The Journal of

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TECHNICIANS, ENGINEERS, SCIENTISTS

SIMULTANEOUSLY with the release of information of the plan for new Colleges in the Universities of Oxford and Cambridge is a minor report in a radio trade journal of a speech criticizing the Government for "spending a lot of money on higher education for technicians, but that is not where you get service engineers. We want the middle run of engineers."

Not for some time has there been such blatant misunderstanding of the definition of a service mechanic, a technician and an engineer.

Engineers are not unaware of their responsibility toward the technician and the mechanic; the operation and maintenance of nearly all engineering equipment increasingly requires mechanics and technicians for the constantly improving standard of technical efficiency. Whilst improved reliability in design and construction may limit the demand for the mechanic, automation will not materially affect the constant recruitment of the technician, particularly in supporting and maintaining the work of the engineer.

Encouraging improved efficiency in the mechanic and technician has for long been a feature of the work of most professional engineering institutions. In the field of radio particularly, our own Institution takes pride in its record of having promoted and participated so actively in the work of the Radio Trades Examination Board. An examining body which, in thirteen years, has issued 2,550 certificates, might be termed as reasonably successful—but it was never established as a means of enabling the successful candidates to describe themselves as service *engineers*. It may not be out of place to suggest that those who take upon themselves the responsibility of advising the trade or industry in speeches might well acquaint themselves thoroughly with the education and training schemes already operating within the industry.

Notwithstanding the promotion of university colleges in other parts of Britain, few will deny the further advantage of the schemes which will enable the two senior Universities to increase the numbers of their students reading for science degrees.

The foundation of the new Oxford College intended to replace St. Catherine's Society will be substantially aided by long term donations of at least £75,000 from industrial sources. This support is an encouraging sign of the new patronage of the universities by industry.

That the new College of Cambridge should be named the Churchill College is indeed a tribute to a man whose vision and support of science did so much for the free countries during the last world war. The Trustees of the appeal say: "Here is an opportunity of training a larger *corps d'elite* which could exert a great influence on technological development. The proposal takes advantage of the Cambridge system, its tradition, its great prestige in science and the arts, and its consequent appeal to young people. And it fits within the framework of the Cambridge pattern without destroying or distorting it."

These tangible efforts to promote excellent opportunity for the future scientist and engineer will remain long after the controversy on the proper use of the word *Engineer* has been resolved—possibly by legislation.

INSTITUTION NOTICES

Obituary

The Council has learned with regret of the deaths of the following members, and has expressed sympathy with their relatives.

Hartley Wilkinson died on 31st March after a long illness. Mr. Wilkinson was a founder member of the Institution, having been elected to Associate Membership in 1926. In more recent years he had been in business on his own account as a radio engineer at Keighley, Yorkshire. He leaves a widow.

* * *

Ronald Brian Evans (Student) died on 11th March after undergoing an operation. He was 25 years of age, and since completing his military service in 1956 had been employed by the Weapons Research Division of A. V. Roe & Company Ltd., Woodford, Cheshire. He was registered as a Student in 1957.

* * *

Prem Das Mahendru (Student) was killed in a road accident in Amritsar on the 17th December. He received a B.Sc. degree in Physics from the East Punjab University in 1949, and was 28 years of age.

British National Committee on Non-Destructive Testing

The Institution has recently become affiliated to the above co-ordinating body which consists of various professional and learned societies concerned with non-destructive testing. The representative of the Brit.I.R.E. on the National Committee will be Dr. A. Nemet (Member), who was for a number of years a member of the Institution's Technical Committee.

International Symposium on Electronics Components

Elsewhere in this issue of the *Journal* appears the final group of summaries of the papers presented at the Symposium on Electronic Components held at the Royal Radar Establishment, Malvern, last September. In response to numerous enquiries it is re-emphasized that these papers are not available from the Institution. Complete papers, together with accompanying discussions, will be published in the Proceedings of the Symposium which will be issued by R.R.E.; as yet no details are available regarding the date or price.

Grants for Sandwich Course Students

With a view to increasing the number of students taking advanced "sandwich" courses at technical colleges, particularly from the medium size and small firms, the Ministry of Education and the Federation of British Industries have reviewed the arrangements for financial assistance.

Many firms already pay students' tuition fees and salaries during the whole of the course, and the F.B.I. recommends its members to continue doing so. Firms are reminded that such payments are treated as normal business expenses for taxation purposes.

It is, however, recognized that the smaller firm may not be able to meet the whole cost of such training. The Minister of Education has therefore recommended Local Education Authorities to give sympathetic consideration to students who are in this position by making awards to cover tuition fees and maintenance.

Permanent Binding of the "Journal"

As members will be aware, a service for permanent binding of their *Journals* has existed for a number of years. Due, however, to the larger size of the *Journal* and to the higher costs of materials and postage, it has been necessary to increase the cost of this service to 15s. 0d., with postage and packing extra as follows:—Great Britain 3s., Commonwealth and other countries 4s.

Members are reminded that complete year's issues and index should be sent with remittance, and that care should be taken to ensure that these are carefully packed to avoid damage in transit.

Monthly Filing Binder for the "Journal"

Wire-type binders are now available from the Institution for filing up to 12 issues of the *Journal* through the year. These binders are fully bound in red leather cloth with *Journal Brit.I.R.E.* blocked in gold on the spines. The cost is 12s. 6d. each, to which postage should be added as follows:—Great Britain 1s. 3d.; Commonwealth 2s.; other countries 5s.

THE DESIGN OF INDUCTIVE POST-TYPE MICROWAVE FILTERS*

by

M. H. N. Potok, M.A., B.Sc., Ph.D. (Associate Member)†

SUMMARY

By application of the results of theoretical analysis and experiment it is shown that microwave filters can be designed to have a desired v.s.w.r. within the pass band and a given insertion loss outside it. Steps leading to an optimum design of three- and four-cavity filters are discussed in detail.

1. Introduction

A considerable amount of literature has appeared in recent years dealing with the design of microwave filters[3, 4, 5]. These, while very helpful, often tend to be either too theoretical in nature or too general in application to allow one to proceed with the design without further collection of empirical data. The design of a waveguide filter to have the desired characteristics is made difficult by two facts: firstly, the obstacles used can rarely be represented by a single susceptance, and secondly, even if an equivalent susceptance or a simple network is found to describe the obstacle fully, this can be true at one frequency only; thus in trying to estimate the bandwidth of the filter one has to consider changes in the susceptances which cannot be covered by reasonably simple mathematical expressions. The paper attempts to resolve these difficulties by a combination of theoretical analysis and empirical results applicable at all microwave frequencies but to one particular type of filter that made up of inductive posts placed centrally along the waveguide at suitable intervals. This type of filter is often used because of its simplicity of manufacture as compared with the multipost or the iris type.

Before proceeding it is thought necessary to clear up one point which has caused some confusion in literature. Almost invariably the design of a filter involves the computation of the "Q"s of the component cavities. The term "Q" borrowed from tuned circuit theory is used freely as a shorthand for the ratio of centre frequency to bandwidth, defined by difference between frequencies at which the power output falls by 3 db. Since the filter is virtually lossless the fall in output is caused by reflection, which at 3 db insertion loss causes a voltage standing wave ratio (v.s.w.r.) of 5·8 at the input to the filter, which could not be permitted because of its effect on the generator or the travelling wave tube. It is arguable what is the permitted v.s.w.r. in the pass band and in the following it is assumed that it must be no greater than 1·05. The consequence of this is very serious because to get, for example, a 3 db bandwidth of 20 Mc/s at 4 Gc/s would call for Q of 200, whereas, if the v.s.w.r. within the 20 Mc/s is to be kept to below 1·05, a Q of 5 is required with single cavity and Q of 58 with a three-cavity maximally flat filter. Thus it is essential to realize—and all too often this has been overlooked—that the relation between the desired bandwidth and Q is not a straightforward one and if the Q is retained it is only because of a certain mathematical facility it affords. In itself it gives little information about the filter. This is very different from the state of affairs at lower frequencies, where matching is not the all-important consideration.

2. The Design of a Single Cavity Post Filter

2.1. *First approximation*

It may be helpful to consider initially the posts as being to a first approximation pure inductive susceptances jb (normalized) separated by a length of waveguide of θ_0 electrical degrees at centre frequency as in Fig. 1. If the input voltage and current were V_1 and I_1, and output

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voltage and current V_2 and I_2, then they can be related together by the equations

$$V_1 = AV_2 + jBI_2$$
$$I_1 = jCV_2 + DI_2$$

(where A, B. C and D are real) which can be written in matrix form

$$\begin{vmatrix} V_1 \\ I_1 \end{vmatrix} = \begin{vmatrix} A, & jB \\ jC, & D \end{vmatrix} \begin{vmatrix} V_2 \\ I_2 \end{vmatrix}$$

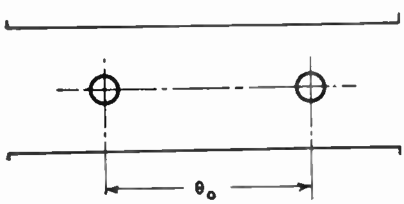


Fig. 1. A two-post cavity filter.

If the network were matched at both input and output so that the generator and the load impedances are, when normalized, both equal to unity, then, since the matrix of the generator series impedance is $\begin{vmatrix} 1, & 1 \\ 0, & 1 \end{vmatrix}$

the relation between generator voltage and current V_0, I_0 and the output voltage and current V_2, I_2 is (see Fig. 2):

$$\begin{vmatrix} V_0 \\ I_0 \end{vmatrix} = \begin{vmatrix} 1, & 1 \\ 0, & 1 \end{vmatrix} \begin{vmatrix} A, & jB \\ jC, & D \end{vmatrix} \begin{vmatrix} V_2 \\ I_2 \end{vmatrix}$$

or

$$\begin{vmatrix} V_0 \\ I_0 \end{vmatrix} = \begin{vmatrix} A+jC, & jB+D \\ jC, & D \end{vmatrix} \begin{vmatrix} V_2 \\ I_2 \end{vmatrix}$$

Hence $\quad V_0 = (A+jC)V_2 + (jB+D)I_2$

but $\quad \dfrac{V_2}{I_2} = 1$

thus $\quad V_2 = \dfrac{V_0}{(A+D)+j(B+C)}$

Now it is clear that without the filter network the output voltage V_2' would be exactly $V_0/2$, hence the insertion of the filter causes a change in the output voltage given by the insertion ratio

$$\frac{V_2'}{V_2} = \frac{A+D}{2} + j\,\frac{B+C}{2}$$

It follows that the insertion phase shift is given by

$$\tan\psi = \frac{B+C}{A+D}$$

and the insertion loss is given by

$$L = \left|\frac{V_2'}{V_2}\right|^2 = \left[\left(\frac{A+D}{2}\right) + j\left(\frac{B+C}{2}\right)\right]\left[\left(\frac{A+D}{2}\right) - j\left(\frac{B+C}{2}\right)\right]$$

$$= \tfrac{1}{4}(A+D)^2 + \tfrac{1}{4}(B+C)^2$$
$$= AD + BC + \tfrac{1}{4}(A-D)^2 + \tfrac{1}{4}(B-C)^2$$

Since the circuit is symmetrical, $A = D$, and by reciprocity theorem $AD + BC = 1$, hence

$$L = 1 + \tfrac{1}{4}(B-C)^2$$

The matrix of a pure susceptance jb is

$$\begin{vmatrix} 1, & 0 \\ jb, & 1 \end{vmatrix}$$

and of a length of lossless waveguide of θ_0 electrical degrees

$$\begin{vmatrix} \cos\theta_0, & j\sin\theta_0 \\ j\sin\theta_0, & \cos\theta_0 \end{vmatrix}$$

hence the matrix of the filter cavity becomes (see also Levy[8])

$$M_{b\theta b} = \begin{vmatrix} 1, & 0 \\ jb, & 1 \end{vmatrix} \begin{vmatrix} \cos\theta_0, & j\sin\theta_0 \\ j\sin\theta_0, & \cos\theta_0 \end{vmatrix} \begin{vmatrix} 1, & 0 \\ jb, & 1 \end{vmatrix}$$

$$= \begin{vmatrix} \cos\theta_0 - b\sin\theta_0, & j\sin\theta_0 \\ j\,[2b\cos\theta_0 + (1-b^2)\sin\theta_0, & \cos\theta_0 - b\sin\theta_0 \end{vmatrix}$$

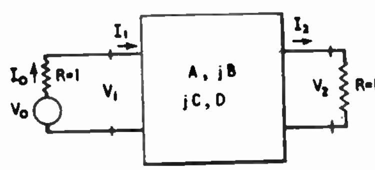


Fig. 2. To illustrate the derivation of insertion ratio.

The insertion loss of this cavity is then

$$L = 1 + \tfrac{1}{4}[2b\cos\theta_0 - b^2\sin\theta_0]^2$$

Since the insertion loss $L = 1$ at the centre frequency it follows that

$$2b\cos\theta_0 - b^2\sin\theta_0 = 0$$

or $$\tan\theta_0 = \frac{2}{b}$$

which defines the distance between the posts. Since b is negative, θ_0 is just less than π. If frequency f_0 changes by δf then the distance θ_0 changes by $\delta\theta$ which can be shown to be

$$\delta\theta = \theta_0\left(\frac{\lambda_g}{\lambda_0}\right)^2 \frac{\delta f}{f_0} = \left(\tan^{-1}\frac{2}{b}\right)\left(\frac{\lambda_g}{\lambda_0}\right)^2 \frac{\delta f}{f_0}$$

where λ_0 is the wavelength in free space, and λ_g is wavelength in the guide. A little trigonometrical manipulation shows that the insertion loss is given in terms of $\delta\theta$ by

$$L = 1 + \frac{b^2}{4}(b^2+4)\,\delta\theta^2$$

provided that b does not change with frequency.

By substituting for $\delta\theta$ from previous expression, Q can be found as corresponding to $L=2$, i.e.

$$Q = \left|\frac{f_0}{2\delta f}\right|_{L=2} = -\frac{b}{4}\sqrt{b^2+4}\left(\frac{\lambda_g}{\lambda_0}\right)^2 \tan^{-1}\frac{2}{b}$$

If variation in b were taken into account, then Reed[5] has shown that the expression for Q would have to be amended to

$$Q = \tfrac{1}{4}\left(\frac{\lambda_g}{\lambda_0}\right)^2\left[\left(\tan^{-1}\frac{2}{b}\right)(-b\sqrt{b^2+4}) + \frac{2b^2}{\sqrt{b^2+4}}\right]$$

The usual approximation for the susceptance of an inductive post of diameter d is given as

$$b = -\frac{2\lambda_g}{a}\,\frac{1}{\ln(4a/\pi d) - 2}$$

where a is the width of the guide.

The expressions given above would allow one to design a single cavity filter of the required Q, and to calculate the insertion loss and phase shift at all frequencies. However, the results would be only approximately correct, and the error would rapidly increase with Q.

2.2. *Second approximation*

Marcuvitz[1] has shown that the equivalent circuit of a post is more nearly a T-network consisting of a shunt inductive susceptance jb and two series capacitive reactances jx, where

$$x = -\frac{a}{\lambda_g}\cdot\frac{(\pi d/a)^2}{1+\frac{11}{24}\left(\frac{\pi d}{a}\right)^2} \qquad \text{.........(1)}$$

and
$$b = -\frac{1}{\frac{a}{2\lambda_g}\left[\ln\frac{4a}{\pi d} - 2 - \left(\frac{\pi d}{2\lambda_0}\right)^2\right] + \frac{x}{2}} \qquad \text{......(2)}$$

where the meaning of the symbols is as in previous paragraph. These formulae are still only approximations, but they hold reasonably well up to $d/a = 0{\cdot}25$ (which as a rule covers most requirements) as can be seen by reference to Fig. 3, which gives the measured and the computed values at 4Gc/s in waveguide 11 (2·37″ by 1·12″). The results given by Marcuvitz[1] at 10 Gc/s are equally satisfactory.

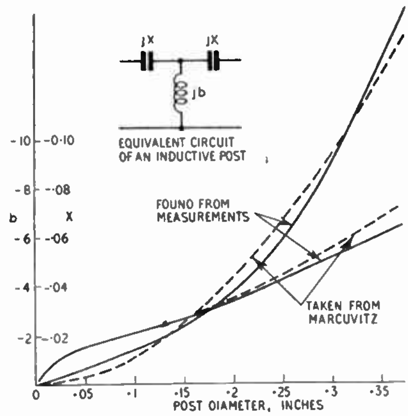


Fig. 3. Shunt susceptance and series reactance of post across a waveguide 2·37″ × 1·12″ at 4 Gc/s.

Assuming this equivalent circuit the matrix of a single cavity consisting of two posts separated by θ_0 electrical degrees now becomes

$$M_{bx} = \begin{vmatrix} A, & jB \\ jC, & A \end{vmatrix}$$

where
$$A = [(1-xb)^2 - (2-xb)xb]\cos\theta_0 - (1-xb)[x(2-xb)+b]\sin\theta_0$$

$$B = 2(1-xb)(2-xb)\,x\cos\theta_0 - [(2-xb)^2x^2 - (1-xb)^2]\sin\theta_0$$

$$C = 2b(1-xb)\cos\theta_0 - [b^2 - (1-xb)^2]\sin\theta_0$$

The insertion loss
$$L = 1 + \tfrac{1}{4}(B-C)^2 = 1 + \tfrac{1}{4}\{2(1-xb)[(2-xb)x - b]\cos\theta_0 - [(2-xb)^2x^2 - b^2]\sin\theta_0\}^2 \qquad \text{......(3)}$$

which leads to

$$\theta_0 = \tan^{-1}\frac{2(1-xb)}{b+x(2-xb)} \qquad \text{......(4)}$$

θ_0 can, of course, be expressed in terms of λ_g, the guide wavelength, by the simple relation $l_0/\lambda_g = \theta_0/2\pi$.

The usefulness of the above expression for θ_0, lies in the fact that it defines uniquely the separation between two posts, which results in no insertion loss (resonance). The question of Q does not arise here, and the accuracy of l_0 depends on the degree of approximation involved in finding x and b. The correctness of this can be judged from Fig. 4, which compares the computed and the measured results at 4 Gc/s in waveguide WG11. The results appear entirely satisfactory and point to one step in the design; the other step is to find Q.

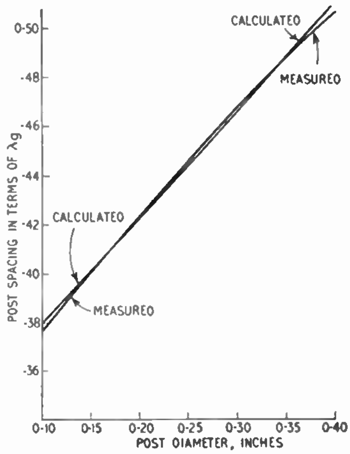


Fig. 4. Post spacing against post diameter to resonate at 4 Gc/s in waveguide WG 11.

The insertion loss of the cavity at any frequency off resonance can be obtained by finding the total differential of L in eqn. (3). This leads to

$$L = 1 + \tfrac{1}{4}\Bigg\{ - [(2-xb)x-b]\,[\sqrt{(2-xb)^2+b^2}]\,\delta\theta + \frac{2x(2-xb)(xb-1)+2b}{\sqrt{(2-xb)^2+b^2}}\,\delta b + \frac{2(2-xb)(1-xb)b+2b^3}{\sqrt{(2-xb)^2+b^2}}\,\delta x \Bigg\}^2$$

Now $$\delta\theta = \theta_0\left(\frac{\lambda_g}{\lambda_0}\right)^2\frac{\delta f}{f_0}$$

$$\delta b = -b\left(\frac{\lambda_g}{\lambda_0}\right)^2\frac{\delta f}{f_0}$$

$$\delta x = x\left(\frac{\lambda_g}{\lambda_0}\right)^2\frac{\delta f}{f_0}$$

By substituting these into the above expression and equating it to 2, one can find Q to a close approximation.

It will be noted that if x is made equal to 0 this reduces to the expression for Q quoted earlier from Reed.

The Q's have been calculated for various sizes of posts at 4 Gc/s and compared with those

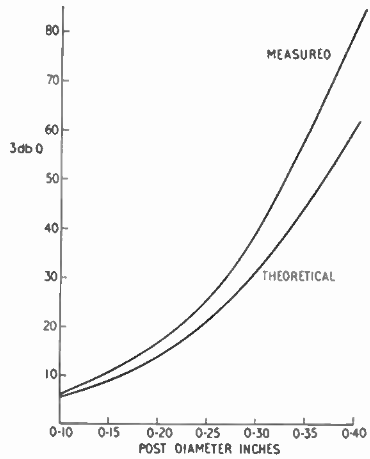


Fig. 5. Q of cavities at 4 Gc/s.

found by measurements in Fig. 5 (see also appendix). The considerable discrepancy indicates that either the simple equivalent circuit of the posts is not complete enough to estimate the effects of frequency change or that the approximations involved in arriving at the theoretical Q are adding up to an appreciable error. It is, therefore, suggested that as a basic approach one should use not the theoretical but empirical Q for the design. In order to make this possible, the quantity $Q/(\lambda_g/\lambda_0)^2$ is plotted not against post diameter but against post spacing l_0/λ_g as in Fig. 6.

Both analysis and measurements at 4 Gc/s and at 7 Gc/s confirm that this curve remains correct irrespective of frequency or waveguide dimensions provided the spacing has been properly calculated.

Thus it is suggested that entirely satisfactory design results from a computation of b and x with the help of eqns. (1) and (2) or by reference

$$Q = \tfrac{1}{4}\left(\frac{\lambda_g}{\lambda_0}\right)^2\frac{1}{\sqrt{b^2+(2-xb)^2}}\Big\{[(2-xb)x-b][(2-xb)^2+b^2]\,\theta_0 + 2b(1-xb)[b-2x(2-xb)] + 2x^2(2-xb)^2\Big\}$$

to the graph given by Marcuvitz followed by calculation of θ_0 from eqn. (4). Q is then found by reference to Fig. 6. Of course, if as is usual one has to start with Q then it is necessary to compute step by step the relation between Q and post diameter for a given frequency and waveguide size.

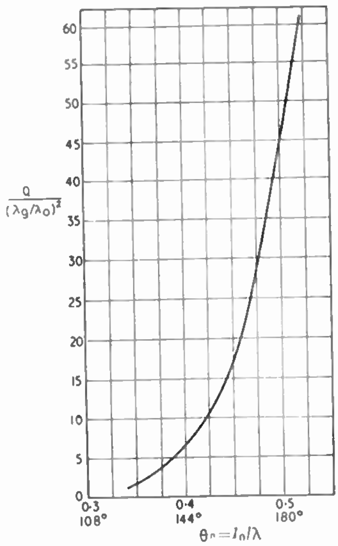


Fig. 6. Relation between $Q/(\lambda_g/\lambda_0)^2$ and post spacing l_0/λ_g.

3. The Design of a Complete Filter

The design of an individual transmission cavity has been given first as being basic to the problem, but in fact, one starts, of course, with a desired v.s.w.r. within and insertion loss outside the nominal pass band of a complete filter. The filter is usually designed to have either a maximally flat or a Chebyshev characteristic.

3.1. *The maximally flat filter*

This type has been popularized and fully discussed by Mumford[2]. For completeness it is proposed to summarize the basic expressions involved in this design. The filter consists of a number of cavities (n) arranged in such a way that the rth cavity has a 3 db Q_r given in terms of the overall 3 db Q_T by the expression

$$Q_r = Q_T \sin \frac{2r-1}{2n} \cdot \pi$$

To find the overall Q_T one has to decide on the maximum v.s.w.r. (σ) which, of course, occurs at the edge of the pass band of width $B = 2\Delta f$. The insertion loss of a maximally flat filter is given by

$$L = 1 + \left(\frac{2\delta f}{f_0} Q_T\right)^{2n}$$

which at band edge $= 1 + \left(\frac{2\Delta f}{f_0} Q_T\right)^{2n}$

and in terms of v.s.w.r.

$$L = \frac{(\sigma - 1)^2}{4\sigma}$$

$$\text{hence } Q_T = \frac{f_0}{2\Delta f}\left[\frac{(\sigma-1)^2}{4\sigma}\right]^{\frac{1}{2n}}$$

Once Δf and σ have been chosen, the reflection loss (i.e. the insertion loss based on the assumption of no filter internal loss, which should be well below 1 db) outside the band depends on the number of cavities used. The author has computed the reflection loss for 1 to 5 cavities and three values of v.s.w.r. The results appear in Fig. 7. For example, if a

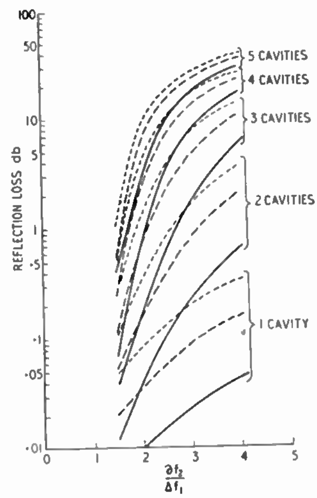


Fig. 7. Ratio of δf_2 for which given reflection loss is desired to Δf_1 at which voltage standing wave ratio is to be below that given by: solid line—1·05; long dashes—1·10; short dashes—1·15.

reflection loss of 30 db is required 40 Mc/s off centre frequency of a 20 Mc/s-band filter then 5 cavities are essential if the v.s.w.r. at band edge is to be kept below 1·15 (in fact, it will be only 1·05 in this case).

Using single post obstacles, the spacing between individual cavities will have to be about $\frac{3}{4}\lambda_g$ to avoid spurious mode coupling. If two neighbouring cavities are respectively θ_1 and θ_2 electrical degrees long, then the spacing between them must be made by $\theta_{12}=\frac{1}{2}(\theta_1+\theta_2+\pi)$.

Thus the design is completed.

3.2. *The Chebyshev-type filter*

Once it is agreed that a certain amount of variation in reflection loss is permitted over the band, the question arises whether the maximally flat filter is the best one to be used, or could one find a better filter in which the v.s.w.r. varies over the pass band but not excessively so. In an ideal Chebyshev filter the insertion loss undulates within the pass band between 0 and 3 db. This is not permissible for communication application of waveguide filters. Recently Levy[8] has shown that a form of Chebyshev characteristics can be obtained but the undulations can be kept down to any desired value by a suitable choice of cavities. The expressions derived by Levy are most useful in determining a suitable design, but they fall short of giving the complete answer because the filter bandwidth appears to be given as the 3 db bandwidth which, as stressed earlier, is inadmissible. However, the necessary information can be derived by taking a few more steps. The treatment can no longer be made universal, and has to be applied to 3, 4, etc., cavity filters, individually.

3.2.1. Three-cavity quasi-Chebyshev filters

Following Levy's notation, let the consecutive cavities have 3 db Q's of Q_x, Q_y, Q_x where

$Q_x/Q_y=k$. Also let $y=\dfrac{2\delta f}{f_0}\,Q_y$.

Then the insertion loss $L=1+16k^4y^2\left(y^2-\dfrac{2k-1}{4k^2}\right)^2$

Thus the steepness of insertion loss outside the band pass will increase rapidly with k. The insertion loss is unity at centre frequency ($y=y_0=0$). It rises to a maximum at

$$y_1=\frac{\sqrt{2k-1}}{2\sqrt{3}\,k}$$

when $L=L_1=1+\dfrac{(2k-1)^3}{27k^2}$ and having then

gone through a minimum ($L=1$) at $y_2=\dfrac{\sqrt{2k-1}}{2k}$ it will rise steadily to infinity, passing through an insertion loss equal to L_1 at $y_3=\dfrac{\sqrt{2k-1}}{\sqrt{3}\,k}$. It is this last quantity y_3 which together with L_1 allows one to arrive at a suitable design. Assuming the maximum permissible v.s.w.r. (σ), the maximum insertion loss L_1 is fixed, hence k, hence y_3 which defines the bandwidth, and since this is known, Q_y can now be fixed:

$Q_y=y_3\cdot\dfrac{f_0}{2\Delta f}$ and $Q_x=kQ_y$. Thus the 3 db Q's are derived and the individual cavities can be designed as in sect. 2.2. For example, if the maximum v.s.w.r. is fixed at 1·05, $L=1{\cdot}006$ hence $k=0{\cdot}59$ and $y_3=0{\cdot}416$. If the desired bandwidth is 20 Mc/s centred on 4000 Mc/s then

$$Q_y=0{\cdot}416\times\frac{4000}{20}=83{\cdot}2$$

and $Q_x=49{\cdot}1$. It will be noted that $k=0{\cdot}5$ gives a maximally flat filter. The corresponding maximally flat filter would have $Q_y'=57{\cdot}3$ and $Q_x'=28{\cdot}7$.

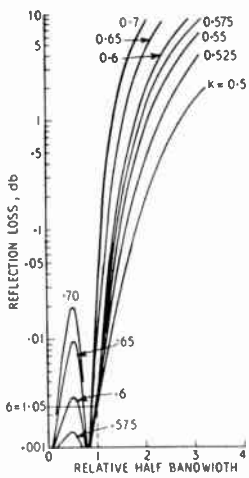


Fig. 8.
Reflection loss of 3-cavity filters with various values of parameter k.

Fig. 8 gives the computed reflection loss for various values of k plotted to the base of normalized half bandwidth showing the considerable increase in insertion loss outside the pass band resulting from increasing k. If $\sigma>1{\cdot}05$ within the pass band were permitted then the graph shows that further improvement in insertion loss is possible. Fig. 9 gives the characteristics of three filters designed on these principles.

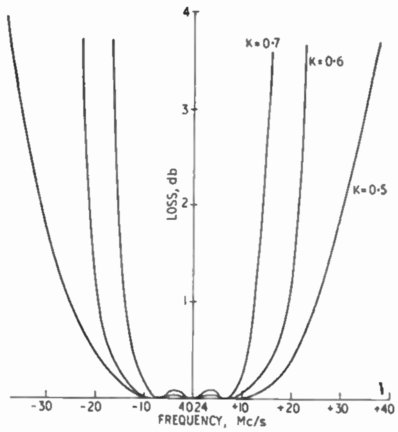


Fig. 9. Insertion loss of three 3-cavity filters.

3.2.2. Four-cavity quasi-Chebyshev filters

Let the 3 db Q's of the individual cavities be respectively Q_x, Q_y, Q_y, Q_x, and let $Q_x/Q_y=k$ and $y=(2\delta f/f_0)Q_y$ as before. Again Levy has shown that $L=1+4y^4[(4y^2-1)k^2-(2k-1)]^2$. The insertion loss is at a minimum at the centre frequency, rising to a maximum

$$L=L_1=1+\frac{(k^2+2k-1)^4}{64k^4}$$

$$\text{at } y_1=\frac{\sqrt{(k^2+2k-1)}}{2\sqrt{2k}}$$

then to a minimum at

$$y_2=\frac{\sqrt{(k^2+2k-1)}}{2k}$$

and beyond rising steadily to infinity and passing the loss$=L_1$ at $y_3=0{\cdot}55\dfrac{\sqrt{k^2+2k-1}}{k}$

This time the optimum design for a maximum v.s.w.r. of 1·05 requires $k=0{\cdot}49$ (the maximally flat filter results with $k=0{\cdot}41$)

hence $y_3=0{\cdot}527$

$$Q_y=0{\cdot}527\,\frac{f_0}{2\Delta f}$$

and $Q_x=0{\cdot}49Q_y$.

Following the example given in previous paragraph

$$Q_y=105{\cdot}3 \text{ and } Q_x=51{\cdot}7$$

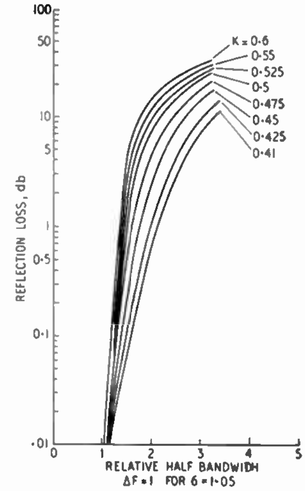
whereas the maximally flat filter would require

$$Q'_y=73{\cdot}4 \text{ and } Q'_x=30{\cdot}5.$$

Fig. 10 gives the reflection loss of a four-cavity filter for various values of k to the base of normalized half bandwidth. By comparison with Fig. 7 it will be seen that a four-cavity filter with $k=0{\cdot}5$ has a much better characteristics than a maximally flat five-cavity filter.

Unfortunately, a five and more cavity filter has two variables and so far it has not been feasible to overcome the large number of computations that are called for to arrive at an optimum design.

Fig. 10. Reflection loss of 4-cavity filters with various values of parameter k.



3.3. $\frac{3}{4}\lambda_g$, $\frac{1}{4}\lambda_g$ *or direct coupling*

With single post cavities it is not advisable to use $\frac{1}{4}\lambda_g$ coupling for reasons given earlier. It has been shown, however, that with three posts across the waveguide it is possible to reduce spurious mode propagation to an extent permitting the use of quarter-wave coupling. There results some saving in space, but at the cost of greater difficulty of manufacture. If necessary one could obtain empirically a curve for 3-post-obstacles corresponding to Fig. 6. The rest of the design procedure would remain unchanged except that the separation between cavities could be $\frac{1}{2}(\theta_1+\theta_2-\pi)$ which is, of course, exactly $\frac{1}{2}\lambda_g$ shorter than that given earlier.

A further saving in space could be obtained by using directly coupled filters. It has been shown that such filters can be derived direct from $\frac{3}{4}\lambda_g$ coupled ones by lumping together the two susceptances of the adjoining cavities into one, such that

$$b'=\tfrac{1}{2}b_1\sqrt{(b_2^2+4)}+\tfrac{1}{2}b_2\sqrt{(b_1^2+4)}$$

This, of course, assumes the obstacles to be pure susceptances and so is only possible in practice to a first order of approximation. Since the common obstacle couples two unequal cavities, it would be rather difficult to arrive at a suitable solution on the lines given in sect. 2.2. It must also be noted that b' is of the order of magnitude b_1b_2, which calls for large susceptances, virtually unobtainable with single posts. Herein lies one advantage of using three-post obstacles, which can give much higher susceptances without excessive post diameter. Of course, irises could also be used as obstacles. Cohn[7] has described recently a method of designing direct-coupled filters using inductive irises. He claims very good results even at bandwidths of 20 per cent. of centre frequency.

It is held by the present author, however, that the essential simplicity and ease of tuning of a single post, $\frac{3}{4}\lambda_g$ coupled filter by far exceeds in its value to the designer the cost of greater size and weight.

4. Phase Shift througth Waveguide Filters

In communication applications it is necessary to avoid excessive group delay changes in the equipment. Some group delay change may occur in the microwave filters, and it is useful to know what to expect. With the knowledge of obstacle susceptance and separation it is possible to compute the phase shift ψ at various frequencies from which the group delay $\partial\psi/\partial f$ can then be calculated.

Assuming pure inductive susceptances it can be shown by suitable manipulation of the matrices that the phase shift ψ is given by the relation

$$\tan\psi = \frac{b(b^2+2)\delta\theta - 4}{2b + 2(b^2+2)\delta\theta}$$

$$\text{where } \delta\theta = \frac{\delta f}{f_0}\left(\frac{\lambda_g}{\lambda_0}\right)^2 \tan^{-1}\frac{2}{b}$$

and δf is the change of frequency from design centre f_0.

It has not been possible to arrive at equally simple expression for ψ when the post is considered as a T-network; however, using the matrix given in sect. 2.2 and remembering that $\tan\psi = \dfrac{B+C}{2A}$, phase shift has been computed for various sizes of posts used in a cavity tuned to 4,024 Mc/s. It has been found, perhaps not surprisingly in view of the narrow band examined, that the computed phase shift based on the more accurate approach was the same as that given by the simple expression quoted above provided the value given to b is the shunt susceptance of the equivalent T-network.

Some results are given in Fig. 11.

Fig. 12 gives the results of measurements as compared with computations on single, three and five cavity filters. The measurements were carried out by using an adaptation of the nodal shift method described by the author elsewhere. There is a good degree of agreement and the curves show that the group delay change is not likely to be troublesome in this type of filter.

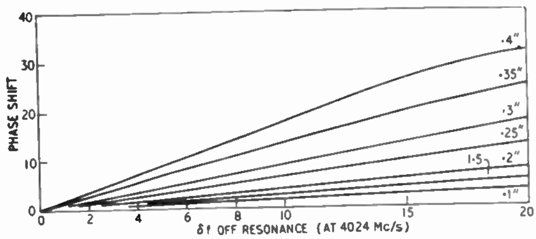


Fig. 11. Phase shift through single-cavity filters made up of two posts of various diameters.

5. Conclusions

It has been shown that a satisfactory design procedure for post-type filters results from the combination of theory and measurements. A good agreement between expected and actual results has been achieved and filters designed

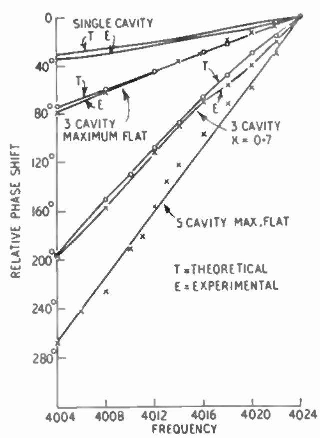


Fig. 12. Comparison between computed and measured phase shift through various filters.

entirely on paper at 4 and 7 Gc/s gave performances close to those predicted.

6. Acknowledgments

The author wishes to acknowledge the help and advice given to him by Mr. J. J. Rudolf, chief microwave engineer at Quy Laboratories of Messrs. Pye Telecommunications Ltd., and to thank Mr. A. D. Jenkins, engineer at these laboratories, who carried out a large part of the measurements.

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8. Appendix: Method of Finding the Exact Length for a Cavity Filter

Consider a shunt susceptance jb (say $b>0$) placed across a uniform waveguide. Let a short circuit be placed on one side of the susceptance at a distance s_1 equivalent to θ_1 electrical degrees and a slotted section be placed on the other side with the probe set to the voltage minimum which is then at a distance s_2 or θ_2 electrical degrees from the susceptance. It is assumed that the susceptance jb lies in the plane 0.

The transfer matrix of a susceptance jb inside two lengths of guide θ_1 and θ_2 is

$$M=\begin{vmatrix}\cos(\theta_1+\theta_2)-b\sin\theta_1\cos\theta_2, & j[\sin(\theta_1+\theta_2)-b\sin\theta_1\sin\theta_2] \\ j[\sin(\theta_1+\theta_2)+b\cos\theta_1\cos\theta_2, & \cos(\theta_1+\theta_2)-b\cos\theta_1\sin\theta_2\end{vmatrix}$$

It can be shown that if this circuit is short circuited at the output then the input impedance

$$Z_{insc}=j\frac{\sin(\theta_1+\theta_2)-b\sin\theta_1\sin\theta_2}{\cos(\theta_1+\theta_2)-b\cos\theta_1\sin\theta_2}$$

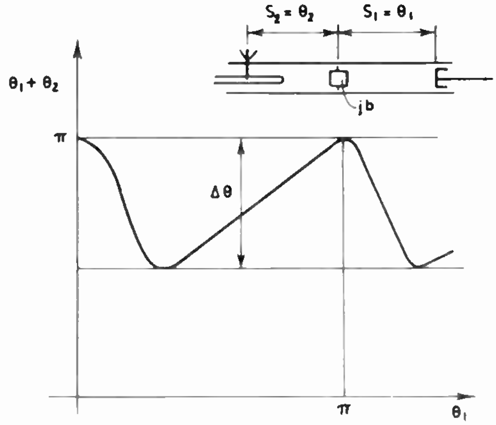

Fig. 13. Nodal shift curve due to a positive susceptance.

and if

$$Z_{insc}=0$$

then:

$$\sin(\theta_1+\theta_2)-b\sin\theta_1\sin\theta_2=0$$

or $\sin\theta_1\cos\theta_2+\cos\theta_1\sin\theta_2=-b\sin\theta_1\sin\theta_2$

hence $\cot\theta_1+\cot\theta_2=b$

which leads to $\tan(\theta_1+\theta_2)=\dfrac{b(1-\cos 2\theta_2)}{b\sin 2\theta_2-2}$

If now $\theta_1+\theta_2$ is plotted against θ_1 one obtains a curve as shown in Fig. 13. This has been studied by Meinke[9], who has shown that if the peak-to-peak size of this curve is denoted by $\Delta\theta$ then

$$\tan\frac{\Delta\theta}{2}=\frac{b}{2}$$

It will be noted that if $b>0$ then the peak of the curve coincides with $\theta_1+\theta_2=\pi$.

(If b were <0 the trough of the curve would satisfy this condition.)

Consider now an actual physical obstacle of finite thickness so that it can no longer be assumed as lying in the plane 0. In that case its equivalent network is no longer a pure susceptance, but it can be considered at one frequency as a pure susceptance inside two lengths of guide. If the obstacle is symmetrical around the plane 0 then the two lengths are identical, say l. This is shown in Fig. 14.

If now a short circuit were placed at a physical distance x from the plane 0 and the corresponding voltage standing wave minimum were found on the other side at a distance y, then by considering the equivalent circuit and referring to the earlier discussion and Fig. 13:

$$l+x=s_1 \text{ and } l+y=s_2$$

The resultant curve must be identical in shape with that in Fig. 13 and its peak in the co-ordinate system referred to the equivalent network must correspond to $(\lambda_g/2+n\lambda_g/2)$ where n can be any integer. But in the co-ordinate system referred to the physical arrangement, the peak will correspond to a set of distances

$$x_p \quad \text{and} \quad x_p+y_p$$

hence it follows that

$$2l=(n+1)\frac{\lambda_g}{2}-(x_p+y_p)$$

Also the peak-to-peak size of the curve Δl is related to the equivalent shunt susceptance jb by

$$\tan\frac{\pi\Delta l}{\lambda_g}=\frac{b}{2}$$

Now if a filter were to be made of two identical

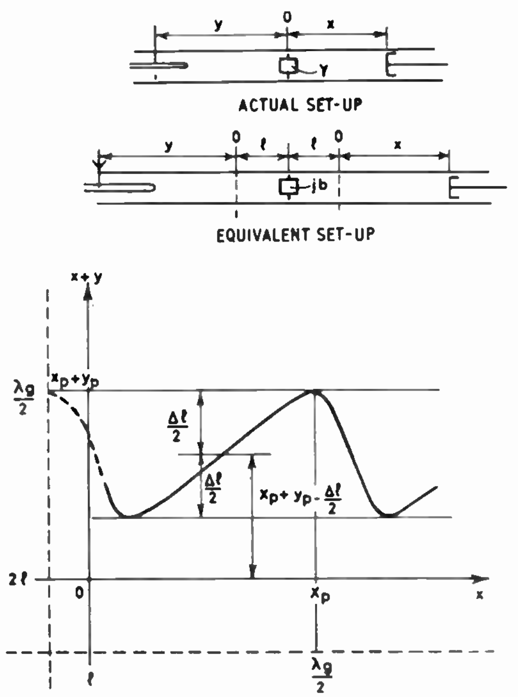

Fig. 14. Derivation of cavity length from nodal shift curve.

pure shunt susceptances jb placed at a distance l_0 apart, then it has been shown that

$$\tan\frac{2\pi l_0}{\lambda_g}=\frac{2}{b}$$

hence

$$\tan\left(\frac{\pi}{2}-\frac{2\pi l_0}{\lambda_g}\right)=\frac{b}{2}=\tan\frac{\pi\Delta l_0}{\lambda_g}$$

$$\text{or } l_0=\frac{\lambda_g}{4}-\frac{\Delta l}{2}\pm\frac{\pi\lambda_g}{2}$$

But the actual obstacle is equivalent to jb within an overall length of guide $2l$, hence the actual physical dimensions of the cavity will have to be

$$l_0-2l=\frac{\lambda_g}{4}-\frac{\Delta l}{2}-(n+1)\frac{\lambda_g}{2}+(x_p+y_p)$$

$$=\left(x_p+y_p-\frac{\Delta l}{2}\right)+\frac{\lambda_g}{4}\pm(n+1)\frac{\lambda_g}{2}$$

The last term in this expression merely serves to reduce the size of the cavity to one nearest to half wavelength. The term $\left(x_p+y_p-\frac{\Delta l}{2}\right)$ is clearly the median of the curve in the physical co-ordinate system.

Thus a correct dimension can be found to make a cavity at any frequency by carrying out the nodal shift measurement on one obstacle only. The results are very accurate provided the positions of the short circuit and slotted section probe can be read to $\pm 0{\cdot}01$ mm, which is not an excessive demand. Also the frequency must be kept well within ± 1 Mc/s of nominal.

An error of $\pm 0{\cdot}01$ mm causes an error in obstacle separation l_0 of $\pm 0{\cdot}04$ mm, while an uncertainty of ± 1 Mc/s in frequency setting causes an error in l_0 of about 0·01 mm, so that the computed separation may be out by about 0·06 mm or 1·5 thousandths of an inch, which is within the tolerances one has to allow for in manufacture and has to be taken up by a central screw.

It is worth noticing that the relative insertion phase shift can be found from the same plot as given by: $\psi=\pi-\bar{\theta}$

where $\bar{\theta}$ is the median of the curve in electrical degrees. This has been dealt with by the author in detail in a paper to appear shortly in *Electronic & Radio Engineer*.

MAGNETIC RECORDING OF TELEVISION PROGRAMMES*

Although the techniques of recording television programmes on film have been greatly improved over the years, there still remain the disadvantages of the relatively long delays involved in processing and the deterioration in quality due to the introduction of an intermediate stage in the sequence of recording.

The recording of the video signals on magnetic tape and their subsequent reproduction as electrical signals has thus many attractions: the major problem has, however, always been the large bandwidth to be reproduced—in the case of the British television signal of 3 Mc/s some two hundred times that necessary in the high fidelity recording of sound. This has implied a need for high tape speeds and extremely small gaps to enable the upper frequencies to be recorded.

Numerous solutions for reducing these technical problems to manageable proportions have been put forward: for instance, one equipment produced in the United States has achieved a tape speed of only 15 in./sec by the use of 2 in. wide tape and four heads rotating at 240 rev/sec, recording laterally across the tape. Such artifices, while assisting reproduction of the higher video frequencies, do not solve the difficulties of reproducing low frequencies, the wavelengths of which will be so long that small outputs will be obtained from the heads. Hence the solution has been to frequency-modulate the whole video band (0–4 Mc/s) on a carrier of 5 Mc/s, and to suppress the upper sidebands.

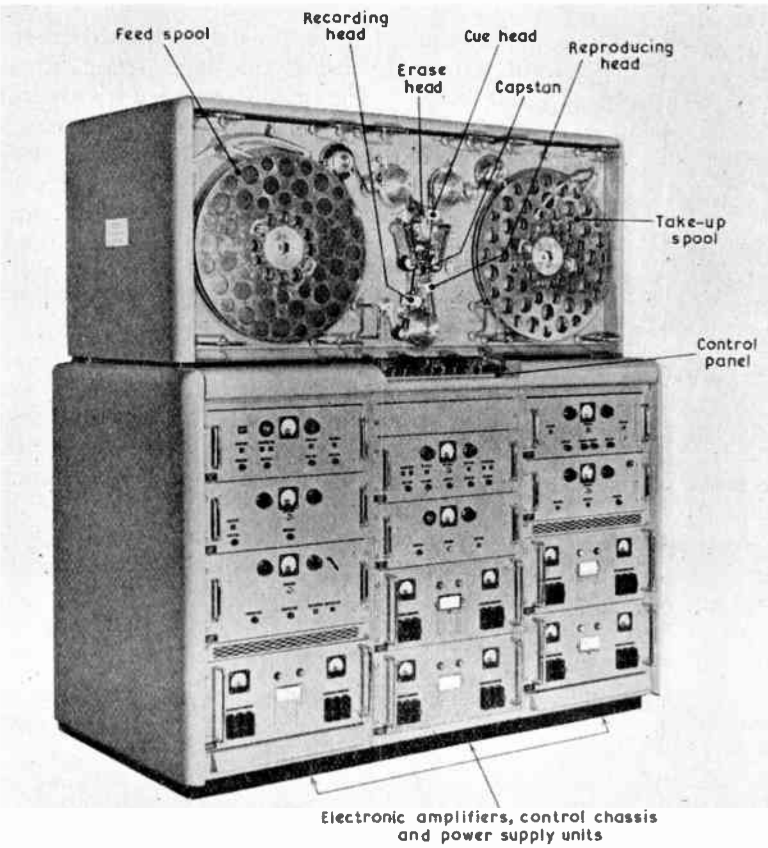

Fig. 1. Front view of the vision electronic recording apparatus.

The approach of the B.B.C. in its recently announced Vision Electronic Recording Apparatus — known as "VERA" — has been somewhat different, although some of the features of the American equipments have been used. A photograph of one of the machines is shown in Fig. 1. The channels which are to be put into service will consist of two such machines which can be controlled from a central control desk. The machine

* Based on information supplied by the B.B.C. Research Department.
U.D.C. No. 621.395.625.3:621.397.3

employs half-inch magnetic tape and a reel (20½ in. diameter) such as those shown in Fig. 1 will accommodate 15 minutes of programme. Continuous recording is possible by the use of two machines and the control desk. The tape speed employed in the present model is 200 in./sec and the magnetic tape used may be a normal thin-base sound recording tape of good quality. Re-wind time is 7 minutes.

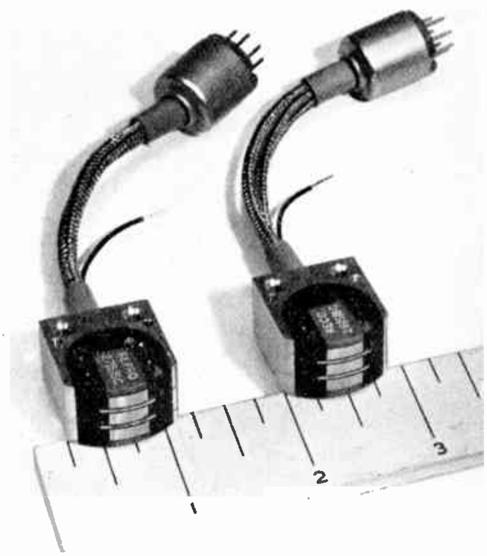

Fig. 2. The 3-track recording and reproducing heads.

The machine employs a three-track system of recording, two of the tracks being devoted to the storing of the video signal and one to the storing of the sound signal. Separate recording and reproducing head-stacks are employed, each stack containing three identical heads separated from each other by copper screens and aligned to the accuracy required in the manufacturing process. The assemblies are encapsulated in epoxy resin to ensure stability. Continuous monitoring of the recorded signal during the process of recording may be carried out.

The heads are illustrated in Fig. 2. They are completely interchangeable and useful lives of up to 100 hours are claimed; after this period the thin mumetal facing is worn down to the core. The designed gap between tape and head is of the order of one-twentieth of a thousandths of an inch (about 10^{-3} mm).

In the tape transport system in the machine most of the power required to drive the tape is supplied by the spooling motors which are arranged to move the tape past the heads at a speed just below the chosen recording speed of 200 in./sec and close to the constant tension required, even when the drive motor is not engaged. This result is obtained by varying the power fed to the spooling motors in accordance with (*a*) their torque/speed characteristic and (*b*) the amount of tape on the reels at any particular moment, the latter determining the speed of rotation required of the reels. When the drive is engaged the drive motor is, therefore, required to supply only a limited amount of power to bring the tape speed up to 200 in./sec. The drive is engaged by lowering two rubber idlers on to a common capstan so that a loop of tape, largely isolated from transient effects in the reels by these idlers and other mechanical filtering elements, is formed. Inside this loop lie the recording and reproducing head stacks. The erasing head is placed at a convenient point which lies outside the loop and precedes the recording head. A "velodyne" system of speed control and correction of the driving capstan is employed. During recording periods the servo driving motor is made synchronous with the mains frequency whilst on reproduction the output of the machine is frame-synchronized to station synchronization signals. The machine is fitted with the usual facilities for braking and for spooling the tape backwards or forwards at a variable speed when the drive system is not engaged.

A block schematic diagram showing the connections of the principal electronic units embodied in the machine is shown in Fig. 3. For storing the video signal the two video tracks are associated, on the recording side, with a band-splitting system in which the video signal is divided into two frequency bands of approximately 0–100 kc/s and 100 kc/s–3 Mc/s. The 0–100 kc/s video band is made to frequency-modulate a carrier and this frequency-modulated carrier is recorded on one track. The low-frequency content of the video

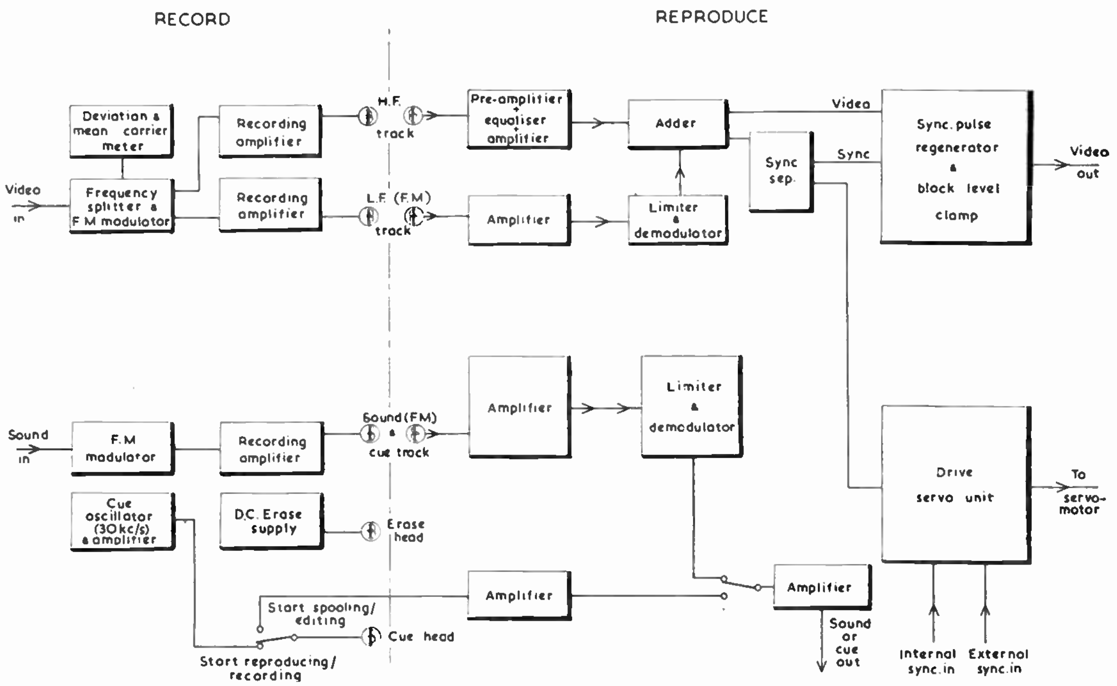

Fig. 3. Block schematic of the principal electronic units.

signal is thereby transferred to a frequency band corresponding to shorter wavelengths so that both the low-frequency and the long-wavelength difficulties inherent in the conventional magnetic-recording system are avoided. In addition, the amplitude-limiting facilities normally associated with the reception of frequency-modulated signals may be incorporated in the reproducing chain to eliminate undesired amplitude fluctuations and overcome almost all "drop-out" difficulties, even when employing thin-base sound recording tape not specifically manufactured for video or instrumentation purposes. The higher vision band, from 100 kc/s upwards, is recorded simultaneously on the second video track in a conventional manner.

On reproduction the output from the frequency-modulated video track is limited, demodulated, and added to the output from the higher-frequency track to reform the composite television waveform. Before transmission to line the synchronization information, including line and frame synchronizing signals and suppression periods, is extracted, reconstituted and added back into the video signal.

It is, of course, obvious that the higher-frequency video band, which employs a conventional recording/reproducing system, will be subject to the same unwanted amplitude-modulation which is being eliminated by the frequency-modulation system of the lower frequency video band. It is, however, an important finding that in practice this does not appear of major importance, for as long as the synchronization signals and the main brightness structure of the picture, represented by the 0–100 kc/s band of the video signal, are maintained intact, reasonable variations in the higher-frequency band do not noticeably degrade the subjective result.

All the low-frequency and long-wavelength difficulties which, in the case of the lower video frequency, are overcome by the use of the carrier system, will also be manifest in the sound channel if a conventional recording of the sound signal is attempted under the higher tape-speed conditions dictated by the video-signal requirements. The difficulties are, however, overcome by an identical technique to that employed to store the lower video frequencies. Accordingly the sound signal is, before recording, made to frequency-modulate another carrier which is recorded on the third track. On reproduction the carrier is limited and demodulated to provide a sound signal of high fidelity exactly synchronous in time with the video information reproduced from the other two tracks.

Simple editing, in the form of replaying extracts from a previously recorded programme, may be achieved by starting the machine at any predetermined point in the recording. This is possible because the machine is equipped with the usual facilities for spooling the tape backwards and forwards to find a desired point in the recording. The method may be extended, as in magnetic sound-recording practice, by cutting and joining extracts from various recordings or different parts of the same recording. Individual frames cannot, however, be examined in a "gate," as in optical film editing, for the tape must be reproduced at the correct speed before a picture can be reproduced on a monitor. A cueing arrangement for the "marking" of editing points has, therefore, been provided. The method adopted is to provide an extra cueing head, lying outside the isolated tape loop, which is fed through a separate recording amplifier from a 30 kc/s oscillator. When the tape is being normally reproduced and the observer wishes to mark some particular point for cutting or starting he presses a CUE key on the control panel of the machine which causes a 30 kc/s burst of signal to be recorded on the sound track of the tape. At this frequency it will not appear in subsequent normal reproduction since it lies well below the frequency-modulated carrier signals which carry the sound programme and any interference effects it might otherwise have will be removed by the limiting process which precedes detection of the television sound signal. However, when the tape is being slowly transported past the reproducing head, using the spooling speed control, at a fraction of the normal speed, the CUE signal will produce an audible note in the loudspeaker or head-phone system so that the point previously marked is found. The cutting and joining of tapes is accurately and quickly carried out by the use of a splicing device provided and the resultant join provides no visible disturbance.

Subjectively the quality achieved with the equipment at its present stage of development is high compared with tele-recordings. One advantage which the magnetic recording shows, compared with film recording, is that rapid movement does not blur, and this would seem to indicate that the decision to allow the upper band of frequencies, i.e. those which are recorded directly and not through the medium of frequency modulation, to suffer possible distortion is justified. The most critical factor in this type of recording is of course the maintenance of constant tape speed, and close examination of the recorded pictures reveals that a certain degree of improvement in the speed control circuits is still required, since a slight waviness of the frame edge is apparent.

From the comparative cost point of view, tape is considerably cheaper than film, even when the present higher cost of the recording and reproducing equipment is taken into account. One great advantage lies of course in the possibility of erasing an unwanted recording and using the tape again. This advantage also gives great possibilities in enabling checks of rehearsals, etc., to be made both cheaply and rapidly. The facility for monitoring the recording *1 frame* ($\frac{1}{25}$th second) after it has been made is also invaluable here and also assisted considerably in speeding-up the development work on the system.

The potentialities of the new system, which is due to come into regular use in the B.B.C. Television Service in the next few weeks, are thus very considerable, although it is emphasized that replacement of film techniques is not anticipated. Film recordings can, of course, be reproduced on other television systems whereas a magnetic recording is tied to the system standard from which it was taken. It is understood that the B.B.C. Research Department, which was responsible for the development of this equipment, see no fundamental reason why it should not be developed along its present lines to enable recordings of systems with wider bandwidths to be made, such as C.C.I.R. or similar systems, or N.T.S.C. colour.

DECTRA: A LONG-RANGE RADIO-NAVIGATION AID*

by

C. Powell†

Read before a meeting of the Institution in London on February 26th.
In the Chair: Air Marshal Sir Raymund Hart, K.B.E., C.B., M.C. (Member).

SUMMARY

The paper first outlines the operational requirement for a navigational aid covering the air routes on an ocean crossing such as the North Atlantic. After a short reference to the Decca phase comparison technique, the tracking and ranging functions of the Dectra system are described with special reference to the time-sharing technique on which the tracking pattern is based. Details are given of the transmitting stations which are common to the Dectra and Decca services, and of the associated airborne receiving and display equipment. A note is included on the prediction of the accuracy and performance of the system. Details of flight trials are summarized in an Appendix.

1. Introduction

Dectra (Decca Track and Range) is a derivative of the Decca Navigator position fixing system for ships and aircraft‡ and is designed primarily to cover long air routes such as those running across the North Atlantic. In that region the volume of traffic is now such that at any given moment there can be well over one hundred aircraft in flight; at present, in the absence of accurate position information, the risk of collision can only be eliminated by widely separating the aircraft in all three dimensions. Prominent among the problems and delays which this separation introduces is the fact that aircraft are frequently prevented from flying at or near the optimum altitude for operating efficiency.

Although Dectra is mainly intended as a solution to the air navigation problems just stated, in which role it would function rather as the basis of an air traffic control system than simply as a source of randomly-used position-fixing data, it can also be used by ships. A large area around the principal trans-ocean air routes is served, an essential feature of any position-fixing system for an air crossing such as that of the North Atlantic being the ability to cover emergency diversions to points remote from the prescribed tracks.

Some of the Dectra equipment is common with that of the Decca Navigator system, in particular the ground stations and the Flight Log pictorial display§ in the aircraft. The latter is important on payload considerations and the use of common transmitting station sites introduces a common element in the geometry of the two systems and facilitates the procedure for changing from one to the other when flying near the route terminals.

While the practical results already gained with the system suggest that no fundamental modification in technique is likely to be required, changes in equipment may be expected to be made in the course of normal evolution and as operational experience is gained. Accordingly this paper is intended rather as a general outline of the Dectra system and of the methods it uses, than as a definitive description of the transmitting and receiving equipment. The results of operational trials are summarized in an Appendix.

* Manuscript received 4th December 1957. (Paper No. 452.)

† The Decca Navigator Co., Ltd., 247 Burlington Road, New Malden, Surrey.

U.D.C. No. 621.396.93

‡ W. J. O'Brien, "Radio navigational aids", *J.Brit.I.R.E.*, 7, pp. 215-248, October 1947.

§ G. E. Roberts, "The design and development of the Decca flight log", *J.Brit.I.R.E.*, **12**, pp. 117-131, February 1952.

2. General Principles

2.1 *The Discriminator*

Since frequent reference is made later in this paper to the phase discriminator circuit which forms the basis of the Decca and Dectra phase comparison technique, the nature of this circuit should first be recalled. Fig. 1 (*a*) is a simplified circuit diagram of the discriminator, in which two alternating voltages V_a and V_b are phase-compared. The phase difference angle thus derived is directly displayed by the associated Decometer indicator.

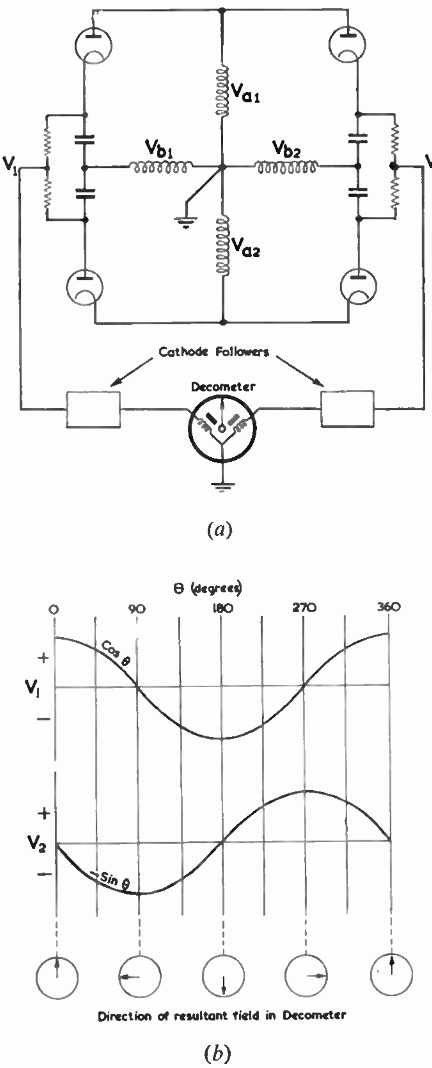

Fig. 1. (*a*) Basic circuit of Decca phase discriminator.
(*b*) Relationship of discriminator output voltages to phase difference angle θ.

One of the signals to be compared (say V_a) is split into two voltages V_{a1} and V_{a2} of equal amplitude but opposite phase, by the use of a centre-tapped transformer coil. The other signal V_b is split into two parts of equal amplitude but in phase quadrature, being derived from the primary and secondary of a double tuned transformer at resonance. These four voltages are applied to the discriminator as shown in the diagram, so that to each diode is applied the sum of two signals of which one is derived from V_a and the other from V_b. The rectified output from each diode is proportional to (and approximately equal to) the vector sum of the two input signals. The net rectified output voltages V_1 and V_2 from the two sides of the discriminator can be shown to vary in the manner indicated in Fig. 1 (*b*) with changes in the angular difference between the phases of the signals V_a and V_b. The output voltages V_1 and V_3 are applied through linear d.c. amplifiers (cathode followers) to respective field coils in the Decometer. The coils are at right-angles and a permanently-magnetized disc aligns itself with the resultant field and turns a pointer to indicate the phase difference angle θ as in Fig. 1 (*b*). Inasmuch as the output voltages V_1 and V_2 can be regarded as proportional to cos θ and sin θ they will be referred to as the cosine and sine outputs of the discriminator. The sine output of a circuit of this type is used in the Dectra receiver to control the phase of a locked oscillator.

2.2 *The Dectra Chain*

Figure 2 represents a Dectra chain of three transmitting stations in which stations A and B combine to generate a "tracking" pattern of hyperbolic position-lines running roughly in the direction of the proposed routes. Station C, at the other end of the crossing, generates in conjunction with the common master station A a second pattern intersecting the first at an obtuse angle and so provides the "ranging" co-ordinate of the Dectra position fix. Thus Dectra is essentially an area-coverage system, with its service area centred on the region through which the main routes to be served will pass.

2.3 *Tracking*

To generate the tracking pattern, station A in Fig. 2 transmits a stable continuous-wave signal of frequency F_1, (about 70 kc/s) continuously except for periodic interruptions lasting a fraction of a second. During each such break, station B transmits a signal of the same frequency F_1, phase-locked to the transmission from A; measuring the phase difference between the two signals in the airborne receiver will locate the aircraft on one of a family of lines of constant phase difference, thus providing a navigational position-line in the form of a hyperbola with A and B as foci. Position lines corresponding to a given phase-difference value will recur at intervals which, measured along the baseline A B, are constant and equal to half a wavelength at frequency F_1. The intervals between adjacent hyperbolae of the same phase-difference value are known, as with conventional Decca, as "lanes".

The transmission from B takes place every few seconds and the position-line indication is given either in terms of a conventional Decometer reading and/or as one of the two co-ordinates of a Flight Log automatic plotter.

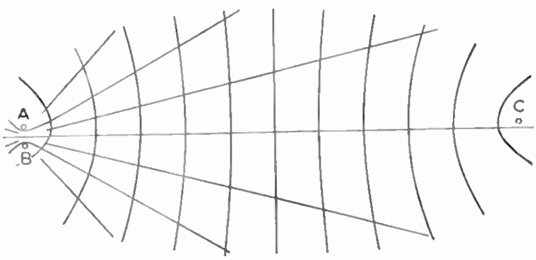

Fig. 2. Patterns produced by three-station Dectra chain.

The use of the same carrier frequency for master and slave by time-sharing the transmissions has certain practical advantages in this instance and represents a departure from normal Decca practice, in which different but harmonically-related carrier frequencies are employed so that master and slave signals can be multiplied in the receiver to a common harmonic value for phase comparison. The basic arrangement of the tracking section of the Dectra receiver is shown in Fig. 3 (*a*) and (*b*). While the master is transmitting, its received signal controls the phase of an oscillator of frequency F_1 as shown at (*a*). During the short

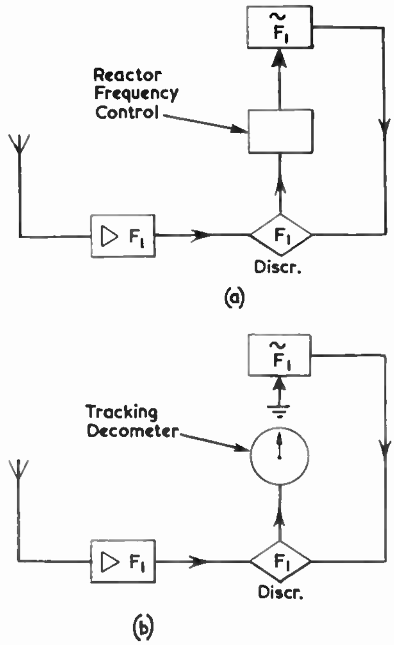

Fig. 3. (*a*) Arrangement of tracking section of receiver during master transmission.

(*b*) Arrangement of tracking section of receiver during tracking-slave transmission.

slave transmissions, the output of the phase discriminator is switched as in (*b*) from the oscillator to the actual tracking display and the latter then indicates the phase difference between the oscillator output (representing the master phase) and the slave signal.

2.4 *Ranging*

The ranging hyperbolic pattern is generated by stations A and C (Fig. 2) whose respective frequencies F_1 and F_2 are each an integral multiple of a common sub-harmonic F_3 of approximately 150 c/s, such that

$$F_1 = nF_3$$

$$F_2 = (n+1)F_3$$

Phase comparison of the slave station C with the master A for the necessary master/slave phase locking, and between A and C at the receiver for deriving the ranging position line indication, is effected at the common frequency F_3. A signal of frequency F_3 emanating from the master station is obtained by feeding the received signal into a ($\div n$) divider, and for station C by subtracting F_1 from F_2 in a mixer circuit. It can be shown that the lanes to which the ranging Decometer responds are of a width corresponding not to the frequency F_3 at which phase is compared, but to the "undivided" frequency F_2. This can be clarified by assigning arbitrary values such as 2 kc/s to F_1, 3 kc/s to F_2, 1 kc/s to F_3 and letting $n=2$. Suppose

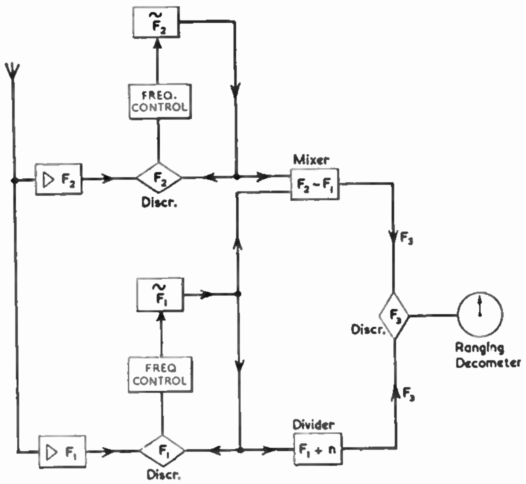

Fig. 4. Arrangement of ranging section of receiver.

now that an observer located somewhere on the interstation baseline AC moves along that line towards station C through a distance equal to half a wavelength at C's frequency F_2, the phase of the F_2 signal at his receiver will be advanced half a cycle while that of the F_1 signal from Station A will be retarded one-third cycle. In the ranging section of the receiver (Fig. 4) the mixer output F_2-F_1 will therefore show a phase change of + five-sixths cycle at frequency F_3 and the phase of the divider output $(F_1 \div n)$ a change of − one-sixth cycle at that frequency. Comparing the mixer and divider outputs in a phase discriminator at frequency F_3 will thus indicate a change of one cycle, so that in making one complete revolution for a movement along the baseline of half a wavelength at frequency F_2 the Decometer is said to record the passage through one lane corresponding to that frequency. Geometrically, the ranging hyperbolic pattern therefore behaves as if phase-locked signals of frequency F_2 were radiated from points A and C.

2.5. *Frequencies*

Typical values of F_1, F_2, F_3 and n are shown in Table 1.

3. The Transmitting Stations

The two stations generating the Dectra tracking pattern are integral with two of the four stations of the "Moose East" Decca Navigator chain in eastern Newfoundland. As indicated in Table 1, a Decca frequency in the "Purple" group is employed and the Dectra master signal is in fact the Purple slave transmission of the Decca chain; the Dectra tracking slave transmissions are the successive short bursts of "Purple" frequency radiated from the Decca master station as part of the normal Decca lane identification service.

Table 1

Frequencies and related data for the North Atlantic Dectra Chain

Function	Frequency
Master and Tracking Slave transmissions	F_1 = 70·3842 kc/s (Decca frequency 2, Purple)
Ranging Slave transmission	F_2 = 70·5375 kc/s (Decca frequency 3, Purple)
Ranging discriminator frequency	F_3 = 153·342 c/s
Division ratio	n = 459

Lanewidths on baselines assuming c = 299,700 km/sec.

Tracking	2129·035 metres	(2328·342 yards)
Ranging	2124·402 metres	(2323·276 yards)

It will be recalled that lane identification involves the generation once per minute of a coarse hyperbolic pattern confocal with each of the three fine patterns in turn, the necessary low comparison frequency *f* of about 14 kc/s being derived by momentarily transmitting signals of Green (9*f*) and Red (8*f*) from the slave whose pattern is to be lane-identified, and master (6*f*) and Purple (5*f*) from the master station. Mixers in the receiver extract signals of frequency *f* for the master and for the slave, which when compared in a phase discriminator provide a hyperbolic position line in the coarse pattern which serves to indicate the lane in which the receiver is located. This enables the normal integrating function of the Decometers to be correctly initiated when the user is uncertain of his starting point, as when entering the coverage of a chain from the open sea, and verified in the event of an interruption in reception.

Inasmuch as Dectra makes use of frequencies which are already radiated for Decca purposes, and which are held in phase synchronism as part of the normal function of the Decca slave equipment, the only major change at the transmitting stations made necessary by the introduction of Dectra is a large increase in the power of the Purple-frequency transmissions so as to secure the required long-range coverage. To this end the appropriate transmitters have been increased in power from the normal 2·4 kW to 12 kW and provision has been made for a further increase, should operational experience indicate that this is necessary. At the same time, transmitting aerials are provided at the two sites having heights of 620 and 450 feet for the Dectra master and slave stations respectively. These aerials are used for the Dectra frequencies only and are supplemented by conventional 165-foot horizontal-top aerials, such as are used on several Decca Navigator chains, to handle the Decca frequencies.

Taking as an example the Dectra tracking slave station, there is a physical separation of about 50 feet between the single-mast "umbrella" aerial carrying the Dectra transmission (at 70 kc/s) and the T aerial transmitting the 84 kc/s Decca master signal. Care has to be taken to avoid losses in either aerial circuit resulting from the proximity of the aerials. The inter-aerial capacitance forms part of the aerial tuning circuit, which can be represented in the form of a bridge as shown in Fig. 5. The inter-aerial capacitance C_e is

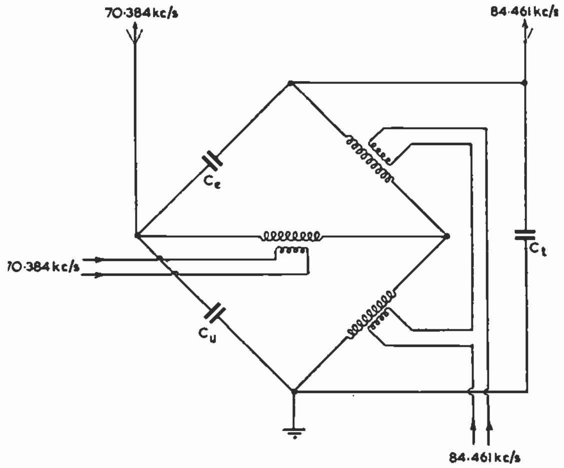

Fig. 5. Basic aerial tuning circuit at Dectra slave station.

in the region of 400 pF, the capacitances of the Dectra and Decca aerials C_u and C_t being respectively about 4000 and 3000 pF. The elements of the bridge, which include the aerial coupling coils and in particular the pair in series across the 84 kc/s output feeder, are so adjusted as to balance-out the voltages at 84 kc/s across the inductance to which the 70 kc/s input is coupled, and also to balance out the 70 kc/s voltage across the Decca aerial.

The only changes in the transmission sequence occasioned by the addition of Dectra to the Decca chain are the repetition of the "Purple" transmissions from the master (i.e. the Dectra slave) at more frequent intervals than the three per minute called for by Decca lane identification. Provisionally a cycle of eight such transmissions per minute has been adopted with eight corresponding breaks in the Dectra master transmission. In addition, each such break in Dectra master transmission is preceded, as shown in the transmission sequence diagram at Fig. 6, by a discrete break lasting about 0·1 seconds. The function of this initial break in master transmission is to operate the trigger circuits in the tracking section of the receiver, as described later.

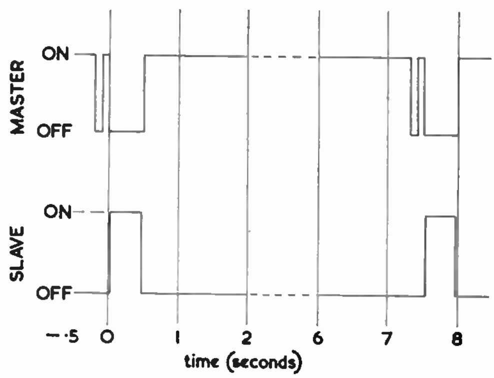

Fig. 6. Master and slave transmission sequence.

Bringing the tracking slave transmission into operation coincidentally with the main break in master transmission has been effected in early trials of the system by the action of the initial break upon triggering circuits, as in the airborne receivers. In the final version of the equipment, however, this operation will be carried out by a self-contained timing device at the slave station comprising a clock monitored by a 14 kc/s crystal oscillator. The achievement of the necessary timing accuracy for correct operation of the sequence shown in Fig. 6 presents no particular problems and this type of synchronization renders the slave station triggering independent of external interference. The onset of the slave transmission is timed to take place about 0·05 seconds after the initiating signal to preclude any possibility of the slave transmission masking the triggering signal through variation in relay operating speed or other causes.

4. The Airborne Receiver

The Dectra receiver is rather smaller than the conventional four-channel Decca Mk. VII or VIII. The case measures 15½ in. deep by 11¾ in. wide and 7¾ in. high and is mounted in standard S.B.A.C. racking. The power consumption is approximately 10 amperes at 24 volts. The reference oscillator and its associated circuits, hitherto included in all but one type of Decca receiver as an internal phase datum and to provide a common reference facility for all users of a chain, is not required in the Dectra airborne equipment; the use of a single time-shared frequency (and hence a single r.f. channel) for the tracking pattern eliminates the problem of differential phase drifts in the r.f. channels the correction of which is the main function of the reference system. In the ranging section of the receiver, where two radio frequencies are employed, the magnitude of the phase errors in the equipment itself is negligible in relation to the accuracy requirement in the ranging co-ordinate.

The basic arrangement of the receiver shown in Figs. 3 and 4, indicates two points of interest: the use of locked oscillators in both r.f. channels, and the time-sharing of frequency F_1 by intermittently switching the output of the phase discriminator from the oscillator phase control circuit to the tracking Decometer or Flight Log display. Before considering these it should be noted that the crystal filters in the r.f. channels have a narrower pass-band than the normal ± 30 c/s at -6 db, being of approximately ± 10 c/s at this level, so as to secure the maximum possible discrimination against noise, while retaining acceptable phase stability, in view of the long operating ranges involved. Two crystal filter circuits are employed in two successive stages in each channel. The sensitivity of the r.f. amplifier is such that the a.v.c. starts operating if the input signal level exceeds approximately 0·5 microvolts.

4.1. *Locked Oscillators*

An important feature of the receiver is the incorporation in each r.f. channel of an oscillator locked in phase to the respective incoming signal. The function of the r.f. amplifiers is solely to control the phase of the oscillators and the latter therefore constitute noise-free signal sources for the succeeding sections of the receiver. Furthermore, the oscillators have a very high frequency-stability and so provide an inertial characteristic which effectively reduces still further the overall bandwidth of the receiver, enabling it to override short breaks or phase variations in the incoming signals. Crystal oscillators are used, oven-controlled at a constant temperature within about $\pm 0{\cdot}1°$ variation. The operating temperature is approximately 60°C, which is some 15° higher than the turn-over point in the temperature characteristic of the crystal used.

Over periods of a few minutes the frequency stability of the oscillator remains in the order of 1 part in 10^7 so that the oscillators in the respective channels will continue to represent the phase of the incoming signals with small errors even in conditions when the r.m.s. level of noise peaks exceeds that of the wanted signal. The frequency control on the oscillator is applied through a reactor (Miller effect) valve which applies a capacitive reactance across the crystal circuit of a value depending on the d.c. bias voltage derived from the sine output of the controlling phase discriminator. The reactor circuit is so arranged that, for full discriminator output, only a moderate degree of instantaneous control is exerted, amounting to about ± 0.02 c/s. The response to greater changes in frequency is very much slower, about 1 c/s in 200 seconds. This prevents the phase of the oscillator suffering violent changes under heavy noise interference. The time-constant must not, however, be too long for the circuit to follow the Doppler frequency-change arising from any movement with respect to the stations that can occur in practice. Under extreme conditions a current type of jet transport aircraft flying at 40,000 feet could carry out the 90 deg. change of course from the directly-approaching to the tangential heading in 27 seconds. Assuming a speed of 0·3 km/sec and a propagation speed of 300,000 km/sec, the phase control for the 70 kc/s oscillator would have to handle a change of 0·07 c/s in 27 seconds, or about 0·0025 cycles/sec/sec, if it was considered necessary that there should be zero phase lag in the oscillator at the end of this drastic manoeuvre; in practice, of course, a lag can be accepted for short periods provided that it does not at any time exceed half a cycle, in which event the oscillator would lose a complete cycle of phase.

4.2. *Triggering*

Figure 7 is a block diagram representing the essentials of the tracking section of the receiver, with particular reference to the time-sharing switching and its triggering by periodic breaks in the master transmission. The triggering signal shown in Fig. 6 is repeated at (*a*) in Fig. 7 and comprises a break in the master transmission lasting 0·1 seconds, followed by a transmission of the same duration which is followed in turn by the 0·5 second break during which the slave station transmits in place of the master. In the receiver the initial 0·1 second break and the onset of the "main" break together comprise a signal which when rectified and differentiated is of the form shown at (*c*).

The differentiated waveform is passed to two multi-vibrator delay circuits each preceded by rectifiers, the latter being connected in opposition; the upper delay circuit in the diagram is triggered by the first (positive-going) pip of the signal and opens part of a valve gate circuit 0·19 seconds later as in (*d*). In the diagram the gate is represented by two switches in series actuated by the appropriate delay circuits: in practice it is a valve whose grid and suppressor are brought above cutoff by the two inputs. The other delay circuit responds to the negative-going differentiation pip at $-0{\cdot}1$ seconds (waveform (*e*)) and opens the remaining section of the gate after a delay of 0·09 seconds. Thus the two sections of the gate open simultaneously but only if the triggering signal is received in its entirety. The gate remains open for approximately 0·05 seconds; the operative trigger pulse (i.e. the second positive-going differentiation pip) admitted during this period to the switching circuit is of approximately 2 milliseconds duration.

The possibility of random triggering by noise pulses is reduced by this arrangement to the comparatively remote chance of successive pulses occurring with the correct relative sign and spacing, or of a strong c.w. interfering signal having a periodicity closely approximating to 0·2 seconds. As a further precaution against noise the minimum and maximum amplitude of the received trigger signal are limited prior to differentiation. The output of the r.f. amplifier with noise superimposed is represented by the waveform (*b*) in Fig. 7 in which the upper dotted line indicates the bias level of the detector; at amplitude less than the level of the lower dotted line, the short-grid-

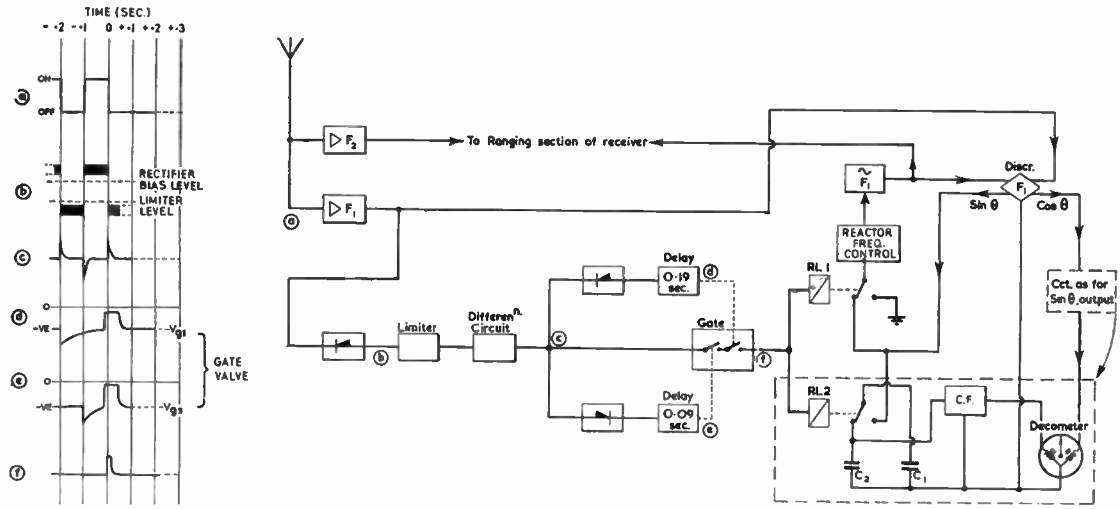

Fig. 7. Block diagram of tracking section of receiver, with special reference to time-sharing triggering circuits.

base limiter valve cuts off and in consequence only the central range of trigger signal amplitudes, relatively clear of the superimposed noise peaks, is accepted for differentiation.

The gated trigger pulse is passed to two pairs of valve/relay circuits, one pair for the output of the phase discriminator corresponding to the sine of the phase difference angle θ and the other for the cos θ output. The two halves of the discriminator/Decometer circuit are identical and only the sin θ section is shown in the diagram. The valve/relay circuits are of a form common in Decca practice in which, after receipt of the initiating pulse, the relay is held operative for a predetermined period with very small variations in this period through changes in the h.t. voltage applied to the associated valve. One such circuit, shown for simplicity as a relay RL1 in Fig. 7, earths the input to the reactor valve controlling the phase of the oscillator (which in normal operation will already be at or near the earth potential, indicating correct locking of the oscillator with the received F_1 signal) for a period of 0·5 seconds. The second relay RL2, with a hold-on period of 0·4 seconds, connects the output of the discriminator to the capacitor C1. During the ensuing slave transmission this capacitor acquires a charge proportional to the sine of the phase difference angle θ between the slave signal and the oscillator output representing the master signal. After 0·4 seconds the relay becomes inoperative and the capacitor C1 is connected to C2, so applying to the sine coils of the Decometer, through the normal cathode-follower output circuit, a voltage proportional to sin θ. Repeating this whole process for the cosine output of the discriminator provides a resultant magnetic field in the Decometer with which the rotor disc aligns itself, moving a pointer to indicate the tracking lane reading or actuating the primary servo in the tracking co-ordinate of a Flight Log as described later.

During the master transmissions, which occupy much the greater part of the transmission cycle, the high input impedance of the cathode follower enables the capacitors C1 and C2 to retain their charge, so preserving a stable Decometer reading pending the next slave transmission. During the slave transmissions the cathode follower remains connected only to C2 so that variations in the discriminator output due to residual noise or short-term phase disturbances do not appear at the display and the latter obtains the phase-difference information, after each slave transmission is completed, from the "clean" source represented by the net charge acquired by C1. The difference in hold-on time for the two relays ensures that reasonable variations of the switching sequence shall not disturb the stored information on the cathode followers.

4.3. *Receiver Controls*

After switching on, the oven containing the crystals for the two oscillators and the four filter circuits takes about ten minutes to reach operating temperature. Thereafter, once the display has been set to the correct co-ordinates, the receiver requires no further manipulation. The control box contains a switch with three positions labelled ST, REF and OP. In the ST or start setting, the oven is warmed up and the initial frequency-locking (prior to phase-locking) of the oscillator takes place. Correct oven temperature is indicated by a flashing lamp operated by the oven temperature-control relays under the stable cycling condition. While the oscillators are in the process of being frequency-locked, which is effected by applying in the ST switch position a frequency-sensitive control voltage to the reactor circuit associated with each oscillator, the Decometers will rotate even if the receiver is stationary until correct locking is achieved. The latter condition is also indicated by the main malfunctioning-warning lights going out. These are operated by relays in response to the cosine output voltages of the respective discriminators; with a steady output of more than −10 volts from the cosine side of the discriminator averaged over about one second, (corresponding to the condition of correct locking of the respective oscillator by the sine output voltage from the discriminator) the lamps are turned off. Under partial locking conditions they flash in response to the successive cycles' change in oscillator phase, and in the absence of signals the respective lamp remains on continuously.

In the REF position, the operation of the phase-control circuits reverts to normal but the sine and consine inputs to the tracking Decometer circuit are connected directly to the output of the associated phase discriminator. With correct phase locking there will be zero volts from the sine output of the discriminator, forming a datum against which the stator of the Decometer can be rotated to produce a tracking reading of zero. When the same Decometer is used for Dectra tracking and Decca "purple," this eliminates the possibility of an index error on Dectra occurring through the stator having been turned from the true zero reading in the process of "referencing" the Decca receiver.

4.4. *Receiving Aerial*

As in normal Decca practice, a "suppressed" aerial is used, comprising a collector-plate flush with the aircraft skin or located behind a dielectric portion of the skin. Two aerials of the latter type have been used in the initial flight trials referred to in the Appendix: one was a circular disc of 280 square inches area in the belly of the aircraft, and the other a metallic foil sheet of 80 square feet forming an inner skin in the "radar nose." The smaller aerial has been found to give adequate signal pickup up to at least 1500 n.m. An essential element in the airborne receiving aerial, of whatever type, is a head amplifier mounted close to the collector-plate, acting as an impedance transformer and having sufficient gain (about 25 times in the Dectra installation) effectively to eliminate noise pickup on the feeder to the receiver.

4.5. *Meter Presentation*

The Dectra position-line information appears as two meter readings which can be plotted on a hyperbolic lattice chart. The Decometers are provided mainly for setting-up and checking purposes and to facilitate changing from Dectra to normal Decca or vice versa when near the route terminals. From the pilot's point of view the information appears as the position of a servo-driven marking stylus on a chart in a Flight Log automatic plotter, as later described.

The tracking Decometer is of the Decca Mark VII Purple type and the purple meter of the aircraft's Decca receiver can serve both purposes. The lane scale is divided into 30 lanes and marked in the conventional Purple lane numbers from 50—80, the gearing between the fractional and lane pointers being such that the latter shows a change of 30 lanes for six "Mark VII" (slave-frequency) or Dectra lanes. A Purple Decometer is employed since the Dectra tracking pattern is confocal with the purple pattern of the Decca chain of which the Dectra stations are part and is derived from the same comparison frequency. The Decometer for the ranging pattern makes one revolution for each lane corresponding to frequency F_2 as already described. Here there is no coincident Decca pattern and the ranging lanes are numbered 0—2,000, the ranging Decometer lane pointers indicating units, tens and hundreds of lanes up to the total of 2,000.

4.6. *Pictorial Presentation*

The Flight Log presents the Decca or Dectra fix pictorially direct to the pilot, records the track flown and enables any desired track (such as the trans-Atlantic route assigned to a particular aircraft) to be followed by steering so as to keep the stylus on the prescribed line. The design and development of this instrument and its associated servo mechanism have been fully described elsewhere,* the following short description being included in order to complete the general picture of the Dectra airborne installation.

The servo system now generally used is based upon impulse or "stepping" motors. The primary element comprises three Decometer movements, one for each colour, which accept the receiver output. In each meter, the normal fractional pointer is replaced by a platinum-iridium contact arm; a slight angular movement of this arm drives it against one or other of a pair of contacts on a disc mounted coaxially with the Decometer, the contacts being connected through slip rings and relays to the forward and reverse windings of an impulse motor. Each time the arm touches one of the contacts the motor moves on or back one step under the control of a mechanical escapement. The impulse motor shaft moves a wiper arm over a series of studs arranged in a ring, and as the arm moves onto a stud it completes the circuit for a second impulse motor in the display head itself which moves the marking stylus or pen in, say the X co-ordinate. Repeating this mechanism for Y co-ordinate completes the basic arrangement, producing a system in which the pen automatically and continuously plots the Decometer information and hence records the track of the aircraft on the "inverse lattice" chart.

Since the instrument will be required to handle charts of many different map-scales, particularly when displaying Decca as opposed to Dectra information, so as to cater for the contingencies of precise air traffic control and airfield approach in limited areas as well as en-route navigation over relatively long distances, the equipment linking the receiver with the Flight Log display head must include a number of alternative input-to-output ratios. This equipment, collectively known as the "computer", must also contain means whereby the correct ratios as well as the correct patterns and senses of rotation for the required area can be set-in instantly. The charts themselves generally form a roll housed within the display head, a portion of the roll measuring about 4 in. by 10 in. being visible; the roll is carried on sprocketed rollers and this movement forms one of the two plotting co-ordinates. The second co-ordinate is provided by the lateral movement of the marking pen on a lead screw.

To reduce manipulation as far as possible, each different chart has a corresponding metal key assigned to it on which the number and position of the wards represents the scale, choice of colour, sense of rotation and other characteristics of the chart; all the keys for a particular chart roll are inserted in a multi-contact switch assembly (the "turret" switch) a single movement of which brings the desired key into operation. The wards of the selected key engage with the corresponding fixed contacts and complete the appropriate relay and other circuits necessary to adjust the computer for the characteristics of the required chart. Five principal Flight Log charts are provisionally allocated for the trans-Atlantic Dectra chain, the overlap between them being so arranged that only two are used on a given crossing, reducing the amount of chart-changing in flight to the minimum. On the reverse side of each chart, a series of range and bearing lines with respect to diversion points (such as the Azores) is drawn which become visible when the back-lighting in the Flight Log display head is turned on. This enables the aircraft to be placed on the optimum diversion track as soon as an emergency arises; an additional small-scale chart, covering the whole service area of the chain and all diversion points, forms part of each roll.

5. Performance

The original predictions of the performance of the Dectra tracking patterns were based largely on the results of long-range Decca trials and in particular on a series of observations by

* G. E. Roberts, *loc. cit.*

J. Th. Verstelle of the Dutch Hydrographic Office on a sea voyage from Denmark to Greenland in May 1950.* These observations comprised a very large number of readings taken by day and night of the normal Decca patterns from the English and Danish chains of stations, and also of the so-called "false" patterns which exist during lane identification. An example of a false pattern is that produced by the signals of master and purple slave frequency radiated together from the master station aerial three times per minute to form the master lane identification signal: geometrically speaking, a Decometer responding to this transient purple pattern should read 0 deg of phase difference everywhere in the coverage. Similarly, phase-locked red and green transmissions are sent from each slave in turn for lane identification and a suitably modified receiver can derive a nominally constant phase-difference reading from each of these.

The behaviour of the false patterns can be a useful index of propagation conditions, since the phase difference value changes only as a function of these conditions and not with the observer's location. Also, the lane identification transmissions provide data on the position-line accuracy that can be expected from a hyperbolic pattern generated by signals transmitted at the same carrier frequency for master and slave; for example, the Decca Purple slave station radiates at 70 kc/s signal as its normal function between the lane identification transmissions, and during the latter a signal of the same frequency, phase-locked with the first, is momentarily transmitted from the master. Given a means of "storing" the intermittent slave signals so that they can be phase-compared with the master in the receiver, a hyperbolic pattern of the type use for Dectra tracking is brought into being. The use of a common frequency for master and slave offers an important advantage in long-range operation, which depends mainly on skywave modes of operation, since it reduces the differential variations on the two transmission paths that would occur if different carrier frequencies were used.

Verstelle's results indicated that at distances up to 1200 miles from the originating stations, phase variations through admixture of the signals received by different modes of skywave propagation would be very unlikely to exceed a total spread of $\pm 0{\cdot}25$ cycles at 70 kc/s; on this basis the contours of 5 and 10 n.m. error at 95 per cent probability were drawn for the double-ended layout (Fig. 8) originally proposed for the chain, which used a tracking pattern at each end of the crossing. Properly speaking these contours refer only to the accuracy of tracking and as such would be expected to be circular; the departure from this form is due to the assumption in drawing the diagram that there will be a considerable increase (arbitrarily taken as two-fold) in the tracking error in regions more than 45 deg each side of the axis of the pattern, due to the increased separation between the skywave-reflection points on the master and slave transmission paths.

The prediction of the accuracy of ranging is difficult because of the strong probability that complete lanes will be "lost" during the course of a crossing. A lane or phase-difference cycle is liable to be lost, with a corresponding whole lane error in the final reading, if the phase comparison process is continuously interrupted for a period longer than that required for the aircraft to pass through a distance equal to half a lane. Such an interruption could take place under the conditions of an unfavourable signal-to-noise ratio and would manifest itself as a loss of field strength in the Decometer coils, i.e. a loss of torque in the meter movement and in the associated Flight Log servo. With the Dectra ranging pattern, lane-loss could occur through cancellation of the principal components in one or both of the ranging signals. It could also occur through unknown phase changes in the two signals of up to about one cycle as the primary mode changes from groundwave to first-hop skywave and vice versa. Whether the lane errors thus introduced would tend to add or cancel at the end of the crossing is conjectural (neglecting here the Flight Log's ability to resolve this uncertainty, at least in part, by the dead-reckoning history it builds up); but the lanes are only some 2,320 yards wide measured along the baseline of the pattern, so that an aircraft on that heading would cross a lane in about 9¼ seconds at what will shortly be the typical cruising speed of

* Unpublished communication.

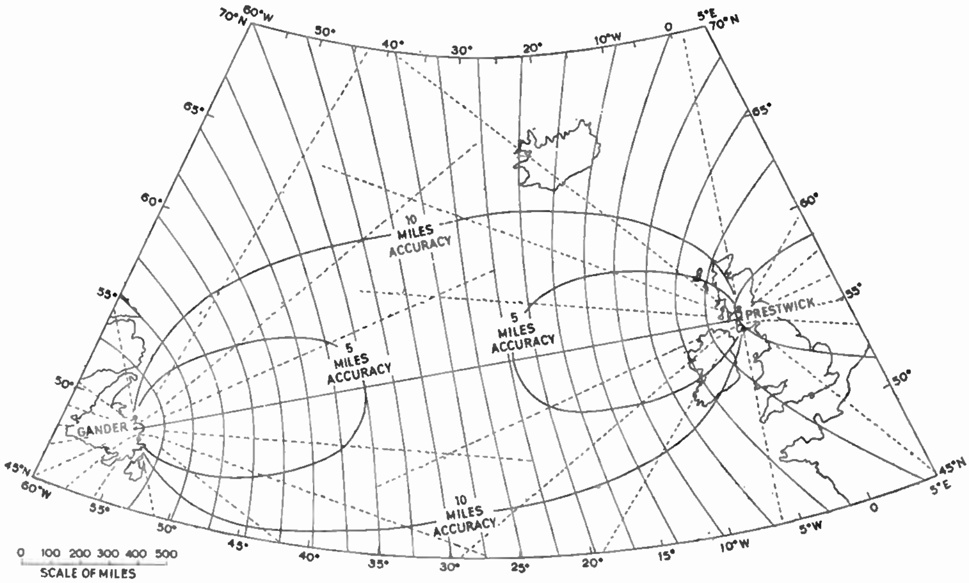

Fig. 8. Predicted accuracy contour diagram for double-ended Dectra chain.

450 kt and a residual error of several complete lanes could therefore be accepted in practice. Provisionally it was assumed that the net error in the ranging measurement would be unlikely to be such as to increase substantially the contour values of Fig. 8 when those values were interpreted as the resultant fix errors rather than errors in the tracking co-ordinate alone. For example, if a ranging error as great as 9 lanes (about 10 n.m.) were present in an aircraft located at the centre of the coverage of Fig. 8, the resulting fix error then would only be $\sqrt{2}$ times the contour value or about 14 miles.

Originally it was proposed to supplement the ranging system discussed above by a "single-signal" method in which the phase of an oscillator in the aircraft, having been adjusted to that of the nearest Dectra station at the time of take-off, would be compared with that station during the ensuing flight and thus provide a direct indication of the distance from the station. The arrangement now used, in which stable oscillators reproduce the phase of the received signals, goes some way to fulfilling the function of the original proposal by preserving lane integration if reception is interrupted for periods of a few seconds; the possibility that occasional gaps may occur lasting minutes rather than seconds, especially when the aircraft passes through the region of skywave/groundwave cancellation, is catered for by the tolerance in lane-loss already mentioned.

The estimation of the power required to ensure a given level of reliability for a given set of receiver and display characteristics is again a complex problem. There is, however, already a considerable quantity of observed data available from long-range trials with the normal Decca chains including those by Caradoc Williams* and Verstelle; on the basis of these and on the available evidence on noise level at the latitudes concerned, it was estimated that a radiated power of 20 kW would be required for the Dectra stations when on a fully operational footing, but that powers of about 2 and 1 kW radiated from the master and slave of the tracking pair of stations would be acceptable for trials purposes. The power radiated by the ranging slave station under the present trials conditions is estimated at 120 watts; the results so far obtained with this transmission are included in the Appendix.

* Caradoc Williams, "Low-frequency radio wave propagation by the ionosphere, with particular reference to long-distance navigation". *Proc. Instn Elect. Engrs.*, **98**, Part III, pp. 81-99, March 1951.

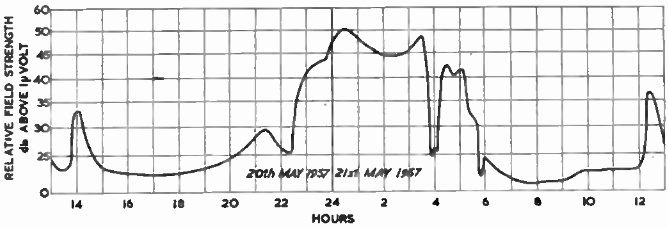

Fig. 9. Record of relative signal strength at New Malden over a 24-hour period.

The reduction of the receiver pass-band to about one-third of that originally assumed, which can be interpreted as a three-fold reduction in the radiated power required to sustain lane integration in a given ambient noise level, is a comparatively recent development and accordingly a lower radiated power than the original estimate of 20 kW may prove acceptable at least for the tracking patterns. The signal strength in U.K. of the Newfoundland transmissions has been sufficient to allow records to be taken at New Malden, Surrey, a distance of 1950 n.m. from the master; Fig. 9 shows typical variations in the relative strength of the master signal received at New Malden over a 24-hour period, with characteristic changes in level resulting from shift, with layer height, in the reflection points of the single and multi-hop components. By day the signal was at or below noise level with the equipment used except for the peaks at 1400 and 1300 hours which are thought to be due to solar flare activity.

While the evaluation is yet at an early stage, the results so far obtained suggest that the single-ended Dectra system at present in operation might replace the dual tracking patterns originally proposed, with a consequent simplification of the geometry and operating procedure; if this were done, an additional Decca chain of normal power might be added in Ireland to reinforce the eastern end of the Dectra coverage. The asymmetrical coverage over the Atlantic thus produced, tapering in towards the western end, would be in conformity with the distribution of the principal route terminals which are more widely spaced in latitude on the European side.

6. Acknowledgments

The author gratefully acknowledges the help of members of the technical staff of the Decca Navigator Company, in particular Messrs. D. A. Hendley and N. G. Fenner of the Research Laboratory and Messrs. H. G. Hawker and H. E. Oliver of the Planning Department. Thanks are also due to Mr. H. F. Schwarz, Managing Director of the Decca Navigator Co. Ltd. for permission to publish this paper.

7. Appendix: A Note on Initial Trials of the Dectra System

The system is at present (March 1958) being tested by the Aeroplane and Armament Establishment, Boscombe Down, on behalf of the Ministry of Transport and Civil Aviation and also by civil air lines. Phase 1 of the Boscombe Down trials has now been completed; its object was to test, in the air, the tracking pattern radiated from Newfoundland, and the trial was carried out in the early summer of last year before the ranging station in the United Kingdom came into operation. Eight sorties were flown, totalling 39½ flying hours. A Valiant jet bomber was used in order to introduce conditions typical of those which will be met by commercial aircraft flying the North Atlantic a year or two hence (i.e. heights from 35,000 to 50,000 ft. and ground speeds between 350 and 550 knots).

To determine the accuracy of the Dectra tracking pattern on the test flights, the true position of the aircraft at short-time intervals throughout the route was found by recording in flight, both manually and photographically, navigational data from several different sources. These included astro, Doppler, Loran and Consol. Every 15 seconds throughout each flight a photograph was taken of an automatic-observer panel, from which positions could be calculated and a datum track constructed retrospectively.

The aircraft installation included a Decca as well as a Dectra receiver. Two aerials were used for Dectra, one an elliptical disc of 280 square inches in the belly of the aircraft, and the other a metallic foil sheet of 80 square feet forming an inner skin to the radar nose. Either aerial could be switched to the Dectra receiver in flight. The Flight Log was used as well as a Decometer and the aircraft was navigated on several of the flights entirely with its aid. In the absence of a ranging pattern, the ranging co-ordinate movement of the Flight Log was simulated by a motor drive.

The fourth test flight in the series serves as a typical example of the Phase 1 trials, and was from Gander to Boscombe Down, by day and at a cruising height of about 40,000 ft. At the start of the flight, Gander Defence Radar was used to position the aircraft at the reporting point known as "Shark," 125 nautical miles from Gander; from then on the Flight Log was used for track keeping, driven from Dectra until longitude 10° W. was reached, whereafter the South-West British Decca chain was used. The pilot continued to use the Flight Log with no other aid until within two miles from the end of the runway at Boscombe. A divergence from the great circle track, between 45° and 44° W., was made deliberately as a check on lane integration. The small elliptical aerial was switched in from take-off until a position was reached approximately 1,500 nautical miles from the transmitters, after which the larger aerial was used.

While random discrepancies between the Dectra and datum tracks were extremely small along the great circle, the mean divergence placed the Dectra track slightly to the north. The run-in from the great circle to Boscombe Down showed the increase in fluctuations that had been recorded on other flights, and the same fixed displacement of the pattern. The cyclic fluctuation was found to be due to too long a time-constant in the discriminator and has since been eliminated; the overall northward shift in the pattern was due to a change in the phasing at the slave station, and occurred at a time when no external monitor receiver was being used.

In general, the results of this flight confirmed those of the previous trans-Atlantic crossing and again showed that the Flight Log could be used with Decca and Dectra for the maintenance of track between Newfoundland and U.K.

While it must be stressed that only tentative conclusions on the performance of the system can be reached at this date, the Phase 1 trials show promise of high accuracy, good reliability, and ease of operation. Allowing for the cyclical and pattern-phasing errors just mentioned, the random errors in tracking had a standard deviation of two nautical miles up to 1,500 n.m. range; up to this distance no lane-loss occurred using the smaller aerial, except when flying directly across the pattern near the transmitters where the lanes are narrowest. The effective range may be less in areas away from the baseline disector, along which the test crossings were made, and this is one of the many aspects yet to be investigated. The tests showed that the equipment was suitable for pilot operation and the Flight Log proved a convenient method of data display.

The next stage was to commence evaluation of the ranging facility and also to carry out flights on which the transmission paths would lie entirely over land instead of over sea. Both these investigations are now in progress, and flights have been made over the North American continent to test the back-cover of the tracking pattern. Of the ranging trials, little can be said yet with certainty, since at the time of preparing this note only four of the 13 flights involved have been analyzed and these not fully; for the four runs, however, within the zone between 10° and 50° W. there appears to be a residual uncertainty of about plus or minus two lanes, or $\pm 2\frac{1}{2}$ nautical miles, in the ranging co-ordinate readings compared with the datum. The signal from the ranging slave station, at present radiating approximately 120 watts, has been received at usable strength at the far western end of the route.

The system is under test in various aircraft at a wide range of flying heights below that of the Valiant, and also in ships. In order to observe the propagational characteristics of the system on a long-term, 24-hour-per-day basis, it is proposed to set up monitor receiving stations at widely spaced locations.

DISCUSSION

Sqdn. Ldr. D. F. H. Grocott, A.F.C., R.A.F.* : I regret that I can contribute nothing to the technical side of the discussion, but I can give some information on the evaluation trials being conducted in a Ministry of Supply Valiant aircraft. I have flown on almost all flights since the ranging stations started operating last autumn and with the Chairman's permission I should like to outline the procedure used on these flights.

From the moment that the aircraft left Boscombe Down until it arrived at its destination in Canada the pilot received no navigational assistance from the two navigators aboard the aircraft. The pilot flew the aircraft with reference to the Flight Log, using a Decca input between Boscombe Down and the west coast of Ireland, Dectra over the main North Atlantic area, and Decca (new Canadian Decca chains) from 50°W to Gander or Ottawa. At all times, the pilot computed his own ground speed and E.T.A.'s from the data provided in the flight log, and passed regular position reports, etc., to the air traffic control authorities. You may be wondering why we carried two navigators aboard the aircraft? The answer is that two navigators are required to monitor the auto-observers and collect other navigational data, which cannot be recorded on film, for the post flight analysis.

Mr. Powell did mention the term "datum track" and said that he had insufficient time to go into the method used to determine this magic line. Perhaps it was just as well that he did not discuss this aspect because it takes two officers more than two weeks to read the auto-observer film, plot all the navigational data, and then determine the datum track from a comparison of the empirical track produced from the Decca, Loran and Consol information and the mathematical track computed from a precise knowledge of heading, Doppler drift and distance gone.

I should like to ask one technical question. Mr. Powell stated that both the ranging and tracking stations are radiating only a fraction of their design power. I wonder if he would please explain what are the advantages and disadvantages of increasing the power immediately up to the design maximum for the respective aerials.

D. A. Hendley (*jointly presenting the paper with the author, in reply*): While the possible need to raise the power of the stations in the future was foreseen from the outset, this will be a considerable undertaking particularly at the ranging slave station. Any increase in power must be consistent with reasonable economy of construction and operation, and it is felt that the evaluation should proceed farther before attempting to specify the radiated power levels for the chain in its permanent form.

I would point out, however, that so far in the evaluation there has been no failure of Dectra attributable to lack of transmitter power. In fact, there are examples of users reporting locking on to the Earl's Hill station (Scotland), which radiates only 120 watts, whilst over Nantucket.

N. G. Anslow (Associate Member)† : B.O.A.C. are just commencing their evaluation of Dectra. Consequently, unlike Sqdn. Ldr. Grocott, I cannot speak from first-hand knowledge of the performance of the system over the North Atlantic. I would therefore like to confine my discussion to asking three questions.

Firstly, is it essential that the Dectra equipment is always associated with Decca? For aircraft operating on long haul routes, i.e. London/San Francisco, where there is at present no Decca coverage on the second part of the route, the carriage of this equipment becomes, from a weight penalty consideration, uneconomic. Could not, for example, the Dectra equipment be "set up" by reference to a position defined by some other aid to navigation such as short range aid already carried by the aircraft for other purposes?

Secondly, in our Trans-Atlantic LF/RTT tests on 120 kc/s we have noticed occasionally "outages" due to anti-phase but coincident amplitude signals arriving by the direct and indirect path. These outages usually last only two or three minutes. Even assuming that the

* Aeroplane and Armament Experimental Establishment.

† British Overseas Airways Corporation.

Dectra will function on a lower signal/noise ratio than LF/RTT, does the author think that these outages will seriously upset the overall system accuracy on a Trans-Atlantic trip where some 6 to 10 outages may occur?

Finally, in view of the general possibilities of loss of Dectra signals due to outages, static or interference of some kind, and the limitations imposed on the overall accuracy of existing Doppler navigation systems because of their dependence on aircraft heading information, would not the ideal be to combine the two and reference the Doppler navigation aid to the Dectra radio pattern, thus avoiding the problems of changes in variation, etc., associated with the heading arrangements for Doppler?

D. A. Hendley (*in reply*): In reply to Mr. Anslow's first question, it is possible to set up the Dectra receiver from other sources of position information, although the likelihood is remote of obtaining from any other single aid a fix accuracy comparable with that of Decca—or sufficient to be certain of setting in the correct Dectra lane. Only a small weight penalty is incurred in having Decca as well as Dectra in the aircraft, as much of the equipment, including the whole of the display, is common to both systems. If Dectra alone were installed, the inability to use the Decca coverage, even if this existed only at one end of the Dectra route, would be a heavy price to pay for a very small saving in weight.

In regard to the "outages" on the 120 kc/s LF/RTT tests, experience to date suggests that this occurs far less frequently with Dectra. The narrower bandwidth and lower acceptable signal/noise ratio undoubtedly help to account for this. In any case, the Flight Log can in effect over-ride short interruptions, and as a result there has been no loss of reliability from this cause.

On the combination of Dectra and Doppler, the two systems have complementary features, and have already been systematically combined in the Decca Integrated Air Navigation (DIAN) system, which uses the Flight Log as a common display for Decca, Dectra and Doppler.

P. F. Cook*: Referring to the reliability of airborne equipment, the more efficient and useful a navigational aid, the greater are the implications of its loss through a defect on any given flight. It would seem that this situation could best be ameliorated by either designing the equipment to possess a great intrinsic reliability or by providing equipments of a more normal reliability but duplicated to increase the probability of continuous functional service, or by providing an alternative standby system. It would be interesting to know which of these (or other) solutions is considered most appropriate to Dectra.

D. A. Hendley (*in reply*): Weight and bulk considerations tend to favour the provision of the highest possible intrinsic reliability. The essential simplicity of the Dectra receiver, and the fact that no valves or other components in it are called upon to operate under exacting conditions, are important factors in this context. Outside the receiver itself there is a useful measure of duplication in that the Flight Log and its associated computer, which together form a large part of the total installation, are supplemented by the Decometers which give the full fix information independently.

C. W. Fowler (Associate Member)†: I would like to ask if the relative stability of the oscillators at the tracking and ranging transmitters is high enough for synchronization periods to be confined to the hours of daylight along the path. If not, and synchronization is a continuous process, are figures available for the diurnal phase changes experienced? Also, what are the extreme positional errors due to this effect?

D. A. Hendley (*in reply*): The locking of the slave station in Scotland with the Master in Newfoundland is on a continuous basis, but the oscillator forming the slave-signal source has a stability better than one part in 10^7 for short periods and can thus maintain acceptable pattern-stability during interruptions in master reception lasting several minutes. Regular monitoring of the transmissions in the U.K. is now in operation, and evidence should shortly become available on which to estimate typical diurnal phase changes at various distances from the stations.

* Air Registration Board.
† Radio Research Station.

THE PROBABILITY OF SPECIFIED LOSSES AT MISMATCHED JUNCTIONS*

by

John H. Craven, M.A.Sc. (Associate)†

SUMMARY

Probability contours are presented for specified losses (<1 db to 6 db) at junctions between networks or lines for a range of v.s.w.r. from 1 to 10.

1. Introduction

In many applications of waveguides and other transmission lines it is useful to be able to estimate the effect on power transmission of joining together two such transmission lines of the same configuration but both more or less mismatched. If the input impedance of both transmission lines is known in amplitude and phase the power loss through the junction can be computed exactly. However, it is more usual to have the mismatch of a transmission line expressed as a v.s.w.r. without any reference to phase. When one such item as the source is connected to the second as a load it is useful to have an indication of the probable power loss due to the mismatch of the junction. Calculations have been made as to the probability of certain specified losses occurring when the v.s.w.r.'s of the transmission lines to be joined assume values from 1 to 10.

2. Derivation

From the consideration of the source–load junction in Fig. 1, it can be shown that the mean power delivered to the load is

$$\frac{E^2 R_b}{|(Z_a+Z_b)^2|}$$

where the source impedance is $Z_a = R_a + jX_a$ and the load impedance is $Z_b = R_b + jX_b$. Maximum power transfer to the load is obtained when $R_b = R_a$ and $X_b = -X_a$ and the maximum power transferred, P_{max}, is

$$\frac{E^2}{4R_a}$$

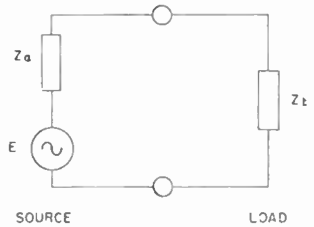

Fig. 1. Source-load junction.

The relationship of the maximum power available to the actual power transferred is

$$\frac{P_{max}}{P} = \frac{|(Z_a+Z_b)^2|}{4R_aR_b} = \frac{(R_a+R_b)^2 + (X_a+X_b)^2}{4R_aR_b}$$

The same result is obtained for a constant current source as for a constant voltage source.

If the source is a transmission line with a v.s.w.r., of S_1, and the load is an identical transmission line with a v.s.w r. of S_2

$$Z_a = Z_0\frac{S_1 + \mathrm{j}\tan\theta_1}{1+\mathrm{j}S_1\tan\theta_1} \text{ and } Z_b = Z_0\frac{S_2+\mathrm{j}\tan\theta_2}{1+\mathrm{j}S_2\tan\theta_2}$$

where θ_1 and θ_2 are the distances in radians from the junction to the voltage maxima in the respective transmission lines and Z_0 is the characteristic impedance of the transmission lines. Therefore:

$$R_a = Z_0 S_1 \frac{(1+\tan^2\theta_1)}{1+S_1^2\tan^2\theta_1}$$

$$R_b = Z_0 S_2 \frac{(1+\tan^2\theta_2)}{1+S_2^2\tan^2\theta_2}$$

$$X_a = Z_0\tan\theta_1\frac{1-S_1^2}{1+S_1^2\tan^2\theta_1}$$

$$X_b = Z_0\tan\theta_2\frac{1-S_2^2}{1+S_2^2\tan^2\theta_2}$$

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* Manuscript first received 18th June 1957 and in final form on 12th December 1957. (Paper No. 453.)

† Radio and Electrical Engineering Division, National Research Council, Ottawa.

U.D.C. No. 621.372.2

$$\text{Thus} \quad \frac{P_{max}}{P} = \frac{\left[\dfrac{S_1(1+\tan^2\theta_1)}{1+S_1^2\tan^2\theta_1} + \dfrac{S_2(1+\tan^2\theta_2)}{1+S_2^2\tan^2\theta_2}\right]^2 + \left[\dfrac{\tan\theta_1(1-S_1^2)}{1+S_1^2\tan^2\theta_1} + \dfrac{\tan\theta_2(1-S_2^2)}{1+S_2^2\tan^2\theta_2}\right]^2}{4\left[\dfrac{S_1(1+\tan^2\theta_1)}{1+S_1^2\tan^2\theta_1}\right]\left[\dfrac{S_2(1+\tan^2\theta_2)}{1+S_2^2\tan^2\theta_2}\right]}$$

$$= \frac{(S_1+S_2)^2 + (S_1^2-1)(S_2^2-1)\sin^2\theta}{4S_1S_2}$$

where $\theta = \theta_1 + \theta_2$

This is a periodic function of θ with a period of π as shown in Fig. 2, which is a plot of P_{max}/P for $S_1=2$, $S_2=6$ (or $S_1=6$, $S_2=2$, since the function is symmetrical in S_1 and S_2). For the special case where $S_1=1$ or $S_2=1$, P_{max}/P becomes a constant with a value dependent on the v.s.w.r. which is not unity.

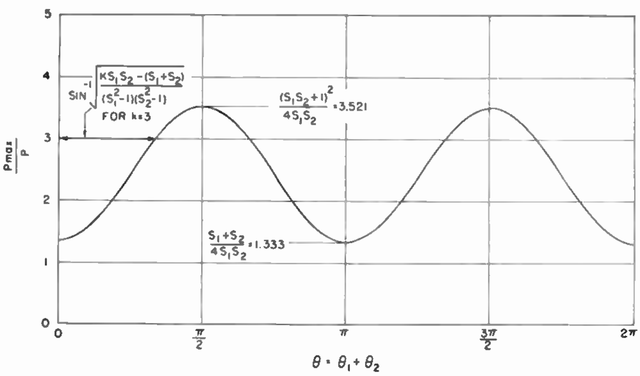

Fig. 2. Plot of P_{max}/P for $S_1=2$ and $S_2=6$ (or $S_1=6$ and $S_2=2$).

Maximum power transfer for given values of S_1 and S_2 occurs when $\theta = n\pi$ ($n=0$, 1, 2) and P_{max}/P is $(S_1+S_2)^2/4S_1S_2$. This is equal to unity for S_1 equal to S_2. Minimum power transfer occurs when $\theta=(2n-1)\pi/2$ ($n=1$, 2) and P_{max}/P is $(S_1S_2+1)^2/4S_1S_2$. For S_1 and S_2 both much greater than 1, P_{max}/P is approximately $S_1S_2/4$ at minimum power transfer. Thus for large equal v.s.w.r.'s the power ratio ranges from unity to $S^2/4$.

It is assumed that the values of both θ_1 and θ_2 have a purely random distribution over the range 0 to π and, since P_{max}/P is a periodic function of θ with period π, θ may be considered as being randomly distributed in the interval 0 to π. Also, since all possible values of P_{max}/P are to be found in the range 0 to $\pi/2$ and P_{max}/P is a symmetrical periodic function of θ with period π, the probability of the value of P_{max}/P being less than a specified quantity "K" is

$$p(K) = \frac{2}{\pi}\sin^{-1}\left[\frac{4KS_1S_2-(S_1+S_2)^2}{(S_1^2-1)(S_2^2-1)}\right]$$

for $(S_1+S_2)^2 \leqslant 4KS_1S_2 \leqslant (S_1S_2+1)^2$ with the restrictions that if $4KS_1S_2<(S_1+S_2)^2$, $p(K)$ is 0, and if $4KS_1S_2>(S_1S_2+1)^2$, $p(K)$ is 1.

A number of different graphs may be drawn to show the relationship between S_1, S_2, K and $p(K)$, but the most useful is a plot of probability contours with respect to S_1 and S_2 for a given value of K. Four such graphs are shown in Figs. 3 to 6 which are plotted for K corresponding to power losses of 1, 2, 3 and 6 db respectively. ($K=1{\cdot}259$, $1{\cdot}585$, 2 and 4.)

In order for the loss due to a mismatched junction of the type under consideration to be certain to give a P_{max}/P of less than K ($p(K)=1$) the product of the v.s.w.r.'s must satisfy

$$S_1S_2 < 2K-1+2\sqrt{[K(K-1)]}$$

In order for it to be certain for the loss at a particular junction to be less than 3 db ($K=2$) then S_1S_2 must be less than $5{\cdot}828$.

It can also be shown that for S_1/S_2 or $S_2/S_1 > 2K-1+2\sqrt{[K(K-1)]}$ it is impossible to have P_{max}/P less than K. That is, for S_1/S_2 or S_2/S_1 greater than $5{\cdot}828$, K must be

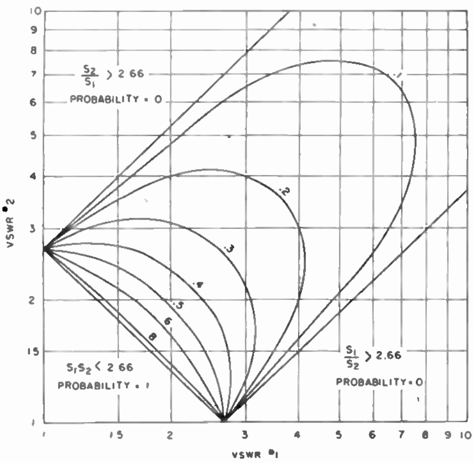

Fig. 3. Probability contours for loss < 1 db (probability of $K<1{\cdot}26$).

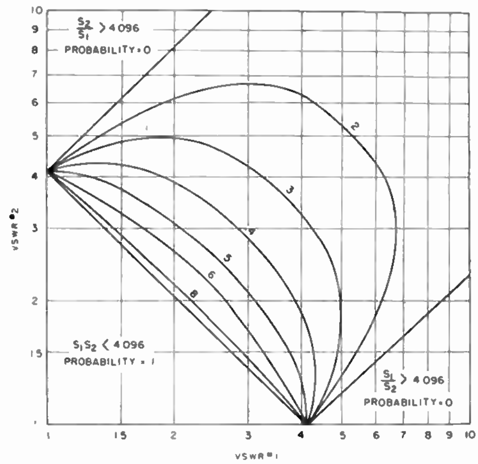

Fig. 4. Probability contours for loss <2 db (probability of $K<1{\cdot}58$).

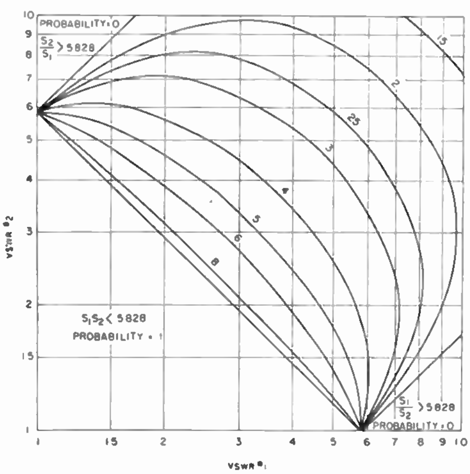

Fig. 5. Probability contours for loss <3 db (probability of $K<2$).

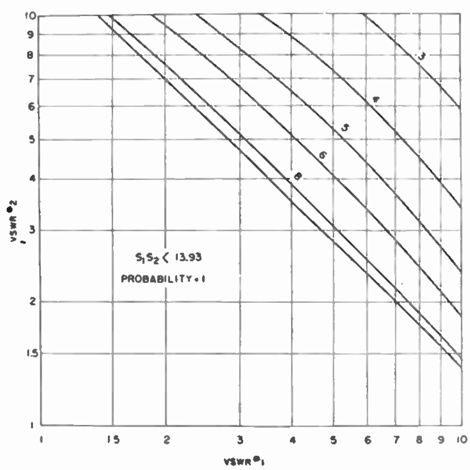

Fig. 6. Probability contours for loss < 6 db (probability of $K<4$).

greater than 2 and the loss must be greater than 3 db.

3. Application

Although these calculations of the probability of specified losses occurring apply quite generally to all source–load combinations, they are particularly useful in the evaluation of broadband receiving systems involving an aerial and a mixer or crystal detector, both of which may have v.s.w.r.'s varying over quite a wide range. The calculation of the exact loss due to the junction of the two components would involve a considerable amount of measurement and computation and would only apply to the specific lengths of line for which calculations were made. On the other hand, by the use of graphs and a few simple calculations, it is possible to obtain an assessment of the probable losses involved regardless of the length of transmission line which might be used between the two points at which the v.s.w.r.'s were measured.

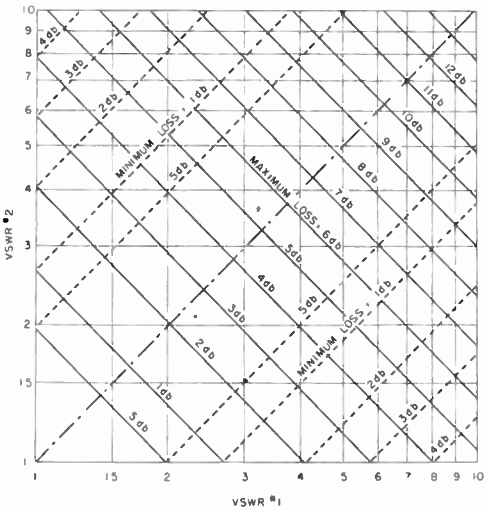

Fig. 7. Probability limits for different values of *K*.

This method may also be applied to the transmitting system such as a power source feeding an aerial provided that the power available from the source is independent of the conditions of loading.

Another useful graph is obtained by plotting the position of the unity and zero probability limits for different values of *K* corresponding to 1 db steps (Fig. 7). For a given pair of v.s.w.r.'s readings of the unity and zero probabilities are obtained. For example, with v.s.w.r.'s of 5 and 3 the unity probability is 6·3 db and zero probability is 0·4 db. This indicates that the loss introduced by the junction will be in the range 0·4 to 6·3 db. It also indicates that if the terminals being joined were separated by a line stretcher or phase shifter an adjustment of the equivalent length of the connecting line could at best give a transmission loss of 0·4 db.

4. Acknowledgment

The author wishes to thank the National Research Council for permission to publish this paper and to acknowledge the encouragement of Mr. W. L. Haney, of the Radio and Electrical Engineering Division.

SHORT CONTRIBUTIONS AND WRITTEN DISCUSSIONS

As was pointed out in the April issue of the *Journal,* the Institution welcomes the submission of short papers which may be either complete in themselves, or comment on previously published work. Contributions should be sent to the Institution for the consideration of the Programme and Papers Committee.

SELECTIVE CALLING SYSTEMS FOR MOBILE RADIO TELEPHONY*

by

N. Sohrabji (Graduate)†

SUMMARY

The principles and applications of selective calling are discussed and the impulse coding and frequency coding methods used are described. A detailed account is given of a modern system using single tones which is adaptable to a wide variety of equipment and applications.

1. Introduction

The rapid expansion of mobile radio-telephony has led to the formation of networks in which a large number of operators are controlled from a single dispatching station. Unless special means are adopted, every operator has to listen to every call in order to receive the few calls which are directed to him. Thus in a fleet of 100 vehicles, a driver would, on the average, have to listen to 100 calls in order to receive one call directed to him. In the frequency bands commonly used at present, there is also the danger of interference from other services within the band. If the operator is to discharge his duties efficiently, some form of selective calling becomes essential.

2. Principles of Selective Calling

Briefly, each receiver in the service is kept mute until a coded signal, unique to each station, is received. On the completion of the call, the receiver is again muted (re-set) in readiness for the next call.

2.1. *Facilities required from Selective Calling Systems*

The minimum facility required is that the wanted receiver is unmuted while the others remain silent. This is most conveniently arranged by switching the screen voltage of the output tube, and has the advantage that the receiver drain is kept low during stand-by. Visual or audible indication, or both, of a call is usually necessary. It should be possible to reset the system at the end of a call. Other facilities can be arranged depending on the method of operation, and the degree to which it is desirable for the calling system to take over control of the service. For example, it can be arranged to lock out the transmitters of the unwanted stations and provide them with a busy signal for the duration of the call. Also a station can be made to respond automatically when called and to send a busy signal when engaged. Facilities similar to those obtained in a line-telephone network can be provided. In some applications, it is desirable to be able to call groups of stations or all stations at once. This is essential for certain services, such as police radio.

2.2. *Performance required from a Calling System*

In order that the installation of a selective calling device may not degrade the service in any respect, the following factors must be considered:—

(1) The system must operate over the whole range of signal strength which allows satisfactory communication, i.e. it must not restrict the operating range.

(2) The performance of the calling device as regards variation of supply voltage, temperature, humidity and vibration must be equal to that of the associated apparatus.

(3) The system must work equally well over land-line or radio links.

(4) The signalling frequencies must lie within the normal audio-frequencies band of the service.

(5) The calling device must not be triggered by speech, interference or noise.

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† Formerly with Redifon Ltd., London, S.W.18, now with Indian Telephone Industries, Bangalore.

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(6) The calling device must not be affected by the operation of the self-quenching and volume controls of the receiver.
(7) The apparatus fitted to the mobile, or "slave" station, must be compact, have low current drain and should be able to work off the supplies already available for the transmitter and receiver.
(8) The calling device should be readily adaptable to the various methods of operation—e.g. simplex, two frequency simplex, and duplex—with a minimum of change.

3. Methods Used in Selective Calling

Broadly, the methods fall into two groups—Impulse Coding, and Frequency Coding.

In the impulse method, a single tone is used and a series of pulses are sent, the total number of pulses being the same. The decoding unit in the remote station employs a stepping switch and when the correct code is received the receiver is unmuted.

In the frequency coding, a number of tones (generally one to four) is transmitted. At the receiving end, high-Q band-pass filters select these tones and if the right combination is received, the receiver is unmuted.

Frequency coding involves very simple decoding apparatus and is hence to be preferred. Impulse coding involves the use of an elaborate decoder, employing a number of relays. The unit is bulky, takes a large amount of power and is more liable to break down. On the other hand, impulse coding is less liable to false operation and is readily adapted to special facilities such as group calling, general calling, or resetting from the control station.

4. Systems Using Impulse Coding

In this method, groups of impulses are transmitted, the total number of impulses being always the same.* The number of channels available is limited by the total number of impulses, the latter being generally from 10 to 25. A four digit code is usual. Dialling is accomplished by means of a standard telephone dial. A block diagram of a basic coding unit is shown in Fig. 1. The telephone dial keys the oscillator, which may have a frequency of say 3 kc/s, and the impulses are fed to the transmitter.

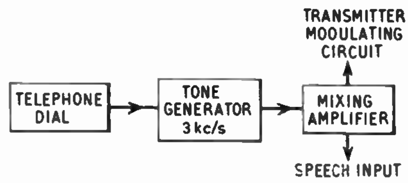

Fig. 1. Block diagram, control station coding unit.

At the receiving end, a decoding unit (Fig. 2) extracts the tone impulses by means of a high-Q filter, which drives a relay. This relay drives a stepping mechanism. If the code received is correct, the stepping relay is driven to its final position, and the receiver is unmuted. If the code is not correct, the de-coding relay releases the stepping relay and the latter returns to its starting position, the transmitter is locked out, and a busy signal appears.

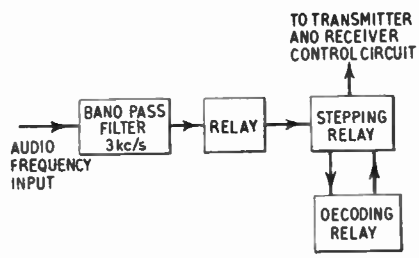

Fig. 2. Block diagram, impulse decoder unit.

A reset mechanism is provided at the control station which sends out a series of pulses and releases all the decoders.

When faster operation is desired, a push button system is used; the operation of a single push button out of the appropriate code[1].

An arrangement has been described[2], in which a dekatron is used for counting the impulses. The correct combination is recognized by a circuit using cold cathode tubes.

5. Frequency Coding Systems

In this system, each slave station is assigned a unique tone or combination of tones. Reception of the correct tone or tones unmutes the

* For example, only numbers such as 4783 and 3964 would be used, in both of which the number of impulses is the same. This simplifies the coding system.

receiver. Other facilities such as lock out, engaged and reply signals, etc., can be provided as required.

With single tone calling, the number of channels is equal to the number of tones. These are selected at equal spacing within the audio-frequency band of the service, which is generally from 300 c/s to 3000 c/s. Single tone calling is not suitable when a very large number of channels is required, because the stability requirements of the tone generating and tone selecting circuits become very difficult. Any simplicity obtained from using a single tone is more apparent than real. For example, with a 100-channel system the number of tone generating networks needed would be 100 for a single tone system and only 20 for a two-tone system. Since the tone generating and filtering circuits are the most critical components of the system, it is advantageous to use multi-tone operation when the number of channels is large.

In multiple tone operation, the signals may be sent either simultaneously, or sequentially. For simultaneous operation, the number of channels available is given by:

$$\text{No. of channels} = {}_nC_r \qquad \text{...........(1)}$$

where n = total number of tones available and r = number of tones per code.

Thus with 10 tones, any two of which may be selected, 45 channels are available.

With sequential operation, the number of channels would be 100 ($=n^r$). For this reason, and also because intermodulation distortion in the system can cause false operation with simultaneous calling, the sequential system is to be preferred. With this method, the reception of the correct tone by the first filter opens up the input to the second filter, and if the second correct tone is now received, the decoder is operated.

In order to avoid triggering by speech, interference and noise, the decoding circuit is so arranged that it responds to a signal of the correct frequency only if the signal is maintained for a certain minimum duration. With high-Q filters and tone durations greater than 200 milliseconds, false triggering can be reduced to negligible proportions.

6. Basic Circuits Used in Frequency Coding

The basic components of a frequency coding system are the tone generators and tone selectors or filters. It is shown above that highly selective filters are necessary, hence the stability of both oscillators and filters is very important. One arrangement for tone generation uses a crystal oscillator associated with a frequency divider[3].

Tone selection (filtering) has hitherto been achieved mainly by means of vibrating reed selectors[4]. A diagram is shown in Fig. 3. Two strips of magnetic material are brazed to a separating block and form a tuning fork. A bar on one of the strips is opposite to a fixed contact wire. The frequency of the tuning fork is obtained approximately by selecting a separating block of the right length. It is then adjusted accurately by bending the tuning bars either towards, or away from, the block. When a current of this frequency flows in the coil, the reeds vibrate and the contact is made for a portion of each cycle. The duration of the contact can be varied by adjusting the gap. The effective mechanical resistance is very low and a high-Q is obtained. The Q is controlled by means of copper sleeves attached to the pole piece. These sleeves act as a short circuited windings and provide the damping which lowers the Q. It can be seen that in order to obtain a stable resonant frequency and Q, very great care has to be taken in the design and manufacture of such units.

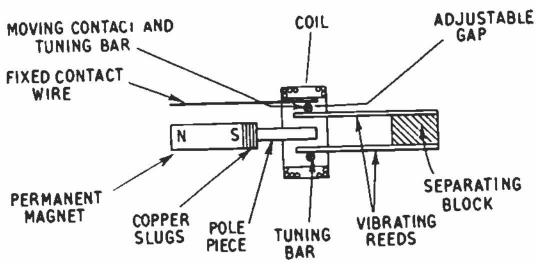

Fig. 3. Vibrating reed selector unit.

In the system to be described presently, tone generation and selection is accomplished by means of resistance-capacitance oscillators and filters. Frequency-selective R-C networks when used in the feedback loop of an amplifier have many advantages at low frequencies. This technique has been used to produce stable, compact units giving a very satisfactory performance.

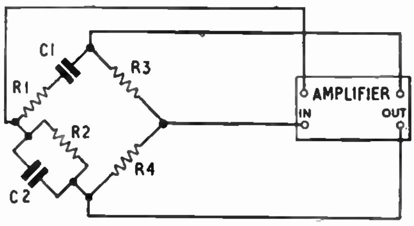

Fig. 4. Basic circuit, Wien-bridge oscillator.

The oscillator circuit: The basic circuit is shown in Fig. 4. When the phase-shift through the amplifier is 360 deg., the frequency of oscillation is given by

$$f_o = \frac{1}{2\pi\sqrt{(R_1C_1R_2C_2)}} \qquad \ldots\ldots\ldots\ldots(2)$$

The gain required is given by

$$A_o = 1 + \frac{R_1}{R_2} + \frac{C_2}{C_1} \qquad \ldots\ldots\ldots\ldots(3)$$

and is equal to 3 times when $R_1 = R_2$ and $C_1 = C_2$.

In order to achieve a stable output and good waveform, a thermistor is used for amplitude control and takes the place of R3. The influence of various factors on frequency and output is shown in the Table. In all cases the figures are the worst obtained in tests on four units.

Distortion is less than 5% under all conditions.

The filter circuit: A parallel-T *R-C* network[5] is used in the feedback loop of an *R-C* amplifier as shown in Fig. 5. By careful design and layout, this circuit can be made to achieve a high and stable selectivity at all frequencies in the operating range.

In order to achieve maximum flexibility, a plug-in arrangement is used for the oscillator and filter networks. The networks are assembled in cans and are sealed after adjustment.

7. A Single Tone Calling System

A single tone system has been designed which incorporates facilities for general call and also for the calling of the control station by the remote operator.

The Master Station calling unit is shown in Fig. 6. In Fig. 7 it is shown connected to the control unit of an f.m. transceiver and through it to a master adaptor unit which is mounted in the transceiver case. The 12-way push button switch S3 (Fig. 6) selects the tone frequency. The 12th button is non-latching and is used as a release button. The oscillator output is fed through a cathode follower (V2a) and a relay contact to the output line, and is finally connected to the transmitter modulating circuit. Part of the output is amplified by V2b and lights a neon lamp Lp13 for the duration of the calling tone. The call switch S2 is pressed for sending tone down the line and operates the gating relay R LA/1 through C6. The relay is a Post Office type 3000 with a heel-end slug and gives a minimum tone duration of 400 msec. Pressing the call button also unmutes the receiver. When none of the oscillator networks are in circuit, the thermistor is supplied with d.c. This keeps the thermistor warm and when a network is connected in the circuit, the

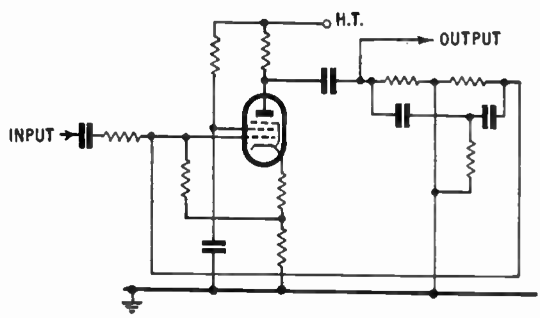

Fig. 5. Basic circuit parallel-T filter unit.

Varying Parameter	Range	Change of Frequency	Change of Output
Supply Voltage	30%	0·05%	10%
Frequency	300 c/s—3000 c/s		5%
Ambient Temperature	20°C to 55°C	0·05%	5%

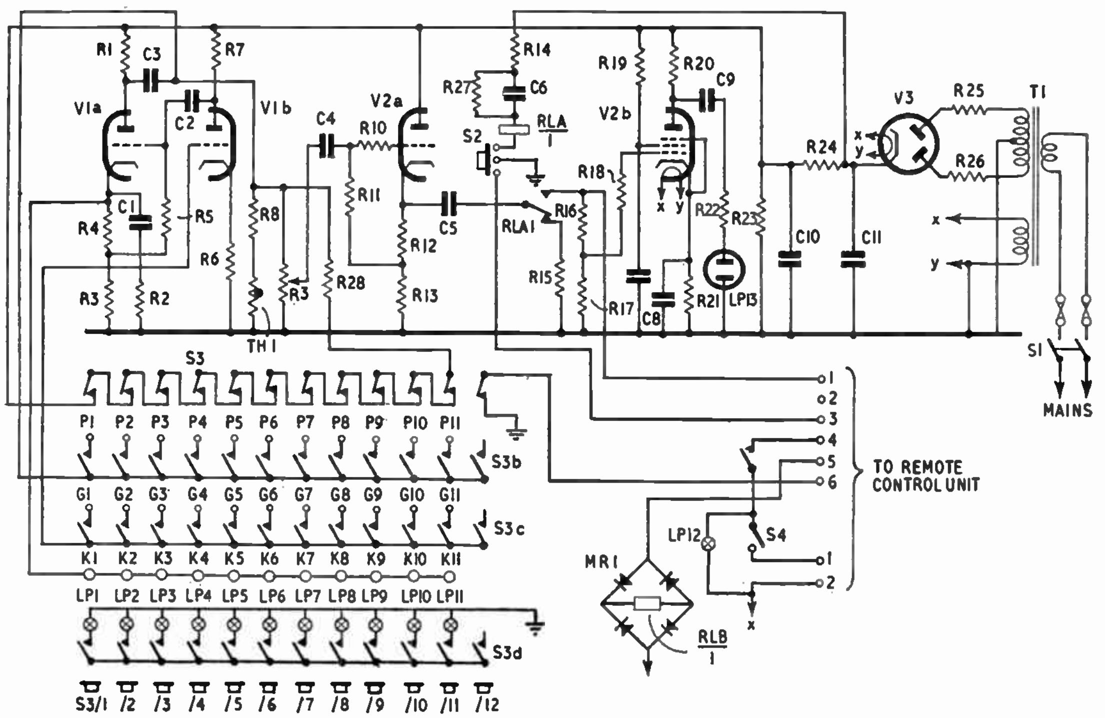

Fig. 6. Circuit diagram, master calling unit.

oscillator output reaches its stable value within 50 msec.

When the control station is called by a remote operator, a signal is sent via the adaptor unit to the master unit which operates RLB/1. Indication of an incoming call is provided by the lamp Lp12. A buzzer is provided for remote indication. The operation of the adaptor unit is very similar to that of the slave unit which is described below.

The Slave Station calling unit is associated with a remote station and takes its supply from it. (Fig. 8.) Two valve assisted filters (V1, F1 and V2, F2) are provided which take an audio input from the receiver at a point after the de-emphasis network but before the "squelch" and volume controls. These feed into two detectors, V3a and V3b, which have a common cathode load R9. The positive voltage across C5 builds up at a rate depending on the time-constant of the charging circuit. This is so arranged that a duration of at least 400 msec is necessary before the current through V4 reaches the value necessary to operate the relay RLA/2. When the relay operates, the positive bias on the cathode of V4 is removed and the relay stays operated, regardless of the voltage across C5. The h.t. voltage is applied to the screen of the receiver output valve and an earth is provided for the indicator circuit. R10 is normally short circuited through the reset switch. At the end of a call when the transmitter h.t. is switched off, the call and reset

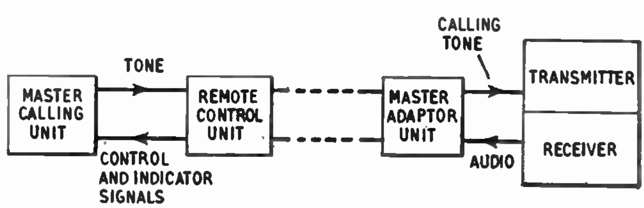

Fig. 7. Connection diagram for typical central station.

indicator operates, showing the operator that he has to reset. When the reset button is pressed, the current through V4 is interrupted momentarily and the relay is released. The positive voltage is again applied to the cathode and the receiver is now muted.

The oscillator section of the slave unit is similar to that of the master unit, except that only one tone is provided. This is because the remote station makes all its calls through the central dispatching station.

The space occupied by the slave unit is less than 1/10th of the total space allotted to the transmitter, receiver and power unit in a typical f.m. mobile installation. Its power consumption is less than 1/6th of the total power used on standby.

The equipment has been designed to function under extreme climatic conditions and attention has been paid to problems peculiar to mobile services.

8. Two Tone Calling System

In this system ten tones are again selected in the audio-frequency band. Two tones are sent out on a sequential basis and the right combination will unlock a remote receiver. The remote station unit for such a system is shown in Fig. 9. The operation is as follows: When the first filter stage V1 F1, receives a tone of the correct frequency and duration, the rectified voltage across the load resistance R9 of V2 is sufficient to unlock the second filter, which is normally biased beyond cut off. If a second tone, appropriate to this filter, is now received, the relay will operate. A central station which controls a large number of remote stations will have its receiver on at all times, hence no provision is made for a calling oscillator in the remote unit.

9. Applications of the Calling System

The flexibility of the units described allows them to be used not only in normal mobile services, but also in a variety of unusual schemes. For example, by using a master unit at each station, a service can be arranged in which any station may call any other in the group. As a second example, a service employing a very large number of remote stations

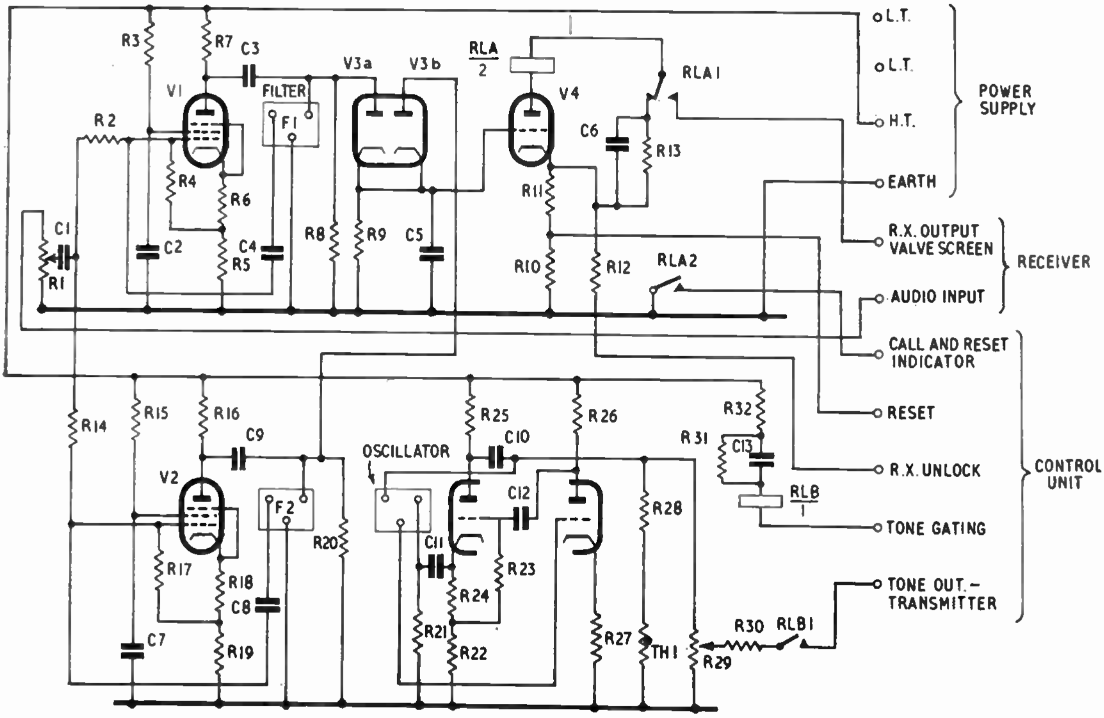

Fig. 8. Remote station unit, single-tone system.

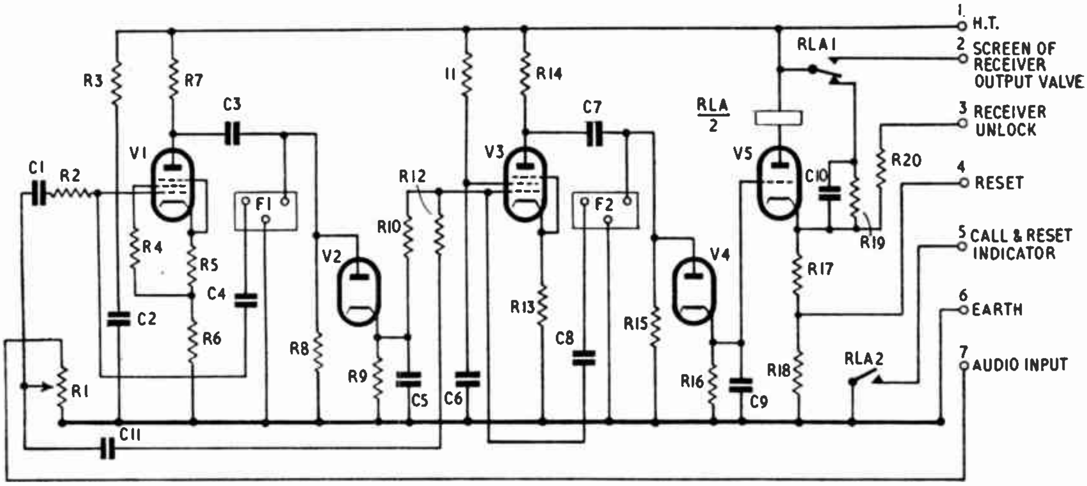

Fig. 9. Two-tone calling system, remote station unit.

such as a thousand, may be considered. Clearly one operator cannot handle all the central station traffic. By using several banks of master units, with indicator lamps wired in parallel, such a service may be successfully operated.

In such schemes problems would be involved, whose nature is administrative, rather than technical. When applying this equipment to an existing system, some changes would be involved, especially in the slave station, since it is desirable to house the slave unit within the transmitter-receiver case. Also, the power drawn by the slave unit, though small, would not be negligible.

10. Conclusions

Selective calling has become a valuable tool for the successful operation of mobile networks on land, sea and air. A simple and reliable tone-calling system has been developed which is capable of great flexibility and can be applied to a wide range of services.

11. Acknowledgments

The author is indebted to Redifon Limited for permission to publish this paper.

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DEVELOPMENTS IN COMPONENT DESIGN

The final selection of summaries of papers read at the International Symposium on Electronic Components held in Malvern in September, 1957

Component Design in Relation to Printed and Potted Electronic Circuits

H. G. Manfield*

Mr. Manfield surveyed the processes leading to failure of potted units either due to sealing, poor design or failure to note the differences in thermal expansions of materials. Some magnetic materials are affected by compression at stress when potted.

Various methods for producing printed wiring were outlined and the requirements for the base materials were stated together with recommendations of line width and spacing and current-carrying capacity of printed conductors. The impact of transistors on the technique was discussed.

Components for printed wiring and for automatic insertion were surveyed and recommendations made as to their suitability or otherwise for the Services. Most have been designed for the commercial market and are unsuitable for use under conditions of shock and vibration. Valve-holders and plug-in connectors are particularly poor and need development.

A complete unit using printed wiring on a 0·1 in. matrix with wired-in valves, heat sinks, etc., has been designed and made. Such a unit could be made by machines but these were large, heavy and expensive.

The effects of contaminants and surface coatings were stressed; none of the latter are very effective against climatic conditions.

Recommendations for standardization of the 0·1 in. matrix were made together with standardization of sizes for leads of components. New and better components and materials are needed to keep pace with the smallness and reliability of the transistor. The problems are simplified by the absence of heat and the low voltages, both of which have been a bugbear in the past.

* Royal Radar Establishment, Malvern.
U.D.C. No. 621.396.69

Some Recent Developments in Miniature Relays

Dr. J. H. Mitchell†

Dr. Mitchell said that in spite of modern switching developments there is a steady demand for relays to meet a variety of new requirements in the electronic field. He discussed five categories of relays to meet various specifications.

In discussing miniature general purpose, sensitive, high-speed and polarized relays, the author confined himself to relays that have been in the process of development for several years and which are now tooled-up and tested.

He said that the equipment designer's need to save space had influenced the design of relays more than any other factor, making it necessary to consider very carefully the basic necessities in the design of a good relay.

In detailing the problems associated with miniaturization of a relay, he discussed several salient features, the most serious being the temperature limitation imposed by the covering of the wire used in coil forming. High-temperature enamelled wire interleaved with terephthalate-based material 0·00025 in. thick and self-supporting windings were under development to meet this limitation.

The author instanced springs operating in an ambient temperature of 200° C. losing their temper in 48 hours of continuous operation, and said this may indicate the need for new spring materials.

Magnetic circuits are more consistent and predictable than in the past, especially in the high-sensitivity and high-speed type of relays which use nickel irons. This left one remaining problem: that of sealing, and by following valve technique some relays are now enclosed in glass envelopes.

The author reviewed available general-purpose relays which follow the well-tried G.P.O. relays but with reduced dimensions. As with the G.P.O.

† Ericsson Telephones Ltd., Beeston, Notts.
U.D.C. No. 621.318.5

relay, none of them meet the vibration requirements for many applications, and this resulted in the development of three relays utilizing a rotary movement.

Several high-sensitivity relays were shown, most of them based on the Type 1A relay developed for the Services. This relay has a very small armature movement which, by a mechanical advantage, operates a single changeover contact with a power of 10 mW.

In the rather limited field of polarized relays the Siemens-Halske, Carpenter and Ericssons polarized relays were discussed. The magnetic circuit and other features of the Ericsson relay were described in detail and the following specific advantages claimed: a change in relative magnet intensities (there are two magnets) due to deterioration or polar gap changes can only affect sensitivity and not symmetry of adjustment; the relay is very stable and has two changeover contacts; the iron circuit, being closed, is not susceptible to outside interference; the armature is mass balanced.

In the field of sub-miniature relays only two designs were described: the Ericsson Type MGS and the E.M.I. Type 1B. The former has an armature across the top of the relay and is held in position by a circular spring. The relay operates in about 3 milliseconds with a contact gap of 4 mils. It has a glass envelope and is fitted with B7G valve base. The latter relay has a butterfly armature suspended on a centre wire, and the single change-over contact required 300 mW to operate.

Nuclear Radiation Damage to Electronic Components

E. H. Cooke-Yarborough*

Mr. Cooke-Yarborough dealt with the detailed effects of various types of nuclear radiation to which electronic components may be exposed under conditions where information concerning their behaviour would be particularly required. Notably he discussed the exposure of certain components to radiation close to a pile, in the interior of a pile, and in the neighbourhood of a nuclear explosion. He pointed out that the severity of conditions in the last case would usually be intermediate between the two other cases, which could be regarded as extremes.

It was frequently asked what would be the effect of nuclear radiation upon components in general. This question was too general and no answer could be given. Details of intensity spectrum and time exposure would be necessary. The spectrum was, physically speaking, of basic importance, ranging from a destructive point of view, from alpha particles with great effectiveness in imparting their energy, to the neutrino which is barely perceptible. The most important particles for the type of questions which appear to arise in practice were the neutrons and gamma rays, though the latter were chiefly effective by exciting beta radiation. The energy distribution of particles could have great effect and, for the present, a crude distribution into high and low energy neutrons could not be accepted as adequate; hence it would be difficult to give experimental results which could be of any precise value. Similarly, the total exposure would vary over an enormous range, perhaps by a factor of 10^{10}, according to whether the component was exposed to low radiation densities such as might appear outside a reactor, or whether it was placed in the interior, close to destruction point.

The most effective way of providing satisfactory variation in conditions is exposure at various points in or near a reactor. Further measurements immediately after exposure, such as are necessary in the study of radioactive decay, would require the manipulation of the unit in a highly reactive state, and it may even be necessary to introduce leads into the reactor.

The kind of radiation which would be found under various conditions may be briefly summarized. Close to the reactor, outside the shielding, the radiation intensity appears to have no destructive effects either on normal equipment or, so far as is now known, upon the human body. At the moment, film and ionization chambers must be used for detecting radiation under these conditions.

In the neighbourhood of a bomb explosion, conditions are more severe and, beyond a range of some 2,300 ft. corresponding to "complete destruction" for a 20 kiloton bomb, the effect falls off relatively slowly for heat and blast, but the specifically nuclear effects of neutron and gamma radiation fall off much more rapidly. Smaller distances would be of no consequence, since the destruction of all equipment would be complete. Exposure to radioactive cloud as is found in carrying equipment through this by aircraft is mainly due to gamma radiation. It was found that at 15,000 ft. an aircraft travelling through the cloud 90 seconds after the explosion

* Atomic Energy Research Establishment, Harwell.
U.D.C. No. 621.396.69:539.16

would still take up a dose of 55 r units, which would be lethal for a human being. Finally, the fall-out effects on the ground must be considered, but data are at present lacking. For larger explosions, the effects in the cloud are increased, mainly due to the increased duration of exposure; effects increase very roughly with the cube root of megatonnage.

Dosages in the interior of a reactor are much more severe, running at the rate of some 10^{18} neutrons per square cm, and even 10^{20} in the very interior, but electronic equipment is unlikely to be exposed here.

In assessing the effect on components of these highly varied doses, we are still further handicapped by our ignorance of the make-up of particular components. Small traces of sensitive elements, which may be unknown to the manufacturer, may well increase the sensitivity to radiation.

Turning to the results of such experiments as have been performed on particular components, much work has been done in America, some of which has been recently reported. The Harwell results may be briefly summarized as follows: A standard may be regarded as the exposure of various components for one second in the B.E.P.O. reactor, giving a dosage of 10^{12} neutrons per square cm (300 r). This was considered likely to be that which might be in practice met near an atomic explosion, but the gamma ray dosage would be lower. The majority of components tested withstood this exposure, though some change was found in electrolytic capacitors, tantalum and aluminium, and units enclosed in glass envelopes appeared to suffer some change. Otherwise the only very vulnerable unit was the transistor, and even here, whilst the properties changed, destruction was not necessarily complete and in certain respects performance might be improved.

The final conclusion was that destruction of electronic components without other large-scale destructive effects does not seem to be very likely, though changes in long-period behaviour must be expected.

Problems in Long-Term Component Reliability

Dr. K. E. Latimer*

Submarine cables present special requirements. Very long life is required under conditions of almost complete quiescence, although during laying, handling shocks may occur and the actual running off of the cable involves a spinning process which will produce large centrifugal forces.

Temperatures after laying are almost constant, but the sealing of equipment calls for particular care to eliminate any possible contaminating vapour. Similarly, internal heating must be kept to a minimum.

Well-proved components (20 years life) will only be abandoned very reluctantly but continuous life tests can be applied to determine possible effects on future components.

Particular causes of trouble to be watched for are all of the "creep" type, but the creeping may be exceptionally rapid.

"Whiskers" growth from soldered joints, electrolytic destruction of transformer windings in which d.c. is being held off, progressive failure of paper capacitors, are all accentuated by inaccessibility.

Only rigorous control of the components during assembly can identify and eliminate noxious contaminants or be prescribed as a way of minimizing these troubles.

* Submarine Cables Ltd., London.
U.D.C. No. 621.396.69:621.315.28

The author then gave a list of some of the more interesting deterioration phenomena:—

Effects of poisons in causing embrittlement of steel by hydrogen; also leakage of hydrogen into all-metal valves.

Voids in copper wire and grain growth.

Effects of organic vapour on relay contacts.

Copper as a catalyst in organic fluids; also burning the enamel off copper wire and quenching in alcohol.

Incipient cracks in carbon rod resistors.

Electrolysis of valve bases through the glass.

Operation of filaments at positive voltage to cathode.

Rapid failure of leaky valves in argon.

Diffusion of helium into valves and possible attack of filament. Other reactions of inert gases.

Effects of magnetic storms on operating voltages of capacitors.

Influence of iron on ceramic formers for pyrolytic carbon resistors.

Electrolysis in transformer and relay windings.

Erosion of glass seals in pickling acid.

Corrosion of silver plated wire by cadmium-bearing solder.

Slow changes of viscosity of silicone oil in the presence of lead oxide and release of gas.

Perforation of thick steel plate by burrowing clams.

AUTOMATIC GAIN CONTROL CIRCUITS IN TELEVISION RECEIVERS FOR NEGATIVE MODULATION SYSTEMS*

by

P. L. Mothersole (Associate Member)†

SUMMARY

The operation and results obtained with a representative selection of a.g.c. gate circuits are described. It is shown that a line-gated circuit is very susceptible to "lock-out" when large signal changes are made due to the loss of synchronization of the line time-base. The development of a simple high-gain gate circuit is described that is free from this fault and, in addition, permits improved video amplifiers to be designed.

1. Introduction

Signals received at the aerial terminals of television receivers are subject to variation in strength due to a variety of causes such as propagation changes, reflections from aircraft, different signal strengths for different transmissions (different programmes) etc. The installation and operation of the receiver is greatly simplified if the amplifier gain is automatically adjusted so that in a wide range of circumstances the picture has satisfactory contrast.

For a receiver to accept a wide range of input signals, without visible alteration of picture drive, a high gain receiver is essential in addition to a high gain a.g.c. system. The rate of control for a typical three i.f. stage receiver controlled on the r.f. and two i.f. stages is about 10 db/volt. Thus to obtain about 80 db of control some 11 volts of a.g.c. bias is required, allowing 3 volts for the r.f. delay. This range of automatic control is only possible with an amplified system.

It is very important however to ensure that the a.g.c. circuits themselves are free from vices. At the same time the performance of the receiver in "difficult" locations (e.g. close to a main road), should still be satisfactory.

This paper is concerned with circuits in receivers designed for use with negative modulation‡ which derive the control potential from the incoming signal. Circuits used in some current receivers are surveyed and finally an improved arrangement is described. This improved circuit derives the a.g.c. potential from the video signal black level. It is completely free from lock-out and in addition permits improved video amplifiers to be designed.

2. Existing A.G.C. Circuits

2.1. *Mean Level Circuit*

The simplest form of a.g.c. circuit is operated from the video detector. The video signal is negative-going, the tips of the synchronizing pulses being the most negative component of the signal and correspond to the maximum output from the transmitter. The negative d.c. component is filtered and used to control the gain of the i.f. and r.f. amplifiers. Such a simple solution however has three serious disadvantages, namely:

(1) The gain of the system is low since the change in a.g.c. potential is proportional to the variation in the video detector signal. Thus a wide range of automatic gain control is not possible.

(2) The a.g.c. potential is not only dependent on the received signal amplitude, but also on the video information. This produces an effect similar to a.c. coupling the video signal. To prevent the video information,

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‡ For a recent paper discussing the problems of a.g.c. with positive modulation, see S. N. Doherty and P. L. Mothersole, "A.g.c. in television receivers," *Journal of the Television Society*, 8, No. 9, pp. 350-367, January/March 1958.

in particular the 50 c/s frame blanking, remodulating the r.f. signal a long time-constant must be used.

(3) The a.g.c. circuit is slow acting, due to the long time-constant, and it cannot follow fast variations of the received signal.

It is common practice in receivers designed for a positive modulation system (i.e. British 405 line) to utilize the d.c. component of the video output signal at the control grid of the synchronizing pulse separator for the a.g.c. bias potential. The video signal at the grid is negatively d.c. restored by grid current and the use of the video output signal results in a higher effective gain in the system. However, with a negative modulated signal, this circuit is very susceptible to "lock-out" when a sudden signal change is made (i.e. by switching to a local station). The mechanism of the lock-out is as follows.

The sudden increase in signal overloads the video amplifier, cutting it off before the a.g.c. potential is able to develop. Since no video output signal is produced no a.g.c. bias is developed, hence a lock-out.

2.2. *Peak Detector Circuit*

A circuit similar in some ways to the mean level circuit is the peak detector, Fig. 1. The tips of the video signal are rectified by the peak detector V2, charging C1 to a d.c. potential equal to the peak signal potential. This potential is used as the a.g.c. potential. A delay potential is often applied to the peak detector diode to prevent it from conducting on small signals. This delay potential may be adjustable to form the contrast control.

The gain of this circuit is still low but the a.g.c. potential is proportional to the received signal potential and not to the video information.

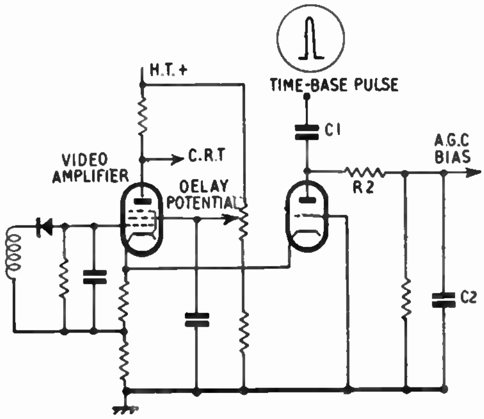

Fig. 2. Cathode coupled a.g.c. gate circuit.

With a negatively modulated system noise pulses often extend above the tips of the synchronizing signal. Such pulses will charge the capacitor C1 to the peak value causing the a.g.c. potential to "ride-up" following a burst of impulsive interference. This results in a lowering of the contrast and brightening of the picture following a burst of interference pulses. The effect may be minimized by using a very long time-constant filter circuit, but then the a.g.c. system will not be able to follow fast variations in signal strength, such as those caused by aircraft.

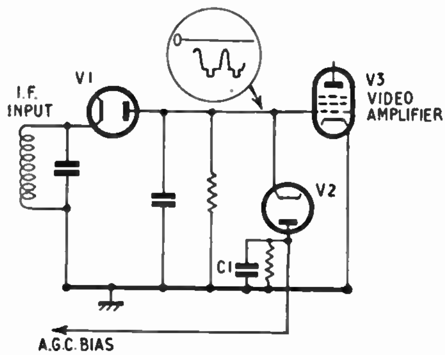

Fig. 1. Peak detector a.g.c. circuit.

2.3. *Cathode Coupled Gate Circuit*

To increase the loop gain of the system a pulse operated amplifier is often used. A simple form employing cathode coupling is shown in Fig. 2. The operation of this circuit is as follows.

The video signal from the cathode of the video amplifier is applied to the cathode of the grounded grid pulse amplifier V2. The amplifier is pulsed into conduction by a positive pulse obtained from the line time-base. The anode current charges the coupling capacitor C1 producing a negative potential at the anode. The anode current is dependent on the grid-cathode potential at the instant of the pulse. Providing the line time-base is correctly synchronized the anode current will be controlled by the d.c. level of the line synchronizing pulses. The negative potential at the anode is smoothed by R2 C2 and is used as the a.g.c. potential.

A positive delay potential is applied to the cathode of the amplifier, V2, to prevent it from conducting and producing any a.g.c. potential with low signal inputs. Variation of this potential enables the level of video signal at which the gate circuit operates to be adjusted (contrast control). In practice this potential may be adjusted by variation of the video amplifier cathode current, often by variation of the screen grid potential.

Noise pulses can only effect the a.g.c. potential if they occur at the instant the amplifier is pulsed into conduction. Thus the noise immunity of this type of circuit is good.

The effective gain of the system, however, depends on the working point of the video amplifier at the tips of the synchronizing pulses. It also depends, for its correct operation, on the synchronization of the line time-base, since this provides the gating pulse. The effects of these factors are discussed later (section 3).

2.4. *Grid Coupled Gate Circuit*

The effective gain of an a.g.c. system may be increased by deriving the control potential from the output of the video amplifier. A typical gate circuit is shown in Fig. 3. The video signal is applied to the control grid from the video amplifier's anode via the resistor R1. This resistor is used to prevent excessive grid current and, in conjunction with the input capacitance of the valve, to remove h.f. noise from the video signal. It also helps reduce the capacitive load on the video amplifier.

The valve is driven into conduction by a large positive pulse from the line output transformer. The anode current charges the coupling capacitor C1 producing a negative potential at the instant of the pulse. Providing the line time-base is correctly synchronized the anode current will be controlled by the d.c. level of the line synchronizing pulses. The negative potential at the anode is smoothed by R2 C2 and is used as the a.g.c. potential.

A positive delay potential is applied to the cathode of the amplifier which prevents the valve from conducting and producing any a.g.c. potential with low signal inputs. Variation of this potential enables the level of the video signal at which the gate circuit operates to be adjusted. If this potential is made more positive the amplitude of the video signal required to operate the gate circuit will be larger. Adjustment of this potential therefore forms a convenient contrast control.

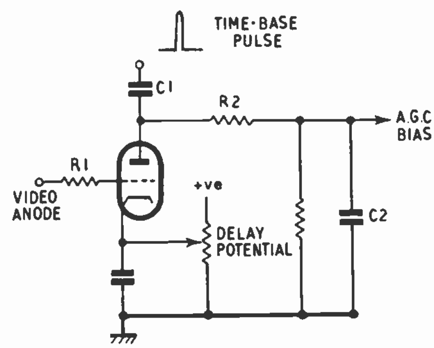

Fig. 3. Pulse operated a.g.c. amplifier.

The effective gain of this circuit also depends on the working point of the video amplifier. The circuit also depends on the line time-base synchronization for its correct operation.

3. Synchronization and "Lock-out"

3.1. *Line Time-base Synchronization*

In the gate circuits mentioned so far the a.g.c. potential that is developed will only be correct if the line time-base is correctly synchronized. If the line synchronizing fails the a.g.c. amplifier will be pulsed into conduction on the video signal. Since the signal at the video anode is more negative than the tips of the synchronizing pulses the a.g.c. potential will be reduced in value, increasing the amplitude of the video signal. If the contrast control had been adjusted to obtain a large video signal this increase in amplitude could overload the video amplifier. Since the video input is negative-going this will result in a reduction of the synchronizing pulse amplitude or the complete removal of the pulses at the anode due to the video amplifier being cut-off by them.

This loss of the synchronizing pulses will prevent the line oscillator pulling into synchronism. The receiver will therefore remain in this condition until manual adjustment of the contrast control increases the a.g.c. bias to remove the video amplifier's overload. This effect is often called lock-out or blocking.

Lock-out may occur each time a receiver is switched on, or when the channel switch is operated. If the full d.c. component of the video signal is applied to the cathode ray tube the overload will result in the picture blacking out. Upon changing stations or switching on therefore, the user may be presented with a blank screen.

3.2. *Prevention of Lock-out*

One method of preventing lock-out due to the cause described above is to ensure that the video amplifier can never be overloaded. This may lead to an uneconomical design, however, since the full current swing cannot be used under normal operating conditions.

Quite apart from this consideration, if the full current swing is used the tips of the synchronizing pulses approach cut-off under normal working conditions. The change in level due to variation in the input signal is therefore small resulting in a poor a.g.c. characteristic.

An alternative method often used to prevent lock-out is a combination of mean level a.g.c., from the video detector, and gated a.g.c., from the video output. If the line time-base synchronization fails, the negative potential at the video detector, which would normally increase and overload the video amplifier, controls the receiver gain.

3.3. *Flywheel Synchronization*

A method often used to ensure that the line hold control is adjusted to the centre of the catching range is to remove the synchronizing pulses when the hold control is adjusted. This technique is very satisfactory for ensuring the line hold control is at the centre of the catching range but it cannot be used with a normal line gated a.g.c. system, since when the line time-base is unsynchronized the a.g.c. potential varies violently and the user will be confused.

4. A.G.C. Gate Circuit Requirements

From the foregoing discussion of some a.g.c. circuits and their shortcomings the requirements of an a.g.c. gate circuit may be summarized as follows:

(1) The a.g.c. bias should be independent of the line time-base synchronization (line hold control).

(2) It should not require critical valve characteristics or component tolerances.

(3) It should have a high gain so that the a.g.c. potential may be produced with a minimum variation in the video signal.

(4) The circuit should be stable under all operating conditions and be completely free from lock-out or blocking.

(5) The circuit should be immune from the effects of impulsive interference.

(6) The a.g.c. circuit should simplify the operation of the receiver.

5. A Novel Gate Circuit

5.1. *Design Considerations*

To obtain a high loop gain the a.g.c. voltage should be derived from the video signal output. Due to the sense of the signal change at the video output, and its d.c. level, some form of pulsed amplifier is necessary. To ensure that the a.g.c. control potential is independent of the line time-base synchronization a pulse amplifier operated by the time-base must not be used as a gate.

An important advantage of a line gated system is its immunity to impulsive interference. If the above considerations preclude the use of a conventional gate then some other method of reducing the effects of impulsive noise such as a cancellation technique must be employed.

Furthermore, an a.g.c. system should maintain constant the black level of the signal at the output of the video amplifier. The problem therefore is to relate in the best possible manner the a.g.c. potential with the black level occurring in the back porch of the video signal.

A simple method of exposing the back porch of the video signal is to cancel the synchronizing pulses with the negative pulses at the anode of the sychronizing pulse separator. The most positive component of the resulting waveform is the black level, which may be measured by

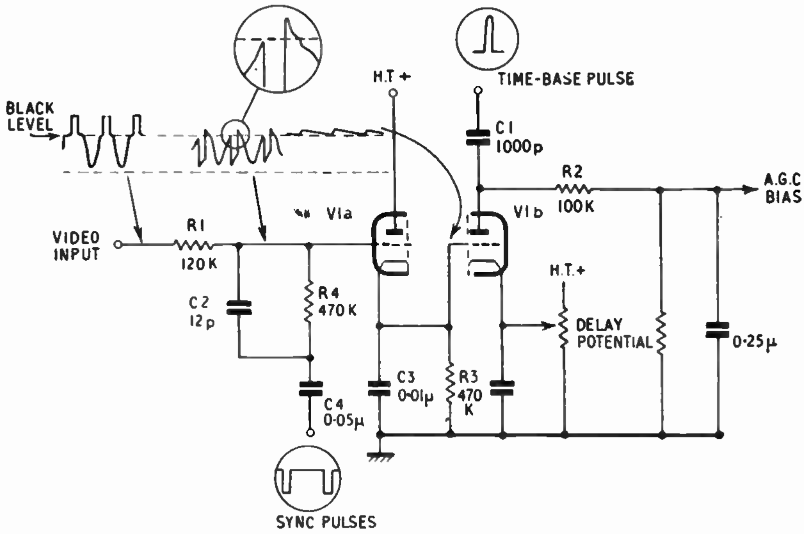

Fig. 4. Synchronizing pulse cancelled gate followed by d.c. amplifier.*

a simple peak detector. The output of the peak detector is a d.c. potential which corresponds to the black level of the video signal. This potential may be used to control a simple pulse operated amplifier.

If the a.g.c. potential is derived from the black level reference information, following the line sychronizing pulses, the loop gain would be slightly lower since black level (or suppression level) corresponds to approximately 70 per cent. of the transmitted power. The loss of gain would not be significant however.

5.2. *Circuit Description*

The circuit diagram of the gate circuit, based on the principles described above, is shown in Fig. 4.* Its operation is as follows.

The video signal from the video output stage is d.c. coupled to the control grid of the peak detector V1A via the resistor R1. The synchronizing pulses are cancelled by the negative-going pulses from the synchronizing pulse separator's anode. These are a.c. coupled to the grid by the network C4 R4.

The peak detector V1A charges the cathode capacitor C3 to the peak potential of the grid waveform (black level). During the line period there is some decay in this potential since the cathode time constant must be fairly short to obtain a fast acting system. Variation of the video signal close to the black level during the line can therefore influence the output of the peak detector. In a non-amplified system the effect would be negligible, but in this circuit the peak detector is followed a pulse operated amplifier. A small differentiated synchronizing pulse is therefore added to the composite waveform at the grid of the peak detector by the capacitor C2. This produces an overshoot on the waveform at the grid in the form of a positive pip on the back porch of the synchronizing pulse (black level), as shown in Fig. 4. The cathode capacitor C3 is therefore charged to a potential corresponding to the peak of this pip or overshoot. Variation of the video signal close to black level is therefore unable to influence the peak detector's output in spite of the small decay of the cathode potential during the line period.

The d.c. cathode potential, which corresponds to the video signal black level potential, is applied to the control grid of the pulse amplifier V1B which operates in the same way as the conventional gate circuit, Fig. 3. The positive delay potential is again applied to the cathode.

The line time-base pulses are only used to operate the d.c. amplifier and therefore the synchronization of the time-base does not affect the a.g.c. bias.

* Patent applied for.

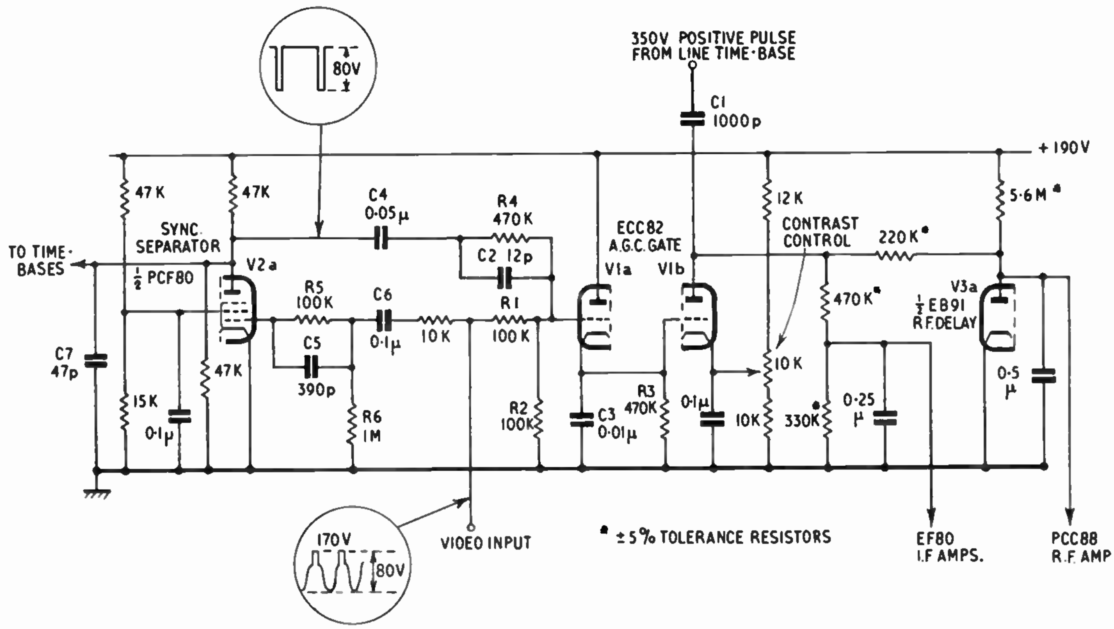

Fig. 5. Synchronizing pulse separator and a.g.c. circuit.

5.3. *Effect of Interference Pulses*

Since the video information to operate the a.g.c. circuit is obtained from the back porch of the video signal the tips of the synchronizing pulses may reach the cut-off point of the video amplifier and the a.g.c. circuit still operates satisfactorily. Interference pulses cannot affect the a.g.c. bias since they cancel in the same way as the synchronizing pulses. Blocking of the synchronizing pulse separator, and the consequent loss of the separated synchronizing pulses, must however be prevented.

This may be accomplished in several ways. Operation of the video amplifier so that the tips of the synchronizing pulses are close to cut-off automatically limits interference pulses to that of the synchronizing pulses. Alternatively a double time constant network may be used in the grid circuit of the synchronizing pulse separator, as shown in Fig. 5 (components R5 C5 and R6 C6).

5.4. *R.F. Delay Circuit*

In order to obtain the best noise factor for the receiver at low signal inputs it is desirable not to control the r.f. amplifier. The r.f. amplifier should only be controlled when the signal level is such that increased noise from the r.f. amplifier does not increase the noise output from the receiver. With a negative modulation system, using frequency modulation for the sound signal, the problem of cross modulation in the receiver is not so serious as for systems with a.m. sound. Sharp cut-off pentode valves (EF80) are normally used in the i.f. amplifier. Only two a.g.c. potentials are therefore required, one for the r.f. amplifier, with a fixed delay, and one for the i.f. amplifiers.

In order to obtain a wide range of automatic control, without cross-modulation, it is very important to ensure that as soon as the r.f. amplifier is controlled, its rate of control is such that the maximum bias is applied to both the r.f. and i.f. amplifiers together. A resistance network designed to provide the correct rates of control is shown, Fig. 6, together with a diagram showing the distribution of the r.f. and i.f. bias with input signal.

The following formulae enable the correct ratios of the resistors to be quickly determined. These formulae are derived in the Appendix.

Let V_r = r.f. bias
V_{rm} = maximum r.f. bias
V_i = i.f. bias
V_{im} = maximum i.f. bias
V_d = delay voltage (V_i at the instant the r.f. amplifier is about to be controlled)
V = input voltage
V_b = h.t. to delay resistor R4

then $$R_4 = R_3 \frac{V_b(V_{im} - V_d)}{V_{rm} \cdot V_d} - 1 \quad \ldots\ldots\ldots(1)$$

$$R_2 = R_1 \frac{V_{im} - V_d}{V_{rm} - V_{im} + V_d} \quad \ldots\ldots\ldots(2)$$

$$V_{max} = V_{im} \frac{(R_1 + R_2)}{R_2} \quad \ldots\ldots\ldots(3)$$

In a practical case using a PCC88 cascode valve and EF80 i.f. valves,

$V_{rm} = 5$ volts; $V_{im} = 5$ volts;
$V_d = 3$ volts; $V_b = 190$ volts.

let $R_1 = 470\ \text{k}\Omega$ and $R_3 = 220\ \text{k}\Omega$,

using equation (1) $R_4 = 5{\cdot}6\ \text{M}\Omega$
,, ,, (2) $R_2 = 330\ \text{k}\Omega$
,, ,, (3) $V_{max} = 12$ volts.

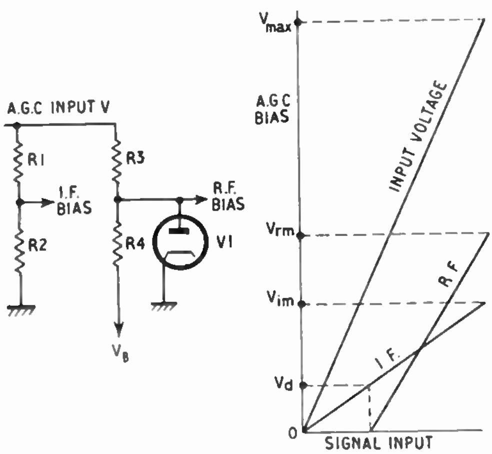

Fig. 6. Distribution network for use with sharp cut-off i.f. valves.

It should be borne in mind that the suitability of the bias potentials for the least cross modulation or modulation distortion must be considered in conjunction with the signal level distribution. Any leakage current through the delay diode V1 would unbalance the control potential ratios and therefore a thermionic diode or a semi-conductor diode with a very high back resistance should be used. It is not good design practice to use the control grid of the cascode valve as a delay diode, since the maximum wattage ratings of the cascode valve may be exceeded.

5.5. *Synchronizing Pulse Separator*

The synchronizing pulse separator associated with the gate circuit may influence the a.g.c. performance, and certain features of its design require to be checked. The grid circuit time constants must be chosen to minimize the possibility of grid blocking, as discussed in section 5.3.

The high gain a.g.c. circuit will maintain the video signal constant for very wide changes in the input signal. It is, however, desirable to operate the synchronizing pulse separator valve with a short grid base to ensure constant amplitude separated synchronizing pulses over a wide range of video signal levels, such as may be obtained by normal adjustment (or maladjustment) of the contrast control.

The pip or overshoot produced by differentiation of the trailing edge of the separated synchronizing pulse will determine the conduction time of the peak detector. The pip is only required to be sufficient in amplitude to prevent video signals near black level from causing the peak detector to conduct, as described in section 5.2. Adjustment of the differentiating capacitor (C2, Fig. 5) during the initial design is best carried out by observing the peak detector current on an oscilloscope. A 1 kΩ resistor connected in series with the anode of the peak detector valve forms a convenient monitoring point. The capacitor should be increased in magnitude until the waveform is just clear of picture signals.

If the trailing edge of the separated synchronizing pulse is very sharp the conduction time of the peak detector will be very short. The conduction time may be increased by slowing the trailing edge of the pulse by additional capacitance at the anode of the synchronizing pulse separator. In a practical circuit (Fig. 5) the conduction time was adjusted to about 2 microsec by the addition of the capacitor C7 (47 pF).

5.6. *Valve Operating Conditions*

The characteristics of the valves used in this circuit are not critical. The peak detector is in effect a cathode follower operating with a high value cathode impedance. The amplifier is pulsed into conduction by a large amplitude pulse to charge the coupling capacitor. The peak currents in both cases are very low. The peak detector current is about 6mA and that of the amplifier about 1 mA peak. The effect of variation in the grid bases of different valves is compensated for by the setting of the contrast control. In an experimental receiver the valve has been changed from an ECC82 to an ECC81, which provides a change in slope of 2:1 and a similar variation in the grid base with no noticeable effect.

In an a.c./d.c. receiver care must be used in placing the valve in the heater chain since hum produced at the cathode will modulate the a.g.c. potential. In practice, using an ECC82 valve, the heater has been connected some 30 volts above earth without producing any perceptible hum effects.

6. Experimental Results

6.1. *Receiver A.G.C. Characteristic*

The complete a.g.c. circuit used in an experimental 625 line C.C.I.R. receiver is shown in Fig. 5. The video signal is applied to the grid of the peak detector, V1A, by a resistive divider, R1, R2 (each 100 kΩ). This ensures that the voltage corresponding to black level at the grid of the peak detector is about 100 volts. The negative synchronizing pulses are a.c. coupled to the grid by the network C4 R4. The control bias was distributed to the r.f. and i.f. amplifiers by a resistive network as discussed in section 5.5. The receiver a.g.c. characteristic is shown in Fig. 7. This characteristic was obtained with a c.w. input corresponding to black level and free running the time-base at their correct frequency. It can be seen from these curves that the full a.g.c. potential may be produced for a change in signal level at the video detector of about 1 volt (2 db).

With a video modulated r.f. signal applied to the aerial sockets, via a 75 ohm switched attenuator, about 55 db of attenuation could be switched in with no visible effect on the picture, except for an increase in the noise level. Unfortunately the maximum output from the generator was limited, the maximum a.g.c. potential that was produced on the PCC88 valve being 3·8 volts. The amplitude of the video signal at the video amplifier's anode was 50 volts pk-pk; the change in its amplitude was 2 db for an input signal change of 55 db.

6.2. *Effects of Interference and Aircraft*

6.2.1. Impulsive interference

With a gated a.g.c. circuit only interference occurring during the synchronizing pulse can effect the a.g.c. potential. However, a series of repetitive pulses, such as those produced by a motor-car ignition circuit, can cause the a.g.c. potential to ride up. This produces a brightening of the picture following a burst of interference. With a high gain system the effect can be very serious. With the gated systems described in section 2, this effect may be prevented by using a long time-constant in the a.g.c. circuit. This, however, prevents the a.g.c. system from following fast variations in the signal strength.

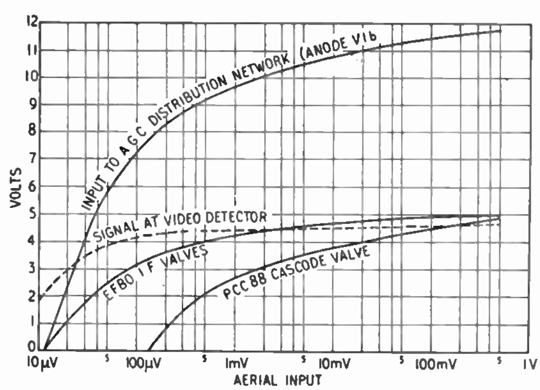

Fig. 7. Receiver a.g.c. characteristic.

With a synchronizing pulse cancelled system the effect is not so serious, since even if the level of the back porch is disturbed the d.c. output from peak detector will still correspond approximately to the black leved, since the charging capacitor will be charged by the black video signals or the front porch of the video signal. Furthermore, any interference pulses which extend into the synchronizing levels of the signal are unable to influence the a.g.c. circuit since they are cancelled in the same way as the synchronizing pulses. When field testing the experimental receiver no sign of ride up in the black level was observed with heavy interference inputs, and a short time-constant in the a.g.c. circuit.

6.2.2. Reflections from aircraft

An aircraft can produce a second signal at the receiver, due to reflections from the metal fuselage. The phase, or time delay, between the direct and reflected signal varies continuously as the aircraft moves. The effect of such a signal on an a.g.c. circuit depends on three factors: (*a*) the rate of change, (*b*) the phase difference or time delay, and (*c*) the relative amplitudes of the two signals.

To study these effects in the laboratory an Aircraft Reflection Simulator was designed.* This instrument incorporates a continuously adjustable mercury delay line, and enables a video modulated r.f. signal to be delayed up to about 300 microsec. This delayed signal is mixed with the non-delayed signal and adjustment is provided for the relative amplitudes and rate of change of phase of these signals.

When the phase difference between the two signals is very small a high gain a.g.c. circuit can reduce the flutter almost entirely. Nevertheless if the amplitudes are almost equal when the two signals cancel, the a.g.c. cannot prevent the complete loss of signal. In order that the a.g.c. control voltage can follow the fast variations in signal a short time constant is desirable.

With a longer time delay the video information is reflected into the back porch. It then depends on the video information how effective the a.g.c. can be in holding the picture steady. With a time delay of about 30 microsec (half line), and equal amplitude signals an a.g.c. system can, in fact, increase the amplitude of flutter. In practice the reflected signal amplitude falls as the time delay increases, and the a.g.c. circuit is then able to reduce the flutter.

In general a high-gain gated a.g.c. circuit reduces aeroplane flutter and it should be made as fast acting as possible.

6.2.3. A.g.c. circuit time constant

With conventional line gated circuit, of the type shown in Figs. 2 and 3, when the line time-base is unsynchronized the beat frequency between the line time-base and the synchronizing pulses is produced by the a.g.c. circuit. This beat frequency modulates the r.f. signal in the r.f. and i.f. amplifiers and a distorted output from the synchronizing pulse separator results. This makes synchronization of the line time-base more difficult. To prevent the effect a long time-constant must be used in the a.g.c. circuits.

With the synchronizing pulse cancelled circuit the sampling level, during the frame synchronizing period, will be the black level. The sampling rate will be 30 kc/s however. This increased sampling rate causes the output of the peak detector to ride-up producing a corresponding change in the a.g.c. potential.

The a.g.c. time-constant requires to be as fast acting as possible, to control fast changes in the signal strength. The time-constant in practice, however, is limited by the distortion that may occur due to the frame synchronizing signal. The sensitivity of the controlled valves to modulation via the a.g.c. line goes down as the control bias is increased. However the sensitivity of the a.g.c. gate circuit increases at a faster rate. The smoothing time-constant of an a.g.c. circuit should therefore be adjusted in a complete design with a strong signal input.

7. Conclusions

The problem of a.g.c. in a receiver designed for a negatively modulated system has been discussed. It has been shown that the dependence of the correct a.g.c. potential on the line time-base synchronization is undesirable, since it can make the receiver more difficult to operate and can also result in a lock-out.

With the simple gate circuit described, the a.g.c. potential is dependent on the black level of the signal and is completely independent of the line time-base frequency (line hold control).

In addition it has the following advantages:

(1) The circuit is d.c. coupled and is completely free from lock-out, paralysis or blocking.

* M. C. Gander and P. L. Mothersole, "An aircraft simulator for television signals," *Electronic Engineering*, **30**, June 1958.

(2) The input impedance is high and no significant capacitive load is added to the video amplifier.

(3) The circuit uses relatively few valves and components and does not depend critically on the valve characteristics or the components tolerances.

(4) The principle of synchronizing pulse cancellation is used in the gate circuit which increases the circuit's immunity to impulsive interference.

The circuit has been field tested in an experimental receiver and these tests confirm the laboratory measurements. From a non-technical user's point of view the advantage in having the picture contrast and brightness completely independent of the line hold control is very considerable.

8. Acknowledgments

The author would like to acknowledge the contribution made by Mr. T. Pascoe in the development of the circuit described in this paper.

The author also wishes to thank the Director of the Mullard Research Laboratories and the Directors of Mullard Limited for permission to publish this paper.

9. Appendix

The formulae for determining the correct ratios of the resistors in the a.g.c. distribution network are determined as follows:

Let V_r = r.f. bias
V_{rm} = maximum r.f. bias
V_i = i.f. bias
V_{im} = maximum i.f. bias
V_d = delay voltage (V_i at the instant the r.f. amplifier is about to be controlled)
V = input voltage
V_b = h.t. applied to the delay resistor R4

The required distribution of a.g.c. bias with input signal is shown in Fig. 6, together with a resistor network which enables it to be achieved. The rate at which the a.g.c. bias controls the i.f. and r.f. amplifiers is determined by the resistor ratios. The rate of control of the i.f. amplifier is $\frac{dV_i}{dV} = \frac{R_2}{R_1+R_2}$ and the r.f. amplifier is $\frac{dV_r}{dV} = \frac{R_4}{R_3+R_4}$

It can be seen from Fig. 6 that the ratio of these control rates is $\frac{V_{rm}}{V_{im}-V_d}$

therefore

$$\frac{R_4}{R_3+R_4} = \frac{V_{rm}}{(V_{im}-V_d)}\,\frac{R_2}{(R_1+R_2)} \quad \text{.........(1)}$$

The r.f. amplifier will start to be controlled when the current in R3 is equal to the current in R4.

This current is equal to V_b/R_4 amps. The input voltage V at this instant is equal to $V_b.R_3/R_4$ volts. The i.f. bias at this instant is equal to V_d.

Therefore

$$V = V_b\,\frac{R_3}{R_4} = V_d\,\frac{(R_1+R_2)}{R_2}$$

rewriting

$$\frac{R_2}{R_1+R_2} = \frac{V_d}{V_b}\cdot\frac{R_4}{R_3} \quad (2)$$

Using equation (2) in (1) and rewriting, the ratio of R_3 to R_4 is determined.

$$R_4 = R_3\,\frac{V_b(V_{im}-V_d)}{V_{rm}\,V_d} - 1 \quad \text{.........(3)}$$

Using equations (2) and (3), and the approximation that $V_d/V_b = 0$ the ratio of R_1 to R_2 can be determined:

$$R_2 = R_1\,\frac{V_{im}-V_d}{V_{rm}-V_{im}+V_d} \quad \text{.........(4)}$$

The maximum input voltage V_{max} is given by

$$V_{max} = V_{im}\,\frac{(R_1+R_2)}{R_2} \quad \text{.........(5)}$$

Equations (3), (4) and (5) are the design equations used to determine the ratios of the distribution resistors.

DISCUSSION ON
"Detection of Pulse Signals in Noise: The Effect on Visual Detection of the Area of the Signal Paint" *

D. C. Fakley†: With certain intensity-modulated displays it is very noticeable that improved detection results from viewing the records at a distance probably due, as Mr. Griffiths suggests, to the integration of the noise background. This observation does not however necessarily conflict with Blackwell's results.[1] In those experiments the noise background was essentially provided by the observer's vision process so that increasing the viewing distance would not decrease the effective noise r.m.s. value of the background to compensate for the decrease in the angle subtended by the "signal" paint. It would seem quite possible that with an actual display under some conditions, the decrease in r.m.s. noise produced by increasing the viewing distance would more than compensate for the decrease in the area of signal paint; this would be so if the initial conditions corresponded to those of large signal area and high illumination as in Fig. 4 of Mr. Griffiths' paper.

I have an idea that the explanation of the subjective experiment on tilting a long chemical recorder trace to reduce the horizontal angle subtended at the eye to improve detection would be forthcoming if Blackwell's experiments were repeated with other than circular signal paints. I would suggest that for such experiments the area of a signal paint would not be a sufficient measure but that the "shape" of the signal area would have to be considered; it would seem to me that serious discrepancies could result from equating a long narrow rectangular signal paint as on a chemical recorder with a circular paint of the same area and brightness against a given noise background. It could possibly be inferred from Blackwell's results that increasing a linear dimension of the signal paint beyond, say, between 30 and 100 minutes of arc depending on the background would result in a very small gain in input threshold. If this is so, foreshortening a chemical recorder trace by tilting it so that the angle subtended by the signal paint was reduced to this order would not result in any serious deterioration in threshold due to area decrease but would yield a 1·5 db depression of the threshold per halving of the effective trace length due to visual integration.

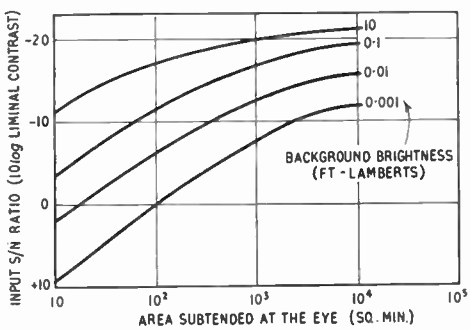

Fig. 4.* Blackwell's results[1] plotted as signal-to-noise ratio versus area for various values of background illumination.

Recently I have analysed a detectability experiment carried out with an intensity-modulated display, the recording medium being Teledeltos paper which was viewed in ordinary daylight. On this display 24 traces were recorded side-by-side. It is a little difficult to define the area of a signal paint on a single trace but it was certainly not less than 0·3 in. × 0·1 in. in the worst case and at times it was as large as 0·3 in. × 1·0 in. It will be seen that these correspond to the large area end—and high illumination—of Fig. 4. Thus it would be expected that the depression of the threshold due to having 24 traces instead of one would owe very little to the effective increase in signal area. This has been confirmed by an analysis based on ideal observer theory in which it has been assumed—effectively—that a visual-integration gain of 1·5 db is obtained for each doubling of the

* J. W. R. Griffiths, *J.Brit.I.R.E.*, **17**, pp. 330-338, June 1957.

† Received 5th February 1958. Mr. Fakley is at the Admiralty Research Laboratory, Teddington, Middlesex.

U.D.C. No. 621.396.963

number of traces and visual-correlation effects have been neglected; the actual analysis is not completely straightford since, for this experiment, the 24 traces were not independent of each other. The correspondence between the estimated thresholds for 50 per cent. detection probability and the experimental results is surprisingly good over a fairly wide range of the display parameters. It would therefore appear that these experiments afford some confirmation that it is possible to get beyond the region where the 2·2–2·5 db per doubling of traces law holds and have reached the state where the simple expression of 1·5 db per doubling of traces can be applied. I cannot be certain of this because it was not possible to vary the number of traces on the display but the evidence is fairly strong.

J. W. R. Griffiths (*in reply*): I think that Mr. Fakley's explanations of the effects described are very reasonable. He suggests, and in my mind quite rightly, that it is not only the area of the signal paint which effects detection, but that the shape also plays some part, and, in fact, I have been hoping for some time to perform some experiments to obtain some quantitative data and so elucidate this point. In the experiments of McGregor,[2] the signal paint was rather elongated, but nevertheless the law of improvement in detectability as the area was increased was similar to that obtained by Blackwell.[1] Since the width of the signal area was not altered during the experiment this evidence is not conclusive, and unfortunately an absolute comparison could not be made for the reasons stated in my paper.

An important point that does arise from these experiments is related to the concept of a critical visual angle, above which the rate of improvement in detectability with increase in angle falls rapidly. In Blackwell's experiment this occurs at about 10-12 minutes of arc. However, in both McGregor's experiment and in that described by Lawson and Uhlenbeck[3] the improvement in detectability is maintained beyond this critical angle. This suggests that, since in both experiments the trace is fairly long and narrow, the linear critical angle may be increased provided that a critical area (expressed in angular units), is not exceeded. As discussed in my paper we can think of the signal paint as consisting of a large number of individual elements, the size of these elements being governed by the size of the noise pulses on the display. Thus a limitation in area rather than linear dimensions corresponds to a limitation imposed by the brain on the maximum number of independent elements with which it can deal. This could still allow the shape of the signal paint to have an effect since the spatial distribution of these elements may effect detection.

References

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2. P. McGregor, "A note on trace-to-trace correlation in visual displays," *J. Brit.I.R.E.*, **15**, pp. 329-331, June 1955.
3. J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals," p. 242. (McGraw-Hill, New York, 1950.)

. . . Radio Engineering Overseas

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621.318.1.002

Technology of manufacturing barium ferrite. R. Lappa. *Prace Instytutu Tele- I Radiotechnicznego (Warsaw),* **2**, pp. 17-43, 1958.

Magnetic properties of barium ferrite are briefly surveyed and the Institute's research work on technology of obtaining barium ferrite reported. Successive stages of technological process are discussed: (1) initial raw materials used, method of mixing them and grinding, and composition of mixtures; (2) preliminary sintering of the mixture with special emphasis on the kinetics of the process; (3) subsequent grinding of the presintered product; (4) pressing and final sintering of shaped pieces of barium ferrite. Results of measurements of magnetic properties of the product are given and investigation of its microstructure reported. Separate stages of the process are discussed from the point of view of achieving the best magnetic performance of barium ferrite while keeping the technological process within reasonable limits of economy of time and means. Isotropic barium ferrite obtained in the process described shows the following properties: remanent induction 2100 gauss, coercive force 1600 oersteds and $(BH)_{max}$ 1 × 10^6 gauss-oersteds.

621.319.43

Coaxial trimmer capacitors with glass or quartz dielectrics. L. Klauz, M. Schmiedhamer and J. Smetana. *Slaboproudy Obzor (Prague),* **19**, pp. 122-128, March 1958.

The capacitors use low-loss lead glass or clear fused quartz tubular dielectric. The change in capacitance is produced by axial shift or rotation of a precisely calibrated pivot inside the tube, changing thereby the surface of overlapping with the stator formed of a baked silver layer on the outer surface of the tube. The method of calibrating the pivot and tube is described and a survey of the capacitor characteristics is presented.

621.372.8:621.3.028.4

A waveguide load made from commercial film resistors. U. v. Kienlin and A. Kurzl. *Nachrichtenteechnische Zeitschrift,* **11**, pp. 138-141, March 1958.

A discussion on various constructions of waveguide loads for microwaves is continued with a description of a design in which commercial cylindrical film resistors are used. Loads of this type have been manufactured for the waveguides 58 mm × 29 mm and 34 mm × 15 mm. The reflection coefficient is less than 1% over a relative frequency band of 15% or 30%. These loads can be loaded with more than 20 watts because of the excellent heat conduction. This new design is compared with known constructions which are manufactured from glossy material of flat films.

621.373.42

The effect of surface resistance of attenuators in travelling wave tubes on gain and stability. W. Eichin and G. Landauer. *Nachrichtentechnische Zeitschrift,* **11**, pp. 131-137, March 1958.

The attenuation as a function of surface resistance on an attenuator in a travelling wave tube with predetermined helix dimensions passes through a maximum. For obtaining a certain attenuation the surface resistance of the attenuator for the helix can be chosen so that the resistance is either higher or lower than the value corresponding to this maximum. The choice of a larger value is more advantageous than the other possibility because the gain and the stability of tubes are increased. The maximum of attenuation for coaxial, cylindical attenuators in the case of usual helix dimensions is found at surface resistances of approx. 2·5 kΩ. In other non-coaxial attenuator arrangements the maximum of attenuation lies at lower values for the surface resistance.

621.373.421

A warble tone generator. R. R. Dutta Gupta. *Indian Journal of Physics,* **32**, pp. 75-79, February 1958.

A warble tone generator using simple phase shift oscillators, is described. Two similar oscillators are used, one having a very low fixed frequency corresponding to the frequency of wobbulation. This low frequency oscillator voltage is applied to the grid of a triode which is used as a variable resistance element in the main oscillator of varying frequency. The relative importance of the different *R-C* sections of a phase-shift oscillator from the point of view of frequency stability has also been discussed. It has been shown, for example, that the variation of the first resistance of the three sections gives maximum percentage variation of frequency.

621.373.431

The production of extremely steep pulse edges in multi-stage non-linear amplifiers. G. Kohn. *Archiv der Elektrischen Übertragung,* **12**, pp. 109-118, March 1958.

A generator is described in which the edge of a multi-vibrator pulse is steepened up to F_{max} in three amplifier stages. The transformers which must be used between the stages for phase inversion cause an exponential droop of the pulse top. By the addition of a complementary exponential pulse this droop can be compensated for. Into a 60-ohm load such a generator can feed either long pulses (of a duration up to 50 μsec) with an amplitude of 100 V and a rise time of < 3 mμsec, or short 50-V spikes, shaped by a section of line, with a base width of < 3 μsec. The repetition frequency can be chosen as desired up to 10 kc/s. The outgoing pulse can be delayed up to 1 μsec with respect to the sync. pulse; jittering is less than 1 μsec.

621.373.431.1

Stable flip-flop circuit for measuring instruments. K. KOSTAL. *Slaboproudy Obzor (Prague)*, **19**, pp. 150-156, March 1958.

First an approximate calculation of the stability of two basic flip-flop circuits is given, dependent on the supply voltage fluctuations and on changes of the tube parameters. The calculations indicate that these circuits are not suitable for more accurate measurements. The design of a flip-flop circuit is suggested, from which the main reasons for instability are eliminated. It follows from the calculations and the measurements described that this circuitry is more stable, and that under certain conditions it is applicable for more accurate measurements.

621.383.4

A survey of picture storage tubes with reproduction of gradations. H. G. LUBSZYNSKI. *Nachrichtentechnische Zeitschrift*, **11**, pp. 115-124, March 1958.

Storage tubes with reproduction of graduations may be classified according to their method of operation. Some of the problems encountered during the design and operation of these tubes as well as the limits for their capacity of storage are described. Recording velocities of 30 to 100 picture elements per microsecond and total reproduction periods from fractions of a second to several hours have been obtained. The limits for the resolution lie between 400 and 2,000 lines. The erase times are still too high—with few exceptions—for television broadcasting. Calculations have revealed that it is possible to increase the capacity of storage and the sensitivity of tubes with photo-conducting storage films when secondary emission multipliers are incorporated. Storage tubes for direct viewing permit the simplification of ancillary equipment required. A number of practical applications for storage tubes are described.

621.385.632.5:621.373.423:621.376.3

The linearization of frequency modulated reflex klystrons. E. SCHUON and H.-J. BUTTERWECK. *Archiv der Elektrischen Übertragung*, **12**, pp. 99-108, March 1958.

The reflex klystron can be modulated in frequency in a particularly simple and wattless manner by superimposing the modulating voltage on to the reflector direct potential of the reflector electrode. The limits of the range that can be utilized for frequency modulation are given by the curvature of the modulating characteristic. This characteristic can be linearized over wide limits by special circuitry which in literature is frequently referred to as "connection of a frequency-dependent load." The study is general and shows to what extent the linearity can be improved or the frequency swing increased.

621.385.2:537.525

The Townsend discharge in a coaxial diode with axial magnetic field. P. A. REDHEAD. *Canadian Journal of Physics*, **36**, pp. 255-270, 1958.

An approximate theory is developed of the breakdown characteristics of a coaxial diode in an axial magnetic field, taking into account the effects of elastic collisions. It is assumed that the electron moves in a constant electric field between collisions and thus the theory is valid only in the appropriate range of magnetic field and voltage. Estimates of transit time and of space-charge effects are also made. Measurements in the pressure range 10^{-3} to 10^{-9}mm Hg are in general agreement with the theory.

621.396.41

Multichannel u.h.f. radio telephone equipment. J. FIEGUTH. *Proceedings of the I.R.E. Australia*, **19**, pp. 43-53, February 1958.

The paper describes multichannel radio telephone bearer equipment which has recently been developed and manufactured for the Australian Post Office. The equipment is frequency modulated and operates in the 900 Mc/s band. It has been designed to carry 12 telephone channels but can be used for up to 36 channels. Some aspects of the performance of the equipment are discussed and it is shown that the international standards for trunk-line radio link equipment are met.

621.396.828:654.19

A new method for the automatic monitoring of broadcasting systems. F. ENKEL. *Nachrichtentechnische Zeitschrift*, **11**, pp. 142-147, March 1958.

An automatic device for the supervision of the electro-acoustic properties of transmission systems, including the high-frequency radiation, is described. Supervision covers the frequency response of the transfer constants, the continuous measurement of noise and the testing of non-linear distortion. Neither a reference programme nor additional bandwidth in the transmission channel are required. Faults are detected with at least the same reliability and accuracy as is possible for a trained listener. The tests are carried out during normal transmissions.

621.397.331.3

Recording television programmes on film in the German Federal Republic. J. GOLDMAN and H. FUNK. *E.B.U. Review*, **48**, pp. 2-9, March 1958.

The different systems used by the broadcasting corporations are described and critically assessed. Operational practice is described and a brief account of measuring and test methods is given.

621.397.62

Method of interlace scanning measurement in television receivers. J. KAMLER. *Prace Instytutu Tele- i Radiotechnicznego (Warsaw)*, **2**, pp. 71-81, 1958.

A method of determining the quality of interlace scanning is described which is based on objective measurement with an oscilloscope. Controlling instruments do not need to be connected inside the receiver and therefore the measurement can be performed during normal operation of the receiver. Definition of interlace scanning coefficient is given as well as mathematical basis of the method. General description of the measuring equipment and the results of measurements for several receivers of various types are also included.

681.142

An electronic differential analyser. A. K. CHOUDHURY and B. R. NAG. *Indian Journal of Physics*, **32**, pp. 91-108, February 1958.

An electronic differential analyser having sixteen operational amplifiers is described. The analyser is specially designed for solving equations arising in connection with research works on circuit theory, servo-mechanisms and electron trajectories. The analyser works both repetitively and non-repetitive. Different sources of error of the computer elements are also discussed. Accuracy of the solutions obtainable by the machine is illustrated by some examples.