, as it were, the resistance rod, the radio frequency after-discharge ultimately selects the secondary gap of the central electrode and choke-cum-resistance path to earth, with the final result that the radio frequency arc collapses because of the damping effect of the resistance. Obviously the main gap must be of such proportions that, once the intensive ionisation caused by lightning diminishes, its impedance to earth despite the lesser ionisation of radio frequency is greater than the impedance of the secondary gap to earth.

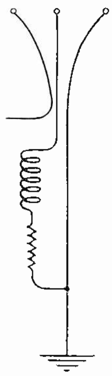


FIG. 2.

Another factor that assists to the desired end is the property of any split or multi-electrode gap, which tends to collapse more quickly than a simple twin electrode gap. This is a well-known fact, but one difficult of explanation apart from capacity effects. It is probable that by far the steepest gradient occurs between the first gap, so that subsequent gaps collapse more readily and in consequence bring about the collapse of the first gap.

When adjusting the Marconi Split Gap Arrester, the first gap, that is the gap between high potential and centre electrodes, should be about one-third the distance between the second gap, that is, the centre and earth electrodes. This is a rough guide to initial adjustment, but obviously final adjustment is a matter of experience on site.

Fig. 1 gives a general idea of the appearance and Fig. 2 shows the circuit of the complete gap.

N. WELLS.

CERTAIN PROPERTIES OF DISSYMMETRICAL ⊤ PURE REACTANCE NETWORKS

(PART I)

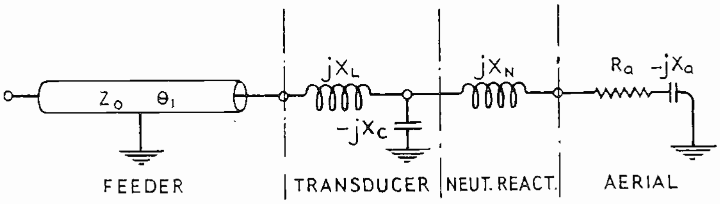
Certain ambiguities may arise in the calculation of voltage and current at the input terminals of an aerial which is coupled to a feeder by means of a dissymmetrical ⊤ reactance network. Two elements of the ⊤ network comprise a transducer, and the other element is an inductance designed to act as an aerial reactance neutraliser at the carrier frequency.

A particular case will be used here to show how easily the ambiguities may fail to be noticed until, after a long and laborious calculation, we are faced with an answer considerably in error. After demonstrating how easily the source of the error may fail to be noticed we shall state what the ambiguities are, and then show how they may be resolved in this and all other cases and the above particular problem will be completed to a correct answer.

The equivalent line technique employed is that due to A. E. Kennelly. The input impedance chart referred to was published in THE MARCONI REVIEW, *Jan.-Feb. and Sept.-Dec.,* 1937.

The article will be concluded in the next issue with a note on the image impedance of dissymmetrical networks, and a group of A matrices which have been generalised for any number of networks has been included.

CONSIDER Fig. 1, which shows an aerial, a transducer and a feeder. The aerial impedance is denoted by $R_a - jX_a$. The reactive component is cancelled at the carrier frequency by an equal and opposite reactance X_N connected in series with the aerial. Thus the transducer, comprised of jX_L and $-jX_C$ is designed to match Z_o at one end and R_a at the other.



f = 183 k.c.
Z_o = 75 ohms, $\theta_1 = j.\ .733$ quarter waves at 183 k.c.
R_a = 1,000 ohms, $X_a = X_N = 400$ ohms, $X_L = 263.5$.
X_c = 284.7.

FIG. 1.

For the purpose of resolving the entire network into an equivalent line in order to compute the magnitude and phase of voltage and current entering the aerial it was considered best to regard X_L, X_C and X_N as comprising a single network since we are not interested in the volts and current entering X_N.

Line Angle of the ⊤ Network.

Thus by the formula for a dyssymmetrical ⊤ network

$$\theta = \tanh^{-1} \sqrt{\frac{Z_A Z_B + Z_A Z_C + Z_B Z_C}{(Z_A + Z_C)(Z_B + Z_C)}}$$

where $Z_A = jX_L$
$Z_B = jX_N$
$Z_C = -jX_C$

we have $\theta_2 = \tanh^{-1} 5.8412$ which by the chart
$= .173 + j\ 1.00$ (hyps & $\frac{1}{4}$-waves)
for the equivalent line angle of the ⊤ network.

Digression on Characteristics of the ⊤ Network.

Now notice that in this result we have an attenuation of 0.173 together with 90 degrees phase-shift in a pure reactance network. The network is therefore acting as a filter in one of its attenuation bands. If L and N are the inductances of reactances X_L and X_N respectively, and C is the capacity of reactance X_C, then this network has three critical frequencies defined by

(i) $\dfrac{\omega^2 LNC}{(L + N)} = 1$

(ii) $\omega^2 LC = 1$

(iii) $\omega^2 NC = 1$

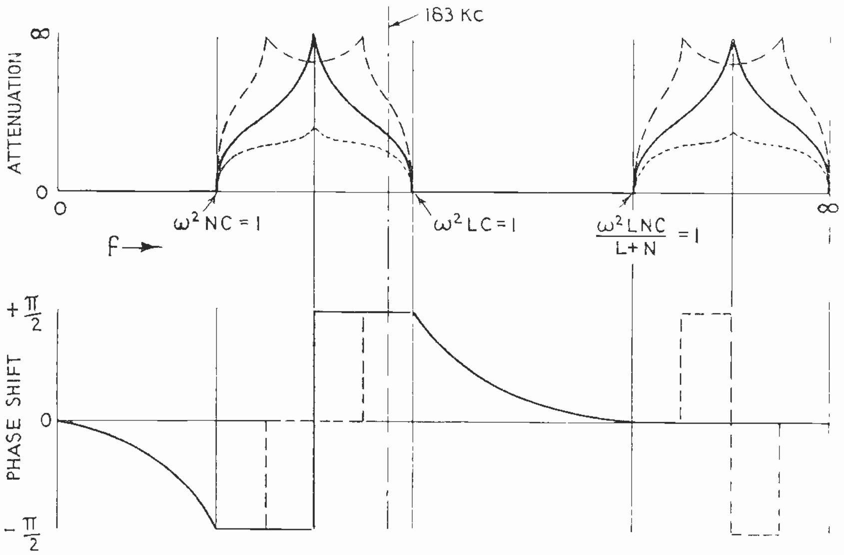


FIG. 2.

The diagrams show more clearly what is the meaning of these conditions. The dotted curves show the possible variation of characteristics. Attenuation is only infinite when

$$\sqrt{\omega C \left(\frac{\omega^2 LN - \left(\frac{L + N}{C}\right)}{(\omega^2 LC - 1)\ (\omega^2 NC - 1)} \right)} = 1$$

Calculation of Image Impedances.

Let *Zs* and *Zd* be short circuit and open circuit impedances respectively.

Then $$Z_{01} = \sqrt{Z_{s1} Z_{d1}} = \sqrt{\left(jX_L + \frac{-jX_C \cdot jX_N}{j(X_N - X_C)}\right) j\,(X_L - X_C)}$$

or $$Z_{01} = \sqrt{\frac{(-X_L X_N + X_C X_L + X_C X_N)\,(X_L - X_C)}{(X_N - X_C)}}$$

$$= \sqrt{\frac{83500 \times (-21.2)}{115.3}}$$

$$= j\sqrt{1.519 \times 10^4}$$

$$= 123.24 \;\underline{/90^\circ}$$

Also $$Z_{02} = \sqrt{Z_{s2} Z_{d2}} = \sqrt{\left(jX_N + \frac{-jX_C \cdot jX_L}{j\,(X_L - X_C)}\right) j\,(X_N - X_C)}$$

$$= \sqrt{\frac{(X_N - X_C)\,(-X_N X_L + X_N X_C + X_L X_C)}{(X_L - X_C)}}$$

$$= \sqrt{\frac{115.3 \times 83500}{-21.2}}$$

$$= \sqrt{-45.45 \times 10^4}$$

$$= j\,674.17$$

$$= 674.2 \;\underline{/90^\circ}$$

Position Angle of Z_a.

Call this θ_D; then since $Z_a = R_a - jX_a = 1{,}000 - j\,400 = 1077 \;\underline{/21^\circ\,48'}$

$$\theta_D = \tanh^{-1} \frac{Z_a}{Z_{02}} = \tanh^{-1}\left(\frac{1077 \;\overline{\backslash 21^\circ\,48'}}{674.2 \;\underline{/90^\circ}}\right)$$

$$= \tanh^{-1} 1.597 \;\overline{\backslash 111^\circ\,48'}$$

By the chart

$$\theta_D = -.174 + j\,1.348 \text{ (hyps \& } \tfrac{1}{4}\text{-waves)}$$

The Equivalent Line.

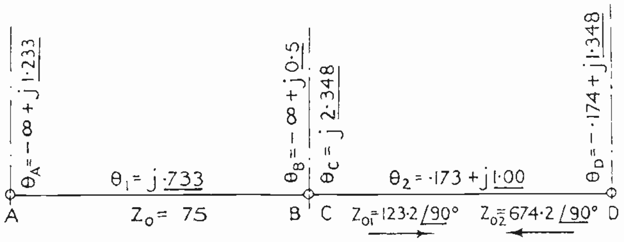


FIG. 3.

Thus we have the data for the equivalent line

$$\theta_C = \theta_2 + \theta_D = .173 + j\,1.00 - .174 + j\,1.348$$

$$= j\,2.348$$

$$\theta_B = \tanh^{-1}\left(\frac{Z_{o1}\tanh\theta_C}{Z_o}\right)$$
$$= \tanh^{-1}\left(\frac{123.2\ \underline{/90^\circ}\ .\ \tanh(j\ \underline{2.348})}{75}\right)$$
$$= \tanh^{-1}\left(\frac{123.2\ \underline{/180^\circ} \times 0.608}{75}\right)$$
$$= \tanh^{-1} 1.00\ \underline{/180^\circ}$$
$$= -\infty + j\ \underline{0.50}\ \text{- - - (hyps \& } \tfrac{1}{4}\text{-waves)}$$

(This value is obtainable from the chart by regarding the chart as the image of its lower, or negative counterpart.)

Examination of these Results.

Now there is an error in these results: For example we know that the impedance of the equivalent line measuring from C to D should match the characteristic impedance of the feeder both in phase and modulus.

That is symbolically

$$Z_o = Z_{o1}\tanh\theta_C \text{ which in our particular case is:}$$
$$75 = 123.2\ \underline{/90} \times \tanh^{-1} j\ \underline{2.348}$$
$$= 123.2\ \underline{/90^\circ} \times 0.608\ \underline{/90^\circ}$$
$$= 75\ \underline{/180^\circ}$$

Thus the impedance of the equivalent line C—D is not matched in phase with surge impedance of the feeder, and in any case there is no such thing as a passive network which can present an angle of 180°. Thus we have seen that it is possible to derive a wrong, if not meaningless, result by what appear at first sight to be correct steps in the process of the calculation. That at least one of these steps is wrong, however, is obvious from the result. The wrong step, in fact, is in the calculation of either Z_{o1} or Z_{o2}. In one of these cases we should have taken the negative root. This brings us to a

Statement of the Ambiguity.

Assuming that in one case the negative root should be taken, there is nothing about the expressions

$$Z_{o1} = \sqrt{\frac{(-X_L X_N + X_C X_L + X_C X_N)(X_L - X_C)}{(X_N - X_C)}}$$

and

$$Z_{o2} = \sqrt{\frac{(-X_L X_N + X_C X_L + X_C X_N)(X_N - X_C)}{(X_L - X_C)}}$$

to indicate which case it should be.

We notice, however, that in our actual problem the factor $(X_L - X_C)$ is $-ve$, while the other two factors are $+ve$. Since therefore this factor appears in the numerator of the expression for Z_{o1} and in the denominator of that for Z_{o2} it would appear that we should write these two expressions as

$$Z_{o1} = +j\sqrt{\frac{(-X_L X_N + X_C X_L + X_C X_N)(X_C - X_L)}{(X_N - X_C)}}$$

and $$Z_{02} = + \frac{1}{j} \cdot \sqrt{\frac{(-X_L X_N + X_C X_L + X_C X_N)\,(X_N - X_C)}{(X_C - X_L)}}$$

or writing
$$a = (-X_L X_N + X_C X_L + X_C X_N)$$
$$b = (X_N - X_C)$$
$$c = (X_C - X_L)$$
$$Z_{01} = + j \sqrt{\frac{ac}{b}}$$
$$Z_{02} = - j \sqrt{\frac{ab}{c}}$$

Let us suppose this is correct and we use these signs in determining the equivalent line :

Position Angle of Z_a.

$$\theta_D = \tanh^{-1}\left(\frac{1077 \,\backslash 21^\circ 48'}{674.2 \,\backslash 90^\circ}\right)$$
$$= \tanh^{-1} 1.597 \,/68^\circ 12'$$
$$= 0.174 + j\,0.65 \text{ - - - (hyps \& } \tfrac{1}{4}\text{-waves)}$$

The Equivalent Line :

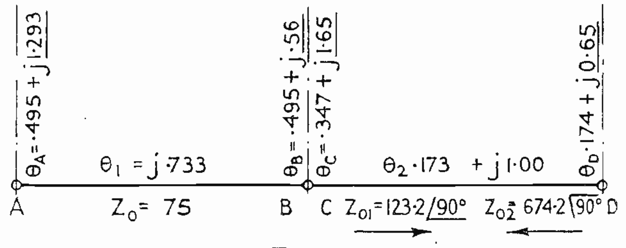


FIG. 4.

$$\theta_C = \theta_D + \theta_2 = .174 + j\,0.65 + .173 + j\,1.00 = .347 + j\,1.65$$
$$\theta_B = \tanh^{-1}\left(\frac{Z_{01}\tanh\theta_C}{Z_o}\right) = \tanh^{-1}\left(\frac{123.2 \,/90^\circ \tanh(.347 + j\,1.65)}{75}\right)$$
$$= \tanh^{-1}\left(\frac{123.2 \,/90^\circ \times .682 \,\backslash 49.8^\circ}{75}\right) = \tanh^{-1} 1.12 \,/40.2^\circ$$
$$= .495 + j\,0.56$$

Now it is quite obvious again that this equivalent line does not represent the network correctly : There is considerable error in the values of θ_A, θ_B, θ_C and θ_D.

Take θ_C for example and test it by the rule that
$$Z_o = Z_{01} \tanh \theta_C$$
we have
$$Z_o = 123.2 \,/90^\circ \cdot \tanh(.347 + j\,1.65)$$
$$= 123.2 \,/90^\circ \times .682 \,\backslash 49.8^\circ$$
$$= 84.1 \,/40.2^\circ$$

which is very different from 75 ohms. Thus the indications suggested on page 30 are misleading. Let us therefore look a little more carefully into this question of the right choice of general number coefficient of the image impedance factors.

Determination of General Number Coefficient.

Using the previous notation we have

$$Z_{d1} = j\,(X_L - X_C) \qquad\qquad Z_{d2} = j\,(X_N - X_C)$$

$$Z_{s1} = jX_L - \frac{jX_C \cdot jX_N}{j(X_N - X_C)} \qquad\qquad Z_{s2} = jX_N - \frac{jX_C \cdot jX_L}{j(X_L - X_C)}$$

$$= j\left[\frac{X_L X_N - (X_L + X_N)\,X_C}{(X_N - X_C)}\right] \qquad\qquad = j\left[\frac{X_L X_N - (X_L + X_N)\,X_C}{(X_L - X_C)}\right]$$

Then when $X_L < X_C < X_N$

$Z_{d1} = -jA_1$, say, and

$Z_{s1} = -jB_1$

Then when $X_L < X_C < X_N$

$Z_{d2} = +jA_2$, say, and

$Z_{s2} = +jB_2$

where A and B are positive numerics

$$\therefore \quad Z_{o1} = \sqrt{(-jA_1)\,(-jB_1)} = -jC_1 \qquad\qquad \therefore \quad Z_{o2} = \sqrt{(jA_2)\,(jB_2)} = +jC_2$$

where C is a positive numeric and is equal to $\sqrt{AB}$

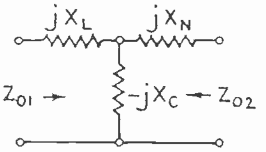


FIG. 5.

The coefficient of C in either case is what we mean by the term "general number coefficient." The condition $X_L < X_C < X_N$ here chosen corresponds with our particular case, but there are obviously other possible conditions. These are set out in tabular form in Table I (page 33), together with the results set out above.

There are many other possible conditions with this network. These will be set out in more compact form together with those of Table I after the solution of our particular problem has been given.

Correct Solution of Particular Problem.

For our particular problem we have

$$\left.\begin{aligned} X_L &= 263.5 \\ X_C &= 284.7 \\ X_N &= 400 \end{aligned}\right\} X_L X_N = 10.53 \times 10^4\,; \quad (X_L + X_N)X_C = 18.88 \times 10^4$$

Hence our case is covered by Condition I (*b*), and by Table I the general number coefficient of C_1 is $-j$, while that of C_2 is $+j$. Therefore

$$Z_{o1} = 123.2\ \backslash 90^\circ \quad \text{and}$$
$$Z_{o2} = 674.2\ /90^\circ$$

Position Angle of Z_a.

$$\text{This is } \theta_D = \tanh^{-1}\frac{Z_a}{Z_{o2}} = \tanh^{-1}\left(\frac{1077\ \backslash 21^\circ\,48'}{674.2\ /90^\circ}\right)$$

$$= \tanh^{-1} 1.598\ \backslash 111^\circ\,48'$$

TABLE I.

CONDITION I: $X_L < X_C < X_N$.	
(*a*) :—$X_L X_N > (X_L + X_N) X_C$	
$\left.\begin{matrix} Z_{d1} = -jA_1 \\ Z_{s1} = +jB_1 \end{matrix}\right\} Z_{o1} = +C_1$	$\left.\begin{matrix} Z_{d2} = +jA_2 \\ Z_{s2} = -jB_2 \end{matrix}\right\} Z_{o2} = +C_2$
(*b*) :—$X_L X_N < (X_L + X_N) X_C$	
$\left.\begin{matrix} Z_{d1} = -jA_1 \\ Z_{s1} = -jB_1 \end{matrix}\right\} Z_{o1} = -jC_1$	$\left.\begin{matrix} Z_{d2} = +jA_2 \\ Z_{s2} = +jB_2 \end{matrix}\right\} Z_{o2} = +jC_2$
CONDITION II: $X_C < X_N < X_L$.	
(*a*) :—$X_L X_N > (X_L + X_N) X_C$.	
$\left.\begin{matrix} Z_{d1} = +jA_1 \\ Z_{s1} = +jB_1 \end{matrix}\right\} Z_{o1} = +jC_1$	$\left.\begin{matrix} Z_{d2} = +jA_2 \\ Z_{s2} = +jB_2 \end{matrix}\right\} Z_{o2} = +jC_2$
(*b*) :—$X_L X_N < (X_L + X_N) X_C$.	
$\left.\begin{matrix} Z_{d1} = +jA_1 \\ Z_{s1} = -jB_1 \end{matrix}\right\} Z_{o1} = +C_1$	$\left.\begin{matrix} Z_{d2} = +jA_2 \\ Z_{s2} = -jB_2 \end{matrix}\right\} Z_{o2} = +C_2$
CONDITION III: $X_N < X_L < X_C$.	
(*a*) :—$X_L X_N > (X_L + X_N) X_C$.	
$\left.\begin{matrix} Z_{d1} = -jA_1 \\ Z_{s1} = -jB_1 \end{matrix}\right\} Z_{o1} = -jC_1$	$\left.\begin{matrix} Z_{d2} = -jA_2 \\ Z_{s2} = -jB_2 \end{matrix}\right\} Z_{o2} = -jC_2$
(*b*) :—$X_L X_N < (X_L + X_N) X_C$.	
$\left.\begin{matrix} Z_{d1} = -jA_1 \\ Z_{s1} = +jB_1 \end{matrix}\right\} Z_{o1} = +C_1$	$\left.\begin{matrix} Z_{d2} = -jA_2 \\ Z_{s2} = +jB_2 \end{matrix}\right\} Z_{o2} = +C_2$

By the chart this is

$$\theta_D = -0.174 + j\,1.348 \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)}$$

The Equivalent Line.

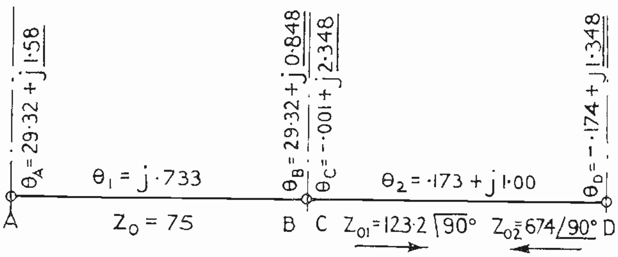


FIG. 6.

$$\begin{aligned}
\theta_C &= \theta_D + \theta_2 = -.174 + j\,1.348 + .173 + j\,1.00 \\
&= -0.001 + j\,2.348 \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)} \\
\theta_B &= \tanh^{-1}\left(\frac{Z_{o1}\tanh\theta_C}{Z_o}\right) = \tanh^{-1}\left(\frac{123.2\ \backslash 90^\circ \cdot \tanh(-0.001 + j2.348)}{75}\right) \\
&= \tanh^{-1}\left(\frac{123.2\ \backslash 90^\circ \cdot 0.612\ /90.15^\circ}{75}\right) \\
&= \tanh^{-1} 1.005\ /0.15^\circ \\
&= 29.32 + j\,0.848 \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)} \\
\theta_A &= \theta_1 + \theta_B = j\,.733 + 29.32 + j\,0.848 \\
&= 29.32 + j\,1.581 \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)}
\end{aligned}$$

This is the correct solution.

Test θ_C by the rule

$$Z_o = Z_{o1} \tanh \theta_C$$

i.e.

$$\begin{aligned}
Z_o &= 123.2\ \backslash 90^\circ \cdot 0.612\ /90.15^\circ \\
&= 75.4\ /0.15^\circ
\end{aligned}$$

That is practically 75 ohms.

Also test θ_A by the rule

$$Z_o = Z_o \tanh \theta_A$$

That is

$$\begin{aligned}
Z_o &= 75 \times \tanh(29.32 + j\,1.58) \\
&= 75 \times 0.9987\ \backslash 0.315^\circ \\
&= 74.9\ \backslash 0.315^\circ
\end{aligned}$$

which also is practically 75 ohms.

Summarising the foregoing we see how easy it is, in approaching such a problem as the one we have chosen, to confuse the general number coefficient of the positive real numerical part of the image impedances of the dissymmetrical network. Also we see how considerably such confusion can affect the results of calculation. In our particular case we knew beforehand what the correct result should be, but it is not difficult to appreciate that there might be many cases in practice in which the correct result would not be known beforehand.

Hence we see the usefulness of Table I which sets out the correct general number coefficient of the image impedances of the particular dissymmetrical network. As already mentioned, these results will be set out in Tables II to V (Part II) in more compact form, together with other cases of dissymmetrical T networks, consisting of pure reactances.

Note on Readings within Negative Image of Chart.

The interpretation of the Input Impedance Chart to yield $\tanh^{-1} 1.598 \overline{\backslash 111^\circ 48'} = (-0.174 + j\,1.348)$ in this example on page 34 introduces us to the resolution of an ambiguity in connection with the chart. If reference is made to our article in THE MARCONI REVIEW, No. 67, Sept.-Dec., 1937, page 8, Ex. (H), the reader will find the following rule given for dealing with cases where the angle ϕ is greater than 90° numerically: Subtract the numerical value of the angle from 180° and use the remainder (after prefixing the sign of the original angle) for the chart in the ordinary way.

To that rule we must now add the following extension. It must be clearly understood that in these cases where the angle is greater than 90° (or less than —90°) the resulting value of α is negative. The practical case here given is an example of all such cases. See also page 30 of this article.

(To be continued.)

PATENT ABSTRACTS

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

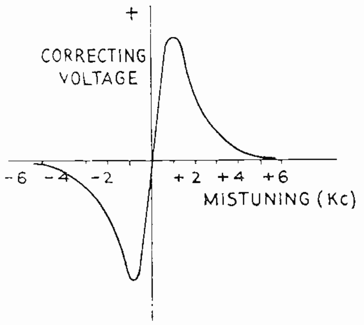
AUTOMATIC FREQUENCY CONTROL SYSTEMS.

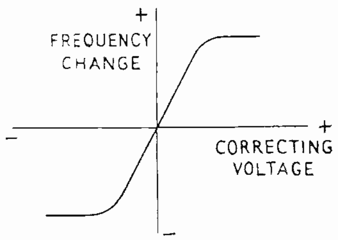
Application date, May 4th, 1937. **No. 495,313.**

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

In discriminator circuits associated with automatic frequency control systems it is usual to arrange for the maximum possible change of correcting voltage between frequency limits ± 5 kc. either side of the intermediate frequency. This tends to cause complete failure to tune a weak station located between two strong ones and may lead to jumping from one station to another, when the desired station signal fades slightly. The first effect may be overcome by disconnecting the A.F.C. when tuning manually, but this does not prevent the latter effect. This specification seeks to correct these faults by reducing the frequency discriminator limits to ± 1 kc. either side of the intermediate frequency (a typical frequency error-correcting volts curve is shown in the first figure) and by using a frequency adjusting circuit which provides frequency correction between certain specified correcting voltage variations and outside these limits causes no variation of oscillator frequency. Such a curve is illustrated in the second figure, and it may be obtained in the case of a valve frequency adjuster by using a short grid base valve, or in the case of a meter operated variable capacitor by specially shaped vanes or sprung stops restricting the travel of the rotor plates.

By this method the A.F.C. system may be left permanently in circuit and satisfactory tuning and holding of a weak station located between two strong stations is obtained.





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BAND PASS FILTERS.

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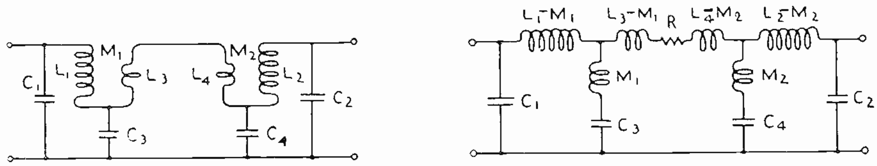
The aim of designers of band pass filters has always been to obtain a flat pass band frequency response with sharp cut-offs and high attenuation outside the band. An earlier patent (No. 410,499) approached this ideal by using two tuned circuits having two series resonant couplings giving rejection frequencies at the edge of the pass band. The first two figures show the actual and equivalent circuit. This method has the manufacturing disadvantage that adjustment of the mutual inductances M_1 and M_2 is critical because the reactance of the series arm between the two tuned circuits must be zero. The difficulty may be overcome by including in the series arm an extra inductance L_5 (in the third figure) and capacitance C_5. If the value of C_5 is chosen to produce resonance of the series arm at the mid-frequency, the reactance of this arm becomes negligible over the pass region.

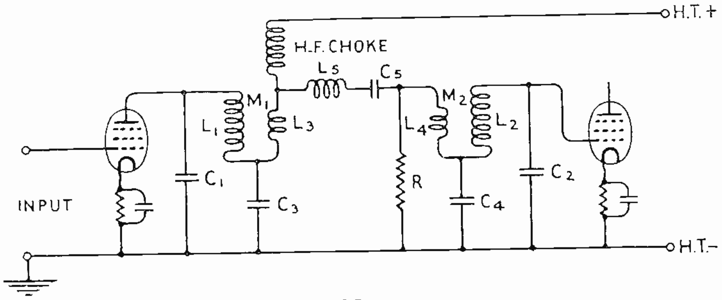


PATENT NO. 495,455.

A suitable circuit for this type of filter is shown in the third figure. The following circuit constants give a pass band of ±7 kc. at a mid-frequency of 465 kc. with 40 db. attenuation at ±12 kc.

L_1	2060 μH	M_1	370 μH	C_1	70 $\mu\mu$F	H.F. choke = 3000 μH
L_2	1370 ,,	M_2	310 ,,	C_2	110 ,,	R = 0.5 M
L_3	240 ,,			C_3	320 ,,	
L_4	140 ,,			C_4	370 ,,	
L_5	375 ,,			C_5	1700 ,,	

L_1 has a higher value than L_2 because a high input impedance is required on the primary side.



PATENT NO. 495,455.

Adjustment of the filter with the aid of a cathode ray tube is simple. C_1 and C_2 are set to give maximum response at the mid-frequency; C_3 and C_4 are next tuned to give the required upper and lower rejection frequencies. In the above example M_1 C_3 gives the lower rejection frequency. C_2 is reduced until a shoulder appears at the high frequency end of the pass band and then C_3 is trimmed to give a similar shoulder on the low frequency side. All three capacitors C_1 C_2 and C_3 are finally trimmed for a flat topped symmetrical pass curve. A method is also described for adjustment using an oscillator and valve voltmeter.

DIRECTION FINDING SYSTEMS.

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This invention relates to an automatic biasing arrangement for Direction Finding equipment which does not destroy the shape of polar curves being received.

It is well known that when using D.F. apparatus if the receiving system operates in too high a field it may become saturated, and in consequence, its directional properties become seriously impaired. It is, of course, not possible to operate an ordinary automatic gain control device on the D.F. receiver proper because such an automatic volume control tends to smooth out the variable signal which it is desired to record. To overcome this difficulty the inventor uses, in addition to an ordinary D.F. aerial, an open aerial connected to a rectifying valve. The output from the open aerial passes through the resistance-capacity circuit, the rectified voltage from which is applied to the gain control of the D.F. receiver proper. It is clear that when weak signals are being received the control valve connected to the open aerial will produce no negative biasing effect on the D.F. receiver, but when the strength of field is increased above a certain critical value, the voltage from the resistance-capacity circuit of the control valve biases back the grid of the amplifying valves in the D.F. receiver, thus reducing the overall gain, the amount of control being independent of the position of the direction finding aerial system.

ELECTRON MULTIPLIER CIRCUITS.

Application date, June 5th, 1937. **No. 496,674.**

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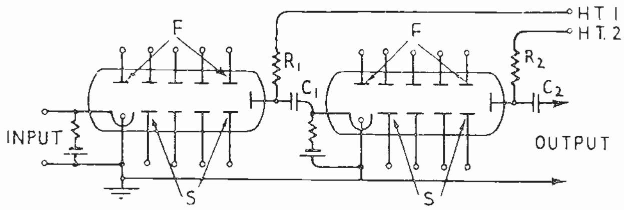
Grid controlled electron multipliers are normally not very efficient amplifiers of very small voltage fluctuations because the D.C. component of current is amplified



PATENT NO. 496,674.

in the same ratio as the A.C. variations. For example an A.C. input voltage may produce a current variation of 1 μA in a D.C. current of 1 mA at the first stage. If the multiplier has a gain of 30 the output current variation is 30 μA in a D.C. current of 30 mA.

This difficulty may be overcome by amplifying through a series of such multipliers each having a limited gain. The figure shows a suggested circuit using R.C. coupling, but tuned circuit-capacitance or transformer coupling may also be employed. The essential feature is that the output voltage change only is passed on to the next multiplier. All the multipliers may, if required, be mounted in the same bulb.

CONTROLLED CARRIER MODULATION.

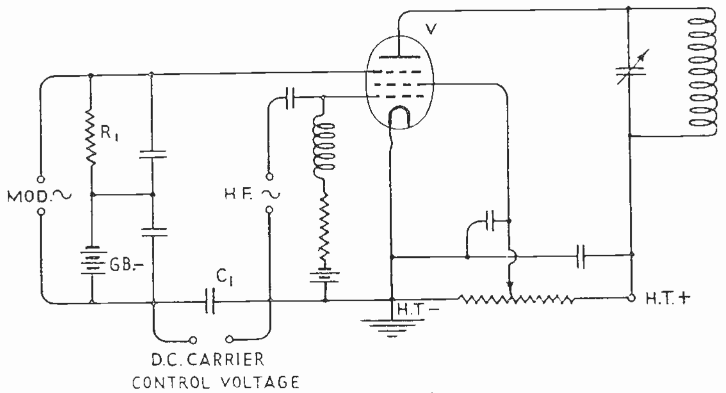
Application date, December 23rd, 1937. **No. 496,845.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and L. T. Moody and C. R. Staines.

A method of using a multi-electrode valve, such as a screened pentode, to provide a simple but effective system of "floating" or controlled carrier modulation is described in this application.

By reference to the circuit diagram, Fig. 1, it will be seen that a pentode valve is used as a driven H.F. power amplifier with normal H.F. circuits connected to its control grid and anode.

Low frequency modulation is applied to the suppressor grid across the grid resistance R_1 and bias supply GB—. The earth return for this circuit is broken by the condenser C_1 across which the D.C. controlling voltage is applied.



PATENT NO. 496,845.

This voltage may be obtained by rectifying a portion of the low frequency modulation signal in any convenient manner.

Thus, a controlling bias is produced across C_1 which is proportional to the amplitude of the modulating signal and is arranged so as to decrease the effective grid bias negative applied to the suppressor grid during periods of modulation and so raise the carrier amplitude in sympathy with the modulating signal requirements.

An alternative arrangement provides for anode modulation of the pentode while the carrier is still controlled in the manner described.

TRANSMITTER MODULATION.

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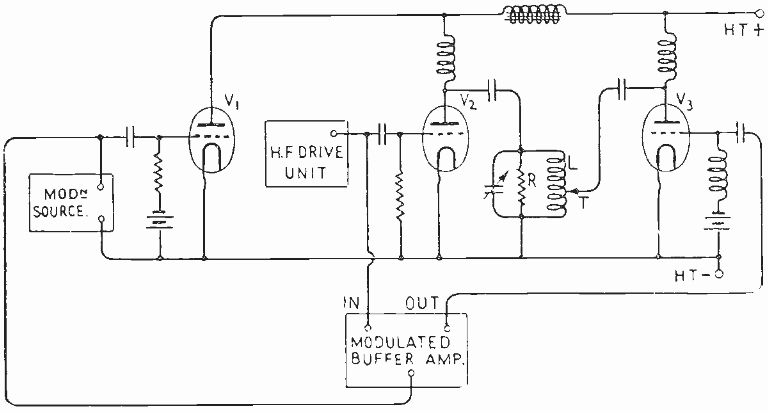
This patent describes a method of obtaining modulation of a radio frequency carrier at high efficiency without the use of large and expensive iron-cored low frequency components as used in conventional Class " B " modulator circuits.

The basic circuit is shown in Fig. 1, V_1 is the modulator valve, V_2 the main H.F. power amplifier valve and V_3 the auxiliary modulated amplifier valve. R_L represents the load due to the antenna circuit.

The modulator valve V_1 is biased to cut-off in the absence of a modulating signal and so is only operative during positive half cycles of the modulating frequency. It thus effects modulation of the H.F. carrier substantially only over the negative half cycles of the envelope.

The main amplifier valve V_2 receives constant drive from the H.F. Drive Unit and operates as a high efficiency Class " C " amplifier the ratio of peak radio frequency voltage to H.T. voltage on its anode being about 0.9.

The drive, which is in phase with that on the grid of V_2, and bias upon the auxiliary amplifier valve V_3 are such that a small anode feed flows to this stage in the zero modulation or carrier condition and since the tap T from the anode of this stage upon the tuned circuit in the anode circuit of the main valve stage is about half way down on the inductance, the ratio of radio frequency peak voltage to high tension voltage on the anode of the auxiliary valve V_3 will be about 0.45 in the



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carrier condition. During positive half cycles of modulation the auxiliary stage is " driven harder " and accordingly takes increased anode feed and delivers increased power to the load and, therefore, the ratio of radio frequency voltage to high tension voltage on the anode of the main valve V_2 is increased and thus this valve tends to take decreased feed. This, however, is resisted by the action of the audio frequency choke which automatically raises the potential at the end connected to the anode of V_2 so as to maintain the feed constant until, with positive peaks of 100 per cent. modulation, the potential at this end of the choke is approximately twice the normal high tension voltage. In this condition the main valve V_2 is delivering very nearly twice carrier power into the load and a similar amount approximating to twice the carrier power is supplied by the auxiliary valve V_3. The choice of the ratio of 0.45 already mentioned as occurring in the carrier condition allows a limiting ratio value of 0.9 to be reached in the 100 per cent. modulation condition.

It is claimed that an efficiency under all conditions of modulation which is rather better than that obtained with the normal Class " B " modulator system can be obtained with this arrangement.

TIME BASE CIRCUITS.

Application date, June 26th, 1937. **No. 497,760.**

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

The blocking oscillator is well known as a generator of large amplitude pulses of short duration and it can be used in the production of a saw-tooth wave shape to trigger a discharge device across a capacitance. Three essential requirements are :—The generated pulses should be steep sided and of short duration, and the oscillator should be independent of output circuit changes.

TABLE I.

CONDITION I: $X_L < X_C < X_N$.	
(*a*) :— $X_L X_N > (X_L + X_N) X_C$	
$\left.\begin{array}{l} Z_{d1} = -jA_1 \\ Z_{s1} = +jB_1 \end{array}\right\} Z_{o1} = +C_1$	$\left.\begin{array}{l} Z_{d2} = +jA_2 \\ Z_{s2} = -jB_2 \end{array}\right\} Z_{o2} = +C_2$
(*b*) :— $X_L X_N < (X_L + X_N) X_C$	
$\left.\begin{array}{l} Z_{d1} = -jA_1 \\ Z_{s1} = -jB_1 \end{array}\right\} Z_{o1} = -jC_1$	$\left.\begin{array}{l} Z_{d2} = +jA_2 \\ Z_{s2} = +jB_2 \end{array}\right\} Z_{o2} = -jC_2$
CONDITION II: $X_C < X_N < X_L$.	
(*a*) :— $X_L X_N > (X_L + X_N) X_C$.	
$\left.\begin{array}{l} Z_{d1} = +jA_1 \\ Z_{s1} = +jB_1 \end{array}\right\} Z_{o1} = +jC_1$	$\left.\begin{array}{l} Z_{d2} = +jA_2 \\ Z_{s2} = +jB_2 \end{array}\right\} Z_{o2} = +jC_2$
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CONDITION III: $X_N < X_L < X_C$.	
(*a*) :— $X_L X_N > (X_L - X_N) X_C$.	
$\left.\begin{array}{l} Z_{d1} = -jA_1 \\ Z_{s1} = -jB_1 \end{array}\right\} Z_{o1} = -jC_1$	$\left.\begin{array}{l} Z_{d2} = -jA_2 \\ Z_{s2} = -jB_2 \end{array}\right\} Z_{o2} = -jC_2$
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By the chart this is

$$\theta_D = -0.174 + j\,\underline{1.348} \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)}$$

The Equivalent Line.

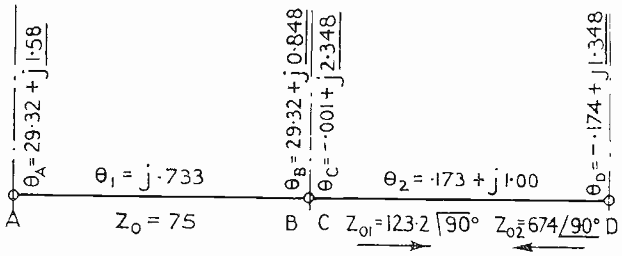

FIG. 6.

$$\theta_C = \theta_D + \theta_2 = -.174 + j\,1.348 + .173 + j\,1.00$$
$$= -0.001 + j\,2.348 \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)}$$

$$\theta_B = \tanh^{-1}\left(\frac{Z_{o1}\tanh\theta_C}{Z_o}\right) = \tanh^{-1}\left(\frac{123.2\ \backslash 90^\circ \cdot \tanh(-0.001 + j2.348)}{75}\right)$$

$$= \tanh^{-1}\left(\frac{123.2\ \backslash 90^\circ \cdot 0.612\ /90.15^\circ}{75}\right)$$

$$= \tanh^{-1} 1.005\ /0.15^\circ$$

$$= 29.32 + j\,0.848 \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)}$$

$$\theta_A = \theta_1 + \theta_B = j.733 + 29.32 + j\,0.848$$

$$= 29.32 + j\,\underline{1.581} \text{ - - - - (hyps \& } \tfrac{1}{4}\text{-waves)}$$

This is the correct solution.

Test θ_C by the rule

$$Z_o = Z_{o1} \tanh \theta_C$$

i.e.

$$Z_o = 123.2\ \backslash 90^\circ \cdot 0.612\ /90.15^\circ$$
$$= 75.4\ /0.15^\circ$$

That is practically 75 ohms.

Also test θ_A by the rule

$$Z_o = Z_o \tanh \theta_A$$

That is

$$Z_o = 75 \times \tanh(29.32 + j\,1.58)$$
$$= 75 \times 0.9987\ \backslash 0.315^\circ$$
$$= 74.9\ \backslash 0.315^\circ$$

which also is practically 75 ohms.

Summarising the foregoing we see how easy it is, in approaching such a problem as the one we have chosen, to confuse the general number coefficient of the positive real numerical part of the image impedances of the dissymmetrical network. Also we see how considerably such confusion can affect the results of calculation. In our particular case we knew beforehand what the correct result should be, but it is not difficult to appreciate that there might be many cases in practice in which the correct result would not be known beforehand.

Hence we see the usefulness of Table I which sets out the correct general number coefficient of the image impedances of the particular dissymmetrical network. As already mentioned, these results will be set out in Tables II to V (Part II) in more compact form, together with other cases of dissymmetrical T networks, consisting of pure reactances.

Note on Readings within Negative Image of Chart.

The interpretation of the Input Impedance Chart to yield $\tanh^{-1} 1.598 \overline{\backslash 111^\circ 48'} = (-0.174 + j\,\underline{1.348})$ in this example on page 34 introduces us to the resolution of an ambiguity in connection with the chart. If reference is made to our article in THE MARCONI REVIEW, No. 67, Sept.-Dec., 1937, page 8, Ex. (H), the reader will find the following rule given for dealing with cases where the angle ϕ is greater than 90° numerically: Subtract the numerical value of the angle from 180° and use the remainder (after prefixing the sign of the original angle) for the chart in the ordinary way.

To that rule we must now add the following extension. It must be clearly understood that in these cases where the angle is greater than 90° (or less than −90°) the resulting value of α is negative. The practical case here given is an example of all such cases. See also page 30 of this article.

(To be continued.)

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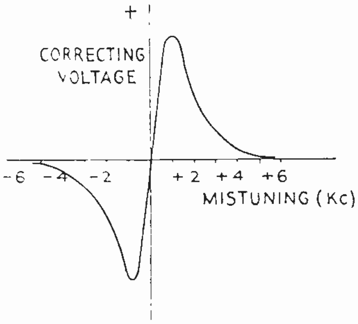

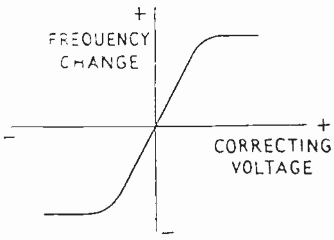

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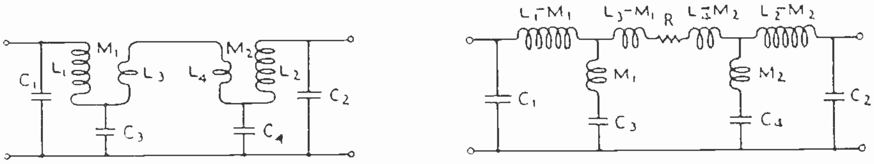

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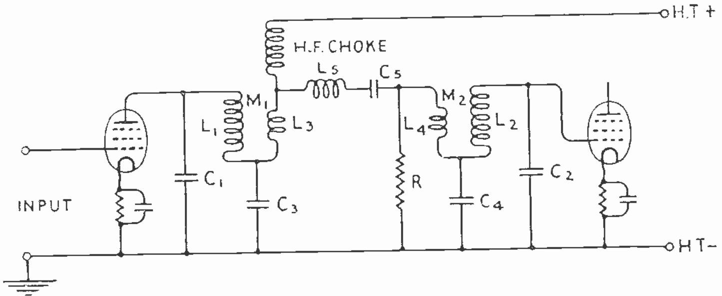

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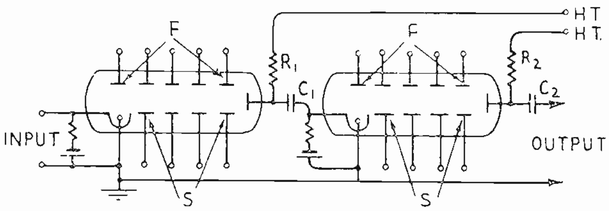

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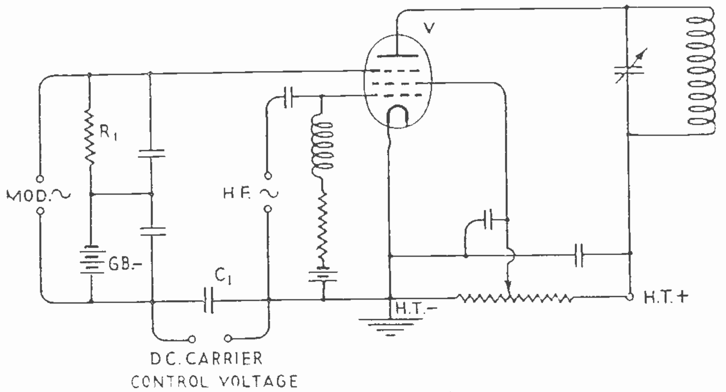

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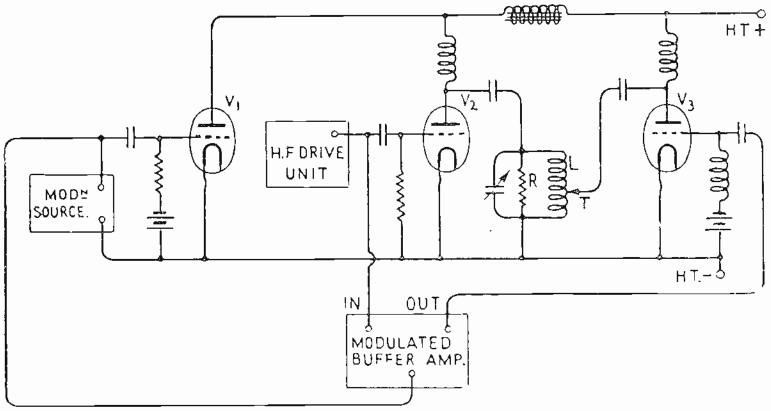
The modulator valve V_1 is biased to cut-off in the absence of a modulating signal and so is only operative during positive half cycles of the modulating frequency. It thus effects modulation of the H.F. carrier substantially only over the negative half cycles of the envelope.

The main amplifier valve V_2 receives constant drive from the H.F. Drive Unit and operates as a high efficiency Class " C " amplifier the ratio of peak radio frequency voltage to H.T. voltage on its anode being about 0.9.

The drive, which is in phase with that on the grid of V_2, and bias upon the auxiliary amplifier valve V_3 are such that a small anode feed flows to this stage in the zero modulation or carrier condition and since the tap T from the anode of this stage upon the tuned circuit in the anode circuit of the main valve stage is about half way down on the inductance, the ratio of radio frequency peak voltage to high tension voltage on the anode of the auxiliary valve V_3 will be about 0.45 in the



PATENT NO. 497,343.

carrier condition. During positive half cycles of modulation the auxiliary stage is "driven harder" and accordingly takes increased anode feed and delivers increased power to the load and, therefore, the ratio of radio frequency voltage to high tension voltage on the anode of the main valve V_2 is increased and thus this valve tends to take decreased feed. This, however, is resisted by the action of the audio frequency choke which automatically raises the potential at the end connected to the anode of V_2 so as to maintain the feed constant until, with positive peaks of 100 per cent. modulation, the potential at this end of the choke is approximately twice the normal high tension voltage. In this condition the main valve V_2 is delivering very nearly twice carrier power into the load and a similar amount approximating to twice the carrier power is supplied by the auxiliary valve V_3. The choice of the ratio of 0.45 already mentioned as occurring in the carrier condition allows a limiting ratio value of 0.9 to be reached in the 100 per cent. modulation condition.

It is claimed that an efficiency under all conditions of modulation which is rather better than that obtained with the normal Class "B" modulator system can be obtained with this arrangement.

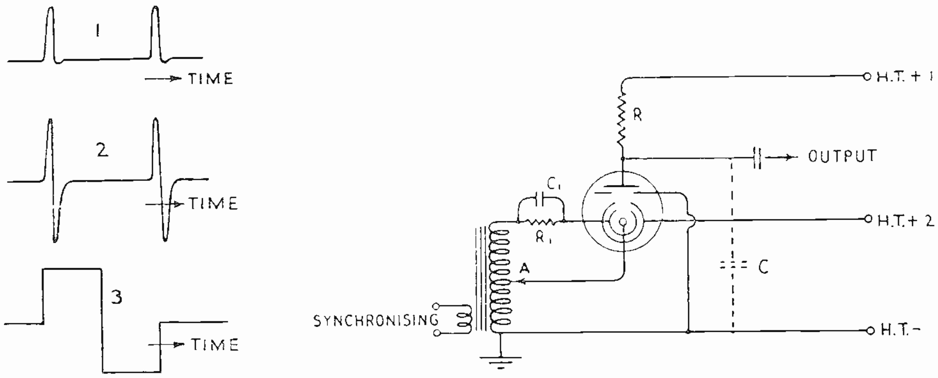
TIME BASE CIRCUITS.

Application date, June 26th, 1937. **No. 497,760.**

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

The blocking oscillator is well known as a generator of large amplitude pulses of short duration and it can be used in the production of a saw-tooth wave shape to trigger a discharge device across a capacitance. Three essential requirements are :— The generated pulses should be steep sided and of short duration, and the oscillator should be independent of output circuit changes.

These may be realised by the use of an electron beam valve as shown in the diagram. The control electrode is a cylinder, with a horizontal slit, encircling the cathode, the anode is a similarly shaped larger cylinder. A suppressor plate is interposed between the output anode and the oscillator. The latter is of the electron coupled type with the cathode connected to a suitable point A on the oscillator coil. Coupled to this is a synchronising coil. The ouput anode is effectively isolated and



PATENT NO. 497,760.

changes of output load have little effect on the oscillator performance. Very steep sided pulses of short duration are developed across R as shown in 1 and these pulses may be made sharper still (see wave shape 2) by replacing R by an L.F. choke. The anode circuit itself may be made to supply a saw-tooth output voltage by connecting the capacitance C from anode to earth. The oscillator pulse causes anode current to flow and so discharges C.

By adjusting the tapping point A on the oscillator coil so as to be close to the control electrode (i.e., near R_1) a square shaped output voltage can be obtained (see 3) and this is useful for providing L.F. switching in cathode ray direction finders.

VARIABLE SELECTIVITY SYSTEMS.

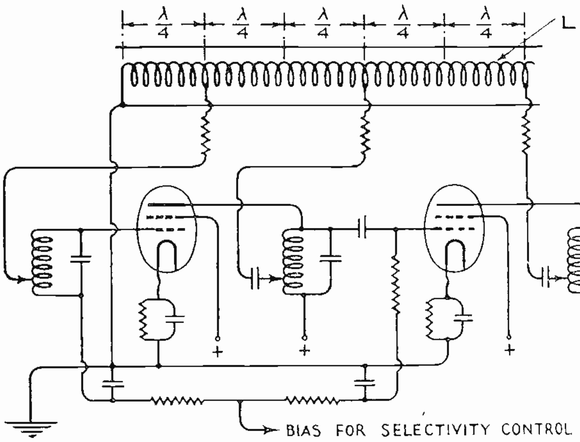
Application date, July 15th, 1937. **No. 498,688.**

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

Reaction can be used for variation of selectivity, but it is difficult to maintain stability because, with normal circuits, the phase relation of the feedback voltages changes with the amount of reaction. In this invention the feedback circuit is obtained by tapping on to different points of a natural or artificial line.

When a line is excited at one end and terminated in a resistance at the other, it has built upon it standing waves with nodes and adjacent anti-nodes at quarter wave intervals. Variation of the terminating resistance alters the amplitude of the standing waves, but does not change the phase relationships of the potentials at different points along the line.

A circuit illustrating the principle is shown in the figure: it consists of a two-valve amplifier with positive feedback applied across both valves. This feedback is provided by tappings on to the natural or artificial line L, a number (e.g., 5) of



PATENT No. 498,688.

quarter wavelengths long. If the mutual conductance of the valves is varied the amount of reaction is varied and the selectivity is changed. The variation of mutual conductance may be effected by changing the bias on the control grids, this bias being controlled manually or automatically in dependence upon either the input signal strength or the interference level.

Circuit arrangements for providing variable selectivity, using positive and negative feedback along lines, supplemented in some cases by negative feedback obtained by means of cathode impedances, are described in the specification.

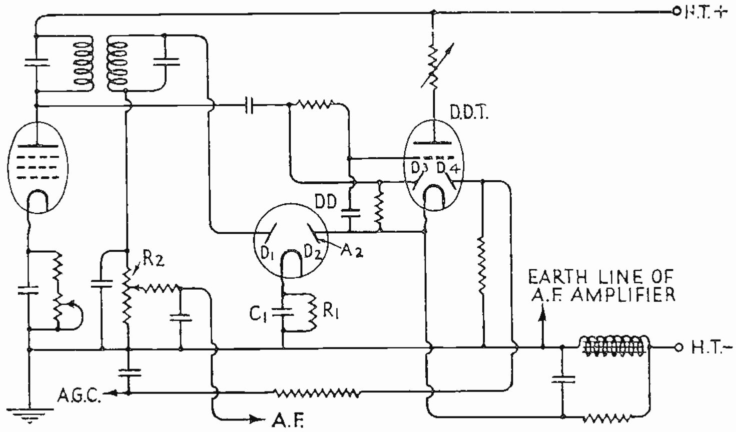
NOISE SUPPRESSION CIRCUITS.

Application date, July 15th, 1937. **No. 498,842.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and K. R. Sturley.

An improved method of "noise suppression" is described in the specification.

One diode D1 of a double diode DD acts as the detector in the receiver, the circuit being conventional except for the inclusion of the resistance R1 bypassed by capacitance C1 between the double diode cathode and earth. A double-diode-triode DDT rectifies the signal (diode D3) for A.G.C. purposes, and the triode portion amplifies the A.G.C. potentials, the diode D4 providing the delay action. The anode A2 of the double diode is connected to the cathode of the double-diode-triode.



PATENT No. 498,842.

With no signal the double-diode-triode cathode is positive with respect to earth, the diode D2 of the double diode conducts and the voltage drop across R biases the detector diode so that it is inoperative. The positive potential of the double-diode-triode cathode falls, with increasing signal, reaching zero when the signal is strong enough.

At this point the detector "suppression" bias is removed and the A.G.C. delay voltage is also overcome. Hence detection and A.G.C. commence simultaneously.

Advantages include :—

(1) Quick removal of suppression bias as suppression point is passed allowing distortionless detection.

(2) No "hum" pick-up from source of A.G.C. voltages and no A.F. "break-through" at zero volume control (R2) setting because one end of the detector load R2 is earthed.

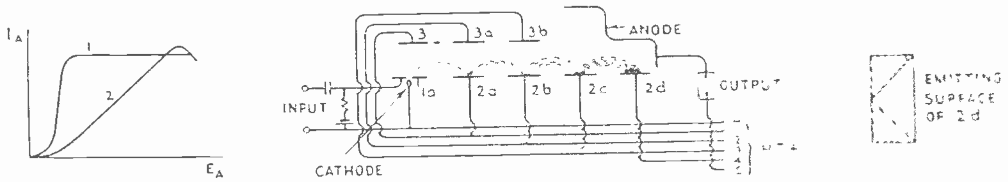
(3) Although the triode portion of the double-diode-triode may be used for A.F. amplification, it is better not to do so, as no distortion due to varying grid bias then occurs.

ELECTRON MULTIPLIERS.

Application date, July 15th, 1937. No. 498,843.

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

The majority of electron multipliers possess an anode or output impedance approaching infinity. This imposes a serious practical limitation in the matching of the impedance of the multiplier to the output circuit. The object of this specification therefore is to enable electron multipliers to be produced having output impedances of desired convenient values.



PATENT No. 498,843.

The normal electron multiplier includes a cathode, a control electrode, a final output electrode, and one or more similar secondary emissive electrodes which are uniformly emissive over their whole operating surfaces.

It is suggested that a non-uniform surface can be given to at least the last emitter electrode in the multiplier, so that this is made more highly secondarily emissive over a predetermined part of its surface, the choice of the selected parts being made in accordance with the operating characteristic required.

If a curve connecting output electrode current and output electrode voltage for a known multiplier is plotted, it will be of the form shown in curve (1) of the first figure, the substantially horizontal part which represents the useful operation of this characteristic curve corresponding to an output impedance of substantially infinity. If, however, the final secondary emitter electrode shown in the modified form of multiplier in the second figure be made more highly secondarily emitting over parts of its surface than over other parts, and if the output electrode and the last two field electrodes, instead of being separate, and insulated from one another, as is normal, are connected together and preferably are integral with one another as shown in the second figure, the multiplier can be made to possess a characteristic curve more suitable to individual requirements such as, for instance, shown in curve (2) of the first figure. The final secondary emitter electrode can be made, for instance, as a composite electrode made partly of silver and partly of some other metal so that when sensitised the emission from the silver portion is considerably greater than from the other portion. This is shown in the third figure, where the shaded part represents the area of lower secondary electron emission and the plain part corresponds to the more highly emissive portion.

In the static condition the electrons strike the electrode at its centre line corresponding to mean output electrode current. Under dynamic conditions the potential of the output electrode varies about the mean value and the number of electrons released varies accordingly: for the point of impact of the electrons on the electrode depends on the output electrode voltage.

If the last secondary emitter electrode is made by reversing the form illustrated in the third figure, a negatively sloping characteristic may be obtained and if the more highly emissive portion is made a trapezoidal shape a higher output impedance may be obtained.

In experimental practice with a final emitter electrode as shown in the third figure, an output impedance of approximately 15,000 ohms has been obtained, a figure which is very much lower than that obtained with normal electron multipliers.

PUSH BUTTON TUNING CONTROL SYSTEM.

Application date, July 20th, 1937. **No. 498,927.**

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

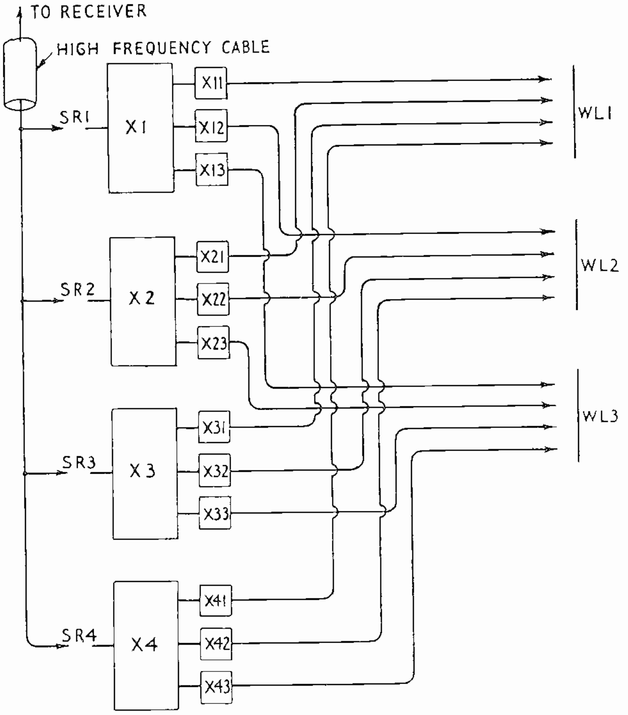
LONG WAVES	SR1	1500 M. DROITWITCH	1648 M RADIO PARIS	1571 M DEUTCHLANDSENDER
400-500 METRES	SR2	449 M N. REGIONAL	463 M LYONS PT.T.	456 M COLOGNE
300-400 METRES	SR3	342 M LONDON REGIONAL	349 M STRASBOURG	357 M BERLIN
200-300 METRES	SR4	261 M LONDON NATIONAL	253 M NICE CÔTE d'AZUR	231 M KONIGSBERG
		WL1 GREAT BRITIAN	WL2 FRANCE	WL3 GERMANY

PATENT NO. 498,927.

The push button system described in this patent enables the provision of a large selection of stations without great mechanical complexity of apparatus. It also avoids the difficulties of mechanical and electrical " back-lash " often met with, especially in systems using a remotely controlled motor. Remote control can easily be arranged, no motor or relay being necessary.

The wavelength range of a receiver is divided into a number of sub-ranges selected by push button switch units. Each sub-range contains a number of pre-determined wavelengths; to select any one wavelength two buttons—a sub-range button and a wavelength button—must be depressed.

The first figure illustrates one way of dividing the wavelength range and shows a simplified control panel. In



X1, X2, ETC. MAIN TUNING REACTANCES
X11, X12, X13, ETC. TRIMMING REACTANCES

PATENT NO. 498,927.

this the stations are so grouped that one in each sub-range belongs to the same country.

Suggested circuit arrangements for straight and superheterodyne receivers are given, and in these operation of a sub-range switch unit selects a number of main tuning reactances, one for each tuning circuit. Associated with each main tuning reactance is a number of trimming reactances which are connected by operation of the wavelength switch units. The tuning circuits can be connected to the receiver by high frequency cables if remote control is required. The second figure shows one such tuning circuit schematically: the switch references correspond to those of the first figure.

The specification includes a description of a control unit giving 48 stations selected by 16 buttons.

ELECTRON DISCHARGE TUBES.

Application date, July 20th, 1937. **No. 499,218.**

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

This specification relates to electron discharge devices capable of generating or amplifying very high frequencies down to approximately 50 cms.

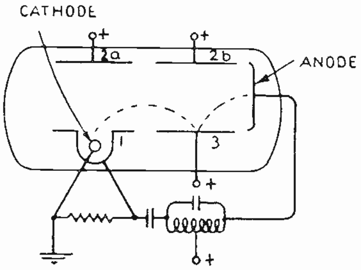
Finite electron transit time from control grid to anode limits the use of normal thermionic valves for very high frequency operation, due to the fact that (A) grid conductance increases and because finally greater than the mutual conductance, and (B) the oscillation periodicity becomes comparable to the electron transit time and correct phase relationship is lost.

This specification discloses an electron discharge device with at least one secondary emissive electrode surface at which electron multiplication occurs, and wherein the electron path and operating parameters are such that the transit time from control electrode to anode is equal to an integral number of periods of the operating frequency.

The first of these properties increases the mutual conductance and the second ensures that, although delays of 360 deg. between control electrode and anode may take place, the grid phase relation is maintained.

In a suggested construction the cathode is surrounded by a fine mesh control grid which in turn is surrounded by a very open mesh grid which acts as an output anode and finally by a cylindrical electrode whose inner surface is rendered secondarily emissive.

Positive potential is applied to the output anode and a lower positive potential is applied to the secondarily emitting electrode.



PATENT NO. 499,218.

In operation electrons emanating from the cathode are controlled by the control grid, proceed through the output anode to the secondarily emitting electrode from which an amplified electron stream returns to and is taken up by the output anode. The complete transit time from the control grid to the secondarily emitting electrode and back to the output electrode is equal to an integral number of periods of the operating frequency.

In a modification of this construction the electron discharge device may be constructed more in accordance with normal electron multiplier construction, as shown in the figure. A linear cathode is positioned inside a curved plate control electrode which has an opening covered by a grid which is in connection with the control electrode and through which the electrons proceed. Opposite the cathode is the first field electrode which is connected to a secondary emitter electrode co-planar with the cathode and opposite this is situated the second field electrode, the two field electrodes being themselves co-planar. At the end of the tube is an output electrode situated at right angles to the field and secondary emitter electrodes. The external circuit may be as shown in the figure. A positive potential may be applied to the first field electrode; higher positive potential to the second field electrode and still higher positive potential is applied through the centre tap on the inductance of the parallel tuned circuit to the output anode. A magnetic field is applied to cause the electrons leaving the grid of the control electrode to pass as indicated in broken lines on the figure to the output anode. The arrangement is such that the total time taken to travel from the grid to the emitter electrode and thence to the output anode is equal to one period of the operating frequency.

It will be appreciated that although one stage only of electron multiplication is shown in the figure, there may be any desired number of stages.

BOOK REVIEW

ELECTRON OPTICS THEORETICAL AND PRACTICAL. L. M. MYERS. Medium 8vo., XVIII, 618 pages, 380 Illustrations and Figures, including 68 half-tone plates, 42s., CHAPMAN & HALL, LTD., 1939.

THIS book by a member of the Research Department of the Marconi Company is rightly claimed as the first in the English language to appear on the subject of which it treats. Its 600 pages, divided into eight chapters, and its eight hundred classified references in the bibliography at the end, give some idea of the labour involved in assembling material for a subject still being intensively exploited in a number of (chiefly) industrial laboratories; and many workers in these places will be grateful to Mr. Myers for making easily accessible so much information hitherto very scattered.

The first two chapters, dealing with analogies between light and electrons, and with the mathematics of the electron trajectory, may be regarded as introductory to the next two, dealing respectively with electron lenses and their aberrations. The next three are more practical, covering in turn the electron multiplier, vacuum technique, and the electron microscope. The eighth and final chapter covers a number of applications of the principles of electron optics, such as the image converter, the cathode ray tube and the iconoscope.

Even this brief summary is enough to indicate the danger, to which this book partly succumbs, of confusing a text book with a monograph. The subject of electron optics is a highly specialised branch of the already highly specialised subject of modern electron physics, and as such its treatment should be able to take a great deal for granted in the reader's equipment. The book is as likely to be useful to the university teacher and the industrial physicist as to the graduate student whom Mr. Myers says in his preface he has had primarily in mind, but all these groups, if they are able to cope at all with the subject of the book, have presumably already some acquaintance with, or at any rate know where to find, the standard theory to be met with in Jeans and elsewhere. This fault is common in a number of books on electronics, particularly in recent American publications. In the present instance, it means that the first sixth of the book covers what might have been better summarised in twenty or thirty pages, thereby making a compact monograph of the book instead of a somewhat lengthy mixture of monograph and text book. The chapter on Vacuum Technique also comes within this criticism. It would have been better to take it for granted rather than endeavour to convey the elements of a complicated practical art in fifty pages.

It is in these chapters, too, that Mr. Myers is not always at his best, being sometimes too didactic and a trifle too florid in what is, again, a highly specialised work, and not a school text-book. Thus: "Coulomb discovered if two quantities of electricity Q_1 and Q_2 . . ." (p. 39), and "We are indebted to the English mathematical physicist Hamilton for the treatment of electron motion as an optical problem" (p. 27)—which is in any case hardly exact. And Mr. Myers is surely quite off his stroke in his unfortunate hortatory introduction to the chapter on Vacuum Technique, with its obscure, not to say tantalising references to "defeatists and obstructionists" (p. 346).

The book, therefore, lives up to itself most admirably in the five chapters remaining outside the foregoing comments. Of these, two (on electron lenses and on aberrations) are the more theoretical part, and precede, logically, the three later ones dealing with practical applications. Any discussion of the latter must turn on the question of emphasis. It seems to us that these portions of the book would have been clearer and their detailed subject matter more accessible to easy reference if they had been divided up into more and smaller chapters. As it is, electron multipliers and electron microscopes, the former not yet much in commercial use, and the latter of, naturally, limited and very specialised application, get separate chapters, while the cathode ray tube and its many versions, and the iconoscope and its variants—both devices of great practical value, come under the final chapter heading of "Further Applications," as if, compared with the electron microscope, they were of less importance. The chapters on multiplier and microscope cover a great deal of ground, however, and are, we believe, as useful a summary as will be found anywhere. The author's treatment of multipliers seems to be somewhat independent of the specific notions of electron optics, as if he were identifying the latter with the more general subject of electronics; but that of course does not minimise the value of this part of the book.

A certain amount of space in the final chapter is devoted to interesting experimental and speculative matters which are as yet by no means "applications." In particular, having regard to the rapid "dating" of technical matters and books dealing with them, it would have been better to deal very briefly indeed with such subjects as " temperature radiating screen for large screen television," but the author's own association with this work may perhaps be regarded as an extenuating circumstance. There is a slight omission from the section on applications to radio valves, where the aligned grid tetrode (not " pentode," as is erroneously stated on p. 577) is treated, but no mention is made of the still earlier " beam valve " developed in the laboratories of the Marconi Company.

The book is very well illustrated, the sixty-eight half-tone plates being particularly worthy of mention. It is these, no doubt, which are partly responsible for the price of the book, for two guineas is high, even for a technical work, which was undoubtedly wanted and which Mr. Myers' patience, industry and knowledge have now placed at our disposal.

G.F.P.