# THE RADIO AND ELECTRONIC ENGINEER

## The Journal of the Institution of Electronic and Radio Engineers

FOUNDED 1925 INCORPORATED BY ROYAL CHARTER 1961

*"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."*

VOLUME 27 APRIL 1964 NUMBER 4

## THE PATTERN OF FUTURE CONVENTIONS

In the provisions of the Institution's Charter it is specified that the first object for which the Institution is constituted is to provide means for facilitating the acquisition of knowledge. The Council is constantly considering the best means by which this object may be discharged, apart from publishing *The Radio and Electronic Engineer* and *The Proceedings*.

One of the most important features of the "learned society" activities of the Institution is the holding of regular meetings of members. Usually these meetings are devoted to the presentation of a single paper, but a recent tendency in London and the Local Sections has been to arrange an afternoon and evening, or a whole day meeting for members to hear and to discuss a group of related papers.

In short, an increasing tendency has been to hold Symposia as distinct from Conventions. The difference between these two types of meeting is that a Symposium deals with a definite theme in depth rather than in breadth, whilst a Convention is broader in concept and may cover several subjects over a period of three or more days.

The Symposium is now a well established activity of the Institution. The duration of some may be one or two days, but others may extend over several days and be residential; the recent Symposium on 'Cold Cathode Tubes and their Applications', reported elsewhere in this issue, is typical of this more ambitious type of Symposium.

Recently the Council has been discussing the form which future Conventions should take and has considered the merits of an approach whereby each of the three or four days should be devoted to particular subjects. These individual days may be regarded as Symposia in their own right. By arranging that the successive themes are related there would be added incentive for those wishing to attend the Convention to stay for two or more days, whereas if the overall programme were not thus related, the exigencies of professional life might militate against spending any time at the Convention.

It is interesting to note that the Joint Organizing Committee for the Symposium on 'Signal Processing in Radar and Sonar Directional Systems' to be held at the University of Birmingham in July is adopting a rather similar approach for the benefit of those who cannot spare four days at the meeting. The first two days will deal mainly with Radar and Radio Astronomy aspects, while the second two days will deal mainly with aspects affecting Sonar, Ultrasonics and Seismology.

The opportunities which a 'multilateral' Convention would provide for engineers from different branches of electronics and radio to meet one another, both in the formal sessions and informally in the congenial surroundings which can be provided by one of our Universities, would be particularly valuable. The Council believes that selection of suitably associated themes to group together in Conventions of this kind would meet the needs of the engineer today to keep in touch with modern developments.

One of the advantages of a Convention is that it also gives greater opportunity for members to meet socially and to exchange individual ideas which may best help them in their professional activity.

The policy of the Institution in this matter must always be subject to change in the interests of the profession generally. The Council has quite an open mind as to future policy and is anxious to secure the views of members before taking a decision. These comments are, therefore, an invitation to members to give their views on the type of meetings which makes most appeal.

G. D. C.



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# INSTITUTION NOTICES

### Symposium on 'Seismic Signals'

A Symposium on 'Modern Techniques for Recording and Processing Seismic Signals' will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, on Wednesday, 13th May, from 10 a.m. to 5 p.m. Papers to be presented will discuss applied seismology, seismometer arrays and associated equipment, and seismic data processing. Advance registration will be necessary; further details and application forms may be obtained from the Institution, 8–9 Bedford Square, London, W.C.1.

### Clerk Maxwell Memorial Lecture

The fifth Clerk Maxwell Memorial Lecture will be given by Sir Gordon Radley, K.C.B., C.B.E., Ph.D., M.I.E.E., on Tuesday, 26th May at 6 p.m., at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1. Sir Gordon is Chairman of the Marconi Company and was previously Director-General of the Post Office; he is a Past President of the Institution of Electrical Engineers. Tickets will be required for this meeting and may be obtained from the Institution.

### The International Conference on Magnetic Recording

The International Conference on Magnetic Recording, which will be held at the headquarters of the Institution of Electrical Engineers, Savoy Place, London, from 6th–10th July 1964, will cover all methods of magnetic recording on moving media. The Conference is being organized jointly by the I.E.R.E., the I.E.E. and the European Region of the I.E.E.E. The nine sessions of the Conference will cover the following:

| | |
|---|---|
| Recording media | Audio-frequency recording |
| Analogue recording | Video-frequency recording |
| Digital recording | General problems including tape-transport systems |

The following broad subjects and applications will be discussed:

Applications of digital and analogue recording techniques in such fields as geophysics, aeronautical engineering, echo sounding; Characteristics of magnetic tape and film; Correlation techniques; Crosstalk suppression; Data acquisition and interchange; Design and characteristics of heads and associated electronic equipment; Drum stores; Editing and indexing; Electro-mechanical design; Erasure; Operational practices; Physical and digital density; Recording by frequency modulation; Recording of colour and monochrome television signals; Stereophonic transmission measurements; Studio equipment.

A programme of technical visits and social functions will be associated with the Conference and items of equipment in support of papers will be exhibited.

Further details and registration forms can be obtained from The International Conference on Magnetic Recording Secretariat, c/o The Institution of Electrical Engineers, Savoy Place, London, W.C.2.

### U.K.A.C. Lectures by a Russian Engineer

Professor Mark Aizerman, Dr.Techn.Sci., of the Institute of Automation and Telemechanics, Moscow, is expected to visit London at the end of May and beginning of June this year, when it is hoped to arrange for him to deliver two or more lectures under the aegis of the United Kingdom Automation Council.

The provisional arrangements are:—

*Thursday*, 28*th May, at* 5.30 *p.m.*, at the Institution of Mechanical Engineers, Birdcage Walk, Westminster, London, S.W.1. Lecture by Dr. Aizerman on "Pneumatic Automation: its Possibilities and its Future".

*Tuesday*, 2*nd June, at* 5.30 *p.m.*, at the Institution of Electrical Engineers, Savoy Place, London, W.C.2. Lecture by Dr. Aizerman on "Non-linear Stability: the Present State of the Art" (including a discussion of the 'Aizerman Problem').

Members of the I.E.R.E., which is a member society of the U.K.A.C., may attend these meetings; those proposing to do so are recommended to communicate with the Honorary Secretary of the U.K.A.C. (Tel.: Covent Garden 1871) shortly beforehand, to ascertain whether these proposed arrangements have been confirmed.

### Symposium on 'Electronics in the Automobile Industry'

Preprints of nine papers presented at the above Joint I.E.E.—I.E.R.E. Symposium, held in Birmingham on 7th April 1964, are now available in booklet form, price 5*s*. including postage, from Mr. G. K. Steel, A.M.I.E.E., Department of Electrical Engineering, College of Advanced Technology, Gosta Green, Birmingham 4. The papers review applications of electronics to automobile research, development, production and traffic control.

### Symposium on 'Signal Processing in Radar and Sonar Directional Systems'

The Institution is joining with the Department of Electronic and Electrical Engineering at the University of Birmingham in organizing a Symposium on 'Signal Processing in Radar and Sonar Directional Systems' at the University of Birmingham from 6th–9th July 1964. A detailed programme and registration form is enclosed with this issue; additional forms are available on request from the Institution, 8–9 Bedford Square, London, W.C.1.



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U.D.C. 621.397.334

# Cross-colour and Luminance Twinkle in N.T.S.C. Colour Receivers with Various Display Devices

By

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**Summary:** It has been observed that under some conditions N.T.S.C. colour receivers employing single-gun tubes exhibit severe cross-colour and luminance twinkle due to the presence of high video-frequency luminance information and noise. From analysis of a simple case it is established that this effect is the result of sampling of the high-frequency luminance and therefore noise components by the colour selection process (necessary to decode the subcarrier chrominance information) which takes place at the screen. Methods of eliminating or substantially reducing this objectionable effect are discussed.

## 1. Introduction

It is well known that in N.T.S.C. colour receivers employing three-gun displays, such as the shadow-mask tube, luminance signal components with a frequency near to that of the colour subcarrier are synchronously detected and result in a visible low-frequency colour beat pattern in the picture known as 'cross-colour'. Similarly noise components near to the subcarrier frequency give rise to low-frequency coloured noise, known as 'parc', which is subjectively more noticeable than the noise on monochrome receivers.[1, 2]

In work on a number of N.T.S.C. receivers employing single-gun tubes, and with video gated operation of such displays, it has been observed that under certain conditions the cross-colour and 'parc' noise are much worse than with a three-gun display and are accompanied by severe small area brightness fluctuations which can best be described as 'twinkling'.

To establish the origin and magnitude of these effects they have been calculated for the very simple case of a luminance signal component near to the subcarrier frequency on a plain white field. The calculation has been carried out for a number of types of single-gun colour receiver configuration and also for a three-gun receiver, as a basis for comparison. Although a simple case is cited for ease of computation, the results also give a general qualitative analysis of the complex noise-beating effects which have been observed.

To simplify the analysis further, N.T.S.C. primary colour phosphors and a gamma of 2 are assumed for all displays.

## 2. Effects with a Three-gun Display N.T.S.C. Receiver

The most visible cross-colour effects which occur in an N.T.S.C. receiver using a three-gun display are those due to luminance (and noise) components near to the subcarrier frequency. It therefore seems reasonable to assume for the purposes of calculation that the chrominance signal may be represented by the expression which is valid over the double sideband region of the $I$ component.[3, 4]

In this case the composite signal may be written:

$$E'_N = E'_Y + 0{\cdot}49(E'_B - E'_Y)\sin\omega_2 t + 0{\cdot}88(E'_R - E'_Y)\cos\omega_2 t \quad \ldots\ldots(1)$$

where $f_2 = \omega_2/2\pi$ is the subcarrier frequency

and $E'_Y = 0{\cdot}30E'_R + 0{\cdot}59E'_G + 0{\cdot}11E'_B \quad \ldots\ldots(2)$

The primes denote gamma corrected signals, so that if we assume $1/\gamma = \frac{1}{2}$, then $E'_R = E_R^{\frac{1}{2}}$, etc.

Now suppose that the luminance signal $E'_Y$ contains no components near subcarrier but that such a signal $E_S \sin\omega_1 t$ is present (either as part of the luminance signal or as an interfering signal) where $\omega_1$ is near to $\omega_2$ such that

$$\omega_1 - \omega_2 = \Delta\omega \quad \ldots\ldots(3)$$

Then the input signal to the receiver decoder may be written

$$E'_T = E'_Y + 0{\cdot}49(E'_B - E'_Y)\sin\omega_2 t + 0{\cdot}88(E'_R - E'_Y)\cos\omega_2 t + E_s \sin\omega_1 t \quad \ldots\ldots(4)$$

To recover the primary red, green and blue signals the composite signal may be synchronously detected along the $(R-Y)$ and $(B-Y)$ axes, followed by appropriate matrixing and amplification. In general other demodulation axes may be employed in the receiver but it may readily be shown that apart from differences in colour difference-signal bandwidths and quadrature crosstalk, neither of which is very relevant to the present study, the end result is the same.

For convenience product-demodulation along the $R-Y$ and $B-Y$ axes will be assumed. Then multiplying by $A\cos\omega_2 t$ the output from the $R-Y$ channel is given by

$$(R-Y) = A[(E'_Y \cos\omega_2 t + 0{\cdot}49(E'_B - E'_Y)\sin\omega_2 t\cos\omega_2 t) + 0{\cdot}89(E'_R - E'_Y)\cos^2\omega_2 t + E_S \sin\omega_1 t\cos\omega_2 t] \quad \ldots\ldots(5)$$

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Applying standard geometrical transforms the output after filtering out the high-frequency components is

$$(R-Y) = \frac{A}{2}[0{\cdot}89(E'_R - E'_Y) + E_S \cos \Delta\omega t]$$

which upon adjusting $A$ to give the required amplitude of the $(R-Y)$ signal gives

$$(R-Y) = (E'_R - E'_Y) + 1{\cdot}14 E_S \sin \Delta\omega t \qquad \ldots\ldots(6)$$

Similarly, multiplying eqn. (4) by $B \sin \omega_2 t$, filtering and adjusting $B$ to give the desired $(B-Y)$ component output gives:

$$(B-Y) = (E'_B - E'_Y) + 2{\cdot}03 E_S \cos \Delta\omega t \qquad \ldots\ldots(7)$$

The required $(G-Y)$ signal (obtained usually by matrixing and inversion) is given by

$$E'_G - E'_Y = -0{\cdot}51(E'_R - E'_Y) - 0{\cdot}19(E'_B - E'_Y) \quad \ldots(8)$$

Hence the $(G-Y)$ channel output is

$$(G-Y) = (E'_G - E'_Y) - 0{\cdot}58 E_S \sin \Delta\omega t - 0{\cdot}385 E_S \cos \Delta\omega t \qquad \ldots\ldots(9)$$

Hence after matrixing with $E'_Y$, either between grids and cathodes of the display guns or earlier, the effective signals applied to the guns are

$$\left.\begin{aligned} [R] &= E'_R + 1{\cdot}14 E_S \sin \Delta\omega t + E_S \sin \omega_1 t \\ [G] &= E'_G - 0{\cdot}58 E_S \sin \Delta\omega t - 0{\cdot}385 E_S \cos \Delta\omega t + \\ &\qquad + E_S \sin \omega_1 t \\ [B] &= E'_B + 2{\cdot}03 E_S \cos \Delta\omega t + E_S \sin \omega_1 t \end{aligned}\right\} \quad (10)$$

Thus in addition to the required signals there is a low-frequency beat $\Delta\omega t$ which will cause a colour beat in the picture. The $E_S \sin \omega_1 t$ term, which is a high-frequency term and the same in each channel, will not produce any colour beat but may give rise to dot patterning and luminance errors. It is usually removed by the subcarrier filtering in the luminance amplifier and may be ignored for the remainder of this section.

Hence, assuming a display tube of gamma 2, set to give illuminant 'C' white, the light outputs in terms of trichromatic units are given by

$$\left.\begin{aligned} [R] &= (E'_R + 1{\cdot}14 E_S \sin \Delta\omega t)^2 \\ [G] &= (E'_G - 0{\cdot}58 E_S \sin \Delta\omega t - 0{\cdot}385 E_S \cos \Delta\omega t)^2 \\ [B] &= (E'_B + 2{\cdot}03 E_S \cos \Delta\omega t)^2 \end{aligned}\right\} \quad (11)$$

where these expressions are defined *only* when the quantities inside the brackets are greater than or equal to zero. For negative quantities the guns are cut-off and the light output remains zero.

It will be apparent that the expansion of eqn. (11) will give rise to spurious terms in $\Delta\omega t$ and $2\Delta\omega t$.

Now if $E_S \sin \omega_1 t$ is in fact a vertical pattern in the picture, $f_1 = \omega_1/2\pi$ is an even multiple of half-line frequency whilst $f_2 = \omega_2/2\pi$ is chosen to be an odd multiple of half-line frequency so that $\Delta f = \Delta\omega/2\pi$ is also an odd multiple of half-line frequency. This means that if at a particular point in the image on one picture the colour beat has a phase $\phi$, the beat at the same point on the next picture (i.e. two fields or 1/25 second later) has a phase $(\phi + 180°)$ and therefore a different colour. The net effect is that the $\Delta\omega t$ terms will tend to be cancelled due to subjective colour fusion and the colour beat pattern seen by the observer will be largely due to the $2\Delta\omega t$ terms since these do not give cancellation because $2\Delta f$ is an even multiple of half-line frequency. However luminance errors due to the $\Delta\omega t$ terms may be large enough to be seen as brightness flicker from picture to picture—i.e. at $12\frac{1}{2}$ c/s.

To simplify calculation even further for the purposes of comparison it will now be assumed that

$$E'_R = E'_G = E'_B = 1$$

so that eqn. (11) becomes

$$\left.\begin{aligned} [R]_1 &= (1 + 1{\cdot}14 E_S \sin \Delta\omega t)^2 \\ [G]_1 &= (1 - 0{\cdot}58 E_S \sin \Delta\omega t - 0{\cdot}385 E_S \cos \Delta\omega t)^2 \\ [B]_1 &= (1 + 2{\cdot}03 \cos \Delta\omega t)^2 \end{aligned}\right\} \quad (12)$$

These expressions have been evaluated for various values of $\Delta\omega t$ and $E_S = 1$ and $E_S = \frac{1}{2}$. From this the resulting colour beat locus for a single field may be plotted as in Fig. 1, which also shows the corresponding luminance at 30 deg intervals of $\Delta\omega t$.

The resultant stationary high-visibility colour beat pattern over two consecutive pictures is shown in Fig. 2, together with the percentage fluctuation in luminance on successive pictures. The cycle rate round the locus

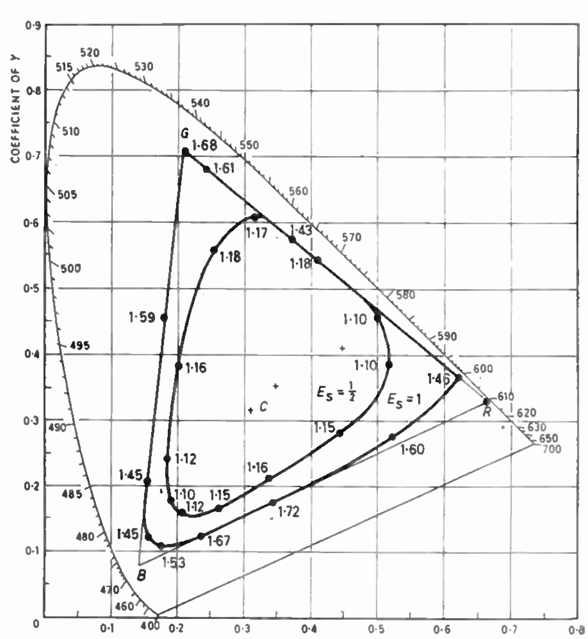



**Fig. 1.** Type I—Single-picture colour beat locus.

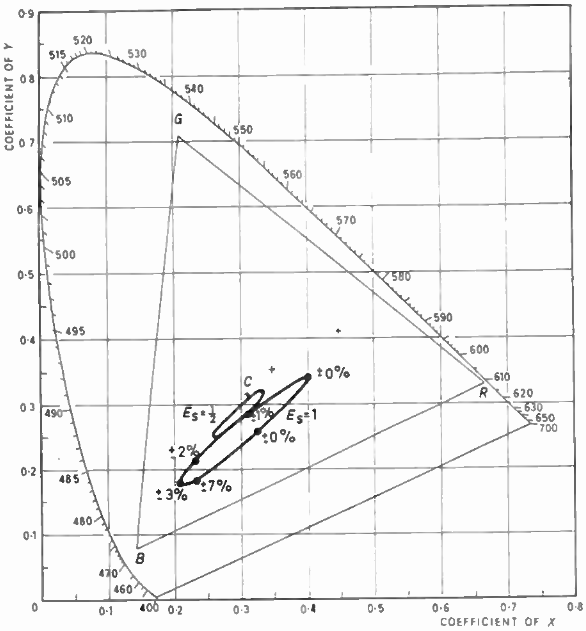



Fig. 2. Type 1—High-visibility colour beat locus and percentage twinkle.

is of course $2\Delta\omega t$, giving a low-frequency colour beat pattern of resolution frequency $2\Delta f$ in the picture. From Fig. 2 it will be apparent that even with $E_S = 1$ the resultant observed colour beat gamut is not large and there is negligible luminance twinkle. This agrees with observation of shadow-mask tube N.T.S.C. receivers. On monochrome transmissions, of course, the chrominance circuits are rendered inoperative and there is no cross-colour or twinkle at all.

For convenience the cross-colour and twinkle described by Figs. 1 and 2 will be termed Type 1.

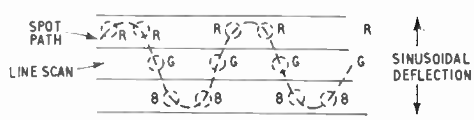
## 3. Effects with Single-gun Display Receivers

### 3.1 *Operation of Single-gun Receivers*

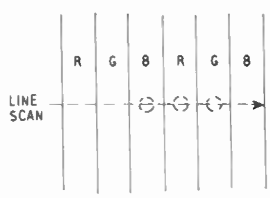
In single-gun displays the primary phosphors are usually arranged in a regular pattern of stripes and the beam current is shared in time sequence between them. Although it is possible to decode the N.T.S.C. composite video signal to red, green and blue and then to 'gate' these to the single gun in synchronism with the passage of the beam over the phosphor stripes, it is usual to drive the tube with a suitable form of signal so that the colour selection occurring at the screen performs the decoding directly.[5] Figures 3(*a*) and 3(*b*) show two possible configurations. In the first the display has a horizontal stripe structure and as the beam is deflected from left to right by the line scan it is sinusoidally deflected up and down giving a colour selection sequence of the form R R G B B G R R G B B ... known as a 'reversing colour sequence' (r.c.s.). Typical of this type of tube are the Chromatron[6,7] and the 'Banana Tube'[8] in which the r.c.s. colour selection may be conveniently made to take place at the constant subcarrier frequency. In the configuration of Fig. 3(*b*) the colour selection takes place automatically as the beam is deflected in the line direction, the sequence in this case being of the type R G B R G B... known as 'continuous colour sequence' (c.c.s.). Typical of this is the beam-indexing tube[9] in which the colour selection frequency is not absolutely constant, but varies with among other things scan linearity. However the selection process is stationary with respect to detail in the picture, i.e. a particular point in the image is always sampled in the same colour, whereas with operation of the chromatron at subcarrier frequency sampling interlace occurs.

Reversing colour sequence displays may be conveniently converted into c.c.s. displays by third harmonic pulsing or by blanking of one of the double frequency excitations of the centre stripe and the form of signal required for c.c.s. operation is not so very different from the N.T.S.C. signal. It can be derived from the latter by means of the operations known as $Y$ to $M$ conversion and elliptical amplification of the subcarrier chrominance signal with translation to the colour selection frequency if different from the subcarrier. The translation may be obtained by heterodyning or demodulation followed by remodulation ('demod-remod'). It may be shown that there is no difference between the results of the two methods.

In theory operation of the single-gun tube may depend upon the assumption of infinitesimal sampling angles and possibly spot size also. In practice the colour errors occurring due to finite angles and spot size cannot be large or the display is useless, so that no great error will arise due to making the same assumptions for the present calculation.




(*a*) Reversing colour sequence (r.c.s.).




(*b*) Continuous colour sequence (c.c.s.).

Fig. 3. Basic forms of sequential colour selection.

### 3.2. *C.C.S. Operation of Single-gun Displays of the First Kind*

#### 3.2.1. Derivation of output signal

Let us consider first of all tubes in which the colour selection rate is constant: typical of these is the chromatron. The most usual form of operation is c.c.s. and the signal then required for the tube may be shown to be

$$E'_S = \tfrac{1}{3}(E'_R + E'_G + E'_B) + \tfrac{2}{3}E'_R \sin \omega_3 t + \\ + \tfrac{2}{3}E'_G \sin\left(\omega_3 t - \frac{2\pi}{3}\right) + \tfrac{2}{3}E'_B \sin\left(\omega_3 t - \frac{4\pi}{3}\right) \quad (13)$$

where $\omega_3$ is the colour selection rate and sampling at the screen occurs at 90, 210 and 330 deg.

Hence if the received N.T.S.C. signal is given by eqn. (4) and 'demod-remod' is employed with the intention of obtaining eqn. (13) above the actual beam current function to be sampled by the screen structure may be shown to be given, in general, by

$$E_S = \Big[\tfrac{1}{3}(E'_R + E'_G + E'_B) + \\ + 0{\cdot}56E_S \sin \Delta\omega t + 1{\cdot}65E_S \cos \Delta\omega t + \\ + \tfrac{2}{3}(E'_R + 1{\cdot}14E_S \sin \Delta\omega t)\sin \omega_3 t + \\ + \tfrac{2}{3}(E'_G - 0{\cdot}58E_S \sin \Delta\omega t - \\ - 0{\cdot}385E_S \cos \Delta\omega t)\sin\left(\omega_3 t - \frac{2\pi}{3}\right) + \\ + \tfrac{2}{3}(E'_B + 2{\cdot}03E_S \cos \Delta\omega t)\sin\left(\omega_3 t - \frac{4\pi}{3}\right) + \\ + E_S \sin \omega_1 t\Big]^2 \quad \ldots(14)$$

Now this expression is sampled at 90, 210 and 330 deg at $\omega_3 t$ and two cases have to be considered.

(*a*) When $\omega_3 \gg \omega_2$: Here the colour selection takes place at a frequency much higher than subcarrier, i.e. out of band, and the $E_S \sin \omega_1 t$ term may be ignored, since if necessary it may be eliminated from the luminance signal before addition to the new chrominance subcarrier, giving no cross-colour or twinkle on monochrome transmissions. In the case of a colour transmission the result of sampling gives

$$[R] = (E'_R + 1{\cdot}14E_S \sin \Delta\omega t)^2$$
$$[G] = (E'_G - 0{\cdot}58E_S \sin \Delta\omega t - 0{\cdot}385E_S \cos \Delta\omega t)^2$$
$$[B] = (E'_B + 2{\cdot}03E_S \cos \Delta\omega t)^2$$

i.e. the same cross-colour and twinkle as for the three-gun receiver (Type I).

(*b*) When $\omega_3 \simeq \omega_2$: The most typical case of this is when $\omega_3 = \omega_2$ (i.e. when the colour selection takes place at the subcarrier frequency). The $E_S \sin \omega_1 t$ term cannot be ignored then as it cannot be removed easily from the signal. Sampling at 90, 210 and 330 deg as before with $E'_R = E'_G = E'_B = 1$, the outputs are

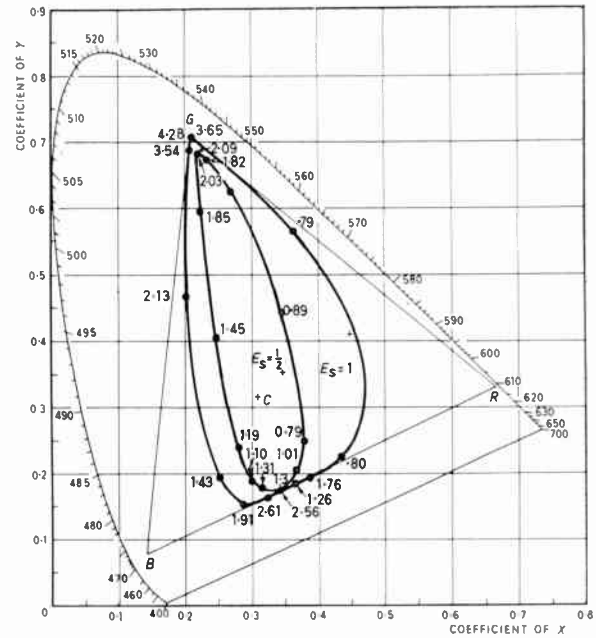


**Fig. 4.** Type II—Single-picture colour beat locus.

now found to be

$$\left.\begin{aligned}[R] &= (1 + 1{\cdot}14E_S \sin \Delta\omega t + E_S \cos \Delta\omega t)^2 \\ [G] &= (1 - 1{\cdot}446E_S \sin \Delta\omega t - 0{\cdot}885E_S \cos \Delta\omega t)^2 \\ [B] &= (1 + 0{\cdot}866E_S \sin \Delta\omega t + 1{\cdot}53E_S \cos \Delta\omega t)^2\end{aligned}\right\} \quad (15)$$

The single picture colour beat loci and luminances for $E_S = 1$ and $\frac{1}{2}$ are shown plotted in Fig. 4, whilst the high visibility colour beat loci and luminance twinkle are shown in Fig. 5. This type of beat will be

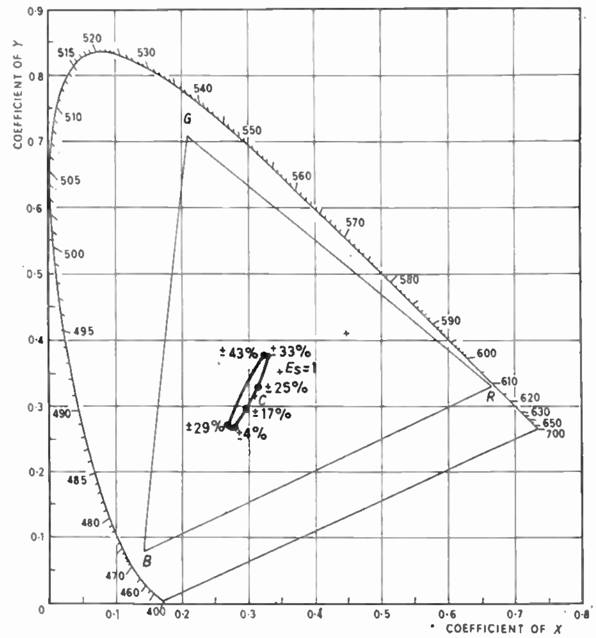


**Fig. 5.** Type II—High-visibility colour beat locus and percentage twinkle.

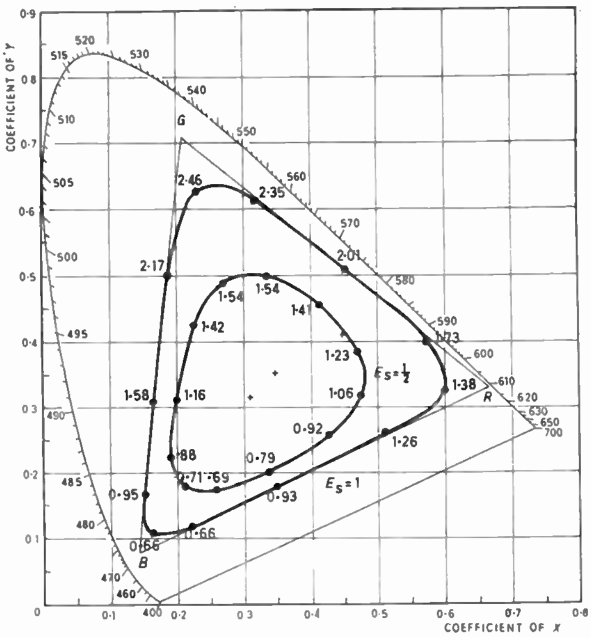



Fig. 6. Type III—Single-picture colour beat locus.

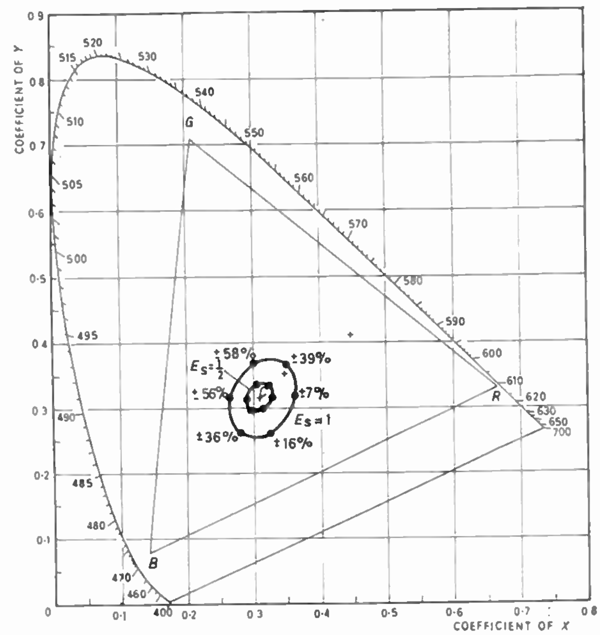



Fig. 7. Type III—High-visibility colour beat locus and percentage twinkle.

termed Type II and it will be immediately apparent that although the high-visibility colour beat gamut is not large there is severe luminance twinkle (up to $\pm 43\%$).

Thus it is clear that the twinkling that has been observed arises as a result of synchronous detection at the screen at subcarrier frequency of luminance detail and h.f. video noise; it is very much worse for the c.c.s.-operated single-gun tube because the required form of drive signal differs somewhat from the N.T.S.C. signal and the demodulation conditions are different.

### 3.2.2. Reverse compatibility (monochrome transmissions)

It can be seen moreover that with $\omega_3 = \omega_2$ and a monochrome transmission, the $E_S \sin \omega_1 t$ component will still be synchronously detected at the screen. In this case the light outputs for $E'_R = E'_G = E'_B = 1$ are

$$\left.\begin{aligned}
[R] &= (1 + E_S \sin \omega_1 t)^2\big|_{\omega_2 t = 90^\circ} \\
&= (1 + E_S \cos \Delta\omega t)^2 \\
[G] &= \left(1 - \frac{E_S}{2}\cos \Delta\omega t - E_S \frac{\sqrt{3}}{2} \sin \Delta\omega t\right)^2 \\
[B] &= \left(1 - \frac{E_S}{2}\cos \Delta\omega t + E_S \frac{\sqrt{3}}{2} \sin \Delta\omega t\right)^2
\end{aligned}\right\} \quad (16)$$

The single-picture and high-visibility loci are shown in Figs. 6 and 7 and from the latter it will be seen that there is still severe twinkle (up to $\pm 58\%$). This effect will be termed Type III.

In the case when $\omega_3 \gg \omega_2$ it might be supposed that the cross-colour and twinkle are the same as for a three-gun display, but in fact luminance components near to frequencies which are a simple subharmonic of the colour-selection frequency can give rise to trouble. In particular the worst effect is likely to occur when $\omega_2 = \frac{1}{2}\omega_3$. The single picture colour

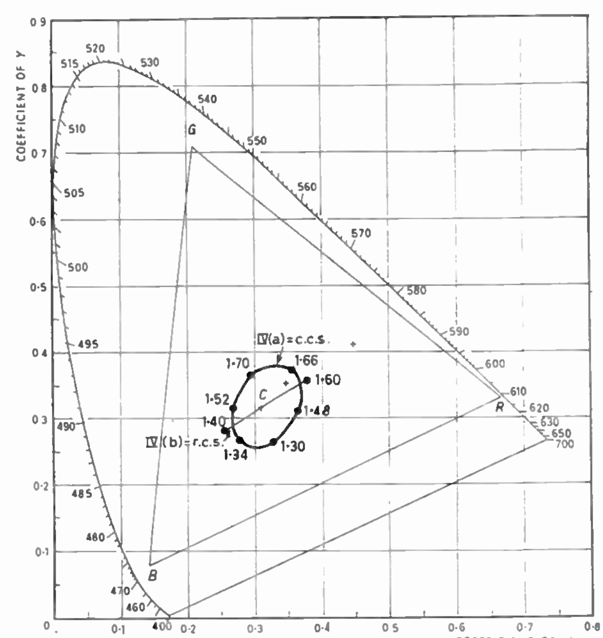



Fig. 8. Type IV—Single-picture colour beat loci ($E_S = 1$, $\omega_2 = \omega_3/2$).



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beat locus on monochrome transmission for the case when $E_S = 1$ is shown in Fig. 8 (Type IV($a$)). (Also shown is the beat locus for r.c.s. operation (Type IV($b$)).) Since $\omega_3 = 2\omega_2$ the colour selection process will be the inverse of dot-interlaced (a situation which will be termed 'anti-dot-interlaced') and the beats will appear as successive stationary vertical bars having colours at specific points on the locus, with a repetition rate in the image corresponding to $f_3 - 2f_1$. When $f_1 = \frac{1}{2}f_3$ the beat pattern will be of a single constant colour.

For a dot-interlaced system the high visibility colour beat over two pictures will vanish but there will be $\pm 13\%$ twinkle for $E_S = 1$. In both cases with $\omega_3 \gg \omega_2$ the effects can be eliminated by inserting a subcarrier notch filter in the luminance channel, in which case the performance becomes the same as for the three-gun receiver.

3.2.3. Summary of c.c.s. operation of first kind of single-gun display

Although the cross-colour and twinkle have been evaluated for c.c.s. decoding on the tube it can be readily shown that the same effects are obtained with c.c.s gating of the output from an R G B decoder—though this mode of operation is not in fact a very practicable one.

The main conclusion which can be drawn is that with c.c.s. operation severe twinkle occurs on both colour and monochrome transmission when the colour selection is performed at the subcarrier frequency because it is not conveniently possible to insert a notch in the luminance channel. It would therefore seem

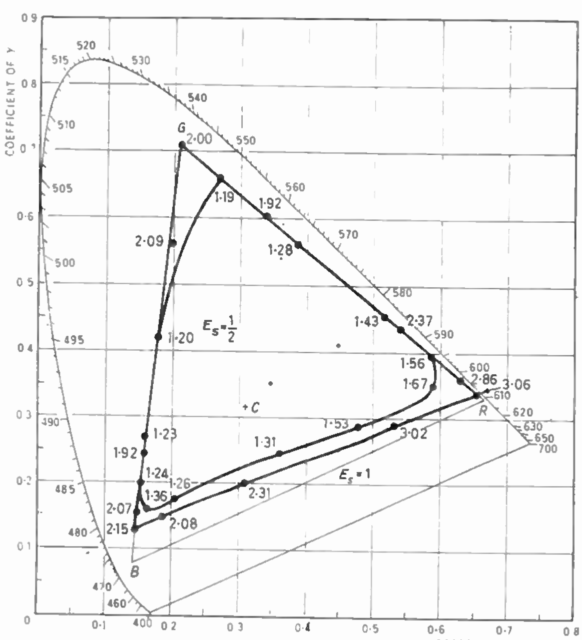


Fig. 9. Type V—Single-picture colour beat locus.

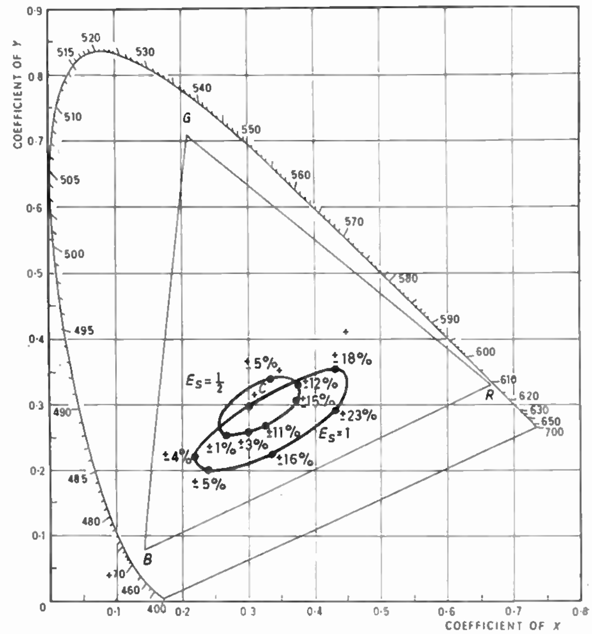


Fig. 10. Type V—High-visibility colour beat locus and percentage twinkle.

necessary to perform the colour selection at a much higher frequency. Twice the subcarrier frequency is a convenient choice and the additional colour-beats which occur due to loss of dot-interlace can be removed by a subcarrier luminance notch, leaving the same beats as for a three-gun receiver. However, for tubes such as the chromatron, the use of twice subcarrier colour selection rate, whilst rendering the dot structure invisible, is probably precluded on the grounds of switching power requirements.

### 3.3. *R.C.S. Operation of Single-gun Tubes of the First Kind*

It will be apparent that the situation with r.c.s. operation will be similar to that described for c.c.s., and as r.c.s. operation is not a preferred method it will not be considered in detail. However it will be evident that if the operating conditions permit subcarrier luminance notching, the effects will again be the same as for a three-gun receiver. Where this is not possible, e.g. when $\omega_3 = \omega_2$ the two cases of monochrome and colour transmission need to be considered.

3.3.1. Colour transmissions with $\omega_3 = \omega_2$

It will be supposed that centre green-stripe operation is employed with sampling at 0 deg for red, 90 deg and 270 deg for green and 180 deg for blue. Then

$$[R] = (E'_R + 1{\cdot}14E_S \sin \Delta\omega t + E_S \sin \omega_1 t)^2\big|_{\omega_2 t = 0^\circ}$$

$$[G] = \tfrac{1}{2}[(E'_G - 0{\cdot}58E_S \sin \Delta\omega t - 0{\cdot}385E_S \cos \Delta\omega t + E_S \sin \omega_1 t)^2\big|_{\omega_2 t = 90^\circ} + (E'_G - 0{\cdot}58E_S \sin \Delta\omega t - 0{\cdot}385E_S \cos \Delta\omega t + E_S \sin \omega_1 t)^2\big|_{\omega_2 t = 270^\circ}]$$

$$[B] = (E'_B + 2{\cdot}03E_S \sin\Delta\omega t + E_S \sin\omega_1 t)^2\big|_{\omega_2 t = 180°}$$

Setting $E'_R = E'_G = E'_B = 1$ these reduce to

$$\left.\begin{aligned}
[R] &= (1+2{\cdot}14E_S\sin\Delta\omega t)^2\\
[G] &= \tfrac{1}{2}[(1-0{\cdot}58E_S\sin\Delta\omega t +\\
&\quad +0{\cdot}615E_S\cos\Delta\omega t)^2 +\\
&\quad +(1-0{\cdot}58E_S\sin\Delta\omega t - 1{\cdot}615E_S\cos\Delta\omega t)^2]\\
[B] &= (1+2{\cdot}03E_S\cos\Delta\omega t - E_S\sin\Delta\omega t)^2
\end{aligned}\right\} \quad (17)$$

The single picture and high-visibility colour beats and twinkle are respectively shown in Figs. 9 and 10 (Type V). The twinkle for $E_S = 1$ is seen to be significant ($\pm 23\%$ max.).

3.3.2. Monochrome transmissions with $\omega_3 = \omega_2$

In this case the outputs may be shown to be

$$\left.\begin{aligned}
[R] &= (1+E_S\sin\Delta\omega t)^2\\
[G] &= \tfrac{1}{2}[(1+E_S\cos\Delta\omega t)^2 + (1+E_S\sin\Delta\omega t)^2]\\
[B] &= (1+E_S\cos\Delta\omega t)^2
\end{aligned}\right\} \quad (18)$$

Figures 11 and 12 show the resultant beats, Type VI, and it is seen that significant twinkle again occurs.

3.3.3. Summary of r.c.s. operation of first kind of single-gun display

It will be evident from the above that r.c.s. operation leads to levels of cross-colour and twinkle comparable with those for c.c.s. operation and the twinkle in both cases can only be eliminated by colour selection at a frequency of the order of twice the subcarrier, possibly together with a subcarrier notch in the luminance channel. As for c.c.s. operation, there is no difference between r.c.s. decoding on the tube and r.c.s. gating of decoded R G B.

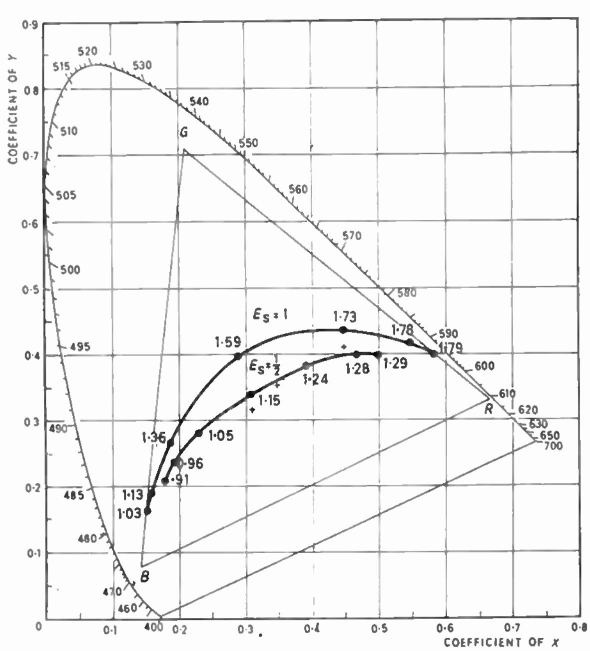


Fig. 11. Type VI—Single-picture colour beat locus.

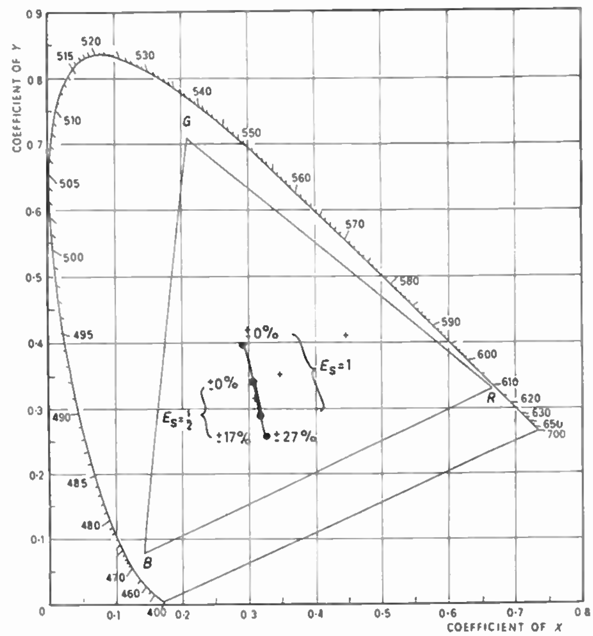


Fig. 12. Type VI—High-visibility colour beat locus and percentage twinkle.

It may be noted that in all cases of colour selection at a frequency in the video band (and therefore near subcarrier) the twinkle may be eliminated (in exchange for a severe single-picture colour beat, which no longer cancels over two pictures) by operating at an anti-dot-interlaced frequency. In practice such a complication would be very impracticable and will not be considered.

3.4. *Operation of Single-gun Tubes of the Second Kind (Beam Indexing Tubes)*

In the beam indexing type of display the colour selection is usually arranged to take place as a direct result of the line deflection of the spot across a vertical stripe structure.[9] For convenience c.c.s. operation is invariably employed. Due to variations in deflection-linearity etc., the colour selection rate is not constant and it is the task of the index circuits to ensure that the chrominance component of the signal applied to the gun is matched in frequency and phase to the stripe structure. The way in which this is performed need not concern us here, but it may be noted that the chrominance signal writing frequency required is typically of the order of twice the subcarrier frequency. Also, since the colour selection is related purely to the location of the phosphor stripes, it follows that it must bear a stationary relationship with respect to vertical detail in the picture. That is, a particular point in the picture will always correspond to a

**Table 1**

Comparison of cross-colour and twinkle for $E'_R = E'_G = E_B = 1$, and $E_S = 1$

| System | Colour Transmission | | | Monochrome Transmission | | |
|---|---|---|---|---|---|---|
| | Beat Type | Cross-Colour | Maximum Twinkle | Beat Type | Cross-Colour | Maximum Twinkle |
| Three-gun and single-gun h.f. colour selection (with luminance subcarrier notch) | I | Moderate | ± 7% | None | None | None |
| Single-gun c.c.s. at subcarrier (no notch possible) | II | Small | ± 43% | III | Small | ± 58% |
| Single-gun at approx. 2 × subcarrier with anti-dot-interlace (no luminance notch) | not determined | | | IV(*a*) c.c.s.<br>IV(*b*) r.c.s. | Small | None |
| Single-gun r.c.s. at subcarrier (no notch possible) | V | Moderate | ± 23% | VI | Small | ± 27% |

particular phosphor colour. Thus although the colour selection rate varies in frequency over the screen, at a fixed point it is fixed in relation to luminance detail.

It follows therefore that the cross-colour and twinkle will be the same as for a display of the first kind (for example a chromatron) where the colour selection takes place with $\omega_3 \gg \omega_2$ at an anti-dot-interlaced frequency. In this case some colour beat patterning can occur due to $E_S \sin \omega_1 t$ and the chrominance component at the subcarrier frequency in the luminance channel but there will be negligible twinkle. Colour beats can be substantially eliminated by means of a luminance subcarrier notch.

## 4. Conclusions

Analysis of a simple case has shown that the cross-colour and severe twinkle due to high-frequency luminance detail (and hence also due to noise) observed in single-gun colour displays arises almost entirely as a result of the use of the subcarrier frequency for colour selection under which conditions it is not possible to employ luminance channel notching.

It appears that in a single-gun receiver display these effects, particularly twinkle, can only be reduced to a small level by employing colour selection at a frequency of the order of twice subcarrier which preferably gives anti-dot-interlace, together with a subcarrier-frequency notch in the luminance channel response. These requirements seem to suit the beam-indexing type of display far more than the type of display employing constrained colour selection, such as the one-gun chromatron, for which colour selection at twice the subcarrier frequency may prove very uneconomical. The important cases are summarized in Table 1, from which it will be seen that under the best conditions the cross-colour and twinkle may be the same as for the three-gun receiver, for which the cross-colour is small and the twinkle negligible.

## 5. Acknowledgments

The analysis of the cross-colour in the case of the three-gun receiver is based upon that of Carnt and Townsend[10] to which due acknowledgment is made.

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

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*Manuscript first received by the Institution on 1st November 1963 and in final form on 15th January 1964. (Paper No. 894/T26.)*

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U.D.C. 621.375.121: 621.397.621.54

# The Design of a Linear-phase I.F. Amplifier for Colour Television Receivers

By

J. O. SCANLAN, B.E.†

**Summary:** The design of a band-pass amplifier with a linear phase response and suitable for colour television i.f. applications, is carried out. The linear phase response is obtained by designing according to a system of pole location. The results obtained from tests on the i.f. amplifier are given, and are found to be in good agreement with the theory.

## 1. Introduction

A system of pole location has previously been described[1] which leads to transfer functions having an approximately linear phase characteristic. The treatment was given in terms of normalized low-pass transfer functions. This system lends itself to the design of amplifiers and filters of low-pass or band-pass characteristics for any given frequency range, and the purpose of this paper is to give an illustrative example of the design of such a band-pass amplifier. The example chosen is a band-pass amplifier centred on $36\frac{1}{2}$ Mc/s and with a $4\frac{1}{2}$-Mc/s bandwidth to 6-dB points on the response curve, i.e. suitable for colour television i.f. amplifier applications.

## 2. Band-pass Amplifier Design

As an example of the design of band-pass amplifiers the following specification will be taken:

| | |
|---|---|
| Centre frequency | 36·5 Mc/s |
| 6 dB bandwidth | 4·5 Mc/s |
| Phase linear within the 6-dB bandwidth | |

Two outputs are required, one centred on 36·5 Mc/s with a 4·5-Mc/s 6-dB bandwidth, and the other centred on 35·1 Mc/s with a 1·8-Mc/s 3-dB bandwidth. An amplifier of this specification would be suitable as the i.f. amplifier in a colour television receiver for British 625-line standards.

In order to achieve the gain required (about 400 μV at the grid of the mixer for 2 V on the vision detector) three stages must be used in the amplifier. Therefore three band-pass interstage networks and a detector stage must be designed. In order to achieve the required phase linearity, it is advisable to make the detector stage wideband and to design the other three band-pass transformers in accordance with the scheme previously mentioned. The three band-pass circuits incorporate six poles. The location of six poles in the normalized low-pass case is shown in Fig. 1. In order to transfer to the normalized band-pass case the pole pattern is moved along the $j\omega$ (frequency) axis until it lies about the desired centre frequency. However, this process introduces a conjugate set of poles at negative frequencies and so the response curves (on a linear frequency scale) will not be the same as for the low-pass case. The situation is shown in Fig. 2.

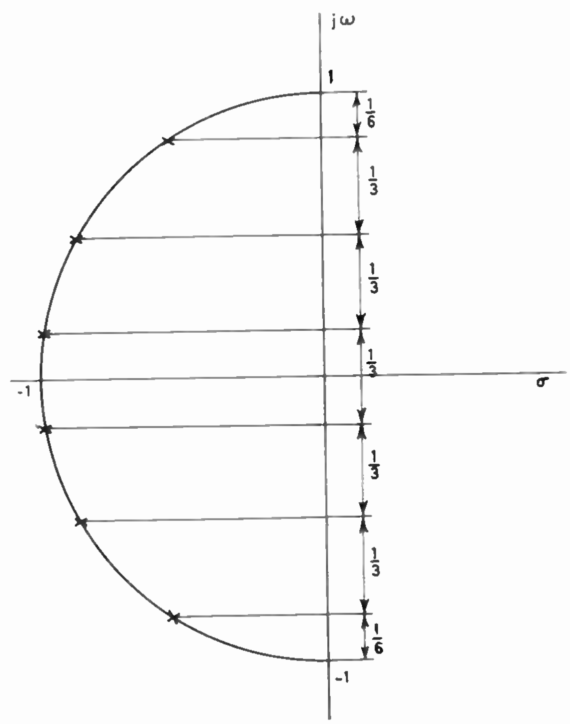


**Fig. 1.** Normalized pole locations for low-pass prototype.

For amplifiers of narrow relative bandwidth the conjugate poles have little effect on the band-pass curve and their effect may be neglected. This is the same approximation as is normally made by putting $\beta = 2\Delta\omega/\omega_0$ instead of $\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)$.

Once the effect of the conjugate poles is neglected, the response curve is symmetrical on a linear frequency scale and is the same shape as the low-pass

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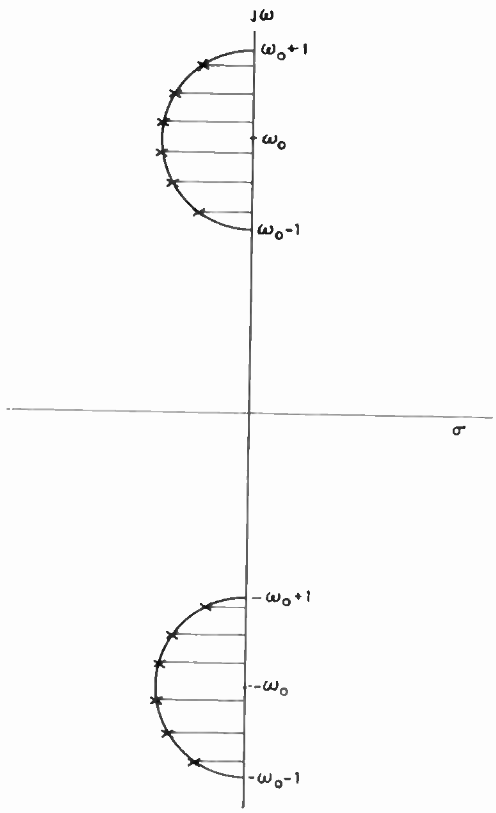


Fig. 2. Normalized pole locations for band-pass prototype.

curve. This enables the design for any given bandwidth to be accomplished by reference to the curves previously drawn for the low-pass case. Figures 3 and 4 show the normalized amplitude and group delay curves for the sixth-order function.

### 3. Band-pass Transformer Parameters

From Fig. 3, which applies to the case where the radius of the locating circle for the poles is 1 rad/s, it is seen that the 6-dB bandwidth in the band-pass case will be 1·5 rad/s. The 6-dB bandwidth required for the design is 4·5 Mc/s, hence a circle of radius 3 Mc/s must be used. The pole locations are then shown in Fig. 5. Thus if the amplifier were to be constructed using six single tuned interstage networks

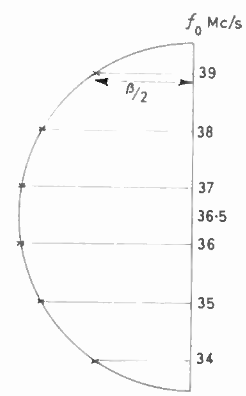


Fig. 5. Pole locations for i.f. amplifier design.

they would consist of three pairs of circuits:

(1) detuned $\pm\ 2\frac{1}{2}$ Mc/s with $B/2 = 1{\cdot}66$ Mc/s ($Q = 11$);

(2) detuned $\pm\ 1\frac{1}{2}$ Mc/s with $B/2 = 2{\cdot}6$ Mc/s ($Q = 7$);

(3) detuned $\pm\ \frac{1}{2}$ Mc/s with $B/2 = 2{\cdot}96$ Mc/s ($Q = 6$).

The detuning is with respect to 36·5 Mc/s and the half bandwidths are given by the corresponding intercepts on the locating circle in Fig. 5. The corresponding $Q$-factors for the three pairs of circuits are (approximately) as shown.

Knowing the detunings and the corresponding $Q$-factors the three band-pass transformers which produce approximately the same response as the six single tuned circuits may be calculated.[2] These transformers are all tuned to the same centre frequency of 36·5 Mc/s but have different couplings and damping. The values are calculated in Appendix 1 and are: (the notation is that of ref. 2)

| | | | |
|---|---|---|---|
| (1) | $k' = 0{\cdot}133$ | $Q_b = 11$ | $Q_p = 60$ |
| | $Q_s = 6$ | $R_s = 2{\cdot}2k$ | $k = 0{\cdot}152$ |
| (2) | $k' = 0{\cdot}08$ | $Q_b = 7$ | $Q_p = 60$ |
| | $Q_s = 3{\cdot}8$ | $R_s = 1{\cdot}2k$ | $k = 0{\cdot}148$ |
| (3) | $k' = 0{\cdot}027$ | $Q_b = 6$ | $Q_p = 60$ |
| | $Q_s = 3{\cdot}1$ | $R_s = 1k$ | $k = 0{\cdot}155$ |

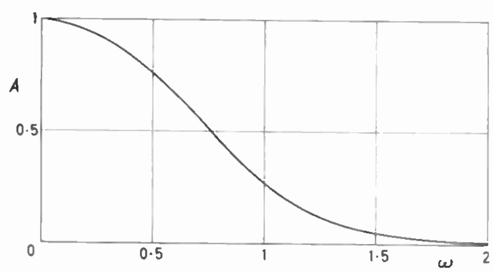


Fig. 3. Amplitude response for normalized 6th-order prototype.

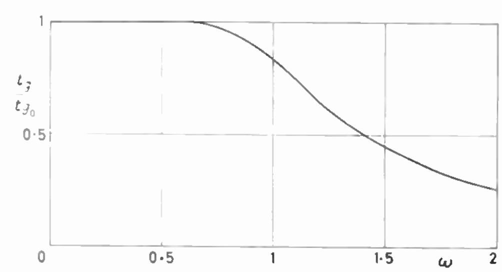


Fig. 4. Phase response for normalized 6th-order prototype.

When these parameters are known the tuned circuits are fully specified, since maximum gain occurs for no added capacitance on either primary or secondary sides and the coils are wound to resonate with the input and output capacitances of the valves respectively.

The alignment of these circuits requires the setting of the correct coupling coefficient $k$ once the appropriate $Q$-factor and centre frequency are set. This can be done by adjusting the coupling to set each individual circuit to a required bandwidth which can be calculated from eqn. (12) of ref. 2 when the $Q$-factors are known. The results of this calculation are:

(1) $B = 7{\cdot}7$ Mc/s
(2) $B = 4{\cdot}7$ Mc/s
(3) $B = 4$ Mc/s

The bandwidths are the 3-dB bandwidths with respect to centre-frequency.

This arrangement will then have a band-pass shape as given in Fig. 3, with a group delay curve of the shape given in Fig. 4. The frequency scale in each case is such that $\omega = 1$ corresponds to 3 Mc/s off centre-frequency.

In order to fully specify the performance, it is necessary to know the centre-frequency gain and group delay. The centre-frequency gains may be calculated from eqn. (8) of ref. 2. They are:

(1) $G_0 = 26$ dB
(2) $G_0 = 30$ dB
(3) $G_0 = 24$ dB

This gives an overall loss in gain of about 3 dB as compared with a system using transitionally-coupled band-pass filters.

The centre-frequency group delay and the group delay variation are calculated in Appendix 2 with the following results for a single pair of band-pass circuits:

$$\frac{t_g}{t_{g0}} = \frac{(1+Q_b^2k'^2)(1+Q_b^2k'^2+Q_b^2\beta^2)}{\{Q_b^2k'^2+1\}^2+2(Q_b\beta)^2\{(Q_bk')^2-1\}+(Q_b\beta)^4}$$

and the centre-frequency group delay $t_{g0}$ is

$$t_{g0} = \frac{4Q_b}{\omega_0}\,\frac{1}{1+(Q_bk')^2}$$

The overall centre-frequency group delay is the sum of the centre-frequency group delays of the individual band-pass circuits. The first equation is valid only for narrow relative bandwidths. Evaluating the centre-frequency group delay for each circuit gives

(1) $t_{g0} = 77$ ns
(2) $t_{g0} = 77$ ns
(3) $t_{g0} = 91$ ns

Total centre-frequency group delay = 245 ns. Thus the group delay curve of Fig. 4 may be labelled in nanoseconds. From this curve it may be seen that at the 6-dB points the group delay variation is greatest and amounts to approximately 5 ns.

### 4. Detectors and Sound Traps

In order that the amplifier as described above may be suitable for use as the i.f. amplifier in a colour television receiver it is necessary to provide outputs for luminance, chrominance and sound signals. This is best done using two detectors, one for luminance and the other for chrominance and sound. This is achieved by adding a wide-band double-tuned detector stage to the amplifier as previously described.

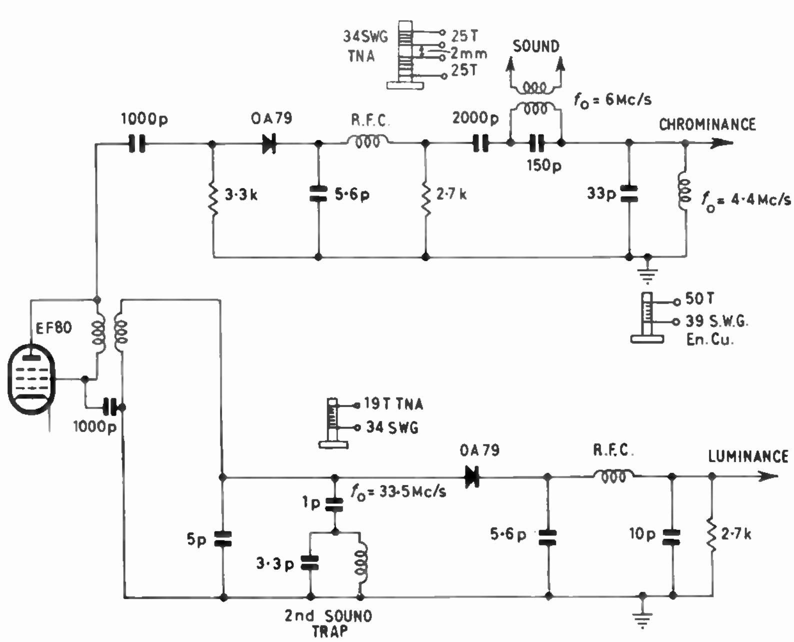


Fig. 6. Detector circuits.



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The luminance output is taken from the secondary side and the chrominance and sound outputs are taken from the primary side.

The coupling of the band-pass transformer is adjusted so that the secondary response is flat, resulting in a double-humped primary response, one of the humps coinciding with the pass-band of the chrominance signal. Two tuned circuits are provided in the chrominance detector, one for the sound output tuned to 6 Mc/s and one for the chrominance output, tuned to 4·4 Mc/s. The capacitance associated with the chrominance tuned circuit is the input capacitance of the following valve (about 30 pF) and the damping resistor is chosen to give the required bandwidth (1·8 Mc/s) for the chrominance signal.

The resistance and capacitance of the luminance detector are chosen to give the required video bandwidth for this signal (5 Mc/s). The value of 2·7 kΩ for the resistance is a suitable characteristic impedance for the delay cable which must be attached to the luminance detector to equalize the delay of luminance and chrominance signals. Figure 6 shows the circuits of these detectors and Figs. 7 and 8 show the measured responses at the two outputs.

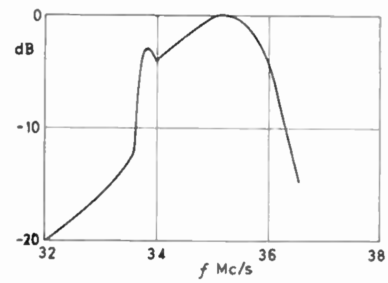
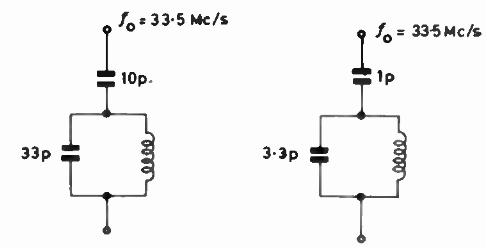



**Fig. 8.** Chrominance detector response.




(*a*) 1st sound trap. (*b*) 2nd sound trap.

**Fig. 9.**

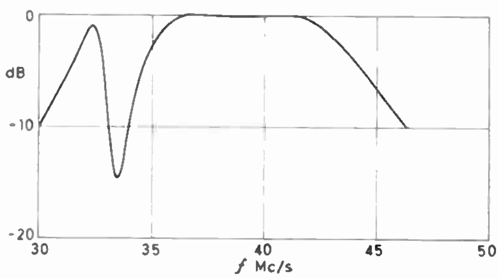



**Fig. 7.** Luminance detector response.

It is necessary to provide two trap circuits in the amplifier in order to achieve sufficient rejection at the sound carrier frequency in the luminance and chrominance outputs. One trap is placed in the coupling loop between the tuner output and the first i.f. valve, and the other is placed across the secondary winding of the detector tuned circuit. Figures 9(*a*) and (*b*) show the circuits of these traps. Figure 10 shows the complete circuit of the amplifier.

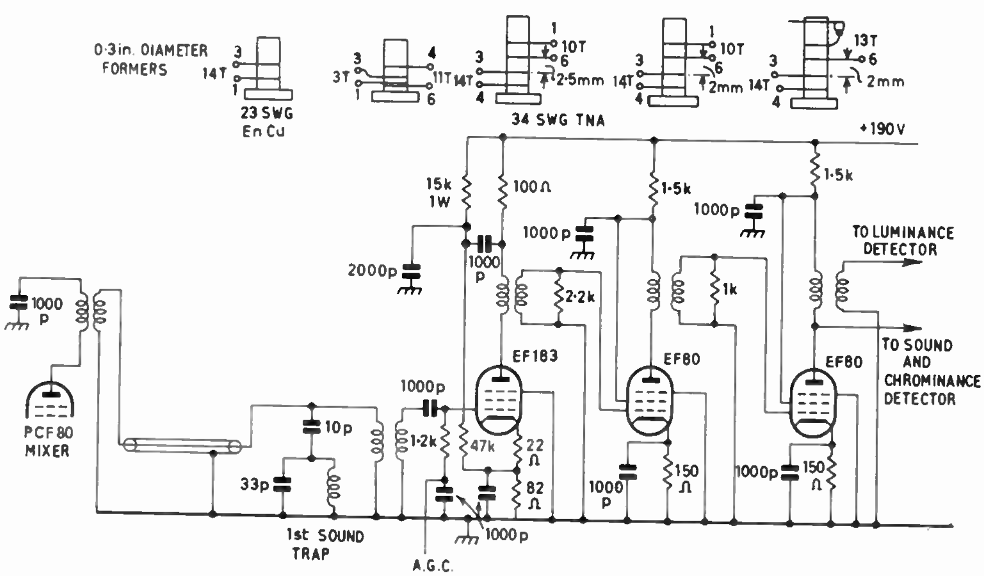



**Fig. 10.**
Complete i.f. amplifier circuit.

## 5. Measured Performance

An amplifier was constructed according to the circuit of Fig. 10 and tested for luminance and chrominance output, and for group delay response. The group delay showed less than 25 nanoseconds variation between 35 and 40 Mc/s, as measured on the Philips group delay meter type GM 2894, using a swept-frequency signal source.

However, the amplifier without its detector stage exhibited no detectable group delay variation between the same frequencies when measured by the same method, which showed very good agreement with the theory. Therefore the 25 ns variation for the complete amplifier was contributed mainly by the detector stage.

As a further check on the phase response, the phase delay of the modulation on a carrier at the vision frequency (39·5 Mc/s) between the input to a crystal mixer and the detector luminance output was measured. The apparatus used for this measurement was the Advance precision phase detector type 205A. This measurement showed that the phase delay varied by less than 30 ns for modulating frequencies up to 5 Mc/s. The latter measurement gives the correct phase characteristics of the amplifier since it takes account of both the vestigial nature of the signal for which the amplifier is intended, and the correct functioning of the detector which requires a fixed carrier frequency.[3]

The amplitude response of the luminance channel is shown in Fig. 11 and the detected output in Fig. 12.

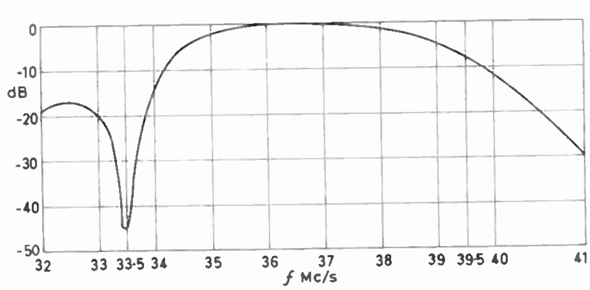



**Fig. 11.** Luminance response.

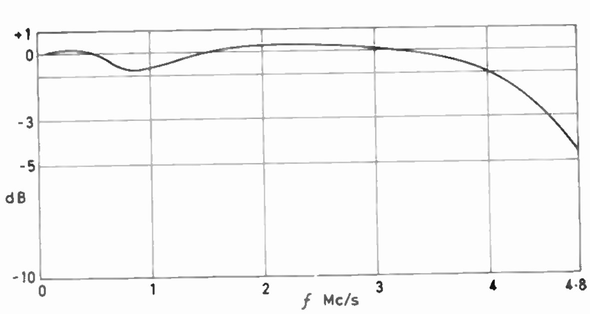



**Fig. 12.** Luminance output.

Figure 13 shows the detected output of the chrominance channel. The group delay in this channel had less than 50 ns variation between 3·5 and 5·4 Mc/s, when measured using both the methods previously described.

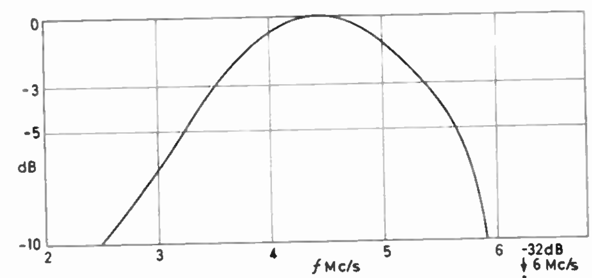



**Fig. 13.** Chrominance output.

These results are in good agreement with the theory, taking account of feedback in the valves and the influence of the detector, particularly on the phase response, each of which have been neglected in the design.

A sensitivity measurement showed that 360 μV at vision carrier frequency on the grid of the tuner mixer valve gave 2 V d.c. on the luminance detector, while 500 μV at chrominance carrier gave 2 V d.c. on the chrominance output.

## 6. Conclusions

An i.f. amplifier having a linear phase response and suitable for colour television applications has been designed, and the practical results obtained are given. These results are found to be in good agreement with the theoretical design.

## 7. References

1. J. O. Scanlan, "A class of transfer functions having approximately maximally flat group-delay characteristics", *Trans. Inst. Elect. Electronics Engrs (Circuit Theory)*, **CT-10**, No. 1, p. 122, March 1963.
2. J. S. Singleton, "Stagger-tuned and bandpass-coupled circuits in wideband i.f. valve amplifiers", *Mullard Tech. Commun.*, **5**, No. 42, pp. 48–59, February 1960.
3. A. van Weel, "Design of detector stages for signals with symmetrical or asymmetrical sidebands", *J.Brit.I.R.E.*, **18**, pp. 525–38, September 1958.

## 8. Appendix 1:

### Calculation of Tuned Circuit Parameters

From reference 2 (p. 53)

$Q_b$ = $Q$ of corresponding pair of single tuned circuits

$$k' = \{k^2 - \tfrac{1}{4}(1/Q_s - 1/Q_p)^2\}^{\frac{1}{2}}$$

$$= \frac{\Delta\omega}{\omega_0}$$



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$\Delta\omega$ = frequency separation between corresponding pairs of single tuned circuits

$\omega_0$ = centre-frequency of overall response.

Hence the following values for $k'$ result:

(1) $k' = \dfrac{5}{36 \cdot 5} = 0 \cdot 133$

(2) $k' = \dfrac{3}{36 \cdot 5} = 0 \cdot 08$

(3) $k' = \dfrac{1}{36 \cdot 5} = 0 \cdot 027$

It will be assumed that for maximum gain all added damping in each band-pass pair is concentrated on the secondary side, i.e. across the grid of the corresponding valve. Hence the $Q$ of the primary circuit will be that due to the output damping of the valve and losses of the coil, together with the output capacitance of the valve and stray capacitance. This results in a primary $Q$-factor $Q_p$ of about 60.

The various secondary $Q$-factors are calculated from the formula[2]

$$Q_b = \frac{2Q_p Q_s}{Q_p + Q_s}$$

This gives:

(1) $Q_s = 6$

(2) $Q_s = 3 \cdot 8$

(3) $Q_s = 3 \cdot 1$

To calculate the damping required on the various secondary circuits one must know the input capacitance of the valve. The first band-pass transformer feeds an EF183 valve with an input capacitance of 11 pF while the other two transformers feed EF80 valves with an input capacitance of 15 pF. This leads to the following damping resistances, $R_s$, on the secondary circuits:

(1) $R_s = 2 \cdot 2$ kΩ

(2) $R_s = 1 \cdot 2$ kΩ

(3) $R_s = 1$ kΩ

The coefficients of coupling, $k$, for the various circuits are calculated from the formula of ref. 2.

$$k' = \left\{k^2 - \frac{1}{4}\left(\frac{1}{Q_s} - \frac{1}{Q_p}\right)\right\}^{\frac{1}{2}}$$

This results in

(1) $k = 0 \cdot 152$

(2) $k = 0 \cdot 148$

(3) $k = 0 \cdot 155$

## 9. Appendix 2:

### Group Delay of Double-tuned Band-pass Transformers

We start from the following equation in ref. 2 (p. 52):

$$\frac{G}{G_0} = \frac{\delta_b^2 + k'^2}{(b - \mathrm{j}\delta_b)^2 - k'^2}$$

The phase angle of $\dfrac{G}{G_0}$, $\varphi$ is

$$\varphi = \tan^{-1} \frac{2\beta\delta_b}{\beta^2 - k'^2 - \delta_b^2}$$

$$t_g = \frac{\mathrm{d}\varphi}{\mathrm{d}\omega} = \frac{\mathrm{d}\varphi}{\mathrm{d}\beta} \cdot \frac{\mathrm{d}\beta}{\mathrm{d}\omega}$$

$$= -\frac{2\delta_b(\beta^2 + k'^2 + \delta_b^2)}{(\beta^2 - k'^2 - \delta_b^2)^2 + (2\beta\delta_b)^2} \frac{\mathrm{d}\beta}{\mathrm{d}\omega} \quad \ldots\ldots(1)$$

If we can approximate $\beta$ by $\dfrac{2\Delta\omega}{\omega_0}$ which can be done for narrow relative bandwidths then $\dfrac{\mathrm{d}\beta}{\mathrm{d}\omega} = 2/\omega_0$ and

$$\frac{t_g}{t_{g0}} = \frac{\{1 + (Q_b k')^2\}\{1 + (Q_b k')^2 + Q_b^2 \beta^2\}}{\{1 + (Q_b k')^2\}^2 + 2(Q_b \beta)^2 \{(Q_b k')^2 - 1\} + (Q_b \beta)^4} \quad \ldots\ldots(2)$$

since $\delta_b = 1/Q_b$

$$t_{g0} = \frac{2\delta_b}{k'^2 + \delta_b^2} \cdot \frac{2}{\omega_0} = \text{centre-frequency group delay}$$

$$= \frac{4Q_b}{\omega_0} \cdot \frac{1}{1 + (k'Q_b)^2} \quad \ldots\ldots(3)$$

*Maximally Flat Group Delay.* The condition for maximal flatness in eqn. (2) is that the coefficients of corresponding powers in numerator and denominator should be equal. This leads to

$$(k'Q_b)^2 = \tfrac{1}{3}$$

*Exact Form of $t_g$.* The exact form of $t_g$ results if $\beta$ is put equal to $\dfrac{\omega}{\omega_0} - \dfrac{\omega_0}{\omega}$ then

$$\frac{\mathrm{d}\beta}{\mathrm{d}\omega} = \frac{1}{\omega_0} + \frac{\omega_0}{\omega^2}$$

$$= \frac{1}{\omega_0}\left(1 + \frac{\omega_0^2}{\omega^2}\right)$$

Thus the centre-frequency group delay will be as given previously, while the expression for $t_g/t_{g0}$ in eqn. (2) is multiplied by $\dfrac{1}{2}\left(1 + \dfrac{\omega_0^2}{\omega^2}\right)$.

*Manuscript received by the Institution on 5th September 1963. (Paper No. 895)*

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U.D.C. 621.396.969.3

# Methods of Distinguishing Sea Targets from Clutter on a Civil Marine Radar

By

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*Presented at a meeting of the Radar and Navigational Aids Group in London on 23rd October 1963.*

**Summary:** Mathematical expressions for the signal received from isolated targets and clutter are analysed to show how the signal/clutter ratio is influenced by the parameters of the radar set. Graphs of signal amplitudes are then used to indicate the characteristics of the signals which can be exploited to improve the observed signal/clutter ratio. The limitations introduced by the display and the techniques used to overcome them are described.

Change of wavelength is often suggested to improve the signal/clutter ratio. It is shown from independent reports that the effect is small and is swamped by the limitations of the display, in some cases at least by non-use of the signal processing controls—swept gain, differentiation and gain.

Other possible techniques are examined, and reasons are given as to why they are not in use on current civil marine radars. It is then shown that such equipment is already operating close to limits set by components and space on board ships, so that improvement is only possible by changes in the processing controls, and better use of them by the operator.

## 1. Basic Principles

A civil marine-radar set which is sensitive enough to display small targets like buoys and boats will also be sensitive enough to display signals from waves, airborne raindrops, snowflakes, and sand particles. Such signals are known as clutter and can be a serious obstruction to the radar observation of buoys and boats. Reports from users indicate that difficulty is sometimes experienced in observing ships in the presence of clutter.

An extended target like sea clutter fills the whole of the radar beam, while a small target like a buoy does not. The equations describing the signal received in the two cases will therefore present certain differences, which indicate how the designer must proceed in order to minimize the clutter signal.

The strength of the signal received from an isolated target is expressed by the equation:

$$S = \frac{P.K.A^2}{4\pi\lambda^2}\,\frac{\sigma}{R^4} \qquad \ldots\ldots(1)$$

where $P$ = peak transmitted power

$K$ = constant expressing aerial characteristics

$A$ = area of aerial aperture

$\lambda$ = wavelength

$\sigma$ = effective echoing area of target

$R$ = range of target.

† Kelvin Hughes Division of S. Smith & Sons (England) Ltd., Dagenham, Essex.

It will be noted that the first part of this expression refers to the characteristics of the radar set, and the second part to the characteristics of the target.

The signal strength received at any instant from airborne rain, snow or sand, is the sum of the individual signals from a large number of isolated targets extending more or less uniformly over the whole cross section of the beam, and over a range interval corresponding to the pulse length. Such signals behave like random noise and the received signal may be expressed by the equation:

$$S' = \frac{P.K.A^2}{4\pi\lambda^2}\,\frac{C.R\theta.R\phi.\tau}{R^4}$$

$$= \frac{P.K.A^2}{4\pi\lambda^2}\,\frac{C.\theta.\phi.\tau}{R^2}$$

where $C$ = effective echoing area per unit volume of clutter

$\theta$ = horizontal beamwidth of aerial

$\phi$ = vertical beamwidth of aerial

$\tau$ = pulse length.

Hence the target/clutter ratio $N$ is given by:

$$N = \frac{S}{S'} = \frac{\sigma}{C.R^2\theta.\phi.\tau}$$

This requires modification in the case of sea clutter which does not extend over the whole of the vertical beamwidth. The equation for the received signal may be re-written



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$$S' = \frac{PKA^2}{4\pi\lambda^2}\,\frac{C'R\theta.\tau}{R^4}\sin D$$

where $C'$ = effective echoing area per unit plan area of clutter

$D$ = angle of depression.

Putting

$H$ = height of aerial

Then $$\sin D = \frac{H}{R} \text{ approximately}$$

since $R \gg H$

Hence $$S' = \frac{PKA^2}{4\pi\lambda^2}\,\frac{C'R\theta.\tau}{R^4}\,\frac{H}{R}$$

$$= \frac{PKA^2}{4\pi\lambda^2}\,\frac{C'\theta.\tau.H}{R^4}$$

from which

$$N = \frac{S}{S'} = \frac{\sigma}{C'.\theta.\tau.H}$$

This makes it quite clear that the only ways of increasing the ratio of target signal to clutter signal are by:

(*a*) increasing the echoing area of the target,

(*b*) decreasing the beamwidth of the aerial,

(*c*) decreasing the pulse length,

(*d*) reducing the height of the aerial above the sea (sea clutter only).

Other changes affect target and clutter equally.

Some of these changes are not completely under the control of the radar designer. For example, putting corner reflectors on buoys is the concern of Trinity House or the port authorities. Then the vertical beamwidth of the aerial must lie between 20 and 30 deg, since aerials are not stabilized against the roll of the ship, and the Ministry of Transport specification requires adequate performance in a roll of $\pm$ 10 deg. Finally the height of the scanner must lie within the limits of the ship's structure; a compromise must be made between good long-range performance and freedom from 'shadow sectors' due to masts, derricks, and funnel, which are both obtained by mounting the aerial as high as possible, and the consequent increase of sea clutter.

In practice, the horizontal beamwidth and the pulse length are the only parameters under design control, and then only within certain limits. There is a practical limit to the size of the scanner even on large ships, while the usable pulse length is a compromise, since component limitations define the minimum, and long-range performance suffers by the use of short pulses and the consequent wide-band receivers.

## 2. Graphical Representation of Signals

A clearer impression of relative signal strengths can be obtained by plotting the variation of mean signal strength against range. Due to wave motion, instantaneous signal levels will fluctuate from scan to scan, but mean levels for a number of targets are illustrated in Fig. 1. Two points are immediately obvious:

(*a*) The signal from a ship is well above the sea clutter at all ranges. Difficulty in distinguishing the ship signal will therefore arise only from processing and display problems.

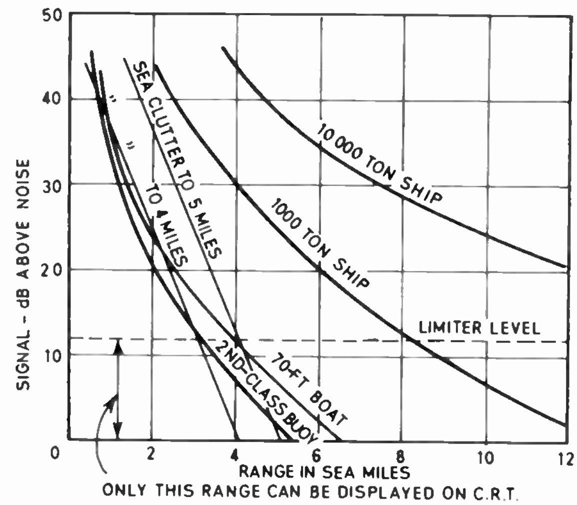


**Fig. 1.** Mean signals from various targets.
3-cm radar with the scanner 120 ft above sea level.
Note constant slope of sea clutter line since this target fills the beam.

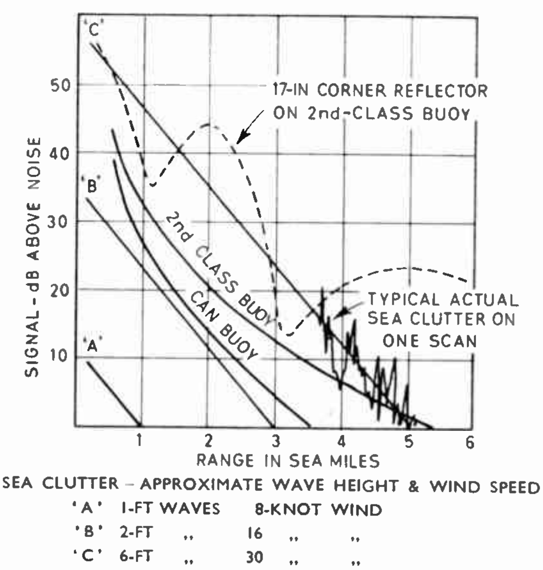


**Fig. 2.** Mean sea clutter and buoy signals.



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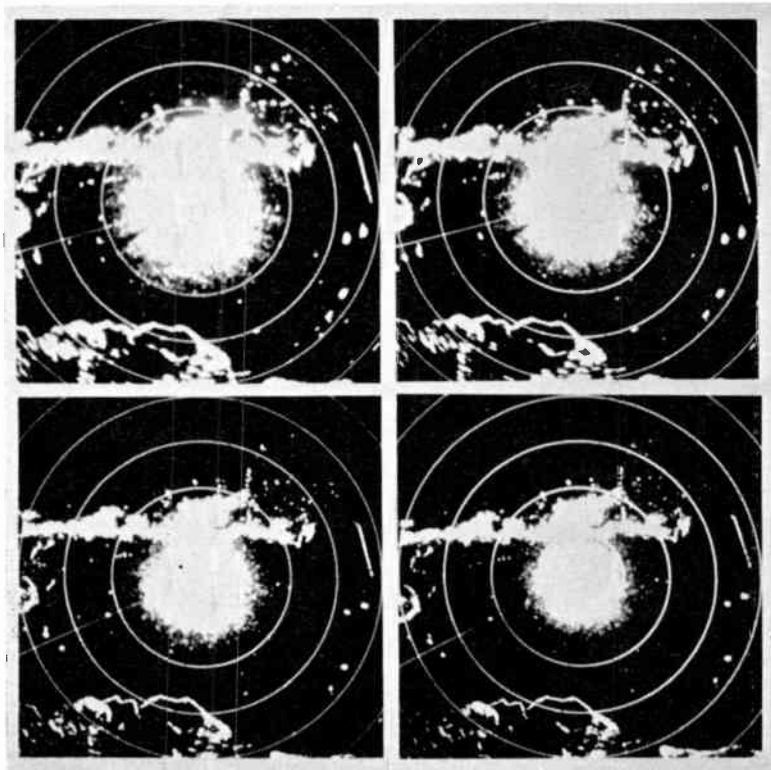

(*a*) 6-ft scanner, pulse length 0·35 μs.

(*b*) 6-ft scanner, pulse length 0·18 μs.

(*c*) 10-ft scanner, pulse length 0·35 μs.

(*d*) 10-ft scanner, pulse length 0·18 μs.

**Fig. 3.** Effect of pulse length on sea clutter.

(*b*) The signal from a boat or buoy may be less than the clutter signal at certain ranges. No practical technique of signal processing can remedy this position for a marine radar. The ratio can only be improved by applying the results of the mathematical analysis in Section 1—that is, either by increasing the echoing area of the target, for example by fitting a corner reflector to a buoy; or by modifying the radar set where possible, for instance by using an aerial with a narrower horizontal beamwidth, or by reducing the pulse length.

The effect of fitting a corner reflector to a buoy is illustrated in Fig. 2. The maxima and minima exhibited by the corner-reflector graph are due to interference between the direct transmission from the aerial to the corner reflector, and the transmission reflected from the sea surface. As the range decreases the path difference increases, producing a minimum at intervals of a half-wavelength.

To illustrate what can be done by altering the pulse length, Fig. 3 shows four pictures of targets lost in sea clutter. The left-hand pictures use 0·35 μs pulses, while those on the right hand use half this pulse length. Note the reduction in radius of the sea clutter.

The improvement produced by using a larger scanner is illustrated in Fig. 4. The display shown in Fig. 4(*b*) uses a 6-ft scanner, the other two use a 10-ft scanner of approximately half the beamwidth. Comparing Fig. 4(*a*) with Fig. 4(*b*), there is a slight increase in the maximum range of sea clutter, but the signal strength from the buoys is markedly improved. Note particularly the two buoys at 140 deg 2·4 miles and 155 deg 2·9 miles. The three targets at about ¾ mile at the 4, 6, and 7 o'clock positions are equally visible in spite of the increase in radius of sea clutter. Range rings are shown at ½-mile intervals. In Fig. 4(*c*) the gain was reduced to give approximately equal signals from the buoys, resulting in a noticeable reduction in sea clutter radius.

Figure 5 is a very diagrammatic representation of signals from rain clutter, which may be taken to represent snow and sand clutter also. Note that unlike sea clutter, the variation of the signal with range is quite random. This signal may be considered to be made up of two components, a varying d.c. level, upon which is superimposed an a.c. component similar to receiver noise. It is shown in Section 3 how these signal characteristics lead to a technique of clutter suppression.

Figure 6 shows the appearance of rain clutter on a 3-cm set at 4 or 5 miles range, while Fig. 7 is a photograph showing rain clutter on a 10-cm air traffic con-



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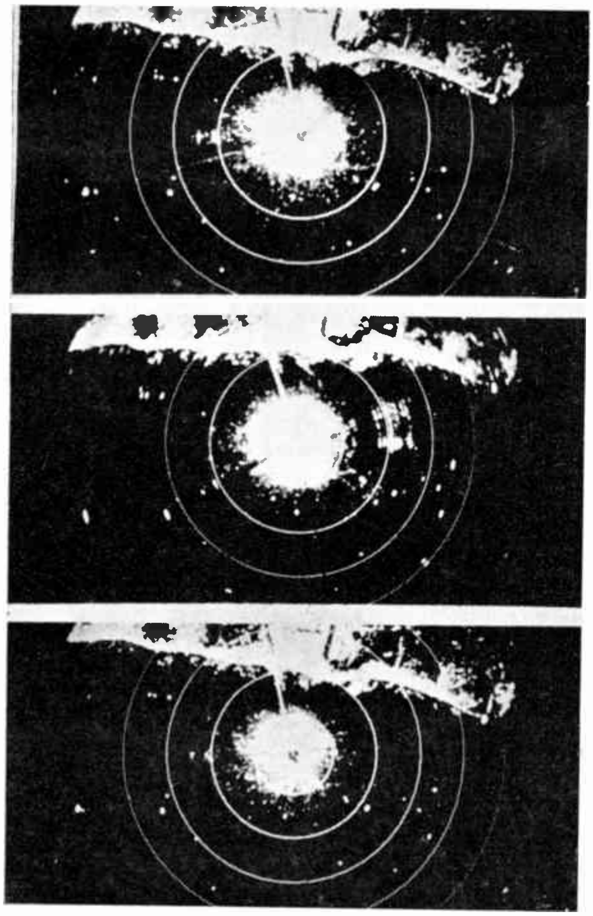
trol radar extending to about 60 miles range. This set uses a wide vertical beam to give the required altitude coverage, and suffers from heavy rain clutter. The technique now being used to reduce the rain clutter is to split the beam into a number of separate narrow beams in a vertical plane, i.e. to reduce the vertical beamwidth of each individual beam, and to feed signals from each beam into a separate amplifier channel. By suppressing the appropriate shorter range section of each beam, echoes below a few thousand feet are eliminated, thus removing from the display the uncontrolled low-flying traffic which is of no interest to the controller, most of the weather echoes, and reducing rain echoes at greater heights. In the marine case, the required signals are immersed in the weather echoes, so they cannot be suppressed completely, but narrower beams and short pulses can be used to reduce rain clutter, as demonstrated with sea clutter.



*Top* (*a*) 10-ft scanner.
*Centre* (*b*) 6-ft scanner.
*Bottom* (*c*) 10-ft scanner (reduced gain).

Fig. 4. Effect of beamwidth on sea clutter.

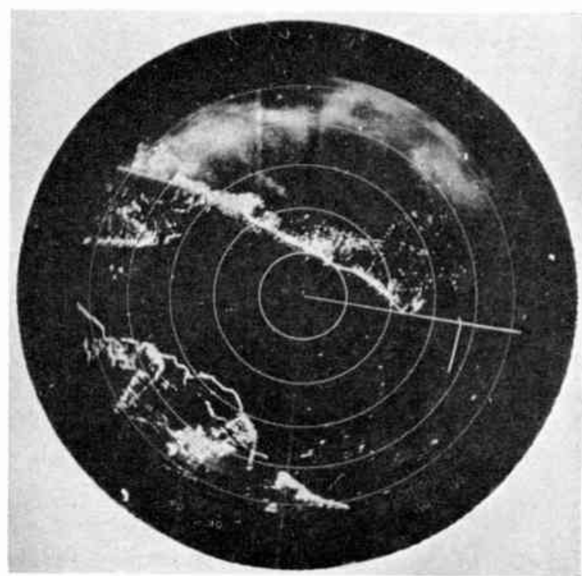


Fig. 6. Rain clutter on a 3-cm radar.

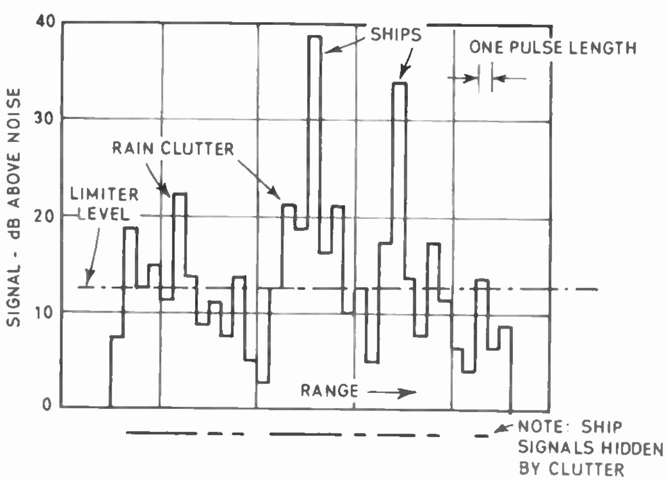



Fig. 5. Rain clutter and ship signals—single scan only.
Raw signals.

Fig. 7. Rain clutter on a 10-cm radar.

In considering the effect of these changes on the overall performance of a radar set, it must be remembered that a reduction of pulse length requires a wider receiver bandwidth, thus increasing receiver noise and *reducing* the maximum range of detection.

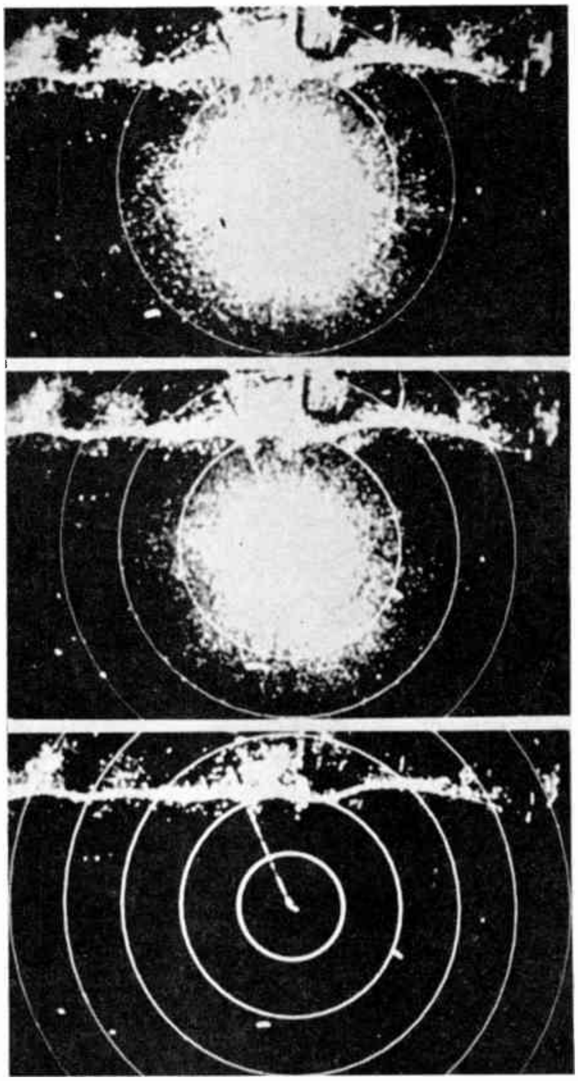
### 3. Signal Processing

The ordinary p.p.i. display on a cathode-ray tube has an important limitation which is a major cause of the difficulties in the detection of targets in clutter. It has been shown from graphs (Figs. 1, 2 and 5) that signal amplitudes from various targets and clutter can rise to many decibels above receiver noise, but the c.r.t. is unable to display more than about 12 dB of brilliance change. We therefore have to find some way of compressing a signal range of some 40–60 dB or more into the display range of 12 dB.

The essential first step is to limit the signals. The range from 0 to 12 dB must be displayed with the highest possible gamma, so that signals only just above noise are visible as clearly as possible. Any signals above this 12-dB level must be clipped, so that the c.r.t. is not driven into defocusing or actual damage. The receiver gain is adjusted so that noise is invisible or only just seen. All signals above the 12-dB level will be presented as of equal amplitude. The dotted line in Fig. 1 marked 'limiter level' indicates where this clipping occurs.

The amplitude differences which would be lost in doing this must now be regained; in the case of sea clutter, the fact that the clutter amplitude falls uniformly with range is used. We arrange that the receiver gain has a low initial value at short range, and increases at a rate corresponding to the fall in clutter amplitude, reaching its normal value at a range where the clutter becomes negligible.

Since the receiver gain is controlled by the bias on a number of stages, this bias can be developed across an RC network of appropriate time-constant. If the initial bias is correctly adjusted the time-constant will allow it to fall at the correct rate, and the signals will then remain within the display range. The law of recovery having been set, all that needs to be controlled is the initial gain reduction, to correspond with the actual clutter amplitude. Figure 8 illustrates this process. For comparison, the effect of halving the



*Top* (*a*) 0·35-μs pulse, no swept gain.
*Centre* (*b*) 0·18-μs pulse, no swept gain.
*Bottom* (*c*) 0·35-μs pulse, correct swept gain.

Fig. 8. Effect of swept gain control.



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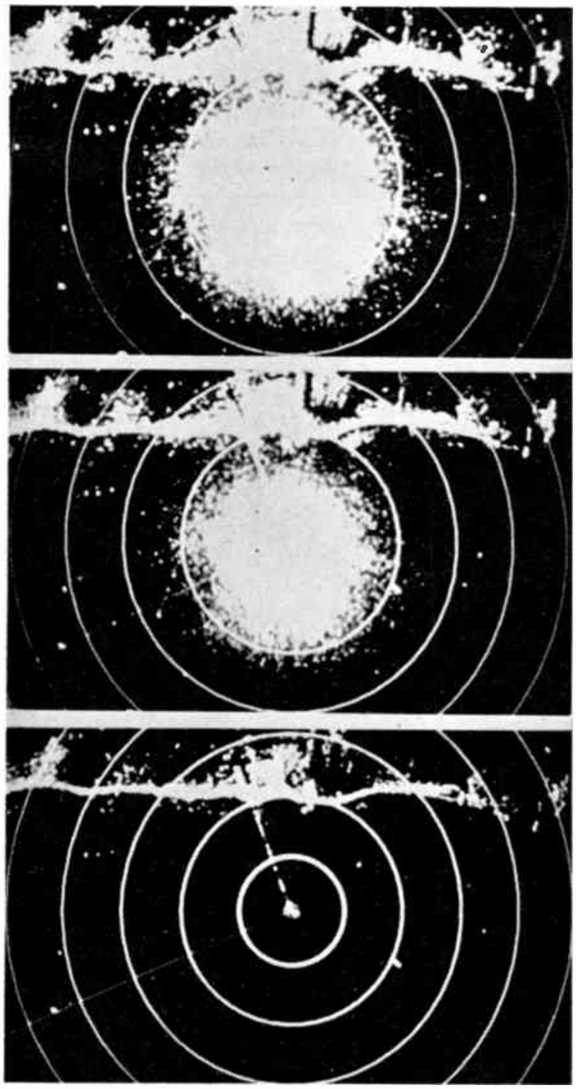


*Top* (*a*) 0·35-μs pulse, no swept gain.
*Centre* (*b*) 0·18-μs pulse, no swept gain.
*Bottom* (*c*) 0·35-μs pulse, too much swept gain.

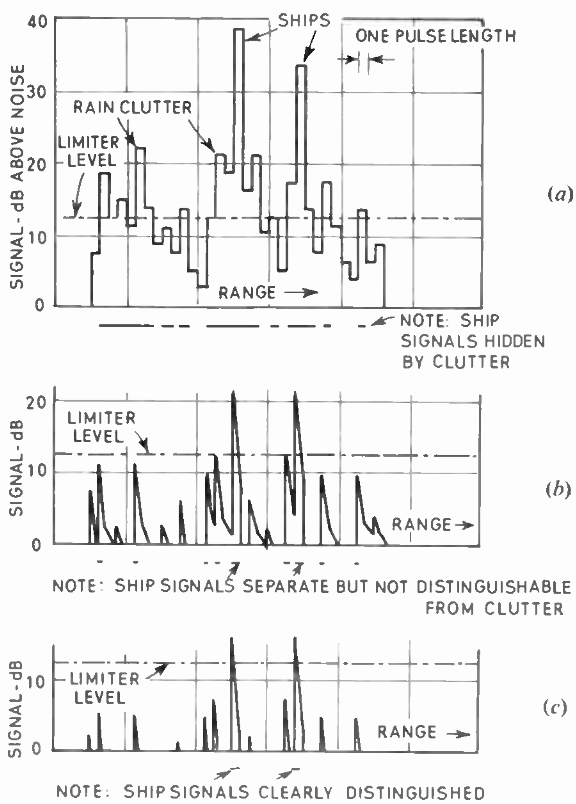
**Fig. 9.** Effect of swept gain control.

pulse length is shown first. The buoy at 1·4 miles 160 deg which just becomes visible when the pulse length is halved is shown clearly in Fig. 8(*c*) where the longer pulse is used but the sea-clutter control is correctly adjusted. Note that a slight trace of sea clutter is left. It is obvious that this processing technique is far more effective than halving the pulse length.

Some judgment must be exercised in the use of this control as can be seen in Fig. 9 in which the sea clutter has been completely suppressed. The small fishing vessel at ¾ mile 70 deg now stands out clearly, as does the pier, but the buoy at 1·4 miles has been completely suppressed along with the sea clutter. Hence while there should be no difficulty in isolating ship targets from clutter, the observation of smaller targets in heavy clutter requires careful adjustment of the controls.

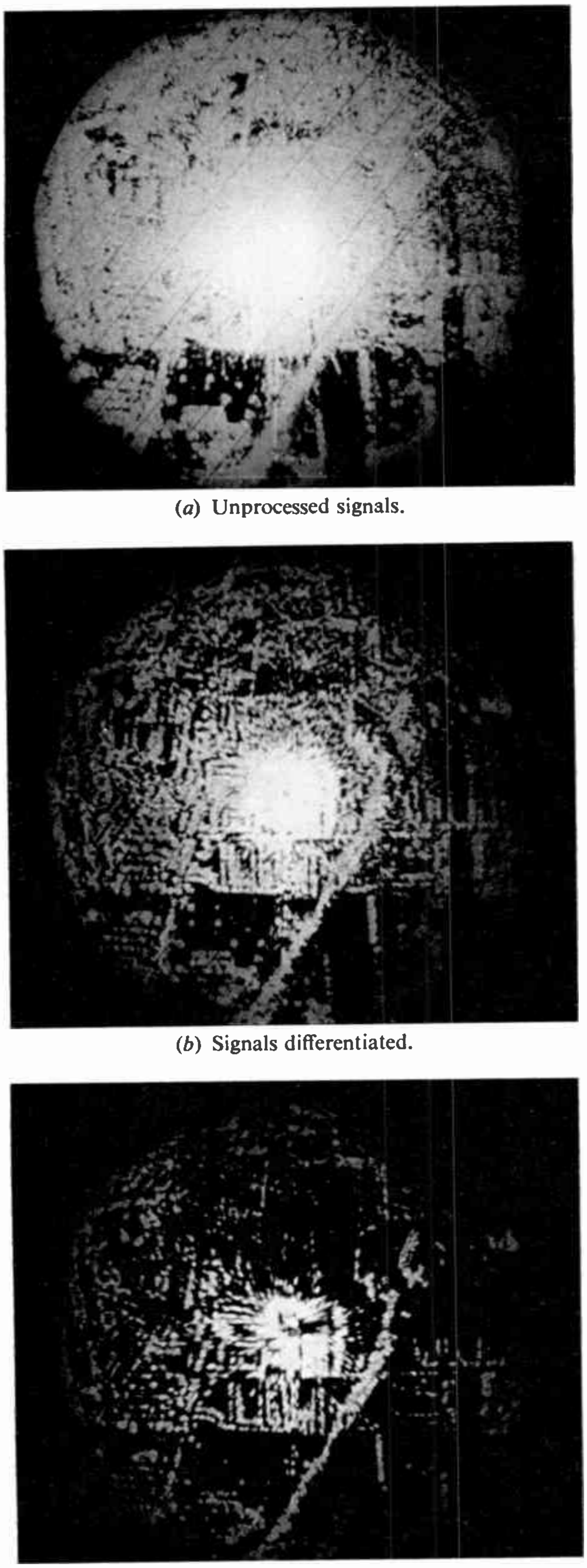
Rain clutter does not fall uniformly with range, but is distributed in a random manner. One cannot therefore use the swept gain control, a point not appreciated by many observers. Instead, the signals are passed through a differentiating circuit with a time-constant shorter than the pulse length (see Fig. 10). This removes the d.c. level, characteristic of these signals, and effectively shortens the pulse length since only the leading edge of each individual signal is used. If this is insufficient to bring the signals within the display range, then the gain must be reduced.

The effect on actual signals is illustrated by Fig. 11(*a*). This shows the area around the factory where the radar set is operating. The set is in the centre of the picture and the factory buildings are grouped around it. To the south (below this) are parallel lines of houses, while to the north lie rough fields and hedges.




(*a*) Raw signals. (*b*) Signals differentiated.
(*c*) Gain reduced by 5dB.

**Fig. 10.** Rain clutter and ship signals—effect of differentiation.



(*a*) Unprocessed signals.

(*b*) Signals differentiated.

(*c*) Signals differentiated and gain reduced.

**Fig. 11.** Factory area.

The diagonal line from the bottom edge towards the top right-hand corner is the embankment of a railway with more streets partly hidden beyond it. These are relatively large targets, and most of the signals reach limiter level so that all detail is lost. After differentiation shown in Fig. 11(*b*), much of this detail is recovered, but some smaller signals still confuse larger ones. The field north of the centre shows echoes from tufts of grass, while the streets to the south are indistinct due to echoes from trees. After reduction of gain (Fig. 11(*c*)) the smaller echoes are subdued and the main outilnes stand out clearly. In exactly the same way ships may be displayed clear of surrounding clutter.

### 4. The Effect of Change of Wavelength

Since the wavelength of the radar set appears in the first part of the equations discussed earlier, it is cancelled out of the expression for the signal/clutter ratio, although of course some wavelength dependence may remain in the clutter reflection coefficient. This result is perhaps surprising in view of the popular belief that 10-cm radar will display targets in clutter much better than 3-cm radar. This impression may have arisen by comparing sets with different sensitivity —what has been compared is the radius of the sea clutter circle without adjustment of the processing controls. Since the 3-cm set is more sensitive to small targets, particularly those low in the water such as sea clutter, clutter is seen at a greater range. If the 10-cm set does not show clutter, this is in fact a warning that it will not show small targets such as sandbanks and low coastlines. On the other hand, if it is sensitive enough to see these, then under bad conditions targets will be obscured by the clutter, unless the operator adjusts the controls. When this is done, the superior definition of the 3-cm set gives it the advantage.

The Netherlands State Commission for Electronic Navigational Aids organized a comparison of 10-cm and 3-cm radar, on the same ship, in the early part of 1962, and have issued a report[2] with their conclusions. Later in the same year the Radio Aids to Marine Navigation Applications Committee (RAMNAC) in this country considered whether 10-cm equipment should be admitted to the British Ministry of Transport Specification, and commissioned their research group to do some comparative trials in order to verify the point. The report[3] of the research group was submitted to the committee in January 1963.

The Netherlands trials used two sets of similar transmitting power and receiver sensitivity, but both scanners were of about 10-ft aperture, so that the 10-cm set had a beamwidth of one and three quarter degrees while the 3-cm set had a beamwidth of about three quarters of a degree, and hence higher aerial gain. The observations of this report are interesting,



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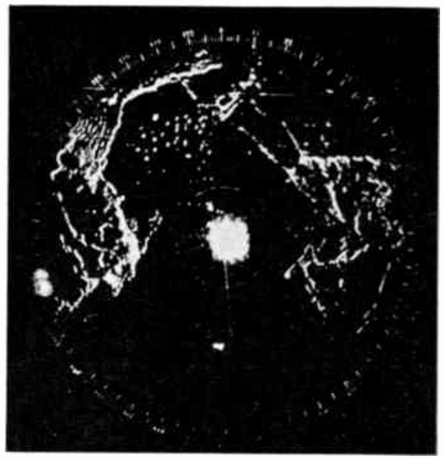
(*a*) 10-cm radar.



(*b*) 3-cm radar.

**Fig. 12.** Radar picture of the entrance to Suez Canal.

since they are all accompanied by photographic evidence.

They are:

(*a*) The maximum detection range of targets was 27% greater on the 3-cm set. This is what would be expected from the higher aerial gain and it is verified by the photographs.

(*b*) Photographs show sea clutter extending to a greater range on the 3-cm set using long pulses, but the reverse on short pulses. No measurement of signal/clutter ratio was attempted, and unfortunately the photographs and the data attached to them indicate that the available processing controls were not fully used.

(*c*) Photographs show rain clutter on the 3-cm set obscuring a target which was clear of clutter on the 10-cm set. The data attached to the photographs indicate that the available processing controls were not used at all.

(*d*) Photographs show the expected higher definition of the 3-cm set (see Fig. 12). No comment is made on the fact that the 3-cm photograph (Fig. 12(*b*)) clearly shows several obstructions at ranges of a few miles which are completely absent from the 10-cm photograph (Fig. 12(*a*)). These obstructions make up the coastline, which is too low for the 10-cm equipment to detect, a point which the RAMNAC report takes up. The Netherlands report finally stated that no definite conclusions can be reached from these results, and suggested further trials.

In the RAMNAC trials measurements of signal strength were made for buoys, ships, sea clutter, and rain clutter for two radar sets, a 10-cm set with a 12-ft scanner, i.e. about $1\frac{3}{4}$ deg beamwidth, and a 3-cm set with 6-ft scanner, i.e. about $1\frac{1}{4}$ deg beamwidth, the other parameters being approximately the same. It would take too long to analyse this report in the detail it merits, but it concludes: "From the results of these trials the 3-cm radar used was significantly more sensitive to small targets in clear conditions. But if there was any great amount of clutter from sea or rain, a small target in the clutter area would generally be more difficult to locate than it would be on a 10-cm radar of equal performance."

The conclusion goes on to state the necessity of increased sensitivity for the 10-cm set, to enable it to pick up low targets such as sandbanks to the specification requirements. The implication of this is that a 10-cm set of such increased sensitivity will display clutter at the same intensity and range as a 3-cm set, and the separation of targets from clutter will therefore be no easier on one set than the other.

It would have been exceedingly interesting if this report had included measurements on a 3-cm set with a 10-ft scanner. The report states: "The signal amplitude of the buoy above the sea clutter would be at least $1\frac{1}{2}$ dB greater on the 10-cm set", while for rain clutter the 10-cm set has a 3 dB advantage. These are the measured figures using a 6-ft scanner on the 3-cm set. If this set uses a 10-ft scanner, i.e. has the same ship fitting problems as the 10-cm set, then the mathematical analysis indicates a 3 dB improvement, making the 3-cm set comparable to the 10-cm set against rain clutter, and $1\frac{1}{2}$ dB better against sea clutter.

We have seen from the graphs that signals from most targets are many decibels above the clutter amplitude. There is, therefore, some doubt as to the importance of the $1\frac{1}{2}$ dB and 3 dB margins discussed. Indeed, one of the conclusions of another authority[1] with regard to anti-clutter circuits is as follows:

"From the steep law of clutter versus range which is to be expected with aerial heights common on ships,



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it is clear that only improvements of say 10 dB are worth while. With lesser improvements the increase of the range interval in which a given target is visible would be so small that no increase in complication appears to be justifiable."

The RAMNAC report also emphasizes the necessity of intelligent use of the controls. The following is only one of several similar comments referring to the 3-cm radar.

"Under rain clutter conditions where the buoys were visible with difficulty with no sea clutter applied, by reducing the gain by 10 dB or the use of a short time constant (i.e. differentiation) they were clearly visible."

The logical conclusion from this evidence, supported by the photograph of rain clutter on a 10-cm set (Fig. 7), is that the visibility of clutter on the display depends on the sensitivity of the radar, transmitter power, receiver noise figure, and so on. The change of wavelength from 3 to 10 cm appears to have negligible effect on the signal/clutter ratio, and the evidence suggests that within this frequency variation, changes in the reflection coefficient of the clutter would be more than cancelled by opposing changes in aerial beamwidth, if the aerial horizontal aperture is maintained constant. The correct adjustment of the processing controls is of immeasurably greater significance in obtaining clearly visible echoes.

## 5. Other Techniques

Many other methods have been suggested to improve the visibility of signals in clutter. The more important of these will be summarized.

### 5.1. *Logarithmic Receiver*

Figure 13 is a graph of receiver output, which illustrates in a different way how the 12-dB output limitation obscures the differences between signals above this level. Logarithmic receivers preserve these differences by compressing all signals within the acceptable range, but at the expense of reducing all displayed differences—not only the signal/clutter differences but also the differences of long range signals above receiver noise. Hence a bigger signal/clutter difference is needed to make the signal visible, but the operator need make no adjustments to achieve this. Correct adjustment of the available controls on a linear receiver will always produce a clearer signal in clutter than such a logarithmic receiver. The logarithmic receiver has the additional disadvantage of increased cost and complexity.

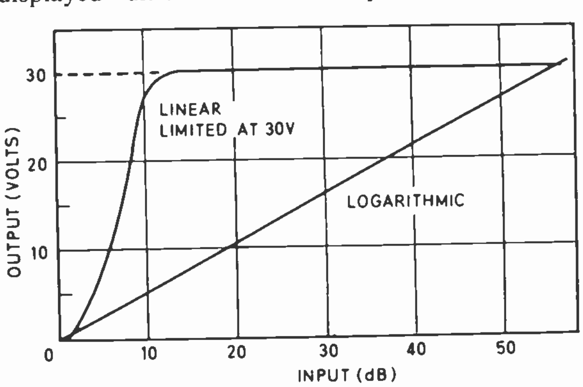
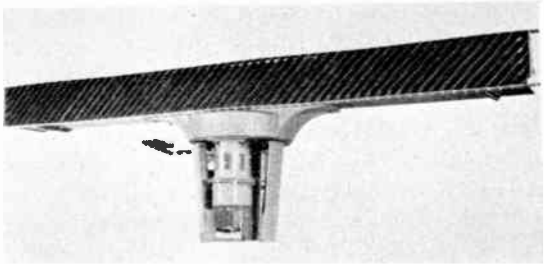



**Fig. 13.** Receiver amplitude characteristics.



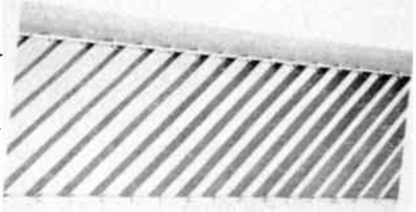


**Fig. 14.** Scanner and quarter-wave plate.

### 5.2. *Polarization*

The choice of horizontal, vertical or circular polarization has little effect on sea or rain clutter, but a technique of conversion to circular polarization can be very effective against rain. The polarization can be converted from linear to circular by the use of a quarter-wave plate, such as the one illustrated in Fig. 14. In front of the radiating aperture is a frame carrying metal strips at 45 deg to the horizontal, so that the radiation traverses the gaps between the strips. Assuming the incident transmitted radiation is horizontally polarized and the direction of the electric vector is 0 deg, this can be considered to be made up of two components at +45 deg and −45 deg. The +45-deg component passes between the strips as if it were in free space, but the −45-deg component is in a waveguide, and its phase velocity is increased. If the dimensions and spacing of the strips are correctly chosen, this component emerges with a 90-deg lead (in time) on the other component, so that they recombine to form circularly-polarized radiation. If this circularly-polarized radiation returns unmodified from a target and traverses the quarter wave plate in the opposite direction, the same component is



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again advanced in phase by 90 deg, giving a total advance of 180 deg. When the two components emerge from the space between the strips they recombine into vertically-polarized radiation.

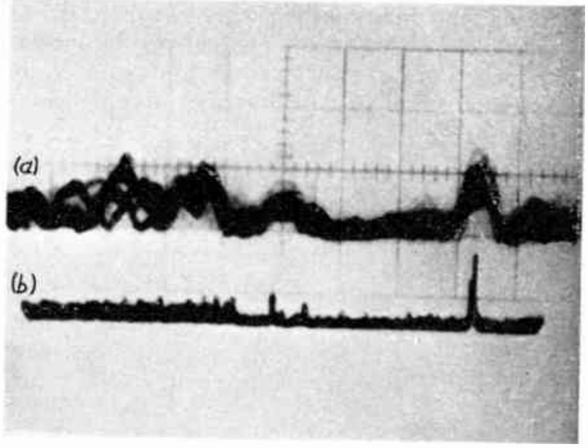
Thus if a quarter-wave plate like the one in Fig. 14 is placed in front of a horizontally-polarized aerial the radiation is circularly polarized, and the scattered wave from a spherical raindrop will still be circularly polarized owing to the symmetry of the target. On returning through the quarter-wave plate its polarization will become vertical, and will not be accepted by the aerial. This is only true for a spherical target like a raindrop, and is not true for an angular object like a ship or buoy. Signals from such targets will still have considerable energy in the horizontal plane of polarization, which will be accepted by the aerial. Thus the raindrop signal is almost reduced to zero while other signals suffer only a slight loss. The claimed advantage against rain clutter is about 15 dB, at the expense of perhaps a 6-dB loss on all signals. Unfortunately, the advantage only applies against rain. The snow, sea, and sand clutter encountered by a marine radar are unaffected. In addition, tests at the Admiralty Surface Weapons Establishment have shown that the performance against certain marine targets, particularly reflector buoys, was poor when circular polarization was used.[4]

Figure 15(*a*) shows the effect of the quarter-wave plate in suppressing the echo from No. 1 Sea Reach

(*a*)

(*b*)

Fig. 15. Suppression of corner reflector buoy by quarter-wave plate.




(*a*) Normal radar, 5-μs pulse. (*b*) Chirp radar.

Fig. 16. The effect of chirp radar on a snowstorm.

buoy, which is fitted with a corner reflector. The buoy lies just outside the range ring as indicated by the arrow in the lower picture taken without the quarter-wave plate. These disadvantages indicate that the marine use of circular polarization is inadvisable.

### 5.3. '*Chirp*' *Radar*

This is a technique of frequency modulation within the pulse, the received signals being passed through a dispersive delay network so that the different frequency components are brought together in time. This is equivalent to the use of a very short pulse.

Figure 16 shows simultaneous photographs of two displays. The upper trace is the signal using a 5-μs pulse, and the echoes in the centre from the Droitwich radio masts are completely obscured by the clutter from a snowstorm. In the lower trace, the pulse is swept through a frequency of 5 Mc/s, and after passing through a dispersive network, becomes the equivalent of a $\frac{1}{4}$-μs pulse. While the amplitude of the aircraft signal on the right is unaltered, the clutter signals are so reduced that the radio masts are clearly visible. The improvement in signal/clutter ratio is about 13 dB.

In this case, the technique requires a receiver bandwidth of 5 Mc/s but it is the equivalent of the use of a high-power short pulse not obtainable by normal means. In a typical marine radar a pulse length of $\frac{1}{20}$ μs requiring about 20 Mc/s bandwidth is usual; to obtain even a small effective reduction of this pulse length by a factor of 3 would require a receiver bandwidth of about 60 Mc/s. A transmitter using a pulsed oscillator and power amplifier would be essential, rather than one using a magnetron, but the cost and complication of such a system renders it unsuitable for civil marine radar.



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### 5.4. *Doppler Effect*

It has been suggested that the movement of the clutter with respect to the target would produce Doppler shifts of the received radiation which might be used to distinguish one from the other. Unfortunately this is not so since the difference in velocity is too small, only 30 c/s for 1 nautical mile per hour at 3 cm, leading to impossible filter requirements. Reference 1 indicates that the technique might be possible for a single target in sea clutter, but a civil marine radar has to display many targets at once, some of which would have the same radial velocity as the clutter and would thus be indistinguishable. Hence a Doppler system would have the serious disadvantage of suppressing some targets as well as the clutter.

### 5.5. *Pencil Beam Stabilized Scanner*

The vertical beamwidth of 20 to 30 deg can be considerably reduced if the scanner is stabilized against ship movement. The mathematical analysis indicates that this would reduce rain, snow and sand clutter considerably—perhaps by about 10 dB. It would, however, have no effect on sea clutter, since the vertical extent of the sea clutter would still lie within the beam. Because of this, and the cost and weight disadvantages, it is not used in civil marine radar.

### 5.6. *Integration*

Reference 1 quotes observations of a single peak of sea clutter, which is found to have a persistence of about 1/100 second. Rain clutter behaves in a similar way.[5] A marine radar scanner with $1\frac{1}{4}$ deg beamwidth rotating at 24 rev/min illuminates a target for about 1/150 second. Thus the clutter peak does not change during the time the radar set records it, and behaves exactly like the signal from a ship or buoy. Hence the clutter signal benefits no more nor less from pulse-to-pulse integration than the target signal does.

The only way integration is of benefit is where long-term integration takes place in the afterglow of the screen. A buoy lost in clutter can sometimes be seen clearly if the tube brilliance is temporarily turned down, appearing in the form of a brighter spot in the clutter afterglow. This is another operational technique apparently known only to a minority of operators.

The large number of elements of area to be displayed separately (about $\frac{1}{4}$ million) makes it impractical to consider any integration other than on the c.r.t. screen. This suggests the need for a new approach.

### 5.7. *High Speed Scanner*

In view of the comments already made on integration, it would appear likely that an improvement in signal visibility could be obtained by rotating the scanner at a higher speed and allowing the signals to integrate on the c.r.t. screen. If the speed is increased until only one pulse is radiated per beamwidth, it would be necessary to modify the aerial, since the main lobe of its radiation diagram will have moved a significant amount by the time the signals are returning from long-range targets. Without such modification the speed limit is about three or four revolutions per second, perhaps as high as it is desirable to go. At this speed the individual clutter signals have time to change and move between successive observations of the same area. If the c.r.t. screen has a decay time-constant of about 10 seconds, some twenty to forty signals will be integrated to produce the brightness of each element of the screen. The improvement in signal visibility will of course depend on the characteristics of the clutter, but 6 dB or more appears probable.

Such a scanner would raise problems of wind resistance and gyroscopic effects, but these are capable of solution by good mechanical design. A more difficult problem may arise because the screen is not observed as a whole. The observer looks at an echo on a small part of the screen. His eyes will, therefore, be receiving pulses of light at a rate of three or four per second. This is close to the frequency of the alpha rhythms of the brain, and dangerous physiological effects may be caused. Hence expert medical advice should be consulted on this stage of any experimental investigation.

### 5.8. *Techniques in Other Fields*

It is often useful to examine methods of dealing with similar problems in other fields. The problem of distinguishing targets in clutter arises basically from the loss of fine detail due to the gross contrast of the signal exceeding the contrast available on the display. This problem arises in x-ray photography, in photography for mapping, in the reproduction of television pictures, in the commercial production of prints from photographs taken in different light conditions on the same film, and in many other fields. The broad principle of 'harmonization' which is applied to all these different processes is to use the local mean signal to modify the signal level in the area so that the available contrast lies on the steepest part of the reproduction characteristic—in just the same way a photocell-controlled iris adjusts the light admitted to a television or cine camera. Reference 6 summarizes much of the information and describes a technique which seems very effective for x-ray use. In an attempt to apply this technique to the processing of radar signals some years ago, a limiter circuit was used in which the gross changes shift the d.c. level of the signals, so that the fine detail is



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retained within the working limits of the display. While this gives a measure of automatic operation if the observer does not adjust the controls, it still passes the fine detail of the clutter, with little reduction of its intensity. It also has a serious effect on the operation of the swept gain control. In the difficult conditions when the signal is not much larger than the clutter, a good observer can obtain a clearer picture by careful adjustment of the controls using a conventional limiter.

## 6. Design Limitations

It has been shown that the practical ways in which the signal can be made more visible in the presence of clutter are by reducing beamwidth and pulse length. It is, therefore, of interest to know to what extent designs can be modified in these directions.

### 6.1. *Cost and Size*

In a competitive market the design engineer must consider the relative expense of different components, methods of manufacture, tools and test gear to achieve a design which meets the specification, at a price the customer can afford. In the marine radar market, the selling price of a complete equipment lies between about £1,000 and £5,000, according to the facilities and accessories provided. The average price is about £2,000, and it is against this figure that the cost of changes must be considered.

Compared with this figure, a stabilized scanner is likely to increase the selling price by at least 50% while 'chirp' radar would be considerably more. The improvement obtained by their incorporation would not justify such expense, and the equipment would be priced out of the market.

A 6-ft scanner accounts for about 20% of the average price. If the alternative 10-ft scanner is used, this increases the price by approximately a further 20%. This not only improves the detection of targets in clutter, but also gives better definition and long range performance. Although 10-ft scanners are currently fitted to one in seven of the equipments in the author's knowledge, there are many relatively small ships using big scanners, so size does not appear to be the limiting factor, suggesting that perhaps it is cost.

The use of 10-ft instead of 6-ft scanners gives about 2½ dB improvement; but a 14-ft scanner will give an additional improvement of only 1½ dB. Still longer scanners provide a further improvement but only by decreasing amounts. A worth-while improvement from the clutter point of view is only obtained by nearly doubling the aerial length.

All manufacturers now incorporate pulse-length switching as a standard feature but the cost of this in itself is negligible. If, however, this is coupled with receiver bandwidth switching to obtain the maximum benefit then the price of the i.f. amplifier increases by about 2% of the total. For this cost, the bandwidth for longer ranges could probably be reduced by a factor of 3, giving a receiver noise figure which is about 4 dB better. However, since this is only evident on long range targets where the inverse eighth-power law applies, the increase in maximum range would be no more than 10%. The inverse eighth-power law is peculiar to marine radar, and occurs when the target and scanner are relatively near a horizontal reflecting surface. The direct ray between scanner and target is subject to interference from the ray reflected from the sea surface. Since there is a phase change of 180 deg on reflection, this interference is destructive when the path lengths are equal or nearly equal, i.e. when the target and scanner are low. This is the limiting case of the signal fluctuations exhibited by the corner reflector buoy illustrated in Fig. 2.

Thus the advantage is small, and the pressure on obtaining reliability by simplicity, and on keeping cost low has prevented the use of bandwidth switching in most equipment.

### 6.2. *Components*

Component limitations apply only to the employment of short pulses. Receivers can be made with any bandwidth required at present, and the only limitation is the magnetron used as the pulse transmitter.

The voltage pulse applied to its cathode can be divided into three regions, the first being the rise of the pulse, the second the flat top, and finally the fall. In the flat-top region the magnetron is oscillating steadily in its normal mode. It will oscillate in the rise and fall periods also, but if the rate of change of voltage is too fast, the magnetron oscillates in a different mode and changes frequency. Having started oscillating in this wrong mode during the rise, it is reluctant to return to its normal mode during the flat-top region of the pulse. This is a very unstable condition of operation and must be avoided.

In the type of magnetron in common use, the practical limit to the rate of rise is about 150 kV/μs. The interval between the voltage at which the magnetron starts to conduct and its correct operating voltage is about 5 kV. Hence the rise must not take place in a shorter time than about 1/30 μs. Allowing the same intervals for the flat top and the fall, the total modulator pulse time is about 1/10 μs. The magnetron is oscillating at full power for about half this time. Hence in practice, the minimum usable pulse is approximately 1/20 μs. This requires a receiver bandwidth in the region of 20 Mc/s, and the higher-definition sets from various manufacturers all have



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parameters of this order. Since the magnetrons are all of the same basic design, although they are produced by different manufacturers this is to be expected.

Summarizing, it can be said that on current production sets, pulse lengths are limited by the available components, while scanners are available to meet any demand to date for narrow beamwidth. In these two respects, different manufacturers' products are very similar. The differences in performance which exist arise mainly in the signal processing controls and in the ease and effectiveness of their operation by the user.

## 7. Conclusions

Targets can be made more easily visible in clutter by:

(*a*) using the biggest scanner (i.e. the one with the narrowest beamwidth); the limitation on this appears to be cost not space;

(*b*) using the shortest pulse consistent with the maximum range performance required; the limitation here appears to be existing component design.

The biggest improvement is obtainable by correct use of the processing controls (gain, swept gain, and differentiation). The engineer must ensure that these are effective and suitably located to facilitate operation. The observer must be trained to adjust these for prevailing weather conditions.

The evidence from carefully conducted trials indicates that 3-cm radar is at least as good as 10-cm radar in displaying targets in clutter provided the same size scanner is used. It is dangerous to regard the 'clutter-blindness' of some 10-cm equipment as an advantage, since such equipment is equally 'blind' to small targets such as sandbanks, and low coastlines.

Finally, the use of higher speed scanners appears to be worth investigation.

## 8. Acknowledgments

The author expresses his thanks to: Admiralty, for permission to quote from unpublished Admiralty Surface Weapons Establishment reports; the Royal Radar Establishment at Malvern for photographs and information on chirp radar; the Radio Aids to Marine Navigation Applications Committee for permission to quote from the papers detailed in the references; and the Netherlands State Commission for Electronic Navigational Aids for permission to comment on their report and the use of the photographs of Suez Bay. The author also thanks his colleagues, particularly Mr. P. Comber who assisted in the preparation of the illustrations, and the Directors of the Kelvin Hughes Division of S. Smith & Sons (England) Ltd., for permission to publish this paper.

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*Manuscript first received by the Institution on 29th October 1963 and in final form on 22nd January 1964. (Paper No. 896/RNA28.)*

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## DISCUSSION

*Under the Chairmanship of Mr. R. N. Lord*

**Mr. R. Thomason:** Would Mr. Harrison agree that most of the trouble resulting from sea-clutter is the inability to distinguish buoys, whilst other targets are in general distinguishable? If this is the case, would it not be possible to attach to the buoys themselves some form of transponder as it is in fact used in civil aviation for air-traffic control?

My second point is concerned with sub-clutter visibility techniques. I understand that the most sophisticated methods of distinguishing targets, whose return is well below clutter level, require the use of correlation techniques. Mr. Harrison did not mention any devices utilizing these methods, unless of course he considers 'chirp' radar to be a form of correlation technique.

Would Mr. Harrison give his views on correlation techniques as applied to marine radars?

**The Author** (*in reply*): It is perfectly correct that some method of enhancing the signal return from buoys would solve this problem, but there is a difficulty in the marine use of beacons which does not arise in the civil aviation case. The cloud and collision warning radar used in aircraft all operate on or very near one frequency (9375 Mc/s), and because of the large distances between aircraft, interference between radars is no problem at present. At sea interference is a problem, and to minimize this, marine radars operate at four selected frequencies differing by a total of 180 Mc/s. Thus the beacon has to receive and respond to a wide range of frequencies.



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Beacons of this type are already in operation, but because of the large space and power supply requirements they are unsuitable for fitting on a buoy.

The more sophisticated forms of correlation technique would be too expensive for marine use, but the use of a high speed scanner and integration on the screen was mentioned in the paper. I would not consider 'chirp' radar as a form of correlation technique since it is really a method of achieving a very short pulse.

**Mr. J. R. Beattie:** Mr. Harrison's paper points to the need for reducing vertical beamwidth to reduce rain clutter and suggests the employment of aerials with vertical beamwidths of between 20 and 30 deg. There are a number of large American, Japanese and British high-gain X-band marine aerials with vertical beamwidths of only 15 deg and I would suggest that this is nearer the optimum for 10- and 12-ft X-band aerials.

He did not mention the employment of different cathode-ray tube phosphors and I have found a preference amongst some users for the double layer type to improve target/clutter ratio. Personally I felt the fluoride phosphor was the best operational compromise.

On the subject of logarithmic receivers it should be noted that some earlier American commercial marine equipments offered both log. and linear receivers, though nowadays the log. receiver was offered as an optional item. One British equipment to-day employs a log. receiver, but I consider the linear receiver was a good operational compromise.

I agree with his reservations on circular polarization as experience has shown there is loss of signal strength on some important marine targets, though it should be noted that it was possible to design corner reflectors for circular polarization.

On the controversial subject of 10 and 3 cm it was only fair to state that a British 10-cm commercial equipment has been developed after a number of 10- and 3-cm trials in the 1950s and as a result of the demand from experienced users for equipment on this frequency. These users were more aware of its advantages, a generally better target/clutter ratio in adverse weather, than its limitations and the former was often decisive in choice of equipment. I recently took part in a discussion on 10- and 3-cm comparisons in which a vessel was fitted with a 10-ft X-band aerial and two 12-ft S-band aerials, one vertically and one horizontally polarized. Heavy rain extended in solid belts to 25 miles on the 3-cm equipment and after the switching to short pulse drastic reduction in gain and severe application of variable differentiation it had to be admitted the picture was barely usable. On the 10-cm equipments rain was hardly discernible and the controls needed no adjustment to give a good picture. To clear the small amount of rain only a small adjustment of differentiation was necessary. When it came to detection of small buoys on the Baltic routes at ranges of about 1 to 2 miles in clutter the 10-cm equipments were superior. This ship has again been fitted with 10- and 3-cm equipment as indeed a fair number of the other Baltic ferries are fitted. I consider the A.S.W.E. report on 10 and 3 cm is not unfavourable to 10 cm equipment. Whilst it gives a useful and much needed extension to our operational knowledge further work is necessary to show clearly the relative merits of the two frequencies.

**The Author** (*in reply*): In conditions of rough weather when clutter is encountered, the ship will be a very unstable platform. This association of conditions was doubtless considered when the M.o.T. specification was written, and led to the requirement for adequate performance in a roll of $\pm$ 10 deg. The increase of gain obtained by reducing the vertical beamwidth below 20 deg leads to a loss of targets when the ship rolls, and I cannot agree that this is beneficial.

It is true that a double layer phosphor can lead to some integration in the slower acting layer, but it has been found very tiring for the operator to observe one of these tubes over any length of time.

I cannot agree that a 10-cm equipment gives a better target/clutter ratio than 3 cm. The tests at A.S.W.E. demonstrate that the difference is small, while the photograph of Suez Bay gives direct evidence that the 10-cm set which is 'blind' to rain and sea clutter is equally 'blind' to targets like low sandy coastlines which present a danger to navigation. Air-traffic control experience also shows that a 10-cm set sensitive enough to display the required targets will also display heavy rain clutter.

**Wing Commander G. H. Pennington:** I have always been interested in the relation between the apparent area and extent of rain clutter as seen on the radar and its true extent. I wonder whether Mr. Harrison could say whether, because of reflections increasing the path length, there is any evidence that the apparent size of a rain belt is greater than the actual? I realize that the reflections will be heavily attenuated.

In his original equations, Mr. Harrison implied that we were interested in *targets in the rain belt.* However, in his comparison of 3-cm and 10-cm equipments he pointed to several examples of targets visible *beyond* the clutter. One is obviously therefore concerned in these cases not only with the size of return from rain but with the attenuation of the target signal as seen through a belt of rain. This point did not seem to be brought out in his comparison, and it seems to me a valid one because I think everyone would accept that a 3-cm wavelength is more attenuated by a given rain belt than 10 cm. There seems therefore to be this essential difference between seeing a target *in* clutter and seeing one in the clear *through* clutter. I believe it is of some importance in practice because very often rain is surprisingly localized.

**The Author** (*in reply*): I know of no evidence relating the apparent area of rain clutter to the extent of the rain. Experiments on the propagation of Q-band radiation at A.S.W.E. (Mr. A. L. P. Millwright) showed little correlation between the attenuation over a three mile path and the rain intensity at either end; which supports the second suggestion that the heavy rain areas are localized. Experimental investigation of this point is lacking, but some theoretical calculations by my colleague Mr. E. L. Crouchman indicate that if this is in fact true as suspected, then the greater attenuation losses suffered by 3-cm radiation in traversing such localized areas are not sufficient to overcome the advantages of narrower beam-



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width which result in higher long-range sensitivity in clear weather.

**Mr. C. Powell:** Mr. Harrison showed how targets can be obscured by mal-adjustment of the swept gain control. Would it not be possible to provide a built-in datum or amplitude reference which would give some guidance in setting the control?

**The Author** (*in reply*): The appearance of the screen itself is the best datum for adjustment of the swept gain control. If the clutter is removed completely from the screen as in Fig. 9, the control should be backed off until the higher peaks of clutter are still visible. This usually entails some compromise, but since the effect of the control is clearly visible this is considered a far better indication of correct setting than any other reference.

**Mr. T. G. Clark:** Mr. Harrison has commented upon a proposal that high speed scanning might alleviate the clutter problem. I would like to remark that if the aerial scan rate is increased without a *pro rata* increase in p.r.f. then there will be a substantial reduction in display gain due to a reduced light output from the phosphor of the tube. Normal phosphors show an increase of light output in the afterglow up to a pulse number of ten to fifteen, depending upon the phosphor, and it is usual to have a p.r.f. rotation rate relationship such that the background of the p.p.i. is continuous at the edge of the tube, thus the display gain increases towards the centre of the display. I would not wish lightly to throw away this advantage by increasing aerial rotation rate without increasing p.r.f. It seems that the apparent clutter amplitude is reduced by reducing the display gain, and this will, also, act upon a signal. It may be that signal/clutter ratio is improved but perceptability of far signals will be decreased.

A further approach to improving signal/clutter ratio that has not been remarked upon by Mr. Harrison, possibly because it is uneconomic and of limited application, was in evidence during the recent harbour radar-television-link demonstrations in Southampton Water. In the conditions of fairly light clutter adjacent to Calshot, objects were visible in the clutter via the storage vidicon camera that were not visible on the raw display. However, in stronger clutter it is likely that the clutter would saturate the storage medium to the detriment of signal/clutter ratio. I agree, also, that the storage tube was at a disadvantage in perceiving moving targets and, also, in integrating the small movement of buoys to produce a target many times larger than the real dimensions.

**The Author** (*in reply*): While it is agreed in principle that the display gain will be higher near the centre of the tube, for the reasons given, it must not be forgotten that these signals will be well above limiter level before processing. A reduction in receiver gain will leave them still above or near limiter level. Thus the brightness of each individual signal will be unchanged. However, it will be in effect a section of the aerial radiation diagram at a higher level, so its angular dimension will be decreased, but is still likely to be in excess of one 3-dB beamwidth. Overlap of successive signals will still occur, and since the major part of the screen integration caused by this takes place in the first few pulses, the loss, due to increased aerial speed will not seriously reduce the display gain. In the particular case of a buoy in sea clutter, this loss is far outweighed by the relative improvement of the buoy signal. It should also be remembered that the use of the swept gain control does not affect the gain at longer ranges, so that distant signals are not reduced.

Mr. Clark has stated fairly the disadvantages of the storage vidicon system, and further comment would be superfluous.

**Dr. D. E. N. Davies:** Regarding Mr. Thomason's suggestion that it should be possible to improve the visibility of small buoys in the presence of clutter by the use of active transponders, in addition to the alarming cost of fitment and maintenance there is also the point that in crowded waterways such devices would create large quantities of high-level clutter. Each wide-band transmission would be able to jam radars other than the interrogating set on both the main beams and side lobes. Although it would be theoretically possible to overcome this problem by the use of directional responders such as active van Atta arrays, the cost and complication would clearly be prohibitive.

I was very interested in Mr. Harrison's remarks relating to the reduction of clutter using a very fast scanning aerial with scan-to-scan integration, thus utilizing the fact that the samples of clutter would then be uncorrelated. Mr. Harrison also stated however, that the scanning rate would have to be kept somewhat below the limit of 1 pulse per beamwidth of scan, otherwise distortion of the beam pattern of the aerial would result. I agree that this is so, but if we consider the case of the beam moving through one-half of the power beamwidth in one pulse repetition period then I suggest that an improved directivity pattern can result. This is because the overall directivity pattern for a target at maximum range would be the product of the transmitting and receiving patterns with a relative angular displacement of one 3-dB beamwidth. I have had cause to calculate the form of such patterns for other purposes, and I think that it is not difficult to appreciate that the overall pattern would have half the beamwidth of the one-way directivity pattern.

This would result in a radar whose directivity pattern varied with the range of the targets. It would be the square of the one-way pattern at short range but the beamwidth would reduce with the range, to a minimum of about half of the one-way beamwidth at maximum range, for the particular scanning rate chosen above. This seems to be an interesting advantage of the system rather than an unwanted distortion. It also partially compensates for the fact that bearing resolution for fixed size targets expressed in terms of linear separation, diminishes with range.

**The Author** (*in reply*): The reduction of beamwidth with range is a consequence of high speed scanning which I had not foreseen, and I am grateful to Dr. Davies for pointing this out. It would not however be accompanied by the increase in aerial gain which accompanies a reduction of beamwidth by conventional methods.



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# RADAR SIMULATOR FOR AIR-TRAFFIC CONTROL

The European organization for the safety of air traffic, Eurocontrol, is to install an air-traffic control simulator intended primarily for use in exploring the complex air-traffic control problems which will occur in the coming era of supersonic flight. The equipment is capable of simulating air situations in which as many as 300 aircraft are involved. It will be used at the experimental centre Bretigny, near Chartres, to ascertain the best methods of control to provide optimum safety and efficiency for air traffic.

**The overall system.** The general introduction of high-speed aircraft, both subsonic and supersonic, is daily making the problems of air traffic control more acute. The difficulties which must be dealt with are well illustrated by the example of the *Concord*, an aircraft which will be in the region of Paris some seven minutes after passing over London. Obviously it will be necessary to permit such aircraft to land without delay and to give them clearances for absolutely clear air routes.

The simulator will show the air situation in an area some 2000 km in diameter (corresponding approximately to the whole of Western Europe). Six simulated primary radars and six simulated secondary radars are geographically associated to give coverage over the whole of the exercise area. The simulator can be set up to take account of the performance and flight plans of aircraft, the weather conditions, the layout of navigation aids, the performance of these aids and the big manoeuvres carried out by the simulated aircraft themselves.

The number of aircraft under control during an exercise can be as many as three hundred. In addition the system allows the simulation of a number of intruding aircraft not under direct control. For each aircraft under control the system generates primary and secondary radar signals corresponding precisely to those obtained with the radar systems actually in service, or likely to come into service in the future.

The system allows a control organization based on five sectors to be simulated; these sectors can be based either on the same control centre or on different centres. Each control position is provided with telecommunication facilities, similar to those of actual control centres, giving simulated communications between controllers and pilots, and between controllers of different sectors.

Twenty-eight controllers can take part in an exercise, each having at his disposal a radar display (with all the facilities at present in existence or likely to come into service in the near future) and a flight progress strip indicator.

**Display and data handling equipment.** The display equipment includes plan displays for the air traffic controllers' positions, tabular displays for the pilots' positions and a special display for the supervisor's position which will provide the necessary information for the overall co-ordination of exercises and experiments.

The design concept provides advanced yet highly flexible equipment which can be readily adapted either as self-contained autonomous displays with individual radar sets, or as units within a complex system where high accuracies and high data handling potential are necessary.

**The computer.** The standard computer, which is fitted with input-output channels usually connected for paper or magnetic tape, carries out the real time operation essential in a complex radar simulator system. A cyclic store has been introduced between the displays and the computer to reduce the number of calls for computer data on both the synthetic and tabular display channels. The cyclic unit stores the necessary data from the computer and itself provides the comparatively high data rate output required for the various display channels. This arrangement leaves the computer free from frequent program interruptions and it is only necessary to provide fresh information when data is actually updated, or demands are made from the keyboards. The cyclic store also stores keyboard and 'rolling-ball' information from the display positions for a limited period and feeds this into the computer periodically.

**The radar simulator.** The six radars which are simulated in the system fall into two categories, corresponding to the degree of realism required. The two radars in the first category have different rotation rates and different synchronisms. The remaining four have the same aerial rotation rate, the same beamwidth and the same synchronization. The radar simulators are capable of providing the following presentation from the output data of the computer:

(*a*) Aircraft echoes with or without m.t.i., taking account of the vertical and horizontal coverage diagrams of the radars and of echo level for various beamwidths.

(*b*) Secondary radar responses with facilities for distinguishing five codes.

The simulators take account of the inter-mixing of signals produced by fixed echo generators moving echo generators and the background noise.

**Synthetic-echo generators.** The moving echo generators in the system are based on the technique of film recording and scanning. A flying-spot tube scanner is used. During recording the video signal is applied to the flying spot and the film is sensitized. During playback the intensity of the spot is no longer modulated and the scanning of the tube analyses the film. The video multiplier cell collects the signals given out by the film and amplifies the video signal. The speed of the film transport mechanism is controlled by the aerial rotation.

**Video map generator.** The video map generator employs a flying-spot tube which is scanned as a p.p.i. display by an unmodulated spot. The storage apparatus comprises two diapositive photographic plates which are scanned by the flying spot. A photo-multiplier cell is used to produce the video signals.

The video signals from each photo-multiplier cell are distributed through ten separate outputs. If necessary one signal can be distributed through twenty output channels, the other being restricted to a single output. In the Eurocontrol simulator project one output is employed for the fixed echo map and twenty independent outputs for the electronic map.



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U.D.C. 621.372.43/5: 621.375

# Antisymmetrical Amplifier Networks

By

A. W. KEEN †

**Summary:** The non-reciprocal activity of the general two-terminal-pair (two-port) transmission network may be attributed to an ideal element characterized by an immittance matrix of the type $[0, \pm jr; \mp jr, 0]$, where $r$ is a real number parameter. It is convenient to generalize this element to the complex parameter case, that is, with $z$ (complex) rather than $r$, including the special case of the actively-represented gyrator, for which the parameter is a pure imaginary. This more general element is called the antisymmetrical amplifier. Its properties are discussed, basic circuits for its realization are given and it is shown that similar properties may be distinguished in certain classes of practical electronic circuits.

## 1. Derivation and Properties

It is well known in electrical network theory that a separation of the general (non-reciprocal) transmission network into active and passive parts‡ may be made by dividing its immittance matrix into its Hermitian and anti-Hermitian components. (See Appendix 1 and ref. 1.) In terms of the short-circuit admittance matrix $[Y]$ these components are given by

$$[Y]_H = \tfrac{1}{2}\{[Y]+[Y]^*\}$$

and

$$[Y]_{AH} = \tfrac{1}{2}\{[Y]-[Y]^*\}$$

respectively, where the * denotes the transpose of the complex conjugate of the $[Y]$ matrix. For the case of the two-port network

$$[Y] = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} = \begin{bmatrix} g_{11}+jb_{11} & g_{12}+jb_{12} \\ g_{21}+jb_{21} & g_{22}+jb_{22} \end{bmatrix}$$

and

$$[Y]^* = \begin{bmatrix} \bar{y}_{11} & \bar{y}_{21} \\ \bar{y}_{12} & \bar{y}_{22} \end{bmatrix} = \begin{bmatrix} g_{11}-jb_{11} & g_{21}-jb_{21} \\ g_{12}-jb_{12} & g_{22}-jb_{22} \end{bmatrix}$$

in which the bar superscript denotes the complex conjugate (i.e. reversal of sign of the imaginary part).

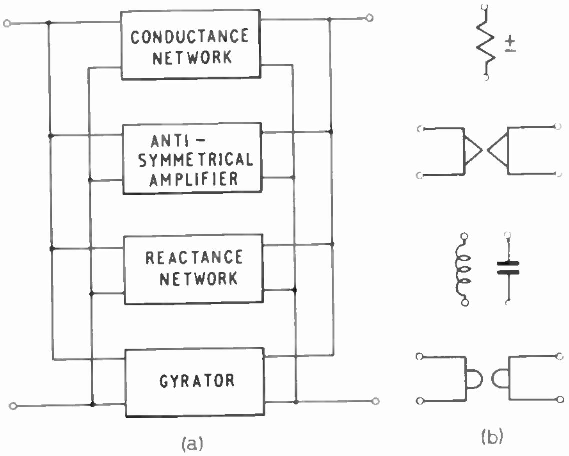


**Fig. 1.**

(*a*) A four-way partition of the general two-port network, introducing the antisymmetrical amplifier.

(*b*) Basic components required for synthesis.

Thence

$$[Y]_H = \begin{bmatrix} g_{11} & \tfrac{1}{2}\{(g_{12}+g_{21})+j(b_{12}-b_{21})\} \\ \tfrac{1}{2}\{(g_{21}+g_{12})+j(b_{21}-b_{12})\} & g_{22} \end{bmatrix}$$

and

$$[Y]_{AH} = \begin{bmatrix} jb_{11} & \tfrac{1}{2}\{(g_{12}-g_{21})+j(b_{12}+b_{21})\} \\ \tfrac{1}{2}\{(g_{21}-g_{12})+j(b_{21}+b_{12})\} & jb_{22} \end{bmatrix}$$

Each of these component matrices may, in turn, be separated into symmetrical (reciprocal) (i.e. having $y_{12} = y_{21}$) and antisymmetrical (non-reciprocal) (i.e. having $y_{12} = -y_{21}$) parts, as follows:

$$[Y]_H = \begin{bmatrix} g_{11} & \tfrac{1}{2}(g_{12}+g_{21}) \\ \tfrac{1}{2}(g_{12}+g_{21}) & g_{22} \end{bmatrix} + \begin{bmatrix} 0 & j\tfrac{1}{2}(b_{12}-b_{21}) \\ -j\tfrac{1}{2}(b_{12}-b_{21}) & 0 \end{bmatrix}$$

and

$$[Y]_{AH} = \begin{bmatrix} jb_{11} & j\tfrac{1}{2}(b_{12}+b_{21}) \\ j\tfrac{1}{2}(b_{12}+b_{21}) & jb_{22} \end{bmatrix} + \begin{bmatrix} 0 & \tfrac{1}{2}(g_{12}-g_{21}) \\ -\tfrac{1}{2}(g_{12}-g_{21}) & 0 \end{bmatrix},$$

resulting in the four-part parallel decomposition illustrated in Fig. 1 (*a*). The basic components needed for synthesis are shown in Fig. 1 (*b*). The symmetrical component of each pair may be represented as a network of positive and negative conductances (e.g. Boucherot networks[2]) in the Hermitian case, or of

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‡ More strictly, into non-conservative and conservative parts (see Appendix 2).



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positive and negative susceptances in the anti-Hermitian case. The antisymmetrical components cannot be represented in terms of conventional two-terminal elements but require the postulation of a pair of basic two-ports, defined by the corresponding matrices. This was first done for the passive case by Tellegen; he called the antisymmetric passive component a gyrator.[3]

The antisymmetrical active element obtained in this way is not available, even approximately, as a single device, but may be resolved into a pair of controlled sources, of equal magnitude but of opposite sign, acting in opposite directions (thereby forming a feedback loop) as shown in Fig. 2. Since $b_{12}$, $b_{21}$ are pure real quantities the controlled source currents are in exact quadrature with the controlling voltages. Such elements have been called transactors.[4, 5] When terminated with admittances $Y_1$, $Y_2$ a feedback loop is completed which has an open-loop transmission of value

$$T = \left(\frac{-\mathrm{j}B}{-Y_2}\right)\cdot\left(\frac{+\mathrm{j}B}{-Y_1}\right) = \frac{B^2}{Y_1 Y_2}, \qquad B = \tfrac{1}{2}(b_{12}-b_{21}).$$

If $Y_1$ and $Y_2$ are real and positive the loop will be a positive feedback loop.

The passive antisymmetrical sub-network, also, is not readily available, either as a single element or in passive form† and is most readily realized in the same form as its active counterpart (see Fig. 2), but with a pair of controlled current sources which are in phase with the controlling voltages. With the same terminations as before the loop transmission is

$$T = \left(\frac{-G}{-Y_2}\right)\cdot\left(\frac{+G}{Y_1}\right) = \frac{-G^2}{Y_1 Y_2}, \qquad G = \tfrac{1}{2}(g_{12}-g_{21}),$$

and the feedback will be negative when the terminations are real.

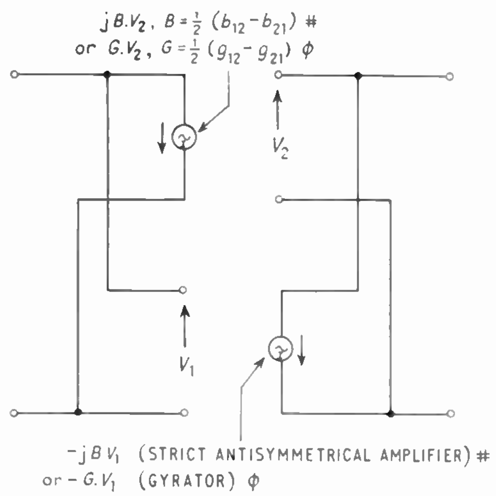


**Fig. 2.** Representation of the antisymmetrical amplifier or gyrator with a pair of controlled sources (transactors).

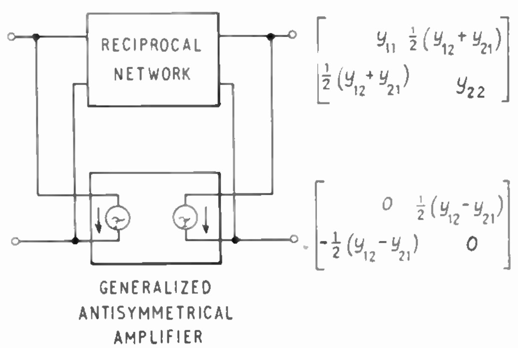


**Fig. 3.** Separation of general two-port network into the generalized antisymmetrical amplifier and a reciprocal network.

It is, therefore, convenient to generalize the antisymmetrical amplifier to have a complex transfer parameter, which reduces to the purely active element at one extreme and the purely passive one at the other. If the two symmetrical (reciprocal) sub-networks are combined together also, one obtains the network partition shown in Fig. 3, consisting of the combination of a reciprocal network, representable by two-terminal elements (possibly including negative conductances), and the generalized antisymmetrical amplifier. Of course, any non-reciprocal active network may be represented in this way and it should be mentioned that Mason[6] has used the dual (i.e. impedance derived) form as a model in feedback amplifier analysis. In this paper attention will be directed to the case where the antisymmetrical part predominates over the symmetrical, so that the entire network is substantially antisymmetrical.

Because the forward and reverse transmittances of the antisymmetrical amplifier are equal in magnitude there is heavy feedback, which is strongly regenerative at one extreme and strongly degenerative at the other. It is, therefore, the impedance transforming, rather than the amplifying (i.e. gain), properties which are of potential usefulness.[7]

In the ideal case (namely, $y_{11} = 0$, $y_{22} = 0$, $y_{12} = -y_{21}$) the input and output admittances are:

$$Y_{\text{in}} = -\frac{(-Y)(+Y)}{Y_L} = \frac{Y^2}{Y_L}, \qquad Y = G + \mathrm{j}B$$

$$Y_{\text{out}} = -\frac{(-Y)(+Y)}{Y_S} = \frac{Y^2}{Y_S}$$

where $Y_L$ and $Y_S$ are the terminating admittances (source and load). Essentially, the driving-point admittance at either side is the inverse of the admit-

† But Hall effect devices, field-effect transistor tetrodes and other types are in course of development.

tance terminating the other. In case $Y = jB$ (active):

$$Y_{in} = -\frac{B^2}{Y_L}, \qquad Y_{out} = -\frac{B^2}{Y_S}$$

which is pure negative inversion, i.e. inversion with respect to a negative constant. In case $Y = G$ (passive):

$$Y_{in} = \frac{G^2}{Y_L}, \qquad Y_{out} = \frac{G^2}{Y_S}$$

which is pure positive inversion.

There are, therefore, two principal fields of application of the (generalized) antisymmetrical amplifier:

(i) Production of negative resistance, by negative inversion of a positive resistance load, in oscillators, $Q$-multipliers, etc.

(ii) Transformation of impedance level, by positive inversion, as in low-output impedance simplifiers, buffer stages, etc.†

If this interpretation is valid one would expect to be able to distinguish antisymmetrical amplifier properties in the action of known practical circuits which have been designed for the purposes just stated, and in Section 3 this will be shown to be so. One might also hope that the fresh insight gained into the action of these circuits may lead to a new design approach and possibly to new circuit configurations.

## 2. Basic Circuits for Antisymmetrical Amplifier Realization

The controlled-source-pair representation of the antisymmetrical amplifier shown in Fig. 2 has the disadvantage of requiring differently-signed sources. This may be overcome by the use of a unit-ratio inverting transformer, but this practice is generally avoided in network synthesis. To explore the possibilities available from a single type of element its transmission properties may be determined for all possible transmission paths. The most readily available controlled source is the admittance-defined transactor:

$$\boldsymbol{Y} = \begin{bmatrix} 0 & 0 \\ Y & 0 \end{bmatrix}$$

which is approximated by the pentode valve or by the transistor, each of which has three access terminal-pairs and may be considered as a transmission network in six ways, any one of which may be obtained from any other by appropriate rotation and/or transposition of the element with respect to its terminations.

The matrices for these different methods of connection are readily obtainable[8] by the corresponding algebraic linear transformations and are as follows:

$$Y_{123} = \begin{bmatrix} 0 & 0 \\ Y & 0 \end{bmatrix}, \qquad Y_{132} = \begin{bmatrix} 0 & 0 \\ -Y & Y \end{bmatrix},$$

$$Y_{213} = \begin{bmatrix} Y & 0 \\ -Y & 0 \end{bmatrix}; \qquad Y_{321} = \begin{bmatrix} 0 & Y \\ 0 & 0 \end{bmatrix}$$

$$Y_{231} = \begin{bmatrix} Y & -Y \\ 0 & 0 \end{bmatrix}, \qquad Y_{312} = \begin{bmatrix} 0 & -Y \\ 0 & Y \end{bmatrix}$$

The subscript numeral sequence denotes the order in which the transactor terminals occur in the transmission path (see Fig. 4).

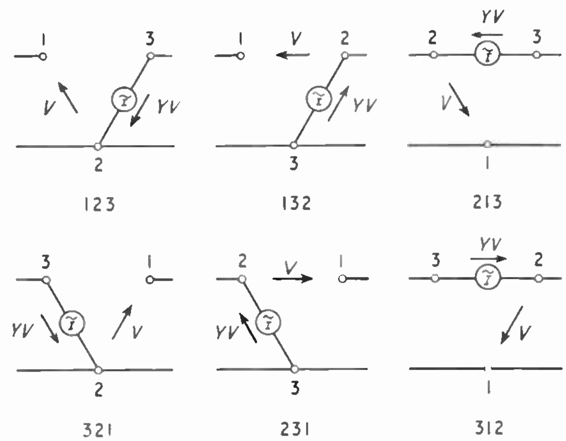


**Fig. 4.** The six orientations of the admittance transactor.

Inspection and comparison of these matrices shows that there are four ways in which they may be added in pairs to produce antisymmetrical resultants, as follows:

$$Y_{123} + Y_{231} = \begin{bmatrix} Y & -Y \\ Y & 0 \end{bmatrix}$$

$$Y_{123} + Y_{312} = \begin{bmatrix} 0 & -Y \\ Y & Y \end{bmatrix}$$

$$Y_{321} + Y_{132} = \begin{bmatrix} 0 & Y \\ -Y & Y \end{bmatrix}$$

$$Y_{321} + Y_{213} = \begin{bmatrix} Y & Y \\ -Y & 0 \end{bmatrix}$$

There are, however, only two distinct circuits, since the circuits corresponding to the third and fourth expressions are reversals of the first and second. It will be noted that in all four cases there is an admittance $Y$ in shunt with one side or the other (see Fig. 5).

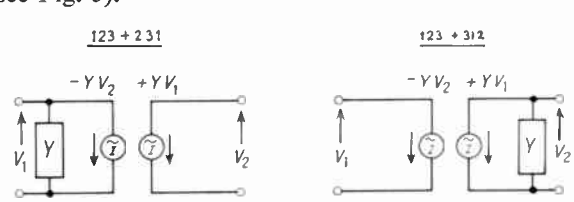


**Fig. 5.** Equivalent circuits of the antisymmetrical transactor pairs.

† A more fundamental application is to inductance simulation by inversion of capacitance. This topic will be dealt with in a separate paper.

The thermionic amplifier valve provides a near equivalent to the admittance transactor in the case $Y$ real, particularly if a pentode type (which may have a negligible $g_a$) is used. Under the assumption of linear negative-grid operation, at frequencies low enough to allow the exclusion of inter-electrode capacitance effects, the admittance matrices for the six paths are as follows:

$$Y_{gca}=\begin{bmatrix}0 & 0\\ g_m & g_a\end{bmatrix},\qquad Y_{gac}=\begin{bmatrix}0 & 0\\ -g_m & (g_m+g_a)\end{bmatrix},$$

$$Y_{cga}=\begin{bmatrix}(g_m+g_a) & -g_a\\ -(g_m+g_a) & g_a\end{bmatrix};\qquad Y_{acg}=\begin{bmatrix}g_a & g_m\\ 0 & 0\end{bmatrix},$$

$$Y_{cag}=\begin{bmatrix}(g_m+g_a) & -g_m\\ 0 & 0\end{bmatrix},\qquad Y_{agc}=\begin{bmatrix}g_a & -(g_m+g_a)\\ -g_a & (g_m+g_a)\end{bmatrix}$$

If $g_a$ is small compared with $g_m$ these approximate closely the matrices for the ideal transactor in the case $Y$ real. Where shunt loading can be tolerated on both sides the antisymmetrical condition can be obtained from a single valve in the *gca* (common cathode) orientation, simply by bridging it (grid-anode) with a conductance of value $\frac{1}{2}g_m$, thus:

$$\begin{bmatrix}0 & 0\\ g_m & g_a\end{bmatrix}+\begin{bmatrix}\frac{1}{2}g_m & -\frac{1}{2}g_m\\ -\frac{1}{2}g_m & \frac{1}{2}g_m\end{bmatrix}=\begin{bmatrix}\frac{1}{2}g_m & -\frac{1}{2}g_m\\ \frac{1}{2}g_m & g_a+\frac{1}{2}g_m\end{bmatrix}$$

It should be mentioned that Shekel[9] has published this network, with negative conductances on each side to cancel the internal end loading, as a realization of the three-terminal gyrator, but such conductances are not conveniently available and the network is of little practical value. A more general case is shown in Fig. 6 (*a*).

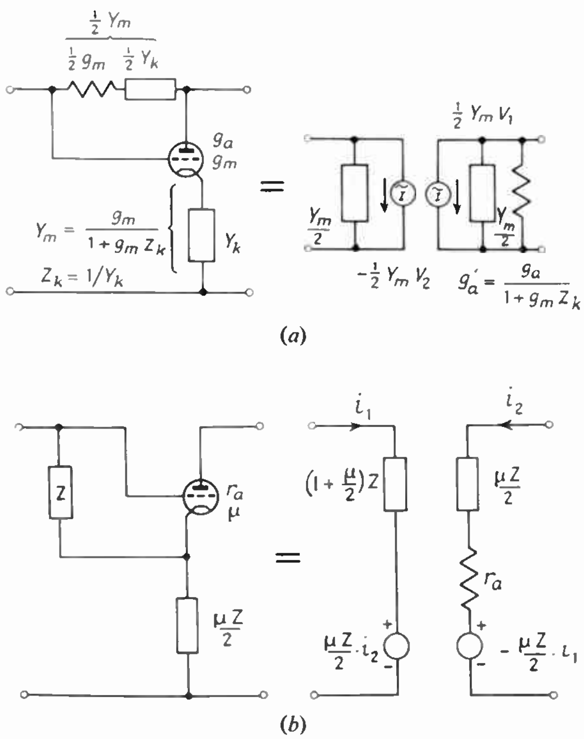


Fig. 6. Single triode antisymmetrical amplifiers.

It will be noted that all of the valve admittance matrices are singular (have zero determinant) and cannot be inverted, so the corresponding $Z$ matrices do not exist. This situation may be remedied by associating an element with the valve in such a way as to make the determinant non-zero. A simple way, which turns out to be useful in the present context, is to shunt grid and cathode with an impedance $Z$ (Fig. 6 (*b*)). The following set of impedance matrices may then be obtained:

$$Z_{gca}=\begin{bmatrix}Z & 0\\ -\mu Z & r_a\end{bmatrix},\qquad Z_{gac}=\begin{bmatrix}r'_a & r_a\\ \mu Z+r_a & r_a\end{bmatrix},$$

$$Z_{cga}=\begin{bmatrix}Z & Z\\ (\mu+1)Z & r'_a\end{bmatrix},\qquad Z_{acg}=\begin{bmatrix}r_a & -\mu Z\\ 0 & Z\end{bmatrix},$$

$$Z_{cag}=\begin{bmatrix}r_a & \mu Z+r_a\\ r_a & r'_a\end{bmatrix},\qquad Z_{agc}=\begin{bmatrix}r'_a & (\mu+1)Z\\ Z & Z\end{bmatrix}$$

where $r'_a=r_a+(\mu+1)Z$. A single-valve antisymmetrical amplifier may then be obtained from the *gca* orientation by inserting an impedance $\pi Z/2$ in series with the cathode lead, with the result:

$$\begin{bmatrix}Z & 0\\ -\mu Z & r_a\end{bmatrix}+\begin{bmatrix}\frac{\mu Z}{2} & \frac{\mu Z}{2}\\ \frac{\mu Z}{2} & \frac{\mu Z}{2}\end{bmatrix}=\begin{bmatrix}\left(1+\frac{\mu}{2}\right)Z & \frac{\mu Z}{2}\\ -\frac{\mu Z}{2} & \left(r_a+\frac{\mu Z}{2}\right)\end{bmatrix}$$

The triode realizations of the $Y$ transactor antisymmetrical pairs are:

$$Y_{gca}+Y_{cag}=\begin{bmatrix}0 & 0\\ g_m & g_a\end{bmatrix}+\begin{bmatrix}(g_m+g_a) & -g_m\\ 0 & 0\end{bmatrix}$$

$$=\underbrace{\begin{bmatrix}(g_m+g_a) & 0\\ 0 & g_a\end{bmatrix}}_{\text{(End loading)}}+\underbrace{\begin{bmatrix}0 & -g_m\\ g_m & 0\end{bmatrix}}_{\text{(Antisymmetrical amplifier)}}$$

and

$$Y_{gca}+Y_{agc}=\begin{bmatrix}0 & 0\\ g_m & g_a\end{bmatrix}+\begin{bmatrix}g_a & -(g_m+g_a)\\ -g_a & g_m+g_a\end{bmatrix}$$

$$=\underbrace{\begin{bmatrix}0 & 0\\ 0 & (g_m+g_a)\end{bmatrix}}_{\text{(End loading)}}+\underbrace{\begin{bmatrix}g_a & -g_a\\ -g_a & g_a\end{bmatrix}}_{\text{(Bridge loading)}}+$$

$$+\underbrace{\begin{bmatrix}0 & -g\\ g_m & 0\end{bmatrix}}_{\text{(Antisymmetrical amplifier)}}$$

and their transposes, the basic circuits of which are shown in Fig. 7, with their equivalents. They have the advantage over the admittance-bridged single-triode form of having shunt loading only on one side. To reduce shunt loading on both sides additional valves are required. A suitable circuit employs cathode-coupled pairs, as shown in Fig. 8. This arrangement provides a good approximation to the ideal gyrator and may be compared with a four-valve circuit which has been published by Sharpe.[10] A three-valve form has been given by Millar.[11]

Addition of the $[Z]$ matrices would require series connection of the corresponding triode circuits which is inconvenient in practice so this possibility will not be considered further.

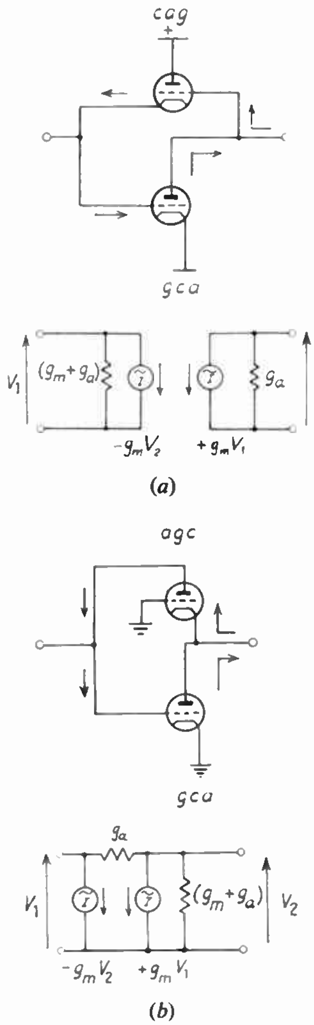

Fig. 7. Antisymmetrical triode pairs.

## 3. Practical Applications of Antisymmetrical Amplifiers

It will now be shown that there are several well-known types of electronic circuit which possess anti-symmetric properties and may be derived from the basic realizations given in Fig. 2. These fall into two classes, corresponding to the two types of feedback: oscillators and low-impedance output stages, the common feature being the heavy degree of feedback incurred by the antisymmetrical condition.

When, in the $Z$-type of single-triode antisymmetrical amplifier (Fig. 6 ($b$)), a pure reactive impedance $jX$

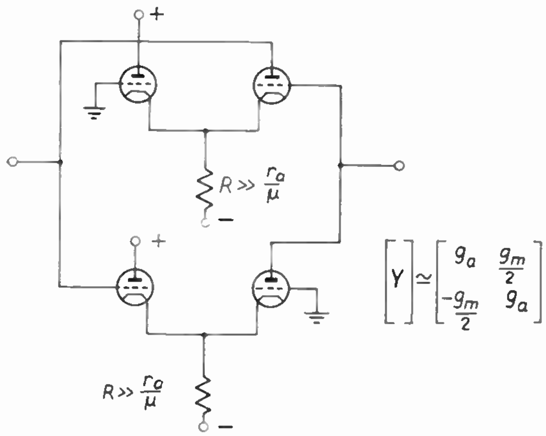

Fig. 8. Antisymmetrical amplifier using cathode-coupled pairs.

is connected across the grid-cathode terminal-pair and a similar impedance of suitable value is inserted in the cathode lead, the input impedance of the amplifier has a negative real part which may be used to neutralize the positive real part of any complex impedance, such as a tuned circuit, shunted across the input terminal-pair, either partly, as in the so-called $Q$-multiplier[12, 13] or completely, as in the class of oscillators which includes the Colpitts and Clapp-Gouriet ($X = 1/\omega C$) and the Hartley ($X = \omega L$).[14, 15]

Adding a reactance of value $k\mu X$ in the cathode lead (Fig. 9 ($a$)) produces the impedance matrix:

$$[Z] = \begin{bmatrix} jX & 0 \\ -j\mu X & r_a \end{bmatrix} + \begin{bmatrix} jk\mu X & jk\mu X \\ jk\mu X & jk\mu X \end{bmatrix}$$

$$= \begin{bmatrix} j(1+k\mu)X & jk\mu X \\ -j(1-k)\mu X & r_a + jk\mu X \end{bmatrix}.$$

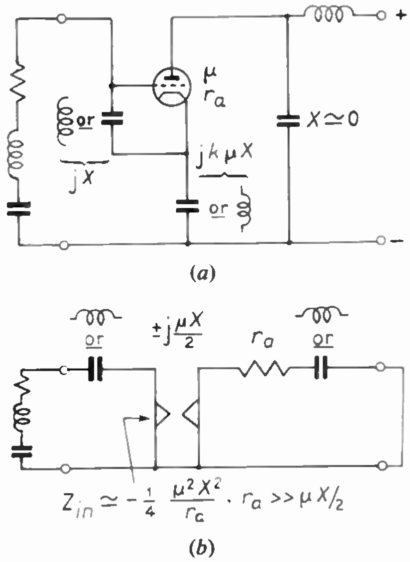

Fig. 9. Antisymmetrical triode oscillators.

When $k = \frac{1}{2}$ this may be interpreted as the equivalent circuit shown in Fig. 9 (*b*), consisting of an antisymmetrical amplifier, with internal loading on both sides. The input impedance at the left-hand side of the amplifier in the general case is

$$Z_{\text{in}} = z_{11} - \frac{z_{12}z_{21}}{z_{22}+Z_L}, \qquad z_{11} = 0$$
$$= -(+\mathrm{j}k\mu X)(-\mathrm{j}[1-k]\mu X)/(r_a+\mathrm{j}k\mu X), \qquad \text{for } Z_L = 0$$
$$= -k(1-k)\mu^2 X^2/(r_a+\mathrm{j}k\mu X)$$
$$\simeq -k(1-k)\mu^2 X^2/r_a, \qquad r_a \gg k\mu X$$

which will have a negative real part in the range $0 < k < 1$, with a maximum at $k = \frac{1}{2}$, which is the strict antisymmetrical condition. The usual analysis for the condition for incidence of oscillation only produces the value $k = 1$, with the condition $0 < k < 1$ for maintenance of oscillation. Usually $r_a$ is so large that $\mathrm{j}k\mu X$ may be neglected and the effect of the antisymmetrical amplifier is to produce from $r_a$ as load a negative input resistance by negative impedance inversion. Of course, the ratio of the two reactances is not critical, and may be varied over a substantial range, but the negative resistance produced is essentially the result of antisymmetrical amplifier action.

A similar approach may be made to the dual of the Colpitts-Hartley class of oscillator, which are sometimes called $Z$ oscillators and have been used with transistors.[16]

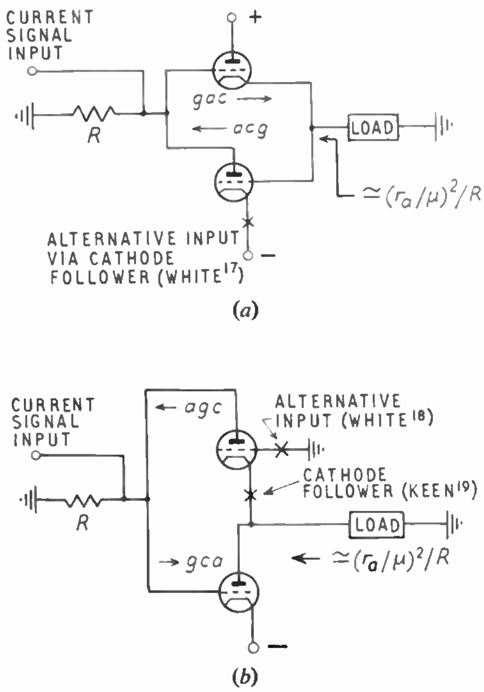


Fig. 10. Antisymmetrical low-impedance output stages.

The usefulness of the antisymmetrical triode pairs for impedance transformation lies in the fact that, because of the high loop gain, they are capable of producing a low output impedance on the side having the internal shunt loading, despite the substantial amount of the latter. For low output impedance application the two basic circuits corresponding to the matrix sums $[Y]_{gca}+[Y]_{agc}$ and $[Y]_{gac}+[Y]_{acg}$ are appropriate (see Fig. 10). The sum matrices are:

$$[Y]_{gca+agc} = \begin{bmatrix} g_a & g_m+g_a \\ g_m-g_a & g_m+g_a \end{bmatrix}$$

$$[Y]_{gac+acg} = \begin{bmatrix} g_a & g_m \\ -g_m & g_m+g_a \end{bmatrix}$$

or, using $g_m = \mu g_a$:

$$[Y]_{gca+agc} = \begin{bmatrix} 1 & -(\mu-1) \\ \mu-1 & \mu+1 \end{bmatrix} g_a$$

$$[Y]_{gac+acg} = \begin{bmatrix} 1 & \mu \\ -\mu & \mu+1 \end{bmatrix} g_a$$

with determinants of value equal approximately to $\mu^2 g_a$. Inverting to obtain the $Z$ matrices:

$$[Z]_{gca+agc} \simeq \frac{r_a}{\mu^2}\begin{bmatrix} \mu+1 & (\mu+1) \\ -(\mu-1) & 1 \end{bmatrix}$$

$$[Z] \simeq \frac{r_a}{\mu^2}\begin{bmatrix} \mu & -\mu \\ \mu & 1 \end{bmatrix}$$

The output impedance for either circuit (with this last approximation), with a load $R$ on the input side is:

$$Z_{\text{out}} = \frac{r_a}{\mu^2} + \frac{(r_a/\mu)^2}{(r_a/\mu)+R} \simeq \frac{(r_a/\mu)^2}{R} \qquad \text{for } R \gg \frac{r_a}{\mu}.$$

It will be clear that a variety of practical circuit variants are feasible. The most important variation is the point at which the signal is injected. If, in the *gac–acg* combination, signal is fed in at the *acg* valve cathode, via a cathode follower (because of the low impedance level there), one obtains the configuration of White's 'super cathode follower',[17] as indicated on Fig. 10 (*a*). Similarly, one may shift the input of the *gca–agc* pair to the *agc* grid, again as shown on the figure, thereby obtaining another circuit due to White.[18] This circuit may be arranged for series or shunt d.c. feed, the former variant being the well-known 'stacked cathode follower'. In another development of the *gca–agc* pair a cathode follower is interpolated between the load and the *agc* cathode.[19]

## 4. Conclusion

It is concluded that the concept of the antisymmetrical amplifier is a significant one in practical electronic circuit design and that it provides an



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appropriate basis for the development of two important classes of network, feedback type $LC$ oscillators and $Q$-multipliers and low-output-impedance amplifier systems.

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## 6. Appendix 1: Hermitian Matrices[1]

Given a matrix $[Y]$ one may obtain from it by interchange of rows and columns a related matrix $[Y]'$ called its transpose. A row vector (i.e. $[1 \times m]$ matrix) transposes into a column vector (a $[m \times 1]$ matrix):

$$[Y] = [y_1 \cdots y_n], \qquad [Y]' = \begin{bmatrix} y_1 \\ \cdots \\ y_n \end{bmatrix}$$

and vice-versa. For a $[2 \times 2]$ matrix:

$$[Y] = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix}, \qquad [Y]' = \begin{bmatrix} y_{11} & y_{21} \\ y_{12} & y_{22} \end{bmatrix}$$

The product of a number of matrices may be transposed by applying the reversal rule, thus:

$$\{[Y_1].[Y_2].[Y_3]\}' = [Y_3]'.[Y_2]'.[Y_1]'$$

Another related matrix, $\overline{[Y]}$, called the (complex) conjugate of $[Y]$ is obtained by reversing the signs of the imaginary parts of all the elements $y_{ij}$ of $[Y]$; e.g. for the $[2 \times 2]$ case:

$$\overline{[Y]} = \begin{bmatrix} \bar{y}_{11} & \bar{y}_{12} \\ \bar{y}_{21} & \bar{y}_{22} \end{bmatrix}$$

where

$$y_{11} = g_{11} + jb_{11}, \text{etc.} \quad \text{and} \quad \bar{y}_{11} = g_{11} - jb_{11}, \text{etc.}$$

The two operations may be carried out in either order, with the same result:

$$\{\overline{[Y]}\}' = \overline{\{[Y]'\}} = [Y]^*$$

which is called the transposed (complex) conjugate of $[Y]$ and denoted by the asterisk superscript; e.g.

$$[Y]^* = \begin{bmatrix} \bar{y}_{11} & \bar{y}_{21} \\ \bar{y}_{12} & \bar{y}_{22} \end{bmatrix}$$

When a square matrix has element values such that $[Y]^* = [Y]$ it is said to be Hermitian.† Alternatively, when the two operations reproduce $[Y]$, but with reversal of sign, i.e. $[Y]^* = -[Y]$, $[Y]$ is said to be anti- or skew-Hermitian. If $[Y]$ is Hermitian, then $j.[Y]$, obtained by multiplying all of the elements of $[Y]$ by $j = \sqrt{(-1)}$, will be anti-Hermitian: conversely, if $[Y]$ is anti-Hermitian, $j.[Y]$ will be Hermitian. For any square matrix $[Y]$ the sum $[Y]+[Y]^*$ is Hermitian, the difference $[Y]-[Y]^*$ is anti-Hermitian, so $[Y]$ may be expressed as the sum of Hermitian and anti-Hermitian components, thus:

$$[Y] = \tfrac{1}{2}\{[Y]+[Y]^*\} + \tfrac{1}{2}\{[Y]-[Y]^*\}$$

as stated at the beginning of the paper. If the anti-Hermitian part is expressed as the product of j into another Hermitian matrix, as follows:

$$\begin{aligned}[Y] &= \tfrac{1}{2}\{[Y]+[Y]^*\} + j \cdot \frac{1}{2j}\{[Y]-[Y]^*\} \\ &= [Y]_{H1} + j.[Y]_{H2}\end{aligned}$$

the matrix analogue of the separation of an individual

† After the French mathematician, C. Hermite (1822–1901).



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admittance element into its real and imaginary parts, $y = g+j.b$, will be obtained. It should be noted, however, that although the principal diagonal elements of $[Y]_{H1}$ and $[Y]_{H2}$ are, like $g$ and $b$, all real, the off-diagonal elements are generally complex. But it will be shown that, like $g$ and $b$, $[Y]_{H1}$ and $[Y]_{H2}$ correspond to the non-reactive and reactive parts of the network, respectively.

## 7. Appendix 2: Hermitian Forms

An expression of the form

$$f = u_1(y_{11}.v_1 + y_{12}.v_2 + \ldots + y_{1n}.v_n) + \ldots + u_m(y_{m1}.v_1 + y_{m2}.v_2 + \ldots + y_{mn}.v_n)$$

in the two variables $u$, $v$ is called a bilinear form. It may be expressed more compactly as the matrix product

$$f = [U]'.[Y].[V]$$

in which $[U] = [u_1 \ldots u_m]$ and $[V] = [v_1 \ldots v_n]$ are column vectors of variables (N.B. $[U]'$ is a row vector) and $[Y]$ a $[m \times n]$ matrix of complex constants. $f$ may be regarded as a $[1 \times 1]$ matrix, $[F]$. In case $[U] = [V]$, so that $m = n$,

$$[F] = \{[\bar{V}]\}'.[Y].[V]$$

whence, by the reversal rule,

$$[F]^* = \overline{[V]'.[Y]'.[\bar{V}]} = [\bar{V}]'.[\bar{Y}]'.[V]$$

so that, if $[Y]$ is Hermitian, $[F]^* = [F]$ and, since $[F]$ is $[1 \times 1]$, $f$ must be pure real; it is called a Hermitian form.

The (complex) power flow into an $n$-port network is

$$P = \sum_{k=1}^{n} (\bar{v}_k . i_k) = [\bar{V}]'.[I]$$

in which

$$i_k = y_{k1}.v_1 + y_{k2}.v_2 + \ldots + y_{kn}.v_n$$

so that

$$P = [\bar{V}]'.[Y].[V]$$

If $[Y]$ is Hermitian, $P$ will be a Hermitian form, hence pure real. A network having a Hermitian short-circuit admittance matrix will therefore be non-reactive (non-conservative) and will consist of purely generative and/or dissipative elements. It will be active or passive according to its determinant being negative or positive. Returning to the result of Section 6, the Hermitian/anti-Hermitian partition of the $[Y]$ matrix will now be seen to correspond to the division of the network into non-reactive and reactive (conservative) parts, respectively.

*Manuscript first received by the Institution on 25th October 1963 and in final form on 8th January 1964. (Paper No. 897.)*

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U.D.C. 535.2:621.373

# Investigation of Relaxation Oscillations in the Output from a Ruby Laser

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*Presented at the Symposium on "Masers and Lasers" in London on 2nd January 1963.*

**Summary:** Measurements have been made on the 'spike' pulses produced by the relaxation oscillation in the output from a laser using a 90 deg ruby crystal. Pulse shape, width and height, the time interval between successive pulses, and the polarization of the pulses have been accurately measured. The rate equations describing the laser oscillation have been solved for operation of the laser close to threshold. The solution predicts a shape for individual pulses of the form $\cosh^{-2}K(t-t_0)$ and this has been verified by the experiments. It is well known that the ruby laser operates in narrow 'filaments'. The solution to the rate equations enabled the diameter and population inversion of the contributing filament to be obtained. Also the measurements on the pulse interval gave the number of filaments active in the crystal at a given time. All the pulses were found to be equally linearly polarized to 99·92%.

**List of Symbols**

| | |
|---|---|
| $n_1$ | number of ions in the active filament in the lower (ground state) energy level |
| $n_2$ | number of ions in the active filament in the upper (excited state) energy level |
| $n$ | population inversion, $n_2-n_1$ |
| $n_{th}$ | threshold inversion |
| $\Delta n$ | $n-n_{th}$ |
| $D$ | maximum value of $\Delta n$ before the emission of a light pulse |
| $q$ | number of photons in the laser cavity |
| $q_0$ | number of photons in the laser cavity at the peak of a light pulse, $D^2/4n_{th}$ |
| $I$ | light output from the laser, $q/t_c$ (photons/s) |
| $I_0$ | light output at the peak of a light pulse, $D^2B/4$ (photons/s) |
| $K$ | constant, $DB/2$ (seconds$^{-1}$) |
| $k$ | constant (seconds$^{-1}$) |
| $\Delta t$ | width of a light pulse at the half-power points (seconds) |
| $t_0$ | time at which the peak of a light pulse occurs (seconds) |
| $W$ | rate/ground state ion at which ground state ions are pumped to the upper level (seconds$^{-1}$) |
| $W_{th}$ | pump rate at the threshold of laser oscillation (seconds$^{-1}$) |
| $\tau$ | spontaneous emission time (seconds) |
| $t_c$ | relaxation time of the laser cavity for the active filament (seconds) |
| $B$ | stimulated emission coefficient (seconds$^{-1}$) |
| $p_m$ | number of modes coupling to the spontaneous fluorescence line |
| $\nu$ | light frequency (c/s) |
| $\Delta\nu$ | fluorescence bandwidth (c/s or cm$^{-1}$) |
| $r$ | refractive index of ruby |
| $c$ | velocity of light *in vacuo* (cm/s) |
| $V$ | volume of active filament (cm$^3$) |
| $d$ | diameter of active filament (cm) |
| $\rho$ | concentration of chromium ions in the laser (ions/cm$^3$) |
| $T_F$ | time interval between successive pulses for one filament (seconds) |
| $T_{AV}$ | average time interval between pulses (seconds) |
| $N$ | number of filaments active in the time of one measurement |

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## 1. Introduction

It is well known that the light from a solid-state laser is not emitted continuously during the pump light pulse but is broken into short pulses or 'spikes' of light (see Fig. 1(*a*) for example). These spike pulses result from a relaxation oscillation in the processes producing the coherent light.[1,2] This paper presents the results of a systematic investigation of the spike pulses for one pulsed ruby laser. Measurements have been made of pulse shape, width and height, of the time interval between successive pulses and of the



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polarization of the light during a pulse. The results are compared with a perturbation analysis of the equations describing the relaxation oscillation. The ruby laser does not act coherently across the whole of its cross-section during any one spike pulse but in a narrow strip or 'filament' parallel to the laser axis. A number of these filaments may become active during the pumping pulse. Comparison of the experimental results with theory is limited by the random operation of these filaments. On the other hand the results, when combined with the theoretical analysis, give information on the size and population inversion of the particular filament active in any one pulse.

## 2. Theory

The basic equations which describe the operation of a ruby laser may be written[3] as

$$\frac{dn}{dt} = 2Wn_1 - \frac{2n_2}{\tau} - 2Bnq \qquad \ldots\ldots(1)$$

$$\frac{dq}{dt} = \frac{n_2}{p_m\tau} - \frac{q}{t_c} + Bnq \qquad \ldots\ldots(2)$$

For a Lorentzian-shaped spontaneous emission line and for a laser operating near threshold, it is well known[4] that

$$B = \frac{1}{p_m\tau}, \qquad p_m = \frac{4\pi^2\nu^2\Delta\nu\, r^3 V}{c^3}$$

The population inversion required for the laser to start emitting, the threshold inversion, is given by

$$n_{th} = \frac{1}{Bt_c}$$

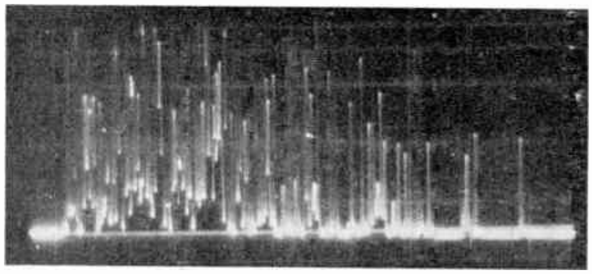
To maintain the laser at the threshold of oscillation, the pump rate must be sufficient to keep in the upper energy level one-half of the chromium ions in the active filament in addition to maintaining the threshold inversion. Therefore the threshold pump rate is

$$W_{th} = \frac{\rho V + n_{th}}{\rho V - n_{th}} \cdot \frac{1}{\tau}$$

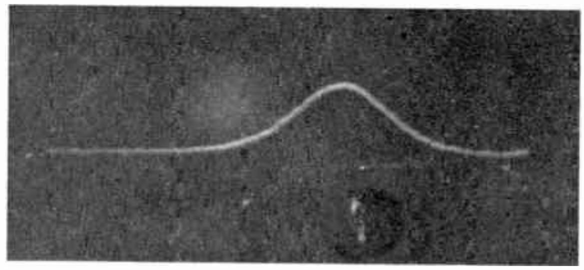
Provided $\rho V \gg n_{th}$, $W_{th}$ can be written as

$$W_{th} = 1/\tau$$

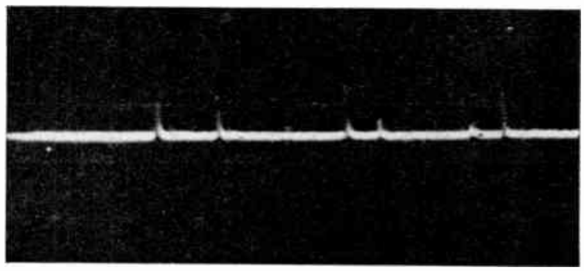
These equations cannot be solved exactly and resort must be made to either numerical solution, or approximation. Numerical solutions using both digital[2,5-12] and analogue[12,13] computers have shown that the light should be emitted from a laser in a regular train of pulses with the population inversion swinging about the threshold value. The amplitudes of successive pulses will, in general, be damped.[9] On the other hand Makhov[9,14,15] has shown that the introduction of a term to describe the cross-coupling between the axial and non-axial modes of the cavity can lead to an undamped solution. These numerical solutions have been employed in the determination of values for the time interval between successive pulses and for the amount of damping.



(*a*) The total laser action, with time reading from left to right. The sweep is 50 μs/cm and the start of the sweep is delayed 500 μs after the start of the pump flash.



(*b*) A single light pulse from the laser. The sweep is 5 μs from end to end.



(*c*) The total laser action when the laser is pumped very close to threshold. The sweep rate is 20 μs/cm.

**Fig. 1.** C.r.o. photographs of the light from a ruby laser.

The measurements to be discussed below have been principally concentrated on measuring the shape of individual pulses. As will be discussed later, the shape was found to be of the form $\cosh^{-2}K(t-t_0)$. Equations (1) and (2) have been examined with a view to predicting the shape of a pulse, and it has been found that when a perturbation treatment is applied a shape is obtained which agrees with experiment. This approximate solution is now described.



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If the laser is operated near to threshold the first two terms of the right-hand side of eqn. (1) are very nearly the same size, with $2Wn_1$ being perhaps 3% larger than $2n_2/\tau$. From the results to be given later in Section 4.2, for the duration of the light pulse (about 1 µs) the third term can be shown to exceed the difference of the first two terms by at least a factor of one hundred. Therefore for the duration of the light pulse it is possible to neglect the first two terms of eqn. (1). This is obviously a first-order approximation and is only valid for operation close to threshold. The first term of the right-hand side of eqn. (2) is $p_m$ times smaller than these terms and since $p_m$ is at least $10^9$ can also be neglected. The only reason for including this term is to indicate the means by which the laser oscillation is started.

The population inversion swings about $n_{th}$ so that eqns. (1) and (2) are best written in terms of

$$\Delta n = n - n_{th}:$$

$$\frac{d(\Delta n)}{dt} = -2B(\Delta n + n_{th})q \qquad \ldots\ldots(3)$$

$$\frac{dq}{dt} = +B\Delta n q \qquad \ldots\ldots(4)$$

The main assumption is now made, namely that

$$\Delta n \ll n_{th}.$$

This assumption will be justified in Section 4.2. Equation (3) can then be written

$$\frac{d(\Delta n)}{dt} = -2Bn_{th}q \qquad \ldots\ldots(5)$$

The following solutions to eqns. (4) and (5) can be obtained:

$$q = q_0 \cosh^{-2} K(t-t_0)$$

$$\Delta n = -D \tanh K(t-t_0)$$

where

$$q_0 = \frac{D^2}{4n_{th}}, \qquad K = \frac{DB}{2}$$

$D$ is the maximum value $\Delta n$ attains before the emission of a light pulse from the laser.

The light output from the laser is thus given by

$$I = \frac{q}{t_c} = I_0 \cosh^{-2}\frac{DB}{2}(t-t_0) \qquad \ldots\ldots(6)$$

where

$$I_0 = \frac{D^2 B}{4}$$

The output light pulse should therefore have a symmetrical shape with a full width between the half-power points of

$$\Delta t = \frac{4\times 0{\cdot}89}{DB} \qquad \ldots\ldots(7)$$

Then, from the definition of $B$

$$\frac{D}{V} = \frac{14{\cdot}24\pi^2\nu^2\Delta\nu\, r^3\tau}{\Delta t c^3} \qquad \ldots\ldots(8)$$

Using the parameters for ruby at room temperature of:

$$\tau = 3\times 10^{-3}\ \text{s}, \qquad \nu = 4{\cdot}32\times 10^{14}\ \text{c/s}$$

$$\Delta\nu = 3{\cdot}3\times 10^{11}\ \text{c/s}\ (11\ \text{cm}^{-1})$$

$$c = 3{\cdot}00\times 10^{10}\ \text{cm/s} \quad \text{and} \quad r = 1{\cdot}764$$

Equation (8) gives

$$\frac{D}{V} = \frac{5{\cdot}29\times 10^9}{\Delta t}\ \text{ions/cm}^3 \qquad \ldots\ldots(9)$$

Equation (9) provides a method of measuring, for the active filament, the amount by which the population inversion swings about the threshold inversion.

It is also possible to measure the volume of the filament from the height of the light pulse, $I_0$:

$$V = I_0 \Delta t^2 \times 2{\cdot}13\times 10^{-10}\ \text{cm}^3 \qquad \ldots\ldots(10)$$

$I_0$ is measured in units of photons/s. Since the lengths of all filaments are identical, 5·08 cm, it is more interesting to calculate the filament diameter, $d$, assuming a cylindrical shape:

$$d = \sqrt{I_0}\,.\,\Delta t \times 7{\cdot}3\times 10^{-6}\ \text{cm} \qquad \ldots\ldots(11)$$

Also, it is simple to measure the time interval between successive light pulses. Unfortunately the only prediction[3,6–8] of this time interval has been based on a linearized solution of eqns. (1) and (2) which does not agree with experimental results. The solution yields a sinusoidal variation of $\Delta n$ and $q$ and gives a time interval of

$$T_F = 2\pi(BV\rho[W - W_{th}])^{-\frac{1}{2}} \qquad \ldots\ldots(12)$$

It should be noted that there are some errors in the equations used in certain of the earlier calculations[6,7] of the time interval but that these do not alter eqn. (12).

Equation (12) gives the time interval for a single filament of the laser. In practice several filaments will be active over the same measurement time to yield a complex pulse pattern from their overlapping pulse trains. The average time interval is then obviously given by

$$T_{AV} = \frac{T_F}{N} \qquad \ldots\ldots(13)$$

## 3. Experimental

The ruby used in the experiments was a 90 deg crystal (this is a crystal with the optic ($c$) axis at 90 deg to the rod axis), 2 in long and with flat ends.† The semi-transparent silver coating was made to have a transmission of approximately 4%. The crystal was

† The crystal was supplied by The A. Meller Company, Providence, Rhode Island, U.S.A.



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pumped at room temperature by six linear flash tubes placed symmetrically about the rod. The light from the laser was detected, after suitable filtration, by means of a vacuum photocell and wide-band amplifier. The bandwidth of the photocell, amplifier and c.r.o. was 1 c/s to 15 Mc/s and was sufficiently wide for negligible distortion of the pulses to be introduced.

In the experiments in which the polarization of the light was measured the experimental arrangement of Fig. 2 was employed. The light patterns from the two photocell channels were compared on a double-beam c.r.o. Much care was needed to ensure that the two patterns corresponded exactly pulse by pulse. For this reason the light was scattered by a diffuse screen before detection in order to mix light from different parts of the laser. The two collimation holes were introduced to remove the angular mode laser light, the proportion of which can vary from pulse to pulse. To ensure that the glass plate, which produced the reference light beam, did not introduce errors in the results its axis was set parallel to the ruby optic axis.

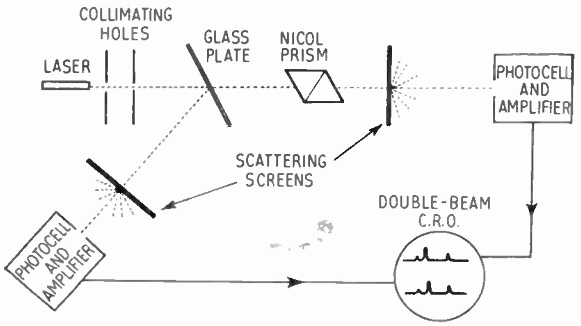


**Fig. 2.** Experimental arrangement for the polarization experiments.

## 4. Results

The relaxation oscillation pulses were photographed on the c.r.o. with a sweep speed between 0·5 μs/cm and 50 μs/cm.

The rate of pumping obtained from the flash tubes was related to the rate required at threshold by measuring the pump light intensity at the completion of the train of oscillation pulses. In this way the pump rates used were shown to lie in the range $1{\cdot}005 \times W_{th}$ to $2 \times W_{th}$.

### 4.1. *Pulse Shape*

For the measurements discussed in this section and in Sections 4.2 and 4.3 the pump rate was between $1{\cdot}01 \times W_{th}$ and $1{\cdot}03 \times W_{th}$. Photographs were then obtained, using the highest sweep speed, of single pulses which did not overlap with any other pulses. A typical trace is shown in Fig. 1(*b*). Each side of a pulse was analysed separately for shape and, as shown in Fig. 3, excellent agreement was found with the shape predicted by eqn. (6). In all, ten pulses have been tested in this way and each fitted the predicted shape. In order to keep Fig. 3 clear only three pulse sides are shown, but these are typical of the accuracy with which the shape is fitted. It was necessary to test each side separately since the pulses were slightly asymmetric. The asymmetry resulted in the fitted width on the fall side of the pulse being 5–10% greater than that on the rise side. The pulse shape has also been measured by Berkley and Wolga[16] who found

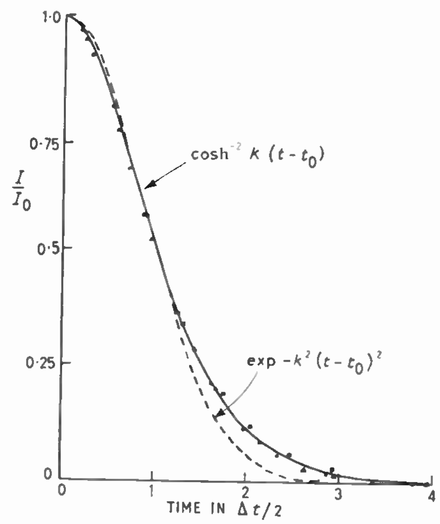


**Fig. 3.** Test of pulse shape for three light pulses.

that a Gaussian amplitude corresponded reasonably well to the pulse shapes they observed. The Gaussian:

$$I = I_0 \exp[-k^2(t-t_0)^2]$$

has been plotted in Fig. 3 and it is seen that it does not describe the experimental shape as accurately as $I_0 \cosh^{-2} K(t-t_0)$.

### 4.2. *Pulse Width*

The pulse widths measured for many pulses ranged between 0·375 and 2·25 μs with a most common width of 0·925 μs. Equation (9) gives values for the amplitude of the population inversion swing in the laser of between $2{\cdot}4 \times 10^{15}$ and $1{\cdot}4 \times 10^{16}$ ions/cm³. It is interesting to compare these values with the concentration of $Cr^{3+}$ ions in the crystal, $1{\cdot}58 \times 10^{19}$ ions/cm³. It is thus found that the swing is only $1{\cdot}5 \times 10^{-4}$ to $8{\cdot}9 \times 10^{-4}$ of the total ions present.

It is necessary to compare the inversion swing to the threshold inversion. The threshold inversion can be obtained from $t_c$, which is determined by the losses in the laser cavity which consist of scattering in the ruby to give[17] a loss of 40% for a double pass of the rod and loss by imperfect reflexion at the two silver surfaces



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of 5% and 15% for the opaque and semi-transparent ends respectively. It is easily shown that $t_c$ is $0{\cdot}99 \times 10^{-9}$s and that the threshold inversion density, $n_{th}/V$ is $1{\cdot}5 \times 10^{18}/\text{cm}^3$. Then by comparision to the values given earlier it is seen that the assumption $\Delta n \ll n_{th}$ is valid.

### 4.3. *Diameter of the Active Filament*

Measurements of pulse height and width combined with eqn. (11) can give the diameter of the active filament. The complete calculation requires an accurate calibration of the photocell sensitivity and since only an approximate figure has so far been obtained the absolute values of the diameters must be treated with caution. The measurements show that the filament diameters vary between 0·18 and 0·97mm. These values are in satisfactory agreement with direct measurements of the filament diameter of 0·15mm obtained from photographs of the end of the laser crystal during oscillation.

### 4.4. *Time Interval between Pulses*

The full range of pump rates available was used for these measurements. The time interval between successive pulses was found to decrease with increase in pump energy. For a given pump pulse the time interval was also found to decrease rapidly when the laser action began and then to increase again at the end of the action. For a pump pulse with which oscillation started at $2W_{th}$ the average time interval was initially 8·0μs, decreased to 2·5μs and then increased to 15μs. The time interval calculated using eqn. (12) with a pump rate of $2W_{th}$ lies within the range of these values. The large changes in the time interval within one pump pulse cannot be explained by this equation. The effect can be explained in terms of different numbers of filaments active at the different times. Equation (13) shows that the inverse of the time interval is related to the number of filaments. If the effect of varying pump power is removed via eqn. (12) by multiplying the measured interval by $\sqrt{(W-W_{th})}$, then $[T\sqrt{(W-W_{th})}]^{-1}$ should be proportional to the number of filaments. In Fig. 4, $[T\sqrt{(W-W_{th})}]^{-1}$ has been plotted (this graph represents the smoothed curve obtained from a widely scattered set of time intervals, the distribution of the measured time intervals are given by the vertical lines) and shows that during laser action the number of filaments increases to a maximum and then decreases again. Statz *et al.*[7] have suggested that this effect can be explained by the focusing effect of the ruby rod on the pump light.

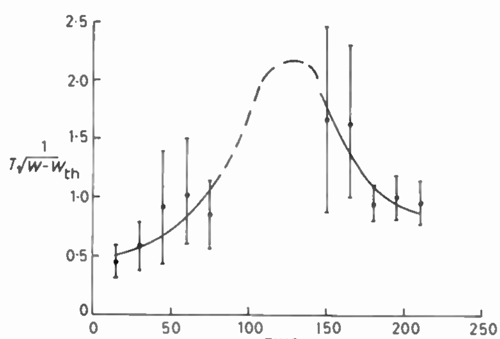


Fig. 4. Variation of the time interval between pulses with the duration of laser action. The graph is plotted as $[T\sqrt{(W-W_{th})}]^{-1}$ and may be considered to show the relative numbers of filaments active at any time. The points shown are averages of many intervals; the range of the intervals used for the average are shown by the vertical lines.

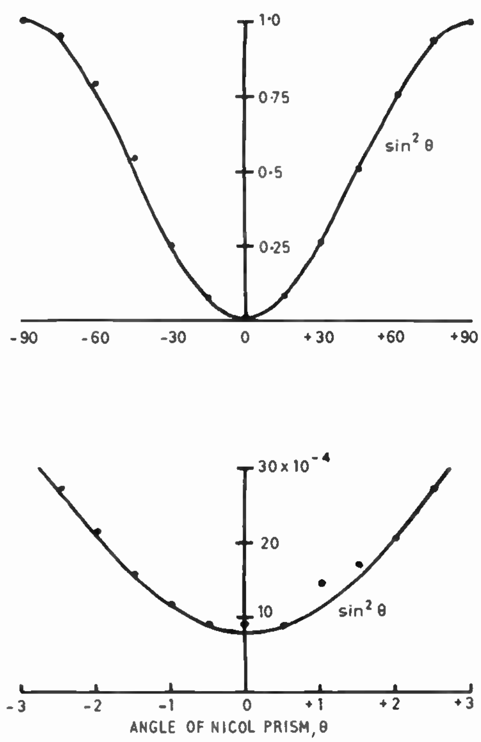


Fig. 5. Test of the polarization of the light from the laser.

### 4.5. *Polarization*

As proved by the earlier experiments of Nelson and Collins[18, 19] the light from a 90deg ruby crystal is nearly completely linearly polarized. The light from a 60deg ruby crystal is also linearly polarized.[18–20] Figure 5 shows that the variation of the light transmitted by the Nicol prism agrees closely with the expected $\sin^2\theta$ curve. Some light is transmitted however when the Nicol prism and the crystal are completely crossed and it amounts to $8 \times 10^{-4}$ of the light incident on the prism. In agreement with this result Collins and Nelson[18] found that the transmitted light was less than $10^{-3}$ of the incident light. The time



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variation of the transmitted light still shows complete correspondence, pulse by pulse, with the reference light reflected from the glass plate. This proves that all regions of the crystal emit light with the same polarization. The small amount of residual light in the crossed position is most likely due to the crystal optic axis having slightly different angles around the rod axis for different positions in the rod. The residual light is not due to imperfections in the Nicol prism since tests of it crossed with another Nicol prism have yielded residual light ratios of $5 \times 10^{-6}$. The full range of pump rates was used for this experiment.

### 4.6. *Pump Level Very Close to Threshold*

If the energy fed to the pumping flash tubes is adjusted so that the laser crystal is very close to the threshold of oscillation then only one or two of the most efficient filaments are excited. The pump rate would be typically about $1{\cdot}005 \times W_{th}$. It is then possible to separate the relaxation oscillations of single filaments. Figure 1(*c*) shows a c.r.o. photograph obtained under these conditions. It is seen that there are two, or possibly three, independent trains of light pulses. Decaying pulse trains of this type have also been observed by other workers both for ruby lasers[3, 21, 22] and for $CaF_2{:}U^{3+}$ and $BaF_2 : U^{3+}$ lasers.[10, 11, 23]

## 5. Conclusions

A detailed study has been made of the pulses produced in the relaxation oscillation of a laser employing a 90deg ruby crystal. All of the parameters which can be calculated by any simple theory have been measured with sufficient accuracy to test the theory. A limitation of this type of measurement is that in order to prevent the individual pulses overlapping the pump power must be set close to threshold.

One fortunate consequence of the necessity of having to pump close to threshold is that the rate equations for laser oscillation prove to be soluble with the single assumption that the amplitude of the fluctuations in the population inversion is small compared to the threshold inversion. The results obtained with the ruby laser show that this assumption is well justified. The solution obtained is valid for any number of cavity modes operating simultaneously provided all modes have the same loss.[8] This condition of equal loss will hold for all axial modes of the same filament but not for different transverse modes. The effect of a number of transverse modes excited simultaneously has not been investigated. Another familiar problem which occurs with ruby lasers is that the temperature of the crystal increases as the pumping continues. This does not affect the theory described provided the alteration of the stimulated emission coefficient with temperature is taken into account. This change is small and has been neglected in making interpretations from the theory.

The solution obtained from the rate equations predicts that the individual light pulses should have a shape of $\cosh^{-2} DB/2(t-t_0)$. Experimental measurements of the shape show that it agrees closely with that predicted. The width of the pulse can be used to determine the excess population inversion, $D$, before the emission of the light pulse. The experimental measurements thus yield values of $D$ between $2{\cdot}4 \times 10^{15}$ and $1{\cdot}4 \times 10^{16}$ ions/cm$^3$. The theory can be extended to include the peak intensity of the light pulse and it can then be used to determine the diameter of the filament which produces the pulse. Obviously implicit in this calculation lies the assumption that only a single filament contributes to any given pulse. Filament diameters between 0·18 and 0·97 mm have been calculated in this way and these agree fairly well with those measured directly by near-field photographs.

The time interval between pulses has also been measured and compared to a linearized solution of the rate equations. Agreement with this solution has been limited by the operation of a large number of filaments. The results have been analysed to show that more filaments are active in the middle of the oscillation than at the beginning or end.

Finally the degree of polarization of the laser light has been measured to great accuracy. The measurements show that the light is linearly polarized to 99·92% and that the light from all pulses is equally polarized.

In conclusion it can be stated that a preliminary investigation has been made of a number of parameters which can be measured for a ruby laser under the conditions of a relaxation oscillation. Although they do not provide a complete understanding of the relaxation oscillation they do provide a starting point from which a more complete model could be constructed.

## 6. Acknowledgments

Grateful acknowledgment is made to Bristol Aircraft Ltd. for supporting this work and particularly to Mr. R. O. R. Chisholm for his unfailing interest and encouragement.

We would like to thank Dr. J. M. Burch of the National Physical Laboratory for pointing out the necessity to align the crystal optic axis and the glass plate in the polarization experiments.

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*Manuscript first received by the Institution on 22nd June 1963 and in final form on 27th February 1964. (Paper No. 898.)*

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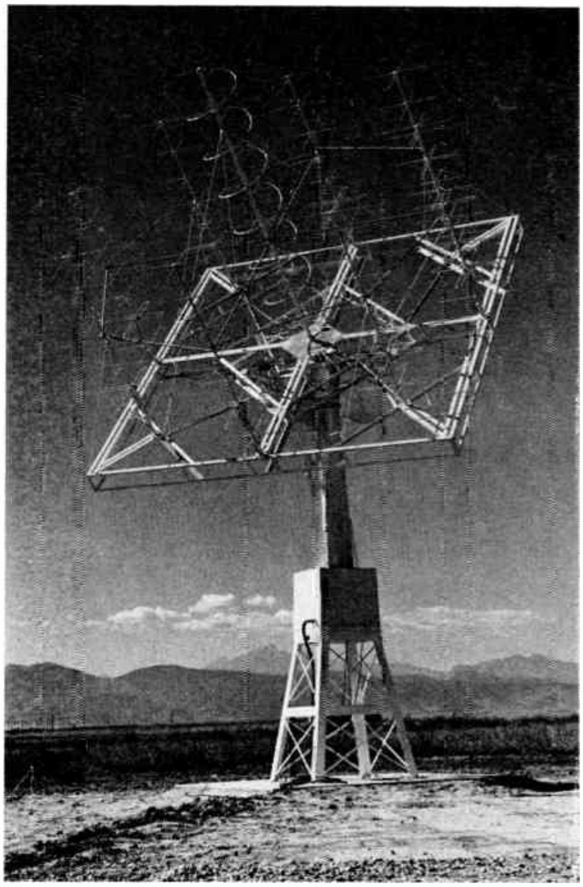
# "TOPSI"—A Forthcoming Ionospheric Research Station

The Central Radio Propagation Laboratory of the U.S. National Bureau of Standards is currently adding three new installations to its world-wide network of radio research stations. A satellite telemetry station, recently completed at Boulder, Colorado, will receive and control information from TOPSI, a topside sounder satellite to be launched by the U.S. National Aeronautics and Space Administration as part of an international programme to probe the ionosphere from above. Two ionospheric vertical sounding stations, one in California and the other in Virginia, will make ground-based soundings of the ionosphere. Data obtained by these two stations will be used to determine the suitability of rocket and satellite launching times for the Pacific Missile Range and the N.A.S.A. Wallops Island Range, respectively.

At frequencies below about 30 Mc/s the ionized layers are studied at many ground stations around the world by means of ionospheric radio soundings using a sweep-frequency device similar to f.m. radar. However, it is in the nature of this technique that it provides no data above the peak of the reflecting layer. Studying the ionosphere from above with this technique was shown to be feasible by a series of topside sounder rocket flights and has been employed systematically since the launching of the *Alouette* topside sounder satellite.

Space scientists will communicate with the second topside sounder satellite TOPSI by means of a chain of telemetry stations, co-ordinated by one such station recently completed at the Bureau's Gunbarrel Hill field station, near Boulder, Colorado. The station consists of an antenna array and electronic equipment in an adjoining small house. Eight Yagi antennas form a highly efficient and directional array to receive the signals telemetered from the satellite. The ninth antenna, a helix mounted in the centre of the Yagi array, is used to transmit commands to the satellite. The entire antenna array is mounted on a movable platform which can be tilted and rotated to follow the satellite as it passes in orbit over the station. The Bureau's ionospheric research scientists expect the station to receive ionospheric data from the TOPSI satellite as it follows the 5000-mile-long section of its orbital path above the Boulder horizon. The station has already been tested successfully on several occasions, receiving data from the *Alouette* satellite on command.

Rocket and satellite experiments in the upper atmosphere and near-space, and the launchings carried out in missile development, require extensive use of ground-based methods by which certain properties of the upper atmosphere may be measured continuously. One such important method is that of sweep-frequency ionospheric sounding, from which the ionization of the high atmosphere (100–300 km) may be monitored. Many rocket launchings must await suitable conditions for optimum experimental results, and these conditions may be detected and even predicted by an ionospheric sounding station.



The new steerable array at the U.S. National Bureau of Standards field station at Gunbarrel Hill, Colorado, which is used to receive telemetered data from satellites.

One of the two new field stations is located at the U.S. Navy's Pacific Missile Range, Point Arguello, California, as a result of an agreement between the Bureau and the Navy. The other, at Wallops Island (Virginia) launching site, followed a similar agreement with the U.S. National Aeronautics and Space Administration.

The Bureau's Vertical Sounding Research Section operates the two stations and undertakes data analysis and consultation for many of the range-using agencies. Developments of improved instrumentation as well as ionospheric studies are constantly carried out at the stations.



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U.D.C. 681.32: 621.318.57

# The Charge Storage Diode as a Logic Element

*By*

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AND

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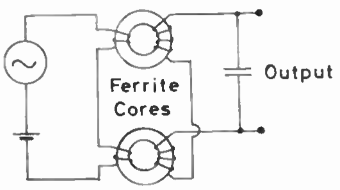
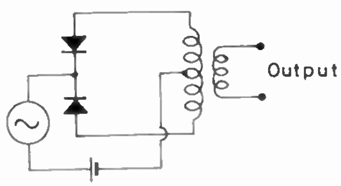
**Summary:** The application of the carrier-storage frequency divider as a parametron-type logical element is described. An analysis of the small-signal phase locking process is presented and a number of basic logical circuits, which have been tested experimentally, are given. The primary limitation to the speed of operation of the device as a logical element is seen to be the drive frequency.

## 1. Introduction

The carrier-storage frequency divider[1] is a second-subharmonic phase-locked oscillator with properties and applications similar to the parametron.[2, 3] In the parametron, the subharmonic oscillation is produced by parametrically pumping one of the circuit reactances at twice the L-C resonant frequency. The reactive element may be either the inductor, using the non-linear magnetization characteristic of a ferrite, or the barrier capacitance of a junction diode, as shown in Figs. 1(*a*) and 1(*b*) respectively. The sub-harmonic output of the carrier-storage frequency divider (Fig. 2) is due to the variation in the recovery delay of a charge-storage diode on alternate cycles of the drive frequency. The output voltage waveform is non-sinusoidal and may contain up to 80 per cent second-subharmonic component depending on the bias and tuning conditions. It is found that tuning is not critical and in this respect the circuit may be superior to the parametric devices which are basically resonant systems.




(*a*) Variable inductance (ferrite core parametron)




(*b*) Variable capacitance (barrier capacitance parametron).

**Fig. 1.** Parametric subharmonic oscillator.

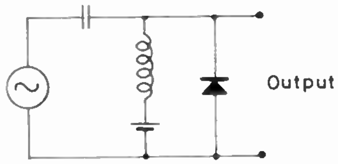



**Fig. 2.** The carrier-storage frequency divider.

Since the second-subharmonic output is phase locked to the input, it can be in one of two distinct states differing in phase by $2\pi$ radians of the drive angular frequency as shown in Fig. 3. These two states may be used to represent the binary digits 0 and 1.[4] Once in a particular state, the oscillator cannot change to the other state unless disturbed by a signal of comparable magnitude. The circuit may thus be said to possess static memory. In a computing system, it is also necessary to be able to transfer information from one element to another. This may be done, with the carrier-storage frequency divider, by

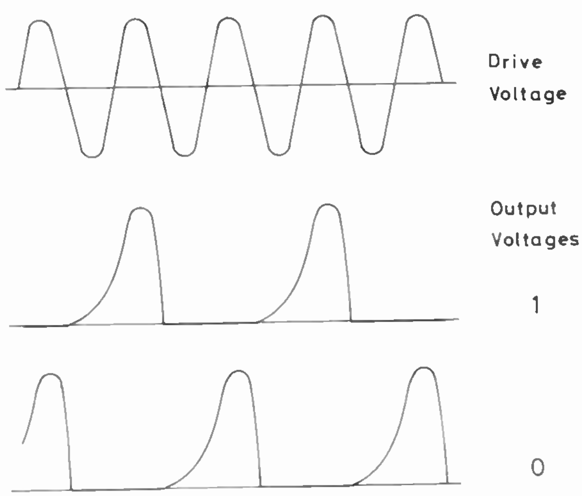



**Fig. 3.** Frequency divider output voltage waveform.

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cyclically varying the bias voltage at a clock frequency which has to be about one-fiftieth of the drive frequency.

## 2. Analysis

A large-signal analysis of the carrier-storage frequency divider is presented elsewhere.[5] Here, only the more important results are given. Referring to Fig. 4, it is assumed that the diode is a sequentially-operated switch which closes when the voltage across it changes from the reverse to the forward direction and opens when depletion of the charge stored at the junction initiates reverse recovery. Circuit conditions are adjusted to make the switch operate at the second-subharmonic of the drive frequency. The circuit performance may be measured in terms of the non-dimensional parameters $\omega_0/\omega$ (the ratio of the L-C resonant angular frequency to the drive angular frequency), $\omega\tau$ (the product of the drive angular frequency and the storage lifetime) and $V/E$ (the ratio of the d.c. bias voltage to the peak amplitude of the drive voltage).

Normally the charge-storage diode is selected having a storage lifetime which makes $\omega\tau = 1$. This value is not critical and diodes having a wide (3 : 1) production spread in $\tau$ may be employed in logic circuits without requiring individual adjustment. It is then possible to predict limits to the bias voltage as shown in Fig. 5. Between $V/E_{min}$ and $V/E_{max}$ experimental and analytical results indicate that the subharmonic output is closely proportional to the value of $V/E$ while the

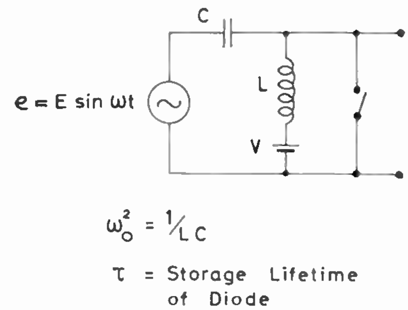


Fig. 4. Large-signal equivalent circuit.

percentage second-subharmonic component remains between 70 and 80 per cent. The limit $V/E_{min}$ is important since it defines a lower threshold to divider operation. Below $V/E_{min}$ the output has no subharmonic component. As $V/E$ is increased through $V/E_{min}$, a subharmonic oscillation builds up. It is found that the phase of the oscillation may be determined by a small locking signal from another divider. A large-signal analysis cannot be used to describe this process and, instead, the divider must be treated as a degenerate parametric amplifier in which the non-linear current-dependent parameter is the diffusion admittance $y_d$.[6]

It can be shown that, for a small-signal variation about a mean current $I_0$, the diffusion admittance of a junction diode is given by,

$$y_d = \frac{I_0}{\delta}(1+j\omega\tau_p)^{\frac{1}{2}} \qquad \text{......(1)}$$

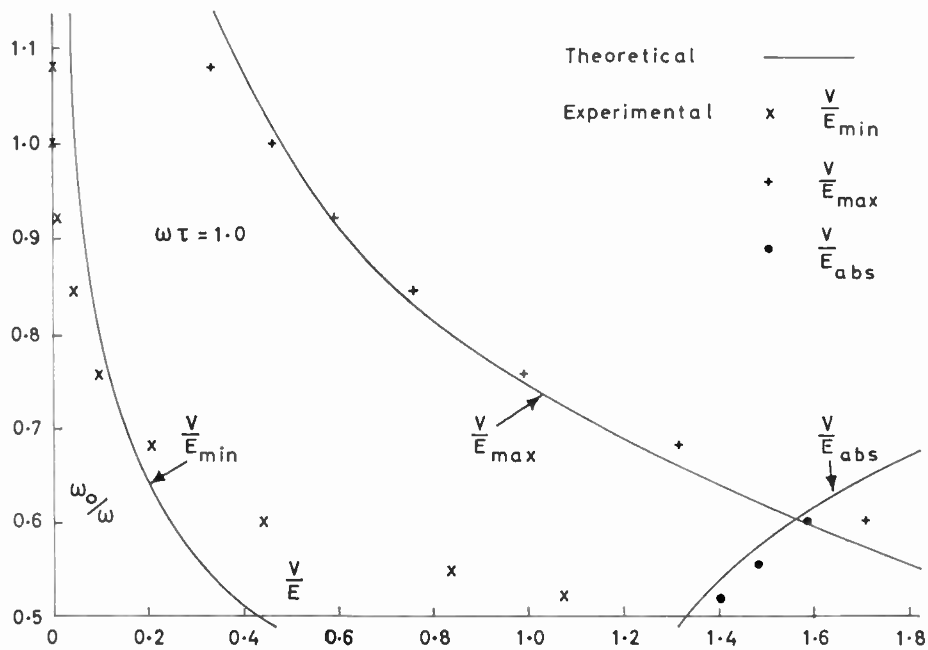


Fig. 5. Experimental and theoretical bias voltage limits.

where $\delta = kT/q = 0{\cdot}026$ volts for germanium at room temperature and $\tau_p$ is the lifetime for holes (assuming a heavily doped p-region). Hence it can be seen that $y_d$ can be written,

$$y_d = g_d + jb_d = (\alpha + j\beta)I_0 \qquad \text{......(2)}$$

since both $g_d$ and $b_d$ are linearly related to $I_0$. The equivalent circuit shown in Fig. 6 can thus be drawn where it is assumed that the signal source is resistive and that the signal current $i_s$ develops a small locking voltage $v_s = V_S \sin(\omega t/2 + \theta)$ across the divider. Equating the subharmonic currents in the several branches to the input current gives,

$$i_s = V_S\left[\left\{G_S + G_L + j\left(\frac{\omega C}{2} - \frac{2}{\omega L}\right) + (\alpha + j\beta)I_0\right\}\sin\left(\frac{\omega t}{2} + \theta\right) + \frac{(\alpha + j\beta)}{2} I_1 \cos\left(\frac{\omega t}{2} - \theta\right)\right] \qquad \text{......(3)}$$

where $I_0$ is the d.c. and $I_1$ the amplitude of the fundamental component of the diode current.

This may be written

$$i_s = V_S\left[\left\{G_S + G_L + j\left(\frac{\omega C}{2} - \frac{2}{\omega L}\right) + (\alpha + j\beta)I_0\right\}e^{j\theta} + {}+j\frac{(\alpha + j\beta)}{2} I_1 e^{-j\theta}\right] e^{j(\omega t)/2} \qquad \text{......(4)}$$

A transducer gain $G_T$ may now be defined such that,

$$G_T = \frac{4G_S(G_L + \alpha I_0)V_S^2}{i_s . i_s^*} \qquad \text{......(5)}$$

where $i_s^*$ is the complex conjugate of $i_s$. Thus,

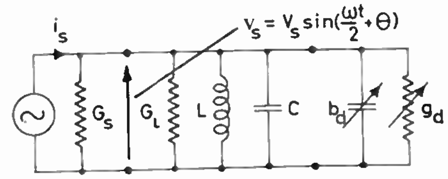


**Fig. 6.** Small-signal equivalent circuit.

$$G_T = \frac{4G_S(G_L + \alpha I_0)V_S^2}{(G_S + G_L + \alpha I_0)^2 + B^2 + \frac{I_1^2}{4}(\alpha^2 + \beta^2) + I_1(G_S + G_L + \alpha I_0)(\alpha \sin 2\theta - \beta \cos 2\theta) + I_1 B(\alpha \cos 2\theta + \beta \sin 2\theta)} \qquad (6)$$

where

$$B = \left(\frac{\omega C}{2} - \frac{2}{\omega L} + \beta I_0\right).$$

This expression for the transducer gain of the divider shows several interesting features. The gain is a function of the susceptance $B$ of the tuned circuit. Maximum gain occurs when $B = 0$, corresponding to $L$, $C$ and the capacitive component of the diffusion admittance resonating at $\omega/2$. The gain also depends on the phase $\theta$ of the locking signal, maximum gain occurring when $\theta = \frac{1}{2}\tan^{-1}(-\alpha/\beta)$, for which

$$\beta \cos 2\theta - \alpha \sin 2\theta = (\alpha^2 + \beta^2)^{\frac{1}{2}}$$

Substituting these conditions into eqn. (6) gives,

$$G_T = \frac{4G_S(G_L + \alpha I_0)V_S^2}{(G_S + G_L + \alpha I_0)^2 + \frac{I_1^2}{4}(\alpha^2 + \beta^2) - I_1(G_S + G_L + \alpha I_0)(\alpha^2 + \beta^2)^{\frac{1}{2}}} \qquad \text{......(7)}$$

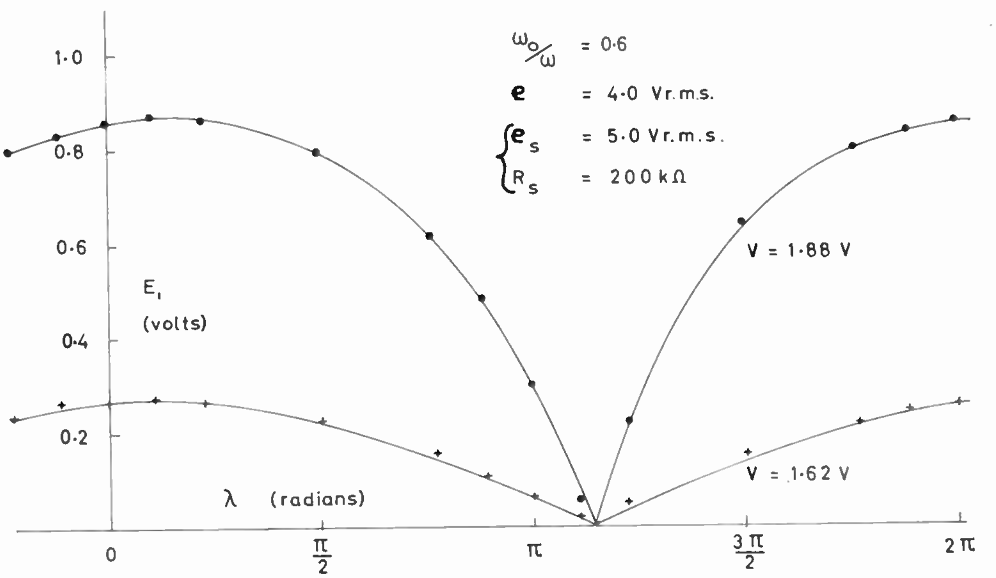


**Fig. 8.** Effect of locking signal phase on output voltage.

This represents the gain of a negative resistance amplifier of the form shown in Fig. 7, where

$$G'_L = G_L + \alpha I_0 \quad \text{and} \quad G_N = (\alpha^2+\beta^2)^{\frac{1}{2}}\frac{I_1}{2}.$$

The condition for oscillation is $G_N > G_S + G'_L$ and, in a lossless circuit in the absence of a locking signal, oscillation commences when,

$$\alpha I_0 = (\alpha^2+\beta^2)^{\frac{1}{2}}\frac{I_1}{2} \qquad \ldots\ldots(8)$$

a condition corresponding to the limit $V/E_{min}$ obtained previously by a large-signal analysis.

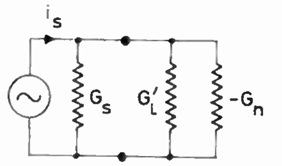


**Fig. 7.** Negative resistance amplifier.

The principal conclusions from the analysis, that below the threshold the circuit gain is phase-dependent and is limited by the parallel susceptance of the tuned circuit and that negative resistance amplification occurs as the threshold is approached, have been verified experimentally. If the phase $\lambda$ of the locking signal $e_s$ is varied with respect to the drive voltage, the amplitude $E_1$ of the divider output voltage subharmonic component varies as shown in Fig. 8. It can be seen that phase locking may be obtained for a wide range of input phase angles. Included in this range is the output phase of a divider operating under optimum bias conditions. Thus phase correction is not normally required in the coupling elements. Figure 9 shows the variation of the subharmonic output as the bias is increased up to $V/E_{min}$, the drive voltage being kept constant. The gain increases rapidly, tending to infinity as the threshold is approached. This permits the in-phase transmission of information from one divider to another which occurs when the bias voltage of the second divider passes through the threshold at $V/E_{min}$. Subsequently the amplitude and phase of the oscillation are independent of the input signal.

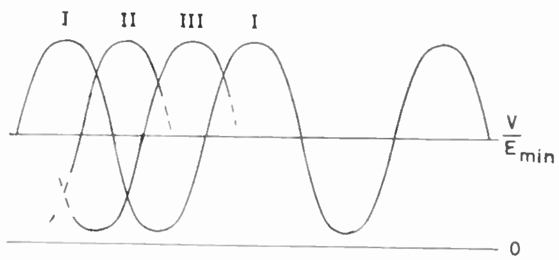


**Fig. 10.** Three-phase bias voltage.

Resistance, capacitance or transformer coupling may be employed between dividers. At higher frequencies transformer or link coupling is preferred. It is simple to connect and a phase reversal, the logical operation of negation in phase script, is easily obtained. Coupling is bilateral and information transfer may occur in either direction depending on the nature of the bias waveform. A unilateral transmission of information may be achieved by employing a three-phase modulation of the subharmonic output voltage of the dividers. With parametrons, this is obtained by a three-phase modulation of the drive

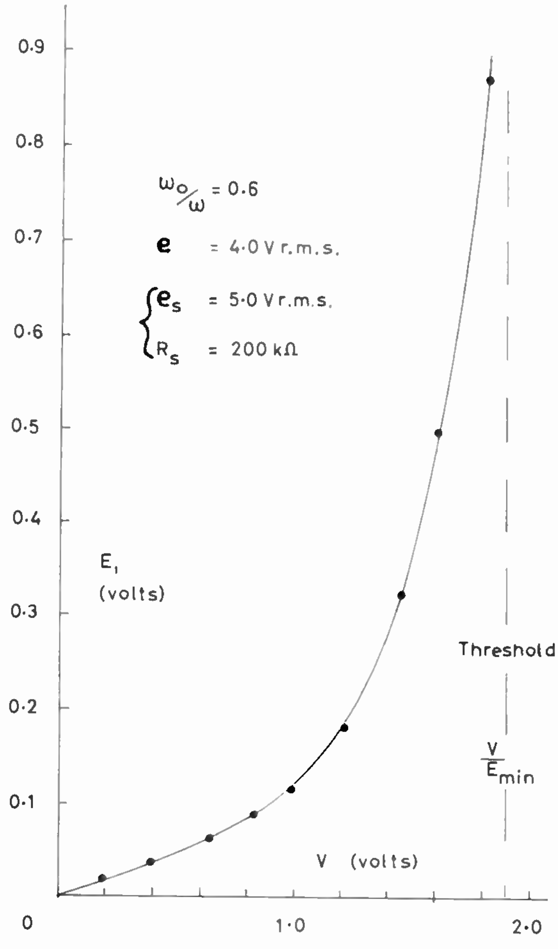


**Fig. 9.** Variation of circuit gain with bias voltage.

voltage. Using carrier-storage frequency dividers it is simpler to leave the drive voltage unmodulated and obtain the modulation of the divider output by means of the three-phase bias voltage waveforms shown in Fig. 10. As the bias to each divider increases through $V/E_{min}$, it becomes phase locked to the divider on the leading phase, the output from the divider on the lagging phase being quenched. If three dividers are



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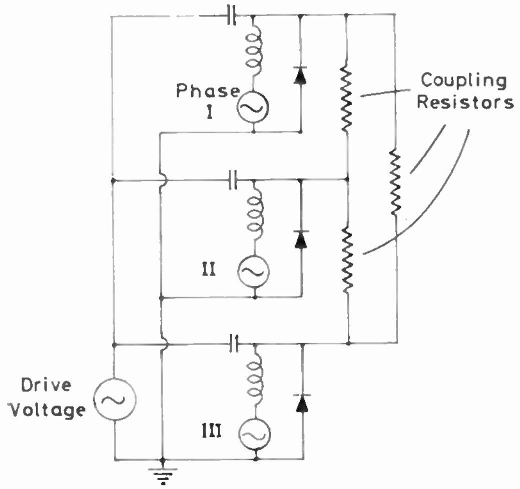


**Fig. 11.** One-bit dynamic memory element.

coupled together and derive their bias in turn from the three phases of the bias supply as shown in Fig. 11, then one bit of information may be caused to circulate continuously. This circuit may be considered to be a one-bit dynamic memory element.

The propagation delay between successive dividers is one-third of the period of the bias frequency. This in turn is limited by the need for reliable phase locking during the interval in which the bias voltage is passing through the threshold region. In practice it is found that a ratio of 50 : 1 between the drive frequency and the bias frequency is necessary. A comparable ratio is required with parametrons. It may be reduced somewhat by operation with values of $\omega_0/\omega$ close to unity and by the use of non-sinusoidal bias voltage waveforms, although any improvement so obtained is unlikely to exceed a factor of two. It would appear to be more promising to investigate operation at higher frequencies than the present 1 Mc/s maximum drive frequency.

**Table 1**

Truth table for majority decision

| a | b | c | d |
|---|---|---|---|
| 0 | 0 | 0 | 0 |
| 0 | 0 | 1 | 0 |
| 0 | 1 | 0 | 0 |
| 0 | 1 | 1 | 1 |
| 1 | 0 | 0 | 0 |
| 1 | 0 | 1 | 1 |
| 1 | 1 | 0 | 1 |
| 1 | 1 | 1 | 1 |

### 3. Majority Logic

It can be seen that if the outputs of an odd number of dividers are added linearly, the resulting signal may be used to phase lock another divider. The ability of a threshold device to respond to the majority signal of a summed input is majority decision and the logic so derived is majority logic. Table 1 is a truth table showing the majority decision of three inputs a, b and c. This may be expressed in Boolean form.

$$d = \bar{a}bc + a\bar{b}c + ab\bar{c} + abc \qquad \ldots\ldots(9)$$

$$= ab + bc + ca \qquad \ldots\ldots(10)$$

and the Boolean connectives AND and OR may be obtained simply by biasing one of the inputs to 0 or 1 respectively as shown in Fig. 12. Here the divider

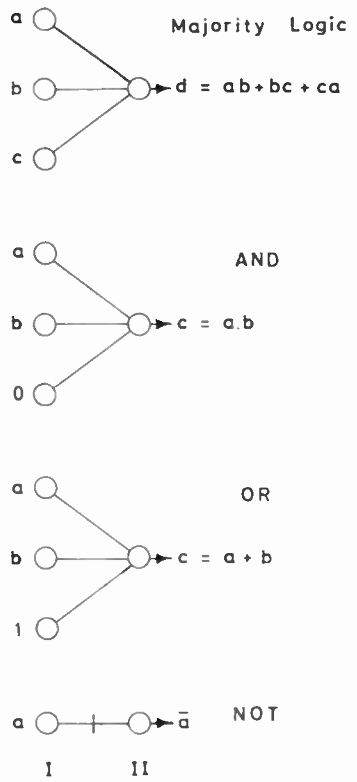


**Fig. 12.** Majority decision, AND, OR and NOT circuits.

elements are represented by open circles and the coupling elements by links between the circles. The third Boolean connective NOT is obtained by phase reversal and indicated by a short bar across the coupling link. The phase sequence of the bias voltage is indicated by I, II, III and the information is transmitted in this direction.

Although the majority decision element may be used in the above way to synthesize any Boolean function since only the operations OR, AND and NOT are necessary to do so, it is more economical to use it

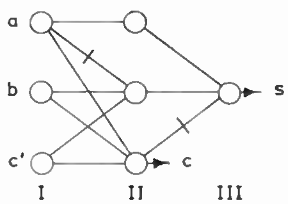


Fig. 13. Three-digit full adder.

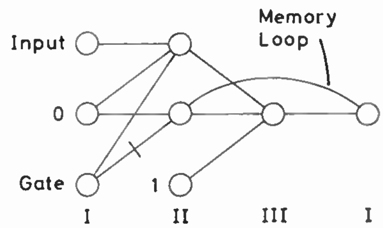


Fig. 15. Flip-flop with gated input.

as a three-variable input device. Synthesizing techniques have been developed which allow some minimization of the number of majority elements used in this way.[7] All these methods are limited in application and practical design is still empirical to some extent.

Some examples of the more elementary logical blocks will now be given. Figure 13 shows a three-digit full adder. The input consists of the addend and augend digits a and b and the carry digit c′ from a less significant stage. The carry c and the sum s are obtained from the first and second logic levels respectively. Figure 14 represents an output selection matrix for four inputs. The appropriate input is selected by the digit group ba. For example, ba = 10 selects input $I_2$ and transmits it to the output. A flip-flop with a gated input is shown in Fig. 15. With the gate at 1, the input is transferred to the dynamic memory loop, and continues to circulate when the gate changes to 0. By extending the number of elements in the loop to $3n$ and gating the input for $n$ clock periods an $n$-digit word may be serially transmitted into the loop which may now be used as a serial word memory block. Finally, a scale-of-four counter is shown in Fig. 16. This consists of two binary counters in cascade, the state of each changing as a 1 appears at its input.

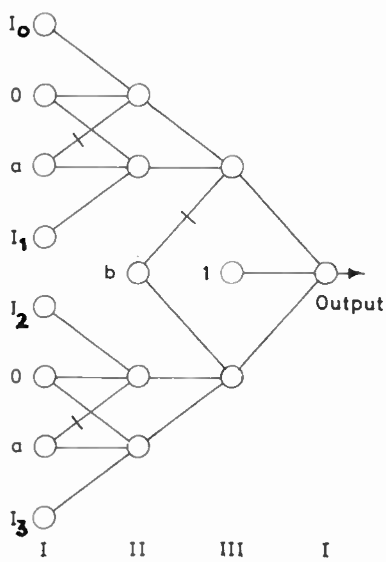


Fig. 14. Output selection matrix.

All the above circuits have been tested experimentally using a drive frequency of 50 kc/s and a clock frequency of 1 kc/s. Some tests have also been made with drive and clock frequencies of 1 Mc/s and 10 kc/s respectively and an extension to at least 10 Mc/s drive frequency appears possible. A study has also been made of the logical design of a small computer using the techniques of phase-script majority logic. This has shown the logical feasibility of a serial arithmetic unit, control unit and memory selection matrices using carrier-storage frequency dividers. As with parametrons, serial memory loops are not economical and alternative memories, such as ferrite core arrays, are necessary.

## 4. Conclusion

It has been shown that the carrier-storage frequency divider may be used to perform logical operations in a manner similar to the parametric second-subharmonic oscillators. It possesses advantages in that the circuit elements are simple and inexpensive and tolerances are less. Both drive and bias voltages are sinusoidal and the power required per logical element is small, typically less than 5 mW. Operating speeds attained so far are about the same as with ferrite-cored parametrons, although an improvement of several orders

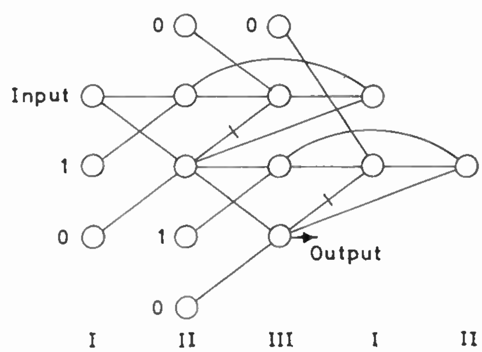


Fig. 16. Scale-of-four counter.

of magnitude is considered possible. This may be achieved by increasing the drive frequency, possibly to 10 Mc/s or higher, by increasing the clock rate to one-half the drive frequency and by using ternary or higher-order logic with phase or hybrid script. Current investigations indicate the practicability of all the above improvements. Nevertheless, it is felt that the carrier-storage frequency divider computing element is more likely to find application in a computing or data processing system where simplicity and economy are more important than speed.

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*Manuscript first received by the Institution on 12th September 1963 and in final form on 19th February 1964. (Paper No. 899/C70).*

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# Brit.I.R.E. GRADUATESHIP EXAMINATION, NOVEMBER 1963

## PASS LISTS

The following candidates who sat the November 1963 examination at centres outside Great Britain and Ireland succeeded in the sections indicated. The examination, which was conducted at 65 centres throughout the world, attracted entries from 443 candidates. Of these 116 sat the examination at centres in Great Britain and Ireland and 169 sat the examination at centres overseas. The names of successful candidates resident in Great Britain and Ireland were published in the March-April issue of the *Proceedings* of the I.E.R.E.

| | *Candidates appearing* | *Pass* | *Fail* | *Refer* |
|---|---|---|---|---|
| *Section A* | | | | |
| Great Britain | 59 | 23 | 26 | 10 |
| Overseas | 85 | 31 | 43 | 11 |
| *Section B* | | | | |
| Great Britain | 57 | 21 | 28 | 8 |
| Overseas | 84 | 17 | 45 | 22 |

### OVERSEAS

**The following candidates have now completed the Graduateship Examination and thus qualify for transfer or election to Graduate or a higher grade of membership.**

AYIVORH, Samuel Clifford (S), *Kumasi, Ghana.*
BASSEY, David Andrew (S), *Nigeria.*
BHOWMIK, Sital Chandra (S), *New Delhi, India.*
BIEGUN, Ephrain (S), *Tel-Aviv, Israel.*
CHADHA, Narindar Nath (S), *Delhi, N. India.*
GAUR, Bibhuti Bhushan (S), *Uttar Pradesh, India.*
GOULDING, Royston David (S), *Lagos, Nigeria.*
HOWARD, Trevor Neville (S), *Salisbury, S. Rhodesia.*
JAYAN, Kulty N. (S), *Bombay 9, India.*
KUMAR, Shanti Deva (S), *Bangalore 15, India.*
NARAYANAMORTAY, Kalambiu (S), *Madras 4, India.*
NWASOKA, Joseph Chukwunwike (S), *Lagos, Nigeria.*
ONOZIE, Baldwin Onyenaucheya (S), *Berne, Switzerland.*
SEAH, Cheng Huat (S), *Singapore 14.*
SETHURAMAIAH, Hanasoge Haranappa (S), *Hyderabad.*
SUBRAMANIAN, N. (S), *Gauhati, India.*
WADHAWAN, Pran Nath (S), *New Delhi, India.*

**The following candidates have now satisfied the requirements of Section A of the Graduateship Examination.**

ANANDU, Verkatassibbiah, *Poona 5, India.*
AVIZUR, Chamabikisis (S), *Vitkin, Israel.*
BARNEA, Joseph, *Afridar, Ashkelon.*
CURTIS, Terryl David Warren (S), *Nova Scotia, Canada.*
DILLEY, Arthur (S), *Nova Scotia, Canada.*
GAVISH, Avraham Givatayim, *Israel.*
JAYARAM, Thali Kottiath (S), *Singapore 19.*
JOHNSTON, Kevin James (S), *Epsom, Auckland.*
KING, William (S), *Adelaide, S.A.*
LASAKI, Muriseli Folrinso, *Lagos, Nigeria.*
LEES, Frank Philip (S), *Salisbury, Southern Rhodesia.*
LEUNG, Wing Kun, *Hong Kong.*
MORMAN, Alexander Rex (S), *Packakariki, N. Zealand.*
NAGARJO-RAO (S), *Mysore, Bangalore 4, India.*
NAGARENDRAN, Lakkenahalli (S), *Mysore State, India.*
NARASIMHA-MURTI, K. N. M., *Bangalore 4.*
OLIVER, Terence John, *Suva, Fiji.*
PADMANABHAN, T. M. (S), *Madras.*
PANDE, Girja Bhushan, *Lucknow, India.*
PANT, Lalit Mohan, *Lucknow, India.*
PARTHASARATHY, Kethandapatti R., *Bangalore 4.*
RAGHAVENDRA-RAO, C. K. (S), *Bangalore 4.*
SEHGAL, Yogindar Mohan, *Delhi 6.*
SHARPE, Michael William Vawser (Assoc.), *Hong Kong.*
SHINDI, Vinayakrao Bhaurao (S), *Cochin 4, India.*
SRIDHARA, Ananthapurma R., *Bangalore 13.*
SUBBARAD, China Maddukun Andhra, *South India.*
SUBRAMANYAM, Seshu, *Punjab 6.*
TREVETT, Eric Henry (S), *Kingston, Jamaica.*
VERMA, Roshan Lal (S), *Bombay 29, India.*
GINSBURG, Amihay, *Zupal, Israel.*

The question papers set in Section A of the November 1963 Graduateship Examination have been published in the March-April issue of the *Proceedings* of the I.E.R.E., together with answers to numerical questions and examiners' comments. Parts 3 and 4 of Section B and Part 5 of Section B will be published in subsequent issues of *Proceedings*.

(S) denotes a Registered Student.



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U.D.C. 621.396.625.3: 621.9

# A Low-cost Magnetic Tape Control System for Machine Tools

*By*

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*(Associate Member)*†

AND

R. L. DUTHIE

*(Associate Member)*‡

*Presented at the Convention on "Electronics and Productivity" in Southampton on 18th April 1963.*

**Summary:** Digital recording techniques are applied to the operation of a die-sinker. The system described uses magnetic recording in preference to punched paper tape where cost, transportation and simplicity are the major design considerations.

Irregular shapes and contours are traced by a stylus which is electrically connected via an arc to a conductive coating on the model. The position of the stylus is related to pulses which are recorded on tape. The recorded tape may be transported to any workshop which employs similar techniques for cutting a work-piece with a permissible tolerance of $\pm 0{\cdot}005$ in.

## 1. Introduction

Industrial requirements are exceeding the pace at which skilled craftsmen can be trained and consequently there is a need to develop machines which can make optimum use of skilled labour. This can be accomplished in two ways: either by enabling a craftsman to work more quickly or by enabling him to transmit his skill in the form of detailed instructions which can be followed repeatedly by relatively unskilled people. The aim of the system is to combine both these techniques in a relatively simple and inexpensive manner.

The method adopted is to enable a skilled craftsman to prepare a programmed tape containing all the information necessary to perform a set job. The operator need only have sufficient skill to thread the tape through its guides, and to follow instructions regarding work-piece material, type of cutter, starting position, etc. To simplify problems of handling and transport, economy has been observed in the amount of tape required to be used.

One machine is used by the craftsman doing the programming and another is used by the man performing the job in hand. The programming machine is not required to do heavy work and so can be of relatively light construction. To save the programmer's time, it is able to program information at a considerably greater speed than the play-back machine uses for the work-piece. This is facilitated by the fact that the programming machine is free from the limitations of cutter loading. Computers are precluded by their expense and the fact that they require skilled programming techniques.

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In a system of this kind it is uneconomic to provide a tool having vastly greater accuracy than is usually needed. Many applications call for accuracies of the order of $\pm 0{\cdot}005$ in. This comparatively modest dimensional accuracy makes low cost, simplicity and reliability easier to achieve.

The system is designed for use in factories or workshops where skilled electrical maintenance may not be readily available; the machine may even be situated in a remote area. Consequently it should be capable of running for long periods (say about five years). To provide such reliability, full consideration must be given to component reliability, freedom from mechanical drift and long-term wear-and-tear (i.e. life expectation of components where 'life' is related to the number of operations performed).

## 2. Basic Conception

It was decided to adopt static-switching techniques, built in removable unit form, throughout the electronic system. The actual division of the circuitry is such that, as far as possible, each unit represents one particular function of the machine. The method has two beneficial factors: (*a*) quick and simple maintenance, (*b*) a customer need only buy those facilities which he actually needs. These considerations led to the adoption of printed wiring boards supported in sheet metal frames.

The front panel of each frame carries a plug for external connections, a socket for a plug-in test unit and spring-loaded fasteners by which the frame is secured to the control cubicle. Direct edge connection to the printed board has been avoided, since it is considered that this type of connection is not as reliable in machine tool applications as the more robust plug and socket combination. The cost and



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potentialities for failure of the soldered connecting wires from the board to the panel-mounted plug were considered to be outweighed by the greater reliability (Fig. 1).

For the purpose of elucidating the system, and in order to prove the prototype, it was decided to make the first application that of a die-sinking machine. This application offers full scope for all three dimensions, and it is hoped that it will overcome the growing shortage of skilled die-sinkers. Those who are skilled in die-sinking are not always skilled in drawing; conversely draughtsmen are not always good at designing dies. By long tradition, die-sinkers are good at making complex shapes, so it seemed logical that the machine should work from models of the shapes required, particularly since in many machine shops these are already in existence. This often applies when it is decided to put a hand-made object into quantity production.

For these reasons, it became apparent that the need was for a three-dimensional copying machine. However, a direct copying machine was not considered to be suitable because it would not allow the 'heavy' work to be done at a time and place both remote from the craftsman. This necessitates some kind of a 'memory' store.

Magnetic tape was chosen for this application because the information packing density attainable is far higher than is feasible with punched paper tape or other similar media. Sprocketed tape was not chosen because stretch on the tape, or a small inaccuracy of the capstan diameter, would tend to cause the tape to climb the sprocket teeth.

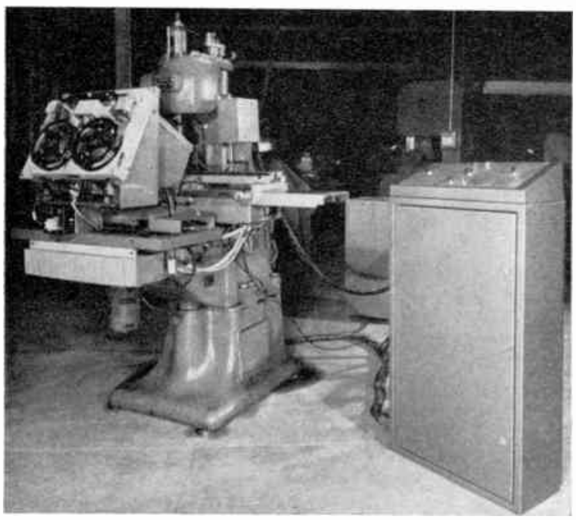
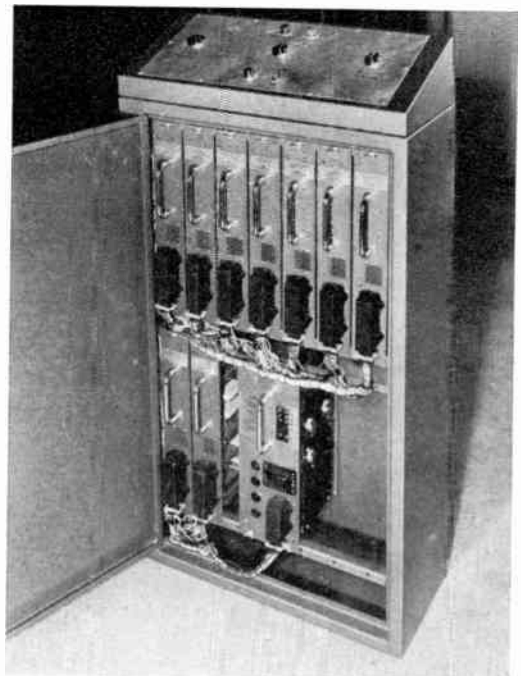
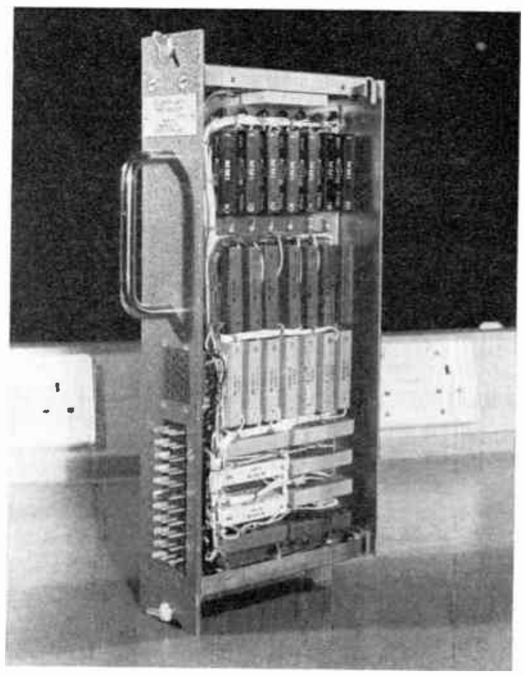


Fig. 1. (*a*) Prototype equipment showing the applications of the controller to a three-dimensional miller. In this version, programming and play-back facilities were combined in one machine.



(*b*) The controller with the door open, showing the method of assembly. The top panel is also hinged, allowing easy access to the backs of the switches. This photograph is of a prototype.



(*c*) A typical sub-panel. The encapsulated units are mounted on a printed circuit card. The twenty-five-way socket is only used for testing.



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For reasons of economy it was decided to use readily available components where possible, so dictating the choice of a sixteen-track system using one-inch wide tape.

The majority of magnetic tapes commercially obtainable are far from being of a reliably homogeneous material. Due to defects in the manufacturing process, minute areas of the tape suffer from a deficiency of magnetic oxide which reduces the sensitivity of the tape for retaining magnetic pulse data. The distribution of such areas is random, but they do not appear at intervals frequent enough to be of a serious disadvantage. The fact that it is not usual to use all channels simultaneously reduces the risk of losing pulses due to these 'drop-outs'. Where a train of pulses is recorded on one channel, the defect is as likely to occur between pulses as upon them. Each channel was duplicated thereby reducing the statistical probability of a work-piece being damaged to an acceptably low figure. Duplicated channels were arranged so that they were never physically adjacent to each other.

The effectiveness of the eight pairs of channels can be increased by employing two or more at once. For example, a pulse whose significance indicates a downward movement might be used in conjunction with a pulse indicating an upward movement. The resulting 'nonsense' command thus implied might be used as a code to perform a different function.

For reasons of economy it is desirable, although not essential, that all the information for any given die should be stored within the same spool. Economy of the amount of tape used is also reflected in cheaper transport, especially when air-freight is used.

It was decided, therefore, not to let the tape run at a constant speed, but to relate the amount of machine movement to the amount of tape travel. This is extremely economical; however, it results in the tape running in an irregular and sometimes jerky manner, and there are times when the tape is completely stationary. Therefore flux-sensitive heads[1,2] are used for reading the information stored on the tape. As a direct relationship between machine-tool travel and tape travel would result in an excessively high tape packing density, a gear ratio of approximately 3:1 is interposed between tape and machine.

A digital system was chosen, because the uneven manner of running the tape does not seriously interfere with a method based on the presence or absence of a pulse. It is, however, more difficult to apply to an analogue technique having relatively fine variations of emphasis.

The fact that the programming and play-back machines are separated both in space and time, makes it impossible to fit an over-all feedback loop to correct for cumulative error between the model and the work-piece. This again favours the choice of a digital system.

### 3. Speech 'Recorded Instruction'

As the machine tool will not necessarily be part of a larger installation, it follows that the loading of the work-piece on to the table will probably have to be done manually. There is no reason why a semi-skilled worker should not be able to set up the machine provided he has the necessary instructions.

There are a number of ways in which these instructions could be conveyed; for instance, printed instructions could be provided with the tape. However, it was felt to be preferable that the instructions should be in the form of speech.

A self-contained auxiliary tape unit is supplied to meet these requirements. When the craftsman, who is doing the programming, decides to record an instruction, he operates a 'record speech' button which stops the machine tool and starts the auxiliary recorder. An instruction code is recorded on the main tape. The button is kept pressed while speaking but when it is released the machine tool resumes its normal functions.

This system is simple and versatile. However, it does require the occasional intervention of an operator, and it carries the possibility of language difficulties.

### 4. Stylus Control in the Vertical Plane

Although occasional commands may be verbal it is essential, if the craftsman's time is to be saved, that the details of the shape of the model should be conveyed to the machine automatically. The contours of the model are traced by moving the table, upon which it is placed, and recording the vertical contours which the stylus has to follow. The movement of the table is of a simple raster configuration sufficiently large to embrace the size of the model.

The machine-tool table moves sideways, from front to rear, and then sideways in the opposite direction; another front-to-rear motion follows. This is repeated a number of times to comprise one raster. In each raster the direction of front-to-rear movement remains constant, but is reversed when the raster is complete.

By convention, movement from side to side of the model or work-piece is referred to as the X-movement. Movement from front to rear is called the Y-movement and movement in the vertical plane is the Z-movement. As the model may be made from a soft, easily worked material such as plasticine, the stylus should not distort its shape by the application of excessive pressure. This is avoided by letting the pressure be



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zero. The stylus is made to hover a nominal distance above the model and in fact does not touch it.

In the programming role, the Z-drive is controlled by a spark proximity detector. When the stylus is high, that is, remote from the model, the open circuit voltage which it carries is used to operate a servo system causing the stylus to move towards the model. When the stylus comes within about 0·002 in of the model, a spark occurs. The voltage on the stylus falls and the servo system causes the stylus to rise until the spark is extinguished. The resultant Z-motion is oscillatory with a peak-to-peak amplitude of about 0·0005 in which is adequate for this purpose. Provided the model is even slightly electrically conductive, a low-energy sparking system can be made to maintain the desired gap. A high-impedance voltage source renders the method safe to the operator. The small amount of hunting which occurs is well within the tolerance of the machine.

If a non-conductive material is used for the model it must be provided with a thin conductive coating. The stylus will follow the model providing that the resistive path offered does not exceed about 100 kΩ.

The mechanical portion of the Z-drive carries a photo-electric rotary switch arranged to provide an output pulse for every 0·002 in of Z-movement. Since the oscillatory motion of the Z-drive could generate spurious pulses, a 'paralysis' circuit is included which inhibits such action. The switch is a single-pole four-way device, with the four photo-electric 'contacts' connected in alternate pairs to the terminals of an Eccles-Jordon flip-flop store. Once contact A is made, the flip-flop is set and further making and breaking of contact A has no further effect. The flip-flop will, however, be reset when sufficient Z-motion has taken place to operate either of the contacts B or D. Since the peak-to-peak amplitude of the oscillatory movement is well within the spacing of the contacts, this will only occur if a general movement of the Z-system has taken place.

The stylus used is selected by the skill of the craftsman doing the programming and should be of the same shape and size as the cutter to be used in making a die, minus the necessary allowance (0·002 in) for the spark gap. If the craftsman introduces a verbal command to change the type of cutter, he should himself make the corresponding change in the type of stylus. However it is anticipated that this will only be an occasional requirement.

In some machining processes it may be necessary to limit the distance travelled in the Z-plane during one particular operation to prevent damage to the tool. The maximum permissible distance depends upon many factors which have to be taken into account by the programmer. Chief among these considerations are the hardness of the metal to be cut and the type and size of cutter to be used. These factors do not limit the *total* depth to which the machine can cut, but only the depth which it can cut during the performance of any one raster.

Using his skill and knowledge of metal cutting techniques, the craftsman makes his choice from four available Z-limit settings, namely 0·032, 0·064, 0·096 and 0·128 in. This limit applies during the whole raster but the programmer can at his discretion select a new limit which will be applicable after the raster is complete.

This system, shown in outline in Fig. 2, consists basically of a binary counter operated by the 0·002-in pulses from the photo-electric switch previously described. The first four stages are binary dividers, the last of which produces an output for every 16

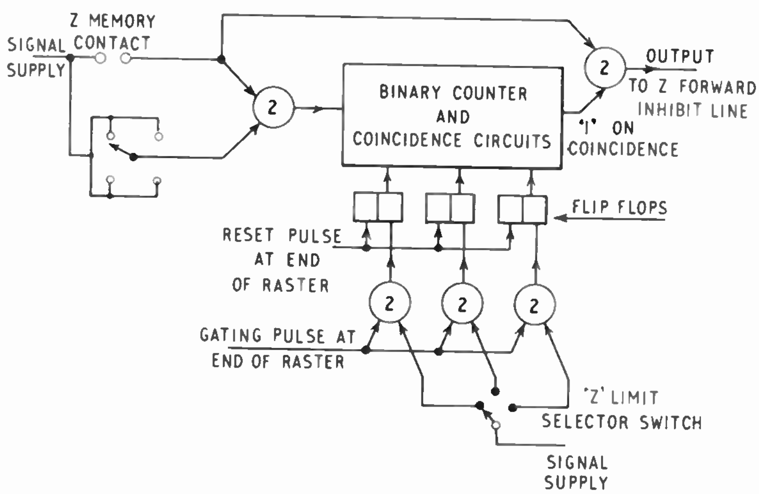


Fig. 2. Outline of Z-limit circuits.



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switch pulses, i.e. for every 0·032 in of Z-movement. This is followed by a further three stages of binary division associated with conventional coincidence detection circuits,[3] which produce an output when they are in the same state as that required for the work in hand. Since the same limit must apply over a complete raster, it is necessary to store the information of the limit required, but the programmer must be able to set the switch for the next limit during a raster. This storage is done by three flip-flops which can be reset to a new value only at the end of a raster. The 'reset' pulse is obtained by the operation of the Y-limit switch, the primary purpose of which is to change the direction of the Y-movement ready for the next raster.

In effect, this system only monitors downward movement of the stylus, and in order to follow the shapes of different dies, the movement of the stylus may be occasionally undulatory. In this instance the memory would have to remember the lowest position reached by the stylus before an upward movement takes place, and should prevent further monitoring in the Z-plane until the stylus returns to that level. A simple but effective electro-mechanical memory is used, which takes the form of a shorting-bar which is friction mounted on the Z-slide. When downward movement first takes place, contacts push against this bar and consequently complete the path to energize the monitoring device. The pressure on the contacts is sufficient to cause the friction-mounted bar to slip on the Z-slide while it is moving downwards. When an upward movement takes place the contacts no longer push against the bar and the circuit is broken. Friction holds the shortening-bar in the same position until a downward movement of the stylus returns the contacts to it. The contacts operate in a trouble-free manner as they have a favourable duty-cycle and are self-wiping. The Z-limit function is effected by causing the coincidence signal and the closure of the electro-mechanical memory switch to block the Z-downward drive.

Another factor to be taken into account by the craftsman is the coarseness of the raster to be used on each occasion, which depends on the hardness of the metal to be cut, the type and size of the cutter to be used and other metallurgical characteristics of the material. He can change the raster setting during programming at his discretion and in each case his choice is recorded automatically on the tape.

### 5. Table Control in the Horizontal Plane

Movement in the Y-plane is initiated by the operation of the X-limit switch which triggers a flip-flop to set the table in motion in the Y-plane. It also inhibits any further movement in the X-plane until the movement is complete.

Control of the direction of Y-motion is automatic, and is determined by a flip-flop which is set to the required condition by the operation of the Y-limit switch at the end of each complete raster. The degree of coarseness of the Y-movement is determined by a Y-increment switch which gives the choice of 2, 4, 8, 32, 64, or 128 thousandths of an inch. Pulses, at 0·002-in intervals from a photo-electric rotary switch, similar to that used in the Z-motion, are fed to a seven-stage binary counter with coincidence circuits on the outputs of all seven stages. The Y-increment switch determines which coincidence circuit operates and hence controls the distance of Y-movement.

The setting of the Y-increment switch is coded on the tape by simultaneous operation of heads on different tape tracks. Up to three pairs of heads may be employed. This block-coding technique avoids the necessity of using separate communication channels for each possible choice of Y-increment. When the movement is in a reverse direction the tape is supplied with a 'Y-reverse' code.

While dies vary greatly in shape, it was realized that, in the majority of cases, most of the movement would be in the X-direction. Consequently, it was attempted to make this information channel as simple as possible. The simplest code is an absence of signal, so it was arranged that the machine should run in the X-direction until prevented by the arrival of a signal which may relate to another axis, whereupon the X-movement would cease. An example of this condition is when the programming machine is required to perform in the horizontal planes. By convention the machine starts at the rear left-hand corner of the die, and operates in the X-plane until the pre-set table stop operates stopping movement in the X-plane and starting movement in the Y-plane. On subsequent operation in the X-plane movement will be in the reverse direction. Additional pulses are recorded on the tape and coded in relation to the amount of Y-movement. When Y-movement is complete the inhibition is removed from X and the next line of the raster commences. When the whole raster is completed, another table stop reverses the Y-movement.

The X-, Y- and Z-motions are all controlled by 'on-off' switching methods in the interests of simplicity and economy, and provide adequate accuracy. The basic circuit controlling the drive to each axis (shown in Fig. 3) consists of a 'stop-go' gate followed by a directional 'forward-reverse' memory, giving the possibility of three-state control of the drive—forward, stop and reverse.

Since some of the machines controlled by this system may use hydraulic drive the standard controller gives an output suitable for the type of hydraulic pilot valve used. When electric drive is used this output is taken to a motor-control unit.



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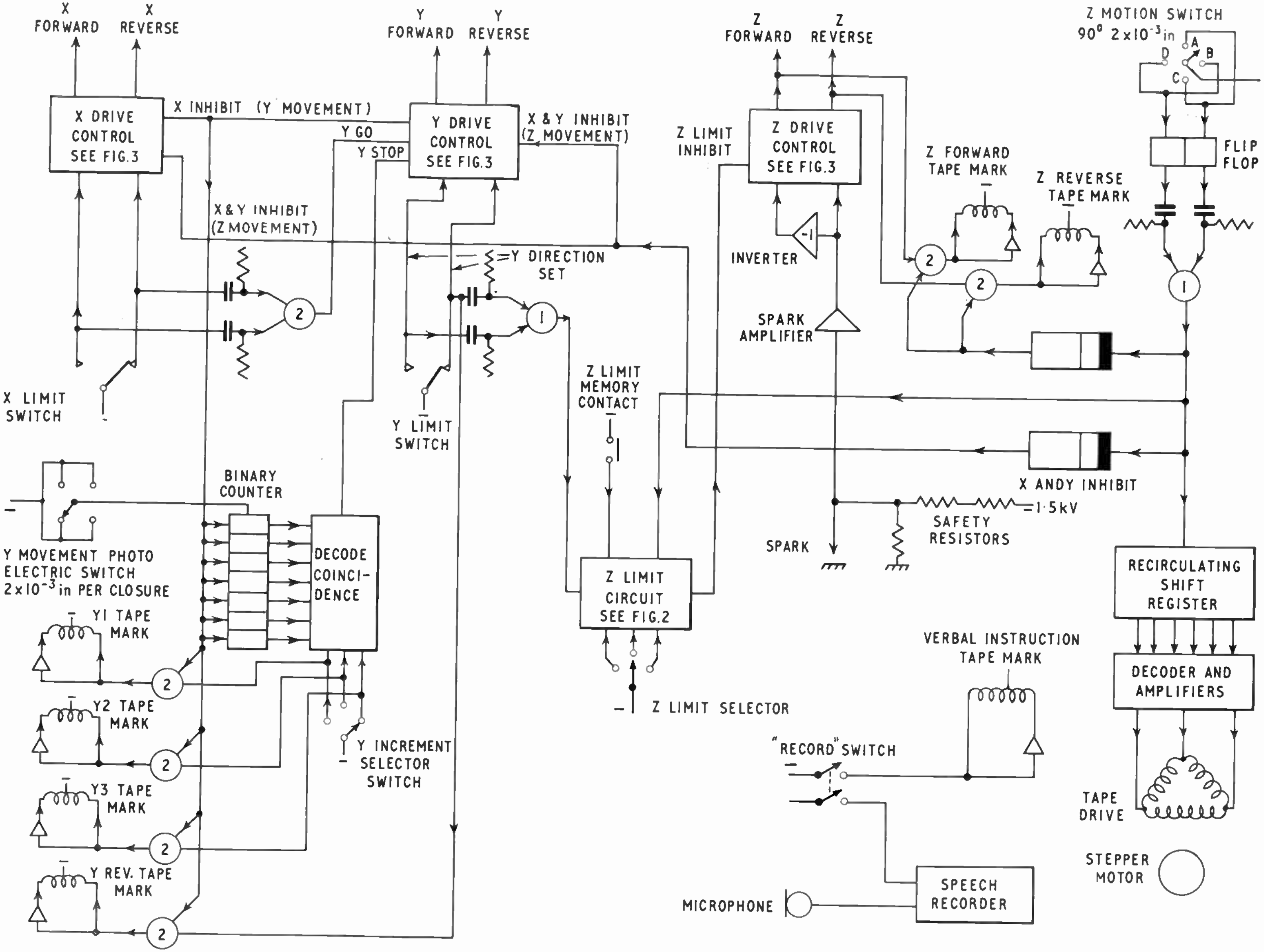


**Fig. 4.** Outline of control system for programming machine.

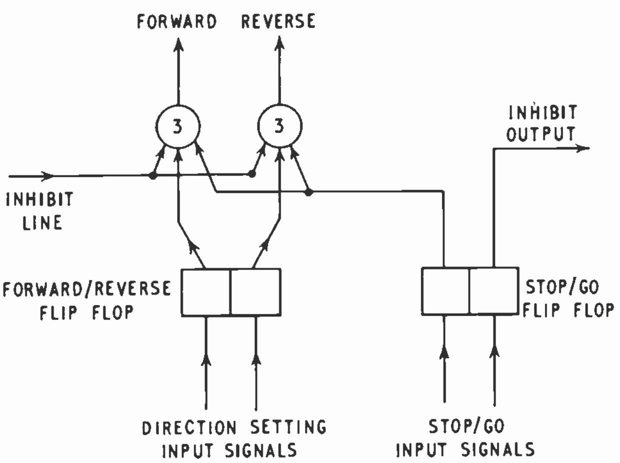


**Fig. 3.** Basic drive control circuit.

## 6. Vertical Profiles

The programming system (see Fig. 4) consists basically of three machine-drive controls interlocked so that the Z-axis takes precedence. The machine may operate in the Y-plane and the Z-plane simultaneously but it can operate in the X-plane only if Y and Z are stationary. Only one axis is normally followed at a time, therefore, sloping surfaces on the model can only be followed by adopting a stepping motion. This does not cause excessive error in the work-piece since the system accuracy is only $\pm 0{\cdot}005$ in. As this degree of approximation is acceptable a considerably simplified circuit can be used.

As previously explained, the programming machine ignores minor hunting of the sparking stylus, but when travel in either the upward or downward direction reaches 0·002 in, it stops motion in the X-plane whereupon pulses corresponding to the Z-axis are recorded on the tape; different channels are used according to the direction of motion. The tape is then advanced so that the pulse on the tape is well clear of the recording head; this is done by an auxiliary tape drive without moving the machine-tool table. A small delay is then allowed to elapse.

If the stylus is adjacent to a vertical profile on the model it will have moved a further 0·002 in before the time delay has elapsed and another Z-pulse will appear on the tape as described above. By repeating this procedure a whole train of such pulses may be formed.

If the stylus is adjacent to a sloping profile on the model it will move 0·002 in and apply one pulse to the tape. However, during the time delay nothing further will happen so that when the delay has elapsed, the table will move again in the X-plane. This will cause further movement in the Z-plane. Alternate X- and Z-movements take place; the length of X-movements vary according to the gradient of the slope.

It is important that the amount of auxiliary tape drive applied between pulses should be the same on the play-back machine as on the programming machine. This is achieved by the use of a stepping motor system to give a fixed pre-determined movement of the tape drive for every command pulse.

From the above it will be seen that tape motion does not only depend upon the raster shape described by the machine-tool table; it is further complicated by stopping and starting due to functions associated with vertical profiles. In practice this leads to the tape motion being somewhat jerky. One simplifying factor is that the travel of the tape is unidirectional.

## 7. Tape Synchronization

Unfortunately magnetic tape has a high thermal coefficient of expansion, and can easily be distorted by mishandling. Another source of trouble is the possibility of a small amount of slip in the capstan drive system even though the grip may be good. It was considered to be unpracticable to expect zero tolerance between the capstan diameter of the programming and play-back machines, hence the cumulative error which might arise would reflect on the finished product. Sprocketed tape would prevent this error becoming cumulative but the likelihood of the tape tending to climb the sprocket teeth and then slip back, especially when the tape has stretched or shrunk, could lead to more erratic results.

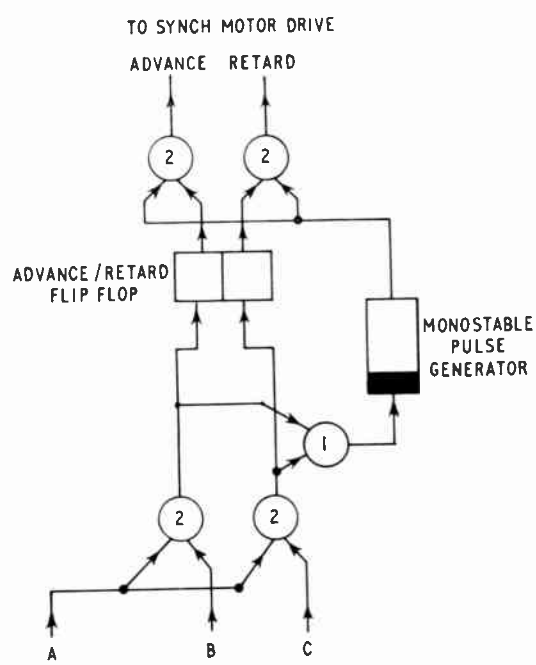


**Fig. 5.** Basic circuitry of synchronizing unit.

The solution to this problem is to use unsprocketed tape with synchronizing pulses recorded on the tape at known discrete intervals during the programming operation. The play-back machine generates electrical signals which should occur just before (B) and just after (C) the tape pulses (A); A, B and C refer to the inputs to the synchronizing circuit (Fig. 5). If coincidence occurs in one sense, the tape is advanced relative to the machine-tool table; if coincidence occurs in the other sense it is retarded. This corrective motion is imparted by means of a small motor which is controlled by drive circuits similar to those used for controlling the major axes of the machine.

## 8. Erasure

It would be convenient if used tapes could be programmed again without too much preparation and even if a new reel of tape was used, there would always be the risk that it may accidentally have become partially magnetized. For these reasons it was decided to over-print the program upon anything which might already be on the tape. During programming all sixteen write heads are energized all the time. When no significant pulse is being applied, the tape is magnetized in a negative sense; a 'command' pulse is recorded by alternating the polarity. In both cases the magnetizing force is sufficient to drive the tape to saturation.

Although the fields on the negatively polarized parts of the tape do not remain uniform over long distances, and the action is a little more complex than it appears, erasure of the unwanted signal is effectively achieved and the signal/noise ratio remains adequate. Hence, the provision of a special erasing system is rendered unnecessary.

## 9. Supplementary Store

It sometimes happens that the machine is required to remember instructions after the pulse bearing the information has passed the head. For example, it is necessary that the machine should remember the direction of travel in the X-plane as during an X-movement the tape carries no command signals; similarly, the pulses which initiate a Y-motion are of very short duration compared with the length of time taken by the machine to carry them out.

The control system depends upon flip-flops for its memory functions in normal operation, but these will hold their information only as long as their supplies are maintained. When power is restored after a loss of supply, the flip-flops are most unlikely to resume the states they were in prior to shut-down. It is therefore arranged that, where such continuity is required, the state of the flip-flop is continuously written into a square-loop magnetic core, where the information will survive the effects of a loss of supply. When the supply is restored, this facility of writing information into the magnetic cores is inhibited and an interrogation pulse is generated. This pulse examines the states of the cores and sets up the flip-flops accordingly. The inhibition is then removed from the 'write' circuits.

It is realized that this is not an elegant method, since it involves using two separate memories. However, it is considered to be simpler and cheaper than conventional computer core store methods and is adequate for this purpose.

## 10. Play-back Machine

The play-back machine is of more rugged construction than the programming machine, and its Z-axis mechanism carries a rotating cutter instead of a stylus. The tape deck is of similar construction to that used on the programming machine, except that a flux-sensitive head is used in place of the more orthodox 'write' head.

To cut a die, all that is required of the operator is that he should be able to thread the magnetic tape through its guides, and correctly position the auxiliary speech tape. When the play-back machine receives a 'recorded instruction' code from the main tape it stops and attracts the operator's attention by sounding a buzzer or flashing a light. By pressing a button the auxiliary tape enables him to hear the programmer's voice and to receive setting-up instructions; these may be replayed as often as required if he fails to understand, or if the environment is noisy. When the machine is correctly set up, normal metal cutting can commence.

The heads are flux-sensitive and are similar in principle to second-harmonic transducers. They are energized at 230 kc/s derived from a crystal-controlled oscillator and amplifier. The output from each head at 460 kc/s is fed to a tuned amplifier, the output of which controls a Schmitt-trigger type output stage. One of these amplifiers is used for each of the 16 channels of the tape to produce a standard-level signal for use in the logic circuits.

Figure 6 outlines the control circuits of the play-back machine, in which a Z-motion directional flip-flop is set forward or reverse by pulses derived from the appropriate tracks of the tape. 'Stop/go' control is exercised by a flip-flop set to 'go' by the onset of a Z-reverse pulse, and reset to 'stop' by a pulse from the Z-axis rotary photo-electric switch when a movement of 0·002 in has been completed. The rotary photo-electric switch is similar to that used on the programming machine.

The Y-motion directional flip-flop is set forward by a signal from any of the three pairs of Y-tracks on the tape, and reset (to reverse) by the combination of any



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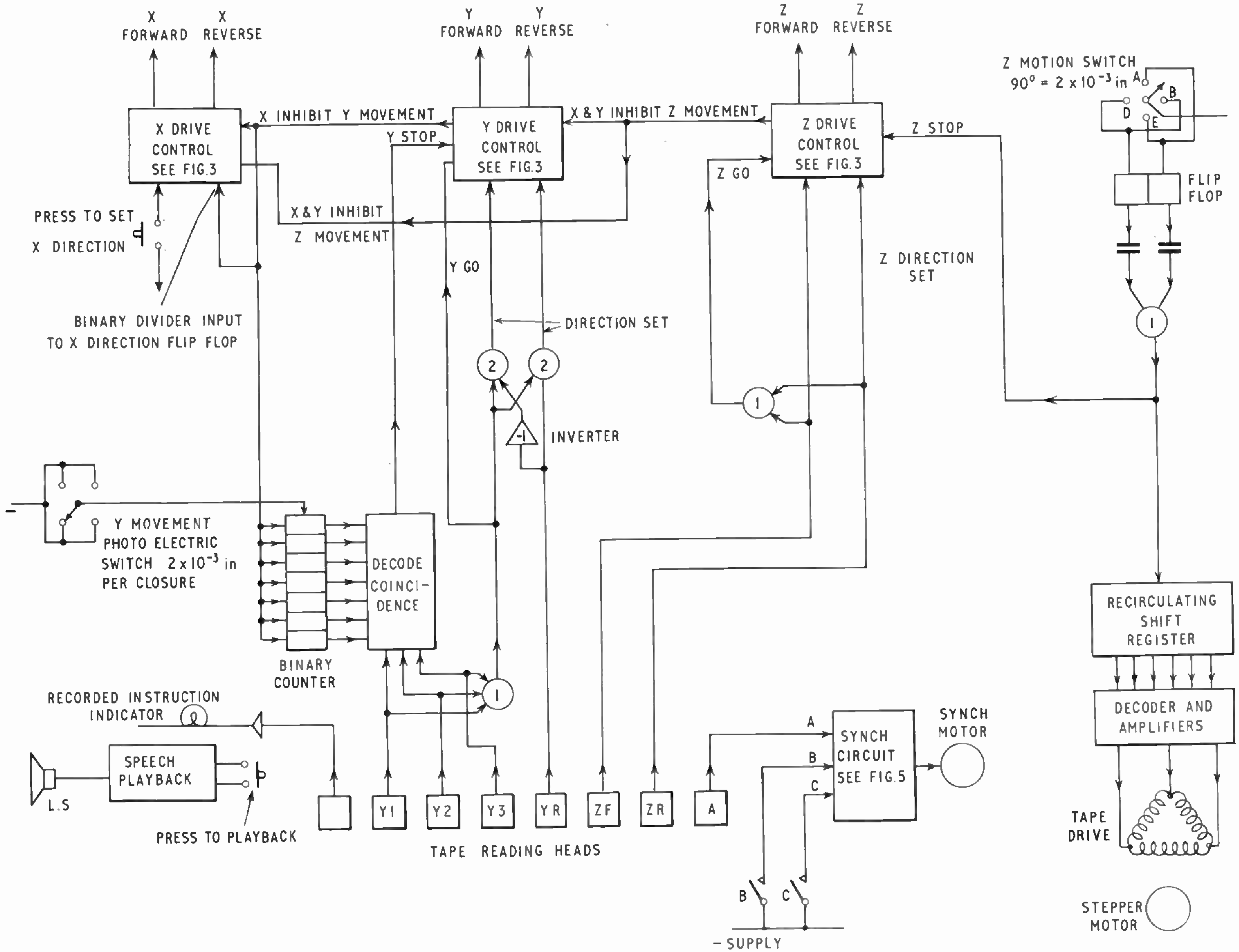


**Fig. 6.** Outline of control system for play-back machine.

one or more of the three Y-signals and the Y-reverse signal. A 'stop/go' gate is connected in such a way that Y-movement can be interrupted by the arrival of a pulse on the Z-axis; otherwise it is controlled by the stop/go flip-flop in the Y-increment circuit, similar to that used on the programming machine.

In the X-drive condition the directional flip-flop is set at the start of the program. Its direction is changed at the end of each line by the operation of the Y-axis. As was the case on the programming machine, the X-'stop/go' flip-flop causes movement in the X-plane when there are no requirements in either the Y- or Z-planes.

On the play-back machine, some safeguard is provided against cutter failure. The rate at which a machine-tool cutter wears out depends upon many factors; in this application it is quite possible that it might have seen service before a particular work-piece was begun. For these reasons it was decided to make the play-back machine self-protective in this respect.

When a die-sinking cutter wears, its reduced efficiency means that, for a given feed rate, it becomes more difficult to turn. This puts greater loading on the motor. On this machine, when motor loading reaches a given magnitude work is suspended and a signal is generated to call the operator's attention. When the cutter has been changed, he can restart the machine which then proceeds normally. Apart from these differences the play-back machine is similar to the programming machine.

## 11. Maintenance

The circuitry has been accommodated in unit blocks so that fault location is simplified and unit substitution prevents long operational delays. Test points are provided on each block, and a simple built-in testing device permits easy location of the faulty unit. The faulty unit can be returned to the makers or to an electronics workshop for repair.

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3. *Mullard Sub-assembly Bulletin*, No. 3, Section 4, February 1961.

*Manuscript first received by the Institution on 9th January 1963 and in final form on 31st October 1963.* (*Paper No. 900.*)

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## DISCUSSION

*Under the Chairmanship of Professor A. D. Booth*

**Mr. J. R. Arrowsmith:** I should like to ask the authors if they could tell us the advantages of their system over ordinary copying machines, as it seems to me to be doing precisely the same job.

As a second query I should like to know how the machine can make a vertical cut. Is it true to say that there is no speed control in the vertical axis and if not how is the feed-rate maintained?

**The Authors** (*in reply*): One of the main features of our system is that it enables skilled craftsmen to be used to greater advantage; they can work more quickly because they are released from mechanical limitations of cutter loading. Another advantage is that metal cutting may be done at a time and place both remote from the model. This is of particular interest to firms having overseas subsidiaries, as it may be much cheaper to transport a reel of tape than a heavy and bulky master die.

Sloping contours are cut in a series of steps, each well within the limits of tolerance of the machine. The vertical increment is fixed at 0·002 in, but the angle of slope depends upon the amount of X-movement associated with each step. A vertical cut is accomplished by letting the amount of X-increment be zero. As explained in the paper, the use of an auxiliary tape drive is then necessary to move the tape on to the next command pulse.

The feed-rate in the vertical axis is included in the initial setting-up instructions, and maintained by the motor characteristics. In general, it is unnecessary to make repeated changes of feed-rate, but verbal instructions to this effect could be included in the programme if desired.

**Mr. J. J. Hunter:** Could the authors give a more detailed account of the factors which led to the choice of magnetic tape as against punched paper tape. This is an important decision for a simple system as it greatly affects the complexity, expense and reliability of the resulting equipment.

**The Authors** (*in reply*): We do not use any X-command signals. When no other commands are present, the machine automatically runs in the X-plane until another requirement arises. This implies that on the play-back machine the length of an X-movement is proportional to the length of tape between signals, i.e. it is being used somewhat in the manner of a tape measure.

Unfortunately both punched paper and magnetic tape are unreliable in this respect. Both have unacceptably large thermal coefficients of expansion, both are liable to mechanical distortion and paper has an unacceptably large hygroscopic distortion factor. Errors due to these causes are particularly important as they tend to be cumulative. Consequently some form of correction is necessary.



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The most unpredictable distortion was that due to mishandling. Experiment showed that only gross abuse produced permanent distortion exceeding 3%, and consequently the machine was designed to cope with errors up to this amount. However 3% of an inch amounts to 0·030 in. This set the limit to the maximum permissible distance between corrective pulses. Using punched paper tape, such packing densities are impossible, whereas on magnetic tape they present no difficulty.

Another factor influencing our choice of magnetic tape was the convenience of having as much information as possible on one spool. Using paper tape, the maximum number of commands per inch on any one track is about ten. Using magnetic tape about two hundred can be stored reliably, although in fact the highest density used in this system is about one hundred and sixty.

**Mr. D. W. Thomasson**: Experience with magnetic-tape flux-sensitive head systems has shown that packing densities greater than 100 characters per inch tend to give trouble when a 'stop-on-character' performance is needed, as in this case. Paper tape, giving a density of 10 characters per inch, would allow an appreciable simplification of the reading arrangements and avoid the 'drop-out' problem. In these circumstances, the use of magnetic tape does not seem to be as desirable as is suggested in the paper.

**The Authors** (*in reply*): We quite agree with Mr. Thomasson that 'stop-on-character' working is troublesome at the packing densities he mentions, but that is not the case in the system described.

On each of our three axes, a 'stop/go' flip-flop is followed by a directional flip-flop. After filtering, amplification and squaring the tape pulses are used to trigger the flip-flops; these retain the memory even if the pulse subsequently disappears due to over-travel of the tape transport.

The second part of Mr. Thomasson's question has a similar answer to that given to Mr. Hunter. We did, in the early stages, have difficulty with thermal effects associated with our read-out, but these have been cured by re-design of the channel amplifiers.

## STANDARD FREQUENCY TRANSMISSIONS

*(Communication from the National Physical Laboratory)*

Deviations, in parts in $10^{10}$, from nominal frequency for **March 1964**

| March 1964 | GBR 16kc/s 24-hour mean centred on 0300 U.T. | MSF 60 kc/s 1430–1530 U.T. | Droitwich 200 kc/s 1000–1100 U.T. | March 1964 | GBR 16 kc/s 24-hour mean centred on 0300 U.T. | MSF 60 kc/s 1430–1530 U.T. | Droitwich 200 kc/s 1000–1100 U.T. |
|---|---|---|---|---|---|---|---|
| 1 | − 150·0 | — | — | 17 | − 150·7 | − 150·2 | + 4 |
| 2 | − 149·9 | — | − 1 | 18 | − 150·3 | − 149·5 | + 5 |
| 3 | − 149·0 | − 149·0 | − 1 | 19 | − 150·0 | — | + 6 |
| 4 | − 149·8 | − 149·9 | − 1 | 20 | − 150·5 | − 151·0 | + 5 |
| 5 | − 150·5 | − 150·4 | − 1 | 21 | − 150·8 | − 150·4 | + 5 |
| 6 | − 150·9 | − 150·8 | 0 | 22 | − 150·5 | − 150·5 | + 4 |
| 7 | − 150·8 | − 150·2 | — | 23 | − 150·6 | − 150·8 | + 5 |
| 8 | − 150·2 | − 150·4 | — | 24 | − 150·7 | − 151·2 | + 2 |
| 9 | − 150·4 | − 150·1 | + 3 | 25 | − 150·3 | − 151·0 | + 3 |
| 10 | − 150·5 | − 150·3 | + 2 | 26 | − 150·5 | − 149·3 | + 5 |
| 11 | − 150·5 | − 150·5 | + 3 | 27 | − 149·1 | — | + 6 |
| 12 | − 150·2 | − 149·1 | + 3 | 28 | − 148·8 | — | + 7 |
| 13 | − 150·8 | − 150·4 | + 2 | 29 | − 149·5 | − 149·1 | + 7 |
| 14 | − 152·0 | − 151·4 | + 3 | 30 | − 149·0 | − 148·4 | + 7 |
| 15 | − 152·3 | — | + 3 | 31 | − 148·8 | − 150·8 | + 9 |
| 16 | − 151·8 | − 150·7 | + 1 | | | | |

*Nominal frequency corresponds to a value of 9 192 631 770 c/s for the caesium F,m (4,0)–F,m (3,0) transition at zero field.*
*The phase of the GBR/MSF pulses will be retarded by 100 milliseconds at 0000 UT on 1st April, 1964.*



World Radio History

# Report on the Symposium on 'Cold Cathode Tubes'

The first major function to be held by the Institution under its new title was the Symposium on 'Cold Cathode Tubes and their Applications' which took place at the University of Cambridge from 16th to 19th March.† Three whole-day sessions dealing with 'Physics of Operation, and Tube Development', 'Circuit Design and Reliability' and 'Applications of Cold Cathode Tubes' were held in the Lecture Theatre of the Cavendish Laboratory. This logical sequence met with general approval, the theoretical papers in the opening session providing a more than adequate background to those in the 'Applications' Session.

The Chair for the opening session was taken by Sir Nevill Mott, F.R.S., Professor of Experimental Physics in the University of Cambridge. Contributions in the opening session included two papers from the University of Swansea and one from the University of Sheffield on the physics of cold electrode emission in gas and the impedance characteristics of glow-discharge tubes respectively. Other papers discussed the design of several of the various types of cold cathode tube currently in use.

The second session revealed some interesting developments in circuit design particularly from overseas authors. Reliability was covered fairly extensively. The failure rate of a group of cold cathode trigger tubes was quoted as being as low as 0·05% per 1000 hours, while one particular life test lasting about 300 000 tube-operating hours revealed no failures at all.

Probably the most interesting session to engineers generally was that held on Thursday when applications of cold cathode tubes were described. These included switching elements in telephone exchange equipment, memory stores in a batch weight ratemeter, supervision of high-power burners for steam generation, the detection of explosions in aircraft fuel tanks and tone generation for electronic musical instruments.

Demonstrations in support of papers presented were given in the adjoining laboratory and attracted a good deal of interest from delegates. One demonstration given at the close of the final session in the Lecture Theatre was that of a portable electric organ described by Mr. H. v. d. Kerckhoff from Holland. The recital given on this instrument by Mr. C. C. H. Washtell was received with some enthusiasm by musically-knowledgeable delegates.

The sessions were preceded on the Monday evening by an informal reception at Downing College followed by an organ recital, also given by Mr. C. C. H. Washtell, in the College Chapel.

The Symposium Dinner was held on the Wednesday evening in the Hall of Downing College. Mr. J. L. Thompson, President of the Institution, welcomed the authors and delegates, particularly those from overseas, and reminded his audience that three of the authors were from Switzerland, the first occasion for many years at which we have heard papers from this country. Previous meetings held at Universities had been entitled Conventions; the President gave the dictionary definition of a 'Convention' as 'A formal assembly of persons for some known object', and a 'Symposium' as 'A convivial meeting for drinking, conversation and intellectual entertainment'. The term 'Symposium' was figuratively given to 'A collection of opinions delivered, or a series of articles contributed by a number of persons on some special topic' and he believed that this Symposium agreed with the latter definition exactly.

Dr. F. A. Benson, Reader in Electronics at the University of Sheffield, replied on behalf of the contributors. While reviewing recent technological advances he commented: 'The transistor, almost a scientific oddity twelve years ago, is now a major part of our industry. Transistor production in the U.S.A. rose from zero in 1952 to about 120 million in 1960. The transistor has replaced the valve for many purposes and the day of certain gaseous devices may almost be over. I was not surprised, therefore, to read from the preview of this Symposium, that the original proposal to arrange this function was not greeted with unanimous enthusiasm. In spite of years of intensive study, however, many unsolved gaseous-electronic systems remain as we are hearing this week.

'As time goes on a man must know more and more before he reaches the frontier and can break new ground. Thus the fraction that only one man can know diminishes so he gets an unbalanced view of the world he lives in. Many advances in technology come from applying ideas and techniques in fields other than those for which they were originally intended. This requires a width of knowledge which is progressively harder to acquire.'

Dr. Benson then congratulated the Symposium Organizing Committee for assembling such a suitable selection of papers.

Mr. A. G. Wray, Chairman of the Organizing Committee, in thanking Dr. Benson, referred to the planning of the Symposium and mentioned that a selection had been made of 31 papers out of a total of nearly 60 offered. He paid particular tribute to the admirable presentation of papers in English by overseas authors.

Other social functions included a cocktail party and buffet supper at which delegates were the guests of Pye Telecommunications.

Of the 133 delegates who attended the Symposium it was particularly encouraging to note that there was a high proportion of visitors from overseas, including France, Germany, Holland, Switzerland and the U.S.A. Average attendance for each session was about one hundred, a number which seems to be the optimum for encouraging discussion contributions. As always, the informal discussions after each session showed the great advantage of a residential meeting of this size.

The Officers and Council of the Institution wish to record their thanks to the Governing Bodies of Downing College and the Cavendish Laboratory for their ready co-operation in the many aspects which go towards the success of such a Symposium.

† Synopses of papers presented were published in the February issue of *The Radio and Electronic Engineer*.



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# Systematic Selection Procedures for Technical Courses

A meeting of the Education Group of the Institution to discuss Selection Procedures was held in London on 20th November 1963, at which representatives of a University, a College of Advanced Technology, the United Kingdom Atomic Energy Authority and the British Broadcasting Corporation discussed their particular methods of selection of students for technical courses.

As the Chairman (Dr. W. A. Gambling) remarked, selection of one kind or another forms an important part of all our lives. Disregarding Darwin's process, for which the time scale is rather long, one might say that the first important selection in British education, a rather controversial one, occurs at the age of eleven. (Incidentally, some of the criticisms of the selection procedures made at the meeting also apply to the eleven-plus barrier.) Selection for some form of higher education takes place at the age of about eighteen followed by selection for a job, and later for promotion, and so on. Selection also occurs in our social lives whether for election to a tennis club or even night club although the qualities on which we are assessed are rather different in these cases. Theologians tell us that selection even follows us to the grave where we face the sternest test of all.

Fortunately, however, the purpose of the meeting was to discuss only selection procedures for technical courses leading to professional qualifications. The first contribution was by Professor B. G. Neal of Imperial College of Science and Technology, London, who has made a systematic study of this problem over the last few years. Student selection in the Department of Mechanical Engineering has been based, for the past four years, on the following five criteria:

(*a*) A test paper in Mathematics and Physics.

(*b*) An essay paper.

(*c*) The number of O-level subjects passed at one sitting.

(*d*) An interview.

(*e*) The Headmaster's report.

A numerical assessment is made under each of these headings and the correlations existing between each of the five predictors and the total mark obtained in the first year examination has been studied. So far data are available for those students who took the 1961, 1962 and 1963 examinations. Using a linear regression analysis it was found that the multiple correlation between the best possible linear combination of the five predictors and the examination total was 0·6 for the 1961 examination, 0·4 for the 1962 examination and 0·26 for the 1963 examination. These, somewhat alarming, figures indicate that even with the optimum combination of predictors the reliability of the selection process decreased dramatically, and unaccountably, over the period studied. The analysis showed that for the 1961 examination the mathematics and physics test was the only significant predictor but for 1962 the headmaster's report was far more significant and for 1963 none of the predictors was useful. Some of the predictors even produced a negative correlation although in most cases the deviation in the results exceeded the mean value.

Professor Neal concluded that success in an engineering course probably depends on many factors which are difficult to define and even more difficult to measure. He suggested that selection procedures should be constantly reviewed and changed, and data on the predictive value of many types of test should be collected even if these are not actually used in selection. Selectors normally also consider factors other than the ability to pass an examination (strengthening the rugger team or college choir?) but nevertheless the failure rate must be kept reasonably low.

Dr. R. T. A. Howell of Brunel College of Advanced Technology also expressed dissatisfaction with existing procedures and outlined some of the inherent difficulties. These include the fact that a student undergoes a sudden transition from school discipline to the college atmosphere of relative freedom and responsibility for his own activities and education, and is perhaps living away from home for the first time. Furthermore, for most students engineering is an almost unknown discipline and they are unable to judge whether or not they are making a wise decision. Dr. Howell suggested that the three main qualities to be sought are academic ability, a firm interest in the subject, and sufficient stamina to maintain the intensive pace of a modern technical course. However, he agreed with the findings of Professor Neal that there is no reliable method of assessing these factors. The usual three yardsticks of potential ability, namely examinations results, headmaster's report and interview are often inconclusive. Perhaps most emphasis should be placed on the interview since it provides direct contact with the candidate and enables the selector to assess, in particular, the candidate's degree of interest in his proposed career. This point was stressed since a strong motivation is necessary to surmount the various difficulties which may arise during the course. Interviewing, however, is a difficult art and the result depends a good deal on the skill of the interviewer.

While the academic representatives were somewhat pessimistic the next two speakers were more hopeful. Admittedly their task is a little easier in that they have to select a relatively small number from a large group of applicants, and they can therefore apply much more detailed tests than a University or College selector. Mr. H. Arthur, of the U.K. Atomic Energy Authority, described the selection of craft apprentices, students sponsored by the Authority for a degree or Dip. Tech. course, and employees for University or sandwich courses.

Craft apprentices are assessed on a headmaster's report, performance in an aptitude test and an interview. There is a certain amount of flexibility in that transfer to a higher (up to National Certificate) or lower grade course is possible. There are initially about 400 applicants of whom perhaps 20 are finally selected by a fairly rigorous process. In the student category considerable importance is attached to the headmaster's report and after interview and an aptitude test about 50 applicants (who have satisfactory A-level results) pass to a final group selection procedure. This lasts a day and a half and includes informal talks, interviews, group discussion and a works visit. The selection of employees for courses of full-time study depends on the recommendation of section and departmental heads who have known the applicant for some time, and an interview. Awards are only made to those who have an outstanding record of previous full-time or part-time education and a clear potential for work at professional level.

The selection of technical trainees for Higher National Diploma sandwich courses, and of employees for Dip. Tech. courses, by the British Broadcasting Corporation, as outlined by Mr. W. K. Newson, is as careful as that described by Mr. Arthur. The failure rate is thus correspondingly low. The required academic qualifications are higher than those normally required by the Colleges and successful applicants must display an active interest in one or more aspects of broadcasting. This need for some degree of motivation had earlier been mentioned by Dr. Howell. Considerable importance is attached to the applicant's personality and character. Various kinds of intelligence test have been experimented with over long periods but none has yet been found to give a satisfactory result. Mr. Newson, after long experience, thinks that the best criteria are A-level results, headmaster's report and interview by a selection board.

The vigorous discussion which followed was opened by Group Captain J. H. Stevens who outlined selection procedures used in the Royal Air Force. For Service personnel many other factors, in addition to those already mentioned, must be taken into account, such as initiative, leadership, etc. These latter qualities are often necessary in civilian life but are not as vital as they are in the Services. During the discussion it soon became evident that there were nearly as many opinions as speakers, but there was certainly agreement that most present-day selection procedures, particularly for College and University courses, are quite inadequate. Few detailed studies have been made and the quantitative data of Professor Neal are most disturbing. Even these are incomplete in that, ideally, it is necessary to know how the rejected applicants would have fared. Short of admitting to courses those whom the selectors think will fail, there is no easy method of obtaining such information. However, the results of such an experiment might well be very significant.

Some of the difficulties of selection have already been mentioned, namely that the applicants are young people undergoing rapid development; a few have reached their academic ceiling, some opt for a higher education who are not really suited to it, some make an unwise choice of subject. Furthermore, attempting to estimate the ability of an immature mind four or five years hence is a formidable task indeed. Even if a reliable selection procedure could be devised which would give a low examination failure rate, and ensure that all those who would profit from a higher technical course are given the opportunity, this would not be the complete answer. The ultimate goal is to predict, not only the performance in a given series of examinations, but the value of an applicant as a professional engineer in later life. We are still a very long way from this enviable state.

In the meantime Universities and Colleges are faced with the problem of what kind of selection procedure to use, bearing in mind the fact that no method will be blessed with any high degree of success. Some Universities depend mainly on the A-level results and dispense with an interview. Since there are often from five to ten applicants for every place, each of whom would take up the time of, say, two staff members for half an hour, this is at least a time-saving solution. It is not practicable, except perhaps for the Oxford and Cambridge Colleges, to use one of the more rigorous interviewing and selection procedures such as that described by Mr. Arthur. This is because with such a method, the high standards set, although giving a high pass rate among the candidates selected, would cause the rejection of many who could profit from a higher education. Most institutions are likely to continue existing methods and one can only echo Professor Neal's plea for a constant review of selection procedures and the keeping and analysing of selection records.

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# A Digital Data Link for Aircraft Communication

By

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*Presented at a meeting of the Institution in London on 14th November 1962.*

**Summary:** The main features of a digital data link system for communication with a number of aircraft over a single radio frequency by time division multiplex are described. Part 1 describes a ground-to-ground link for conveying the coded messages from the originating centre to the radio transmitter site. Six telephone lines are used, each line carrying 19 parallel 50-baud voice frequency telegraph channels. The equipment is engineered with transistors and printed circuit boards and some figures of fault incidence and reliability are given. Part 2 describes the airborne decoder. Message recognition and means to extract the digit synchronization are described and the design considerations for filtering and coupling the message from the receiver to the decoder are given. Two digital-to-analogue conversion circuits are outlined, one to provide a 400-c/s voltage proportional to the digital quantity, the second a means to generate an angular quantity as a shaft position within the decoder. Brief mention is made of the engineering methods adopted to give the necessary reliability.

## 1. Introduction

The rapid growth of aviation is placing increasing demands on the existing air traffic control system. By introducing modern methods of automation traffic may be made to flow more efficiently and accident risks reduced; modern radars, special displays, semi-automatic data extraction and data computers are gradually being used by the system and each contributes to improved overall performance. One of the limiting features in the present system is the voice communication which is used between ground and aircraft; this makes inefficient use of the limited radio frequency spectrum, and imposes a considerable load on the aircrew, much of it of a routine character. This paper describes a digital communication system which can be used to supplement a voice system and mitigate these undesirable features.

An air terminal could be expected to accept one aircraft for landing and one for take-off every minute, so that the data processor and link should have a minimum capacity for 120 aircraft per hour. Some capacity is required for aircraft passing over, and, to allow for growth, the system should be capable of handling about 500 aircraft. The equipment to be described has this order of capacity and speed.

The first part of this paper is devoted to the ground equipment whose task it is to take the aircraft messages from the processor and pass them over telephone lines to the site of the ground transmitter. The second part is concerned with the equipment used to decode the digital information which is derived from the aircraft receiver. This involves synchronization, ability to recognize addresses, storage of the message and provision of a suitable display.

Messages are sent to all aircraft on one frequency and each aircraft selects its message from the complete message train by recognizing its own address. A digital system has the great advantage that the digit rate may be made faster or slower, depending on the bandwidth of the communication channel, and the designer is therefore in the position to select the speed for the particular application he has in mind, while keeping to the same basic coding plan. Speeds ranging from 5000 bits per second to 750 bits per second are suitable for a short-range link which would use u.h.f. or v.h.f. while 25 bits per second would be suitable for a long-range h.f. link. A fixed format message is used and time is divided between air-to-ground message and the ground-to-air message. When an aircraft recognizes its address a following message frame is available to it for sending back its own message to the ground.

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## Part 1. GROUND-TO-GROUND DATA LINK

### 2. Data Transmission over Telephone Lines

It was decided at the outset of the project to carry the information on telephone lines and to utilize to the full the very extensive G.P.O. network. This meant that it was essential to select a modulating means acceptable to the present system and capable of working over all kinds of plant. Two choices were available, firstly, a carrier could be selected within the 3 kc/s speech band and this carrier modulated as fast as possible, until the side-bands which spread out from the carrier could no longer be faithfully reproduced. This is known as the serial stream system. The second method was to use a number of carriers side by side within the band and modulate each carrier somewhat slower; this is the parallel voice frequency telegraph technique.

The group or envelope delay for a typical telephone circuit has a minimum value around the mid-band frequency and increases near the upper and lower frequency limits. Typical average delays for 100 miles of circuit including amplifiers are as follows: audio circuit, up to 15 ms; carrier or coaxial path, less than 1 ms. The spread for the two cases is approximately 2 ms for the audio circuit and 1 ms for the carrier circuit where the terminal equipment contributes largely to this figure.[1] Pulse distortion may be calculated for a given delay distortion or spread in envelope delay,[2] but the type of signal and circuitry determines the amount which the system can tolerate. In most cases the bit length must be greater than one third of the delay distortion,[3] for satisfactory operation. A 400-mile circuit with 4 ms of delay distortion would limit the maximum bit rate to approximately 750 bits per second. Clearly a 50-baud channel will be unaffected by delay distortion of a normal telephone circuit. A serial system is limited in speed by the delay distortion but the capacity of a parallel voice frequency telegraph is determined by the number of 50-baud channels that may be stacked side by side within the telephone band. A 24-channel equipment is available commercially and can therefore, offer a capacity of 1200 bits per second, which is nearly twice that of a serial system operating without special circuits. Where a message is distributed over 24 channels some compensation for the time of arrival of the digits from channels at the edges of the band may be necessary, but this is a problem associated with the digital part of data handling and not with the distortion of individual digits.

If we assume that the noise on the lines has a uniform spectral response, then the noise power in a channel is proportional to the bandwidth. Dividing the total bandwidth by the number of channels $n$ and dividing the signal power by $n$ produces identical channel signal/noise ratios for the serial and parallel systems. Good circuits have a noise level of $-47$ dBm so that for a 100 μW signal to line, the signal/noise ratio is 37 dB in both cases. A narrow band frequency shift system will theoretically produce an element error rate,[4]

$$p = \tfrac{1}{2}\exp(-S/N)$$

and for $p = 10^{-6}$, the value of $S/N$ is 11 dB. This means that the circuit would have to be 26 dB worse than normal to produce one bit in error per million bits. Good results can be expected from both systems with uniform noise and in practice the serial system should have an advantage, as the peak phasing of the multi-tone telegraph limits the power to line to 100 μW. The serial system may use the full power of 1 mW.

A telephone system has associated with it contact noise as well as thermal noise and for normal operating conditions the most significant causes of errors will be the impulsive portion of the interference. Before detailed analysis can be made it is necessary to get a picture of the distribution of peaks above various voltages, their duration and energy content. Unfortunately, this type of data can not be obtained with any consistency and an experimental procedure must be adopted for comparision of various systems. This may be accomplished by recording samples of actual interference on magnetic tape and injecting into a noiseless channel.

In general the effect of reducing bandwidth is to reduce the peak magnitude of the impulsive interference in direct proportion and to spread out the impulse in time also in proportion to the bandwidth. Filtering is liable to produce tails on the impulses and if these occur close together a build-up can occur which will produce errors. Wide-band f.m. limiters are usually more effective against impulse noise than those using just the minimum of bandwidth. It would appear, therefore, that unless there is bandwidth to spare or unless the designer is prepared to increase the digit length the serial or parallel systems are likely to give comparable results.

Three types of modulation—amplitude modulation, frequency modulation and phase modulation—have all been successfully demonstrated. The two angle systems have either two tones or one tone present at all times and are, therefore, likely to be better than a symmetrical amplitude modulation system which produces an error for interference exceeding half the difference between the amplitude of the carrier for marks and spaces. The best choice is probably phase modulation of a single tone, but due to the likelihood that an impulse will produce a double error, any error checking



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code used must be capable of detecting double errors.

A complete data system must employ a means of digit synchronization, and the start and end of message must be marked out in time. These requirements tend to impose special limitations on the form of the message for the serial stream system but the parallel system can use one of the available channels for this purpose, quite separate from the message.

The study tended to conclude that, for a data system to transmit about 1000 bits per circuit over a distance of 400 miles, a parallel system employing about 200 bits per channel using either frequency shift or phase modulation would be reliable. A fully developed voice frequency telegraph system, recently transistorized, was available and the complexity of using a large number of 50-baud channels was accepted. The basic receivers and transmitters were admirably suited to supervisory control and a full range of monitoring and automatic line switching equipment was available. Nineteen channels are used, one of these carries the start pattern exclusively and the message is formed into a block nineteen by four digits.

### 3. Description of the Link

The transmitting terminal consists of a multiplexer which accepts aircraft messages and prepares them for transmission down one of three telephone lines. Each telephone line is modulated by a voice frequency telegraph equipment (v.f.t.), which consists of nineteen 50-bit, 120-c/s wide tone channels. The receiving terminal is very similar except that the multiplexer is now called a demultiplexer and the v.f.t.'s contain mainly receiving units.

Figure 1 shows a simplified diagram of the link. The messages are received from the data processor, in parallel on 66 wires. These are accepted on demand by the input unit and are stored in the input register. The first thing that happens is that the message is circulated in the input register and 6 parity digits are added. These parity digits stay with the message for checking purposes throughout the system.

Having inserted the parity digits, a line register is selected out of a choice of three and the 72 digits are transferred to form a block of $4 \times 18$ digits. A start pattern '1' and three '0's is added in channel 4 to give a total of $4 \times 19$ digits. Each row of the line register is connected as a shifting register and each row has its 50-bit voice frequency telegraph channel. Four shift pulses are applied to the 19 shift registers, spaced 20 ms apart and the voice frequency telegraph channels are all modulated simultaneously to produce the frequency shifted tones. These are ultimately filtered and combined to provide a composite signal to the telephone line.

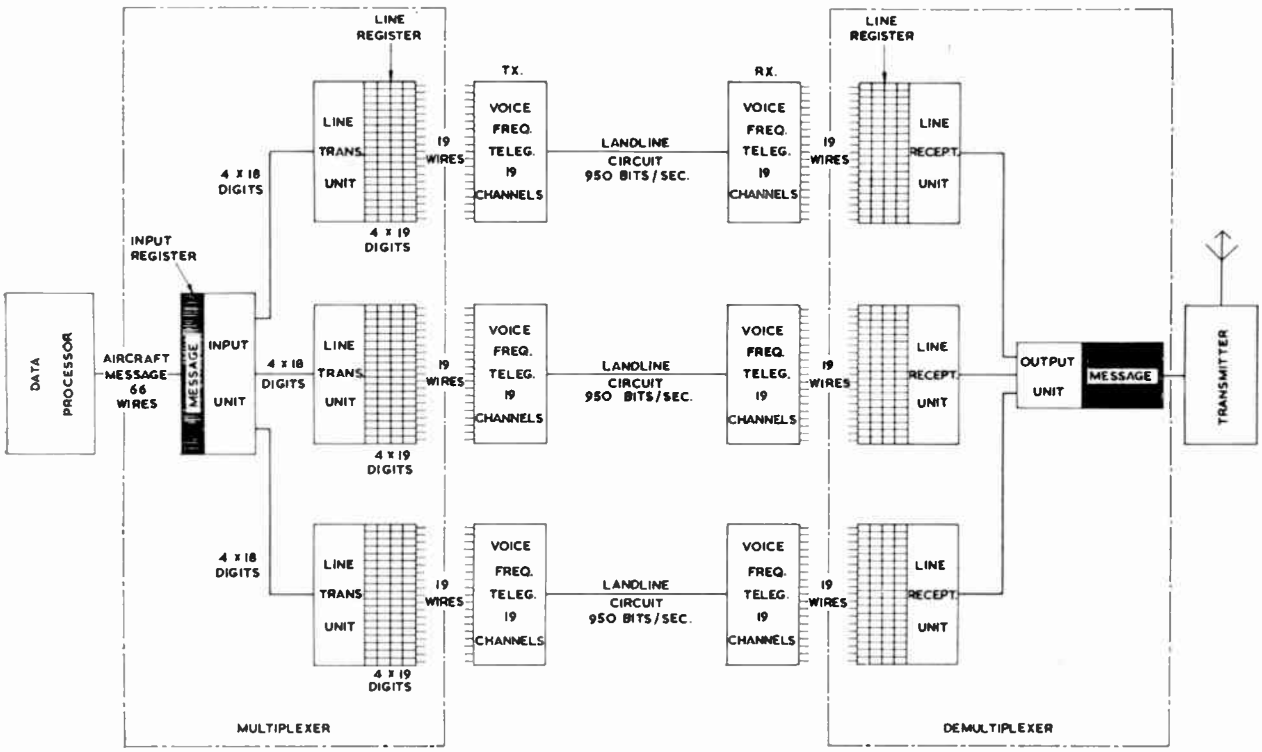


**Fig. 1.** Block diagram of the ground-to-ground link.

Directly the input unit has transferred its message into a line register it may accept another message and transfer it into the next line register. When everything is operating correctly three landline circuits are used simultaneously, but in the event of failure the system may be reduced to a single circuit.

On arrival of the message at the receiving voice frequency telegraph rack each narrow band channel receiver has to decide by measuring the frequency of the received tones whether a '1' or a '0' has been sent. Clearly any modulation or demodulation processes en route must not introduce frequency errors. To overcome this difficulty the v.f.t.'s inject a 300-c/s tone into each line and when this is received a correction is applied to the complete set of channels to compensate for any unwanted frequency shift.

It is the reception of the first '1' in the start channel which commences the decoding. Each channel is inspected 10 ms later at the mid-point of the digit and a decision is made as to whether a '1' or '0' has been received. Four shift pulses are again generated and the corresponding line reception register block should then contain the identical digit pattern originally held by the line transmission register. Channel delay measurements on typical routes have shown a maximum overall variation of $\pm 2{\cdot}5$ ms when both the telephone lines and the v.f.t.'s were considered together. It was possible to choose the start channel with an average delay time and acceptable operation was obtained without the need for separate compensation of delays for each channel.

Once the line reception register is full the four start pattern digits are checked together with the site address. Then the block is formed into a serial message and a check is made on the parity relationships, to see if the message has suffered during its transmission over the landlines. When this is completed the output unit is informed that a correct message is available for transmission by the radio transmitter.

Two controlling crystal clocks are used in the system. The first controls the rate at which messages are accepted from the data processor and the second the rate at which messages are transmitted by the output unit to the radio transmitter. By accepting messages at a slightly slower rate than that sent out over the air, messages cannot pile up at the transmitter and a little spare time is always available.

It is the output unit which co-ordinates the three separate lines and takes care of the timing. Under ideal conditions the message blocks follow each other in rotation down the three telephone lines. In practice differences in time delays could occur on the lines due to different distances involved. This lack of symmetry is accommodated by a queueing circuit which is actuated by the ready pulses initiated by the checking programmes. For every 150 output messages that the output unit wishes to transmit only 149 arrive. The queue enables this 'spare' time to be accumulated until a complete slot is available. This spare slot is filled by a choice from two test messages and it is these messages which may be decoded on demand in the aircraft to check that the equipment is operating correctly. The queue selects a particular line unit and the message is transferred into the output register. Just prior to transmitting the message the synchronizing burst and start pulse for the ground-to-air message is generated and then the actual message follows. During read-out there is a final parity check. If this proves to be wrong no action is taken beyond illuminating alarm lamps. In addition to the message output a control waveform is provided to switch the transmitter on.

### 4. Equipment Design and Reliability

Both the multiplexer and demultiplexer are housed in six-drawer cabinets. The drawers pull out on runners and all connections are accessible. The basic printed wiring board is inserted into two 15-way sockets and special keyways prevent the insertion of a wrong board. Germanium transistors are employed throughout and the circuit design will tolerate a wide range of drift in the individual components before upsetting the operation. The v.f.t. equipment is also completely transistorized and as the receiving channels are not required to drive a teleprinter, the output relay has been replaced with a transistor drive circuit and the reliability of the equipment improved. The v.f.t. is a thoroughly developed equipment and its maintenance is straightforward, but to help diagnose faults in the data handling a test rig has had to be designed. This instrument supplies test patterns and accepts digital information from various parts of the system. The patterns appear on indicator tubes and by observation a fault can in general be located to a single board. The operator of the link is given a console which shows diagrammatically the flow of data, together with switches and spare telephone lines. A status report on the lines and radio transmitter is sent back from the remote sites, using single telegraph channels and these serviceability lights appear on the console. Standby equipment and lines may be substituted in cases of failure and messages re-routed if necessary.

One 2850-bit system employs four thousand transistors and during the development stage detailed records of faults were kept. The results shown in Table 1 have been divided into two equal time periods each of 6500 hours. Reading the top line we find that eight transistors failed in the first 6500 hours and only a further two failed in the second 6500 hours. The percentage failures per 1000 hours for the transistors is approximately $0{\cdot}02\%$.



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**Table 1**

| Component | No. in system | Total faults at 6500 hrs | Total faults at 13 000 hrs | Faults % per 1000 hrs for 13 000 hrs |
|---|---|---|---|---|
| Transistors | 4234 | 8 | 10 | 0·018 |
| Diodes | 4635 | 1 | 1 | 0·0017 |
| Capacitors | 7891 | 1 | 1 | 0·001 |
| Inductors | 460 | nil | nil | nil |
| Resistors | 19 069 | nil | nil | nil |
| Plugs | 750 | nil | nil | nil |
| Transformers | 692 | 1 | 1 | 0·011 |
| Board connectors | 744 | 1 | 3 | 0·03 |
| Rectifiers | 72 | nil | nil | nil |
| Potentiometers | 240 | nil | nil | nil |
| Soldered joints | 115 000 | 5 | 9 | $6 \times 10^{-4}$ |

The junction type diodes have proved very reliable and printed board connectors fairly satisfactory provided the printed board is accurately made. Soldered joints are a major source of trouble with nine failures in 13 000 hours, giving a fault rate of $6 \times 10^{-4}$% per 1000 hours. Dip soldering is not employed on these boards. Since the laboratory development stage, slight improvements in engineering and inspection have been made and it is now thought that a mean time between failures of 1000 hours will be attained for each 2850-bit system. With regard to the time taken to locate and correct a fault, this is expected to be 15 minutes but enough operational experience has not yet been obtained to substantiate this figure.

### 5. Messages Lost due to Landline Faults

Together with equipment faults, messages lost due to landlines have been recorded. One circuit used an audio line 100 miles in length carrying only the v.f.t. tones and the other a carrier circuit of similar length carrying a number of unrelated tones besides the v.f.t. tones; neither were switched circuits. The number of complete breaks was quite high especially outside normal working hours when repair and maintenance work was probably in progress. These have been conveniently ignored and the total of messages lost has been calculated for both parity and address faults that occurred with an effective duration of less than three quarters of a second or resulted in a loss of less than ten consecutive messages. Most of the faults occurred during the working day which suggests that they were caused by cross-talk, switching and dialling impulses. The audio line was twice as good as the carrier circuit. One message was lost per 3 000 000 messages for the audio line and one message was lost per 1 500 000 messages for the carrier line.

## Part 2. GROUND-TO-AIR LINK

### 6. Aircraft Data Decoder

The data decoder receives the digital messages from the aircraft receiver and converts them into analogue form for display on the aircraft's instruments. Messages are individually addressed to particular aircraft and each message has an initial synchronizing signal plus a unique start sequence to enable the decoding to be accomplished.

Figure 2 shows a block diagram of the decoder. The first operation the decoder has to perform is to recognize that a message is present and to establish digit synchronization. Once digit synchronization has been established the digits are sampled at their greatest amplitude and the program which follows has then the best chance of completing the decoding correctly. These two initial processes may be combined if the message preamble consists of a number of digit reversals. A simple tuned circuit accomplishes the recognition and the phase of the signal established within the tuned circuit may be used to generate a set of pulses for sampling the digits. When one isolated message is received there is likely to be a d.c. component present either due to differences in the oscillator frequencies or to overload. This d.c. component must be removed otherwise the sampling will be unreliable. In this application where the modulation process is frequency shift keying the output from the receiver discriminator is connected to the decoder via a capacitor and a pair of diodes set at $\pm 5$ V. The signal which is itself $\pm 5$ V is driven between the defined voltages and the restoration is established about zero voltage. Clearly this is satisfactory for the reversals during the synchronizing period but may not be for the message with noise on top of the pulses. However, it was found possible to select the time constants to give reliable operation for the fixed message length used.

A low-pass filter with a cut-off set at half the bit rate is incorporated after the a.c. coupling.[3] This filters a rectangular pulse to produce a shape very close to Gaussian. Compared to a circuit which would integrate the pulse over its duration, the loss in signal/noise ratio is negligible. Also the 10% overshoot has little or no effect on the discrimination process for the subsequent pulses. Noise is thus limited within the signal channel to near optimum and as the synchronizing channel consists of a tuned circuit with a $Q$ of 10 or more, noise is further eliminated, until a condition is reached where the synchronization is still satisfactory after the messages become unreliable.

The sampling pulses from the synchronizing circuits are generated at the position in time when the output



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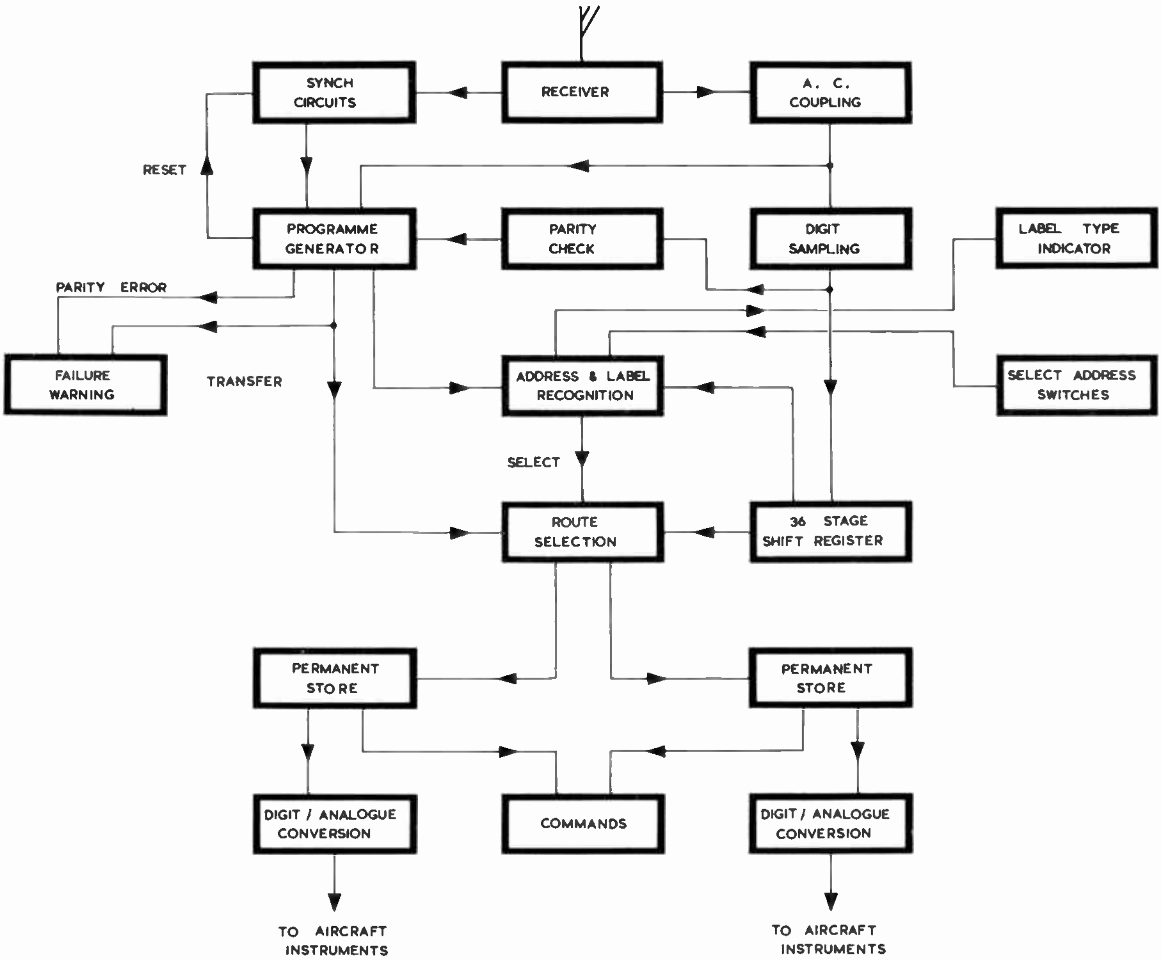


**Fig. 2.** Decoder block diagram.

pulse from the filter is at its maximum amplitude. Sampling pulses and signal are combined within a sampling gate and a decision is made as to whether a '1' or '0' has been received. The '1's and '0's are then passed into a shift register. Just previous to this operation, the program is given a 'go-ahead' sign by the receipt of a start pulse. At a time when the aircraft address and label are present in known positions in the register the address and label circuits come into operation and if both are correct the route selection gates are set to route the information into one of two permanent stores. Information proceeds through the shift register until a full message is positioned in it. Included within each message are parity check digits and these must be correct before a read pulse is generated which transfers the message to its appropriate store. To the stores are attached the digital-to-analogue conversion units which provide 400 c/s a.c. voltages to actuate the aircraft instruments. Discrete commands are fed to relays within the decoder and these in turn operate a voice box which draws the pilot's attention to the command by a pre-recorded sentence.

Without the confidence given by voice contact from the ground, some means of checking that the equipment is working correctly must be provided. Firstly, a universal test message which commands all the instruments to read a definite value is sent from the ground at regular intervals and this may be selected by the pilot. Secondly, a failure warning circuit is incorporated which will warn the pilot if he is not receiving good messages or if one or more of the power supply phases has failed. Parity errors in two consecutive messages or the absence of a good message for a preset time will also give a warning. This preset warning time may be one of several values ranging from 30 seconds to approximately four minutes.

### 7. Digital-to-Analogue Conversion—Linear

A number of digital-to-analogue conversion units have been designed for connection to the permanent



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stores. One such unit provides an output at 400 c/s whose voltage is proportional to the numerical value of the digital information. This has been designed using transformers wound to an accuracy of 1 part in $10^3$. Figure 3 shows a set of transformers with their secondaries connected in series. At any instant one half of the primary is energized, the choice being made by switches which operate from the permanent binary storage stages into which the digital message has been read. Each transformer therefore gives an output which in relation to the primary voltage may be either in phase or anti-phase, but which has a constant amplitude. A half digit is subtracted and a half digit is added. The turns ratio of each transformer is arranged to give the correct weight and, together with the lower transformer, provides an output equal to 127/128 to zero of the reference voltage. The reference voltage is fed into the decoder and the output is the analogue quantity referred to the reference voltage. Figure 3 shows a simulator for checking the analogue voltage against a calibrated potentiometer. The reference is fed across the potentiometer and the potentiometer is turned until zero error is obtained. In practice a servo motor is actuated by the error detector and the numerical value of the digital information appears on a scale. The design has been made for a secondary load of not less than 10 000 ohms and at balance this is, of course, very high. The transformer switching arrangement is exactly the same for all transformers and a symmetrical type transistor is used to do the switching. It has been found that in production this type of digital-to-analogue conversion circuit gives a resultant error from all causes not exceeding one part in five hundred and the quadrature error is small enough to cause negligible errors.

## 8. Digital-to-Analogue Conversion—Angular

A second type of conversion generates the required angle as a shaft position within the decoder itself. A synchro is mechanically coupled to the shaft and the angular information transmitted to the aircraft instrument in the normal way.

One revolution is divided into 256 fine parts, so that the three most significant digits of the eight digit binary coded message selects a particular 45 deg sector from eight possible sectors, the remaining five digits an angle within the chosen 45 deg sector.

Figure 4 shows the circuit where the horizontal stator winding is the cosine winding and the vertical winding the sine winding of a synchro resolver. If then, it is possible to put a 400 c/s voltage proportional to the cosine of the demanded angle (between 0° to 45°) on the cosine winding and a voltage proportional to the sine on the sine winding, and if the voltage appearing on the rotor winding shown parallel to the

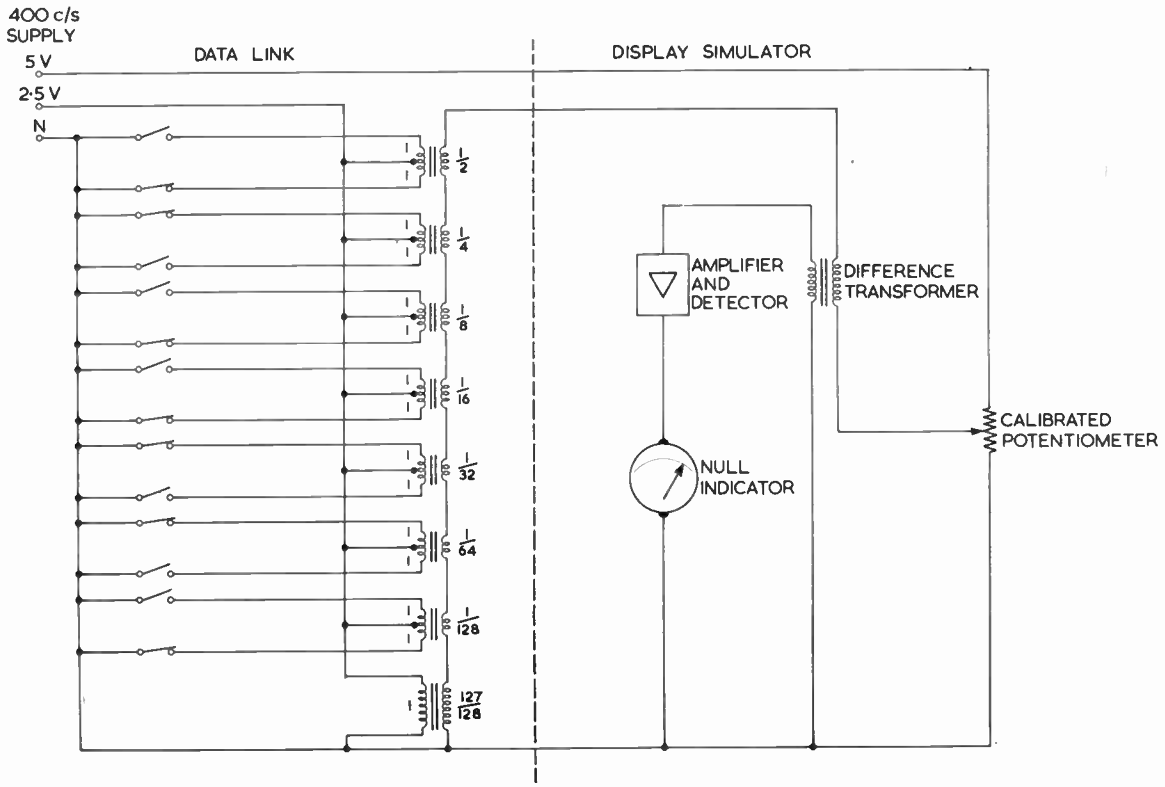


Fig. 3. Converting a binary coded quantity to a 400-c/s analogue voltage.



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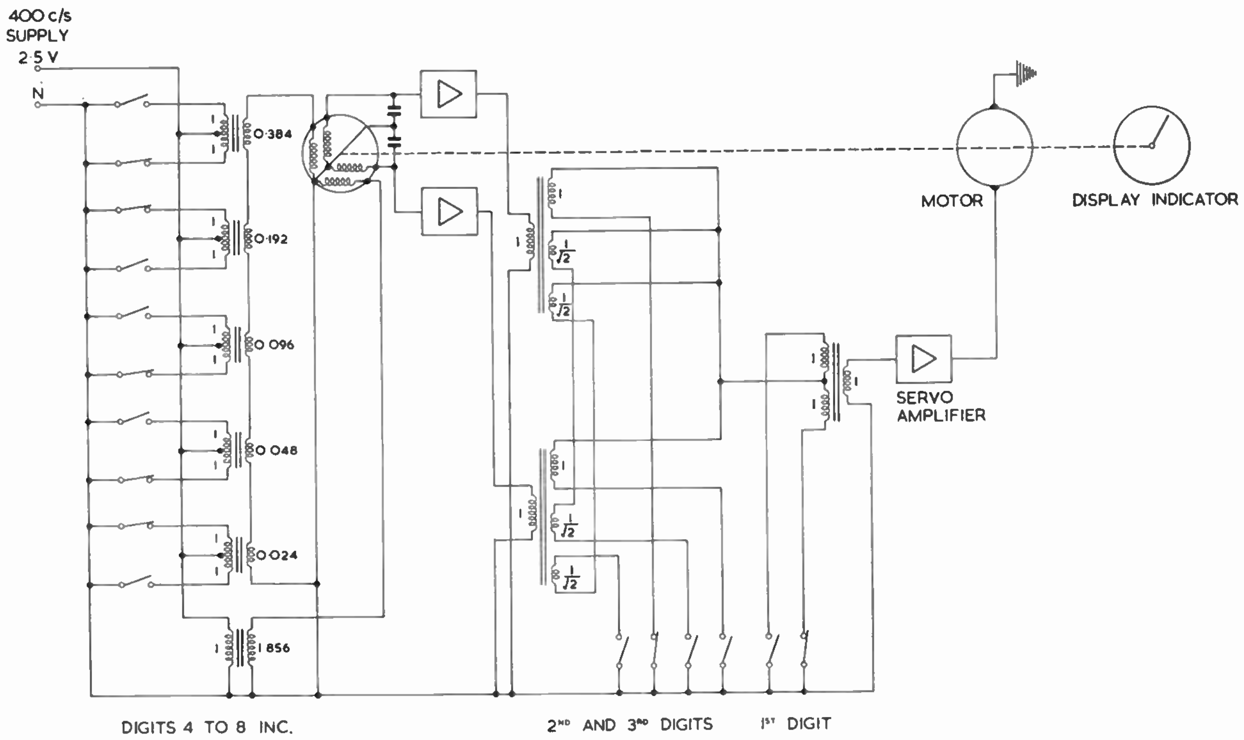


**Fig. 4.** Setting a shaft to a binary coded angle.

sine winding is applied through a servo amplifier to a motor which will rotate the winding, the motor will stop the resolver shaft at the demanded angle. The vertical drawn rotor winding traverses 0 deg to 45 deg, the horizontal winding 90 deg to 135 deg and the vector sum of the two windings, 45 deg to 90 deg. Reversing the sense of the horizontal winding on the rotor and adding it to the other will set the angle between 135 deg and 180 deg. It is possible by combinations of the rotor windings, either alone or in pairs, to select the angle between 0 deg and 180 deg and the most significant digit will select the correct half of the circle.

The transformers select the windings and their polarities are actuated by the transistor gates driven from the permanent stores. When vector addition takes place with the two rotor windings the servo gain is adjusted by a $1/\sqrt{2}$ in the transformers.

Now, fortunately, over an angle of $\pm$ $22\frac{1}{2}$ deg

sin $\theta$ is approximately equal to $\theta$ and
cos $\theta$ is constant.

The mechanism is, therefore, one of addition for $\theta$ using the same transformer technique as was used in the linear case. This voltage is applied to the sine winding of the resolver and the cosine winding voltage is constant.

The actual angle defined is exact at the angles 0 deg, $+22\frac{1}{2}$ deg and $-22\frac{1}{2}$ deg. Maximum errors of $27\frac{1}{2}$ min occur at the angles $\pm$ $11\frac{1}{4}$ deg. This is adequate accuracy for the smallest increment of 1·4 deg.

## 9. Airborne Equipment Design

The equipment has been designed using germanium transistors with a certain number of silicon diodes employed in the decoder. Circuits are designed to operate satisfactorily over a temperature range of +65° C to −25° C, and where the environment is such that these figures are likely to be exceeded the decoder is packaged in a pressurized double skin cylindrical container and cooling air from the aircraft system is blown between the outer and inner skins to ensure maintenance of the operating temperature.

The circuitry is arranged on a hexagonal frame around a central air duct, (Fig. 5). Radial fins from the duct divide the frame into six sectors and the sub-assemblies and printed circuit boards are shaped to fit into these sections, thus giving a very rigid and compact assembly without in any way affecting accessibility. Cabling is carried in these vertical ducts formed by the shape of the boards and sub-assemblies. At the base of the central air-duct is a fan which circulates air up through the duct returning over the layers of circuit panels. The air in the container is dried and maintained under pressure of about five pounds per square inch. The boards are wrap-wired to connecting bars on the frame.



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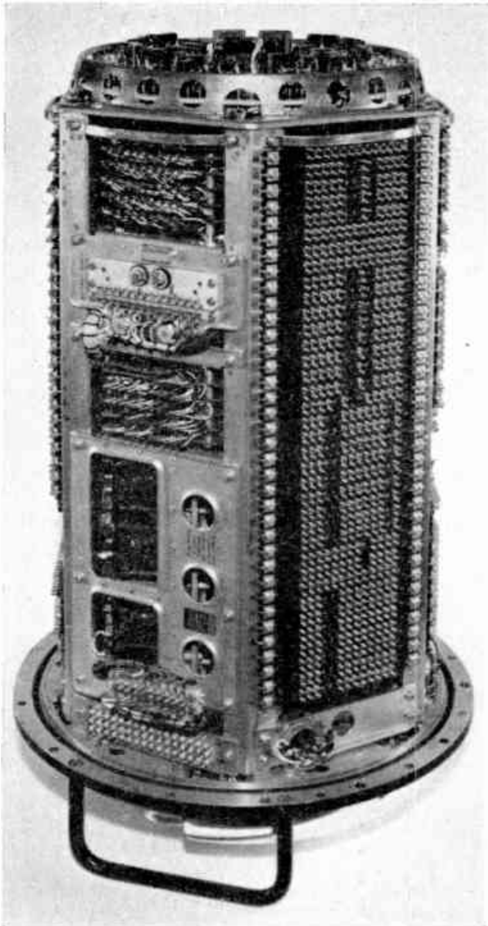

Fig. 5. The airborne decoder.

A complete set of test equipment for checking the performance of the data decoder is provided. In its basic form the test equipment is mounted on a trolley which is wheeled out alongside the aircraft so that the airborne decoder is checked *in situ*. The same test equipment can be used to test the decoder after removal from the aircraft, for example, in a repair workshop.

## 10. Future for Aircraft Data Links

A high speed ground-to-ground and ground-to-air data link has been designed using the most modern techniques available to the designer. It is not expected that civil aviation would use this equipment in its present form as extensive repackaging would have to be carried out to meet the requirements for installations in civil aircraft. The bit rates, for example, will be different and for economy reasons the total digital storage would be very much less. One advantage of the digital system described is that the basic data rate is easily adjusted to various conditions. It should, therefore, be possible to provide a system with the ability to handle communications to aircraft at long range, together with aircraft close to the air terminal, by the use of a basic information coding plan.

## 11. Acknowledgment

The equipment described here has been designed by The General Electric Company Limited on behalf of the Ministry of Aviation. Certain aspects of the equipment are the subject of pending patent applications.

## 12. References

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*Manuscript first received by the Institution on 9th October 1962 and in final form on 13th November 1963. (Paper No. 901.)*

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# Radio Engineering Overseas . . .

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## FRAME DIFFERENCE TECHNIQUES IN TELEVISION

One proposed method of reducing the channel capacity requirements for the transmission of television signals over long distances is known as frame difference signal (f.d.s.) coding. This method implies the recoding of television signals so that only those parts of every frame are transmitted which are different in two consecutive frames. In order to assess the possible savings in channel capacity resulting from this technique, it is first necessary to measure and record the probability density of f.d.s. areas in actual television programme signals. An Australian engineer has shown that by using a vidicon camera tube as a storage and subtracting device, f.d.s. signals can be produced which are suitable for statistical analysis. The generated signals are further processed to obtain the frame-by-frame integral of the f.d.s. areas and the ancillary equipment required to carry out this operation is described. The data are recorded for subsequent evaluation.

"An experimental frame difference signal generator for the analysis of television signals", A. J. Seyler. *Proceedings of the Institution of Radio Engineers Australia*, **24,** No. 11, pp. 797–807, November 1963.

## NEW TRIODE-OSCILLATOR EQUIPMENT CIRCUIT

A new equivalent theoretical circuit diagram has been introduced in Germany when the triode is operated in class A, B and C conditions. The equivalent admittances are calculated from the geometry of the elliptical dynamic characteristic on the constant current diagram. The self-excitation conditions for a triode oscillator with a circuit such as is usual for u.h.f. triode oscillators with disk-sealed triodes is derived for class A, B and C operation.

"A new equivalent circuit diagram of the triode and its significance for u.h.f. triode oscillators", A. Sander. *Nachrichtentechnische Zeitschrift*, **16,** No. 11, pp. 595–604, November 1963.

## RADAR TRACKING SYSTEMS

One modern method of radar angular tracking is to use the monopulse principle. Although much has been written about monopulse systems there has been no proper theoretical comparison of the sensitivities of the conical-scan and the monopulse systems, resulting from the complex nature of the monopulse beams and the different signal treatment in the two systems. A Swedish engineer has set out to determine the fundamental relationship between the different beams and their resulting effect on the total sensitivity of the systems with regard to the signals for range and angular tracking. A conical-scan and a monopulse system with a paraboloid reflector of the same size can be made to have similar signal/noise ratios in the surveillance mode. In the tracking mode, however, the monopulse system then has 6 dB better signal/noise ratio in the range channel, and 3–6 dB better signal/noise ratio in the angular error channel when compared with the conical-scan system.

"A comparison between the sensitivities of radar monopulse and conical-scan systems", Tore Fjällbrant. *Ericsson Technics*, **19,** No. 2, pp. 229–50, 1963.

## SPEECH-PROCESSING TECHNIQUES

An improved pitch extractor for a vocoder, which operates by speech sounds over commercial telephone lines band-limited from 300 c/s to 3400 c/s, has been developed in Japan.

The principle of the pitch extractor is based on the fact that in voiced sounds the same waveforms are repeated in each pitch period, that is to say, the short term auto-correlation of voiced sound waves has high maxima in each pitch period; the pitch periods are between 2·2 ms and 12·5 ms in length. On the other hand, unvoiced sound waves are at random as noise, that is to say, the short term auto-correlation of unvoiced sound waves cannot build up to a high level over a threshold level.

"Auto-correlation type pitch extractor", K. Maezono. *Journal of the Institute of Electrical Communication Engineers of Japan*, **46,** No. 8, pp. 1083–91, August 1963.

## VIDEO STORAGE

Because of the nature of its target mechanism, the vidicon camera tube operates as a storage device, capable of storing both optical and electrical video inputs as a charge distribution for times well in excess of one television frame period. An Australian engineer has discussed the factors which are of main importance when the vidicon is used as a store (as opposed to normal camera usage) with particular reference to the circuit requirements dictated by these factors. Practical circuits of interest in meeting these requirements are also described.

"On the use of the vidicon camera tube as a video storage device", J. B. Potter. *Proceedings of the Institution of Radio Engineers Australia*, **24,** No. 12, pp. 855–65, December 1963.



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