No. 68

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January-March, 1938

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THE MARCONI REVIEW

No. 68. *January-March, 1938.*

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EDITOR'S NOTE

THE MARCONI REVIEW, now in its tenth year, is indicating its more mature outlook by the introduction from the present date of certain new features which we believe will serve the technical need of the present time better, perhaps, than other features which they must displace from this magazine. In future it will be issued as a quarterly instead of every two months, and the number of pages will be increased more or less in proportion, as found necessary.

The articles we hope to publish will, on the whole, have a greater technical bias than hitherto ; and articles on communication problems involving advanced mathematical treatment, which an Editor for reasons of extra cost, and because they appeal to a more limited public, may often hesitate to accept, will be given adequate space in our pages.

We have arrived at this decision in view of the fact that refinements in performance called for in communication apparatus to-day necessitate a greater use of analytical methods in design, and a familiarity with such methods must therefore be encouraged to the utmost.

Under the heading "Patent Abstracts" we hope to present to our readers a summary of the more recent Marconi patents, in which, for the benefit of our licensees, the particular feature useful for practical application as distinct from the claims of each patent will be stressed. The diagrams illustrating these abstracts will be drawn as wireless engineers view the circuits discussed, which is often very different from the point of view of the Patent Office draughtsman.

The section devoted to News and Notes will be discontinued. It is with a feeling of regret that we close down this feature, which has served to keep our friends informed of the use of Marconi apparatus in so many widely divergent fields of activity, but we realise that our personal feelings must give way to the interests of our technical readers. As we shall cease to publish News and Notes, it is clear that we can no longer devote space to more personal matters, such as staff activities, obituaries and information generally which more properly belongs to a House Journal.

Finally, although the production costs of the magazine are likely to be increased, and in a periodical of this character one does not expect such a factor to be offset by greatly increased sales, the Marconi Company has decided, as a friendly gesture to our readers, to reduce the annual subscription to 10s. 6d. (inclusive of postage), or 2s. 6d. per copy.

Under these new conditions, and in the belief that the changes will be understood and appreciated by our readers, THE MARCONI REVIEW now embarks on a more staid, but possibly a more useful career wherein research and technique will be given increased space for the recording of up-to-date practice in the science and art of wireless communication engineering.

H. M. D.

THE WIRELESS PILOT

Marconi's Wireless Telegraph Company Limited recently received an opportunity, through the good offices of Capt. Philip Bailey, Director of P.B. Deviator, Ltd., to instal and test on his Puss Moth aeroplane a wireless direction finder with linking apparatus for connection to the machine's automatic pilot. The following article gives a brief account of the experiments made with the object of steering an aeroplane along a predetermined course by causing the wireless signals received by the direction finder to actuate the rudder through a relay chain and the gyroscopically controlled servo motors of the automatic pilot.

The Automatic Pilot.

FROM time to time mention is made in the technical and daily Press of the flight of aircraft being controlled by the automatic or robot pilot.

Some of the inventions relating to gyroscopic control date back many years, and to-day several types of such mechanism have been developed to a high pitch of efficiency so that the employment of the automatic pilot has now become commonplace.

To those who desire to study the subject of gyroscopically controlled flight, a paper read before the Royal Aeronautical Society in February last, is commended.* For the present it will suffice to explain that in most systems auxiliary steering motors or servo motors are provided which may be mechanically connected at will to the flying controls of the aircraft. These motors are actuated by air or oil under pressure, the flow of which is controlled by valves operated by the gyroscopic mechanism when the machine deviates from its course in azimuth or pitch.

For our experiments a form of automatic pilot marketed by Messrs. P.B. Deviator, Ltd., was already provided in the Puss Moth aeroplane placed at our disposal. In the design of this device use is made of a single gyro spinning on an axis in line with the line of flight. A moment's consideration will show that a gyroscope so placed can be made to exercise control of lateral deviations and also of pitch. It is only the former control with which we are at present concerned.

In the P.B. Automatic pilot the gyroscopic rotor is driven by a shaft via a ball and cup. This form of coupling is designed to permit of continuous slip between driving and driven members. The continuous slip has the effect of permitting relative movement of the rotor to occur about the driving spindle with the minimum of friction.

In line with the driving shaft, but on the other side of the rotor and attached to it, is the valve actuating spindle.

The gyro when spinning takes up a set position in space and when the light valve triggers are moved against this spindle by the yawing of the craft they will at once be operated by it, and will cause a compensating movement of rudder or elevator to give a course correction.

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* "Aeroplane Stability and the Automatic Pilot," by F. W. Meredith, B.A., and P. A. Cooke, O.B.E., M.C.

It is necessary for the pilot to be provided with a convenient means of altering course. In the case of the P.B. system an extremely simple and effective electrical control is arranged. This form of control is achieved by attaching to the end of the gyro spindle, which operates the valve mechanism, a soft iron armature which normally takes up a position in the centre of a " pack " of four radially disposed

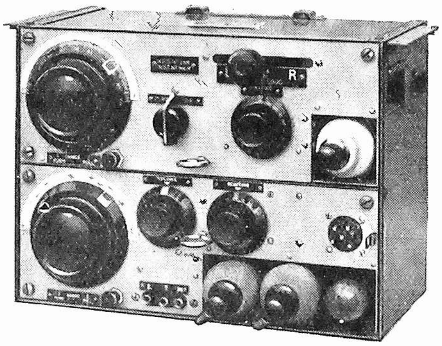


FIG. 1.

FIG. 2.

electro magnets. By energising any one of these four magnets (one pair is placed vertically for " left-right " control, and the other pair horizontally for pitch control), the armature is attracted, the gyro rotor displaced or precessed, and the corresponding valve gear operated.

By including a variable resistance in series with the 12 volt supply to the electro magnets the rate of turn can be conveniently adjusted to suit requirements. Four press buttons are provided on the dash board labelled " left," " right," " up " and " down."

The Direction Finder.

In selecting the type of direction finder for this experiment, two considerations were borne in mind. Firstly the accuracy of steering would be dependent upon the " goodness " of the direction finder, as such. Secondly, for the effective working of the control and for easy observation of results, bearings must be taken on stations emitting a continuous carrier. It was therefore decided to employ for the initial experiments, a D.F. which would give the best practical performance when receiving from a long wave broadcasting station, and which would give visual indication of course deviations.

Of the several available types the AD.5062B—the performance of which was known—was selected.

The AD.5062B apparatus, a photograph of which is shown in Fig. 1, comprises a four valve receiver consisting of two stages of H.F. amplification, a detector, and one low frequency amplifier, and in which provision is made for the formation of a cardioid diagram—by combining the signals from the loop and a trailing aerial. A hand switch for reversing the currents in the trailing aerial, in relationship to those in the loop, is provided as an integral part of the receiver.

Connected to the receiver proper is the type 626 Visual indicating device, Fig. 2, which consists of a motor-driven commutator, the function of which is to reverse at about 100 p.p.s. the H.F. currents in the trailing aerial in relationship to those in the loop (in the same manner as with the hand switch) and synchronously to reverse the rectified L.F. output from the receiver to the Visual indicating meter—a centre reading microammeter.

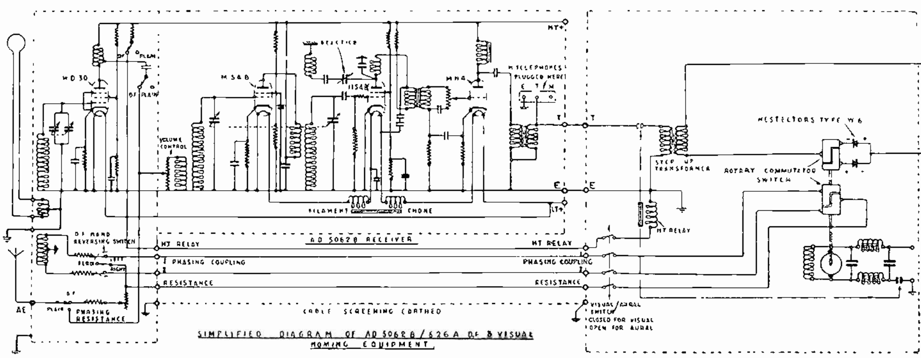


FIG. 3.

The pointer of the microammeter therefore records differences in amplitude between pulses of opposite sign, the system being designed with sufficient lag so that the pointer is incapable of following changes between individual pulses, but merely records the aggregate result.

It will be seen that when on course with the frame orientated at right angles to the line of flight, pulses of equal amplitude but opposite sign will be received and the needle will not depart from its central position. A sufficient deviation to the right or left will bring about a corresponding increase in amplitude in the pulses of one of the trains and a decrease in amplitude of the pulses in the other, thus causing a meter deflection to the right or left as the case might be.

With the apparatus is used a 13 inch diameter screened rotatable frame and a 200 feet trailing aerial—for " Vertical "—for the formation of the cardioid diagram. Fig. 3 gives a schematic diagram of connections.

The Linking Apparatus.

The connection between the direction finder and the automatic pilot consists of a sensitive moving coil relay which is connected in parallel with the Visual indicating meter, and is arranged so that the tongue of the relay follows precisely the movement of the needle of the meter.

On either side of the relay tongue is a contact connected to the winding of one of two subsidiary relays, one of which is therefore actuated when an off-course reading is registered. The subsidiary relays are each provided with contacts, which, when actuated, complete the circuit of the appropriate electro magnet of the automatic pilot control pack.

The sequence of operations is therefore :—

1. The inducing of an E.M.F. in the D.F. loop, depending on its angle of divergence.
2. An off course deflection of the visual indicator accompanied simultaneously by—
3. The marking to the right or left of the master relay.
4. The operation of the corresponding subsidiary relay.
5. The application of current to the corresponding electro magnet of the automatic pilot control pack.
6. The precession of the gyro by the energised magnet.
7. The operation of the corresponding value trigger by the gyro spindle.
8. The admission of oil through the valve to the corresponding servo motor to operate the rudder.

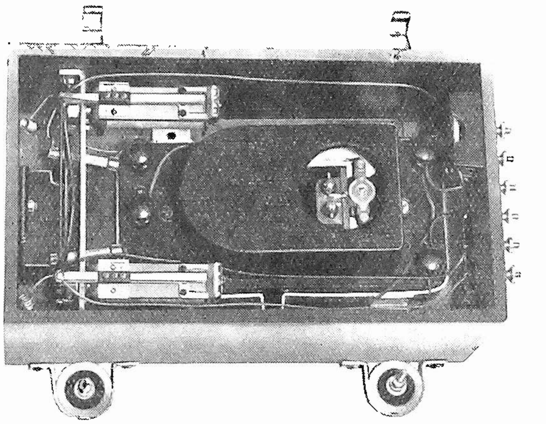


FIG. 4.

Two small supervisory lamps have been connected in parallel with the automatic pilot electro magnets, to show when the control is in operation. A variable resistance has also been fitted to control the amount of current flowing through the energised magnet. The adjustment of this resistance regulates the rate of turn of the aircraft.

Fig. 4 shows an illustration of the linking apparatus.

Results.

Having installed the D.F. apparatus in our small aeroplane and attended to such matters as screening and bonding of the engine and air frame, the machine was taken to the compass bed for the calibration of the D.F.

The error curve when tuned to Droitwich station was plotted and found to be quite reasonable, the maximum error being of the order of 4 degrees. Afterwards one or two preliminary flights were made to watch the general operation of the device and to make adjustments. One or two interesting features soon emerged. For example, such was the response to the control that it was found quite easy to steer the aeroplane merely by rotating the loop hand wheel—if the loop is offset from the zero a few degrees at a time the head of the aeroplane will follow by a like amount—again it was found that the control eliminated all possibility of the 180 degrees ambiguity, for if the loop connections to the receiver are correctly applied for heading towards a station, and if at the moment of application of the control it so happens

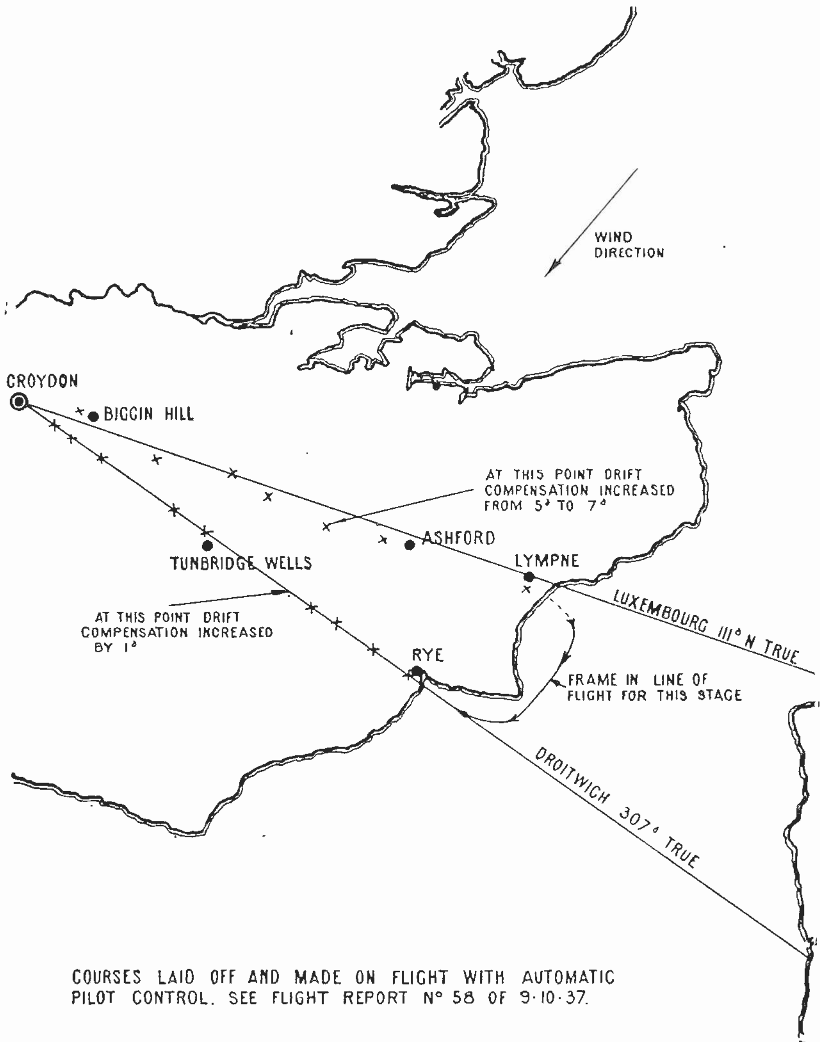


FIG. 5.

that the head of the aeroplane is pointing in the opposite direction, as soon as an off-course divergence is registered the control will take charge and rotate the aircraft gradually through 180 degrees, i.e., until head-on to the station.

At first we found that there was a tendency for the aeroplane to follow a " weaving " course, i.e., to get into slow oscillation. This was found to be due to the momentum of the aircraft on the turn, causing the frame to swing through the minimum sufficiently far on the other side to cause the control to be actuated, thus producing a to-and-fro motion. This defect was readily cured on our experimental flights by including an adjustable resistance, or " rate of turn control," in the gyro magnetic control circuit.

When preliminary adjustments had been made we decided to occupy a few flying hours in ascertaining the accuracy of steering obtainable. This we suspected would be of a fairly high order owing to the practically instantaneous reaction of the rudder to any off-course deviation registered by the direction finder. It will

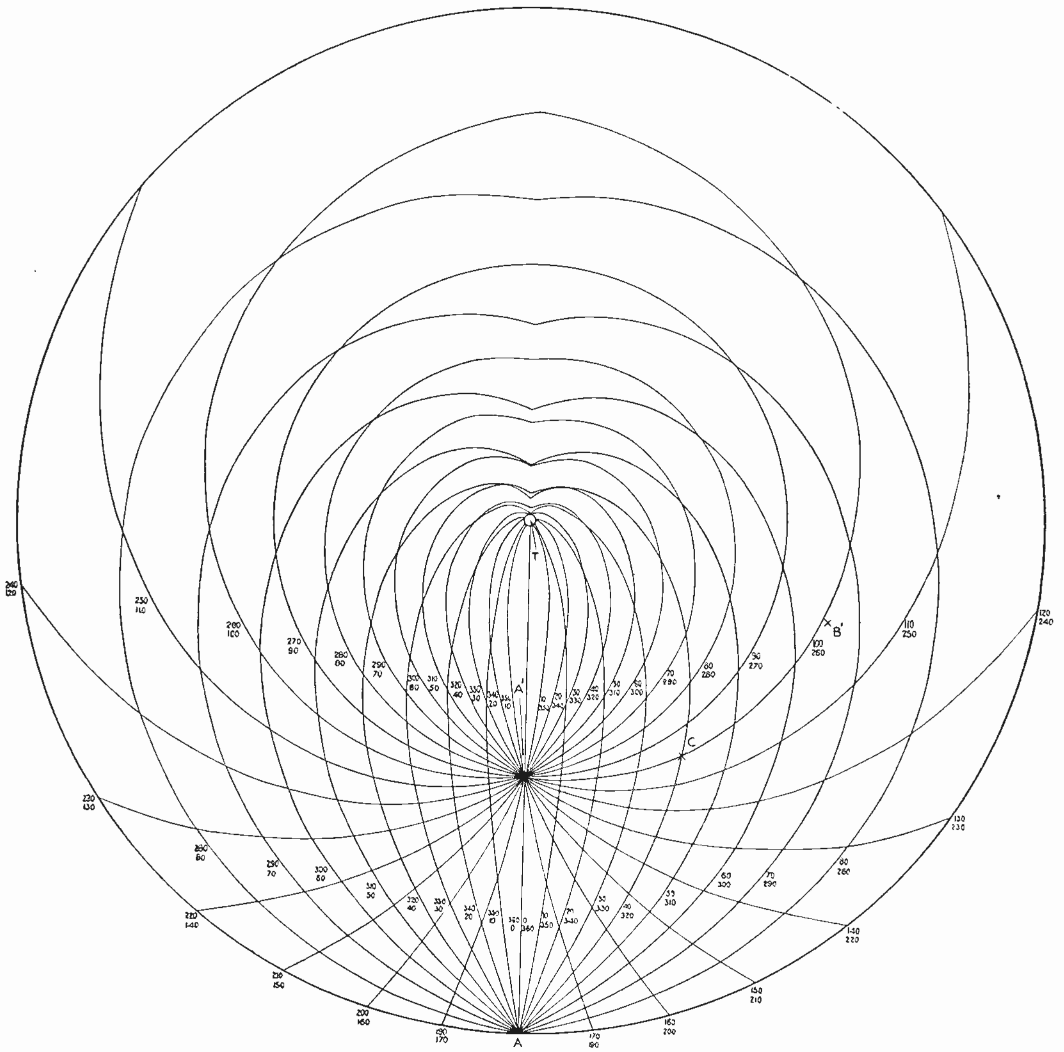


FIG. 6.

be appreciated that human errors and lag due to mental and physical reactions of the pilot are entirely eliminated by the automatic action of the device.

It was found that a remarkably accurate course could be made if drift compensation was accurately assessed. Fig. 5 shows two courses laid off and made, one on Luxembourg and another on Droitwich, the latter being almost exactly correct.

The Orientator.

It will be appreciated that if the loop aerial is orientated so that its plane is at right angles to the line of flight, the course followed will be a straight line (neglecting the effect of " drift "). If the loop be orientated so that its plane is parallel to the fore-and-aft direction of the aircraft, the theoretical course will be a circle having the transmitting station at its centre. For intermediate angles of orientation, intermediate curves between the straight line and the circle would be followed.

For convenience of reckoning a set of curves, which are actually equal angle spirals, has been prepared in the form of the orientator shown in Fig. 6, which is used in conjunction with a map or chart of suitable scale to enable the pilot or navigator easily to determine the correct setting of the loop for any given course from one place to another, neither being necessarily the location of the radio transmitter.

It is intended that the orientator be made up in the form of a circular or semi-circular disc of transparent material marked out with lines as shown. In use the orientator is placed over the appropriate map or chart so that the point T is positioned over the transmitting station which is being employed for guidance, and the orientator is rotated so that point A or A^1 is over, or as nearly as possible over, the point of departure. A selection of scales of maps and orientators is indicated if wide coverage is required.

Suppose it is desired to proceed from point A to point B^1. Then the aircraft is navigated along the 40 degrees 320 degrees curve, with its loop orientated at this angle (ignoring drift setting) until the point C is reached, whereupon the frame is re-orientated to the angle indicated by the curve 100 degrees 260 degrees.

Application.

It would seem that the machine steering achieved will provide means for following a somewhat more rigid course than could otherwise be steered in ordinary daily flying. Several applications have been proposed, e.g. :—

Where traffic requirements make it desirable for incoming and outgoing aircraft to keep to sharply defined lanes.

For aerial photography, where it is necessary to follow prescribed courses with great precision.

For the approach to an aerodrome under conditions of poor visibility.

More experience is necessary, however, before the value of this device as an aid to aerial navigation can be truly assessed, but the means are to hand whereby an aircraft can be set to follow of its own accord a predetermined route, the pilot watching progress, and making small course adjustments from time to time to compensate for varying degrees of drift.

On a recent flight in calm air, 1 degree (on Droitwich from over London) was found to be sufficient movement of the loop orientation to cause the corrective control to be actuated.

Thanks are due to Capt. Bailey, who piloted the machine, and to colleagues in the Research, Engineering and Aircraft Departments of the Marconi Company, for valued advice and assistance rendered during the progress of the experiments.

J. M. FURNIVAL.

FREQUENCY SELECTIVE FEEDBACK APPLIED TO THE DESIGN OF BAND-PASS AMPLIFIERS

It is the purpose of this paper to show that with an intermediate frequency band-pass amplifier employing the normal arrangement of overcoupled transformers an improved response can be obtained by the use of frequency selective feedback. Negative feedback only is considered. The method is particularly well adapted for obtaining variable band width. A method of design is developed from the use of generalised response curves and confirmed by experimental results.

THE almost universal use of overcoupled transformers for intermediate frequency amplifiers is in itself sufficient testimony to their advantages over other systems at present available. But while their advantages are well known their limitations are equally well defined.

It will be recognised therefore that any attempt to improve the response by increasing the efficiency of the circuits results only in the production of pronounced "humps" without contributing greatly to the rate of cut-off. But by using high efficiency coils in conjunction with cathode feedback applied by means of tuned circuits connected in the cathode lead of one of the valves, a level response can be obtained in the pass range together with a substantial increase in the rate of cut-off. Consider this in relation to Figs. 1, 2 and 3.

Fig. 1 represents the ideal for certain purposes. That is, it fulfils the following requirements :—

(i) The amplifier response must be substantially flat in the pass range.

(ii) The response must have fallen off by a given amount at some fixed frequency on either side of the pass range.

(iii) It must preserve this degree of attenuation at all frequencies in the cut-off range.

Fig. 2 shows the kind of response that might be obtained by the use of overcoupled transformers alone, using high efficiency circuits. The desired curve is shown again, dotted.

Fig. 3 shows the amount by which the full line curve of Fig. 2 falls short of the ideal.

It will be seen, therefore, that if the response of Fig. 3 can be produced by the cathode feedback circuits the required ideal response curve is at once obtained, for the overall response will be the addition of the response due to the cathode circuits acting alone, to the response of the overcoupled transformers acting alone, provided there is no mutual interaction.

It will now be shown how closely these conditions can be met using practicable arrangements.

Take the case of a general impedance in the cathode lead of a valve. The feed-back is approximately proportional to the value of the impedance, and is negative,

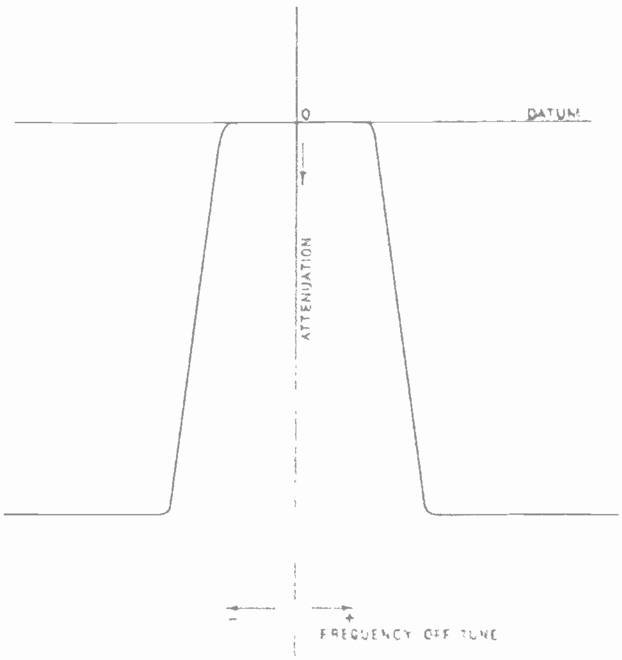


Fig. 3.

that is, as the impedance rises the gain of the valve decreases, and, conversely, as the impedance falls, the gain rises. Therefore, to obtain the type of response shown in Fig. 3, by means of the cathode circuits acting alone, these circuits must have an impedance characteristic which is the inverse of that curve. Such an impedance characteristic is given by several two-terminal networks, but they are not all practicable because of other considerations; moreover, as they can be shown to be exactly equivalent provided the required circuit values can be achieved, one only need be considered. The one that has been chosen comprises two parallel tuned circuits connected in series with one another as in Fig. 4. The circuits are staggered with respect to the centre of the pass band, one being tuned to the frequency f_2 which is above the central frequency f_0 and the other to the frequency f_1 which is below f_0. The frequencies f_2 and f_1 are chosen to be the frequencies at which maximum attenuation is required and are the two frequencies at which the combined cathode circuit has maximum impedance.

In order to design the circuits so that the required conditions of flatness of

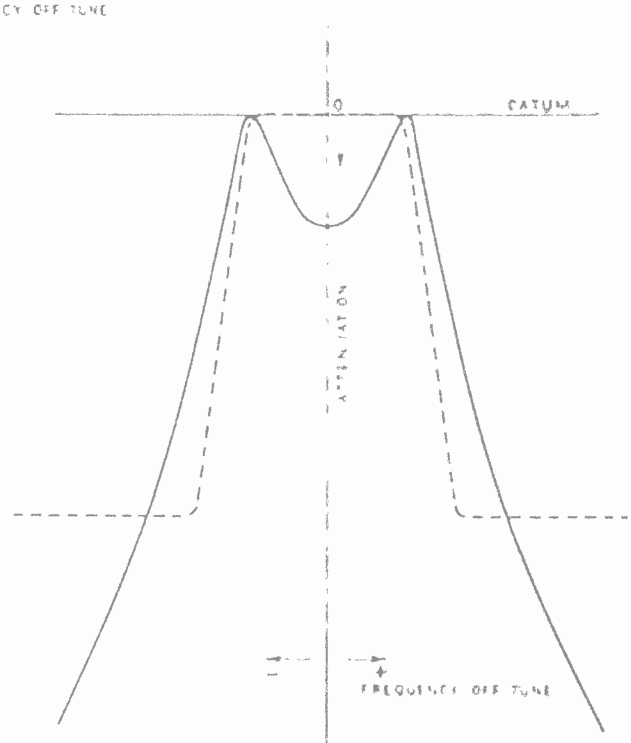


Fig. [illegible]

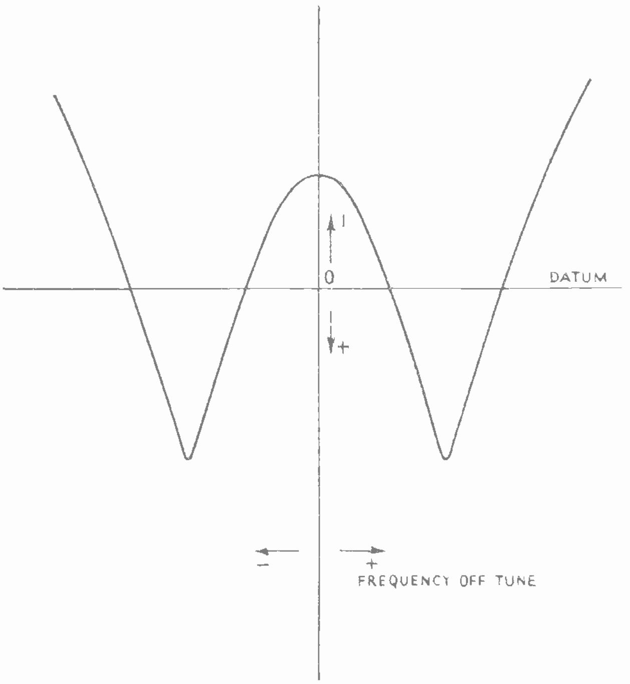


Fig. 3.

overall response in the pass range, and sharpness of cut-off are fulfilled, it is desirable to be able to predict the precise shape of the response produced by the overcoupled transformers and that produced by the cathode circuits separately, and, conversely, from a given shape of response to be able to determine the values of the component elements.

This work would be exceedingly tedious if it were necessary to determine the shape of response for every particular set of conditions and so the design is greatly facilitated by the use of generalised response curves. From a family of these curves the shape of the response of any overcoupled transformer can be determined, and from another such family of curves the shape of the response of any pair of cathode circuits can also be determined.

The application of generalised response curves to the design of band-pass filters was described by R. T. Beatty in an article entitled "Two element filters," W.E., Oct., 1932, and the development of the argument will be briefly repeated here, but in a slightly different form, and be extended to include the cathode circuits.

Consider Fig. 5 (A). This is a symmetrical four terminal network. It can be shown that Fig. 5 (A) is equivalent to Fig. 5 (C) which is really a two terminal network or equivalent shunt impedance Z_L which will replace the network of Fig. 5 (A).* Moreover :

$$\left|\frac{Z_L}{Z_L \text{ max.}}\right| = \frac{2\beta}{\sqrt{4K^2 + (1 - K^2 + \beta^2)^2}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

Where β is a parameter equal to $\frac{\omega M}{R}$

and $K = \frac{2Q \triangle f}{f}$

$Q = \frac{\omega L}{R}$ = the efficiency of each coil

f = mid-band frequency.

$\triangle f$ = frequency off-tune, considered +ve above mid-band frequency and —ve below mid-band frequency.

* The derivation of these equations is given in the Appendix

If the tuned transformer which constitutes the network be connected in the anode circuit of a valve whose internal impedance can be considered as infinite then the gain of the stage will be proportional to the equivalent impedance Z_L.

The ratio $\left|\frac{Z_L}{Z_L \text{ max.}}\right|$ is therefore a measure of the voltage response with respect to the max. voltage gain of a single tuned transformer in the anode circuit of a valve.

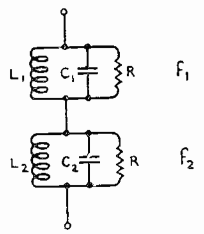


FIG. 4.

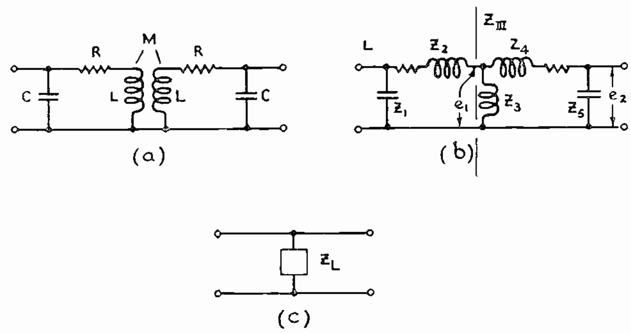


FIG. 5.

By plotting 20 log 10 $\left|\frac{Z_L}{Z_L \text{ max.}}\right|$ against K on a logarithmic scale for different values of β we have a family of curves from which it is possible to obtain a direct reading decibel-frequency relative response curve for any symmetrical pair of coupled circuits simply by a slight manipulation of the horizontal scale.

Since the scale of K is logarithmic and $K = \frac{2Q\triangle f}{fo}$ the operation of dividing K by $\frac{2Q}{fo}$ in order to obtain the corresponding values of $\triangle f$, is performed merely by sliding the scale horizontally through the required distance with respect to the curves and then regarding it as a scale of off-tune frequency.

A convenient way of shifting the scale in practice is to draw the complete graticule on tracing paper and to move it along over the curves until it takes up the required position.

This may be illustrated with reference to Fig. 6 which shows the curves of 20 log 10 $\left|\frac{Z_L}{Z_L \text{ max.}}\right|$ plotted against K for different values of β.

To convert this horizontal scale of K into a scale of kilocycles off-tune it is first necessary to evalute the constant $\frac{2Q}{fo}$ for the particular pair of coupled circuits in question. As an example say the coils had a Q of 100 and they were each tuned to resonate at 450 kilocycles per sec. then $\frac{2Q}{fo} = \frac{2 \times 100}{450} = 0.445$. It is now only necessary to shift the scale horizontally so that the position marked 1 on the new scale occupies the position previously marked 0.445. The new scale then reads kilocycles off-tune directly.

It follows by a similar argument that the decibel scale is purely relative and could equally well be considered from any other datum level than the one shown merely by shifting the tracing paper scale vertically.

In general there will be more than one transformer in the amplifier and to obtain the overall response we must add the individual response curves of the various transformers. If they are all the same it is only necessary to multiply the decibel scale by the number of transformers in the amplifier.

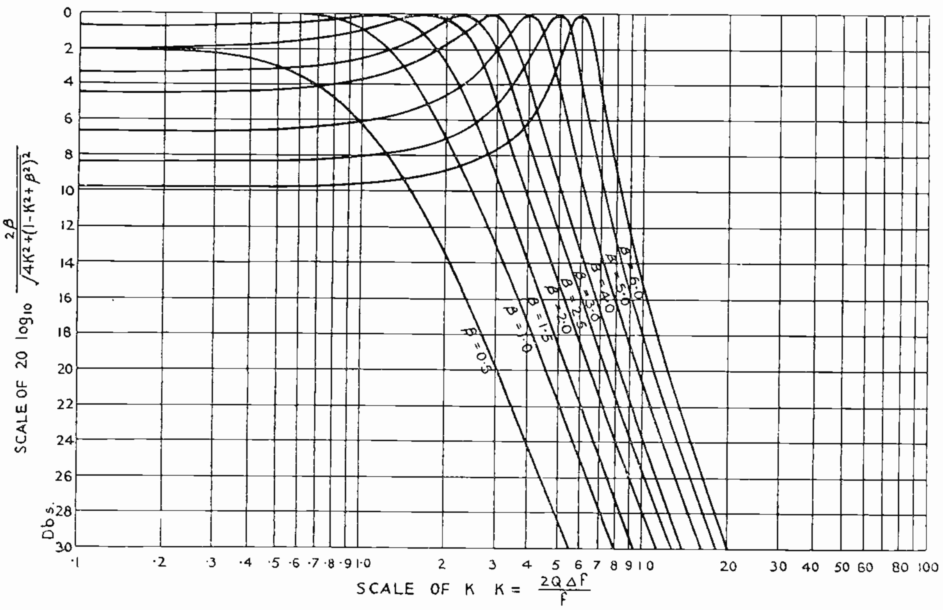


Fig. 6.

The method of using a sheet of tracing paper to carry the movable scale is extremely useful when superimposing different curves. Suppose, for instance, that we have a family of generalised curves of the responses due to the cathode circuits alone and that these curves are traced off on to a separate sheet of tracing paper, we can now place these curves over the curves due to the coupled circuits and move them horizontally or vertically with respect to the latter until we find a pair which when added together give the required *shape* of response. We can now superimpose the frequency scale and move it horizontally until the pass band ends at the required frequency, and read off opposite the 1 kilocycle mark the values of $\frac{2Q}{fo}$ for both the coupled circuits and the cathode circuits, from their respective scales of K.

It remains therefore to put the response produced by a pair of cathode circuits in the form of generalised curves with K as the abscissæ.

Consider the circuit of Fig. 4.

If one circuit be tuned to resonate at a frequency f_1 and the other to a frequency f_2 such that the mid-band frequency fo is given by the expression

$$fo = \sqrt{f_1 f_2}$$

then : $$\frac{f_1 - fo}{fo} = \frac{fo - f_2}{f_2} \doteq \frac{\Delta fo}{fo}$$

The resonant frequencies of the circuits can be said to be " staggered " with respect to one another, and $\Delta\, fo$ is the semi-stagger.

Assume that the ratio $\frac{R}{WL}$ is the same for both circuits, that is that $Q_1 = Q_2 = Q$ say—

Then the impedance of the two circuits in series is given by the expression :—

$$Z_c = \frac{1}{\frac{n}{2R}\left[\frac{1}{1 + K^2} + \frac{1}{n} \pm jK\left(\frac{1}{1 + K^2} - \frac{1}{n}\right)\right]} \quad \ldots \quad (2)$$

" n " is a parameter and is given by the equation $n = \left(\frac{2Q\Delta fo}{fo}\right)^2$

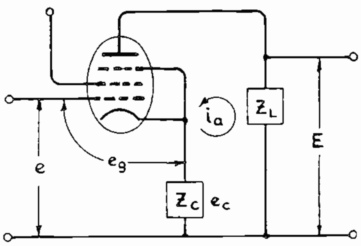


FIG. 7.

R is the impedance of either of the parallel circuits at resonance (the impedance at their respective resonance frequencies are assumed equal).

The response due to the feedback is determined by the effect of the impedance Z_c when connected in the cathode lead of the valve. (See Fig. 7.)

If g is the mutual conductance of the valve without feedback and g_1 is the equivalent mutual conductance of the valve with feedback,

Then $$\frac{g}{g_1} = 1 + g\, Z_c \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

But the relative response due to the cathode impedance is determined by the ratio $\frac{g}{g_1}$ and therefore by the equation (3).

Substituting equation (2) in equation (3) we have

$$\frac{g}{g_1} = 1 + \frac{g}{\frac{n}{2R}\left[\frac{1}{1 + K^2} + \frac{1}{n} \pm jK\left(\frac{1}{1 + K^2} - \frac{1}{n}\right)\right]} \quad \ldots \quad \ldots \quad (4)$$

It is now necessary to assign some value to R in terms of g. This can conveniently be done by fixing the allowable reduction in the effective mutual conductance of the valve plus feedback, at the mid-band frequency.

That is when $K = 0$.

A suitable reduction is 3 db., or $\frac{g}{g_1} = 1.414$.

Therefore

$$1.414 = 1 + \frac{g\, 2\, R}{n + 1}$$

and so

$$2R = \frac{0.414\,(n + 1)}{g} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

Substituting equation (5) in equation (4) we have that

$$\frac{g}{g_1} = 1 + \frac{1}{\frac{n}{0.414\,(n+1)}\left[\frac{1}{1+K^2} + \frac{1}{n} \pm jK\left(\frac{1}{1+K^2} - \frac{1}{n}\right)\right]} \quad \ldots \quad \ldots \quad (6)$$

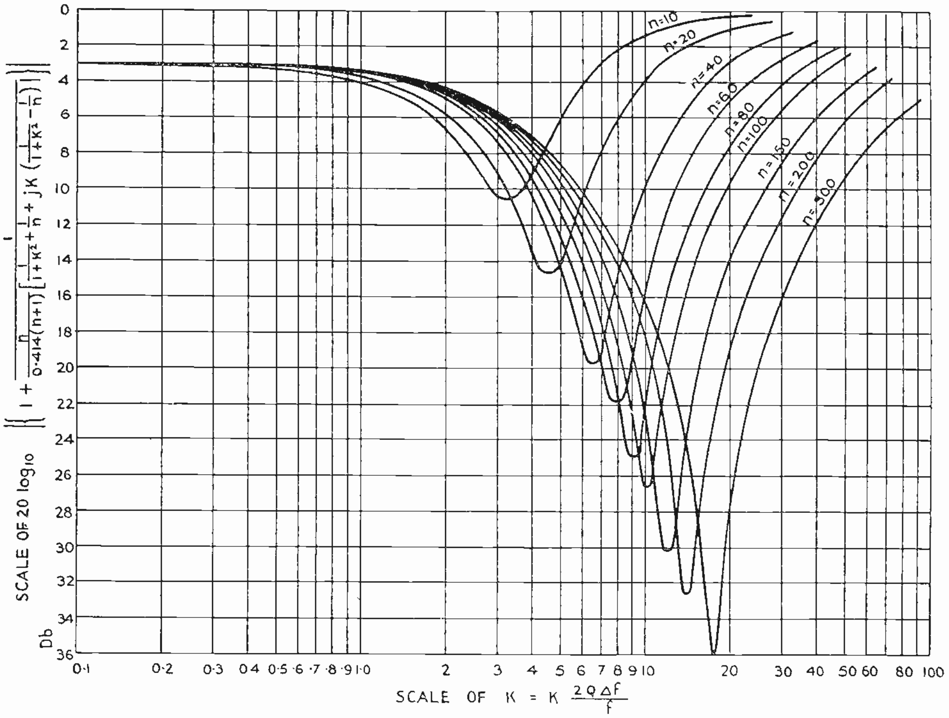


FIG. 8.

Just as equation (1) represented the response with respect to K of a pair of over-coupled circuits in conjunction with a valve, β being a parameter, so equation (6) represents the response due to a pair of cathode circuits with respect to K, n being a parameter. Fig. 8 shows

$$20 \log_{10} \left|\left\{1 + \frac{1}{\frac{n}{0.414\,(n+1)}\left[\frac{1}{1+K^2} + \frac{1}{n} + jK\left(\frac{1}{1+K^2} - \frac{1}{n}\right)\right]}\right\}\right|$$

plotted against K for various values of n.

These are generalised response curves of the cathode circuits in conjunction with the valve and can be used in precisely the same way as the previous generalised curves of Fig. 6.

As before the scale of K is converted into a scale of frequency by shifting it horizontally by an amount which is equivalent to dividing the values by $\frac{2Q}{fo}$. This Q, however, is the Q of the cathode circuits and fo is the mid frequency between the respective resonance frequencies of these two circuits.

In considering the use of the curves of Fig. 6 together with those of Fig. 8, let us choose a particular value of "β" in the one case and "n" in the other; we then have only two curves to deal with. The one represents the response due to the overcoupled transformers or anode circuits acting alone and the other represents the response due to the feedback or cathode circuits acting alone. Also we have seen that both these can be considered as plotted on a scale of off-tune frequency by sliding the scale horizontally with respect to the curve by an amount which is

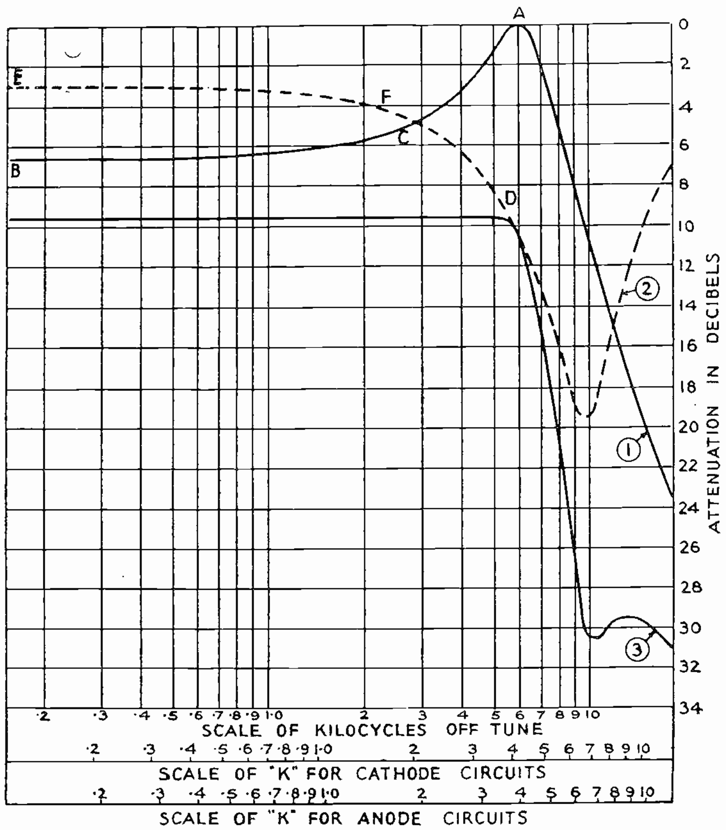


FIG. 9.

determined entirely by the Q value of the respective circuits and the mid-band frequency. This is, of course, equivalent to sliding the curves horizontally with respect to a fixed scale.

Now, if for a given mid-band frequency, we consider the scale fixed and to be a scale of off-tune frequency the effect of moving the curves horizontally is the same therefore as that of changing the Q of the respective circuits. As an example take Fig. 9. Curve (1) represents the curve of response of the anode circuits acting alone. Curve (2) represents the response of the cathode circuits acting alone, plotted on the same scale of frequency off tune. Moving curve (1) horizontally with respect to the frequency scale is equivalent to changing the Q of the anode circuits and moving curve (2) horizontally is equivalent to changing the Q of the cathode circuits. The position of the scale of "K" of each curve with regard to the fixed frequency scale gives the absolute value of the Q's (since $K = \frac{2Q}{f} \Delta f$).

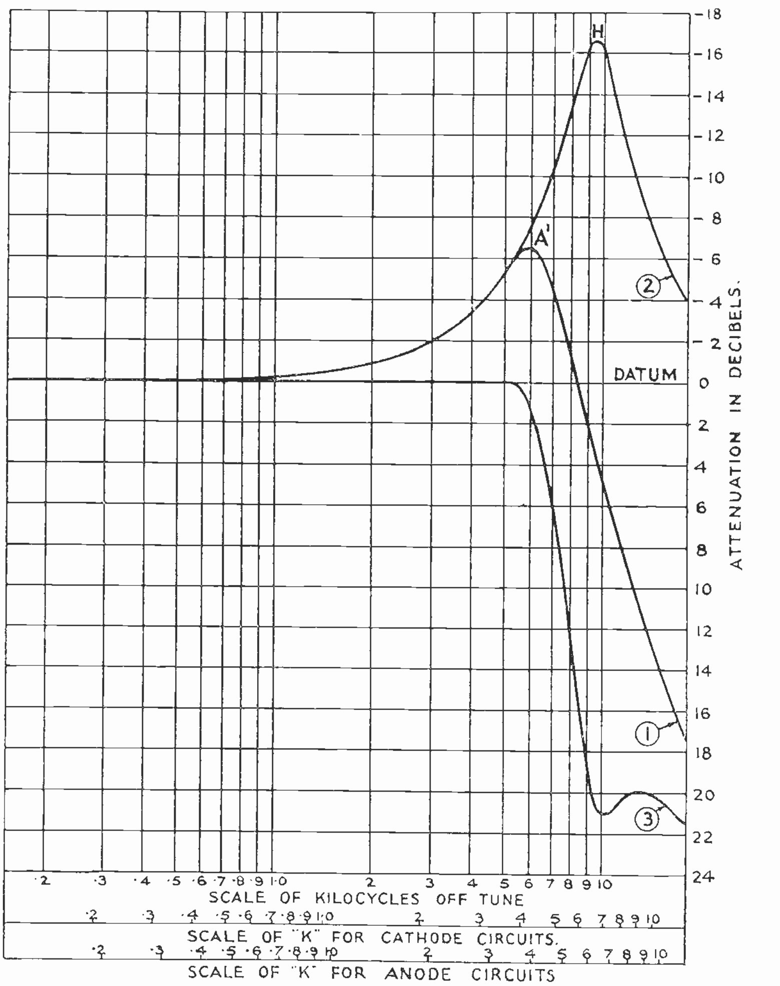


FIG. 10.

Curve (3) is obtained by adding curve (2) to curve (1) and therefore represents the overall response of the amplifier with the anode and cathode circuits acting together. The horizontal position of curve (2) with respect to curve (1), i.e., the Q of the cathode circuits with respect to Q of the anode circuits has been chosen so that the resultant curve has constant attenuation in the pass range.

It will be noted that the frequency of the end of the pass range on curve (3) is practically the same as the frequency of the point "A" on curve (1). And so, for a particular value of β, since the width of the pass band (which is fixed by the specification of the amplifier) fixes the position of the point A, and this fixes the position of curve (1) with respect to the frequency scale, it therefore, by the previous argument, also determines the Q of the anode circuits.

In order that the resultant response curve (3) of Fig. 9 may be flat in the pass region, the shape of the portion B C A of the curve (1) must be the exact inverse of the shape of the portion E F D of the curve (2). That is by inverting the curve (2) and sliding it vertically into the required position it will coincide exactly with the curve (1) over the portion B C A E F D. Moreover, since we are concerned only with the relative response of the complete amplifier with respect to an arbitrary datum line, the overall response curve is given now by the difference between the two curves.

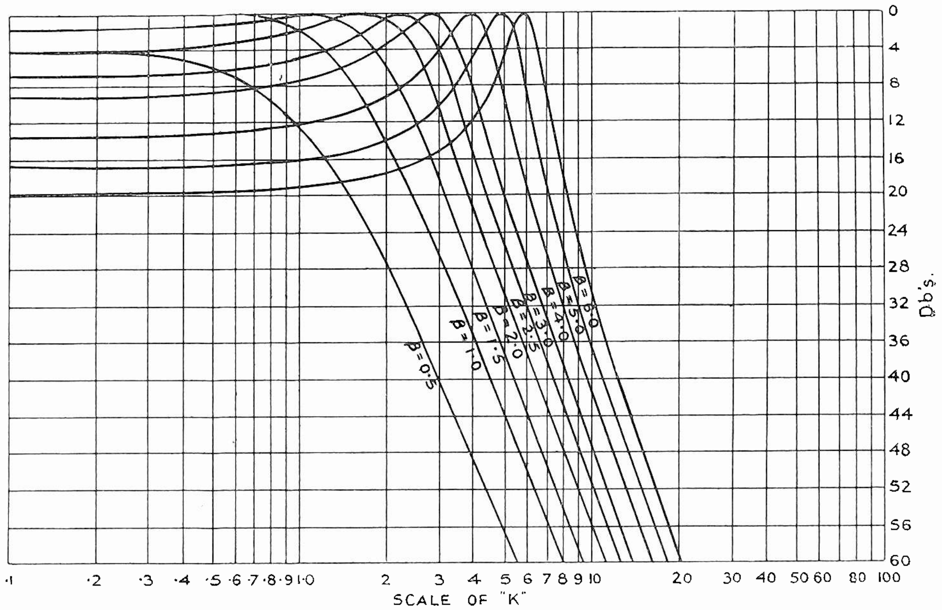


FIG. 11.

This is shown in Fig. 10, where curve (1) is equivalent to curve (1), Fig. 9; curve (2) is the inverse of curve (2), Fig. 9; and curve (3) is the overall response curve considered with respect to the new datum level. This makes the design very simple, for all we have to do is to slide the equivalent of curve (2), Fig. 10, about on the paper, horizontally and vertically, until a reasonable fit is obtained with the curve corresponding to curve (1) of Fig. 10 up to the point A^1. The resultant response curve of the amplifier will then be flat in the region of the pass band.

The next consideration is the position of the point of maximum attenuation due to the cathode circuits. This point, which is the point marked " H " on curve (2) of Fig. 10, might conveniently be termed the dip frequency, and is dependent upon the value of " *n* " : in fact, it defines " *n*." There are, however, practical limits to the value of " *n* " due to the limiting values of Q, which can be obtained in the cathode circuits : so that in practice the highest convenient value of " *n* " would be chosen first, and then the best value of "β" for the anode circuits to give a flat response over the pass range.

Let us consider now a practical example of the design of a complete amplifying stage embodying cathode feedback.

Suppose we require an amplifier which will give uniform response over a band of $\pm$ 6 kilocycles per second and which will present the maximum rate of attenuation between $\pm$ 6 kilocycles off-tune and $\pm$ 10 kilocycles off-tune and not rise to any serious extent at frequencies further off-tune than $\pm$ 10 kilocycles. Suppose also it is a single valve stage and that the mid-band frequency is 460 kilocycles per second.

Associated with a single I.F. valve stage we can have two overcoupled transformers, one in the grid circuit of the valve, and one in the anode circuit of the valve.

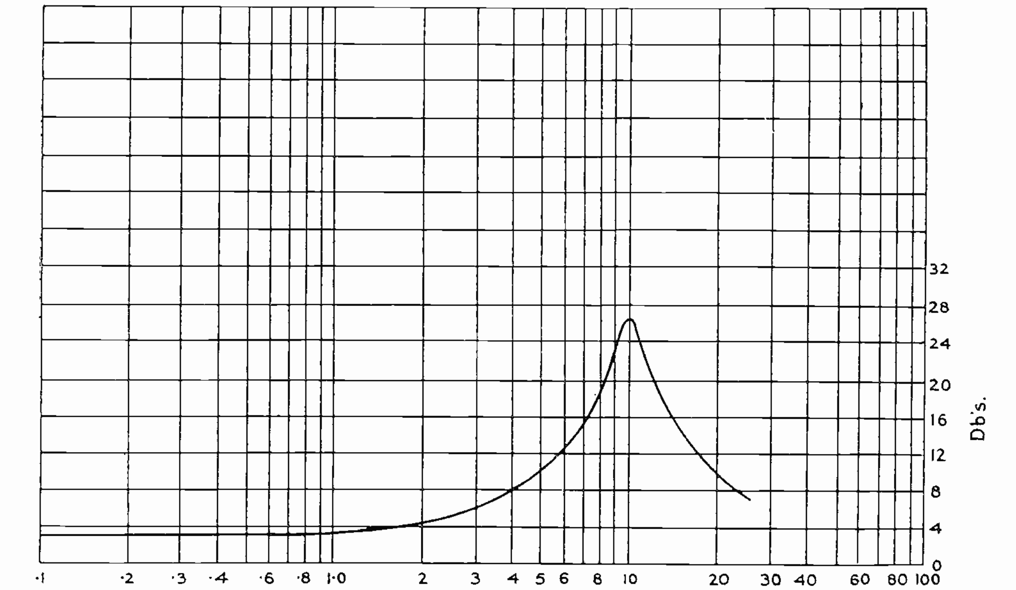


FIG. 12.

Assume that these two are identical. Then Fig. 6 represents the generalised curves for the two transformers in cascade if we simply multiply the ordinate scale by two. This has been done and the curves reproduced in Fig. 11.

Now consider the cathode circuits, generalised curves for which are shown in Fig. 8. The dips are to occur at $\pm$ 10 kilocycles per second. Assume a practical limit to the cathode circuit Q's of 250 (provisionally) we have that $n = \left(\frac{2Q\triangle fo}{fo}\right)^2$ where $\triangle fo$ is the frequency off-tune at which the dip occurs. Therefore $n = \left(\frac{2 \times 250 \times 10}{460}\right)^2 = 118.$

If we actually choose the value of n as 100 instead of 118, we have that Q must be 230 if the dip frequencies remain the same.

The curve of $n = 100$ must now be inverted and superimposed on Fig. 11. But in fixing "n" and fixing the frequency off-tune of the dips we have automatically fixed the position of the frequency scale with respect to the cathode circuit curves from the expression $K = \dfrac{2Q\Delta f}{fo}$ (because by so doing we have fixed Q and fo). This

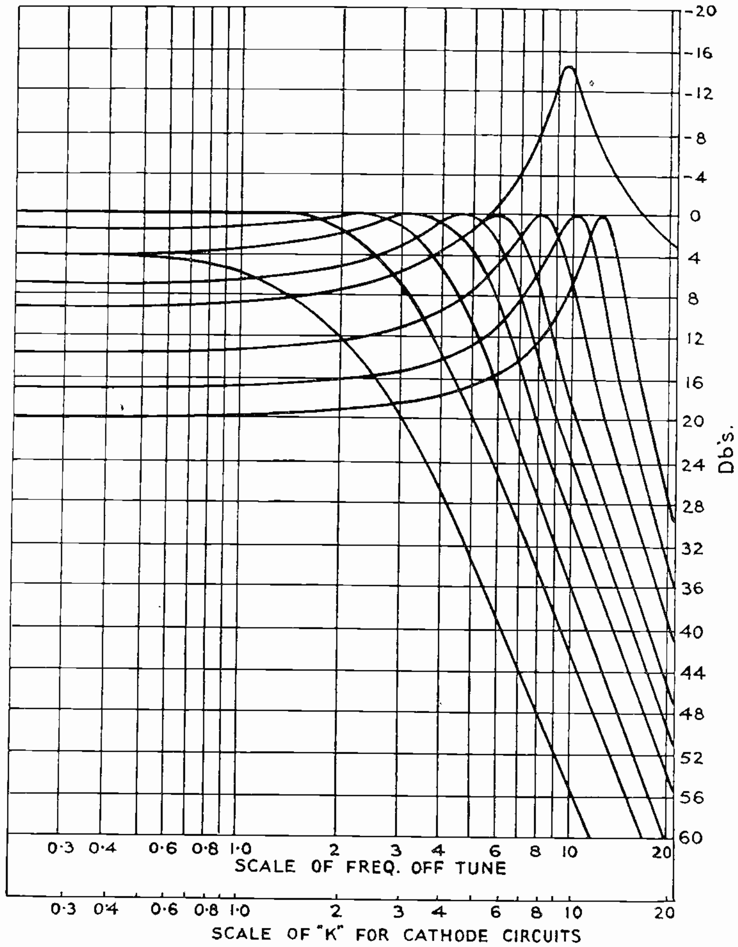


FIG. 13.

curve is therefore shown alone inverted in Fig. 12 on a *frequency* scale, and the *curve and frequency scale* are to be superimposed on Fig. 11 by transferring Fig. 12 to a piece of tracing paper.

We now move this curve and scale about until we have a reasonable fit up to 6 kilocycles off-tune. This will be found to occur for the curve for $\beta = 3$. The position of $f = 1$ is then that previously held by $K = 0.45$ Fig. 13, therefore $\dfrac{Q}{fo} = 0.45$ and so $Q = \dfrac{0.45 \times 460}{2} = 103.5$, say 104, *which is the Q required for the overcoupled transformer circuits.*

In order to complete the design it is now necessary to know the inductance of the overcoupled transformer coils and the mutual conductance of the valve.

In most cases it is desirable to design the amplifier for maximum gain, which means that the transformer coils must present as high an impedance as possible at resonance.

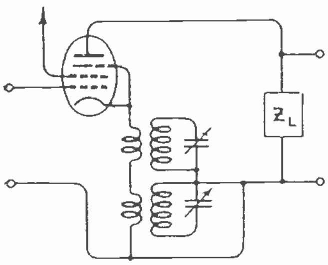


FIG. 14.

The resonant frequency impedance is however given approximately by the expression $Z_o = w_o LQ$. w_o is fixed and Q has been determined, therefore in order to make Z_o high we have to make L high, but the practical limit here is set by the self capacity of the coil, either in that it becomes greater than the capacity required to tune the coil to f_o or because it is a lossy condenser and by forming too great a proportion of the total tuning capacity makes it impossible to achieve the required Q. By careful design, L can be made to be about 700 μh at 460 kilocycles with a Q of 104, when actually coupled in the circuit (by using dust cores and paying attention to dielectric losses). A maximum trimmer capacity of 150 pf is therefore required but can be partly fixed.

With a type V.M.P.4 G valve having a fixed bias resistance of 250 ohms, the value of the mutual conductance has been measured for a number of actual valves and found to be fairly accurately 2 milliamps per volt.

Now from equation (5) above we have that for a 3 db. loss at mid-band frequency due to the cathode circuits $2R = \frac{.414\,(n - 1)}{g}$. Where R is the resonant frequency impedance of each of the cathode circuits

$$\therefore\ R = \frac{.414\ (100 - 1)}{2 \times 2} \times 1000 = 10{,}450 \text{ ohms.}$$

Now, as an approximation we can say that $R = w_o LQ$ where "L" is the inductance required in the cathode circuit

$$\therefore\ L = \frac{R.}{w_o Q} = \frac{10{,}450}{6.28 \times 460 \times 10^3} \times \frac{10^6}{230}\ \mu h$$
$$= 15.7\ \mu h$$

It will be appreciated that to attain a Q of 250 with coils of only 15.7 μh is no mean task, so that some other way of achieving the same end is necessary. Such means is attained by the use of a pair of transformers in the cathode circuit instead of simple circuits. The circuit arrangement is shown in Fig. 14. A mathematical justification for substituting transformers for simple circuits would be rather involved and will not be given. The final test, however, is in their practical application and this has been proved to be entirely satisfactory provided certain precautions are taken. Close coupling is required between primary and secondary windings, so that too high a ratio of turns is undesirable. A practicable value has been found to be about 4 : 1, so that the secondary inductance becomes $4^2 \times 15.7 = 250\ \mu h$, a value at which a Q of over 230 is easily obtained using a dust core and litz wire. Moreover the trimmer condensers required need only now have a maximum capacity of .0005 μf; not an unusual or difficult value to obtain with a mica trimmer.

Variable selectivity or variable band-width can be obtained without upsetting the flatness of response in the pass-range simply by varying the coupling of both the overcoupled transformers simultaneously in the usual way.

Consider Fig. 13 again. If the Q of the overcoupled circuits remains the same and the cathode circuits are unaltered, then the frequency scale remains fixed, and varying the coupling in the overcoupled transformers merely changes the value of "β." By moving the cathode circuit curve vertically through the required amount

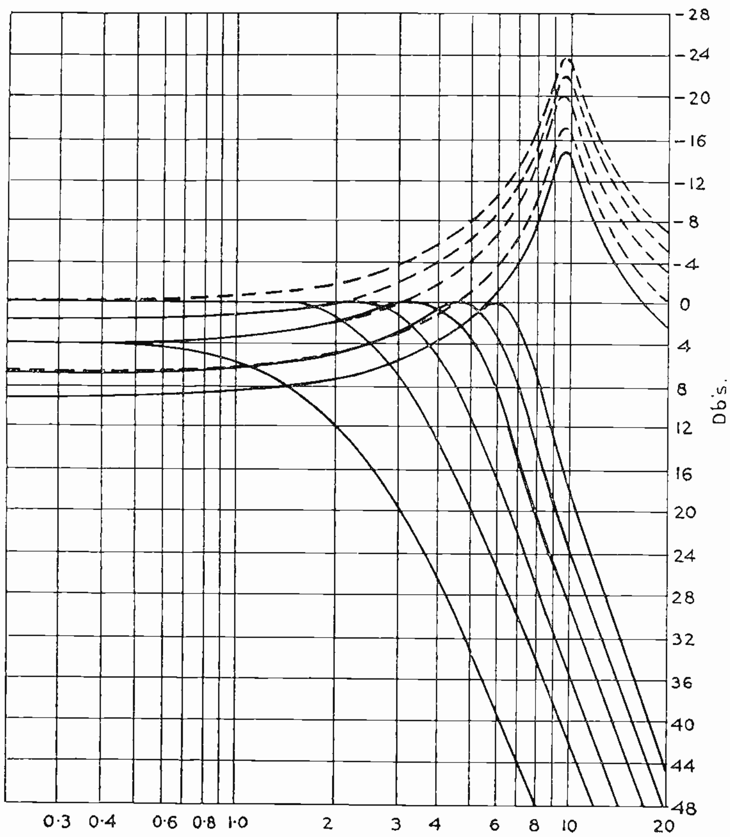


FIG. 15 (A).

it will at once be seen that the curves fit within the pass-band for other values of β less than $\beta=3$ but that the extent of the pass-band is reduced as is shown diagrammatically in Fig. 15 (A). The corresponding overall curves are shown in Fig. 15 (B).

It will also be seen that as the pass-band is narrowed the total attenuation with respect to the pass range at 10 Kc./sec. off tune is rapidly increased, which is a very desirable feature when attempting variable selectivity.

Some undesirable effects are introduced by the use of cathode transformers instead of simple circuits and also by the grid-cathode capacity of the valve. These, however, can be completely overcome by slight modifications to the circuit. The complete circuit arrangement is shown in Fig. 16. The inclusion of a small inductance in series with the two cathode transformers and of a small variable capacity across them ensures complete control of the symmetry of the characteristic, while

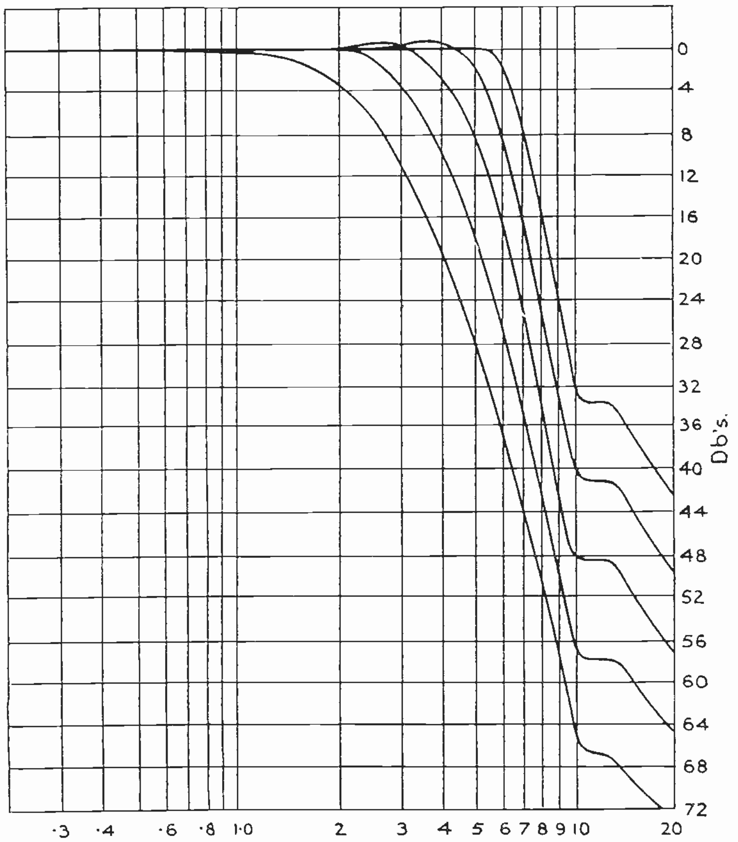


FIG. 15 (B).

the use of a slight step down from the secondary of the input transformer to the grid of the valve completely overcomes the difficulty of the grid cathode capacity with the loss of only a few dbs.

The question of the overall gain of such an amplifying arrangement as that shown in Fig. 16 is naturally an extremely important one, and it might be argued that the use of high Q overcoupled transformers produces only a pronounced double hump and that in order to flatten the overall curve we are deliberately throwing away this extra gain by means of the cathode circuits, without substantially increasing the gain at mid-band frequency. The effect might appear even to be to decrease the gain at mid-band frequency. Fig. 17 has, therefore, been included to show the way in which the mid-band gain, that is, the level of the bottom of the trough between the two humps of an overcoupled transformer, varies with its coil Q for a given fixed band-width. It will be seen that up to Q = 100, for a fixed band-width of

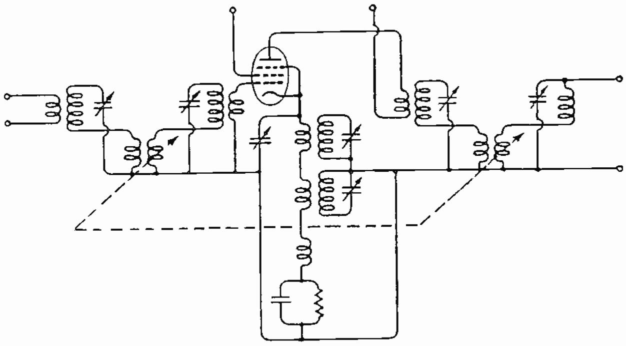


Fig. 16.

± 5 kilocycles and a mid-band frequency of 460 kilocycles per second, there is quite an economic increase in gain.

This means that, for a given band-width, a higher overall gain is obtained, by using high Q coils and cathode feedback than could be obtained using critically

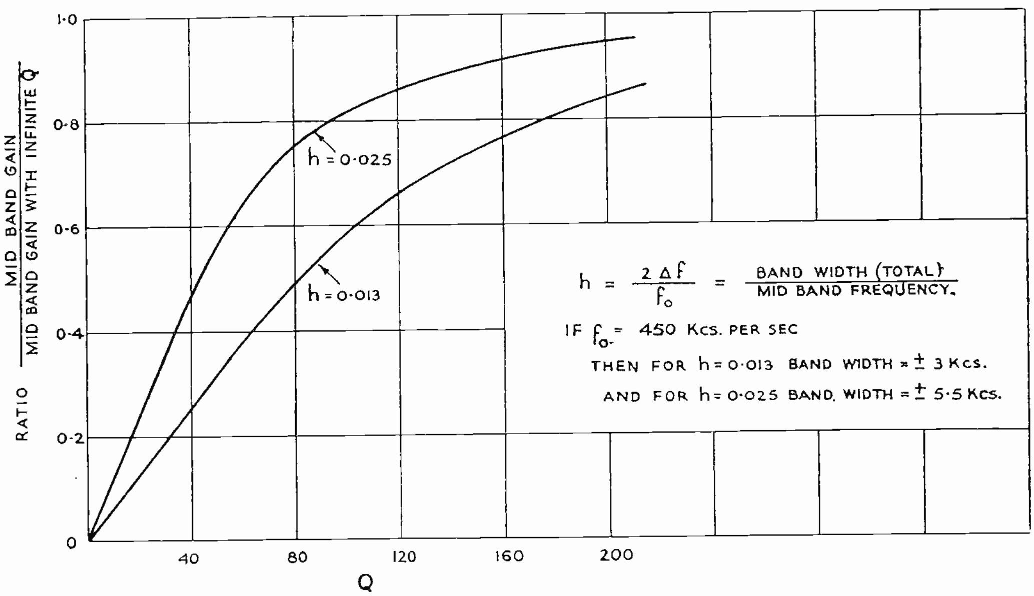


Fig. 17.

coupled circuits without feedback. When this is considered in relationship to the tremendous improvement in cut-off, the ease with which an almost ideal variable band-width characteristic can be obtained, and other advantages, it would seem that the use of cathode feedback circuits is more than justified.

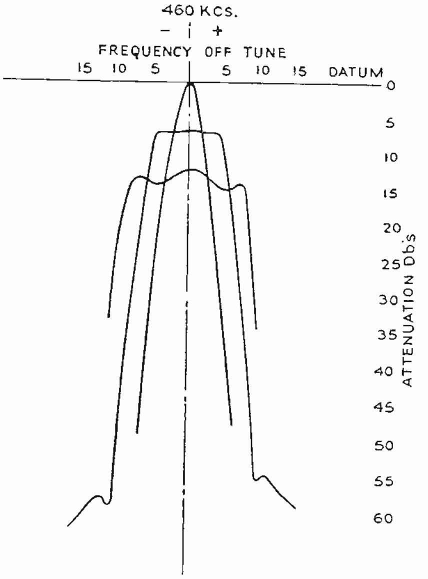


FIG. 18.

Finally, in Fig. 18 are reproduced some measured curves which show the kind of results that can be expected in practice from an amplifier designed along the lines given in the example above.

APPENDIX.

Impedance of a simple series circuit containing inductance (L), Capacity (C) and resistance (R).

If Z be the impedance at any frequency equal to $\frac{\omega}{2\pi}$

$$\text{Then } Z = R + j\omega L - j\frac{1}{\omega C}.$$

Let ω_o be the value of ω at resonance

$$\text{Then } L\omega_o = \frac{1}{C\omega_o}$$

Therefore

$$Z = R + j\omega L - j\frac{\omega_o^2}{\omega} L.$$

Near resonance

$$\omega = \omega_o \pm \triangle\omega$$

$$= \omega_o\left(1 \pm \frac{\triangle\omega}{\omega_o}\right)$$

$$\text{Therefore } Z = R + j\omega_o L\left(1 \pm \frac{\triangle\omega}{\omega_o}\right) - j\frac{\omega_o^2 L}{\omega_o\left(1 \pm \frac{\triangle\omega}{\omega_o}\right)}$$

$$= R + j\omega_o L\left(1 \pm \frac{\triangle\omega}{\omega_o} - \frac{1}{1 \pm \frac{\triangle\omega}{\omega_o}}\right).$$

If $\frac{\triangle\omega}{\omega_o}$ is small compared with 1

$$Z = R + j\omega_o L\left(1 \pm \frac{\triangle\omega}{\omega_o} - 1 \pm \frac{\triangle\omega}{\omega_o}\right)$$

$$= R \pm 2\,j\omega_o L\frac{\triangle\omega}{\omega_o}$$

$$= R\left(1 \pm 2\,j\frac{\omega_o L}{R}\frac{\triangle\omega}{\omega_o}\right).$$

But $\frac{\omega_o L}{R} = Q$ the efficiency of the circuit.

Therefore if we put $K = 2Q\frac{\triangle\omega}{\omega_o}$

$$Z \text{ becomes } = R(1 \pm jK) \quad \ldots \quad (7)$$

Equivalent impedance of two mutually coupled circuits forming a single tuned transformer.

Consider Fig. 5A. This can be shown to be exactly equivalent to Fig. 5B.

Let Z_{111} be the impedance of Z_3 shunted by Z_4 and Z_5 in series.

Referring to Fig. 5B,

$$e_1 = i \frac{Z_1}{Z_1 + Z_2 + Z_{111}} \cdot Z_{111}$$

Therefore

$$\frac{e_2}{i} = \frac{Z_1 Z_{111}}{Z_1 + Z_2 + Z_{111}} \cdot \frac{Z_5}{Z_4 + Z_5} \quad \cdots \quad (8)$$

$$\text{Now } Y_{111} = \frac{1}{Z_3} + \frac{1}{Z_4 + Z_5} = \frac{Z_3 + Z_4 + Z_5}{Z_3 (Z_4 + Z_5)}$$

$$\therefore Z_{111} = \frac{Z_3 (Z_4 + Z_5)}{Z_3 + Z_4 + Z_5} \quad \cdots \quad (9)$$

$$\text{Therefore } Z_1 + Z_2 + Z_{111} = Z_1 + Z_2 + \frac{Z_3 (Z_4 + Z_5)}{Z_3 + Z_4 + Z_5}$$

$$= \frac{(Z_1 + Z_2)(Z_3 + Z_4 + Z_5) + Z_3 (Z_4 + Z_5)}{Z_3 + Z_4 + Z_5}$$

$$= \frac{(Z_1 + Z_2 + Z_3)(Z_3 + Z_4 + Z_5) - Z_3^2}{Z_3 + Z_4 + Z_5} \quad \cdots \quad (10)$$

Substituting (9) and (10) in equation (8) we have

$$\frac{e_2}{i} = \frac{Z_1 Z_3 Z_5}{(Z_1 + Z_2 + Z_3)(Z_3 + Z_4 + Z_5) - Z_3^2} \quad \cdots \quad (11)$$

If $Z_1 = Z_5$ and $Z_2 = Z_4$ then

$$\frac{e_2}{i} = Z_1^2 \frac{Z_3}{(Z_1 + Z_2 + Z_3)^2 - Z_3^2} \quad \cdots \quad (12)$$

Now $Z_1 = -j \frac{1}{\omega C}$; $Z_2 = R + j\omega (L - M)$; $Z_3 = j\omega M$.

Therefore as shown for equation (7)

$$Z_1 + Z_2 + Z_3 = R (1 \pm jK).$$

Substituting these values in (12)

$$\frac{e_2}{i} = Z_1^2 \frac{j\,\omega M}{R^2 (1 \pm jK)^2 + \omega^2 M^2}$$

$$= \frac{Z_1^2}{R} \cdot j \frac{\omega M}{R} \frac{1}{(1 \pm jK)^2 + \frac{\omega^2 M^2}{R^2}}$$

Putting $\beta = \frac{\omega M}{R}$

$$\frac{e_2}{i} = Z_1^2 \cdot \frac{\beta}{R} \cdot \frac{1}{\pm 2K - j (1 - K^2 + \beta^2)}$$

If we are not concerned with the phase relationship between e_2 and i we can put this in the form

$$\left|\frac{e_2}{i}\right| = Z_1{}^2 \frac{\beta}{R} \cdot \frac{1}{\sqrt{4K^2 + (1 - K^2 + \beta^2)^2}}$$

Upon differentiation we find that the maximum value of $\left|\frac{e_2}{i}\right|$ occurs when $K = \pm\sqrt{\beta^2 - 1}$ or $1 \pm \sqrt{2 - \beta^2}$

$$\therefore \left|\frac{e_2}{i}\right| \max = \frac{Z_1{}^2}{2R}$$

$$\therefore \left|\frac{e_2}{i}\right| \Big/ \left|\frac{e_2}{i}\right| \max. = \frac{2\beta}{\sqrt{4K^2 + (1 - K^2 + \beta^2)^2}}$$

But $\left|\frac{e_2}{i}\right| = \left|Z_L\right|$ in Fig. 5 (c)

$$\therefore \left|\frac{Z_L}{Z_L \max.}\right| = \frac{2\beta}{\sqrt{4K^2 + (1 - K^2 + \beta^2)^2}} \quad \cdots \quad \cdots \quad (13)$$

Impedance of two parallel tuned circuits connected in series with one another.

Consider Fig. 4 :

Let the resonant frequency of $L_1 C_1 R$ be f_1 and the resonant frequency of $L_2 C_2 R$ be f_2

Then $\dfrac{f_1 - f_o}{f_o} = \dfrac{f_o - f_2}{f_2} \doteq \dfrac{\triangle f_o}{f_o}$ say.

f_o is therefore the geometric mean between f_1 and f_2 and the resonant frequencies of the two circuits can be said to be "staggered" with respect to one another; $\triangle f$ is the "semi-stagger."

Assume $Q_1 = Q_2 = Q$

and $$2Q\frac{\triangle f_o}{f_o} = K_o$$

Then we can write with sufficient accuracy that

$$Y_1 = \frac{1}{R}\Big[1 + j(K_o \mp K)\Big]$$

$$Y_2 = \frac{1}{R}\Big[1 - j(K_o \pm K)\Big]$$

$$\therefore Z_1 = R \cdot \frac{1 - j(K_o \pm K)}{1 + (K_o \pm K)^2}$$

$$Z_2 = R\,\frac{1 + j(K_o \mp K)}{1 + (K_o \mp K)^2}$$

The total series impedance is $Z_1 + Z_2 = Z$ say.

Therefore :

$$Z = R\left[\frac{1}{1+(K_o \pm K)^2} + \frac{1}{1+(K_o \mp K)^2} + j\left\{\frac{(K_o \mp K)}{1+(K_o \mp K)^2} - \frac{(K_o \pm K)}{1+(K_o \pm K)^2}\right\}\right]$$

$$= R\left[\frac{1}{A} + \frac{1}{B} + j\left(\frac{C}{B} - \frac{D}{A}\right)\right] \text{ say} \quad \ldots \quad (14)$$

Where
$$A = 1 + (K_o \pm K)^2 = (1 + K^2 + K_o^2) \pm 2KK_o$$
$$B = 1 + (K_o \mp K)^2 = (1 + K^2 + K_o^2) \mp 2KK_o$$
$$C = (K_o \mp K) \text{ and } D = (K_o \pm K)$$

Therefore
$$AB = (1 + K^2 + K_o^2)^2 - 4\,K^2K_o^2 \quad \ldots \quad (15)$$

and
$$A + B = 2\,(1 + K^2 + K_o^2) \quad \ldots \quad (16)$$

$$(AC - BD) = (K_o \mp K)\,\{1 + (K_o \pm K)^2\} - (K_o \pm K)\,\{1 + (K_o \mp K)^2\}$$
$$= K_o \mp K + (K_o \mp K)(K_o \pm K)^2 - K_o \mp K - (K_o \pm K)(K_o \mp K)^2$$
$$= \mp 2K + (K_o \mp K)(K_o \pm K)(K_o \pm K - K_o \pm K)$$
$$\therefore \quad (AC - BD) = \mp 2K\,(1 - K_o^2 + K^2) \quad \ldots \quad (17)$$

Substituting (15) (16) and (17) in (14) we have

$$Z = \frac{R}{(1+K^2+K_o^2)^2 - 4K^2K_o^2}\left[2\,(1+K^2+K_o^2) + j\left\{\mp 2K\,(1+K^2-K_o^2)\right\}\right]$$

$$= \frac{2R}{(1+K^2+K_o^2)^2 - 4K^2K_o^2}\left[1 + K^2 + K_o^2 \mp jK\,(1 + K^2 - K_o^2)\right]$$

$$\therefore\ Y = \frac{(1+K^2+K_o^2)^2 - 4K^2K_o^2}{2R}\left[\frac{1 + K^2 + K_o^2 \pm jK\,(1 + K^2 - K_o^2)}{(1 + K^2 + K_o^2)^2 + K^2\,(1 + K^2 - K_o^2)^2}\right] \quad \ldots \quad (18)$$

But rearranging the denominator

$$(1 + K^2 + K_o^2)^2 + K^2\,(1 + K^2 - K_o^2)^2$$
$$= (1 + K^2 + K_o^2)^2 + K^2\,(1 + K^2 + K_o^2 - 2K_o^2)^2$$
$$= (1+K^2+K_o^2)^2 + K^2\,(1+K^2+K_o^2)^2 - 4K^2K_o^2\,(1+K^2+K_o^2 - K_o^2)$$
$$= 1+K^2\,[(1+K^2+K_o^2)^2 - 4K_o^2K^2] \quad \ldots \quad (19)$$

Substituting this in (18) we have

$$Y = \frac{(1+K^2+K_o^2)^2 - 4K^2K_o^2}{2R}\left[\frac{1+K^2+K_o^2 \pm jK\,(1+K^2-K_o^2)}{(1+K^2)\,\{(1+K^2+K_o^2)^2 - 4K_o^2K^2\}}\right]$$

$$\therefore\ Y = \frac{1}{2R}\left[\frac{1+K^2+K_o^2 \pm jK\,(1+K^2-K_o^2)}{1+K^2}\right]$$

$$= \frac{1}{2R}\left[1 + \frac{K_o^2}{1 + K^2} \pm jK\left(1 - \frac{K_o^2}{1 + K^2}\right)\right]$$

Putting $K_o^2 = n$ and rearranging

$$Y = \frac{n}{2R}\left[\frac{1}{1 + K^2} + \frac{1}{n} \mp jK\left(\frac{1}{1 + K^2} - \frac{1}{n}\right)\right] \quad \ldots \quad (20)$$

or
$$Z = \frac{1}{\frac{n}{2R}\left[\frac{1}{1+K^2} + \frac{1}{n} \mp jK\left(\frac{1}{1+K^2} - \frac{1}{n}\right)\right]} \qquad \ldots \quad \ldots \quad (21)$$

Effect of a General Cathode Impedance Z_c on the Effective Mutual Conductance of a Valve.

Consider Fig. 7. The voltage gain of the valve is $\frac{E}{e}$. But the voltage gain of the valve alone is equal to $\frac{E}{e_g}$, therefore the ratio of the gain of the valve acting alone to gain of the valve with feedback is $\frac{e_g}{e}$.

Now if the impedance of the valve is high compared with Z_L and Z_c we can write

$$i_a = ge_g = g_1 e$$

Where g is the mutual conductance of the valve itself and g_1 is the equivalent mutual conductance of the valve with cathode feedback.

Therefore $\qquad \frac{g_1}{g} = \frac{e_g}{e}$

But $\qquad e_g = e + e_c$ and $e_c = - i_a Z_c = - ge_g Z_c = - g Z_c (e + e_c)$

$$\therefore \quad e_c = - \frac{gZ_c e}{1 + gZ_c} \qquad \ldots \quad \ldots \quad (22)$$

Now
$$\frac{e_g}{e} = \frac{e + e_c}{e} = 1 + \frac{e_c}{e} \qquad \ldots \quad \ldots \quad (23)$$

Substituting (22) in (23) we have

$$\frac{e_g}{e} = 1 - \frac{gZ_c}{1 + gZ_c} = \frac{1}{1 + gZ_c}$$

Therefore
$$\frac{g_1}{g} = \frac{1}{1 + gZ_c} \qquad \ldots \quad \ldots \quad (24)$$

J. D. BRAILSFORD.

COMPARISON OF PARALLEL AND SERIES COUPLING CIRCUITS FOR TRANSMITTERS

When using inductive coupling on transmitters the load circuit will usually be series tuned, for a low resistance load (<100 ohms) and parallel tuned for a high resistance load (>500 ohms). For intermediate values each case must be judged on its merits. The behaviour of both types of circuit under varying conditions of loading and frequency is summarised in tables. Amongst other points it is shown that, if X_L, X_C are the reactances of coupling coil, tuning condenser and R the load resistance—

(1) $X_L = X_C$ *gives the maximum loading for either circuit. For the series circuit this setting gives zero reactance, but not for the parallel circuit, and the discrepancy increases with the ratio of* X_L *to* R.

(2) *The series circuit will be used when it is convenient to make* $X_L >> R$, *and the parallel circuit when convenient to make* $X_L << R$.

(3) *When* $X_L = X_C = R$ *the loading obtained by series or parallel circuit will be approximately the same. To increase the coupling beyond this point* X_L *must be increased for the series circuit and decreased for the parallel circuit.*

(4) *To cover a wave range, the series circuit tends to work with a constant value of capacity working at a variable voltage, while the parallel circuit requires a variable condenser working at constant voltage.*

WHEN using inductive coupling on transmitters or elsewhere the two simplest circuits are those shown in Fig. 1.

Behaviour of Series Circuit.

Fig. 1 (a) shows the *series* circuit with load R_1 coupling coil L_1, in which the EMF is induced as indicated by the alternator. C_1 is the tuning condenser. Such a circuit will usually be used when R_1 is small, giving a low damped circuit. This allows full loading to be easily obtained. Values of R_1 up to 100 Ω can be easily handled in this way. The characteristics of such a circuit are simple.

The resonance point of the circuit is found by varying C_1 until the current in the load R_1 is a maximum. This will always occur when $L_1\omega = \frac{1}{C_1\omega}$, so that the reactance of the circuit is zero. The value of current will be $\frac{E}{R_1}$. Theoretically for any value of L_1 it will be possible to find a value of C_1 to fulfil this condition.

Behaviour of Parallel Circuit.

Fig. 1 (b) shows the parallel circuit with load R_2 and tuning condenser C_2 in parallel with it. L_2 is the coupling coil with internal EMF indicated by an alternator.

This circuit will normally be used for high values of load resistance, giving a low damped circuit, allowing full loading to be easily obtained. Values of R_2 down to 500 Ω can be handled in this way. The characteristics of this circuit are simple when $R_2 >> \frac{1}{C_2\omega}$, but become more complicated at the lower values of R_2. When

$R_2 >> \frac{1}{C_2 w}$ the resonance point will occur when $L_2 w = \frac{1}{C_2 w}$ and the reactance of the circuit will be nearly zero.

At low values of R_2 if the circuit is tuned by varying condenser C_2 to give the maximum value of current in the load, this setting will not give zero reactance.

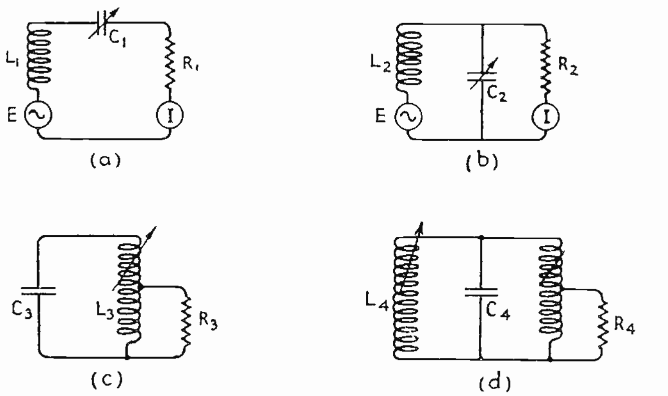


FIG. 1.

Further if L_2 is increased beyond a certain value no real tuning point will be obtained though the current will pass through a maximum value.

The behaviour of the circuit is best studied in connection with Figs. 2 and 3. Fig. 2 (a) shows the actual circuit with

R_P = load resistance.
X_P = capacity reactance of parallel condenser C.
X_L = inductive reactance of coupling coil L.

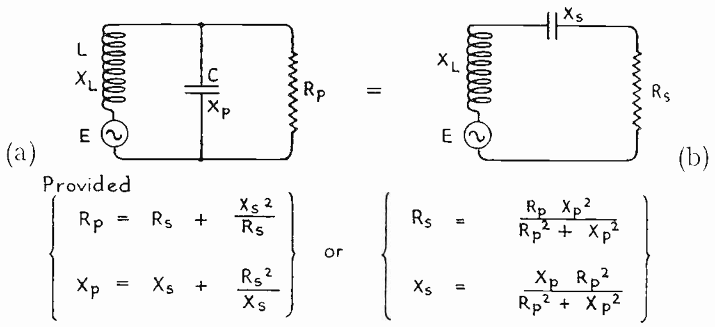


FIG. 2.

Fig. 2 (b) shows the circuit with R_P and X_P replaced by their series equivalents R_S and X_S. These are connected with R_P and X_P as shown below.

$$R_S = \frac{R_P X_P^2}{R_P^2 + X_P^2}, \qquad X_S = \frac{X_P R_P^2}{R_P^2 + X_P^2} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

or the alternative form

$$R_P = \frac{R_S^2 + X_S^2}{R_S}, \qquad X_P = \frac{R_S^2 + X_S^2}{X_S} \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

When $R_P >> X_P$ we get from (1)

$$R_S = \frac{X_P^2}{R_P} \text{ and } X_S = X_P$$

Resonance occurs when $X_L = X_S = X_P$ and the effective series resistance $R_S << X_P$.

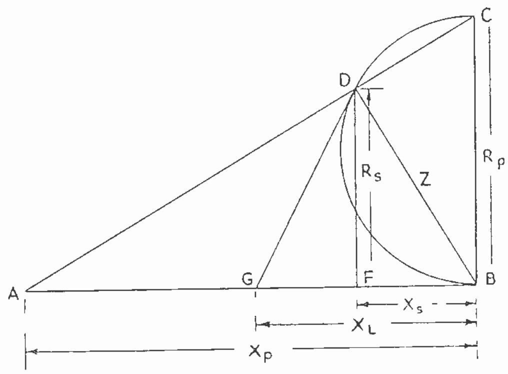


FIG. 3.

The conversions shown in (1) and (2) are the same as those employed for the reactance transformer ; the form of the circuit being identical. The curves there derived can be used to study this case, but a more useful method is as follows. In Fig. 3, AB is drawn equal to X_P, and BC is perpendicular to AB equal to R_P. AC is joined, and BD is drawn perpendicular to AC, and DF perpendicular to AB. Then it can be shown geometrically that

$$FD = R_S \text{ on the same scale.}$$
$$FB = X_S \quad ,, \quad ,,$$
$$BD = \sqrt{R_S^2 + X_S^2} = Z$$

where Z = impedance of R_S and X_S in series or R_P and X_P in parallel.

A semi-circle described on BC will pass through D. Hence when X_P (the reactance of the tuning condenser C_2) is varied the locus of D will be this semi-circle described on CB.

In Fig. 3 mark off BG along BA equal to X_L and join GD.

Then $GF = GB - FB = X_L - X_S$ = resultant reactance of circuit.

Hence $\overline{GD} = \overline{GF} + \overline{FD}$ = resultant impedance of whole circuit.

A common problem will be that R_P is given and suitable values of X_L and X_P have to be found. This can be done by taking different values of X_L, and varying X_P to find the resonance point. This may be defined in various ways.

(1) *Find value of* X_P *to give circuit zero reactance,* i.e., $X_S = X_L$.

This gives the construction shown in Fig. 4 (A) and (B). BF is marked off equal to X_L and a perpendicular drawn through F to cut the semi-circle on BC.

(A) $X_L = \frac{1}{2} R_P$.

The perpendicular through F will touch the semi-circle at D. Join GD and produce to cut BF in A.

Then $AB = X_P = R_P = 2\, X_L$

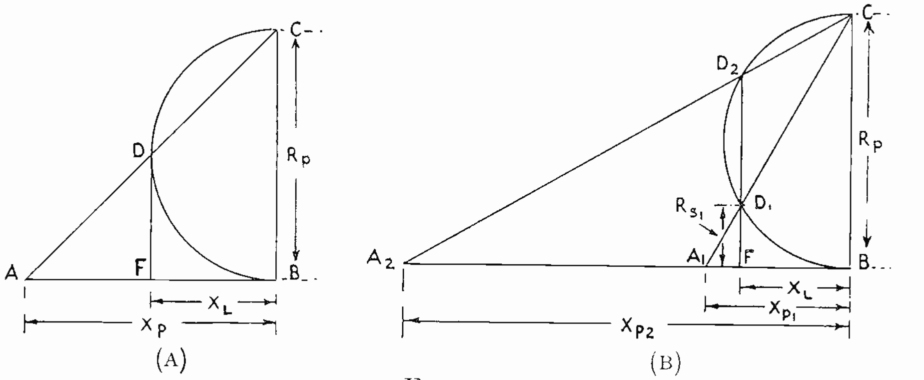
and $FD = R_S = \frac{1}{2} R_P$.

(B) $X_L < \frac{1}{2} R_P$.

The perpendicular through F will cut the semi-circle at two points D_1 and D_2. Join C to D_1 and D_2 and produce to cut BF in A_1 and A_2. Then with X_P equal to either A_1B or A_2B the circuit will have zero reactance, but the effective series resistance will be FD_1 and FD_2 respectively. For coupling purposes the lower value of R_S, i.e., FD_1 will be the only one of use.

(C) $X_L > \frac{1}{2} R_P$.

The perpendicular through F will not cut the semi-circle and the condition of zero reactance cannot be obtained with any value of X_P.

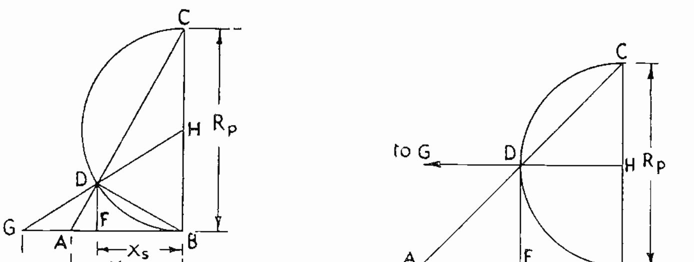


(A) (B)

FIG. 4.

(2) *The Value of* X_P *to give maximum current in the coil* X_L.

Referring to Fig. 3, GD is the resultant impedance of the whole circuit. Hence the current in the coil will be a maximum when GD is a minimum. This gives the construction shown in Fig. 5 (A) in which BG is marked off equal to X_L and joined



(A) (B)

FIG. 5.

to H, the mid-point of BC, GH cuts the semi-circle in D, and GD is a minimum. Join C to D and produce to cut BG in A. Then AB is the required value of X_P.

Note X_P is always less than X_L.

If X_L is very large GH tends to become parallel to AB as shown in Fig. 5 (B) and X_P tends to equality with R_P.

(3) *The value of* X_P *to give maximum current in the load* R_P. This is the most usual requirement.

Referring to Figs. 2 and 3 we find :

Resultant impedance of circuit = GD.

" " X_P and R_P in parallel = BD

$\therefore$ Current in coil $= \frac{E}{GD}$

Volts across $R_P = E \cdot \frac{BD}{GD}$.

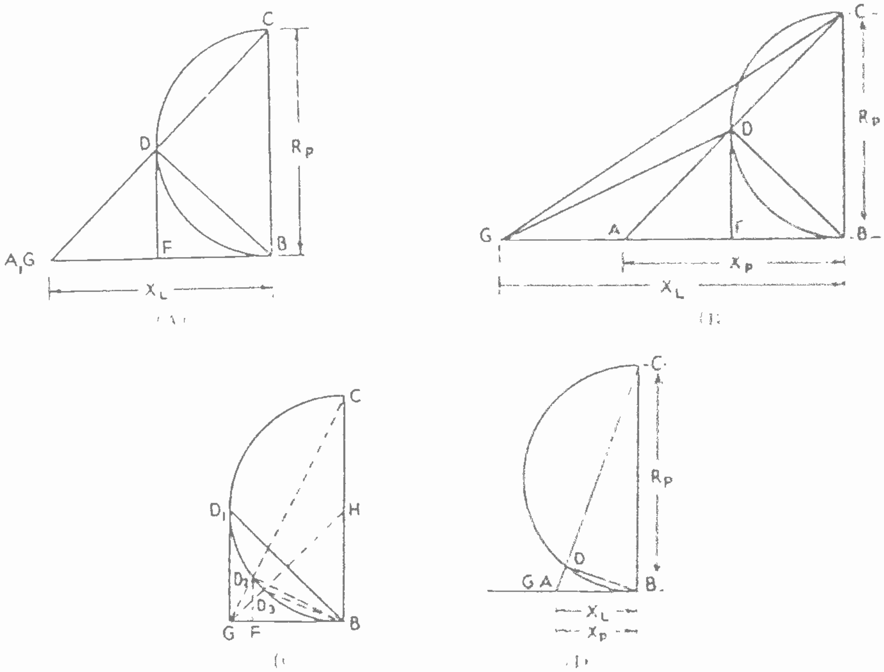


Fig. 6.

This will be a maximum, and the current in R_P will be a maximum when $\frac{BD}{GD}$ is a maximum.

The current in $R_P = \frac{E}{R_P} \cdot \frac{BD}{GD}$.

The exact solution is not easy to obtain graphically in the general case, but an approximation can be obtained by inspection and trial and error.

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

Make BG = X_L = R_P and BA = X_P = R_P so that A coincides with G. Join AC, cutting the semi-circle in D. Join DB.

Then in this position BD = GD and for any other position of D (fixed by value of X_P) BD<GD. The value of X_P (= R_P = X_L) chosen makes $\frac{BD}{GD}$ a maximum,

and is therefore correct. These are the values of condenser and inductance which in a series circuit would give resonance (maximum current and zero reactance), but the phase is 45 degrees in the parallel circuit.

(B) $X_L > R_P$. *Fig.* 6 (B).

Try $X_P = R_P$. Then $GD > DB$ and the initial movement of D as X_P is varied is along FD ($\perp^2$ to AB). From this it follows that the ratio $\frac{DB}{GD}$ is increased by moving D upwards towards C, and the displacement required to reach a maximum value will increase as X_L is increased.

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

Trying three different values of X_P

(i) $X_P = R_P$ point D_1, $\frac{BD_1}{GD_1} = \underline{1.414}$.

(ii) $X_P = X_L$ point D_2.

$$CG = \frac{R_P}{2}\sqrt{2^2 + 1^2} = \frac{\sqrt{5}}{2} R_P$$

$$CD_2 = R_P \frac{2}{\sqrt{5}} \qquad BD_2 = R_P \frac{1}{\sqrt{5}}$$

$$GD_2 = CG - CD_2 = \left(\frac{\sqrt{5}}{2} - \frac{2}{\sqrt{5}}\right) R_P = \frac{1}{2\sqrt{5}} R_P$$

$$\frac{BD_2}{GD_2} = \underline{2}$$

(iii) X_P to give maximum current in *coil* point D_3.

$$GD_3 = GH - GD = \frac{\sqrt{2} - 1}{2} R_P = .207 R_P$$

$$BD_3 = R_P \sin 22\tfrac{1}{2} \text{ degrees} = .382 R_P$$

$$\frac{BD_3}{GD_3} = \frac{.382}{.207} = \underline{1.84}.$$

These results indicate that the maximum current in the load occurs for values of X_P close to $X_P = X_L$ but that the circuit so adjusted becomes increasingly inductive as X_L is increased. When $X_L = R_P$ the phase angle is 45 degrees.

From this it is clear that X_L and X_P must be less than R_P for the circuit to be useful and preferably less than $\frac{1}{2} R_P$.

(D) $X_L << R_P$. *Fig.* 6 (D).

By inspection it can be seen that $\frac{BD}{GD}$ is a maximum when $X_P = X_L$ and A coincides with G as shown in Fig. 6 (D).

(To be continued.)

PATENT ABSTRACTS

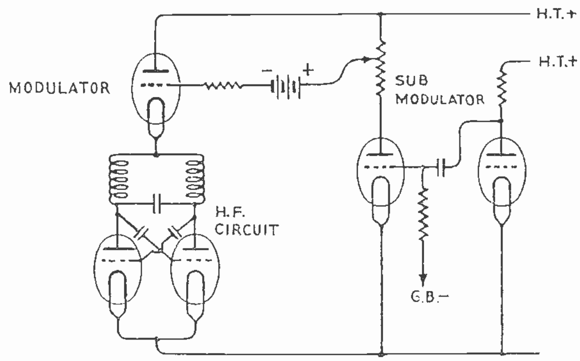
Under this heading it is intended to give abstracts in each issue of this Journal of a selection from the most recent inventions originating with the Marconi Co. These abstracts will stress the practical application of the devices described.

CARRIER WAVE MODULATION SYSTEMS

Application date, February 14th, 1936. No. 470,421.

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and N. H. Clough and E. Green.

This patent deals with a series modulated carrier system. The idea of series modulation of a carrier wave, that is a system in which the modulator valves are arranged in series with the H.F. valves, was one of the earliest put forward, but curiously enough no practicable series modulation circuits were developed until some few years ago, when the Marconi Company used a series modulation system on broadcast transmitters of large power, the Droitwich Station being one of the first introducing this principle. In this case the modulation filaments were connected to the negative end of the H.T. supply and earthed, but the H.F. valve and circuit was statically above earth although earthed to H.F. through a condenser.



Patent No. 470,421.

The present invention shows an improved series modulation system in which both the H.F. valves and the Controlling Modulation circuit are at earth potential. These are important features, and make such a system easy to design for even short wave H.F. circuits and simplify the modulation system. The circuit arrangements are as follows :—

The filaments of the high frequency valves are earthed and connected to the negative terminal of the high tension supply. The H.F. circuit may be of the usual centre tapped tuned anode type, and the modulation valve proper is connected between this tap and the H.T. positive supply point. The sub modulation valve, including a series anode resistance, is arranged in shunt across the H.T. supply, a tapping on the resistance being taken through a squegger resistance to the grids of the main modulators. Such an arrangement would not give 100 per cent. modulation owing to the fact that even when the main modulator acquires its maximum positive grid bias, i.e., when the sub modulation is cut off, the volts across the main modulator valve cannot fall to zero (usually only to about 20 per cent. of its normal H.T.). In consequence, an additional bias voltage either in the modulator grid circuit or in series with the sub modulator cathode, is added to make up this deficiency in volts, and enables 100 per cent. modulation to be obtained.

Another feature of the modulation is that the circuit arrangement lends itself without much alteration to a floating carrier system whereby the strength of the carrier produced is a function of the modulation amplitude.

The figure shows a schematic diagram of the series modulation circuit described.

CATHODE RAY TUBES

Application date, February 3rd, 1936. **No. 470.004.**

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

In television systems employing a cathode ray tube for the picture reproduction, modulation is obtained by varying the brilliancy of the light spot, but no change should take place in the size of spot.

In many cathode ray arrangements an image of the cathode is focussed on to the screen, but in this case halo effects are chiefly predominant when alteration is made to the light intensity.

In the present invention the method of obtaining a uniform size spot is to form an electrode gun consisting of a cathode, in front of which is an apertured disc or stop, which forms a control plate, followed by an electrode lens system of dimensions which are critically specified, and this in turn may be followed by a second stop. By this means an image of the aperture in the control electrode, which is in front of the cathode, rather than the cathode itself, is formed on the screen.

The stop nearest the cathode has an aperture small compared with the cathode area behind it, and the lens system following consists of a cylindrical electrode with apertured end discs.

With such a system no alteration of spot dimension occurs when the beam intensity is modulated.

DIRECTION FINDING SYSTEMS

Application date, November 20th, 1935. **No. 467.892.**

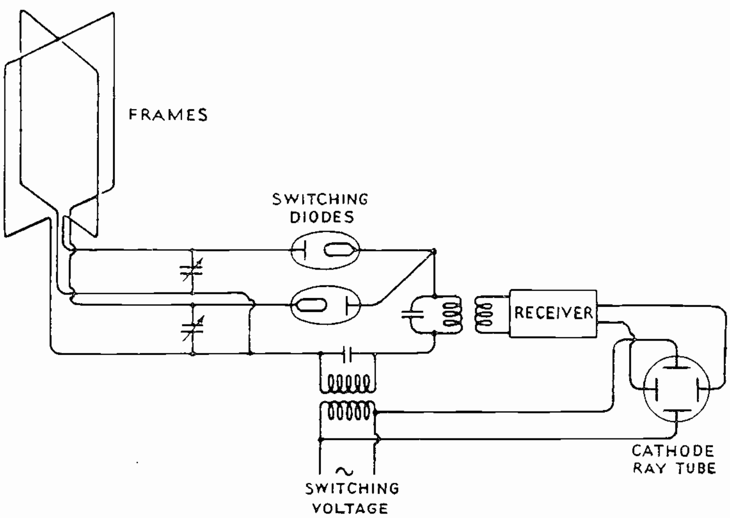
Patent issued to Marconi's Wireless Telegraph Co., Ltd., and G. M. Wright.

This invention discloses an equisignal receiving D.F. of the visual type that is visual in the sense that an equality of signals is judged by means of a cathode ray picture.

In one arrangement the outputs from two frames at right angles are rapidly switched in succession to a receiver. The output of the receiver is connected to one pair of plates of a cathode ray oscillograph, and the other pair is connected to the switching voltage. By this means the successive signals from each frame produce a picture which varies from a wide mark followed by a thin line, i.e., when one frame is receiving maximum signals and the other nothing, to a continuous mark, i.e., when the frames are each receiving the same amplitude signal.

It will be observed that since the equal signal zone is when the rate of change of signal strength is very great, the equal picture condition will be very sensitive to rotation of the aerial system, and in consequence a sharp indication of direction will be obtained.

An important feature of this system is that since the picture gives an indication of the type of signal being received, it is easy to distinguish between signals of various types, and the system does not therefore suffer the disadvantage that seems a feature of vision systems in general, namely, false readings.



Patent No. 467,892.

The electrical switching system adopted is novel. Each frame is connected to the output tuned circuits through a diode valve, one diode facing one way and the second the reverse way, and in the common lead from the output tuned circuit back to the frames is connected a transformer winding (shorted for H.F.), the primary of which is fed from a low frequency source. This means that each diode becomes conductive in turn at every alternate half cycle of low frequency, and in consequence each frame is coupled to the common output circuit successively.

The figure attached illustrates the patent.

LIGHT VALVES

Application date, November 20th, 1935. **No. 466,031.**

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

The type of light valve disclosed in this specification is actuated from the scanning beams of a television cathode ray tube and consists essentially of a reflecting or refracting surface the reflecting or refracting property of which can be varied at any point over the surface by means of a charge imparted to it by means of the Cathode Ray beam.

Specifically one such device consists of a very thin layer of crystals which are normally in optical contact with a plain glass surface owing to adhesion effect. If this adhesive force at any point be cancelled by an electric charge on the crystals so that optical contact between the crystals and the glass, which may form the totally reflecting surface of a prism, ceases, the prism will become totally reflecting at that point. At other points the optical contact will render the prism non-reflecting, and if the charge is imparted by a scanning cathode ray beam modulated by television signals the reflecting qualities of the prism at the individual elements of its operating force will be varied in accordance with the television signals.

In a second application fine carbon particles are scattered over the face of a totally reflecting prism, this face forming the end wall of a cathode ray tube. The scanning cathode ray beam charges the carbon particles positively and pulls them away from the end wall of the tube.

In a third arrangement a very thin sheet of mica coated with colloidal graphite is normally in optical contact with the glass surface, but is pulled away at any point from the surface by electric forces set up at that point by the scanning ray.

In another modification, the end wall of a Cathode Ray tube is formed of a plain surface. Inside the tube and parallel to the end wall is a sheet of mica, the space between which and the end wall is filled with asymmetrical crystals and the face of the mica nearer the electron gun of the tube is rendered refractive by sputtered or ruled conductive but separated particles. When the crystals are in repose, their flat surface will rest on the end wall and light from a light source which passes into the end wall from outside will not reach the reflective surface of the mica plate. When, however, a cathode ray beam scans the mica plate, charges are set up and the crystals are oriented in such a manner that light passes through them, is reflected, and passes out again. Accordingly as the cathode ray beam, modulated in accordance with picture signals, passes over the plate the crystals are oriented in accordance with the television signals and picture reproduction can be effected.

TELEVISION RECEIVERS

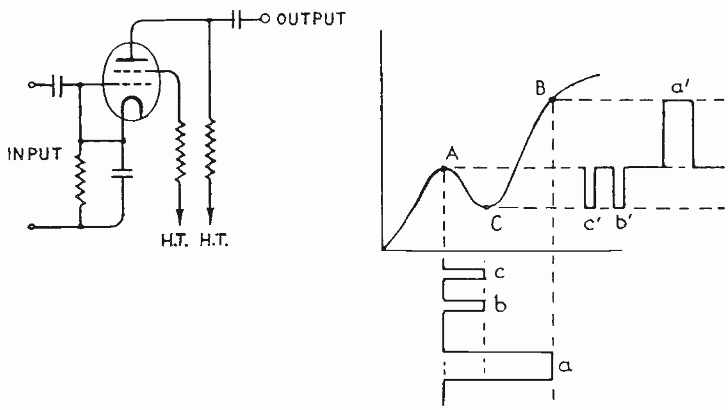
Application date, April 29th, 1936. **No. 472,923.**

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

This specification deals with the separation of line and frame synchronising impulses in television receivers. Normally line and frame synchronising signals are produced at the television transmitting station as impulses having relatively short and long durations respectively and they may or may not have in addition differing amplitudes.

Receiver circuits relying solely on the differing duration of the impulses for discriminating between them are, in general, neither reliable in operation nor simple in construction. Methods whereby the impulses are first given differing amplitudes and are then separated on this basis are therefore preferable.

Where the impulses have the same initial amplitudes but differ in duration they are, in accordance with this specification, applied to a screen grid valve having a time constant circuit in its cathode circuit, as shown in the first figure, so designed



Patent No. 472,923.

as to accept the relatively long framing signals but heavily to attenuate the short duration line signals. The screen grid valve is operated at such a point on its characteristic that the attenuated line signals produce a diminution in anode current whereas the relatively non-attenuated frame signals produce a rise in anode current as in the second figure.

In a modification of this system the time constant circuit may be inserted in the screen grid circuit of the screen grid valve and the valve may be so adjusted as to give a rise in anode current for the short duration line signals but a diminution of anode current for the longer duration frame signals.

In either of these methods signals of different duration but of the same initial amplitude and polarity may be transformed into signals of opposite polarity making subsequent separation exceedingly simple.

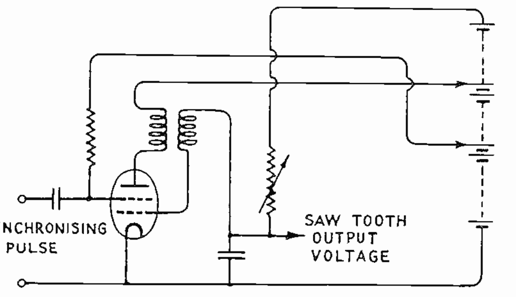
In a third modification the mixed impulses are applied to a triode, the resistance of which in parallel with a condenser forms the time constant circuit. A screen grid valve incorporated with this circuit functions in a similar manner to that described above. The operating point of the screen grid valve is so chosen that small increases of anode voltage caused by line signals produce only small changes in anode current, whereas frame signals produce relatively large changes. The screen grid valve may be replaced by a gas discharge tube such as a neon lamp and the triode may be replaced by a high resistance.

TELEVISION RECEIVERS

Application date, August 2nd, 1935. **No. 463,625.**

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

This specification describes a development of the well-known "trigger" oscillator as a generator of saw-tooth wave form and square-topped negative pulses, which may be used to modulate a cathode ray tube for suppressing the "flyback." An important feature of the invention is the small synchronising pulse required and its application to an electrode other than the grid or anode of the oscillator.



Patent No. 463,625.

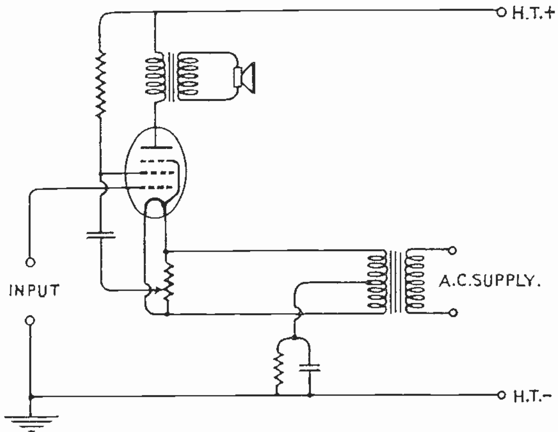
A tetrode or pentode valve is used with inductances in anode and grid circuits tightly coupled so as to produce violent oscillation under suitable conditions. The grid circuit is completed by a capacitance to cathode and a variable resistance to a high positive voltage. The resistance allows variation of the free frequency of the saw-tooth wave produced across the grid capacitance, and linearity of wave form is obtained by returning the resistance to a high positive voltage. The screen circuit contains a resistance connected to a suitable positive voltage and the synchronising pulse, which may be as low as one-fiftieth volt, is applied to the screened grid by a capacitor. Oscillation occurs when the charge on the grid capacitance approaches zero and grid current charges the capacitance negatively, thus negative bias is rapidly developed and oscillations cease. The "trigger" action causes a positive pulse of screen current—this becomes a negative pulse of voltage between screen and earth. The pulse is substantially square-topped and of duration equal to the flyback time of the saw-tooth voltage wave. The saw-tooth voltage may be used for electrostatic or, in combination with the square-topped wave, for electromagnetic deflection of the beam of a cathode ray tube.

RECEIVING CIRCUITS

Application date, October 24th, 1935. **No. 464,790.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and J. D. Brailsford.

This patent describes a method of cancelling the hum produced by the heaters of directly or indirectly heated tetrode or pentode valves by applying a neutralising hum voltage in the screened grid circuit. The screen is decoupled from the H.T. supply by a resistance and capacitance and the latter is returned to the movable arm of a potentiometer connected in parallel with the heaters. The centre tap of the mains transformer heater winding is connected to earth or to the positive end of the self-bias resistance. Adjustment of the potentiometer reduces the hum to



Patent No. 464,790.

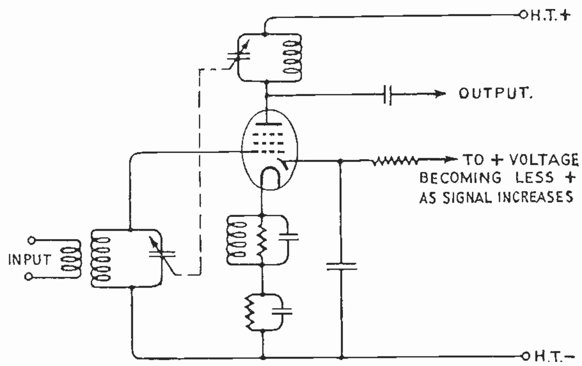
almost negligible values. Hum voltages due to inadequate H.T. smoothing may also be substantially reduced by this method.

GAIN CONTROL CIRCUITS

Application date, December 16th, 1935. **No. 467,430.**

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

The specification points out the advantages of obtaining R.F. gain control by negative feedback in the simultaneous reduction of signal voltage and distortion and



Patent No. 467,430.

gives various methods of obtaining feedback control. The most important is by means of a variable resistance in the cathode circuit of the controlled valve. This cathode resistance must be shunted by a high A.C. impedance having low D.C.

resistance in order to prevent variations of D.C. bias on the valve. The shunt impedance must not produce phase shift of the signal frequencies and a parallel L.C. circuit is recommended tuned either to resonate at the centre of the frequency band or to the signal frequency.

Methods are discussed of obtaining automatic variation of the feedback resistance and the importance of using a device having a linear resistance characteristic is stressed. A circuit is given in which a combination of negative feedback and variable-mu gain control is employed. Negative feedback control is obtained by a diode biassed from the A.V.C. line and in shunt with the cathode L.C. circuit.

ELECTRON BEAM VALVE

Application date, December 16th, 1935. **No. 467,573.**

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

This invention is concerned with the type of valve in which the electron stream is focussed in a beam, and its object is the production of more linear I_aE_g characteristics and also a negative slope on the "flat" part of the I_AE_A curves. Cross modulation effects are thus reduced and the selective properties of a tuned circuit connected to the anode may be improved.

The special features are the use of deflector plates following the normal control, focussing and shielding electrodes, a slotted anode and, behind this and at a higher voltage, a collector electrode. One deflector plate is connected to cathode and the other to a voltage derived from the anode or the total current. This voltage deflects the beam so that it falls partly on the anode slot to be collected by the electrode behind the anode. By suitable proportioning of the slot the I_AE_g characteristic curve may be made to approach linearity over most of its working range. Other forms such as flat-topped curves may also be obtained if desired. The deflector plate voltage may be derived from a resistance in the cathode circuit or a certain proportion of the anode voltage may be fed back.

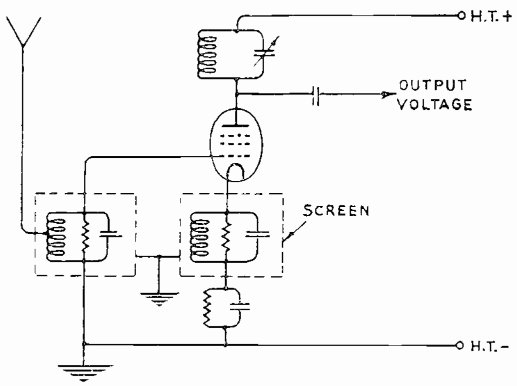
The negative slope on the I_AE_A characteristics is obtained by using the slotted anode as an electron lens such that increase of anode voltage concentrates the beam in and around the slot.

RECEIVING CIRCUITS

Application date, December 10th, 1935. **No. 468,784.**

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

It is almost impossible to arrange direct aerial coupling, suitable for all types of aerials, to the first sharply tuned circuit of a receiver without reducing selectivity or disturbing ganging. The invention described in this specification overcomes these difficulties and in addition reduces cross modulation effects to a minimum. The first sharply tuned circuit of a receiver is coupled in the anode circuit of a valve which has a flatly tuned grid circuit and negative feedback. The aerial is coupled to a tapping on the grid circuit. Negative feedback is obtained by placing in the cathode circuit a parallel R.L.C. circuit of low Q value, tuned to the centre of the particular frequency band required. The value of R is so chosen that a band pass effect is obtained. If the suggested values of L = 2,000 μhys, C = 12.5 $\mu\mu$F and R = 12,500 ~ for the 200—550 metre range are used, the effective grid voltage is reduced to one-eleventh of the applied grid voltage. Negative feedback falls off outside the band pass range of the cathode circuit, but the frequency characteristics of the grid tuned circuit, which is similar to that in the cathode circuit, cause attenuation of frequencies outside the pass band. An appreciable gain may be obtained from the valve if the first tuned circuit of the receiver is sharply tuned, and the frequency-sensitivity characteristic of this stage can be adjusted to produce in conjunction with the receiver an almost constant overall sensitivity.

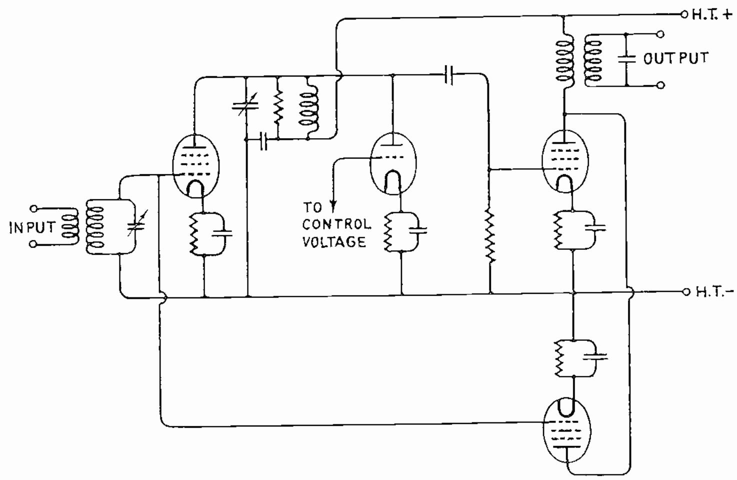


GAIN CONTROL CIRCUITS

Application date, February 3rd, 1936. **No. 469,895.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and J. D. Brailsford.

The method of automatic gain control disclosed in this patent involves the use of a variable impedance and does not depend on the curvature of the anode current-grid voltage characteristic of a valve. Modulation distortion and cross modulation can be reduced to a minimum since the value of control impedance can be made independent of the voltage applied to it. The input signal is supplied to two channels, the outputs from which are connected in opposition. The output



valve in one channel is preceded by an intermediate valve, the gain of which is varied by an impedance in parallel with its anode circuit. By suitably adjusting the initial gain of this channel a rising or falling gain characteristic can be obtained by variation of the control impedance. If the gains of the two channels are equal any variation of control impedance increases the output voltage, whilst the control impedance can produce either increase or decrease of output when the two channels have different initial gains. The variable impedance may consist of a variety of devices, but for automatic operation grid bias derived from the input signal may be used to control the anode impedance of a valve.

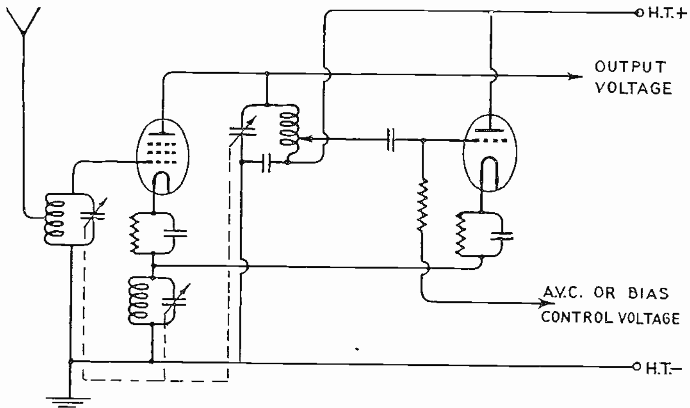
The variable impedance valve is shunted across a tuned circuit in the anode of the intermediate valve and normally no difficulties due to incorrect phasing of the output voltages occur. When a large degree of control is required (about 40 dbs.) certain phase conditions must be satisfied and these are indicated in the specification.

GAIN CONTROL CIRCUITS

Application date, February 3rd, 1936. **No. 469,896.**

Patent issued to Marconi's Wireless Telegraph Co., Ltd., and J. D. Brailsford.

Control of gain of an R.F. amplifier by negative feedback is the principle underlying this invention. Three circuits tuned to the incoming signal frequency are used in the grid, anode, and cathode circuit of an R.F. valve. The cathode tuned



circuit, which normally develops a voltage in opposition to the applied grid voltage, is also in the cathode circuit of a control valve, the grid of which is connected to a tapping point on the anode tuned circuit of the R.F. valve. The control valve—its anode is connected direct to H.T. positive—provides positive feedback and its grid bias, which may be controlled manually or automatically, is initially adjusted so that it cancels the negative feedback due to the R.F. valve. Increase of this grid bias decreases the positive feedback and so reduces the output voltage from the R.F. valve. With high slope valves a control of 60 dbs. may be obtained.

The advantages of the system are low distortion—gain control does not depend on valve curvature—output voltage almost independent of input voltage so that succeeding R.F. stages can be designed for small signal voltages and a high degree of control by small variations of bias voltage.