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# FIELD STRENGTH MEASUREMENTS

*The Marconi Company has recently introduced auxiliary equipment to its Type 476 Field Strength Measuring Set to meet a demand for apparatus capable of measuring very low or very high field intensities over certain wave bands.*

*A description is given below of a vertical aerial adaptor unit which enables the measurement of field intensities of the order of 0.15 to 0.25 micro-volts per metre on the short wave bands and of a loop aerial adaptor unit for dealing with field intensities of the order of 2 volts per metre over the broadcast wave bands.*

THE increased sensitivity of modern short wave radio receivers has led to the necessity for the measurement of field intensities of a much lower order than was previously required. Hitherto it was sufficient to be able to measure intensities of the order of 2 or 3 micro-volts per metre. It is now, however, often required to measure fields at less than one-tenth of this strength.

As an aid in the reception of these very weak field intensities, a vertical aerial, owing to its greater effective height, is much more efficient as an energy source than a loop aerial. The effective height of a tuned loop is limited by the fact that it (the effective height) is directly proportional to the number of turns and area of the loop and inversely proportional to the wavelength. The turns and area are, of course, limited by the tuning range. A gain of as much as 10 : 1 in effective height can be obtained by the substitution of a half wave vertical aerial in place of the loop and the advantage to be obtained from such a change for the reception of very weak signals is therefore obvious. Before field strength measurement is possible, however, it is necessary to know the effective height of the vertical aerial.

The effective height of a vertical aerial is not amenable to accurate calculation as is the case with a tuned loop aerial, but it is quite a simple process to calibrate its effective height by comparison with that of a known loop. This calibration is, in effect, a measure of the ratio of the E.M.F. induced in the vertical aerial (by a signal radiated from a distant source from which the direct or ground ray is received) to that induced in the loop by the same source. For this purpose a small oscillator placed at a distance of not less than ten wavelengths from the measuring set may be used as the distant calibrating source.

The most usual method of field strength measurement and that employed in the Marconi Type 476 equipment is by what is known as the substitution method, that is, by the elimination of the signal being measured and by substituting a local calibrating signal of known magnitude. In the case of a loop aerial the incoming signal is quite easily eliminated by the orientation of the loop to the zero signal position. The local calibrating signal is then introduced free of interference from the external signal. With the vertical aerial the external signal cannot be so easily excluded. To achieve this the aerial must be entirely disconnected and, to preserve equal receiver input conditions, an equal non-receiving load must be substituted at the receiver input whilst the calibrating signal is being introduced.

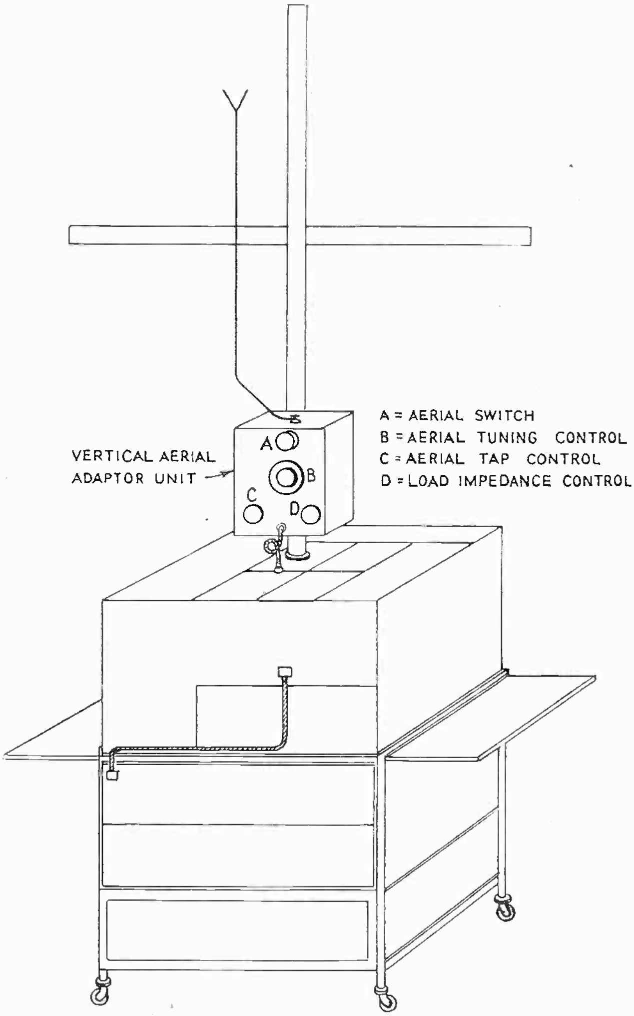


FIG. I.

It will be obvious that the use of the vertical aerial is not limited to the measurement of weak signals. Once it has been calibrated under a given set of conditions it may be employed for all field intensity measurements within the scope of the apparatus as a whole.

In comparing the results of field measurements made on the loop and vertical aerials of a signal arriving from a source outside the limits of direct ray reception, it should be borne in mind that, owing to the various possible types of ray polarisation, the two types of aerials may not at times be proportionately energised. Consequently quite large apparent differences of field strength may occur. It frequently happens that the signal is at a maximum in one type of aerial when it is at a minimum in the other type and vice versa.

**Vertical Aerial Adaptor Unit.**

A Vertical Aerial Adaptor Unit has been specially designed to operate in conjunction with the Marconi Type 476 Field Strength Measuring Set. (The Type 476 equipment was described in THE MARCONI REVIEW, No. 36).

This Adaptor Unit takes the form of a small metal case carrying the necessary circuit arrangements and capable of easy attachment to the lower part of the vertical standard of the loop aerial frame. Connection is made to the receiver input by the insertion of a pair of plugs into the sockets normally occupied by the loop aerial plugs. The only additional connection to be made (apart from the attachment of the vertical aerial) is an "earth" for the adaptor to the case of the main unit. No alteration to the circuits of the main equipment is required. A sketch of the complete equipment showing the adaptor unit in position is given in Fig. 1. Theoretical circuits are given in Fig. 2.

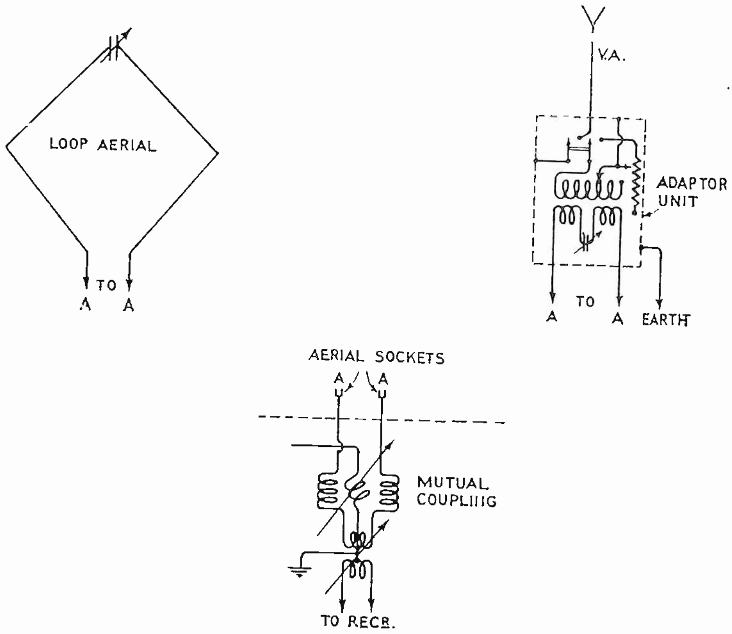


FIG. 2.

The full wave range of the vertical adaptor unit is from 14 metres to 115 metres.

The vertical aerial gain varies, of course, with the length of the aerial, which, for best results should not exceed the half wave of the signal being observed. The gain will also vary to some extent with local conditions. At a wavelength of 20 metres as compared with an aerial 7.5 metres in length, the gain varies by some 10 per cent. to 20 per cent. per metre increase or decrease in aerial length.

Using an aerial of a fixed length of 7.0 metres over the full wave band the average gain in sensitivity, as compared with the loop aerial, is some 15 to 20 decibels. This increase in sensitivity permits the measurement of field intensities down to levels of the order of 0.15 to 0.25 micro-volts per metre. A representative vertical aerial gain factor calibration curve covering the full wave band is given in Fig. 3.

Full operating and calibrating instructions are supplied with the adaptor unit.

### High Field Intensity Measurement.

The maximum field intensity capable of being measured by a field strength set is, other things being equal, inversely proportional to the effective height of the aerial of the set. A reduction in the effective height of the aerial, therefore, results in a proportionate increase in the maximum field intensity with which the set is capable of dealing. Advantage is taken of this fact to extend the high field range of the Marconi Type 476 Field Strength Measuring Set by the introduction of a

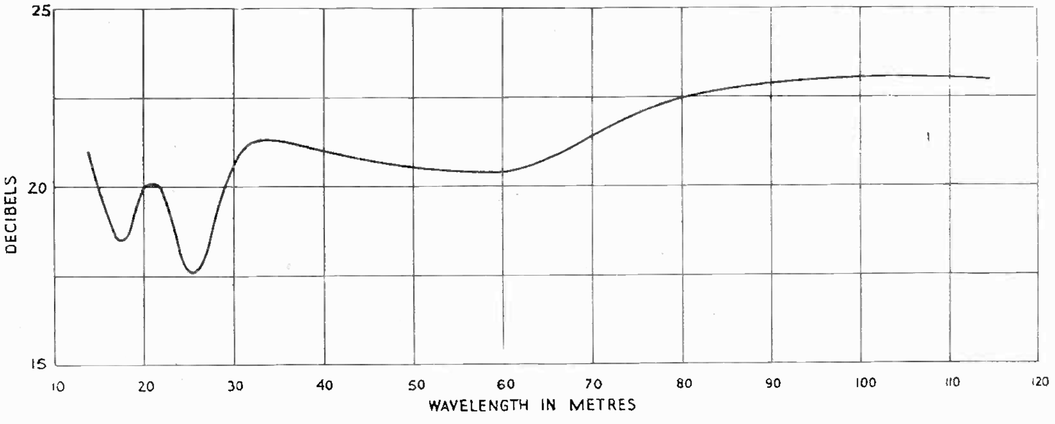


FIG. 3.

specially shielded adaptor to carry a loop aerial of much smaller effective height than that supplied with the standard equipment. By the addition of suitable loading facilities, very nearly the same L/C ratios to those employed on the standard loops are maintained, so that the wave ranges covered by the latter are approximately the same on the smaller type loops.

The high field loop aerial adaptor is fitted to the receiver by plug and socket connection in the same manner as the normal standard loop aerial frame.

### Maximum Field Voltage Ranges.

Using the high field aerial adaptor and a series of three special aerial loops in conjunction with the Marconi Type 476 Field Strength Measuring Set, field intensities up to and over 2 volts per metre at any part of the wave band between 200 metres and 2,000 metres can be measured.

F. M. WRIGHT.

# THEORY OF REACTANCE TRANSFORMER

*Under certain circumstances it is desirable so to manipulate radio circuits that a resistive load of value R is converted into a load nR or vice versa. In the following article the theory of a transformer suitable for this purpose is described, and curves given to aid the practical design of such a device.*

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THE reactance transformer is used where it is necessary to convert a load of $nR_1$ into a load of $R_1$ or vice versa. The first step in the construction of such a transformer is to put such a value of reactance X in parallel with the load $nR_1$ (as shewn in Fig. 1a) that the equivalent series circuit is $R_1$ in series with a reactance $X^1$ (as shewn in Fig. 1b). To this we now add a series reactance $-X^1$ (as shewn in Fig. 1c) and this cancels out $X^1$ giving the required load $R_1$ (Fig. 1d). This result is summarised in Fig. 1e. Conversely, if a primary load $R_1$ is connected across the terminals BC it will be quivalent to $nR_1$ across AB as shewn in Fig. 1*f*. The parallel reactance X has been shewn as a capacity and the series reactance $-X^1$ as an inductance as this is usually the case in practice, but they could be reversed.

For the calculation of values we proceed as follows.

Admittance between A and B of $n$R and X in parallel $\Big\} = \dfrac{1}{nR_1} - \dfrac{1}{jX} = \dfrac{X + jnR_1}{nR_1X}$

Impedance between A and B $\Big\} = \dfrac{nR_1X}{X + jnR_1} = \dfrac{nR_1X\,(X - jnR_1)}{X^2 + (nR_1)^2}$

Dividing this into resistance and reactance terms we get

Resistance term $\Big\} = \dfrac{nR_1X^2}{X^2 + (nR_1)^2} = R_1$ by hypothesis .. .. .. (1)

Reactance term $\Big\} = \dfrac{-j\,(nR_1)^2X}{X^2 + (nR_1)^2} = jX^1$ .. .. .. .. .. (2)

$$\text{From (1)}\quad nX^2 = X^2 + (nR_1)^2$$

$$X^2 = \frac{(nR_1)^2}{n - 1}$$

$$X = \frac{n}{\sqrt{n-1}}\,R_1 \qquad \text{.. .. .. .. .. .. (3)}$$

Substituting this value for X in (2)

$$-X^1 = \frac{(nR_1)^2 \cdot nR_1}{\sqrt{n-1}\,(nR_1)^2\left(\dfrac{1}{n-1} + 1\right)} \qquad \text{.. .. .. .. (4)}$$

$$= \frac{nR_1}{\sqrt{n-1}\;\dfrac{n}{n-1}} = R_1\sqrt{n-1}$$

$$X = \frac{nR_1}{\sqrt{n-1}}, \quad -X^1 = \sqrt{n-1}\; R_1$$

X is infinite when " $n$ " is infinite and also when $n = 1$.
Differentiating to find the minimum value of X

$$\frac{d}{dn}\, n\,(n-1)^{-\frac{1}{2}} = (n-1)^{-\frac{1}{2}} - \tfrac{1}{2} n\,(n-1)^{-\frac{3}{2}} = (n-1)^{-\frac{1}{2}} \left\{ 1 - \frac{1}{2}\,\frac{n}{n-1} \right\} =$$

$$= (n-1)^{-\frac{1}{2}} \left\{ \frac{2n-2-n}{2\,(n-1)} \right\}$$

This vanishes when $n = 2$ and $X = 2R_1$.

For large values of " $n$ " $X \simeq X^1 \simeq \sqrt{n}\, R_1$

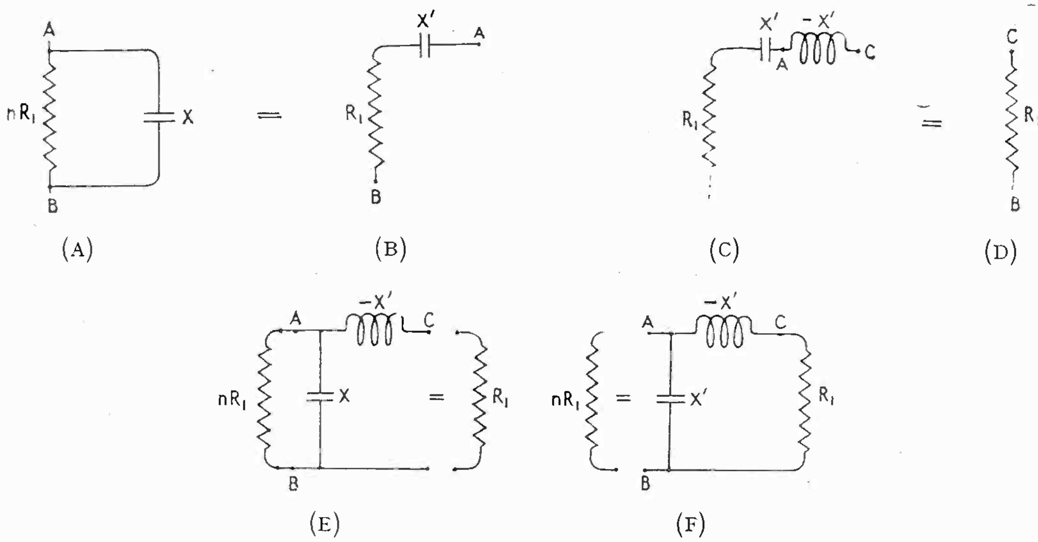


FIG. 1.

| $n$ | $\sqrt{n-1}$ | $\frac{n}{\sqrt{n-1}}$ | $n$ | $\sqrt{n-1}$ | $\frac{n}{\sqrt{n-1}}$ |
|---|---|---|---|---|---|
| 1 | 0 | ∞ | 14.0 | 3.6 | 3.9 |
| 1.16 | .4 | 2.9 | 26.0 | 5.0 | 5.2 |
| 1.5 | .707 | 2.12 | 50.0 | 7.0 | 7.1 |
| 2.0 | 1.0 | 2.0 | 82 | 9.0 | 9.1 |
| 3.0 | 1.41 | 2.12 | 101 | 10.0 | 10.1 |
| 5.0 | 2.0 | 2.5 | 122 | 11.0 | 11.1 |
| 7.0 | 2.45 | 2.86 | 145 | 12.0 | 12.08 |
| 10.0 | 3.0 | 3.33 | | | |

The values of X and $X^1$ have been computed in the above tables and curves are shewn in Fig. 2

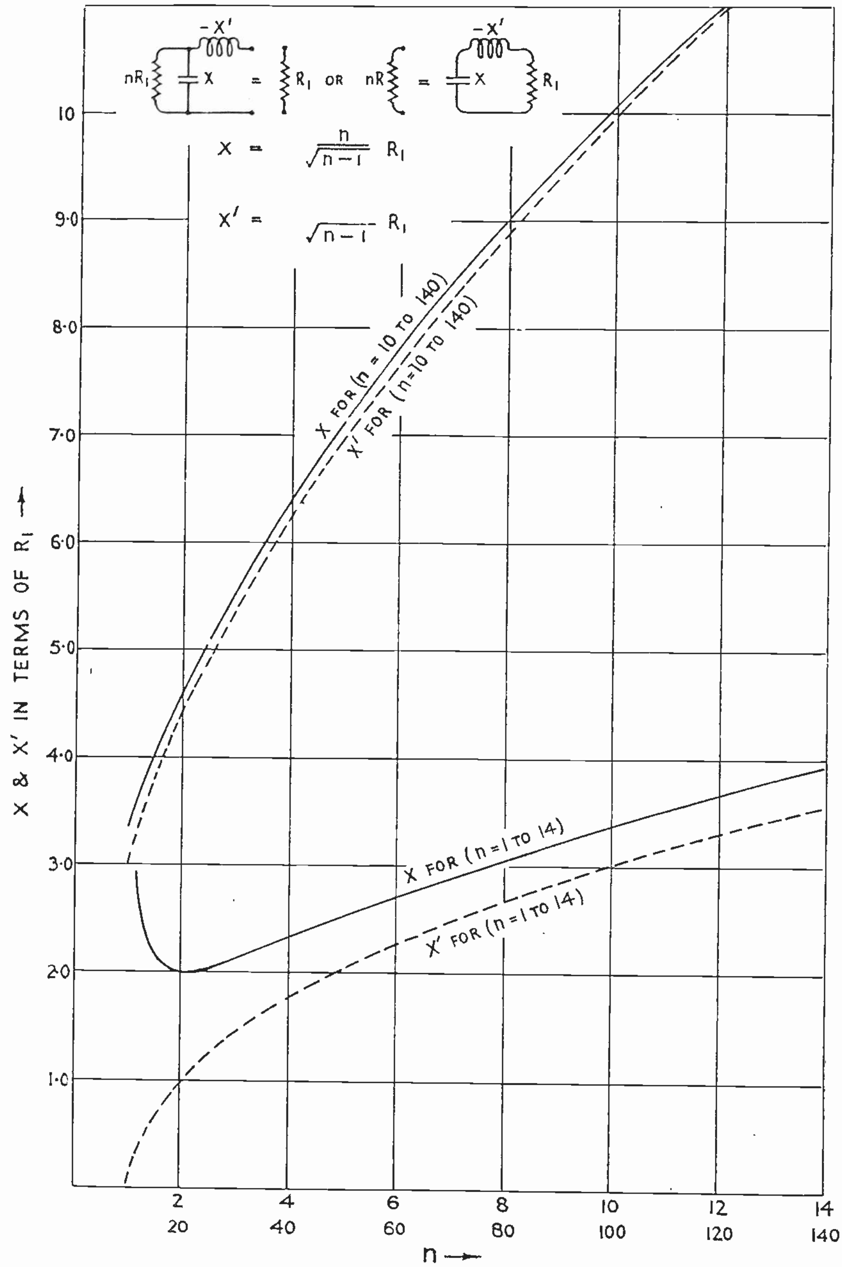


FIG. 2.

One point that stands out is the small range of capacity and inductive reactance required to deal with a large transformation ratio.

It is hoped to follow up this article with one dealing more exclusively with the practical side of the question of reactance transformation. E. GREEN.

# A STROBOSCOPE "FLICK" CIRCUIT

*When using a gas-discharge tube for stroboscope illumination it is frequently found difficult to make the flash sufficiently short to give a sharp image, particularly if it is desired to see stationary images at multiples and sub-multiples of the fundamental frequency. This paper describes one method of obtaining an extremely short flash and points out some of the limitations of the circuit.*

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THE circuit to be described has been evolved from the ordinary Abraham-Bloch multivibrator, but as the action of this device is not universally known it would seem desirable to attempt a short explanation.

On reference to Fig. 1 it will be seen that the multivibrator consists of a simple two-valve resistance-fed capacity-coupled amplifier, deliberately back-coupled from output to input. This, of course, results in a form of "motor-boating," but it is necessary to follow its action a little more closely in order to understand it fully.

Assume the circuit to have been at rest and some slight disturbance to have made G1 become a little negative (say—$\triangle e$).

Let the voltage amplification of V1, in combination with R1 and R4 be called $\mu$.

Then it will be seen that G2 will be given a potential of $-\mu\triangle e$. Thus, as G1 becomes more and more negative, G2 should become more positive, at $\mu$ times the rate.

On the other hand, applying the same reasoning to V2, it would appear that G1 must go negative $\mu$ times as rapidly as G2 is driven positive. This is clearly only possible when $\mu = 1$, when both grids will change their potentials at an equal rate.

Here, then, is the key to the problem. There exist between electrodes and between the various circuit components, small self-capacities. Hence, at a certain very great rate of change of potential, $\mu$ does equal 1.

This state of affairs holds until V1 has been cut off to zero anode current, but the stray capacities associated with this valve continue to become charged, thus causing G2 to go more positive, which in turn drives G1 more negative to an extent corresponding to a little less than the anode supply voltage.

Referring now to Fig. 2, it will be seen that G1 has travelled from *a* to *b* and G2 from *c* to *d*. The circuit is now more or less passive, as V1 is now not conducting and therefore no amplification can occur round the network.

C1 and C2, however, are holding charges which R3 and R4 are leaking away, quasi-exponentially. This falling away of the potentials on G1 and G2 is not *exactly* exponential as it is aided by the amplification of V2 and modified by its characteristics and its grid current.

G1, travelling from "b" to "c," eventually arrives at that point where V1 again begins to amplify and the foregoing process is then repeated in the reverse direction.

The question often arises as to how the frequency of a multivibrator is calculated. Exact calculation is impossible as, obviously, the frequency depends not only upon the values of the four resistances and two condensers but also upon every known and unknown characteristic of the valves employed, and is therefore outside the realms of practical calculation.

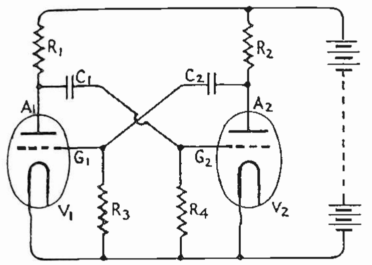


FIG. 1.

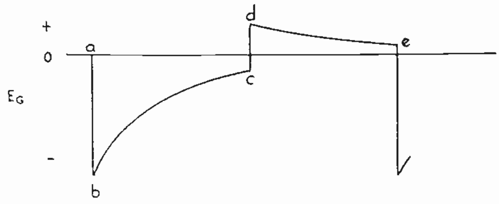


FIG. 2.

A rough guide is that :—

$$f \text{ is of the order of } \frac{2}{(C1+C2)\ (R1+R2+R3+R4)},$$

but this is very approximate and may give a result two or three hundred per cent. in error.

It is believed that the " flick over " from " a " to " b " takes only a fraction of a micro-second, as it is possible to make the circuit oscillate at one or two megacycles per second (or, of course, at as low a frequency as is desired. The writer has had one going at about three cycles per minute).

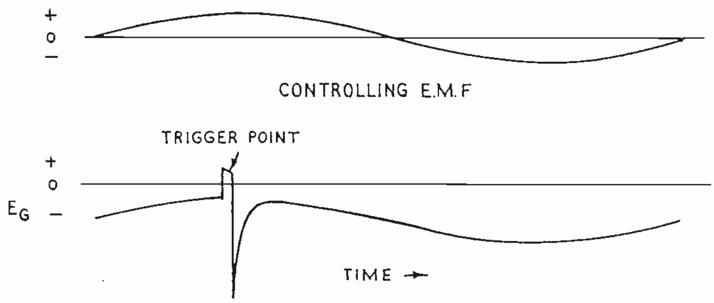


FIG. 3.

A point of interest is that, as shewn above, the multivibrator will oscillate without the inclusion of any inductance whatever. It is commonly stated (and it is believed that the idea originated with Van der Pol) that a minute amount of inductance is necessary to maintain oscillation. An analogy used to support this view is that the steam engine will not work without at least a small flywheel. It is impossible to find a mechanical analogy to support the other view but, at any rate, the ordinary vacuum brake steam driven pump does seem to manage without a flywheel.

**The Modified Circuit.**

The above description has been of an ordinary multivibrator without any grid bias. If negative grid bias be applied, the portions of the curve *b—c* and *d—a* in

Fig. 2 will be lengthened until, when the bias is great enough, the valves no longer amplify and therefore do not oscillate.

In no case, so far, have we made the wave very asymmetric, which is one thing we desire, but this *can* be accomplished by biasing one valve only.

Further, we do not desire the circuit to oscillate freely ; we merely wish it to give one sharp impulse once during each cycle of a controlling frequency. This may be done by biasing one valve only until the circuit just does not self-oscillate. The controlling frequency is then superimposed and trips the biased valve at the peak of every wave, thus giving rise to a single rapid oscillation in a time determined by the circuit constants.

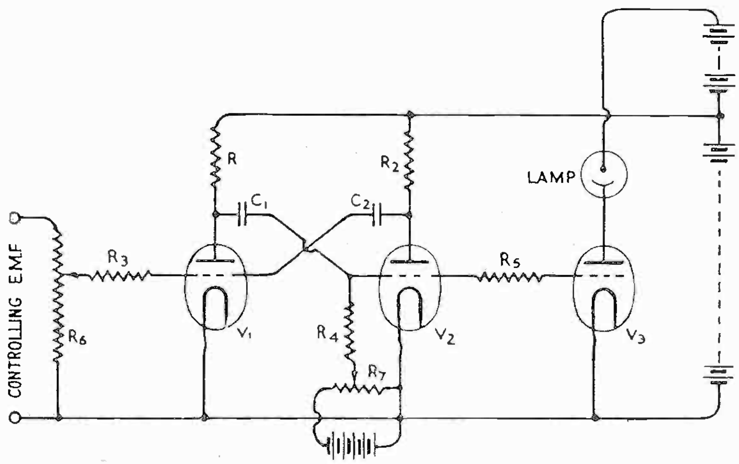


FIG. 4.

The grid voltage waveform of the controlled grid is more clearly demonstrated in Fig. 3.

In Fig. 4 is shewn the final circuit arrangement. It will be observed that the controlling *e.m.f.* is impressed not on the biased grid but on G1. This is to gain advantage of the amplification of V1 and so make the device more sensitive. The resistance, R5, coupling V2 and V3, is to limit grid current in V3. The resultant lamp current is shewn in Fig. 5.

Resistances R1, R2, R3, and R4 may be almost anything, but a good value to start from is 10,000 Ω. R5 may be about 100,000 Ω. R6 and R7 may be less than 5,000 Ω. C1 and C2 depend entirely upon the length of time of flash required and must be found by experiment.

It should be borne in mind that an extremely short flash means poor average illumination.

The length of flash may, if desired, be made as short as one micro-second, provided that the controlling frequency is not too low.

**Limitations of the Circuit.**

Although this device gives delightfully good results when properly handled it is not fool-proof.

If the signal is weak and the valve slightly over biased it may flash at one half or even at one third of the controlling frequency. On the other hand, if the signal is too strong, unless the bias be increased, it will flash more than once during the cycle, giving rise to multiple images.

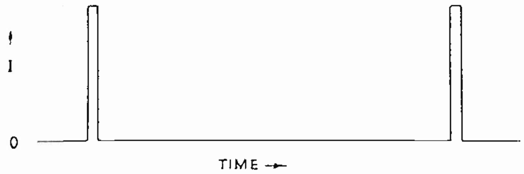


FIG. 5.

Should the operator be the fortunate possessor of a musical ear, he may, by connecting a pair of 'phones in circuit, be able to make good use of these multiples and sub-multiples, but such an arrangement can hardly be called fool-proof. It is usually quite easy to determine whether or not it is flashing at the fundamental frequency by varying the input potentiometer.

If the controlling frequency is to be varied over a very wide band, illumination will vary accordingly, as the average illumination is proportional to frequency.

This can be overcome, to a certain extent, by switching the condensers to give various times of flash.

In spite of these drawbacks, however, it would seem that the device may be useful in certain classes of work, such as frequency measurement.

W. S. MORTLEY.

# NOTE ON A MODIFIED KERR CELL FOR 240 LINE OPERATION

*It is necessary to modify the design of a Kerr cell for operation in high definition television with 240 line scan. The modifications discussed here embrace the use of diverging electrodes and the application of the circular polariscope, which ensures that both beams encompassed by the electrodes are circularly polarised. In this manner it is possible to set up the required retardation in the cell. It is recommended that the cell be biassed electrically half a wavelength in the positive direction, and biassed optically a quarter wavelength in the negative direction, which procedure serves to lower the operating voltage and therefore the power consumed by the cell.*

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UP to the present it has been the practice to use a multiplate type of Kerr cell in mechanical optical systems, but owing to the high capacity of this type it is obvious that some means of reducing this capacity must be entertained to enable the cell adequately to perform its functions for high definition television of 240 lines and upwards.

The first step in reducing the capacity is to adopt the diverging plate cell. This form of cell was suggested by Wright (described in the "I.E.E. Journal," Vol. 70, p. 340, 1931). For this type of cell $d^2$, for the retardation expression, must be replaced by $ab$, where $d$ is the gap between the parallel plate electrodes, and, for the diverging cell, $a$ is the gap at the centre, and $b$ is the gap at the ends.

As it stands, this cell is not suited for operation in a double image polariscope as described by N. Levin in MARCONI REVIEW, No. 44, 1933. In the first place it will be necessary, as in Fig. 1, to form an image of the first aperture between the polarizer and the analyser, and where this image is formed diverging plate electrodes will be positioned, but in the double image polariscope there are two images formed here, these images touching one another if the polarizer and analyser are of equal thickness.

Fig. 2 shows the vibration directions of each image. In order to effect the retardation, therefore, it will be necessary to embrace the two beams by electrodes so oriented that their normal would be inclined 45° to the line drawn in the centre of the two images. This would mean that the retardation will be small owing to the increased size of the gap. In order that the cell might be most effective, the electrodes will have to occupy the position of the dotted lines. It is obvious, therefore, that if this position is to be adopted, the vibration directions of the two beams must be modified. It would appear on first consideration that the only possible solution would be to transform the linear vibration of the light of both beams to either circular vibration or to linear vibration at 45° to the original. Both of these modifications can be carried out in practice.

In order to change the linear vibrations into circular vibrations, we might adopt the circular polariscope first suggested by Airy in 1833. The theory of this form of

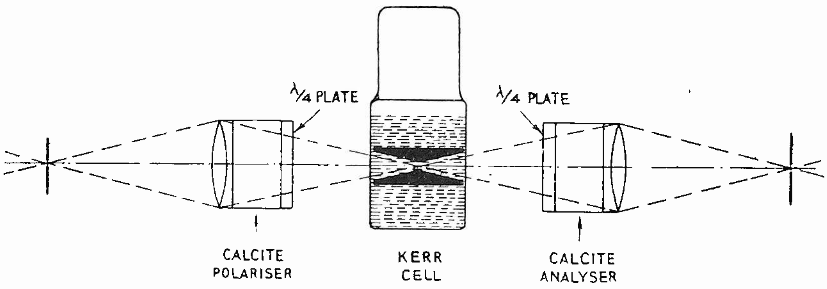


FIG. 1.

polariscope is given in most advanced text books (see Coker and Filon, " Photo Elasticity," page 73). Briefly, the device consists of the circular polarizer comprising a single image plane polarizer to which is affixed a quarter wave plate whose vibration directions are disposed at 45° azimuth to that of the plane polarizer. This ensures that the light leaving

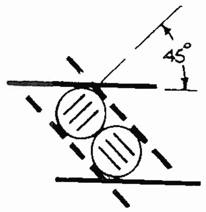


FIG. 2.

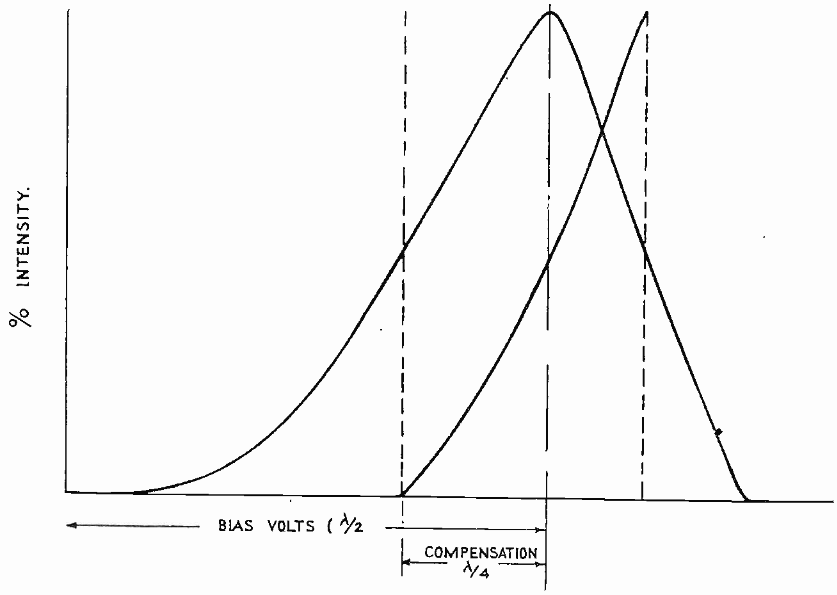


FIG. 3.

the combination is circularly polarized, and if the fast direction of a quarter wave plate is oriented in a right-handed manner from that of the plane polarizer, the light leaving the combination will be left-handed. The analyser comprises a similar quarter wave plate fixed to a plane analyser in such a manner that the light emerging from the quarter wave plate vibrates perpendicularly to the vibration direction of the plane analyser. In this manner it will be seen that extinction will take place, and this is the case for all relative orientations of circular polarizer and circular analyser.

When a bi-refringent section is introduced in such a polariscope, retardation takes place irrespective of the relative orientation of the section to either polarizer or analyser. We turn this property to account by embracing both beams of light within the Kerr cell by one pair of electrodes so that the gap between these electrodes is a minimum. If the quarter wave plates are cemented to their respective polarizer and analyser there should be no fear of extra light loss.

The second method to which we referred above is to rotate the vibration direction of the beams through 45°. This can be done by first taking the double image circular polariscope and cementing a second quarter wave plate to a polarizer. The orientation of this second plate should be such that the linear vibration emerging from it (the incident vibration being circular) should be 45° to the original vibration direction. A similar quarter wave plate would have to be added to the analyser. We might add, however, that this arrangement would somewhat complicate matters, and it is possible that complete extinction would not take place.

A second method of reducing the power required to drive the cell is to operate the cell on such a portion of its intensity curve that breakdown will not result. It does not seem expedient to operate the cell on the second slope for reasons of breakdown, but it is possible to bias the cell electrically to give quarter wave retardation and then to compensate optically for the quarter wave plate so that the negative half of the signal will reduce the visual field to extinction. This will be understood from Fig. 3. Owing to the nature of the intensity curve working on this part, it serves considerably to reduce the operating voltage.

# CORRECTION CIRCUITS FOR AMPLIFIERS

*Due to the rapidly increasing use of amplifiers, not only in the field of radio engineering but in other fields, as for instance, that of medical research, the demands made upon the designer become more and more exacting. A few years ago only two types of amplifier were in general use, viz., those operating at a spot frequency which could be selected in the radio frequency range, and those operating over the audio frequency spectrum. Since then the requirements of the television engineer have had to be met, and more recently those of research workers in the medical field, these latter frequently presenting problems not met with in radio engineering. In the medical field one is inclined to suspect that much work has been done with apparatus which, owing to its limitations, particularly that of frequency response, was unsuitable for the purpose in that it was only capable of providing part of the information sought.*

*It is thought that it may be of use to summarise the various methods that have been used to extend and control the response of an amplifier beyond the audio frequency range. Frequencies below and above this range will be dealt with separately. In this issue, low frequency circuits will be treated : in a subsequent issue means will be discussed for extending the response of an amplifier to include high frequencies. Transformer coupled amplifiers will not be considered since they are in general unsuitable for use outside the audio frequency spectrum (30—10,000 cycles per sec.), and only inside this range (say 100—3,000 cycles per sec. approx.) if phase shift of serious magnitude cannot be tolerated. It should, however, be mentioned that iron cored transformers can be used economically at frequencies up to about 100 kcs. provided they are required to operate over a relatively narrow band.*

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## Amplifier Response at Low Frequencies :

(1) *The Normal Uncorrected Circuit.*

THE stage represented in Fig. 1A can, as is well known, be represented by the network of Fig. 1B in which the valve is replaced by a generator, providing a potential $E_0 = \mu e_g$ and having internal resistance equal to the internal resistance of the valve $R_0$. The anode load is represented by $R_1$, the grid coupling condenser by $C_2$ and grid leak of following stage by $R_2$. This network may be represented as in Fig. 1C provided the impedance $\frac{1}{\omega C_2}$ of the coupling condenser is small compared with the resistance $R_2$ at the lowest frequency it is desired to handle. The anode load $R_1$ and grid leak $R_2$ may be lumped and represented by $R = \frac{R_1 R_2}{R_1 + R_2}$.

This approximation is only justified when the impedance $\frac{1}{\omega C_2}$ of the coupling condenser is less than one-third of the resistance $R_2$, since the proportion of the voltage developed across $R_1$ which appears across $R_2$ is

$$\frac{R_2}{R_2 - j\,\frac{1}{\omega C}}$$

and providing this limiting ratio is not exceeded, the output will be $\frac{3}{3 - j\,1} = \frac{3}{3.16} = .95$ of the input to the grid circuit, equivalent to .5 dB. loss. The maximum phase shift will, under these conditions, not exceed $\tan^{-1}\frac{1}{3} = 19°$.

We have then, for conditions under which the approximation of Fig. 1C holds :

$$\text{Voltage output} = E_o \frac{R}{R_o + R} = e_g \frac{\mu R}{R_o + R}$$

For frequencies at which the impedance $\frac{1}{\omega C_2}$ is greater than $R_2 \frac{1}{3}$ (Fig. 1B)

$$\text{let } Z_1 = R_1, \qquad Z_2 = R_2 - j \frac{1}{\omega C_2}$$

then $Z_1$ and $Z_2$ in parallel represent the load impedance $Z$, i.e.,

$$Z = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

Volts developed across the load $E_L = E_o \frac{Z}{R_o + Z}$

Volts developed across $R_2 = E = E_o \frac{Z}{R_o + Z} \cdot \frac{R_2}{Z_2}$

$$= E_o \frac{Z_1 R_2}{R_o (Z_1 + Z_2) + Z_1 Z_2}$$

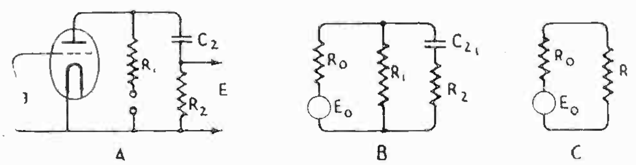


FIG. 1.

Putting $Z_1 = R_1$ and $Z_2 = R_2 - j \frac{1}{\omega C_2} = R_2 (1 - j \frac{1}{\omega T_2})$

where $T_2 = C_2 R_2$ the time constant of the grid circuit,

$$E = E_o \frac{R_1 R_2}{R_o R_1 + R_o R_2 (1 - j \frac{1}{\omega T_2}) + R_1 R_2 (1 - j \frac{1}{\omega T_2})}$$

Dividing throughout by $R_1 R_2$

$$E = E_o \frac{1}{1 + \frac{R_o}{R_1} + \frac{R_o}{R_2} - j \frac{1}{\omega T_2} (1 + \frac{R_o}{R_1})} \quad \ldots \ldots \quad (1)$$

which may be written

$$E = E_o \frac{1}{a - jb} = E_o \frac{1}{A \angle -\theta} = E_o \frac{1}{A} \angle \theta$$

i.e., the phase shift is always positive.

It will be seen that if $R_2$ is large compared with $R_1$ (not less than ten times) $1 + \frac{R_o}{R_1} + \frac{R_o}{R_2} \fallingdotseq 1 + \frac{R_o}{R_1}$ so we may write

$$E = E_o \frac{1}{1 + \frac{R_o}{R_1}} \cdot \frac{1}{1 - j \frac{1}{\omega T_2}} \quad \ldots \ldots \ldots \ldots \quad (1A)$$

The first of these fractions determines the normal stage magnification, and the second the shape of the frequency response curve. It is convenient to plot

$\dfrac{1}{1 - j\,\dfrac{1}{\omega T_2}}$ expressed in dBs. on a linear scale against frequency on a logarithmic scale for some convenient arbitrary value of $T_2$, say $.01 = T_0$, i.e., plot $-20 \log \left[1 + \left(\frac{1}{\omega T_0}\right)^2\right]^{\frac{1}{2}}$ against frequency. This is done in Fig. 2. It will be seen that at a frequency of 10 cycles per sec. $= f_0$ there is a loss of 5.5 dB. If the curve is transferred to a piece of transparent paper and a vertical line drawn through $f_0 = 10$ and a horizontal along the zero dB. line, then by sliding the paper with the

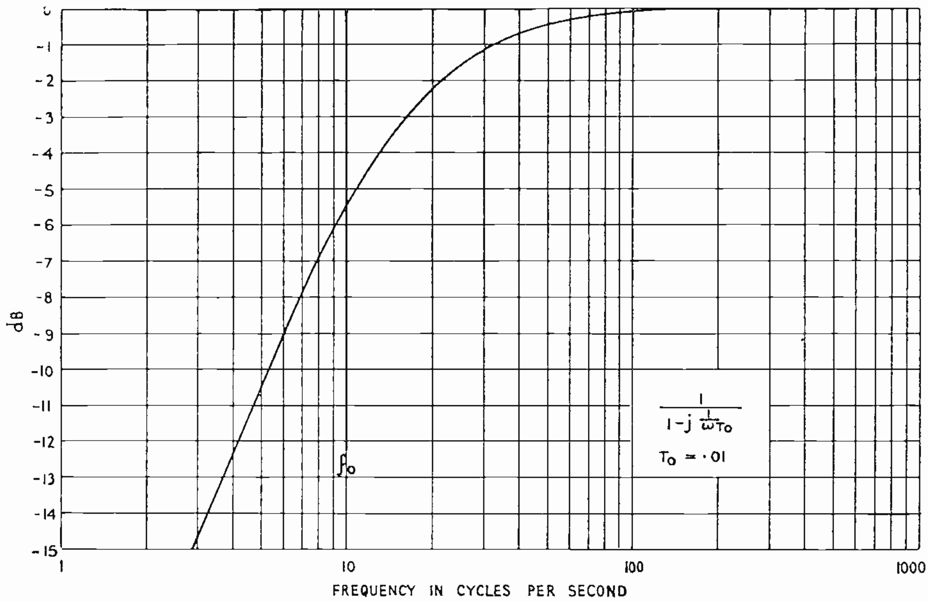


FIG. 2.

zero dB. line always coinciding with the scale, the response for any amplifier may be read off directly. For instance, suppose $T_2 = 1.0$ (i.e., 1 mfd. and 1 megohm) $T_2 = 100 \times T_0$ and $f_0$ must therefore coincide with $\dfrac{f_0}{100} = \dfrac{10}{100} = .1$ on the frequency scale, i.e., for any value $T = mT_0$ the vertical frequency line must be placed on $\dfrac{f_0}{m}$.

If the approximation of equation (1A) is not sufficiently accurate, equation (1) may be written

$$E = E_0 \frac{1}{1 + \dfrac{R_0}{R_1} + \dfrac{R_0}{R_2}} \cdot \frac{1}{1 - j\,\dfrac{1}{\omega T_2 \cdot \dfrac{1 + \dfrac{R_0}{R_1} + \dfrac{R_0}{R_2}}{1 + \dfrac{R_0}{R_1}}}}$$

$$= E_o \frac{1}{1 + \frac{R_o}{R_1} + \frac{R_o}{R_2}} : \frac{1}{1 - j\frac{1}{\omega T'_2}} \quad \ldots \ldots \ldots \quad (1B)$$

$$\text{where } T'_2 = T_2 \frac{1 + \frac{R_o}{R_1} + \frac{R_o}{R_2}}{1 + \frac{R_o}{R_1}}$$

i.e., $T'_2 = m' T_o$ and the vertical frequency line must be placed on $\frac{f_o}{m'}$.

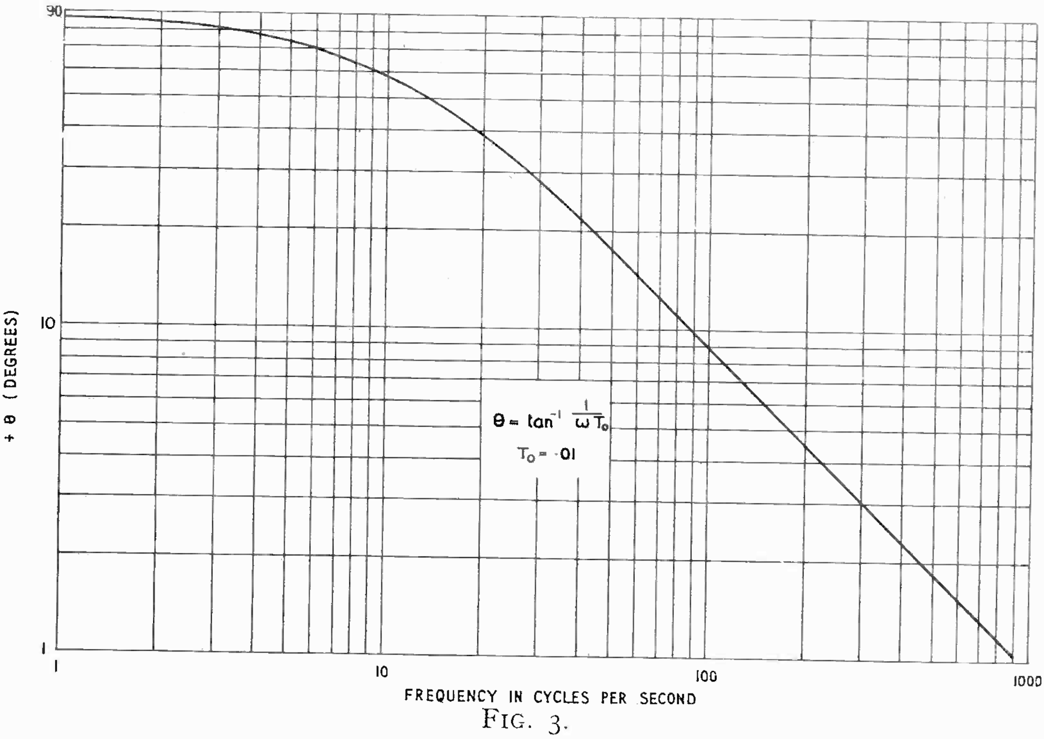


FIG. 3.

A similar method may be adopted with regard to phase shift. The phase shift is given by $\theta = \tan^{-1} \frac{1}{\omega T'_2}$. By plotting $\theta$ on a logarithmic scale against frequency on a logarithmic scale, the curve is straight over the most useful portion of the range. In Fig. 3 such a curve is shown for $T_o = .01$ and it is seen that at 100 cycles per second the phase shift is 9°, and at 10 cycles per second 58°. Either or both of these frequency lines may be conveniently selected for $f_o$. Then, as in the case of frequency response for any particular value of $T'_2 = m' T_o$, the frequency line is set on $\frac{f_o}{m'}$, and the phase shift read direct.

In either case, by reversing the procedure, i.e., by deciding on the shape of the response curve or the amount of phase shift, the correct value of $T'_2$ can be readily ascertained.

It might be thought that as good a response at low frequencies as desired can be obtained by making $T_2$ sufficiently large. Although this is true in some cases, practical considerations tend to set a limit to the maximum value. For instance, the input resistance of the grid circuit of the following valve (which is in parallel with $R_2$) does not in general exceed 5 megohms, and is in some cases less, unless valves are specially selected. This sets a top limit to $R_2$ of .5 megohm. The author prefers not to exceed .1 megohm whenever possible, as providing a bigger factor of safety. For amplifiers required to operate over a wide frequency spectrum, increasing the value of the condenser $C_2$ will inevitably increase its bulk, and this as will be shown later, will increase the stray capacities and will thus tend to reduce the maximum frequency at which the amplifier will operate without appreciable loss. Reliable paper condensers of .1 mfd. can be made in a reasonably small size so that a value of $T_2 = .01$ for this class of amplifier represents a normal value. From the curves, the response for this value of $T_2$ at 20 cycles per sec., is 2.2 dB. down and the phase shift 39°. Before methods of correction are discussed, the effect of auto-bias arrangements on frequency response will be noted, since any correction will of necessity have to compensate for any loss in the auto-bias circuit.

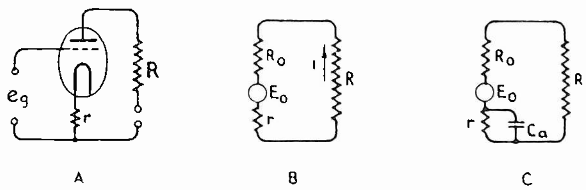


FIG. 4.

(2) *The Auto-bias Circuit.*

Consider the circuit of Fig. 4A and its equivalent network Fig. 4B, in which $r$ is the bias resistance. Let a current $i$ flow in the network, then the voltage of the equivalent generator $E_o$, which is the voltage between grid and cathode, is

$$E_o = \mu(e_g - ri)$$

$$\text{but } E_o = (R_o + R + r)\, i$$

$$\therefore\ \mu\,(e_g - ri) = (R_o + R + r)\, i$$

$$\therefore\ i = \frac{\mu\, e_g}{R_o + R + (\mu + 1)\, r}$$

$$\text{Output voltage } E = Ri = \mu\, e_g \frac{R}{R_o + R + (\mu + 1)\, r}$$

$$= \mu\, e_g \frac{R}{R'_o + R}$$

$$\text{where } R'_o = R_o + (\mu + 1)\, r$$

i.e., the internal resistance of the valve may be regarded as being increased by $(\mu + 1)\, r$. For normal values of bias resistance $R'_o$ may be twice the value of $R_o$, i.e., the normal voltage output may be halved.

If, now $r$ be shunted by a condenser $C_a$ (Fig. 4C), clearly the impedance of the cathode network will tend to zero at some high frequency and tend to $r$ at some low frequency. In other words, the response of an auto-biassed stage will fall to some constant value as frequency is reduced in a manner similar to Fig. 5A, which represents the ratio of $R_o$ to $R'_o$ in dB. against frequency for a screen grid valve of internal resistance 350,000 Ohms, $\mu = 1{,}000$ and $r = 350$ Ohms, $C_a = 50$ mfd. The phase shift introduced is shown in Fig. 5B. It will be observed that the use of the cathode network will cause positive phase shift at the output.

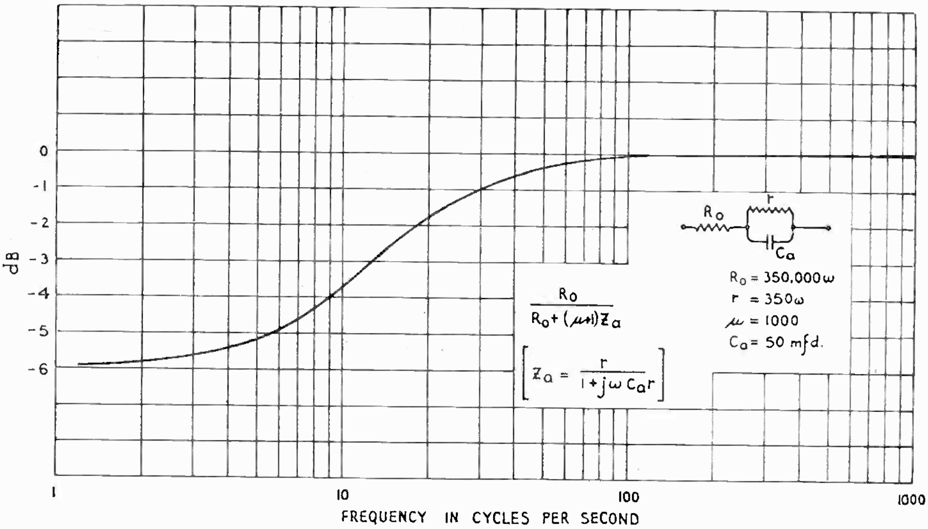


FIG. 5A.

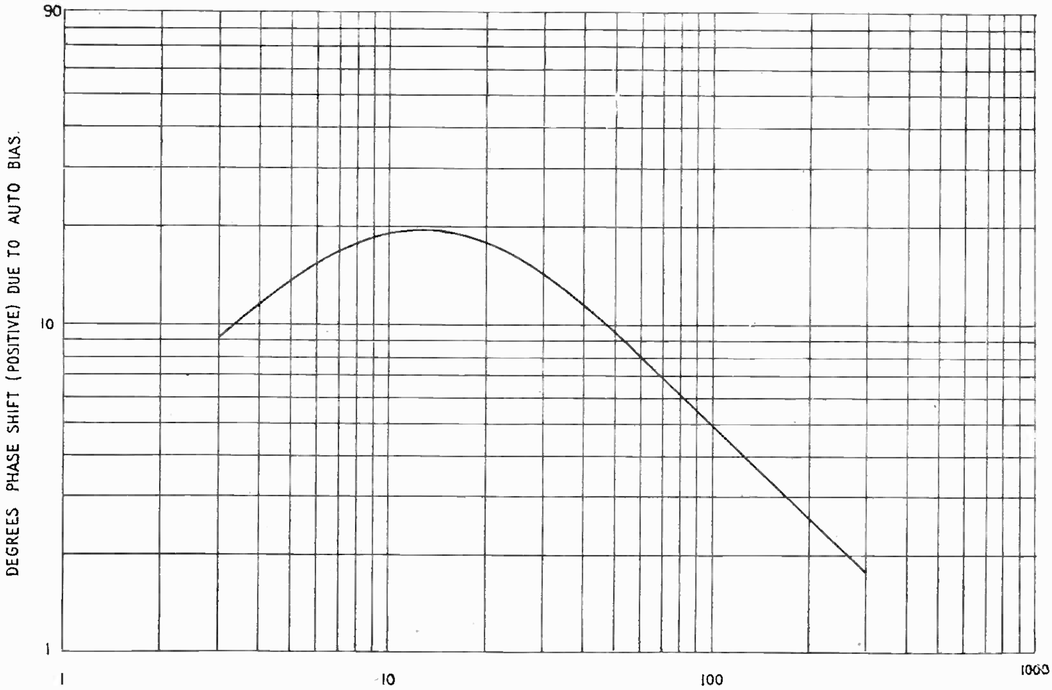


FIG. 5B.

In the above $R'_o = R_o + (\mu + 1)\, Z_a = R_o + (\mu + 1)\, \dfrac{r}{1 + j\omega T_a}$. This curve may also be traced and shifted along the frequency axis to correspond to variations of $C_a$, values for the other components remaining constant.

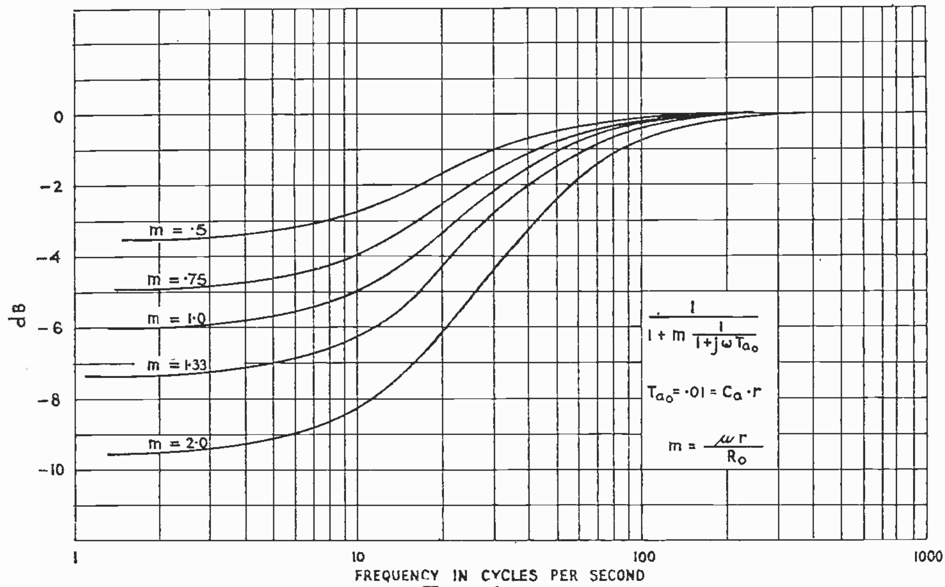


FIG. 6A.

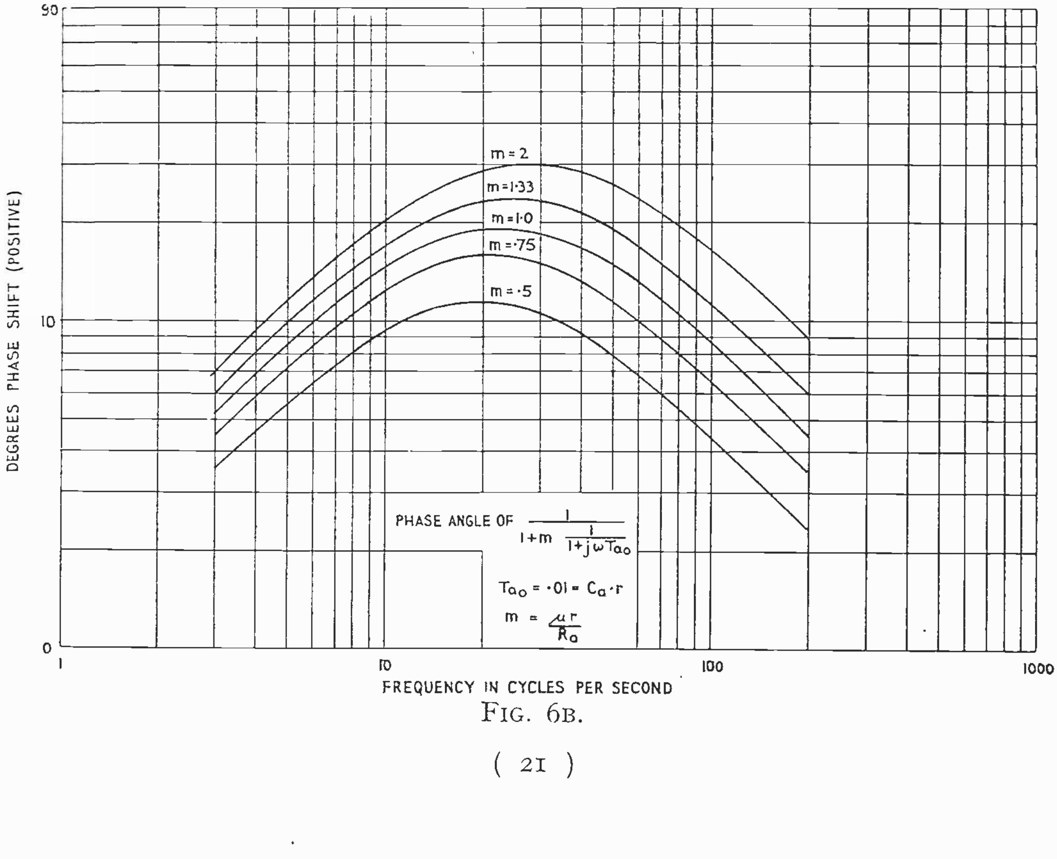


FIG. 6B.

The ratio $\dfrac{R_0}{R_0 + \dfrac{(\mu + 1)\,r}{1 + j\omega T_a}}$ may be written $\dfrac{1}{1 + m\,\dfrac{1}{1 + j\omega T_a}}$ where $m = \dfrac{\mu r}{R_0}$.

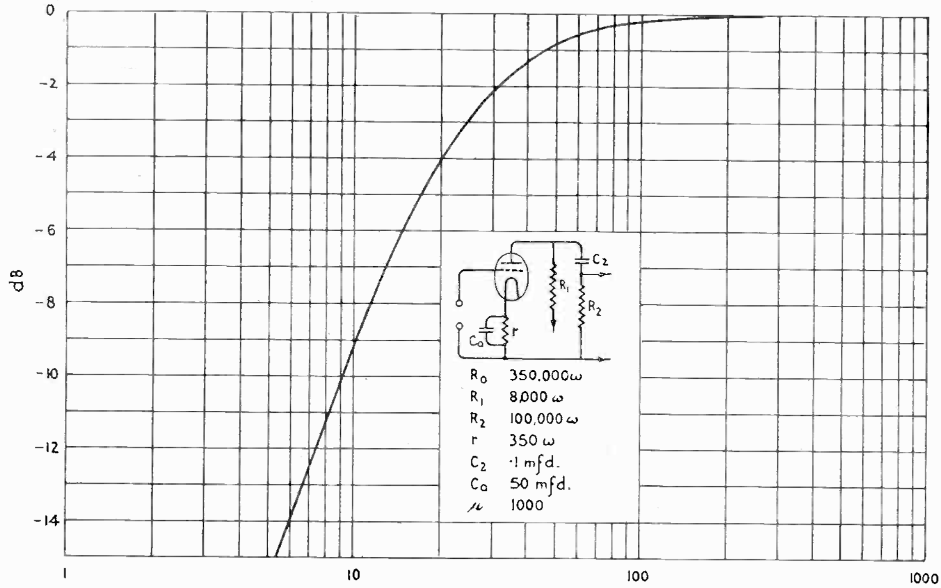


FIG. 7.

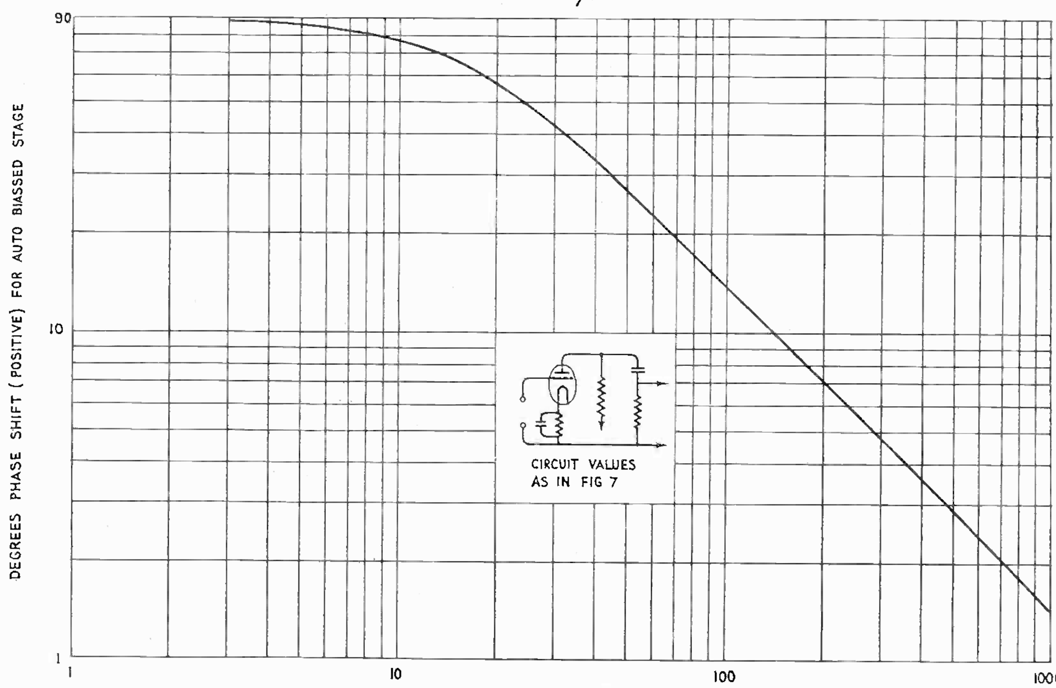


FIG. 8.

Curves for this expression are shown in Fig. 6A, while Fig. 6B shows corresponding values of phase shift. These curves are drawn for $T_{ao} = C_a r = .01$ with $m$ as parameter. As an example of their use, suppose a valve is employed in which $\mu = 400$, $R_o = 100{,}000$, $r = 500$, then $m = 2$. Then the curve for $m = 2$ will represent the response when $C_a = \frac{T_{ao}}{r} = \frac{.01}{500} = 20$ mfd. If the cut-off is too great then the curve may be traced and shifted to the left after the manner described in connection with Fig. 2, the value of $C_a$ being increased in the same proportion as $f_o$ (which may be chosen arbitrarily) is reduced.

We can now estimate the frequency response and phase shift of an auto-biassed stage. Using the valve above with anode load = 8000 Ohms, $C_2 = .1$ mfd., $R_2 = 100{,}000$ Ohms, $T'_2$ is sufficiently close to $T_2 = .01$ to use Figs. 2 and 3. These must be added to Figs. 5A and 5B respectively, and the resulting curves are shown in Figs. 7 and 8 (this procedure is only justified when the load impedance $R_1$ is small compared with $R_o$ and $R_2$).

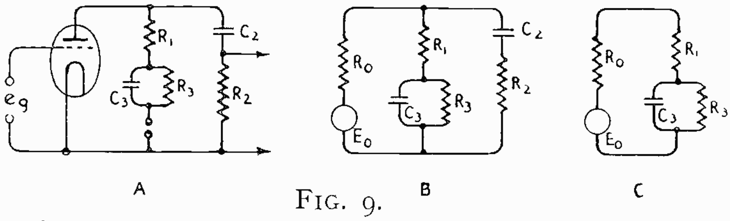


FIG. 9.

It will be found by superposing Figs. 2 and 3 on Figs. 7 and 8 respectively that no exact fit can be obtained, as might be expected, but fair correspondence can be obtained over a big range. The respective $f_o$ lines do not, however, overlie the same values of frequency. This indicates that a circuit designed to correct the curve and providing a gain curve and negative phase shift curve similar in shape to Figs. 2 and 3 will not compensate for both frequency loss and phase shift exactly. It is usual to effect a compromise by allowing say $\pm .5$ dB. and $\pm 20°$ or less in the range required.

Correction circuits are discussed in the next section.

**Low Frequency Correction :**

(1) *Correction in Anode Circuit.*

The main object of a correction circuit is to increase the amplification available over a range of frequencies at which loss occurs in curves of the type shown in Fig. 2 and Fig. 7 without affecting the amplification over the flat portion of the frequency response characteristic. If the anode load is caused to increase over this range, the desired effect will be attained. This may be performed by inserting in series with the anode load a resistance and capacity in parallel and adjusting the time constant $T_3$ of this circuit to cause it to operate in the desired range. Fig. 9A shows such a circuit, and Fig. 9B the equivalent network. It will be seen that the circuit $C_3$ $R_3$ is similar to an anode decoupling circuit, and may, indeed, be used as such. It will be appreciated, however, that this circuit will not provide effective decoupling at frequencies below that at which correction begins to take effect. If decoupling is necessary at these frequencies an additional decoupling circuit must be used, but care taken to ensure that its time constant is such that its impedance is negligible at the lowest frequency it is desired to handle. The shape of the response curve will thus be unaffected in the required range.

Disregarding for the moment the circuit $C_2R_2$, i.e., consider Fig. 9C (this is permissible if $R_2$ is assumed large compared with $R_1$), the voltage $E_L$ developed across the total anode load can be written

$$E_L = E_0 \frac{1 - j\,\dfrac{1}{\omega T_3}\,\dfrac{R_1 + R_3}{R_1}}{1 + \dfrac{R_0}{R_1} - j\,\dfrac{1}{\omega T_3}\left(\dfrac{R_0}{R_1} + \dfrac{R_3}{R_1} + 1\right)}$$

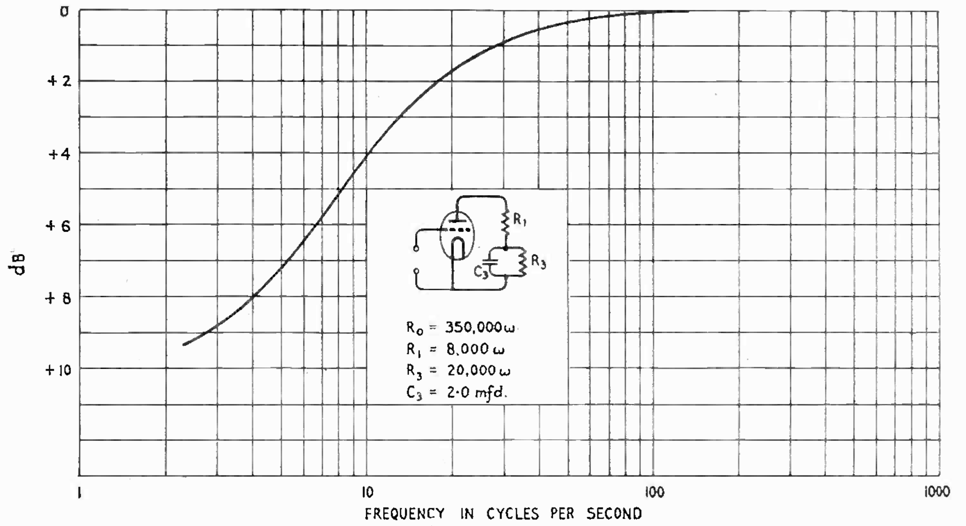


FIG. 10.

This may again be transformed and written

$$E_L = e_g \frac{\mu R_1}{R_0 + R_1}\left(\frac{1}{1 - j\,\dfrac{1}{\omega T''_3}}\right) \div \left(\frac{1}{1 - j\,\dfrac{1}{\omega T'_3}}\right)$$

$$\text{where } T''_3 = T_3 \times \frac{R_0 + R_1}{R_0 + R_1 + R_3}$$

$$T'_3 = T_3 \times \frac{R_1}{R_1 + R_3}$$

The first fraction determines the normal amplification of the stage. The second fraction may be read off in dBs. from Fig. 2 after shifting the curve, as can the third fraction. The dB. values of the third fraction must be read as positive owing to the division sign, those of the second fraction being negative.

The algebraic sum is determined, the result being positive, i.e., a gain is obtained as frequency is reduced. This curve is most conveniently plotted as in Fig. 10 with positive dBs. downwards. It can be shifted along the frequency axis to correspond to changes in $C_3$ provided the other components are not altered. It cannot be used to give changes of $R_3$, since a change of $R_3$ will alter the amount of gain available due to the correction circuit, whereas $C_3$ determines the frequency range over which this gain is available. In a similar way the phase shift figures may be derived from Fig. 3, those for the third fraction being read as negative, and for the second as positive. As the larger value is negative, the resultant phase shift is negative. Fig. 11 shows the final curve. This is best plotted with negative phase shift upwards. Figs. 10 and 11 are plotted for $R_o = 350{,}000$, $R_1 = 8{,}000$, $R_2 = 20{,}000$, $C_2 = 20$ mfd. The similarity between these curves and those for auto-bias should be noted.

Having now determined the frequency characteristic of the anode circuit, the grid coupling components may be chosen. The curves of Fig. 2 and Fig. 3 are superposed upon Figs. 10 and 11 respectively until they overlie. An exact fit will be found impossible and it will also be noted that for the best fit for frequency

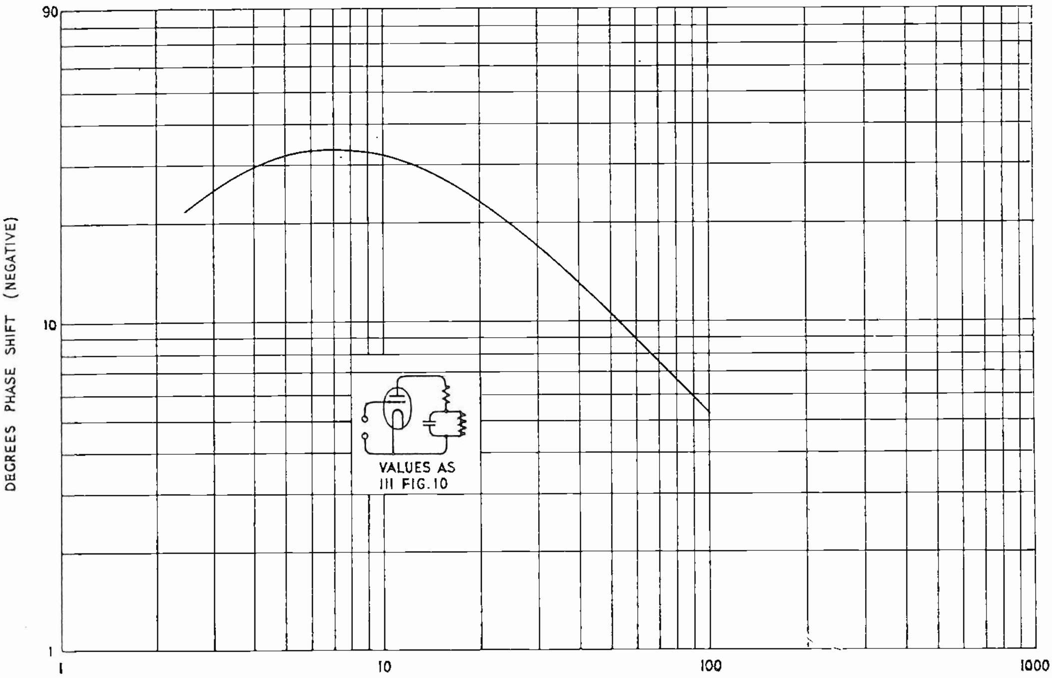


FIG. 11.

response or phase, $f_o$ overlies two different frequencies. One can therefore choose a value of $f_o$, which of course must be the same for both curves, to provide approximately linear response while tolerating a certain amount of phase shift, or approximately zero phase shift while tolerating a slight variation in frequency response, or again, a compromise may be effected by permitting a slight variation from the ideal in each case. The use to which the amplifier is to be put will decide which solution should be adopted.

As an example, the phase shift curve is chosen to give the best fit. This is accomplished when $f_o = 10$ overlies 6 cycles on Fig. 11. The difference between the two curves is plotted on Fig. 12B. In a similar way, the difference between the curves of Fig. 10 and that of Fig. 2 transposed for $f_o = 10$ to overlie 6 cycles is shown in Fig. 12A. These curves will therefore represent the overall curves of the stage, provided $R_2$ is high (say 10 times the anode load). The time constant of the grid circuit is $T_o \times \frac{f_o}{6} = .0167$. If $R_2$ is chosen as 330,000 Ohms, $C_2 = .05$

mfd., any variation from the curves due to the shunting effect of the grid leak will be negligible.

It will sometimes be found that even with a compromise, the variation from the ideal is greater than permissible in the frequency range required, i.e., the trial value of $T_3$ is too small. The curves Fig. 12A and B may be shifted to the left so as to bring the minimum frequency it is desired to handle within the range of permissible variation, after which $T_2$ and $T_3$ must be multiplied by the ratio of the two frequencies through which any one point on the curve passes before and after shifting, e.g.. for a 12° phase shift at 5 cycles (Fig. 12B) $T_2$ and $T_3$ must be multiplied by 2. This will mean that $C_3$ must be increased by this ratio and the product

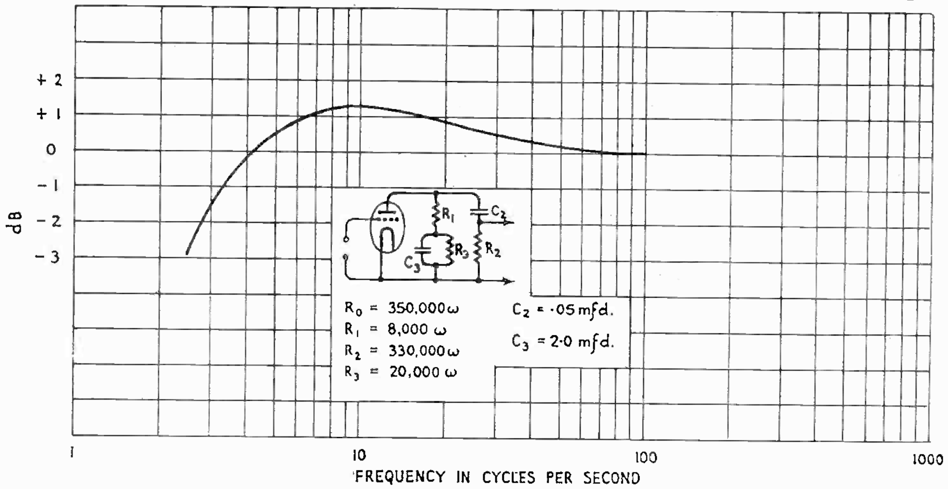


FIG. 12A.

$C_2 R_2$ by the same amount (it being immaterial which component of $T_2$ is increased). If, on the other hand, the required minimum frequency is well inside the range of permissible variation, $T_2$ and $T_3$ may be decreased if desired, thus effecting an economy in components.

If the value of $R_3$ is comparable with $R_2$, it is desirable to make a check on this last curve. This is due to the fact that the shunting effect of the grid coupling circuit is not taken into account in the foregoing method. This is best done by evaluating the expression given below (equation 2) for a frequency corresponding to a drop of say between 2 and 4 dBs. on the curve, or a phase shift of 10 to 20 degrees.

The voltage developed across the resistance $R_2$ in Fig. 9B may be shown to be

$$\left.\begin{aligned} E &= E_0 \left(\frac{1}{a - jb}\right) \div \left(\frac{1}{1 - j\dfrac{1}{\omega T'_3}}\right) \\ \text{where } a &= 1 + \frac{R_0}{R_1} + \frac{R_0}{R_2} - \frac{1}{\omega T_2 \omega T_3}\left(1 + \frac{R_0 + R_3}{R_1}\right) \\ b &= \left[\frac{1}{\omega T_3}\left\{\frac{R_0}{R_1} + \left(1 + \frac{R_3}{R_1}\right)\left(1 + \frac{R_0}{R_2}\right)\right\} + \frac{1}{\omega T_2}\,\frac{R_0 + R_1}{R_1}\right] \\ T'_3 &= T_3 \frac{R_1}{R_1 + R_3} \end{aligned}\right\} \quad (2)$$

The first fraction must be calculated and the second may be read off Figs. 2 and 3 after curve shifting, the dB. values being read as positive and the phase shift as negative owing to the division sign.

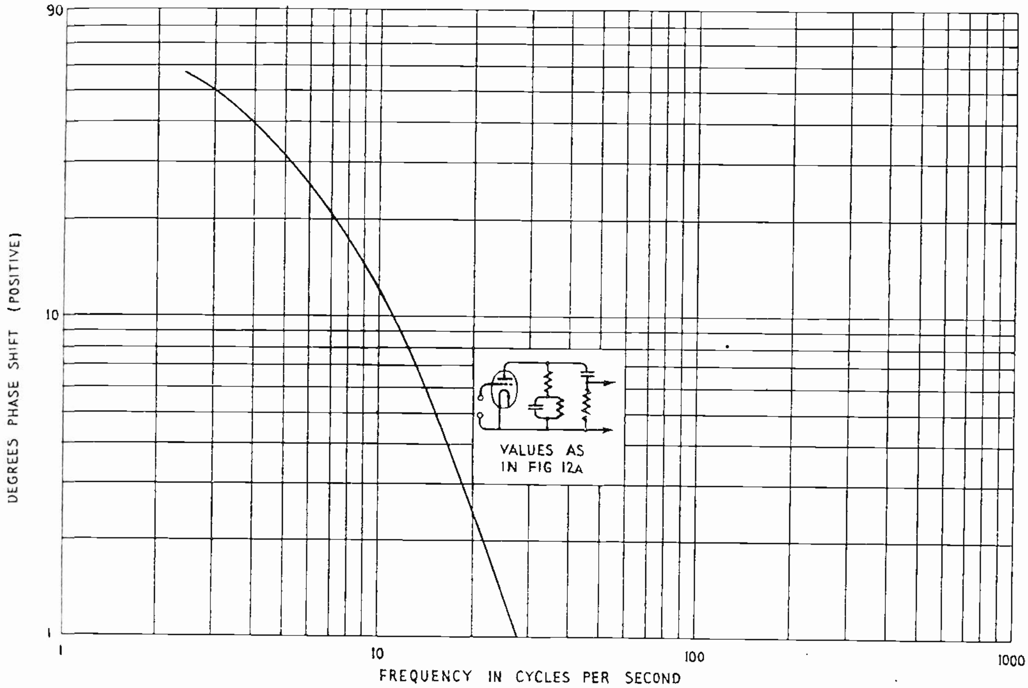


FIG. 12B.

It is realised that in many cases it may be quicker to evaluate equation 2 for a number of frequencies, but the graphical method is given, as in the first place it will quickly provide a rough estimate for suitable values of $T_2$ and $T_3$ in any particular amplifier, and secondly, calculation from equation (2) can be very laborious unless a slide rule of the Davis-Grinstead type is available.* This rule transforms expressions of the type $a + jb$ to $r \angle \theta$ with one setting. That used by the author

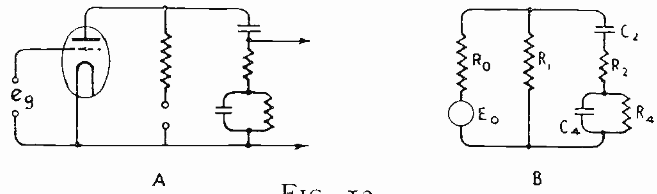


FIG. 13.

was engraved in the Company's Works and has 10 inch scales, the upper (A) being two decades, and the lower (D) one decade. Values of $\theta$ from 0.6 degree upwards can be determined. It has proved invaluable.

It will be understood that although no provision was made in the foregoing example for auto-bias, the procedure is similar, the auto-bias curve of the type

* Inst. of Post Office Electrical Engineers Printed Paper No. 110, p.7.

shown in Fig. 5 being combined with curves similar to Figs. 10 and 11 before selection of $T_2$.

Correction as described, has the disadvantage that it involves an increase in anode battery voltage to compensate for the drop in the resistance $R_3$.

(2) *Correction in the Grid Circuit.*

It is possible in some circumstances to modify the grid circuit to provide control of the frequency response characteristic. The circuit that may be employed is shown in Figs. 13A and B.

The circuit depends for its operation on the fact that at normal frequencies the effective anode load comprises $R_1$ and $R_2$ in parallel but below some predetermined frequency, the effective anode load comprises $R_1$ and $(R_2 + R_4)$ in parallel. The circuit will thus operate successfully when $R_1 \geq R_2$.

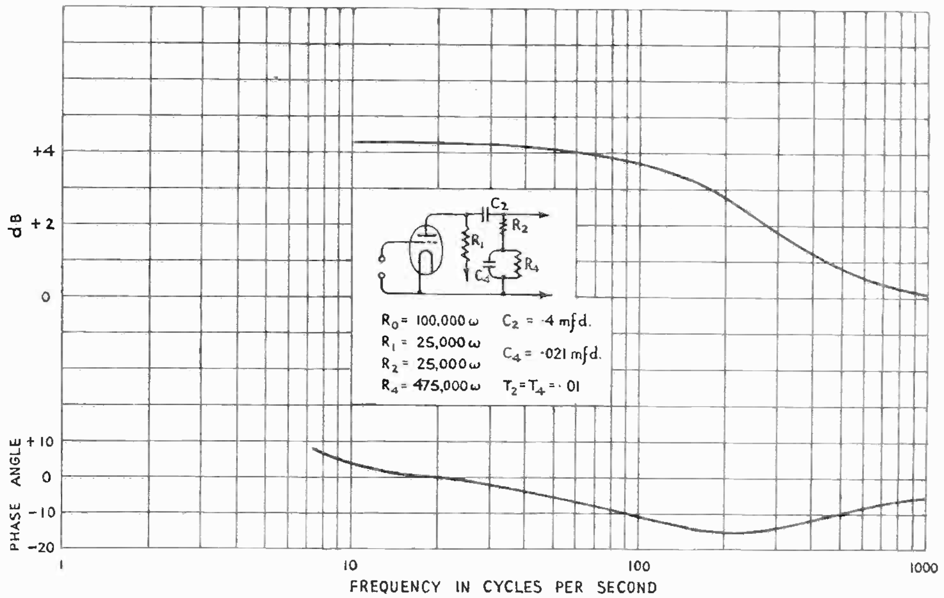


FIG. 14.

If, observing this condition, $T_2$ is made large and $T_4 = C_4R_4$ small, then a rising characteristic will be obtained as frequency is reduced. If, however, $T_4$ is such that the impedance of the circuit $C_4$, $R_4$ does not increase as frequency is reduced until the impedance of $C_2$ becomes comparable with $R_2$, then the frequency characteristic will at first fall slightly, then remain substantially level over a range determined by $T_4$, and finally merge into the characteristic given by $C_2\ (R_2 + R_4)$. It will be appreciated that if $C_2$ and $R_2$ were given their maximum permissible values, with a view to obtaining best low frequency response, no advantage is to be gained by using this type of circuit since the input impedance of the following value will not permit of any sensible improvement, unless special care is taken to keep this high.

Fig. 14 shows an application of the circuit for obtaining a correction effect below 1,000 cycles per second, it being assumed that a loss occurs over this range in some preceding apparatus. Note that the phase shift is negative over a considerable range, and that correction for phase can thus be obtained since loss in response is generally accompanied by positive phase shift. These curves are calculated from equation (3) below. An analysis of the network Fig. 13B shows that the output voltage E is

$$\left.\begin{aligned} E &= E_0 \left(\frac{1}{a - jb}\right) \div \left(\frac{1}{1 - j\,\dfrac{1}{\omega T'_4}}\right) \\ \text{where } a &= 1 + \frac{R_0}{R_1} + \frac{R_0}{R_2} - \frac{1}{\omega T_2 \omega T_4}\left(1 + \frac{R_0}{R_1}\right) \\ b &= \left[\frac{1}{\omega T_4}\left\{\frac{R_0}{R_2} + \left(1 + \frac{R_0}{R_1}\right)\left(1 + \frac{R_4}{R_2}\right)\right\} + \frac{1}{\omega T_2}\left(1 + \frac{R_0}{R_1}\right)\right] \\ T'_4 &= T_4 \frac{R_2}{R_2 + R_4} \end{aligned}\right\} \quad (3)$$

The first fraction must be calculated and the second may be read off Figs. 2 and 3 after curve shifting, the dB. values being read as positive and the phase shift as negative owing to the division sign; the dB. values of the first fraction will be negative and the phase shift positive. The algebraic sum of the corresponding values will give the response and phase shift curves.

## SYMBOLS.

| | |
|---|---|
| $E_0 = \mu e_g =$ | equivalent generator voltage. |
| $E$ | Output voltage to following stage |
| $E_L$ | Voltage across anode load. |
| $R_0$ | Internal resistance of valve. |
| $R_1$ | Resistance of anode load. |
| $R_2$ | Resistance in grid circuit. |
| $R$ | Effective anode load, i.e., $R_1$ and $R_2$ in parallel. |
| $R_3$ | Anode correction resistance. |
| $R_4$ | Grid correction resistance. |
| $r$ | Auto-bias resistance. |
| $C_2$ | Grid coupling condenser. |
| $C_3$ | Anode correction condenser. |
| $C_4$ | Grid correction condenser. |
| $C_a$ | Auto-bias condenser. |
| $T_2 = C_2R_2$ | Time constant of grid coupling circuit. |
| $T_3 = C_3R_3$ | ,, ,, ,, anode correction circuit. |
| $T_4 = C_4R_4$ | ,, ,, ,, grid ,, ,, |
| $T_a = C_ar$ | ,, ,, ,, auto-bias circuit. |
| $T_0$ | Arbitrary time constant of .01 selected as a standard of comparison. |
| $f_0$ | The frequency selected for use with $T_0$ (usually 10 cycles per second) as a standard for curve shifting. |

(*To be continued.*)

# MARCONI NEWS AND NOTES

## WIRELESS COMMUNICATION FOR AFGHANISTAN.

THE Afghanistan Government has placed a contract with the Marconi Company for the supply and erection of five wireless stations in the most important centres in Afghanistan.

The installation of an up-to-date wireless system of communication will be a valuable contribution to the development of Afghanistan's commercial and social relations with other countries and an equally important factor in the country's internal communication service.

The most powerful of the five new stations will be situated near Kabul, and the other four at Maimana, Khanabad, Khost, and Diyazungi.

The Kabul station will be fitted with a short-wave transmitter of the Marconi SWB.11 type, suitable for the transmission of telegraphy and telephony, and two receiving installations, one of the Marconi R.C.52 type and the other of the Rg.28 type.

The Marconi SWB.11 transmitter covers a waverange of 15-80 metres with an output of 5-6 kilowatts on telegraphy, and $3\frac{1}{2}$-$4\frac{1}{2}$ kilowatts on telephony, to the aerial feeders.

The receiving and transmitting sites will be separate, and will be about 10 miles from the City of Kabul, where the central telegraph office for the control of the wireless stations will be located.

At Maimana, Khanabad, Khost, and Diyazungi, Marconi S.8A transmitters and Rg. 28 receivers will be installed, and these installations will be used for internal communication.

**Aircraft Wireless.**

THE Marconi Company, as the leading British wireless company, showed a comprehensive range of aircraft wireless apparatus at the Static Exhibition which was held in connection with the Flying Display arranged by the Society of British Aircraft Constructors at Hendon on June 29th and July 1st. This exhibition coincided with the British Royal Air Force Flying Display at Hendon, and many visitors were present from countries throughout the world.

In addition to types of aircraft wireless installations which have proved their reliability by long service on the British Empire air routes and in connection with


*Marconi Exhibit at Hendon Flying Display.*

civil and military air services in all parts of the world, a number of new designs based on the Marconi Company's wide experience of flying services were included.

Equipment for aerodromes and wireless D.F. "Homing" apparatus which has proved of the greatest value in navigation over long distances and under difficult conditions, were also shown.

The exhibits included the following types of equipment :—A.D.37/38 (medium and short wave, 170 watts) ; A.D.41/42 (medium wave, 170 watts) ; A.D.43/44 (intermediate wave, 20 watts) ; A.D.45/46 (medium wave, 20 watts) ; A.D.49/50 (medium wave, 65 watts) ; A.D.52 direction finding and "Homing" attachment ; and two other sets which were exhibited for the first time. These were a light-wave aircraft receiver and an airport receiver.

**Light Aircraft Receiver.**

The Marconi Light Aircraft Receiver is a small and economical wireless receiver of the superheterodyne type, designed particularly for the service of owners of light aeroplanes who may wish to fit a wireless receiver to intercept the weather broadcasts detailed in British Air Ministry Notice to Airmen No. 63 (11th June, 1935), transmissions from the aeronautical stations operated by the Air Ministry and similar transmissions sent out in other countries.

A further use for these sets is their installation on air liners where broadcast entertainment is desired. Attendants on these air liners can, by means of this set, switch on entertainment broadcasts for passengers equipped with headphones, as in the case of the railway broadcasting service. Gramophone records can also be used to entertain passengers, and a microphone can be incorporated in the circuit so that the officer in charge can make announcements to passengers, give information with regard to landing times, or describe the country over which the aeroplane is travelling.

**The Marconi Airport Receiver.**

This receiver is intended to supply a demand which has arisen in connection with flying clubs, municipal air ports and private aerodromes where there is no Air Ministry wireless station but where airport officials and club members wish to receive weather reports and other Air Ministry transmissions such as information regarding the positions of machines on various routes, etc.

By means of this receiver it is also possible for members of flying clubs to receive broadcast entertainment or entertainment from gramophone records.

The receiver is a 9-valve superheterodyne radio gramophone of the latest type designed in accordance with the long and specialised experience of the Marconi Company in broadcasting and aircraft wireless matters. It has a waverange of from 200 to 500 metres and from 750 to 1,700 metres. A junior model of this equipment is available operating with five valves instead of the nine valves used in the senior model already described.

## Brussels Exhibition.

BRITISH wireless exhibits figure prominently in the " Alberteum," Palace of Science, at the Brussels Exhibition, a series of exhibits representing varied modern applications of wireless and high-frequency technique, contributed by the Marconi Company, occupying one of the principal sections.

A great part of the display in the " Alberteum " is devoted to the progress of wireless, and Marconi apparatus which is shown includes a complete broadcasting studio equipment, a portable military field station, diathermy apparatus, and a representative collection of historical sets and relics illustrating the principal phases of development from Marconi's earliest experiments. The Société Anonyme Internationale de Télégraphie sans Fil, the International Marine Sounding Device, Société Anonyme, and the Comité International Radio-Maritime also have comprehensive exhibits.