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Radio as a Course for Graduation?

IN the course of casual conversation one sometimes hears the suggestion made that perhaps Universities ought to provide a degree course in Radio, and by this is meant, presumably, a course for graduation and not an additional course and second degree, subsequent to graduation. Although it is scarcely to be credited that such suggestions can be serious and weighty, nevertheless it is worth while to consider what is involved in this strange idea and to consider what is involved in the concept of graduation itself.

In this country, graduation denotes the satisfactory completion of a course of study whose central purpose is to teach people how to learn systematically for themselves and to attain an awareness of mind towards 'good learning'; it is the first step on the endless road of self education in intellectual ideas and logical thought. The discipline for this first step requires that much use must be made of associated facts and sciences, but they are used only by way of example and exercise towards the main purpose; which can be attained through a variety of exemplary systems, or degree courses as they are called. Undergraduates are apt to take it for granted that the first purpose of their studies is to attain specific knowledge about a particular subject; thus mistaking the means for the end. A dozen years may pass before it slowly dawns on a graduate that what he really did, in undergraduate days, was to educate and train his mind and that the substantial store of knowledge which he gained from his courses was by way of an extra benefit, thrown in in the process. Correct and precise knowledge in some subject provides the equivalent of an A.B.C. for writing exercises, set by way of examining the state of mental education which has been attained.

Engineering Science can be utilized as a vehicle for a graduation course and those who intend to become Professional Engineers commonly ride in this vehicle in preference to, say, mathematics or physics. Some nine-tenths of those who take a degree in engineering do in fact become professional engineers, but a much smaller proportion of those who graduate through mathematics or physics concentrate the work of their life in the active practice and advancement of these subjects. Therefore it is entirely proper that graduation courses in engineering should be particularly appropriate to a life to be spent in our learned profession. Accordingly the courses ought to include the most inherent fundamentals of those parts of the physical sciences which an engineer may get called upon to study during his professional career but, in the main, going only just far enough to equip him for further study if needed. In Electrical Engineering, the scope of these possibilities is getting wider at an ever increasing rate.

In addition to this initial grounding in the fundamental concepts, an engineering course *must* include a severe discipline in the numerical solution of the widest possible variety of academic problems having some bearing on engineering apparatus. A vague general idea of how things work, in broadest outline, or a perception that a given problem must be capable of solution by those who know how to do it is not good enough. Ceaseless striving to obtain quantitative results must be inculcated throughout the course; the discipline for this is the working right out of endless numerical problems and it does not detract from the main purpose if many of them become rather pedagogic in the process of making up an example which can be solved completely. Competence in the technique of numerical solutions of the academic type of problem constitutes a major difference between the discipline of an engineering course and a course in the Natural Sciences and is one of the many justifications for its separate existence.

Let us now see what a graduating course would have to cover if it was to be restricted particularly to Radio. We will presume the undergraduate starts at the University with a real competence in Newtonian mechanics and the elements of Higher Mathematics. The graduation course would have to include the whole of electrostatics and the electromagnetism of steady currents and the whole theory of d.c. and of a.c. circuits and the association with them of thermionic, photo-electric and gas-filled valves and cathode-ray tubes; and it would include the full theory of the transmission line; presumably it would not omit the magnetic qualities of iron or the permittive properties of dielectrics. So far we have suggested nothing which is not in a normal engineering course. It would have to include some thermodynamics in order to deal with thermionic emission and gaseous discharge; but might then fall short of the normal course by omitting the thermodynamics of heat engines.

A full development of the electromagnetic theory, including the theory of aerial design, would be a prominent part of a radio course because this is the only thing peculiar to Radio; virtually the whole of this is absent in the normal course and ought not all to be absent. The preliminary work for the electromagnetic theory would necessarily bring back the mathematics associated with inverse-square law; it is much to be regretted that most of this was squeezed out of the normal course, some thirty years ago.

No doubt considerable mental training could be provided within the scope of this field, which is vastly narrower than that of the normal course. At this stage we ask if it is particularly suited to those who would become professional engineers practising in radio communications. And we should see it is not suited until it had included the study of the air-gap magnetic circuit, iron-cored transformers and, at very least, the operational behaviour of rotating electrical machinery; some teaching in mechanisms, calculation of mechanical stresses, some elements of heat engines and something about power cables. And so on, and so on, till we come back nearer and nearer to the normal course. And this must be so inevitably because there can be no such thing as a Radio Engineer but only a Professional Engineer whose practice may be particularly associated with Radio. There can and must be Radio Technicians, and Special Radio Courses are provided for them by certain Educational Institutes.

Radio courses might possibly contain some elements of molecular physics, crystal structure and quantum mechanics; but these ought to be included in normal courses because they are very difficult subjects to embark on after graduation. Normal courses commonly include a close study of the performance of rotational machinery and they do so because this offers quite peculiar opportunities for mental training in exact thought and the tracing out of the net effect of a succession of contributory causes whose separate magnitudes are not known precisely. A Radio course would probably omit all this and be much the poorer thereby.

If we have in mind education for those who will be professional engineers then surely the very idea of a graduation course concentrated on radio is complete anathema; for it would but open, inevitably, a fresh gate for detailed and pedestrian technological instruction of the 'immediately useful' variety.

Far from providing a narrower course, we will hope that Universities will be quite fearless in providing a normal course which is even wider than today in fundamental teaching but without loss of the firmly established strict discipline of quantitative working. To make room for fresh fundamental ideas some details which are not really as important as they seem will have to be thrown out; but the difficulty is that there are so very few of these details still left in; it is quite surprising how few can be found when a course is searched through systematically. But may the mistake never be made in this country of getting over the difficulty by concentrating on Radio.

E. B. M.

NEGATIVE-FEEDBACK AMPLIFIERS

Conditions for Maximal Flatness

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Introduction

WHEN negative feedback is applied to an amplifier having two or more stages it is well-known that, once a certain small amount of feedback is exceeded, the frequency characteristic develops peaks at the edges of the band. These peaks get higher and sharper as the feedback is increased until, in an amplifier of three or more stages, self-oscillation commences. There are few direct references in the literature to these peaks in the frequency characteristic. Fritzinger[1] suggests removing them by filters outside the feedback loop and Terman and Pan[2] make the same suggestion in an experimental study of a three-stage amplifier.

The present analysis shows that it is possible to control the shape of the external frequency characteristic without using any additional components either inside or outside the feedback loop. In particular, a substantially flat response over the maximum possible frequency band can be obtained with any amount of feedback merely by properly distributing the gain between the various stages of the amplifier.

1. Basic Relations

Consider an amplifier having a voltage-amplification A; with an input voltage e, the output voltage will be Ae. If now a fraction β of this output voltage be returned to the input circuit so as to produce negative feedback, the input must be increased to $e + A\beta e = (1 + A\beta)e$ in order to maintain the output voltage at its original value. The *external amplification* A_x obtained with feedback is, then

$$A_x = A/(1 + A\beta) \quad \ldots \quad \ldots \quad (1.1)$$

The sign convention is to be chosen so as to make A and β both positive quantities when the feedback is negative.

Since A is a complex quantity (1.1) includes both magnitude and phase effects; only the case in which β is independent of frequency is considered. The present discussion is concerned with the 'frequency characteristic' of the external amplification; i.e., with $|A_x|$ as a function of frequency. It is convenient to express this by the variable $a = |A_o/A_x|$, where A_o is the value of A in the region where it does not vary significantly with frequency.

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The 'feedback factor' F is defined as $1 + A_o\beta$. From (1.1)

$$1/A_x = \beta + 1/A$$

$$A_o/A_x = A_o\beta + A_o/A = F - 1 + A_o/A$$

hence $$a^2 = |A_o/A_x|^2 = |F - 1 + A_o/A|^2 \quad (1.3)$$

In this expression, the frequency enters through the term A_o/A which represents, in both magnitude and phase, the frequency characteristic of the amplifier alone; i.e., without feedback.

2. Single-Stage Amplifier

The case of the single-stage amplifier is almost trivial and there is no question of controlling the distribution of the gain, but the analysis illustrates the general procedure and the notation.

The type of amplifier considered is shown in Fig. 1. At very high frequencies the amplification is determined by the mutual conductance g_m and stray capacitance C; it is independent of the coupling resistance R and is given by

$$A_a = g_m/\omega C \quad \ldots \quad \ldots \quad \ldots \quad (2.1)$$

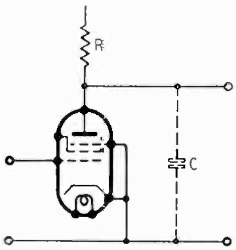

Fig. 1. *Single-stage amplifier.*

A plot of (2.1) is the amplification asymptote; it is impossible, in a low-pass system, to produce a frequency characteristic lying measurably above this at very high frequencies.

The amplification at any frequency is given by

$$A = g_mR/(1 + j\omega CR)$$

whence $$A_o/A = (1 + jx) \quad \ldots \quad \ldots \quad (2.2)$$

where $A_o = g_m R$; $x = \omega CR$

The latter form of (2.2) is the convenient one to use; A_o is the low-frequency amplification and x is the frequency-variable.

The external amplification with feedback applied is given by

$$1/A_x = \beta + (1 + jx)/A_o$$

whence $$A_o/A_x = F + jx \quad \ldots \quad \ldots \quad \ldots \quad (2.3)$$

The external amplification frequency-characteristic is

$$A_o/FA_x = A_{xo}/A_x = 1 + jx/F \quad \ldots \quad \ldots \quad (2.4)$$

Comparing (2.4) with (2.2), it will be seen that the effect of the feedback is simply to increase the bandwidth in the same ratio as it decreases the gain.

3. Two-Stage Amplifier

Consider first an amplifier consisting of two identical stages like Fig. 1. Then the frequency characteristic is given by

$$A_o/A = (1 + jx)^2 \quad \ldots \quad \ldots \quad \ldots \quad (3.1)$$

The more general case of two stages of different characteristics must be considered. Since the amplification asymptote is independent of the distribution of the amplification between the stages, the notation used in the analysis must make this clear; this is achieved by writing the frequency parameters for the two stages as

$$x_1 = bx\,;\ x_2 = x/b \quad \ldots \quad \ldots \quad \ldots \quad (3.2)$$

Where x^2 = the parameter defining the asymptotic amplification for the two stages.

Then $A_o/A = (1 + jbx)\,(1 + jx/b)$

$$= 1 - x^2 + j\,(b + 1/b)\,x \quad \ldots \quad (3.3)$$

It is convenient to replace b by the 'staggering'* coefficient S, where

$$2S = b + 1/b \quad \ldots \quad \ldots \quad \ldots \quad (3.4)$$

For two identical stages, $S = 1$; as the stages are 'staggered' by increasing the amplification of one (by increasing the coupling resistance) and, simultaneously, decreasing that of the other to keep the total amplification constant, the bandwidths are staggered in the same ratio and S increases.

Then (3.3) becomes

$$A_o/A = 1 - x^2 + j2Sx \quad \ldots \quad \ldots \quad (3.5)$$

then $A_o/A_x = 1 + A_o\beta - x^2 + j2Sx \quad \ldots \quad (3.6)$

$$a^2 = F^2 - 2(F - 2S^2)\,x^2 + x^4 \quad (3.7)$$

The condition for a maximum or minimum in the frequency response can be found by equating to zero the differential coefficient of (3.7) with respect to x^2, so as to obtain the value x_p of x corresponding to the peak.

This gives, $x_p{}^2 = F - 2S^2 \quad \ldots \quad \ldots \quad (3.8)$

and corresponds to a minimum of a and so to a maximum in the amplification.

Inserting this value of x_p into (3.7) gives the value of a at the peak.

$$a_p{}^2 = 4S^2(F - S^2) \quad \ldots \quad \ldots \quad (3.9)$$

Eliminating F between (3.8) and (3.9) gives the locus of maxima,

$$a_p{}^2 = 4\,S^2(x_p{}^2 + S^2) \quad \ldots \quad \ldots \quad (3.10)$$

* This use of the word 'staggering' is unusual but convenient and is obviously derived from the r.f. practice of staggering the tuning of a number of resonant circuits on either side of some reference frequency. Here it is applied rather to the time constants of successive RC circuits: the values in different stages are adjusted to be above and below a reference value.

The height of the peak relative to the level part at low frequencies is given by

$$(a_p/a_o)^2 = (4S^2/F^2)\,(F - S^2) \ldots \quad \ldots \quad (3.11)$$

If $F = 2S^2$ (3.7) or (3.11) shows that there is no peak in the frequency response, which reduces to

$$a^2 = F^2 + x^4 \quad \ldots \quad \ldots \quad \ldots \quad (3.12)$$

This elimination of the terms in x^2 from the frequency characteristic gives the condition of *maximal flatness*, (m.f.) a term introduced by Landon[3]. This condition defines the largest value of F (for a given S) at which the frequency characteristic is free from maxima and minima.

(3.12) can be written

$$|A_{xo}/A_x|^2 = 1 + x^4/F^2 \quad \ldots \quad \ldots \quad (3.13)$$

which shows that, with m.f., the effect of the feedback is (i) to change x into $x/\sqrt{F}$ and (ii) to sharpen the knee of the curve. The universal m.f. frequency characteristic is plotted in Fig. 2.

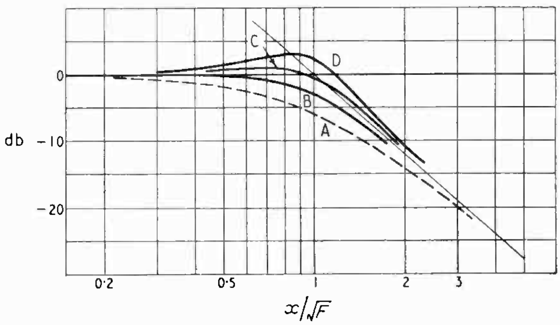

Fig. 2. Frequency characteristics of two-stage amplifier. Curve A is without feedback and curves B, C and D are respectively for maximal flatness, a 1-db peak and a 3-db peak.

The response near the knee can be improved (raised) if a small peak can be tolerated, by a slight departure from the m.f. condition. The height of the peak above the level part can be defined by $1/\alpha$

$$\text{where } a_p/a_o = \alpha \quad \ldots \quad \ldots \quad \ldots \quad (3.14)$$

With the aid of (3.11) the required staggering to produce a peak of this height is found to be,

$$S^2 = (F/2)\,(1 - \sqrt{1 - \alpha^2}) \quad \ldots \quad \ldots \quad (3.15)$$

The frequency characteristic, with this relation between S and F, becomes

$$\left|\frac{A_{xo}}{A_x}\right|^2 = 1 - 2\frac{x^2}{F}\sqrt{(1 - \alpha^2)} + \frac{x^4}{F^2} \quad \ldots \quad (3.16)$$

The curves of Fig. 2 illustrate the type of response which can be obtained by using slightly more feedback than is required to give maximal flatness.

4. Three-Stage Amplifier

For three identical stages the frequency characteristic is given by

$$A_o/A = (1 + jx)^3 \quad \ldots \quad \ldots \quad \ldots \quad (4.1)$$

When the stages are staggered there are, in the general case, two independent staggering coefficients; the asymptote is retained by writing

$$x_1 = b_{1x} \,;\; x_2 = b_2 x \,;\; x_3 = x/b_1 b_2 \quad \ldots \quad (4.2)$$

Then $A_o/A = 1 + jK_1 x - K_2 x^2 - jx^3$ .. (4.3)

where

$$K_1 = b_1 + b_2 + 1/b_1 b_2 \,;\; K_2 = b_1 b_2 + 1/b_1 + 1/b_2$$

Then, on the application of feedback, the frequency characteristic becomes

$$A_o/A_x = (F - K_2 x^2) + jx(K_1 - x^2) \quad \ldots \quad (4.4)$$

$$a^2 = F^2 + (K_1^2 - 2FK_2)x^2 + (K_2^2 - 2K_1)x^4 + x^6 \quad \ldots \quad (4.5)$$

The ideal m.f. conditions, making the coefficients of x^2 and x^4 both zero, are

$$K_1^2 = 2FK_2 \,;\; K_3 \equiv K_2^2 - 2K_1 = 0 \quad \ldots \quad (4.6)$$

Consider the second of these; inserting the, values of K_1 and K_2 from (4.3) it becomes

$$K_3 = b_1^2 b_2^2 + (1/b_1^2) + (1/b_2^2) = 0$$

which is impossible.

The practical m.f. conditions are therefore

$$K_1^2 = 2FK_2 \,;\; K_3 \equiv K_2^2 - 2K_1 \text{ minimum} \quad (4.7)$$

Minimizing $K_2^2 - 2K_1$ with respect to b_1 and b_2 gives $b_1 = 1 = b_2$ and $F = 1.5$

This amount of feedback is quite useless; the problem is, therefore, to chose b_1 and b_2 for any specified value of feedback in such a way as to make $K_2^2 - 2K_1$ small.

It is convenient to write,

$$\left.\begin{array}{l} b_1 b_2 = 1/B^2 \\ b_1/b_2 = r^2 \end{array}\right\} \quad \ldots \quad \ldots \quad \ldots \quad (4.8)$$

With this notation the m.f. conditions become

$$B^4 + (1/B^2)(r^2 + 1/r^2) = 2(F-1)\{1/B^2 + (r + 1/r)B\}$$

$$K_3 = (1/B^4) + (r^2 + 1/r^2)B^2 \text{ minimum} \quad (4.9)$$

Inspection of these makes it clear that B must be greater than 1, otherwise K_3 [the coefficient of the x^4 term in (4.5)] becomes large whatever the value of r. Physically this means that the geometric mean bandwidth of two of the stages must be greater than that of the third. In the simple case when two of the stages are identical this may be summarized 'two wide, one narrow.' In the literature Terman[4], Day and Russell[5] and Wermann[6] all recommend the use of 'two narrow, one wide,' the opposite of the present case. These writers agree in giving a value for the *maximum* feedback which can be applied without self-oscillation and it is easily verified that their expression is correct. With the arrangement now proposed, however, there is no limit to the amount of feedback which may be applied, not merely without risk of self-oscillation but with a frequency characteristic of controlled shape throughout its range.

When $(r + 1/r)/B_3$ is negligible compared with 1, (4.9) becomes approximately,

$$B^3 = 2(F-1)(r + 1/r) \,;\; K_3 = B^2(r^2 + 1/r^2) \quad (4.10)$$

When r is not far from unity, the approximations in (4.10) are within 1% for $B \geqslant 2.4$ (i.e., for $F \geqslant 4.5$) and so are acceptable for all useful amounts of feedback. Remembering that $r^2 + 1/r^2 = (r + 1/r)^2 - 2$, r can be eliminated between (4.10), giving

$$K_3 = B^8/4(F-1)^2 \quad \ldots \quad \ldots \quad \ldots \quad (4.11)$$

Therefore, for a given amount of feedback, B must be made as small as possible; the first of (4.10) shows that this requires $(r + 1/r)$ to be made as small as possible, and this is achieved by setting $r = 1$. Physically this means that two of the stages are best made identical in bandwidth; this reduces the number of independent parameters to one, B, and is a very welcome simplification.

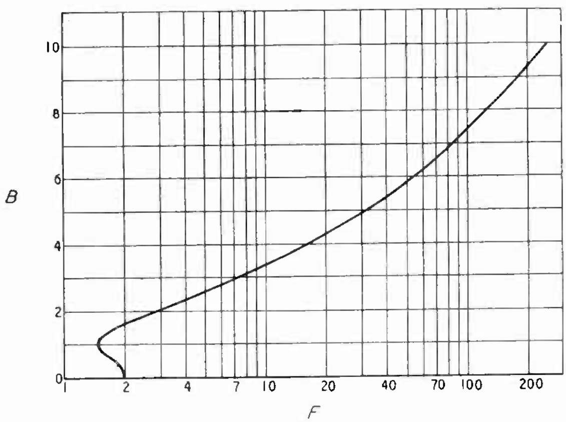

Fig. 3. *Relation between the staggering factor B and the feedback factor F for maximal flatness in a three-stage amplifier.*

5. Three-Stage Amplifier, 'Two Wide, One Narrow'

When $r = 1$, the frequency characteristic is

$$A_o/A = (1 + jx/B)^2 (1 + jB^2 x) = (1 - K_2 x^2) + jx(K_1 - x^2)$$

where $K_1 = B^2 + 2/B$

$$K_2 = 2B + 1/B^2 \quad \ldots \quad \ldots \quad \ldots \quad (5.1)$$

On the application of feedback, the external amplification is given by

$$A_o/A_x = (F - K_2 x^2) + jx(K_1 - x^2) \quad (5.2)$$

The frequency characteristic is

$$a^2 = F^2 + (K_1^2 - 2K_2F)x^2 + (K_2^2 - 2K_1)\ x^4 + x^6 \quad \ldots \quad \ldots \quad (5.3)$$

The m.f. condition is

$$F = K_1^2/2K_2 = \frac{1}{2} \cdot \frac{(B^3 + 2)^2}{2B^3 + 1} \quad \ldots \quad \ldots \quad (5.4)$$

This relation between the feedback factor F and the staggering coefficient B is plotted in Fig. 3 for $0 < B < 10$. The values of $B < 1$ correspond to the 'two narrow, one wide' state; it will be seen that the amount of feedback which is required to produce m.f. is too small to be of any practical use. The range of values of interest is $2 < B < 10$ corresponding to $3 < F < 250$ (roughly 10 to 44 db).

When $B \geqslant 2$, (5.4) may be approximated within 1.4% by

$$F \approx 1 + B^3/4 \quad \ldots \quad \ldots \quad (5.5)$$

The coefficient of x^4 in (5.3) is

$$(1 + 2B^6)/B^4 \approx 2B^2 \quad \ldots \quad \ldots \quad (5.6)$$

This approximate value is within 1% for $B \geqslant 1.92$; using it the m.f. frequency characteristic becomes

$$a^2 = F^2 + 2B^2x^4 + x^6 \quad \ldots \quad \ldots \quad (5.7)$$

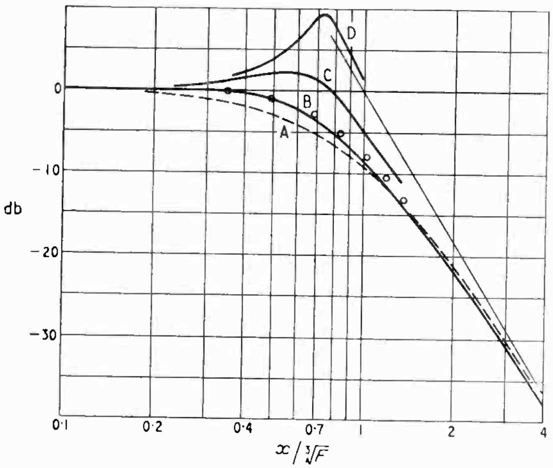

Fig. 4. Frequency characteristics of three-stage amplifier. Curve A is without feedback and curves B, C and D are respectively for maximal flatness and with feedback equal to two and four times the m.f. value. The circles indicate the m.f. curve for $F = \sqrt{10}$.

5.1 *Universal M.F. Curve.*

Dividing (5.7) by F^2 leads to a universal curve which gives the m.f. frequency characteristic.

$$(a/F)^2 = 1 + 2(B/F)^2x^4 + (1/F^2)x^6 \quad (5.8)$$

From (5.5) $B^3 = 4(F - 1) \approx 4F$

Using this to eliminate B in (5.8)

$$(a/F)^2 \approx 1 + 2^{\frac{7}{3}}(x/F^{\frac{1}{3}})^4 + (x/F^{\frac{1}{3}})^6$$
$$\approx 1 + 5y^4 + y^6$$

where $y \equiv x/F^{\frac{1}{3}}$ $\quad \ldots \quad \ldots \quad \ldots \quad (5.9)$

This curve is plotted in Fig. 4. Points computed without approximation for $F = 10$, $10^{\frac{3}{2}}$, 10^2, fall, for practical purposes, on the line. Points computed for $F = 10^{\frac{1}{2}}$ are plotted as small circles in the figure.

The effect of applying m.f. feedback is illustrated by this figure, which includes, for comparison, the response curve of an amplifier having three identical stages and no feedback. The effect of the feedback is similar to that found for a two-stage amplifier; it increases the bandwidth roughly in proportion to $\sqrt[3]{F}$ and sharpens the knee of the curve.

5.2 *Effect of Increased Feedback.*

In the case of the two-stage amplifier, an improvement in the region of the knee of the curve was found by increasing the feedback slightly above the m.f. value so as to produce a small, flat peak in the response. A similar result is to be expected in the case of the three-stage amplifier but, since self-oscillation is possible in this case, care must be exercised. It is possible that a further improvement might result from giving all three stages different bandwidths when the feedback is not m.f. The attempt to analyse this general case on the lines of (3.8) to (3.11) leads, however, after a great deal of tedious algebra, to a series of complex and intractable expressions from which it is difficult to draw any definite conclusion.

A procedure which is relatively simple, both analytically and in practice, is to set up the m.f. condition and then examine the effect of increasing the feedback. We can first find the amount of feedback required to produce self-oscillation by inspecting (5.2). The two conditions required are (i) a phase-shift of 180° and (ii) an $A\beta$ loop amplification of unity; this corresponds to $A_x = \infty$. The conditions are

$$x^2 = K_1$$
$$F = K_2x^2 \quad \ldots \quad \ldots \quad \ldots \quad (5.10)$$

The critical amount of feedback to produce self-oscillation is therefore

$$F_c = K_1K_2$$
$$= 5 + 2B^3 + 2/B^3 = 8F_f - 3 \quad (5.11)$$

where F_f is the m.f. value of feedback factor.

The 'stability margin' of an m.f. three stage amplifier thus increases from about 17 db with 10 db feedback to about 18 db when the feedback is very large.

For a given value of B, (5.3) can be used to give the frequency characteristic resulting from a change in the feedback from the m.f. value. It is now necessary to distinguish between F_f, the m.f. value of feedback factor for the given amplifier, and F, the feedback actually applied.

Using (5.4), (5.3) becomes

$$a^2 = F^2 + K_1{}^2(1-F/F_f)x^2 + 2B^2x^4 + x^6 \quad (5.12)$$

Using (5.1) and (5.5.) to express the coefficients in terms of F and F_f only, the universal frequency characteristic corresponding to (5.9) is

$$\left(\frac{a}{F}\right)^2 = 1 - 2^{\frac{2}{3}}\left\{\frac{2F_f-1}{F}\left(\frac{F}{F_f-1}\right)^{\frac{1}{3}}\right\}^2\left(\frac{F}{F_f}-1\right)y^2 + 2^{\frac{7}{3}}\left(\frac{F_f-1}{F}\right)^{\frac{2}{3}}y^4 + y^6 \quad .. \quad (5.13)$$

When $F_f \gg 1$ this can be further simplified to

$$\left(\frac{a}{F}\right)^2 = 1 - 2^{\frac{8}{3}}\left(\frac{F_f}{F}\right)^{\frac{4}{3}}\left(\frac{F}{F_f}-1\right)y^2 + 2^{\frac{7}{3}}\left(\frac{F_f}{F}\right)^{\frac{2}{3}}y^4 + y^6 \quad .. \quad .. \quad (5.14)$$

As a check on the inaccuracy of the approximations in (5.14) we may insert $F = 8F_f$, which, by (5.11), should yield a zero minimum value for the right-hand side. The value obtained is -0.1 which shows that the approximations are not such as to introduce errors of any practical importance.

Curves of (5.14) for $F/F_f = 2$ and 4 are plotted in Fig. 4.

6. Effective Feedback

An amplifier to which feedback near the m.f. value is applied has a substantially flat external frequency characteristic which extends nearly up to the amplification asymptote. The phase change in the amplifier so near its cut-off frequency must be considerable; we may therefore say, loosely, that the flat frequency characteristic is obtained by balancing the rise which the feedback would produce (because of this phase shift) against a fall due to the attenuation in the narrow stage of the staggered amplifier. This means that the amplification without feedback must have fallen considerably by that frequency at which the knee of the m.f. response curve is reached.

The quantity which is effective in reducing noise, distortion and fluctuations due to changing valve and circuit parameters is $|1 + A\beta|$; this may properly be called the 'effective feedback' and indicated by the symbol F_e. Since only the case in which β is independent of frequency is being considered, F_e will fall as A falls and will become negative (corresponding to positive feedback) when A suffers sufficient change of phase.

The fact that the *shape* of the frequency characteristic depends upon the amount of feedback also shows that the effective feedback must be small at the edge of the band. The frequency band over which the effective feedback does not fall much below its low-frequency value is that part over which the response remains level when the feedback is changed.

The frequency at which A, and therefore F_e, begin to fall can be located when we remember that the frequency characteristic of A is controlled (except at very high frequencies) by the 'narrow' stage of the amplifier. The value to which F_e has fallen at the edge of the band can easily be computed directly.

It is, of course, inevitable, in any feedback amplifier, that the effective feedback falls as the asymptote is approached. The point is of unusual importance in m.f. amplifiers because the useful (i.e., substantially flat) part of the frequency characteristic also extends nearly to the asymptote.

6.1 *Two-stage Amplifier.*

The frequency at which A begins to fall in an m.f. amplifier with large feedback can be found from (3.3) which, for $x^2 \ll 1$ and $b^2 \gg 1$, becomes approximately.

$$A_0/A \approx 1 + jbx \quad .. \quad .. \quad .. \quad (6.1)$$

which is, of course, the expression for the narrow stage alone.

A will, therefore, have fallen 3 db at

$$x = 1/b \approx 1/2S = 1/\sqrt{2F} \quad .. \quad .. \quad (6.2)$$

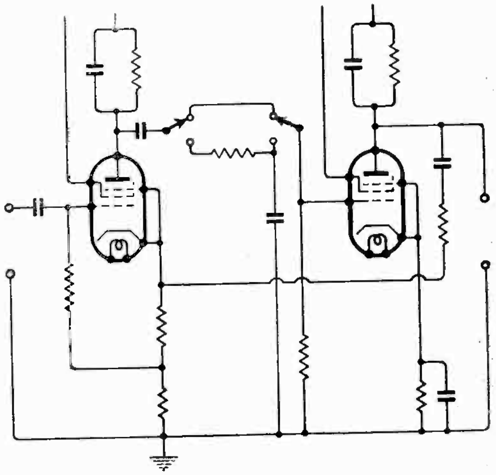

Fig. 5. *Amplifier for experimental verification.*

The level part of the frequency response extrapolates to meet the asymptote at $x = \sqrt{F}$, and the m.f. amplification has fallen by 3 db at this point, which may therefore be taken as defining the knee of the curve. The frequency at which A falls by 3 db is, therefore, about $1/\sqrt{2F}$ times the 'knee frequency.'

The effective feedback at the high-frequency edge of the band can be found by putting $x = \sqrt{F}$ and the m.f. condition ($F = 2S^2$) into (3.5), which gives,

$$A_o/A = 1 - F + j\sqrt{2F} \approx F(-1 + j\sqrt{2}) \qquad (6.3)$$

$$\text{Now } 1/A\beta = (A_o/A)/A_o\beta \approx A_o/AF$$
$$\approx -1 + j\sqrt{2} \qquad (6.4)$$
$$\therefore A\beta \approx (-1 - j\sqrt{2})/3$$
$$1 + A\beta = 2/3 - j\sqrt{2}/3$$
$$F_e = |1 + A\beta| = \sqrt{2/3} \qquad (6.5)$$

We may therefore say, roughly, that the feedback vanishes at the edge of the band.

6.2 *Three-Stage Amplifier.*

Following the lines of (6.1) to (6.5) we obtain, for the frequency at which A has fallen 3 db,

$$x = 1/B^2 \approx (4F)^{-\frac{2}{3}}$$

The m.f. response is 3-db down at $x = (2/3)F^{\frac{1}{3}}$, where the frequency is about $1.7F$ times that which gives a 3-db fall in A.

The effective feedback near the knee of the m.f. frequency characteristic, at $x = (2/3)F^{\frac{1}{3}}$, is obtained from (5.1), using the approximation $B^3 \gg 1$,

$$A_o/A \approx 1 - 1.33F + j2.37F \approx F(-1.33 + j2.37)$$
$$A\beta \approx -0.19 - j0.34$$
$$1 + A\beta \approx 0.81 - j0.34$$
$$F_e \approx 0.87$$

These results closely resemble those for the two-stage amplifier; for both two and three m.f. stages we may say,

(i) The effective feedback begins to fall at a frequency about F times lower than the knee.

(ii) The effective feedback factor is about unity (no feedback) at the knee.

7. Low-Frequency Response

In the simple case when the frequency characteristic, at low frequencies, is entirely controlled by the grid-circuit coupling capacitance C_g and resistance R_g, all the arguments developed above for the high-frequency region apply unchanged, provided the frequency variable is given the appropriate meaning; i.e., $x = 1/\omega C_g R_g$. In most practical amplifiers the response at low frequencies is also affected by the time constants of screen, cathode and anode decoupling and by-pass circuits, as well as by unwanted coupling through the h.t. supply system.

8. Band-Pass Amplifier

A band-pass network can be constructed from any low-pass network by tuning all the reactors of the low-pass network to the desired mid-frequency of the band-pass network, tuning the inductors of the low-pass network to series resonance and the capacitors of the low-pass network to parallel resonance.

A well-known network theorem states that, when this has been done, the bandwidth of the band-pass network, at any specified response-level relative to that at the mid-frequency, is the same as that of the low-pass network at the same response level relative to zero frequency.

In the amplifiers considered above, this change to a band-pass characteristic is obtained simply by adding an inductor, in parallel with the coupling resistance, to resonate with the circuit capacitance at the desired mid-frequency.

All the arguments developed for RC-coupled amplifiers thus apply equally to tuned-anode amplifiers. The 'staggering' parameters have the same meaning (ratio of bandwidth); the various frequency-response curves apply as they stand, provided it is remembered that the frequency-variable now refers to the total bandwidth for any specified response level.

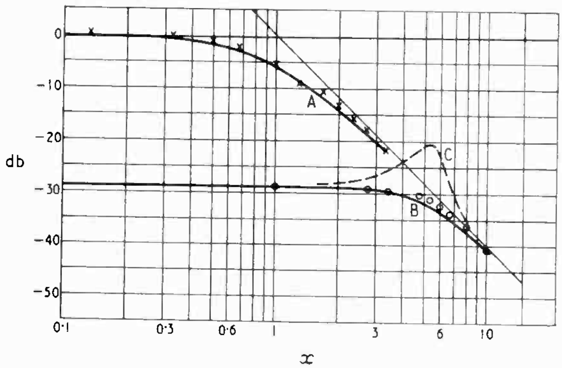

Fig. 6. *Two-stage amplifier. Curves A and B show the calculated performance, A: for identical stages and no feedback and, B, for m.f. stagger and 29-db feedback respectively. The points marked are the measured values of an experimental amplifier. Curve C shows the measured results with 29-db feedback and no stagger.*

9. Experimental Verification

A simple verification of the m.f. conditions was made with an amplifier having the circuit of Fig. 5. To avoid uncertainty as to the value of the circuit capacitances, capacitors of the order of 0.001 μF were connected across the anode loads. To avoid any change in d.c. conditions, staggering was carried out by changing the values of these capacitors, the asymptote being retained by keeping the product of their values constant. A three-stage amplifier was simulated by adding a third RC circuit in the intervalve coupling.

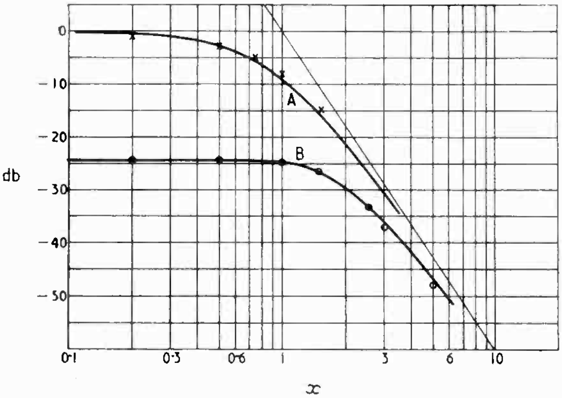

Fig. 7. *Three-stage amplifier. Curve A is for no feedback and curve B for 24.5-db feedback and m.f. stagger. The curves are calculated and the points experimental.*

Ordinary commercial components with tolerances of ±10 to 20% were used. Fig. 6 is for the two-stage case, the upper curve being for two identical stages with no feedback and the lower full line curve for two stages m.f. staggered with 29-db feedback. The curves are theoretical and the points experimental. The agreement is sufficiently good in the 'no feedback' case. The points lie above the m.f. curve at the knee; this is probably because the staggering was not quite accurate. The dotted curve is experimental, for 29-db feedback applied without staggering the amplifier.

The results for the three-stage case are given in Fig. 7. The upper curve is for the amplifier with no feedback; the lower for 24.5-db feedback with m.f. staggering. The curves are theoretical and the points experimental. It is not possible in this case to give a curve for the original amplifier with feedback applied but no staggering, since 24.5 db considerably exceeds the feedback necessary to produce self-oscillation in this case.

It is considered that these experimental results give sufficient verification of the theory.

10. Acknowledgment

The writer is indebted to Marconi Instruments Ltd. for permission to publish this paper and to R. V. Greenham, who carried out the experimental work described in Section 9.

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THEORY OF ERROR DISTRIBUTION

Application to Radio Navigational Aids

By P. F. Duncan

THE accuracy of measurements made in the observation of experiments in wireless engineering is not normally of a sufficiently high order to warrant the application of the theory of errors. Indeed, in the more general employment of radio, there is often an initial uncertainty as to the accuracy of the measuring instruments available; to such an extent, in fact, that there would be little point in attempting to refine the results by statistical methods.

When we turn to the application of wireless to navigation aids, however, the matter is seen in a quite different light. We find that when endeavours are made to assess the true accuracy of low-frequency c.w. systems such as Consol and Decca or of pulse systems like Gee and Loran (in all of which the desired fix data are derived from the intersection of two independent position lines) recourse must be made to a more precise method. This is provided by the theory of errors of observation.

Much of the relevant technique of the theory has already been given[1,2] but it would nevertheless appear worth while to present the basic principles and some of the more important formulae in a simple and concise form suitable for those who may be concerned with determining the accuracy of the type of radio aids to navigation already mentioned. It is accordingly proposed to show how the theory of normal distribution of errors may be applied to two dimensional problems such as the degree of accuracy of a fix dependent upon the intersection of two position lines.

MS accepted by the Editor, November 1947

1. Uncertainty of Measurements

Whenever measuring instruments are used to obtain the magnitude of a quantity, it is rarely possible to secure more than an approximation; the true value evades us. The reason for this, as far as practical applications go, is the fallibility of the operator (the human error) and the limitations of the measuring instrument.

More profoundly, there is also the consequence of Heisenberg and Bohr's epistemological principle of indeterminacy. This is a principle based on the inadequacy of spatio-temporal relations which implies that any physical determination whatsoever involves a reaction between the thing to be measured and the agent attempting to measure it—uncertainty in the result is unavoidable.

Despite allowance for constant instrumental inaccuracies (systematic errors), it is inevitable that uncertainty will exist in all measurements made in connection with radio engineering because, even if the principle of indeterminacy be ignored, there remains the human error. Incidentally, it is not without interest to note that the very term 'accuracy' tends to evade definition; it is indeed a comparative term and no more.

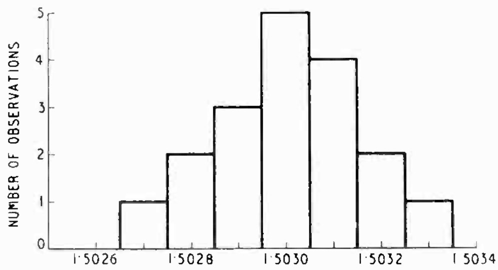

Fig. 1. *The stepped curve.*

2. Example of Critical Measurement

It will be found helpful if we first consider, as an example of a critical measurement, the form of error distribution which would be encountered in so elementary an operation as determining the diameter of a steel bar. If such a task is to be attempted with the precision commonly required for engineering purposes, the measuring instrument may, in all probability, be a travelling microscope, the scale of which we can suppose graduated in thousandths of an inch. A number of measurements—say, eighteen or twenty—would be taken and a graphical analysis of the results made.

Let us assume that all the readings fall between 1.5027 in and 1.5033 in. Taking as ordinates the number of readings between 1.50265—1.50275 in, 1.50275—1.50285 in, etc., and plotting against the appropriate intervals on the microscope scale, the result will be a stepped curve as in Fig. 1.

If, however, a greater number of readings is taken, and at the same time the intervals are made smaller by a finer subdivision of the microscope scale, then the curve will lose its step-like character and, in the limit, will become a smooth curve, as in Fig. 2.

Neglecting the systematic error represented by imperfect graduation of the microscope scale, the errors disclosed in the curves are the result of incorrect alignment of the cross-hairs of the microscope and the edge of the steel bar or of incorrect estimation of the fractional part of a scale division. These (human) errors may equally well be positive or negative.

A more significant result of increasing the readings and decreasing the intervals is that there will be a smaller number of observations which will differ greatly from the true value. The smooth curve therefore tends to become symmetrical about its maximum, and this maximum will represent the true value as distinct from the most probable value. (The most probable value of quantity is the arithmetic mean of the observed values, and from the preceding statement it will be noted that the most probable value approaches the true value as the number of observations is increased.)

If the true value as determined from the maximum of the smooth curve is subtracted from the values shown on the abscissa of Fig. 2, the curve can equally well be depicted as in Fig. 3, namely, number of observations v. error in inches.

3. Normal Law of Error

An error curve in the form of Fig. 3 is common to many types of measurement, and may be represented by the Gaussian or 'normal' law of error:

$$\delta N = \frac{Nh}{\sqrt{\pi}} \exp\left(-h^2x^2\right) \delta x \qquad \ldots \quad \ldots \quad (1)$$

where δN is the number of readings having errors between x and $x + \delta x$, N is the total number of readings (which must be very large) and h is a constant known as the precision constant.

In practice this law will not prevail absolutely because it implies the possibility of errors extending from plus infinity to minus infinity. Whereas in measuring the steel bar, for instance, errors will not exceed, say, one inch because of the finite travel of the microscope lead screw. Nevertheless, in many instances the law is followed quite closely for errors which are not too great; it is sufficiently accurate for all practical purposes.

Several terms in common use in the theory of errors will now be defined and their expression in terms of the precision constant h will be given.

Probability.—The probability that an observation will have an error between x and $x + \delta x$. This, by definition above, is $\delta N/N$.

Mean Error.—The mean of all errors without regard to sign. Also known as the Mean Deviation. Since, for an ideal distribution of errors, the curve is symmetrical, mean error is also the mean of the positive and the mean of the negative errors. From equation (1), the sum of the errors between x and $x + \delta x$ is sensibly equal to $\frac{Nh}{\sqrt{\pi}} \exp(-h^2x^2)x\,\delta x$ so that the sum of all errors of positive sign is $\frac{Nh}{\sqrt{\pi}}\int_0^\infty \exp(-h^2x^2)x\,dx$ or $N/(2h\sqrt{\pi})$. Similarly, the number of errors having positive sign is $N/2$ and hence the mean positive errors is $1/hN$, while the mean negative error is also $1/hN$. Therefore

$$\text{Mean error} = 1/(h\sqrt{\pi}) = 0.5642/h \quad \ldots \quad (2)$$

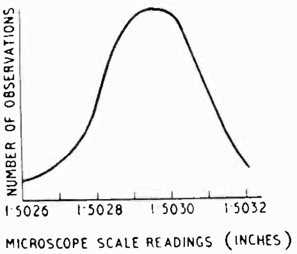

Fig. 2. *Smooth curve.*

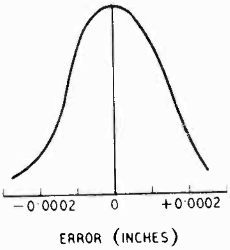

Fig. 3. *Error curve.*

4. Error of Mean Square

The square root of the mean of the squares of the errors. Also known as the Standard Deviation. By a process similar to that employed in deriving the value of mean error, it may be shown that

$$\text{Error of Mean Square} = 1/(h\sqrt{2}) = 0.7071/h \quad \ldots \quad (3)$$

Probable Error.—The probable error is the value of x for which, if positive, the number of greater positive errors will be equal to the number of lesser positive errors; and, if negative, the number of greater negative errors will be equal to the number of lesser negative errors. Knowing the value of x for which

$\frac{h}{\sqrt{\pi}}\int_0^x \exp(-h^2x^2)\,dx$ is $1/4$, then

$$\text{Probable Error} = 0.4769/h \quad \ldots \quad (4)$$

5. Practical Applications

In what has been discussed so far, it has been assumed that (ignoring systematic errors) the mean of the readings is the true value. For a limited number of observations, however, this is not the case and the formulae given above for error of mean square and probable error require modification.

In regard to mean error, it may be found in practice that there is a slight difference between the sum of the positive and the sum of the negative errors. This being so, the mean of the two sums will give the mean error.

In practice, the error of mean square should be calculated from the modified expression

$$\text{Error of Mean Square} = \sqrt{\frac{\text{Sum of squares of errors referred to mean}}{N - 1}} \quad \ldots \quad (5)$$

and the probable error should be calculated from the expression

$$\text{Probable Error} = 0.6745 \times \sqrt{\frac{\text{Sum of squares of errors referred to mean}}{N - 1}} \quad \ldots \quad (6)$$

It is obvious in connection with each of these modified expressions that for a reasonably large number of observations there is little significant difference between a denominator N and a denominator $N - 1$, the former being strictly applicable to the limiting case where N is infinitely large.

Reverting to probable error, it must be remembered that the mean of a finite number of observations is, by the very nature of the limitation, itself subject to error. It is, in fact, this probable error of the mean in which we are interested. This is given by tne probable error of a single observation divided by the square root of the number of observations, and its significance lies in the fact that if a great many means are obtained in successive experiments, nearly one-half will differ from the overall mean by an amount greater than the probable error. Hence

$$\text{Probable Error of the Mean} = 0.6745 \times \sqrt{\frac{\text{Sum of squares of errors referred to mean}}{N(N - 1)}} \quad \ldots \quad (7)$$

The computation of the mean of a number of observations and the calculation of the three precision indices (mean error, error of mean square and probable error) is generally a tedious process. A considerable saving of time can be effected if an approximate mean is taken, as follows.

If a series of observations produces numerical values for $X_1\ X_2 \ldots X_N$, then the true mean is $\mathbf{X}$, and approximation $\mathbf{X}'$ is chosen by inspection. It can be shown that $\mathbf{X}$ is given by

$$\mathbf{X} = \mathbf{X}' + \frac{\Sigma(X - \mathbf{X}')}{N} \quad \ldots \quad (8)$$

and the error of mean square (σ) by

$$\sigma^2 = \frac{\Sigma(X - X')^2}{N} - \left(\frac{\Sigma(X - \mathbf{X}')}{N}\right)^2 \quad \cdots \quad (9)$$

Reference to equation (6) shows that the probable error is related to σ by the expression

Probable Error = 0.6745 .. .. (10)

The quantities $X - \mathbf{X}'$ are relatively small—hence the simplified computation.

To decide whether an observed error distribution follows the normal law, for practical purposes it is generally sufficient to calculate the precision constant from the formulæ given above and to superimpose the normal curve on the curve of observed error distribution. The degree of correspondence then becomes a matter of personal opinion.

6. Application to Radio Navigation Aids

We now turn to the application of the theory of distribution of errors to radio aids to navigation in which the position of the aircraft, ship or vehicle is to be determined from the intersection of two position lines and in which the accuracy of the resulting fix is subject to doubt because the position lines are themselves subject to error.

A non-statistical method which is sometimes adopted in order to assess the error of fix consists of drawing position lines as determined from the radio system employed and then adding limits for some assumed maximum error. The result is an arbitrary quadrilateral figure which is taken as defining the error of fix.

It is unlikely, however, that any radio-aid system exists in which the error of position line can be said to have definite limits; the quadrilateral of error, as defined above, can have only an artificial meaning. Instead, errors of position lines are subject to a greater or lesser extent to the normal law of errors. Practical analyses of the Gee and Loran systems have already indicated that they follow the normal law to quite a close extent. The statistical method of error assessment will therefore now be considered.

For simplicity, let us assume that the errors of a d.f. system are subject to the normal law and that the bearing error at a point P from one transmitter is θ_1 while that from a second transmitter is θ_2. The probabilities of each error are then given by

$$\frac{h}{\sqrt{\pi}}\exp\left(-h^2\theta_1^2\right) d\theta_1 \text{ and } \frac{k}{\sqrt{\pi}}\exp\left(-k^2\theta_2^2\right) d\theta_2$$

where h and k are the precision constants for the two transmissions and may, of course, be the same. The probability that errors θ_1 and θ_2 occur simultaneously is given by the product

$$\frac{hk}{\sqrt{\pi}}\exp\left(-h^2\theta_1^2 - h^2\theta_2^2\right) d\theta_1 d\theta_2 \quad \cdots \quad \cdots \quad (11)$$

With the point P as origin, setting up local Cartesian co-ordinates x and y enables it to be shown that expression (11) can be re-written

$$\frac{\sqrt{AB - H^2}}{\pi}\exp\left(-(Ax^2 + 2Hxy + By^2)\right)\delta x\, \delta y \quad \cdots \quad \cdots \quad (12)$$

where A, B and H are constants for the given position P and are expressed in terms of the bearing of P with respect to the two transmitters and the distance between them.

If a large number of observations are taken, it may be shown that $1/n$ of the observations lie outside an ellipse defined by

$$Ax^2 + 2Hxy + By^2 = \log_e n \quad \cdots \quad \cdots \quad (13)$$

The 50 per cent ellipse, given by substituting $n = 2$, is appropriate and accurate as a measure of fix accuracy and is such that of a large number of observations 50 per cent will lie within and 50 per cent without the ellipse.

Curves of constant-fix accuracy may be derived from the data secured in this way and expressed in terms of constant area of the 50 per cent ellipse or in terms of the ratio of the major to the minor axis of the ellipse.

Although these results may not appear very convincing at first sight, it may be stated that trials have shown that of some hundred observations taken, fifty per cent did in fact lie inside and fifty per cent outside the error ellipse calculated for the small area in which the trials were conducted. It may be added that the results of such analysis played an important part in subsequent operations depending on the use of the radio navigational aid in question. In the future it will no doubt be found that there is greater recourse to the theory of error distribution in assessing the accuracy of certain types of radio aids to navigation.

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EDITORIAL ADVISORY BOARD

On his appointment to the International Radio Consultative Committee as vice-director for broadcasting, L. W. Hayes, O.B.E., M.I.E.E., has resigned his position as head of the B.B.C. Overseas Engineering Information Department. He has also resigned from the *Wireless Engineer* Editorial Advisory Board, to which P. A. T. Bevan, B.Sc., A.M.I.E.E., of the B.B.C. Engineering Division, has been appointed in his place.

NEW YEAR HONOURS

A. J. Gill, B.Sc.(Eng.), has been created a Knight Bachelor. Since 1947 he has been engineer-in-chief of the G.P.O. and was recently appointed a vice-president of the I.E.E.

R. T. B. Wynn, M.A., has been appointed a C.B.E. He is assistant chief engineer of the B.B.C. and a vice-chairman of the Radio Section of the I.E.E. for the present session.

TRIODE INTERELECTRODE CAPACITANCES

By E. E. Zepler, Ph.D., M.Brit.I.R.E., and J. Hekner

(Communication from Electronics Dept., University College, Southampton.)

SUMMARY. Changes in the grid-cathode and grid-anode capacitances of triode valves, as functions of the working conditions, were measured. The influence of the mutual conductance, the amplification factor and the supply voltages are shown. A theory is given which is in fair agreement with the experimental results.

1. Introduction

INVESTIGATIONS of the frequency stability of valve oscillators have to take into account changes of the interelectrode capacitances with changing voltages at anode and grid. By using limiting devices and carefully choosing the feedback factor the amplitude of oscillation can be kept small, so that the working conditions vary little over a cycle. Further by using stabilized supply voltages, changes in interelectrode capacitances can be kept low, so that at present the temperature constancy of the tuned circuits is the factor of first importance. It is, however, to be expected that with improved circuit components increased attention may have to be paid to the valve capacitances.

Measurements of interelectrode capacitances have been carried out at radio frequencies with various valves under different working conditions. Results of the measurements of one previous paper have been largely confirmed[1], while those of another fairly recent paper differ widely from the results given here.[2] * Inter-relations between the changes of the grid-cathode capacitance c_{gk} and the grid-anode capacitance c_{ga} have been studied, and a theory is suggested to explain these interrelations.

2. Apparatus and Method of Measurement

The type of bridge used by Jones was adopted. Certain modifications made it possible to measure c_{gk} and c_{ga} separately if desired, as will be shown later. The two resistances R_1 and R_2 in Fig. 1 were of equal magnitude; the variable capacitor C_2 was calibrated, and equilibrium of the bridge was obtained by adjustment of C_2. Changes in the valve capacitances, under the influence of varying working conditions, were found by the changes in C_2 required to restore the equilibrium. The resistances R_1 and R_2 were chosen as small as possible and the junction of the two resistances was earthed. Thus earthed capacitances of the batteries are rendered harmless, because these earth capacitances are either across the detecting branch or across the very small resistances. It was found that capacitances up to 100 pF could be connected across either R_1 or R_2 without any noticeable effect on the balance of the bridge. A further advantage of using low resistances will be seen when the measurements of c_{gk} and c_{ga} are discussed. Small loops in conjunction with movable iron-dust cores, connected in series with R_1 and R_2, served to adjust the phase; they are not shown in the drawings. Once the phase was adjusted for one of the three capacitance measurements, a further adjustment in the course of the particular measurement was not required. The measuring frequency was usually 0.5 Mc/s, occasionally 1 Mc/s and 1.5 Mc/s were used to verify certain assumptions (see later) and to test the reliability of the bridge.

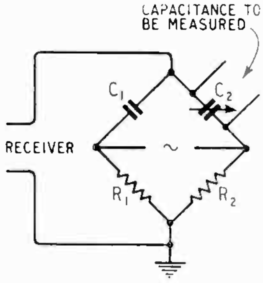

Fig. 1. *Simple de Sauty bridge.*

The capacitances C_1 and C_2 were approximately 15 pF when the cathodes of the valves were cold. The r.f. signal was derived from a signal generator through a carefully screened transformer using an earthed electrostatic screen between primary and secondary. To improve the signal-to-noise ratio in the detecting receiver, a tuned circuit which was matched to the receiver input was connected across the detecting branch.

Changes in c_{gk}, c_{ga} and $(c_{gk} + c_{ga})$ were measured. The values for c_{ga}, deduced from the measurements of c_{gk} and $(c_{gk} + c_{ga})$, agreed with those measured directly within 20%. When

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* The method adopted in the latter paper is quite different from that used by Jones and by the authors of this paper. From the description given very severe damping of the oscillator circuit must occur, particularly in the circuit of Fig. 3(b), and it cannot be seen where the influence of this on the frequency of oscillation has been taken into account. In addition, Miller effect might easily cause capacitance changes of the order of those measured.

it is considered that with the MH4 valve the change in c_{ga} is only about 2% of that in c_{gk}, the results give a good indication of the accuracy obtained. Changes of the order of a few per cent in capacitance were apt to take place within about half an hour, therefore corresponding measurements of c_{gk} and $(c_{gk} + c_{ga})$ had to be carried out within a few minutes. This was made possible with the help of a multi-point switch. The cathodes were heated for 20 minutes before taking measurements.

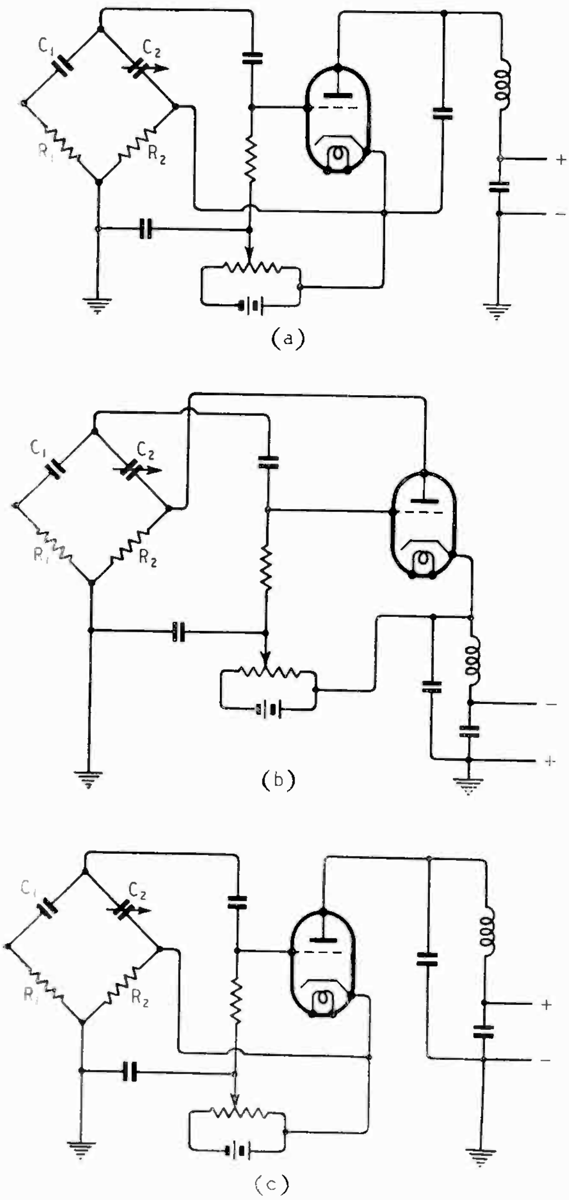


Fig. 2. The connections for the measurement of $(c_{gk} + c_{ga})$; c_{ga} and c_{gk} are shown at (a), (b) and (c) respectively.

Fig. 2(a) shows the circuit used for measuring $(c_{gk} + c_{ga})$. The anode is connected to the cathode through a 1-μF capacitor so that neither the anode impedance nor the anode-cathode capacitance of the valves can enter into the balance of the bridge. R_1 and R_2 were both either 1 or 3 ohms. The results were the same in both cases.

Fig. 2(b) gives the circuit used for measuring c_{ga}. In this case the anode impedance and the anode capacitance of the valve are in parallel with R_2. It is easy to see, however, that this effect can be neglected because of the small value of R_2.

c_{gk} was measured with a circuit shown in Fig. 2(c). In this case a small correction of the result is required, if it is to serve as a check on the two preceding measurements. A simple calculation shows that the anode current of the valve causes the p.d. across R_2 to fall by the fractional amount $\frac{\mu + 1}{2}\frac{R_2}{r_a} \approx \frac{1}{2} g_m R_2$

where r_a is the anode impedance and g_m the mutual conductance of the valve. To restore balance, C_2 must be increased by $\delta C_2 = C_2 R_2 g_m$; therefore, to obtain the correct result for the increase in c_{gk} the value $C_2 R_2 g_m$ must be added to the capacitance change measured. C_2 naturally refers to the total capacitance of the arm in the final position of balance. The values obtained for δc_{gk}, with R_1 and R_2 equal to 1 or 3 ohms, differed by the calculated amount, and the correct δc_{gk} was derived by extrapolation. To avoid grid current the grid bias was always negative.

3. Results

As in Jones's paper, c_{ga} was found to decrease when the valve became conducting, while c_{gk} always increased; the change in c_{gk} was considerably larger than that in c_{ga}. Three types of valves of widely differing parameters and under varying working conditions were used. In the graphs the zero point of reference for the change of capacitance is with the valve working with a hot cathode, but heavily biased far beyond cut-off. It was found that under these conditions the capacitances were independent of the individual values of grid and anode voltage. The main results to be seen from Figs. 3, 4 and 5 are as follows:

1. c_{gk} increases with increasing anode current, but the rate of change eventually becomes very small.

2. c_{ga} decreases with increasing anode current. Eventually the rate of change becomes very small and even tends to become positive.

3. For equal anode currents the change is largest with low anode voltage.

4. The change in $(c_{gk} + c_{ga})$ is approximately μ times as large as the change in c_{ga}.

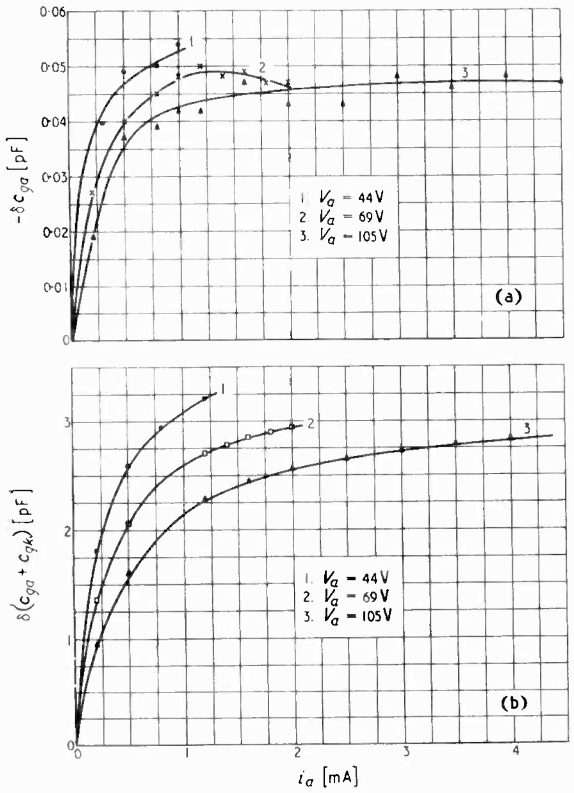


Fig. 3. *Variations of δc_{ga} and $\delta(c_{ga} + c_{gk})$ with anode current for the Marconi and Osram MH4 valve.*

In order to see the connection between the changes in the capacitances and the mutual conductances of the valves, the changes are plotted in Figs. 6, 7 and 8 as a function of g_m. At first the changes are almost proportional to g_m, but with larger anode currents this proportionality no longer holds. For given values of g_m, the changes are always largest for low anode voltages.

As will be explained later, a close relationship was expected to exist between the ratio $\frac{\delta(c_{gk} + c_{ga})}{\delta c_{ga}}$ and the amplification factor of the valves. For this reason the amplification factor μ was measured by a bridge method under the various working conditions and is plotted, together with $\frac{\delta(c_{gk} + c_{ga})}{\delta c_{ga}}$, in Figs. 9, 10 and 11 as a function of i_a. The agreement, though not exact, between the ratio and μ is too striking for this to be considered a pure coincidence. However the deviations in the cases of the MH4 and ML4, particularly with the larger anode currents, cannot be dismissed as errors in measurement in view of the fact that the ratio $\frac{\delta(c_{gk} + c_{ga})}{\delta c_{ga}}$ is always found to be larger than μ.

It was thought that a slight expansion of the electrodes caused by the heating effect of the anode current might account for the discrepancies. For this reason c_{ga} was first measured with a small anode current. Then a heavy anode current was allowed to flow for several minutes after which the valve was brought back instantaneously to the previous low value of anode current and measured again. The second measurement showed however, the same value of capacitance as before, so that the possibility of a thermal effect had to be dismissed.

However, there seems to be another effect of the order of magnitude required to remove largely the discrepancies in Figs. 9, 10 and 11. Changes in $(c_{gk} + c_{ga})$ were carefully measured in the neighbourhood of the cut-off point and the

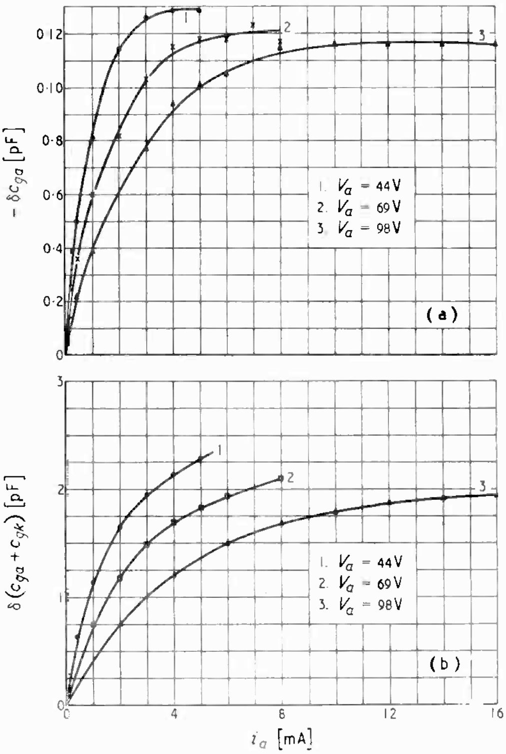


Fig. 4. *Variations of δc_{ga} and $\delta(c_{ga} + c_{gk})$ with anode current for the Marconi and Osram ML4 valve.*

results are given in Fig. 12. It will be seen that $(c_{gk} + c_{ga})$ increases when one approaches the cut-off point of the valve, before anode current flows. The effect appears to be caused by a movement of the space charge as a whole. In the following we neglect interaction between electrons and consider plane parallel electrodes. We assume that an electron leaves the cathode with the velocity v and that at the cathode there exists a constant field E tending to drive the electron back. Then the distance s which the electron travels before returning is found from the simple relation $\frac{1}{2} mv^2 = Ees$, where m is the mass and e the charge of an electron. The average distance s' of the electron during the time of its flight is found by a simple integration to be

$$s' = \frac{mv^2}{3Ee}.$$

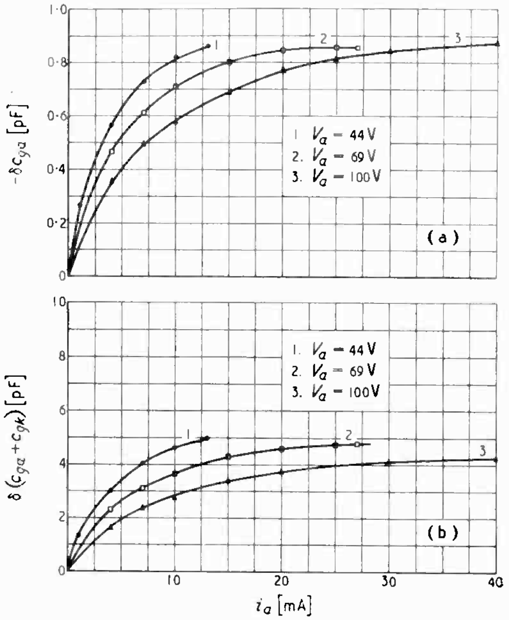


Fig. 5. *Variations of* δc_{ga} *and* $\delta(c_{ga} + c_{gk})$ *with anode current for the Marconi and Osram PX4 valve.*

If now an alternating voltage is applied between grid and cathode the average distance of the electrons forming the space charge from the cathode varies with the applied voltage and the current caused by this variation is $\Sigma e \frac{ds'}{dt}$, the sum to be taken over all the electrons leaving the cathode. In the above case $E = E_1 + E_2 \sin \omega t$, where $E_2 \ll E_1$. Hence the current caused by the movement of the space charge is proportional to $\frac{E_2 \omega \cos \omega t}{E_1^2}$; i.e., it is in the nature of a capacitive current. With the PX4 the increase in $(c_{gk} + c_{ga})$ was negligibly small and is therefore not shown in the graph. A simple explanation is that the PX4 is a directly-heated valve, in contrast to the ML4 and MH4. Because of the voltage drop across the filament, the emission starts at the negative end. Hence, when anode current begins to flow, the larger part of the cathode is well beyond cut-off and the effect due to the movement of space charge is still very small. As can be seen from Fig. 11, the agreement between μ and the ratio $\frac{\delta(c_{gk} + c_{ga})}{\delta c_{ga}}$ is fairly good in the case of the PX4, so that a correction factor is hardly required.

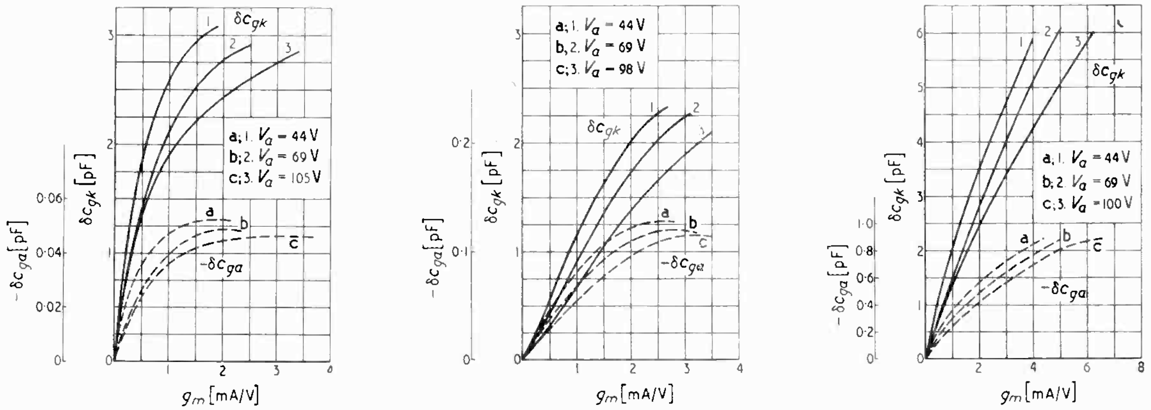
When measuring c_{ga} near the cut-off point it was found that the capacitance was in no way affected until actual emission took place. This is not surprising since the movement of the space charge constitutes a current between cathode and anode which is not recorded by the bridge. Therefore, in comparing the changes in $(c_{gk} + c_{ga})$ and in c_{ga} due to the effect of conduction current alone, one has to deduct from $\delta(c_{gk} + c_{ga})$ the change caused by the movement of the space charge. It is seen that the agreement with μ is then much better than is indicated in Figs. 9 and 10.

4. Theoretical Considerations

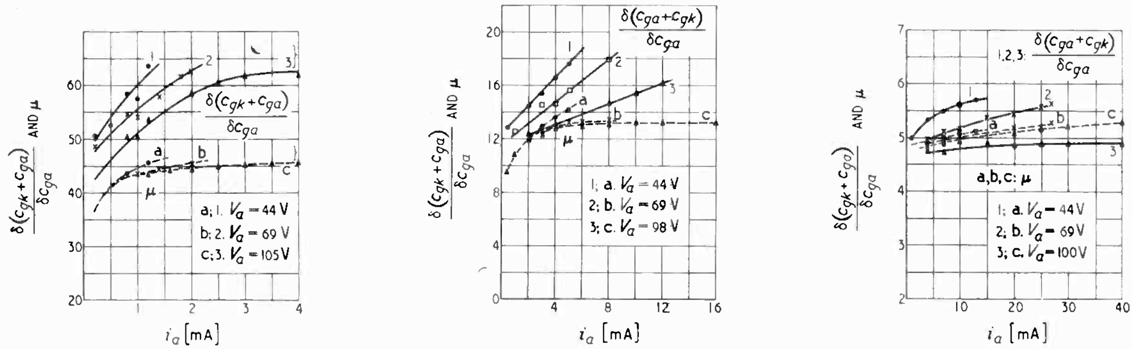
As an explanation of the above results, a simple theory is put forward which seems, qualitatively and quantitatively, to meet the facts. Assume that a plate capacitor is charged to a potential difference V by an electric charge of magnitude q; then the ratio q/V is equal to the capacitance between the two plates, if other conducting surfaces are not taken into consideration. If now a negative charge is brought near to the outer surface of the negatively charged plate the potential difference between the two plates is increased, so that the measurement of q and V shows an apparent decrease of the capacitance. If the negative charge is brought near to the outer surface of the positively charged plate an apparent increase of capacitance is experienced, while additional charge between the plates leaves the capacitance substantially unchanged.

The first of these cases occurs when the grid-anode capacitance is measured [Fig. 2 (b)]. At balance of the bridge, grid and cathode are at the same alternating potential, and the effect of the applied alternating voltage is merely to alter the potential of the anode with respect to the cathode. When, for example, the alternating potential of the grid is positive with respect to the anode the emission is lowered, since in effect the potential of the anode is lowered with respect to the cathode. The reduced density of the electron stream between grid and cathode affects mainly the charge on the grid and only little that on the anode. Thus, for a given applied voltage between grid and anode the charging current is reduced and a decrease in capacitance is measured. The electrons between grid and anode are likely

In the circuit shown in Fig. 2(a) the variations of anode current are μ times those in the circuit of Fig. 2(b), because in Fig. 2(a) the alternating potential is between grid and cathode of the valve, whereas in Fig. 2(b) it is between anode and cathode. This fact suggested the idea that the change in $(c_{gk} + c_{ga})$ might turn out to be μ times the change in c_{ga}. In contrast to the circuit of Fig. 2(b), the electron density increases in the circuit of Fig. 2(a) when the potential of the grid is positive with respect to the anode (and, of course, to the cathode). Although the change in electron density which causes the increase in capacitance takes place between the electrodes concerned, this case is different from the general



Figs. 6–8. *Variations of δc_{gk} and δc_{ga} with mutual conductance g_m; Fig. 6, for the MH4 valve; Fig. 7 for the ML4 valve; Fig. 8, for the PX4 valve.*



Figs. 9–11. *Changes of $\delta(c_{gk} + c_{ga})/\delta c_{ga}$ and μ with anode current; Fig. 9 for the MH4; Fig. 10, for the ML4; Fig. 11, for the PX4.*

to have little effect, partly because their density is smaller, partly because they are between the two electrodes concerned. When the measuring frequency was varied between 0.5 and 1.5 Mc/s the results were unchanged. This shows that the change in dielectric constant due to the existence of free electrons is not the cause of the measured decrease in c_{ga}.

one discussed above, because here one electrode is the source of the electrons. If, for example, the grid voltage is raised with respect to the cathode, the increased number of electrons surrounding the grid will attract an additional positive charge to the grid, but they will not, as would happen in the general case, attract a corresponding positive charge to the cathode, since the effect of the altered grid potential consists in transferring electrons from the space charge into the space near to the grid. Actually the experimental results seem to indicate that, in the case of $(c_{gk} + c_{ga})$, the electrons between

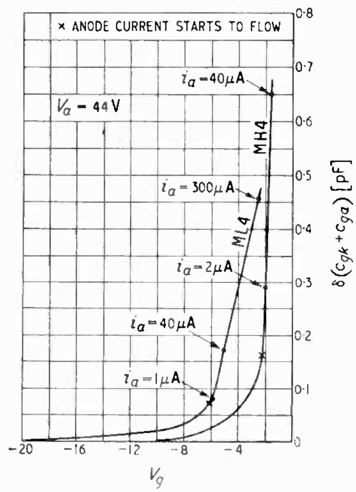


Fig. 12. *Variation of $\delta(c_{gk} + c_{ga})$ with grid bias for the MH4 and ML4 valves.*

grid and cathode have hardly any effect on the cathode; otherwise it would be difficult to understand why the change in $(c_{gk} + c_{ga})$ is μ times that of c_{ga}. The electrons between grid and anode should have little effect in either of the two cases, for reasons given above.

It seems plausible that the changes in capacitance, as shown by the results, are not proportional to g_m, because primarily it is not the change in anode current but the change in electron density at the grid which produces the changes in capacitance. For the same values of g_m the changes in capacitance may be expected to be largest when the velocity of the electrons on passing the grid is smallest. The question may be raised whether variations in the electron density at the grid with an applied alternating voltage could decrease while g_m increases. Such a possibility seems indicated by the results shown in Fig. 6.

Another factor of some importance may be the path taken by the electrons. When the grid is very negative the main sources of emission are those parts of the cathode not shaded by the grid. Most electrons will, therefore, pass the grid well away from the grid wires, and their effect is smaller than might be the case with low bias. This is in accordance with the measurements which show, for a given g_m, the least change of capacitance when the grid bias is large and the anode voltage correspondingly high. The measurements of μ, under different working conditions, also suggest such an effect, μ being smallest with large anode voltage.

Variations in c_{ga} appear to be a better indication of the effect of conduction current than variations in c_{gk}, because c_{ga} does not react to the apparent enlargement of the cathode. A means of separating the effects of the conduction current and of the enlargement of the cathode has not yet been found. It is tempting to take the changes in c_{ga} as the starting point and to obtain the effect of the movement of the space charge on c_{gk} by deducting $\mu(\delta c_{ga})$ from the values found for $\delta(c_{gk} + c_{ga})$. But the basis for such a procedure cannot be considered sufficiently established.

From the theory put forward one may expect the changes in the anode-cathode capacitance to be positive and of a similar magnitude to those of c_{ga}. Such measurements are likely to meet with great difficulty because of the valve impedance being in parallel with the capacitance to be measured. For this reason these measurements seem hardly worth the trouble.

Further experiments which are in progress will include the measurement of the input impedance at very high frequencies which, it is hoped, will give an indication of the relative transit times between cathode and grid under various working conditions.

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[2] S. C. Mitra and S. R. Khastgir. "Inter-electrode Capacitances of Triode Valves and their Dependence on the Operating Conditions." *Indian Journal of Physics*, June 1946.

INDEX TO ABSTRACTS

Orders can now be accepted for the *Wireless Engineer* Index to Abstracts and References 1948. It is priced at 2s. 8d. (including postage) and will shortly be available. It includes a list of most of the journals abstracted together with the addresses of their Publishers or Editorial Offices, and the form of abbreviation of their titles used in the Abstracts.

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TEST GEAR EXHIBITION

Marconi Instruments Ltd., St. Albans, Herts, are exhibiting a range of communications test equipment in their London Showrooms, 109, Eaton Sq., S.W.1.

The exhibition will be open from 21st February until 4th March and will include demonstrations. Applications for tickets should be made to the Company.

FREQUENCY STABILIZATION OF V.M. VALVES

By H. Borg, Lic. ès Sc., Ing. E.S.E., A.M.I.E.E.

(Admiralty Signal and Radar Establishment)

1. Introduction

HIGH stability of microwave oscillators is an essential requirement in apparatus making use of the high-Q circuits used in centimetre-wave work. Communication systems with better stability improve efficiency by restricting the frequency band; if the production of extremely accurate and stable carrier frequencies becomes a simple matter, the possibility of several hundred channels in a given frequency band can be realized. It is obvious that special applications requiring comparison of radio frequencies or phases such as f.m. or Doppler radar presupposes high stability of the sources; in particular, random frequency changes within the pulse duration in such systems may become of primary importance. Very stable frequencies are also required for laboratory measuring apparatus in the u.h.f. and s.h.f. regions (e.g., measurement of the Q of a high-Q cavity), and in the study of molecular absorption of microwaves by gases.

The object of these notes is to survey the methods and some of the problems in connection with frequency controlled oscillators; attention will be largely concentrated on the principles rather than on the details. Investigation is particularly concerned with c.w. oscillators in the microwave region. A description and a discussion of a new method of stabilizing by electronic means the frequency of a v.m. valve at 9,360 Mc/s by comparison with a crystal is included. A short term stability of the order of $\pm$ 100 c/s, relative to the crystal, has been achieved by means of a double discriminator.

2. Methods of Stabilization

If $s(t) = A \cos \phi(t)$ represents a signal of oscillatory character, the angular frequency of the oscillation is, in fact, the instantaneous frequency, defined by the derivative of the argument $\phi(t)$ with respect to time t. As very few of the so-called fixed parameters are physically independent of time, the frequency of an oscillator is not constant over a given interval of time.

MS accepted by the Editor, October 1947

The production of electromagnetic waves involves in principle—

(a) A source of energy of which the magnitude and rate of flow is not always constant.

(b) A circuit or resonator excited and maintained in oscillation by the action of this energy. The physical characteristics of the circuit can be assumed to be constant only as a first approximation.

(c) An external load or sink of energy, subject also to variations and coupled to the oscillator with the result that any modification in the properties of the sink system will be reflected back in the source; in theory, variations in radiation and propagation would affect the oscillator.

Absolute stability does not exist—the term 'stable frequency source', referred to in these notes, applies to sources which are nearly stable; to fix ideas let us say it applies to sources with an overall stability better than 1 part in 10^7. Various methods may be used, separately or simultaneously, to reduce the inherent instability of frequency sources; they can be classified as follows.

2.1. *General Methods*

General methods are well known; they involve the regulation of the power supplies, care in the design, the use of materials of low temperature coefficients of expansion (quartz, invar), temperature control (thermostat) and temperature-compensation of the circuits and their components by the use of two materials of different thermal coefficients of expansion, special valves or circuits (e.g., Franklin oscillator), control of the reactive component of a tunable load for oscillators having their frequency especially conditioned by the external load circuit (this method has been successfully employed in the stabilization of magnetrons).

2.2. *Crystal Control*

Mechanical vibrating systems have in general a high Q and, therefore, a better frequency stability than electrical systems. Tuning forks and magneto-striction oscillators have been entirely superseded by piezo-electric crystal oscillators. Cuts at various oblique angles with respect to the crystallographic axis have almost completely eliminated the effects of temperature over a small range of variations, and also prevent undesired coupling effects between the different modes of vibration. In the h.f. range high-stability frequency sources are obtained when these crystal resonators are used as frequency-determining elements. It is difficult to obtain directly a fundamental crystal frequency higher than 30 Mc/s,* but by the use of multiplier stages it is easy to reach the u.h.f. region. One very important feature of the crystal-controlled oscillator is that, besides being independent of small temperature variations, it can be put into a temperature-regulated oven and a very good long-term stability can be achieved.

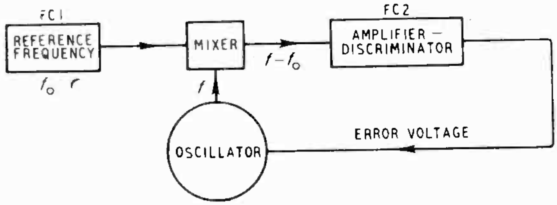


Fig. 1. *General form of a stabilizer embodying a stable reference oscillator.*

2.3. *Locking to External Source*

This is the case where a low-power stable source is used to control a less stable high-power oscillator.

The phenomenon of synchronization of oscillators is known. When the injected frequency is approximately equal to the frequency of the unstabilized source (sub-multiple frequencies not being considered here), the external signal affects both the instantaneous amplitude and the instantaneous frequency of the oscillator. The amount by which the frequency to be controlled is pulled increases with the magnitude of the injected signal; the 'pulling range' is a function of the circuit characteristics† and the respective stability of the two oscillators. Even when the injected signal is not large enough to produce synchronization, the frequency is disturbed and the pull-in effect on the oscillator exists outside the locking range. In general, the tendency towards synchronization lowers the expected beat frequency between the two oscillators and produces strong harmonic distortion.[6, 8]

For v.m. valves such as type 723A/B with an output of 20 mW and a loaded Q of the resonator of the order of 100, an incident power of 0.5 mW injected from a CV87 (higher stability source; $Q_l = 1000$) produced a pulling range of about 1 Mc/s.

The interaction of oscillators often occurs accidentally, and this supplementary effect has to be borne in mind in the microwave region, where a cavity resonator may be situated in the field of an adjacent oscillator.

2.4. *Frequency-control System*

This is the method used by the author; it is an indirect method similar in principle to the one used in a.f.c. systems for receivers.

The first requirement is a reference-frequency element FC1 which provides the reference frequency f_0. It can consist of either an active network, including a 'stable-frequency source' such as a crystal-controlled oscillator, or a passive network, such as a tuned circuit or resonator embodying all the previously mentioned precautions to ensure that its resonant frequency is stable within the order of approximation required for the 'stable-frequency source.'

The second requirement is a discriminator FC2 and it is usually associated with an amplifier. It is used in such a way that the unstabilized signal develops an error voltage dependent upon the algebraic difference between the frequency of the applied signal and the frequency of the reference element. This error voltage is then applied to an appropriate electrode of the oscillator in order to reduce this difference of frequency.

Such a frequency-control system can be thought of as an error-measuring device consisting of a discriminator-amplifier, combined in a feedback circuit. The frequency f of the oscillator to be stabilized is compared with a reference frequency f_0, supposed constant in a first approximation.

In the case of an active network this comparison is made by means of a frequency mixer. An i.f. amplifier is generally used after mixing, and the frequency f_0 differs from f by this intermediate frequency. Over a period of time, the beat frequency $(f - f_0)$ varies only by the amount of instability. The arrangement is sketched in Fig. 1.

When a passive reference circuit, such as a four-terminal network, is used the mixer is not essential. The general arrangement is then of the form shown in Fig. 2(a). The two frequency control elements FC1 and FC2 are not necessarily separate and could be shown as a single frequency-control element, consisting for instance of a single resonator followed by a rectifier and a d.c. amplifier. When this is done we obtain the most general representation of an f.c. system; i.e., a feedback loop through a four-terminal network

* 'Modal crystals' oscillating at a mechanical harmonic have produced starting frequencies of this order.

† Theoretically some non-linearity is necessary.

as shown in Fig. 2(b). This is also the general definition of a servo system.

Systems exist where a single frequency control element is used. A priori, this may be an advantage because the additive effect of the errors of FC1 and FC2 is avoided.

In all cases, the phase and amplitude of the error voltage must be kept related to the fluctuations of the oscillator frequency f in a unique way. A one-to-one correspondence between variations of frequency and variations of the error voltage is required. This condition is not always easy to achieve when circuits are complicated.

Besides the reference-frequency element the two important parts of the f.c. system are the amplifier and the discriminator. The former will be dealt with later ; the latter is basically an f.m.-a.m. converter. Frequency variations of an applied carrier are transformed into algebraic voltage variations ; Fig. 3 is a typical f.m. discriminator curve. This curve assumes constant amplitude of the input signals. Zero output corresponds to the reference frequency f_0. For input frequencies above f_0 a positive output voltage is obtained ; for frequencies below f_0 a voltage of opposite sign is produced. The detection may be square-law or linear. A linear law is desirable for most applications, but this is only obtained over a limited interval. Linearity is relatively unimportant when the discriminator is used in f.c. systems. When the discriminator is used either as an f.m. demodulator, or as the controlling element on the reactance valve of the usual a.f.c. system for receivers, the frequency range is usually defined by the projections, $f_1 f_2$, on the frequency axis of the linear part of the curve.

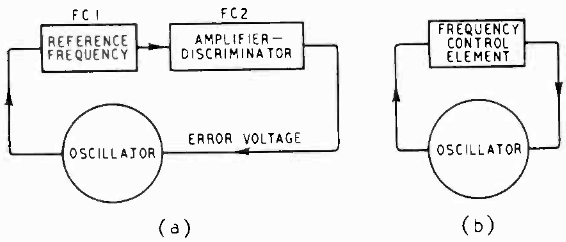


Fig. 2. *A stabilizer of the passive type takes the form shown at (a) and can be generalized to (b).*

This band can be now extended to include the whole working range, as we are no longer concerned with the linearity of the operation. In practice, the band must be broad enough to accept most of the received energy.

At intermediate frequencies the tuned circuits are conventional and the two main types of discriminators (amplitude type with three tuned circuits and phase of Foster-Seeley type) are well-known. If very narrow bandwidth and great stability of the tuned circuits are required, a quartz discriminator can be used. In microwave techniques, tuned-cavity resonators replace the tuned circuits*; a few specific examples of their use will be given later.

The sensitivity of a discriminator is defined by the slope of the voltage/frequency curve ; for the linear part generally considered, it is measured by

$$k = \frac{\text{output volts}}{\Delta f \text{ in Mc/s}} \quad \quad (1)$$

for a given amplitude of the input signal.

In order to minimize residual error, it is necessary to employ discriminator circuits of high Q and therefore difficult to adjust.

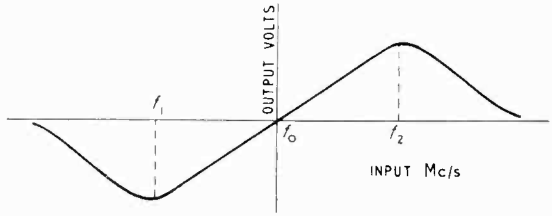


Fig. 3. *Typical discriminator characteristic.*

3. Stabilization of a V.M. Valve

In the following a reflex microwave klystron will be mainly considered, but the results are not confined to this part of the spectrum, and most of them apply to other arrangements.

The v.m. reflex oscillator is extremely useful as a local oscillator. One of the chief reasons for its general use is the fact that it can be tuned over a useful range of frequencies by varying an operating voltage ; mechanical tuning and sometimes thermal tuning can be used to supplement the electronic tuning. For most of the frequency stabilization systems it is necessary for the oscillator to be continuously variable in frequency through the whole range of stabilization, and therefore the *electronic frequency range*[11] is a limiting factor of primary consideration in the design of an f.c. system. The important parameters are listed below.

(a) The output power range expressed as a function of frequency. In the design of an oscillator it is necessary to make a compromise between output power and electronic tuning range.

(b) The electronic tuning range measured between the half-power points ; this range is roughly equal to the 3-db bandwidth Δf of the loaded resonator, provided that the

‡ Since writing this article the use of 'spectrum-line discriminators' based on resonant absorption properties of gases has made possible f.c. elements substantially independent of temperature (Cf. Ref. 14).

oscillator is loaded for optimum output and the circuit efficiency is good (over 60%).

$\Delta f/f \approx 1/Q_l$ .. .. (2)

The range is not independent of the applied voltages as it might seem, since the choice of a higher mode (by lowering the reflector voltage) will require a lower Q_l to obtain the optimum output.

(c) The tuning rate defined by

$$\rho = \frac{\text{frequency shift}}{\text{variation of reflector potential}} = \frac{\Delta f}{\Delta e} \quad \ldots \quad (3)$$

This ratio can be deduced from the reflector characteristics.

The frequency of a v.m. valve is a rapidly varying function of the reflector voltage. Large variations Δf can be obtained with relatively small values of Δe without restricting the output power. For the 723A/B the tuning rate ρ is of the order of 2 to 5 Mc/s per volt according to the conditions (mode, frequency, voltage, Q). A large increase in frequency range can be obtained by working beyond the range defined by the half-power points, but if ρ departs much from constancy, difficulty might occur in the operation of the control system.

It is thought, quite often, that the high-Q resonators existing at u.h.f. lead to very stable oscillators. This, however, proves not to be the case, and the relative natural stability of a v.m. valve is not much better than that of a normal h.f. oscillator. In fact, a small shunt conductance is needed due to the fact that the 'starting current' (minimum current below which the valve will not oscillate) depends on the conductance of the circuit[3] and not on its Q. The resonant cavity can be made with a very high Q, ten times better for example than that of an h.f. tuned circuit, but such a high Q cannot be used for an oscillator; the field and current distributions required across and around the narrow gap of the oscillating cavity are not obtained with a very high Q. Furthermore, when the oscillator is made tunable, as in the case of valves used in f.c. systems, the Q must be reduced to satisfy relation (2).

An oscillator, whether of the space charge or the v.m. type, has its fundamental frequency of oscillation determined by two conditions with respect to the feedback path :—

(a) The amplitude condition; theoretically a gain$>$1 is necessary; in practice that means that high-quality circuits are required.

(b) The phase condition; with a high-Q circuit phase variations are very rapid near the working frequency; such a circuit produces stable frequencies, but the range is very much restricted. The Q involved in a practical case can be found with the help of (2); if the electronic tuning range is $\Delta f = \pm 25$ Mc/s for a v.m. valve, such as the 723A/B used for the experiments at 10,000 Mc/s, $Q_l \approx 200$. In fact, owing to the variations of reflector voltage the condition for equation (2) is not fully satisfied with the result that the obtained value for Q_l is optimistic. With reference to this question it may be mentioned here that in a.f.c. stabilization schemes using reference cavities it has been found that a very high $Q(>1400)$ produces discontinuities in the oscillating frequency unless great attenuation is used between the oscillator and the cavity.

3.1 *Mechanism of Stabilization*

We may ask: What is the mechanism of stabilization of a v.m. valve? How does the variation of the reflector potential affect the frequency of oscillation?

An electronic oscillator consists of an electron beam exciting the parallel-resonant circuit formed by the tuned cavity. The electron stream passes first through a radio-frequency field across a gap between two electrodes, then into a drift space in which there is a retarding field (produced by the reflector), and finally returns through the radio-frequency field across the gap. The drift time, function of the voltage, is the fundamental parameter in electronic tuning[4,5]. Oscillations occur at the frequency at which the largest voltages can be developed across the resonator gap; this is in general realized when the total susceptance B at the gap vanishes. Action and interaction between the electron beam and the circuits form the basis of all v.m. valve theory. Because of velocity modulation and drift action the electron stream produces an admittance across the gap; for any voltage this electronic admittance Y_e can be obtained by dividing the radio-frequency electron current by the radio-frequency gap voltage. Three main admittances are considered in the conventional theory and contribute to the total admittance Y, at the gap of the resonator.

Let

$Y_r = G_r + jB_r$ be the admittance of the resonator.

$Y_l = G_l + jB_l$ be the admittance of the load.

$Y_e = G_e + jB_e$ be the admittance of the electron beam, all these admittances, which depend on frequency, being measured at the gap.

The susceptance B_e of the electron stream is defined by the Bessel functions of the bunching parameter. It is very small outside the resonance range, and its value can be supposed constant with constant operating voltages; it becomes very important at the frequency of oscillation.

The general condition for oscillation is given by

$$Y_r + Y_l + Y_e = 0$$

which leads to the two equations

$$G_r + G_l + G_e = 0 \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

of use in connection with problems of power transfer, and

$$B_r + B_l + B_e = 0 \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

defining the frequency f of oscillation.

The second condition, which is of interest in our present problem, means that the frequency of oscillation is determined by the condition that the electronic susceptance of the beam B_e, is to be equal, in absolute value, to the total circuit susceptance $(B_r + B_l)$, all susceptances being measured at the gap.

Any variation of susceptance, mechanical (tuning), electrical (voltage or load variations) or electronic (beam loading), will affect the frequency of oscillation. With reference to electrical variations of susceptance, it is generally found difficult to tune any oscillator by varying the susceptance of the load; the 'pulling effect' in the region of resonance leads, unless precautions are taken to limit the rate and the range of these susceptance variations, to frequency jumps and discontinuities in the output of the oscillator. If the coupling between the oscillator and the load is sufficiently lowered, the discontinuity will disappear but so also may the power output.

The frequency of oscillation is defined by the condition

$$B(\omega) = 0$$

where $B(\omega)$ is the total susceptance $(B_r + B_l + B_e)$ seen by the electron stream at the gap; and frequency instability occurs whenever

$$B(\omega) = 0 \quad \text{and} \quad \frac{dB(\omega)}{d\omega} = 0$$

are satisfied simultaneously.

All stabilizing schemes are based on equation (5) which must always be satisfied. Any accidental change $\Delta B(\omega)$ produces a variation $\Delta\omega$ in frequency: the way to reduce or cancel $\Delta\omega$ is to act in such a way as to balance it by an equal and opposite variation of susceptance. This is usually done by appropriate variations of the d.c. 'steady' operating voltages of the electrodes which produce the required variations in the beam susceptance B_e.

4. General Principles of F.C. Systems

We shall now outline, without going into the details, two theoretical aspects of the problem. Simplified and general theories will be used to give some insight into the behaviour of the circuits in an f.c. system.

4.1. *Stabilization Factor*

The most general form shown in Fig. 2(b) will be considered. If an unstabilized oscillator takes the desired frequency f at the instant t and produces a frequency drift Δf at the instant $t + \Delta t$, the range of the rate of variation $(\Delta f/\Delta t)$, to be covered in practice, puts a severe requirement on the system. Time constants must be exceedingly small and the corresponding bandwidths wide enough to accommodate fluctuations of the order of a fraction of a microsecond, such as are often encountered in practice. D.C. coupling is also required, after the discriminator, in order to keep the steady voltage at the right level at times when $\Delta f = 0$.

If the delays introduced by the circuits are neglected, a general expression can be deduced which is, however, applicable only when the duration of the frequency variations of the oscillator is much longer than the time constant of the circuits. The basic form of an f.c. system comprises two parts: an oscillator to be stabilized and a frequency-control element producing an appropriate error voltage. As a start it is assumed that the error voltage is not applied to the oscillator so that the two parts can be considered separately. When the desired frequency is supplied to the f.c. element, the error voltage is zero. If the applied frequency varies by df the error voltage produced is de, with

$$de = k\,g\,df$$

where g is the total voltage gain of the f.c. system, consisting mainly of the gain of the amplifier.

k is the sensitivity of the discriminator, previously defined (1) by the slope of its Volt/Freq. curve, for a given amplitude of the applied signal.

If such an error voltage de is applied to the frequency-control electrode of the oscillator, it will produce a frequency variation of

$$\rho de = k\,\rho\,g\,df \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

where ρ is the rate $\Delta f/\Delta e$ [Mc/s, V] deduced from the characteristics of the reflector curve (3).

If now the error voltage is applied and the circuit is closed, any frequency variation Δf of the oscillator feeds back an error voltage of the appropriate sign, which acts to oppose the initial Δf but, of course, does not completely correct it ; if Δf is the value of frequency drift when the oscillator is uncontrolled, after correction a residue of, say, δf, still remains. With the feedback loop completed, when the system disturbed by the initial Δf has reached equilibrium (it is supposed that the frequency variation is unique, or that sufficient time has elapsed for all effects of previous deviations to have disappeared) the frequency residue δf acting back through the f.c. element maintains this equilibrium by producing the error voltage δe applied to the reflector. Using (6) we obtain

$$\delta f = \Delta f - k \rho g \delta f \quad \quad (7)$$

This important relation is classical in the theory of automatic-control systems.[2] It has been derived here in the particular case of a system in equilibrium and does not take account of time ; i.e., of the transmission properties of the system. Therefore it cannot be used for a dynamical analysis of the system. The position of equilibrium, within the range of validity of the previous reasoning, is given by (7), but it is difficult to find because it depends on the unknown Δf, and furthermore, the mechanism of control becomes inoperative in the vicinity of this position. In fact, the small residue δf is essential for the operation of the control.

The stabilization (or control) factor can be deduced from (7) and is defined by

$$\Delta f / \delta f = 1 + k \rho g \quad \quad (8)$$

where Δf is the initial frequency variation of the oscillator if it were uncontrolled, and δf is the residual frequency variation of the oscillator remaining after correction.

Since a stable and non-oscillating system of control is required, in practice, the zeros and singularities of the function $(1 + k\rho g)$ are of great importance, since they determine the characteristic of the response. Thus three important parameters g, k and ρ determine the action of the f.c. system ; all of them can vary and can only be assumed constant as a first approximation.

The amplifier is a very important part of the system. Its gain g is in fact complex, and consequently its phase characteristic becomes important, but as the whole system is essentially a negative-feedback device, the restrictions imposed upon the amplifier may not be so severe. In practice its design is a compromise between bandwidth, gain and phase-shift characteristics ; in most cases a superheterodyne method is used to change the frequency of operation and make the use of a stable wide-band i.f. amplifier possible. The gain of the amplifier is then the product of the i.f. and d.c. gains ; the total amount of gain required and its distribution over the amplifier stages is a matter of design. When wide bands of the order of, say, 20 Mc/s are required, the i.f. stage gain is lower than that of a d.c. amplifier. Bulk and stability are also factors to be considered. When two discriminators are used, as in the scheme described later, or greater efficiency is desired, the characteristics of the amplifier must satisfy more stringent conditions if self-oscillation of the whole system, throughout the complete range of operation, is to be avoided.

4.2 *Time Delays and Stability*

Relation (7) was established without taking into account the transmission characteristics of the f.c. element. The effect of a single frequency drift Δf takes a certain time to go through the circuits, and the action of the error voltage is not synchronized with the cause producing it. The product ($k\rho g$) in equation (7) is complex and depends on the frequency ; therefore the amplitude and phase characteristics must be carefully controlled in order to avoid instability. These have not yet been taken into account, and it is now necessary to outline a more general analysis.

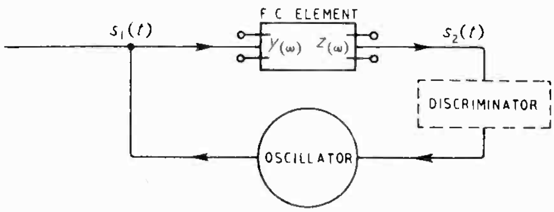


Fig. 4. *Generalized form of an f.c. system, taking phase shift into account.*

Referring to Fig. 4, a fraction of the signal voltage $s_1(t)$ produced by the oscillator is impressed on the input of the f.c. network, assumed to be linear. The discriminator and other non-linear devices of the system are not included in this analysis. The corresponding voltage output as a time function will be expressed as $s_2(t)$.

Let $Y(\omega)$ be the complex transfer admittance defined by $i_2(\omega)/s_1(\omega)$, and $Z(\omega)$ the output load impedance. The most general form of the output voltage is

$$s_2(t) = s_1(t)\, Y(\omega)\, Z(\omega) \quad \quad (9)$$

Replacing $s_1(t)$ by its Fourier integral

$$\int_{-\infty}^{+\infty} S(\omega)\, e^{j\omega t}\, d\omega$$

and applying the principle of superposition to the network, assumed to be linear, the output current can be obtained

$$i_2(t) = \int_{-\infty}^{+\infty} S(\omega)\, Y(\omega)\, e^{j\omega t}\, d\omega$$

where $Y(\omega)$ is the transfer admittance function depending on the network and its terminal impedances. This function has to be assumed known for all frequencies, since the integral covers an infinite frequency range.

The output voltage $s_2(t)$ appearing across the load impedance $Z(\omega)$ is deduced in a similar manner.

$$s_2(t) = \int_{-\infty}^{+\infty} S(\omega)\, Y(\omega)\, Z(\omega)\, e^{j\omega t}\, d\omega \quad .. \quad (10)$$

$[Y(\omega)\, Z(\omega)]$ is the transmission factor of the network and also, from (9), it is the generalized gain of the amplifier. This complex function can be replaced by

$$A(\omega)\, .\, e^{j\phi(\omega)}$$

in order to show both the amplitude and phase characteristics of the transmission system. Then (10) takes the form

$$s_2(t) = \int_{-\infty}^{+\infty} S(\omega)\, A(\omega)\, e^{j[\omega t + \phi(\omega)]}\, d\omega \quad .. \quad (11)$$

This well-known general form[1] is very useful in relating phase changes to time delays in the theory of networks. It is easy to verify that if $A(\omega)$ is replaced by $A(\omega)\, .\, e^{j\omega t_0}$ the output signal $s_2(t)$ is also replaced by $s_2(t + t_0)$, thus any linear phase variation in the transmission function

$$Y(\omega)\, Z(\omega) = A(\omega)\, e^{j\phi(\omega)}$$

of the network is equivalent to a time displacement in the response of the network. More generally any variation of the phase properties of the transmission network amounts to a variation in the time displacement. As $A(\omega)$ is, by definition, a frequency function, the error voltage δe produced by the action of $s_2(t)$ on the discriminator, is not synchronous with the cause Δf producing it. If a succession of frequency perturbations Δf is considered, self-oscillation of the control system is possible. This can be avoided by one of the following methods:—

(a) Making $Y(\omega)\, Z(\omega)$ less dependent on frequency and, if possible, making the system aperiodic.

(b) Controlling the phase and amplitude characteristics.

(c) Introducing time derivatives of the error voltage; with correct signs and amplitudes the condition for critical damping can be attained.

It is generally found that phase reversal of the feedback occurs at high frequencies, and therefore that oscillations of the control system take place; a capacitance across the reflector control circuit of the valve can prevent the trouble, although it is obvious that this reduces the range of action of the control system.

5. Figures of Merit

Besides the reliability of the system as a whole, several other factors are important in assessing the value of a frequency-stabilization scheme.

(a) Efficiency factor; i.e., how much r.f. power has been used to achieve the result. This is of special interest in microwave techniques where the amount of power is limited. The efficiency of the operation can be defined by the ratio

$$\frac{\text{Power usable after stabilization}}{\text{Total power available before stabilization}}$$

A good automatic controller extracts negligible energy from the output of the controlled system.

(b) The degree of stability obtained, measured by the maximum frequency deviation still existing after stabilization. It may be difficult to ascertain, in practice, the accurate value of this frequency excursion. Measurement of the absolute stability is a fundamental problem in itself. Relative instability measured against a sub-standard, or by comparison with some known system, is normally referred to. It comprises the short-term instability occurring over periods of the order of a second or less. The causes here are ripples and variations in the supplies, action of external stray a.c. fields or variations in the load impedance (a serious matter and one difficult to avoid), erratic mode changes, frequency jumps and scintillations. . . . These are phenomena occurring especially in microwave oscillators.

Long-term instability occurring usually over periods of time of over a minute and mainly due to slow variations in supply voltage and temperature effects in the oscillator circuits. Any instability over periods of time longer than a few seconds would be considered here as a long-term effect.

(c) The rate of change of any residual frequency fluctuation. This is a very important parameter in some special applications. If $A \cos \phi(t)$ is the oscillating signal, the instantaneous angular frequency is $d\phi(t)/d\omega$ and the drift rate will be defined by $d^2\phi(t)/dt^2$. Most of the schemes of frequency control are slow and remove only frequency drifts occurring in a time interval of a few seconds or more (long-term instability).

(d) The operating frequency range in Mc/s of the f.c. system. This is usually smaller than the electronic frequency range.

(e) The stabilization factor or control factor defined by the relation (8).

(f) The resetting accuracy of the system has also to be considered as in all control systems; upon it depend the frequency range of operation and the flexibility of the system.

6. Frequency Control at 10,000 Mc/s

The following considerations are confined to c.w. oscillators of the v.m. type§. When all the usual precautions regarding the regulation of power supplies and care in design have been taken the resulting stability of a v.m. valve, at 10,000 Mc/s is seldom better than 2 or 3 parts in 10^6; in other words, the assumed spot frequency of the oscillator covers a band of not less than 20 kc/s. In addition, frequency drifts due mainly to temperature effects may produce frequency variations of a few megacycles per second. Stabilization with an f.c. system invariably includes an amplifier and a discriminator, and the only difference between the two existing methods, which will be described now, lies in the choice of the reference frequency element.

6.1. *Stabilization by Reference Cavity*

Resonant cavities have been used extensively for frequency stabilization and various methods have been employed. The addition of an external resonant cavity to an oscillator for the double purpose of tuning and stabilizing has been used for magnetrons, and can give a stabilization factor of the order of 7, but only the case of a cavity used in an f.c. system will be considered here.

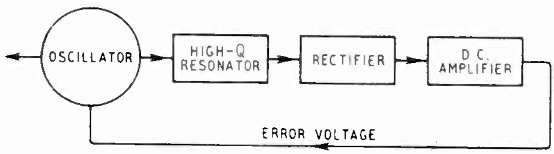


Fig. 5. *Form of stabilized system using a cavity resonator.*

In an f.c. system the cavity provides the reference frequency f_0 (frequency element FC1). Furthermore, the resonance curve of the cavity plus amplifier may be used as the second part FC2 to complete the frequency-control system. Clearly the direct use of the cavity does not give this result, for no discrimination is possible for frequencies in the neighbourhood of the resonance point. If, however, the mean frequency of the oscillator to be controlled is adjusted on the side of the resonance curve a control system can be devised on the basis of frequency variations translated into amplitude variations. A crystal rectifies the energy transmitted by the cavity, and the crystal current, suitably amplified, can be fed back to control the reflector of the v.m. valve (see Fig. 5). It can be shown that due to the characteristics of the available valves and crystals in the 10,000 Mc/s region a voltage amplification of the order of 70 db is necessary. Microwave discriminators have been successfully developed by R. V. Pound[7, 9, 10].

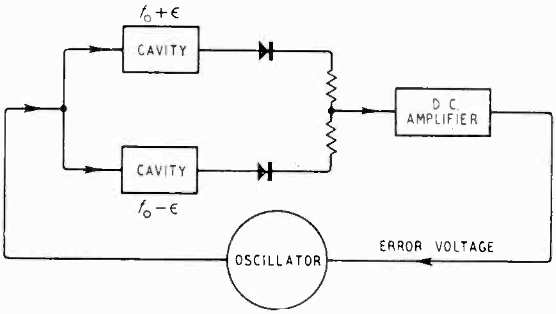


Fig. 6. *Double cavity-resonator for frequency stabilization.*

Several other schemes have been devised.

(a) A single-tuned cavity used with additional modulation added to the crystal or to the cavity itself in order to obtain a discrimination effect.

(b) In another scheme two cavities, tuned to two different frequencies on each side of the frequency f_0, are employed. The two rectified currents in opposing each other in a common resistive load produce the same result as described previously for a discriminator (Fig. 6).

(c) Use can also be made of the symmetry of balanced circuits. A system now in use working on the basis of a phase discriminator, consists of a single cavity resonator in conjunction with two crystals and the operation of a magic tee in which the length of the side arms differ by $\lambda/8$. It has been shown that for such a system a signal of the order of 1 milliwatt at 10,000 Mc/s, combined with cavities with unloaded *Q*s of 30,000, can give $gk = \frac{3}{4}$ volt/Mc/s. The sensitivity of the discriminator and the factor of efficiency of the f.c. system are low compared with the values obtained by the author (see 7.2). All these d.c. systems require a large proportion of the power to be

§ For a more general analysis see Ref. 12.

diverted through the f.c. system. A further disadvantage occurs in the d.c. amplifier which suffers from 'shift' and hum pick-up.

(d) Another system of cavity stabilization has been devised which eliminates the d.c. amplifier. An oscillator working at the intermediate frequency modulates in amplitude the h.f. current of one of the crystals of the magic tee. The sidebands produced are mixed with the original signal in the other crystal, and the intermediate frequency thus obtained has its sign and magnitude dependent on the amplitude and phase of the reflection coefficient of the controlling resonant cavity. Amplification and detection in a phase mixer (mixing with the original i.f. local oscillator) produce a unidirectional voltage identical in character with the output of a discriminator.[7] This i.f. system gave better results than the d.c. system in all respects.

(e) A discriminator based on the change in position of the standing-wave pattern in a line terminated by a resonant cavity has also been used. Two fixed probes A and B, spaced the distance of a quarter-wavelength apart, are adjusted so that their outputs are equal at resonance. For frequencies either higher or lower the output of A increases, that of B decreases, or vice versa. When the rectified outputs of two crystals connected to the probes are subtracted a curve of the discriminator type is produced.

The efficiency of all the above schemes is bound up with the properties of the cavities. Many disappointments have been encountered in using thermally-compensated cavities. Because variations of temperature and pressure affect the dielectric constant, cavities evacuated and sealed by means of glass windows placed over the coupling apertures have been designed. It may be noted here that the performance of materials such as invar or quartz, coated with a high-conductivity surface has been found to be inferior in respect of temperature compensation to that achieved by the use of two materials having opposite temperature coefficients of expansion. A copper cavity with a copper diaphragm attached to an invar rod, fixed at its upper end to the copper cylindrical cavity, also results in a high order of compensation. The phenomenon of temperature-hysteresis in tuning occurs in many metals, and limits the resetting accuracy of cavities; solids, it is known, have their properties affected by residual strains and by the previous mecnanical and heat treatment to which they have been subjected.

Simplicity is a very important feature of the method of frequency stabilization by reference cavities. It will, no doubt, be possible to achieve in the near future a degree of temperature compensation of the same order as that of the crystal. With the increase of machining accuracy the frequency tolerance of the cavity should become comparable with that obtained from a crystal. Furthermore, the cavity can be continuously tunable, and the valve will follow the cavity frequency to the extent of its tuning range. A short-term stability of 10^{-8}, and a long-term stability of the order of 1 part in 10^5 of the frequency of resonance of the cavity have been obtained by R. V. Pound.[7]

General considerations and some preliminary investigation have shown the difficulties in assessing the stabilization factor or even the maximum frequency drift of any system, as no standard frequency source exists in the 10,000-Mc/s region. Some information on the degree of stabilization can be obtained by beating the frequencies of two identical stabilized systems and comparing the purity and the stability of the beat notes thus obtained. In other beating methods a comparison is made between those two frequencies and that of a third intermediary oscillator; the result can be made independent of this comparison frequency, but interaction and consequent synchronization of the oscillators is the real difficulty in all these methods. For general purposes and for measurement a crystal-controlled source in the 10,000 Mc/s region, even of low power, was considered essential, and the second general method of stabilization was used by the author.

A complete comparison between stabilization by means of resonant cavities and stabilization by means of a crystal-controlled frequency has not been possible. Of the six parameters defined in Section 5 it has been possible to evaluate only the degree of stability and the stabilization factor. Further information and systematic comparison of the various systems on the basis of the figures of merit may lead to some reassessment of the different methods.

6.2. *Stabilization by a Crystal-source*

The method is well-known and widely used in h.f. and v.h.f. work. The piezo-electric crystal is, up to now, the most precise and most reliable among the known frequency-determining elements; its performance is improved by the choice of special cuts, and also by the use of special circuits. The fundamental frequency of oscillation is of the order of a few megacycles per second, but this frequency can be multiplied up to the region of 200 Mc/s in conventional valve-multiplier stages. If precautions are taken to avoid large multiplication factors, especially in the initial stages, the harmonic content of the final frequency is reasonably good for the application in view. Although with doublers and treblers many stages of amplification are necessary, the system is flexible and not difficult to put into proper adjustment. The amplifier valves needed for such a chain are available for frequencies up to 3,000 Mc/s. Multi-resonator klystrons can be used as power amplifiers and frequency multipliers at 3,000 Mc/s, but no 10,000-Mc/s klystron amplifier has yet been produced. Otherwise, by injecting a small amount of a 3,000 Mc/s crystal-controlled oscillation into such a klystron, it would be possible to obtain a crystal-controlled 10,000-Mc/s source directly.

The only stable crystal frequency that could be used was therefore the one obtained by successive multiplying stages. Consequently, frequency variations of the unstabilized oscillator had to be controlled by an f.c. system. The production of a 10,000-Mc/s signal from a crystal-controlled 3,000-Mc/s oscillator was first considered, but the advantages and the possibilities of utilizing a greater step for this harmonic generation were recognised to be important enough to warrant some research. The non-linear characteristics of the crystal are known to give rise to a rich harmonic content, and it was finally found possible to construct a suitable harmonic generator giving high-order harmonics.

The harmonic generator was then utilized to produce a laboratory model of a 10,000-Mc/s frequency-stabilized system which took the form of an improved a.f.c. system locking a 10,000-Mc/s local oscillator to the 46th harmonic of a 2.3-Mc/s crystal-controlled signal; a short term stability of the same order as that given by a crystal was eventually obtained.

Before describing the system it may be noted here that in this method the stability is not entirely defined by the crystal; variations of the discriminator circuit will affect the stability to the same extent as crystal variations. The design problem is therefore of importance. Another disadvantage is the effect of spurious u.h.f. or i.f. signals which can take over control in any system of this kind. On the other hand, the system can be controlled at any desired harmonic frequency of the crystal, or at any other chosen frequency within the controlling range, and furthermore, by superimposing a modulation voltage on the discriminator or the i.f. amplifier, a frequency-modulated output signal can be obtained within the stabilization range. In describing the performance of the system we shall not consider long-term instability, as no special precautions were taken to obviate this effect. Fluctuations caused by temperature variations were numerous and would be difficult to evaluate separately. It was found possible to keep the controlled oscillator, with all its parts exposed to the air, within $\pm$ 100 c/s of the crystal frequency, for a few hours, after temperature equilibrium of the gear had been reached.

The a.f.c. system employed required some manual tuning to supplement the control system in the initial stage. Referring to Fig. 1, it is obvious that if the beat frequency $(f - f_0)$ moves beyond the limits of the i.f. pass band no error voltage will be produced and the f.c. system will become inoperative. A.F.C. systems with full automatic action, operating as soon as the set is switched on, usually imply large frequency corrections (10 to 15 Mc/s, at least, at 10,000 Mc/s for the transient period). Some kind of slow mechanical searching or electronic scanning is thus required to shift the frequency throughout the frequency range: when the oscillator passes through the correct frequency, a.f.c. starts to operate and takes over the control of frequency. For the v.m. valve used (723A/B type) the coefficient $\left(\frac{\Delta \text{ freq.}}{\Delta \text{ temp.}}\right)$ was approximately $-$ 0.2 Mc/s per degree centigrade, and the maximum frequency deviation between cold and warm was greater than the control range (5 Mc/s) provided by the f.c. system. Therefore either some initial tuning had to supplement the controlling action, or the set had to be switched on some time before use.

Another consideration is the efficiency of the control as regards slow or rapid variations in frequency. A motor-driven or any other mechanical device allows only for slow variations.

In conclusion the main requirement for a stabilization scheme involving a crystal-controlled reference frequency seemed to reside in the ability of the system to follow all frequency variations, however rapid they might be. This necessitated very short time constants for the circuits, which is not incompatible with the wide frequency band to be used.

7. Characteristics of the Scheme Employed

As no thought was given to the mechanical layout, a detailed description of the equipment will not be made. The American v.m. valve 723A/B used as an oscillator was stabilized against a 9,360-Mc/s signal of the order of 10^{-9} watt having its frequency controlled by a 2.8-Mc/s crystal.

The work consisted of two main parts.

7.1 *Harmonic generation with crystal distorter*

A crystal-controlled transmitter with a power output of about 0.5 watt at 200 Mc/s was used. The nominal crystal frequency was 2,826 kc/s. The temperature of the crystal was maintained at 50 ± 2 degrees C by means of a temperature-controlled oven. Five successive multiplication stages in the order 2 × 2 × 3 × 2 × 3 raised the frequency to 2.826 × 72 = 203.472 Mc/s. Mismatch being avoided as far as possible, this signal was fed to a crystal distorter consisting essentially of a conventional 10,000-Mc/s mixer in which the crystal was mounted axially along the *E*-vector in the centre of a rectangular waveguide. A 10,000-Mc/s radial choke fitted to the crystal tip eliminated interaction between the waveguide section and the v.h.f. source. The 46th harmonic of the input signal, corresponding to a frequency of 203.472 × 46 = 9,360 Mc/s, was chosen. In the early stages of the work a 406.94-Mc/s signal was used, but identical results were obtained later with half this frequency. The signal was detected by a receiver tuned to 9,360 Mc/s (wavelength = 3.2 cm) preceded by a high-*Q* calibrated transmission-type filter (cavity resonator of the H_{01} type) in order to reject all the harmonics except the one required. Both the signal and the local oscillator were fed into a mixer (CV113) and the resulting i.f. output was passed through a conventional i.f. amplifier. As harmonic multiplication of frequency also takes place in the mixing crystal, the high power level of the local oscillator (milliwatt region) was fed through a variable attenuator (15 db max) in order to minimize the distortion thus produced. By signal noise considerations the amount of 9,360-Mc/s power produced was estimated to be of the order of 10^{-9} watts; i.e., well above the noise power level ($KT\Delta f$) of a 5-Mc/s bandwidth receiver.

Silicon crystals of the CV103, CV113, CV253, CV257, 1N23B and J-1N26 types (Sylvania and Western Electric) were used. The performance and possibilities of these crystals were not fully investigated. Good and consistent results were obtained with the 1N23B, CV257 and some experimental crystals at 24,000 Mc/s. A systematic study may correlate d.c. characteristics, low-frequency rectification efficiencies and u.h.f. properties with the amount of harmonic content. Second-order parameters, such as method of preparation, presence of impurities, construction and assembly details, as well as the existence of secondary peculiarities in shape or curvature of the characteristic curves, might provide some information on the efficiency of the harmonic power generation.

Suitable biasing was found necessary depending on the type of crystal utilized; a self-bias, produced by a crystal current of 5 to 15 mA, was the average required. This was not critical although one or two dead spots throughout the range were repeatedly noticed. The experimental procedure eventually adopted consisted in varying the crystal current by altering a series resistance in order to locate the optimum working point; the optimum value of this resistance did not exceed 10 kΩ. Beyond this value very little difference

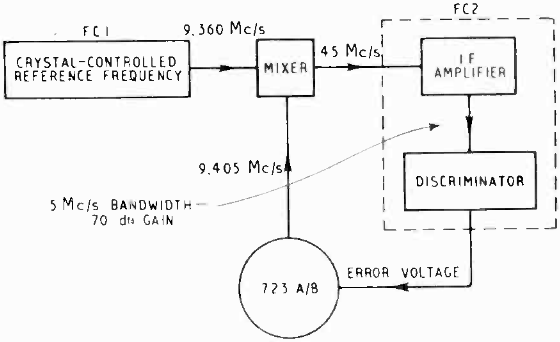


Fig. 7. *Block diagram of the experimental stabilizer.*

was noticed and the r.f. output remained practically constant from 5kΩ to infinity. Considerable improvement in the performance of the crystal distorter resulted from this practice, for the blank spots mentioned often led to almost complete obliteration of the signal in the early stages of the work.

In spite of the success in producing a fair amount of 9,360-Mc/s power with these relatively low frequencies—no doubt a frequency as low as 20 to 30 Mc/s will be used later as these techniques improve—further investigation is needed on the following points:—

(a) Measurements and correlation of the different parameters of the crystal with their performance as harmonic frequency multipliers. Conditions of optimum performance.

(b) Law of distribution of the harmonic content of a given crystal distorter. The comparison between the output of the 23rd and that of the 45th or the 46th harmonics did not lead to any conclusion. The efficiency of an harmonic power generator in decibels, is given by

$$10 \log \frac{\text{Available } n^{th} \text{ harmonic power}}{\text{Input power (fundamental)}}$$

but the direct measurement of power is always difficult to make at u.h.f. and the problem is even more complicated here because some method must be used to select only power of a given harmonic.

(c) Investigation of new crystals, especially of the germanium type. Basis for selection. Germanium mixers were not available at the time of these trials (September–October 1946).

(d) Noise properties of such harmonic crystal distorters.

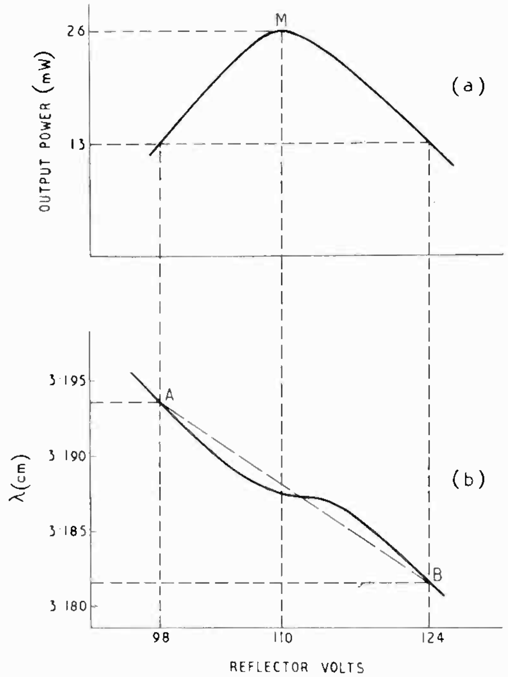


Fig. 8 (above). *Curves of output power (a) and wavelength (b) against the reflector voltage.*

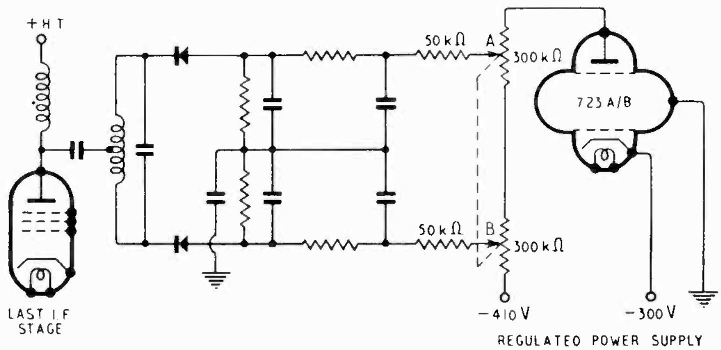


Fig. 9 (right). *I.F. discriminator and its connections.*

7.2. *System of Frequency Control*

The crystal controlled 9360-Mc/s oscillation of the order of 10^{-9} watt was used as FC1 for the stabilization of the reflex v.m. valve 723A/B. The general block diagram is shown in Fig. 7.

The 723A/B (Western Electric) which is a reflector type of v.m. valve was used as a c.w. oscillator. The beam produced is velocity modulated during its 1st transit and the bunched electrons, after being reflected, give up energy to the resonator during their 2nd transit. A frequency variation of $\pm$ 200 Mc/s (i.e., $\pm$ 2%), by mechanical tuning was available, and an additional $\pm$ 25 Mc/s approx., due to electronic tuning, obtained by variation of the reflector voltage. The -110-volts reflector mode was used (-110 V with respect to cathode potential) as the valves available gave better oscillations with this mode. The total input power was of the order of 10 watts, with a maximum output of approximately 25 mW with the resonator at 300 V and a total current of 20 mA. The output was taken through the base of the 723A/B by means of a short coaxial line feeding directly into a waveguide. An attenuation of the order of 12 db reduced the crystal current in the mixing crystal (CV113) to approximately 1 mA. It may be noted here that the same crystal used as harmonic generator could also be used for mixing. It was, however, thought preferable to separate the two functions at this preliminary stage.

After determining the centre M of the mode (Fig. 8), the reflector voltage was varied and the resulting frequency shift and power output were measured. The curves are shown in Fig. 8(a) and (b) for the valve used in the final equipment. These curves are the characteristics required for the design of the f.c. system. The maximum frequency shift obtained from one end A to the other B was approximately 35 Mc/s; this defines the useful electronic tuning range. Assuming AB to be a straight line, the coefficient

$$\rho = \Delta f/\Delta e \approx 2 \text{ Mc/s/V}$$

With the input signal employed (0.5 mW for the local oscillator and 10^{-9} watt for the crystal-controlled signal) 1-Mc/s variation in the uncontrolled oscillator frequency produced an error voltage of about 10 V at the output of the amplifier-discriminator. With g and k, defined as before

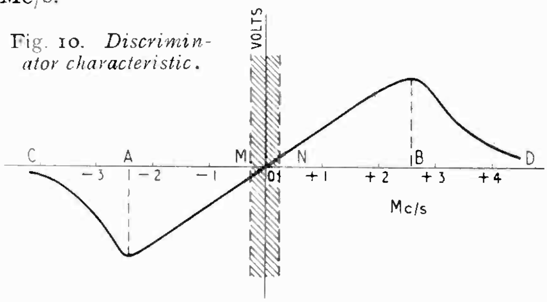
$$gk = 10$$

and the stabilization factor defined by (8) is 21.

Two germanium high-level detectors were used in the i.f. phase discriminator to ease the problem of the d.c. connections required between the discriminator and the reflector. The amount of error voltage applied to the reflector was controlled by varying the two ganged potentiometers A and B of Fig. 9.

The overall response (i.f. amplifier + discriminator) is given by the curve of Fig. 10. The origin 0 corresponds to the mid-band frequency of 45 Mc/s.

Fig. 10. *Discriminator characteristic.*



An additional two stage d.c. amplifier for the error voltage produced by the discriminator was used at the start, but this was found unnecessary when a greater amount of 9,360 Mc/s was produced. The 45-Mc/s i.f. amplifier is conventional. The bandwidth of the amplifier and the discriminator first used was of the order of 15 Mc/s to cover the whole of the electronic tuning range. Later, however, attention was mainly concentrated on finding out how nearly the stability of the oscillator approximated to that of the crystal, and an i.f. amplifier and discriminator of only 5-Mc/s bandwidth were employed.

8. Results of Initial Experiments

As already mentioned long-term stability was not measured, all the effort being devoted to achieving high short-term stability. Part of the output of the i.f. amplifier was fed to a display unit giving a visual indication of the signal on a screen, over a wide frequency range. The maximum range of frequency sweep was of the order of 1 Mc/s ; the centre frequency of 5.25 Mc/s was made to correspond to the i.f. signal of the Hallicrafter S27 used for reception. Frequency deviations of the order of 10 to 20 kc/s could easily be seen on the screen ; smaller deviations were more accurately measured by beating the i.f. output with the 45-Mc/s crystal-controlled frequency derived from the frequency-multiplier gear. The scheme described gave a short-term stability of the order of $\pm$ 6 kc/s with respect to a 45-Mc/s crystal frequency.

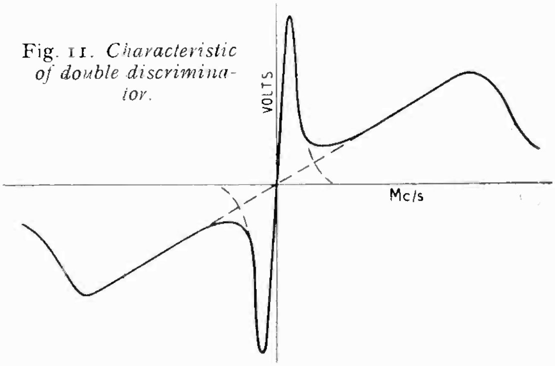
To improve the performance a better discriminator was required : the residual error decreases when the slope $\dfrac{\Delta \text{ output volts}}{\Delta \text{ frequency}}$ increases. A steeper discriminator curve can be achieved by

(a) an increase in the sensitivity of the discriminator. This can be done, for example, by decreasing the bandwidth : the benefit thus obtained would not normally exceed a few decibels.

(b) an increase in the gain of the amplifier, resulting in an increase in the output voltage ;

(c) a better design of the discriminator in order to improve the slope at the cross-over point. With single-tuned circuits this can only be achieved by raising the Q, and consequently at the expense of the bandwidth.‖ With coupled circuits it is theoretically possible to steepen the sides of the resonance curve : good slope can be associated with moderately wide band but, in practice, the mutual reactions of these various tuned circuits make the adjustment intricate. The ideal form of discriminator required is, of course, a combination of two adjacent perfect band-pass filters, the d.c. components being reversed by the action of the two rectifying crystals.

Fig. 11. *Characteristic of double discriminator.*



9. Use of a Second Discriminator

It was then thought that a separate discriminator having a narrow bandwidth could be added to take up the fine adjustment near the origin. In principle, the first discriminator, with a frequency range AB = 4 Mc/s (Fig. 10) presents an 'uncertainty zone' which occupies a frequency band MN centred around the origin 0. It must be noted that the working range of the discriminator is wider than AB. The limiting points C and D are not, however, accurately defined, because the range CD, of the order of 8 Mc/s, suffers from a serious decrease in power when the curve approaches its asymptote.

With the single wide-band discriminator the error voltage pulled the v.m. valve to within $\pm$ 6 kc/s of the reference frequency; i.e., the

‖ A quartz discriminator can produce a stability of a few c/s for a bandwidth of a few hundred cycles, the central frequency being of the order of 100 kc/s.

uncertainty zone MN was approximately 12 kc/s. A second discriminator with a curve of adequate shape could cover this region MN and produce in its turn an 'uncertainty zone' very much smaller (an approximate estimate deduced from laws of similitude gives a zone of the order of 40 c/s). In fact, the two discriminator curves combined together produce a curve of a general shape, as shown in Fig. 11.

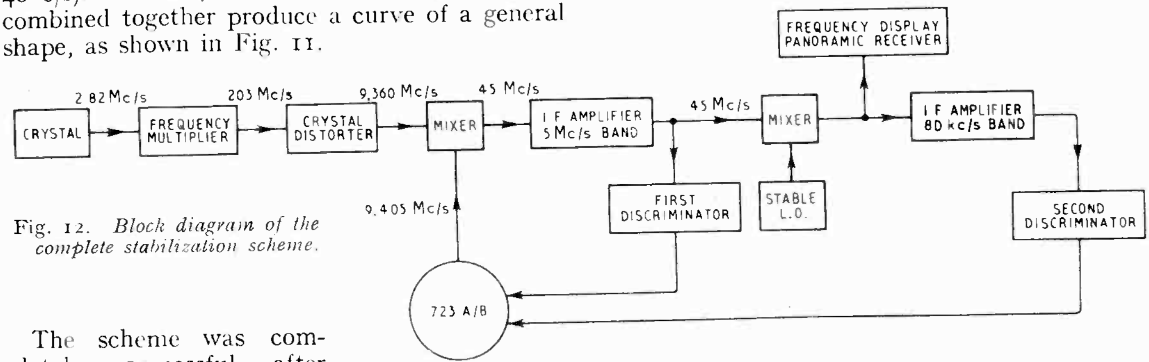


Fig. 12. *Block diagram of the complete stabilization scheme.*

The scheme was completely successful after some effort had been devoted to arranging the circuits and the time constants in such a manner as to prevent interaction of the discriminators and hunting due to the presence of the new control element. The sensitivity of the second discriminator was greatly increased by the addition of an extra 3-stage i.f. amplifier. An S27 Hallicrafter receiver, slightly modified to feed the panoramic receiver and tuned to 45 Mc/s, was used; the selectivity switch was at the 'broad' position (approximate bandwidth 80 kc/s) variable gain of maximum value 40 db). Results confirm that increase of sensitivity produces better stability, as a consequence of the better slope of the characteristic curve of the second discriminator. The complete scheme is illustrated in Fig. 12.

No amplitude-limiting system was used for the signals, although this was found to be desirable in order to keep the characteristics of the discriminator constant, as the 'uncertainty zone,' which is a function of the slope at the cross-over point of the curve of the discriminator, varies with the input signal. It may be noted, however, that such a limiter is incorporated in the S27, in the f.m. detector system.

The results with this double discriminator were very good. The 723A/B was stabilized within $\pm$ 100 c/s of the frequency of the crystal; this was checked against a 45-Mc/s harmonic of the controlling crystal. The crystal harmonic itself was not stable to $\pm$ 100 c/s, but this does not affect the validity of the result. The 9,405-Mc/s oscillator was, therefore, locked to the crystal in such a way that the overall variation of frequency was contained in a band less than one part in 10^8 wide, with respect to the crystal reference frequency.

10. Applications

During the course of development several possible applications of the scheme were envisaged, but only those which appear to have practical value are listed here. The laboratory set-up described can be used for precision measurements involving stable frequencies in the microwave region (high *Q*s of cavities, absorption spectra of selective phenomena, observation of the fine structure which may be involved in such phenomena). The model described can also be used as a reference system for other systems of stabilization. Application to the communications field presents a double advantage; accuracy in transmitter carrier frequency and stabilization of the local oscillator in the receiver. In the system of communication known as 'cross band' at each station the transmitting oscillator is also used as local oscillator for the receiver[13] and consequently, only one oscillator at each end needs to be stabilized. The possibilities and advantages in multi-channelling have already been mentioned.

The modulation of stabilized oscillators is possible. Direct reflector-voltage modulation may be used with low-frequency stabilization only. The f.c. system is d.c. coupled to the reflector and an a.c. modulation could be connected at the same point, but if this is done the highest frequency for which frequency stabilization could operate would necessarily be less than the lowest modulation frequency. Frequency modulation about the stabilized frequency can also be achieved, with stabilization acting through the period of the modulating cycle. A frequency-modulated signal can be superimposed on the discriminator, or the modulation directly applied to the crystal-controlled frequency. Alternatively the i.f. amplifier can be amplitude, phase or frequency modulated; the same methods can be applied to the crystal (distorter or mixer) bias voltage, or even to an additional crystal in the waveguide circuit.

11. Acknowledgments

All the original work described in the present paper was done in Admiralty Signal and Radar Establishment and permission to publish is gratefully acknowledged. The author is indebted to Dr. J. Thomson for many useful suggestions and for his interest in the work, and especially to F. Pugliese for his active collaboration and his invaluable help through critical discussion, during the whole course of the investigation.

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[6] "A Study of Locking Phenomena of Oscillators." R. Adler. *Proc. Inst. Radio Engrs*, June, 1946.

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[12] "Microwave Mixers", R. V. Pound, M.I.T., Radiation Laboratory Series, Vol. 16, Mc Graw Hill, 1948—Chapter 7 by E. Durand on Frequency Control of Local Oscillators.

[13] "A Duplex System of Communications for Microwaves", R. V. Pound, *Proc. Inst. Radio Engrs*, July 1948.

[14] "Frequency Stabilization of Microwave Oscillators by Spectrum Lines", Smith, De Quevedo, Carter and Bennett, *Journal of Applied Physics*, December 1947.

CORRESPONDENCE

Correction

SIR,—I very much regret to advise you of an error, and several minor mistakes in my paper, "Mutual Impedance of Parallel Aerials," published in the November 1948 issue of *Wireless Engineer*.

In the practical example of the curves I gave on page 352, the lines 28–30 should be read as follows:

$$d/\lambda = 0.075 \qquad \theta_{11} = 10^\circ$$

we have $F_0/F_{\lambda/4} \approx 1.78 \qquad R_0 \approx 4.6$ ohms.

Therefore the diagram of Fig. 12(b) no longer refers to the example and should accordingly be corrected to the new values.

The following corrections should also be made: P. 344, first column, line 5, instead of $\mathbf{I}_y = I_0 \sin my$ read $\mathbf{I}_y = I_1 \sin my$. In the second member of equation (8) the factor $1/2\pi$ is missing. P. 346, first column, line 14, instead of $m\Delta' l_1 = 45^\circ$ read $m\Delta' l_1 = 5^\circ$. P. 350, second column, last line instead of $l_1/\lambda = 0.5$ read $l_0/\lambda = 0.5$.

Please accept my apologies for my carelessness in correcting the proofs.

G. BARZILAI.

Syracuse, N.Y.

NEW BOOKS

A Textbook of Radar

By the Staff of The Radiophysics Laboratory, Council for Scientific and Industrial Research, Australia. Pp. 579 + v, with 31 plates and 347 diagrams. Chapman & Hall Ltd., 37, Essex Street, London, W.C.2. Price 50s.

Like most other books on radar, this is a collective work. The twenty chapters have been written by as many named authors, with the acknowledged assistance of others unnamed. Unlike most collective works, however, it has been co-ordinated to a degree which is not always achieved even by a single author, and which reflects very great credit on the unnamed editor as well as on the contributors. The style is consistently good, and there is hardly any overlapping.

Nor has space been wasted on matters that should be part of the equipment of any professional radio engineer. The historical side of radar is only briefly sketched in, and no more than due attention has been given to the earlier metre-wave types. As a result of this freedom from padding, all but a small proportion of the book is available, and has been used, for expounding microwave radar.

Among the subjects treated most fully are aerials, microwave transmission lines (especially waveguides) and cavities, klystrons, modulators, and display circuits. Details of actual equipments are included sparingly, only where desirable to illustrate principles and practice. In this way the disadvantage arising from the fact that all authoritative sources of information on radar were organizations devoted to the 1939-45 war effort has been minimized in the present publication. Although civil radar as such is only briefly treated, and apparently without much basis of experience, the contributors have concentrated mainly on those principles most likely to be applied in post-war developments.

Unfortunately the contributions appear to have been written before the general publication of work on radar, in this country at least. Except for one or two in the last chapter, there appear to be no references dated later than 1945; and the majority of them are therefore to unpublished reports of Government or industrial organizations. In particular, it was evidently too soon for Part IIIA of the I.E.E. Journal, and as a result there are some obvious thin patches on the U.K. side. Terminology, too, is often American rather than British. It is good, however, to have some information on the Australian contribution to radar, which hitherto seems to have had less attention than its due.

It is to be hoped that writers of textbooks in this country will, like those in America and Australia, overcome their conservatism to the extent of adopting the rationalized M.K.S. system on units. The book under review is a good example of how unnecessary impediments can thereby be removed from the path of the student, and clarity—especially in the treatment of transmission of microwaves—be increased.

M. G. S.

The Measurement of Stress and Strain in Solids

Pp. 114 + xvi with 41 illustrations. The Institute of Physics, 47, Belgrave Square, London, S.W.1. Price 17s. 6d.

This book consists of a collection of papers read at a conference at the University of Manchester in July 1946. There are four on the characteristics and use of resistance strain gauges, three on photoelasticity, two on X-rays and one on acoustic methods.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 2/- each.

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

598 287.—Frequency-discriminating circuit for applying both positive and negative feedback to control the tone of an audio-frequency amplifier.

E. K. Cole Ltd. and E. L. Hutchings. Application date 25th April, 1939.

AERIALS AND AERIAL SYSTEMS

598 165.—Frame aerial in which the wire turns are embedded in a frame of high dielectric-constant and magnetic permeability, such as synthetic resin impregnated with iron-dust.

Masteradio Ltd. and R. Pollock. Application date 17th October, 1945.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

599 700.—Cyclically-operated tuning-means for operating cavity-controlled resonators used for radiating a blind-landing beam.

Sperry Gyroscope Co. Inc. (assignees of D. F. Folland, J. O. Schock and F. A. Jenks). Convention date (U.S.A.), 27th May, 1942.

599 889.—Radiolocation system arranged to give a plan-position indication of the terrain under observation, and provided with means for temporarily arresting the normal drift of the display.

F. C. Thompson. Application date 15th October, 1945.

599 922.—Feeding and inter-coupling arrangement for an aerial array of three dipoles for radiating overlapping beam signals.

Marconi's W.T. Co. Ltd. (assignees of G. H. Brown). Convention date (U.S.A.), 21st April, 1943.

600 073.—Radiolocation device for indicating at a given station the identity of objects located at different other points, by correlating the responses of separate interrogating units.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 18th March, 1944.

600 101.—Construction designed to increase the directive characteristic of an electromagnetic horn, and for conveniently coupling it to a waveguide.

A. B. Pippard. Application date 29th September, 1945.

600 117.—Navigational system in which a craft is steered along a given course by comparing the reception of synchronized signals, radiated on different carrier frequencies, from two spaced transmitters.

H. G. de France. Convention date (France) 27th October, 1939.

TRANSMITTING CIRCUITS AND APPARATUS

(See also under Television)

600 024.—Receiver for frequency-modulated signals wherein the carrier wave is first converted into square-topped pulses which are then counted to derive the detected signal.

Philco Radio and Television Corpn. Convention date (U.S.A.). 24th October, 1944.

600 076.—Muting or noise-limiting arrangement for frequency-modulated receivers, particularly when tuning between stations.

Philco Radio and Television Corpn. Convention date (U.S.A.) 2nd September, 1944.

600 299.—Receiver capable of detecting either amplitude or frequency-modulated signals by first converting them into an equivalent train of time-modulated pulses.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 24th May, 1943.

600 384.—Superhetrodyne receiver in which a number of switch-selected crystal oscillators are coupled to the local oscillator for the purpose of station selection and to prevent frequency drift.

J. P. Wykes. Application date 13th October, 1945.

600 750.—Mounting a radio receiver on the dashboard of a motor-car so that the loudspeaker is tilted downwards in order to economize space.

C. W. Eggleton and Smith's Motor Accessories Ltd. Application date 10th November, 1944.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

599 747.—Television scanning system in which a variable-velocity traverse is used to prevent the formation of undesirable potential gradients on the surface of the mosaic screen.

J. D. McGee. Application date 20th July, 1945.

600 018.—Television system in which an iconoscope tube is arranged to generate signals that represent the instantaneous light intensity and also changes in its average or background value.

Farnsworth Television and Radio Corpn. Convention date (U.S.A.) 10th January, 1944.

600 096.—Method of mounting the c.r. tube in a television receiver so as to reduce the normal dimensions of the cabinet.

D. Jackson and Pye Ltd. Application date 27th September, 1945.

TRANSMITTING CIRCUITS AND APPARATUS

(See also under Television)

599 625.—Coaxial-line feeder with quarter-wave stubs for coupling a short-wave transmitter and receiver to a common aerial.

Standard Telephones and Cables Ltd. (assignees of A. G. Kandoian). Convention date (U.S.A.) 27th June, 1944.

599 748.—Scanning head with cylindrical copy-holder and traversing mechanism for a facsimile telegraph system.

Marconi's W.T. Co. Ltd. (assignees of C. J. Young). Convention date (U.S.A.) 1st August. 1944.

599 978.—'Anti-fatigue' arrangement of the desk and control panel used in the prolonged supervision of broadcast transmissions.

The British Broadcasting Corpn, F. C. McLean and L. E. H. O'Neill. Application date 26th September, 1945.