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Journals Under Attack

TODAY there are pressures of all kinds on every learned and professional institution and society, and in particular on their publishing activities, which may not be appreciated or fully understood by many members. Some of these pressures are financial and ultimately could well have a serious effect on the future prosperity and indeed very existence of all institutions. Paradoxically, members often themselves unwittingly contribute to this erosion of the viability of their professional bodies.

There are two trends which threaten the continued existence of learned journals such as our own. The first is perhaps the less obvious, namely the photocopying of articles from current issues. For many years it has been legal to make a single copy of a paper *provided* that it is for private study and that it is obtained through an authorized library, i.e. one not established for profit (and thereby excluding the libraries of industrial and commercial organizations). However, the excessive use of photocopies, to the extent of obtaining these rather than by ordering reprints from the publisher (which, it should be noted, are printed to a standard identical with the original article and moreover cost little if any more than a photocopy) is clearly a direct loss of revenue to the publisher. More significant is the effect on subscriptions to the journal since the former multiple subscriber reduces his order to one copy only, fulfilling requests to see papers in individual issues by furnishing photocopies.

But while the use of photocopies obtained through authorized libraries is within the law, the information service supplying copies of papers which have been produced in quantity and thereby are outside the terms of the 'fair copying' system, is breaking international copyright laws. So too is the purveyor of unauthorized cover-to-cover reproductions. It is in fact these practices, the first most rife in the USA and the second mainly occurring in Eastern Europe, that are particularly worrying, and possible remedies are one of the principal concerns of the Association of Learned and Professional Society Publishers, of which the Institution is a founder member. Some form of combined action may well have to be adopted to deter the multiple copiers.

Turning now to another abuse to which some members are perhaps more consciously a party, journals (and other similar but non-periodical publications) are by custom supplied as one of the rights of membership or at a preferential 'members' rate'. The practice of passing these to a library, who would normally pay the full rate, is clearly a loss in revenue to the Institution. It is equally clearly to the advantage of every member that subscriptions to the Journal should increase in number and therefore help to keep down membership subscriptions.

Solely financial considerations are not however the main criterion in these matters: whether the journals of our own and similar societies continue to exist and flourish is the concern of the whole scientific and technological community. These journals fulfil functions which cannot realistically be effected in any other information distribution system so far proposed. The independence from commercial and similar organizations and the practice of refereeing all contributions provide an essential filter that will keep the stream of knowledge relatively free from spurious, outdated, trivial and obscure material. Of course, the system is not faultless—no man-made system can be—but it is the basis on which scientific and technological progress depends. Its defence from the erosion caused by financial pressure can be sustained by the help and vigilance of all members.

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A v.h.f. surveillance receiver adapted for the reception of suppressed-carrier double-sideband transmissions

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SUMMARY

The paper describes the adaption of a v.h.f. surveillance receiver for reception of suppressed-carrier double-sideband signals by means of a synchronous demodulator of the $2F$ type employing a digital tracking filter. A theory of operation of the demodulator is developed and its performance under various conditions described. A discussion of the choice of automatic gain control is included. The performance is compared, both in theory and practice, with the performance of the receiver when using an a.m. and a f.m. demodulator.

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1 Introduction

Amplitude-modulated signals in which the carrier reduces to a level below that of the sidebands are familiar to h.f. broadcast listeners. Interest in the practicality of successfully recovering the sidebands without over-modulation distortion was made a subject of study by the author, and these rather fortuitously coincided with proposals for power conservation and improved system performance for the telecommunication requirements of the Home Office.[1] This paper describes an experimental v.h.f. radio receiver which incorporates a diminished carrier reception mode and is applicable to radio telephony and data transmission.

The usual forms of v.h.f. radio telephony employ either amplitude or frequency modulation. A single sideband with reduced carrier v.h.f. mobile system has also been described.[2] Double sideband (d.s.b.) is a more general nomenclature for amplitude modulation, but is usually specified more completely as: d.s.b.-s.c. (suppressed carrier), d.s.b.-d.c. (diminished carrier), and a.m. (100% carrier). The distinction is necessary when discussing the nature of the transmission, but as far as reception is concerned, the d.s.b. mode of reception refers to all classes of d.s.b. The receiver to be described also has other modes of demodulation which can be switched in for comparison purposes; these are f.m., a.m. and c.w.

2 Receiver Arrangement

Since the receiver was to be essentially an experimental apparatus, a modular approach was selected.† This approach has enabled the receiver to be employed in a number of different system situations which are referred to below. A functional block diagram of the receiver is shown in Fig. 1.

2.1 *Tuning Head*

The tuning head is modular and heads that cover the range 30–1000 MHz are possible but experiments in the v.h.f. band only have been attempted to date, and the aerial circuit passband was usually narrowed to ~300 kHz. The first local oscillator, normally tuned manually, could be injection locked to an external reference frequency such as a crystal-controlled oscillator.

2.2 *I.F. Amplifier, F.M. and A.M. Detectors*

These are fixed gain units preceded by selectable passband crystal filters at 21·4 MHz. Demodulation of f.m., a.m. and c.w. signals occurs at 1·65 MHz. Bandwidths of 10 kHz and 20 kHz only are discussed in this paper. The 1·65 MHz signal is tapped off this board and feeds the d.s.b. demodulator and an i.f. waveform monitoring terminal.

2.3 *Normal A.G.C.*

The receiver a.g.c. is normally derived from the a.m. detector (envelope) and may have various attack and decay times, which are discussed later. The a.g.c. circuit also drives a signal strength meter. The f.m. discriminator output is taken to a tuning meter.

† Astro Communications Laboratory receiver frame type SR-209.

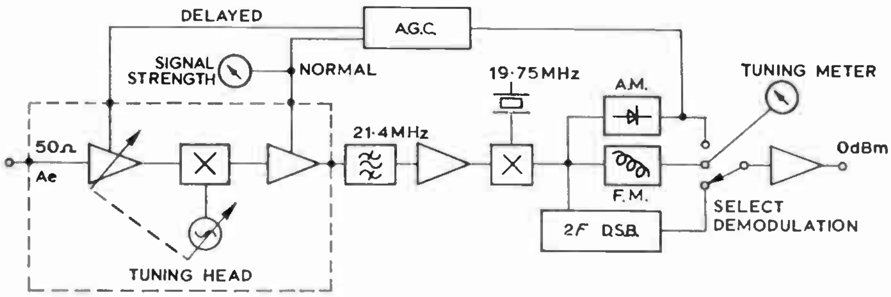

Fig. 1. Experimental receiver schematic.

2.4 *D.S.B. Demodulator*

This is contained on boards similar to the previous circuit boards and is described in detail in Sect. 3. It also has its own a.g.c. arrangement, if desired.

2.5 *Other Functions*

Internal stabilized power supplies (± 12 V and $+24$ V), audio amplifier, mute/squelch circuits are also available. The audio amplifier cuts the audio response by 3 dB at 300 Hz and 3000 Hz.

3 D.S.B. Demodulation

The choice of demodulator depends somewhat on the form of transmitted signal, and the frequency tolerance of the system. It was decided to design for both diminished and suppressed carrier, and also to meet the present v.h.f. mobile radio practice of a transmitter/receiver offset tolerance of approximately ± 1 kHz. These facts suggested, for example, that the use of diminished carrier recovery techniques as were employed in the s.s.b. system already referred to[2] were precluded.

A further difference between s.s.b. and d.s.b. is that the latter requires both exact frequency and phase coherence, or a tightly specified offset. Practical offsets are between approximately 10–40 Hz, but this is even more stringent than for s.s.b., so clearly is not a recommended method of demodulation at v.h.f.[3]

The best known method of demodulating d.s.b. is that described by Costas in 1956.[4] A u.h.f. receiver incorporating a Costas loop has also been described.[5] A loop operating at 1·65 MHz was developed in integrated circuit form,[6] but appeared to be sensitive to frequency jitter which was present in the receiver when operating in the free tuning mode, and so was not incorporated in the receiver.

Diddy and Lindsey[7] have shown, however, that the Costas loop and the alternative method of recovering d.s.b., the '2*F*' system, are theoretically identical under certain conditions and since it is the latter system which is employed in the receiver, it was not felt that the performance was degraded unduly by this choice.

3.1 *The 2F Method*

This method of carrier recovery of the diminished carrier in an a.m. signal has been known for many years.[8] Briefly, if a d.s.b.-s.c. signal is multiplied by itself (squared), it becomes a 100% amplitude modulated signal. The 2*F* component is then selected, or enhanced, by filtering, divided by two to bring it back to carrier frequency, and then used to coherently demodulate the incoming signal. The 2*F* process is illustrated by means of Fig. 2.

This explanation is satisfactory for many signal conditions, but very often some remanent carrier remains in the transmitted signal, which, when the modulation (sidebands) reduces, becomes comparable to the sidebands. At the same time this remanent carrier may well be out of phase with the sidebands and the following discussion indicates some of the difficulties which can arise with the 2*F* demodulator unless certain precautions are observed.

For example, let the received signal be a single-tone-modulated d.s.b.-d.c. signal, where the carrier is out of phase by an angle ϕ with respect to the two equal amplitude sidebands. (This last assertion does not detract from the results obtained.) Then

$$s(t) = A\cos(\omega_c t+\phi)+B\cos(\omega_c\pm\omega_m)t \qquad (1)$$

where A and B are arbitrary amplitudes. Then

$$\begin{aligned}[s(t)]^2_{\text{at }2F} &= \tfrac{1}{2}A^2\cos(2\omega_c t+2\phi)+B^2\cos 2\omega_c t \text{ (carrier 2F)}\\ &\quad +AB\cos[(2\omega_c\pm\omega_m)t+\phi] \text{ (first removed sidetone)}\\ &\quad +\tfrac{1}{2}B^2\cos[2(\omega_c\pm\omega_m)t] \text{ (twice removed sidetone)}\end{aligned} \qquad (2)$$

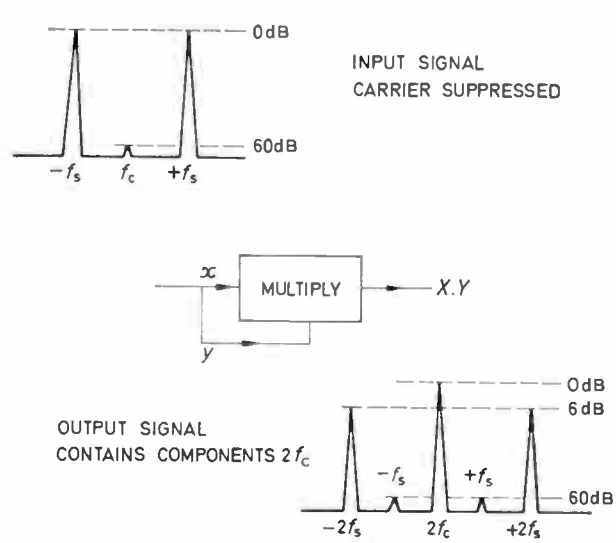

Fig. 2. The frequency doubling process.

$2F$ demodulators in the main rely on the fact that the $2F$ carrier component exceeds other components in amplitude which are generated by the $2F$ squarer. We therefore, examine equation (2) for this condition.

(i) *Carrier in-phase, $\phi = 0$*

The condition we seek is either

$$(A^2/2)+B^2 \geq AB \tag{3}$$

or

$$(A^2/2)+B^2 \geq B^2/2 \tag{4}$$

Condition (4) is illustrated in Fig. 2, and we note the presence of a carrier merely helps the situation. However, not so condition (3). The first removed sidetones are placed closer to the wanted $2F$ carrier and a detailed examination of condition (3) shows that these first removed sidetones are only 3 dB below the carrier if $A \simeq 1{\cdot}3$ B, i.e. $s(t)$ has just exceeded the 100% a.m. condition. This is the reason why it is much more difficult to obtain proper demodulation (lock) with a 100% a.m. signal than a carrier, or d.s.b.-s.c. tone, respectively.

(ii) *Carrier out-of-phase, $\phi \neq 0$*

Again considering the competition between the $2F$ carrier and the first removed sidetones, and taking the components in-phase with B only in equation (3). Therefore

$$A^2 \cos 2\phi + B^2 \geq AB \cos \phi. \tag{5}$$

Let

$$A = nB.$$

Then

$$n^2 \cos 2\phi - 2n \cos \phi + 2 \geq 0.$$

The critical condition is

$$n = \frac{\cos \phi \pm \sqrt{\cos^2 \phi - 2 \cos 2\phi}}{\cos 2\phi}. \tag{6}$$

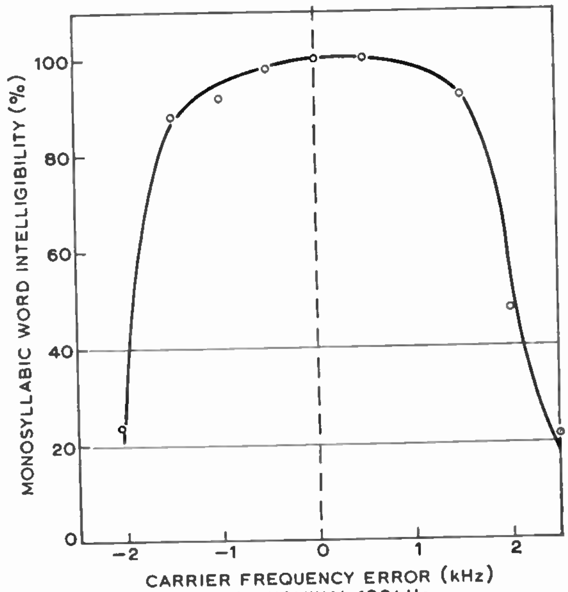

Fig. 4. Intelligibility of detector using $2F$ coil (Pinches, M. C., Private communication).

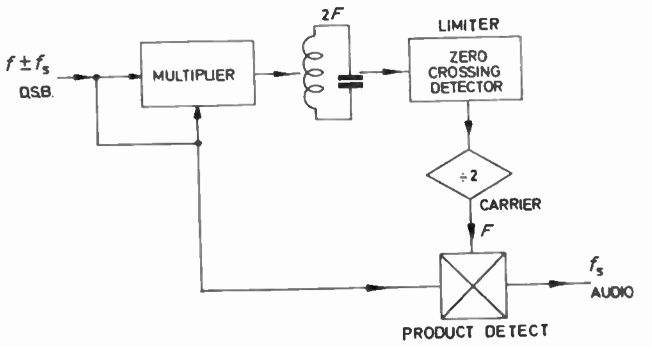

Fig. 3. Coil $2F$ d.s.b. demodulator.

If n becomes real, and the sidetones will have the same amplitude as the carrier component when

$$\cos^2 \phi - 2 \cos 2\phi = 0$$

or

$$\cos \phi = \sqrt{\tfrac{2}{3}} = 0{\cdot}816$$

or

$$\phi \simeq 36^\circ$$

and

$$n = \frac{\cos \phi}{\cos 2\phi} \simeq 2{\cdot}42, \text{ i.e. approx. } 83\% \text{ a.m.}$$

This condition can occur quite easily, especially under selective fading conditions and indicates that the $2F$ system will have a difficult time deciding which is the carrier amongst the sidetones unless the carrier is sufficiently suppressed.

The two effects that occur in practical systems are (i) the hold range of the demodulator is much reduced, and (ii) demodulation fails as the out-of-phase carrier passes through the critical conditions just derived. Good carrier suppression of course removes both these effects.

3.2 *Enhancing the 2F Component*

The simplest method of enhancing the $2F$ carrier component is to use a single tuned circuit. A zero-crossing detector, or limiter, then removes the remaining amplitude modulation. The signal is then divided by two to provide the coherent demodulating signal as shown in Fig. 3. An analysis of the operation of this circuit is given in Appendix 1. This concludes with a result one might intuitively expect, namely, an optimum Q-factor for the $2F$ tuned circuit exists, which combines reasonable sensitivity with reasonable frequency tolerance. Ideally the tuned circuit should eliminate envelope minima, but this is not entirely practical for the lowest modulation frequencies, namely 300 Hz. A coil Q-factor

$$Q \simeq \frac{\text{centre frequency}}{\text{twice average modulation frequency}}$$

is a useful rule of thumb.

Figure 4 shows an experimentally determined monosyllabic word score test versus tuning error for a $2F$ coil system at 100 kHz with Q-factor of approximately 50.

Although this simple demodulation is effective, the resultant speech has an interesting 'burbly' sound, whilst the circuit naturally has a noise threshold which occurs too early for most radio systems. (This is due to the fact that the demodulating carrier is derived in a wideband system, i.e. the coil.)

3.3 *Dynamic Tracking Filter*

The last two disadvantages of the basic $2F$ demodulator indicate the need for selecting the $2F$ component with a high degree of accuracy. For example, selecting the carrier in a 100 Hz bandwidth compared to the incoming signal (i.f.) bandwidth of 10 kHz gives a carrier-to-noise advantage of 20 dB. Since detected signals with signal/noise ratios of less than 10 dB are not of too great practical interest, this indicates the $2F$ component in such a system could be recovered down to carrier/noise levels below those of practical interest. Another important feature of the dynamic tracking filter approach is that phase coherence of the recovered carrier is also maintained, which is essential for d.s.b. demodulation.

The dynamic tracking filter employed in the demodulation was based upon a shunt switched N-path filter.[9] The basic properties of this filter are summarized in Fig. 5.

The fundamental frequency component selected by the network is $1/N$th the clock frequency; the selection bandwidth has an equivalent Q-factor of $NRCf_0$, where C and R are the passive components indicated. Using diode pairs driven directly from JK flip-flops, centre frequencies approaching 1 MHz, and selectivity bandwidths of a few tens of hertz are practical.

In order to make this filter dynamically track the input, the filter output is multiplied N-times and used to injection lock the clock waveform generator. The scheme is shown in Fig. 6, integrated within the $2F$ detection system. The arrangement provides the narrow band selection circuit for the $2F$ component in such a way that the phase and amplitude of the demodulating carrier are almost independant of frequency drift over a small range of frequency.

3.4 *Details of Actual D.S.B. Demodulator*

At first sight, Fig. 6 appears to be free of tuned components. Unfortunately the N-path tracking filter

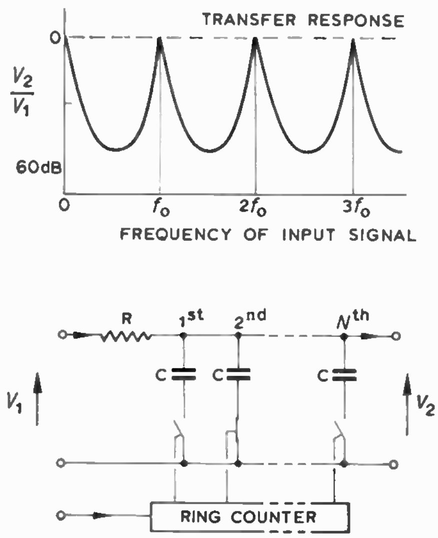

Fig. 5. Basic properties of N-path filter.

responds equally as well to harmonics of its fundamental response, and also generates harmonics because of its signal sampling action. Fixed tuned coils are, therefore, used to select the wanted fundamental at $2F$. At the same time it is also convenient to select a third harmonic, $6F$ for the purpose of synchronizing the clock waveform. A third harmonic is prominent because of the square wave from the first limiter. The position and effect of these coils are illustrated in Fig. 7. In the main the operation of the circuit is exactly identical to the $2F$ coil

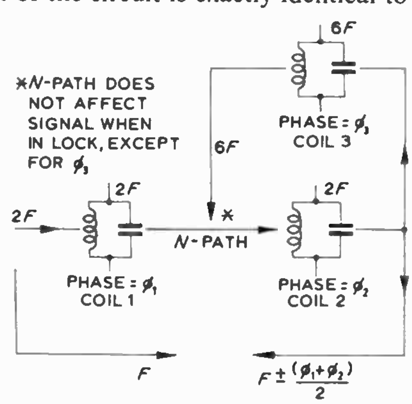

Fig. 7. The tuned components in the N-path demodulator.

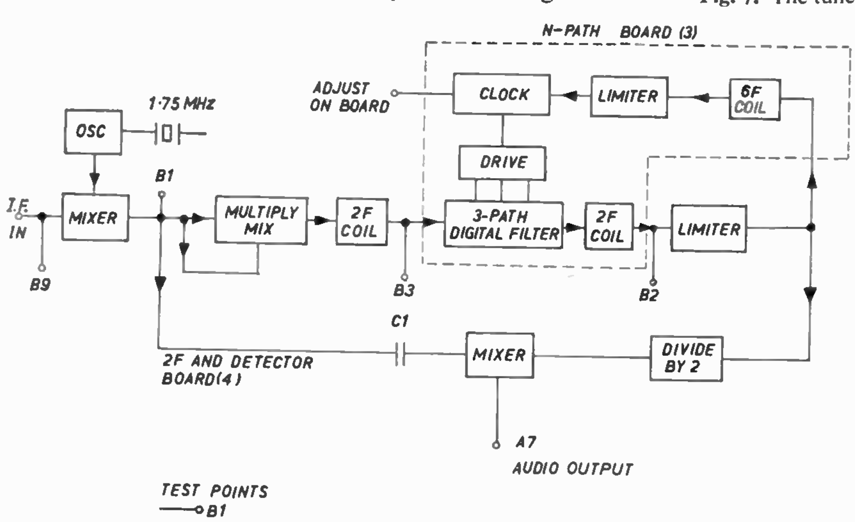

Fig. 6. D.s.b. detector arrangement.

case (Appendix 1). Thus the first $2F$ coils will determine the point at which the outputs 'hops' over to a sideband output, i.e. on single-tone modulation a position where the audio output is twice frequency can be found each side of the nominal tuning position. The only difference is that the N-path filter picks out whichever component is presented to it, effectively free of noise.

The loss of output of the demodulator, due to unequal phase balance between the incoming signal and the restored carrier, will depend on the aggregate of both $2F$ coils. If, on the other hand the $2F$ coils are not unduly selective, the dynamic tracking filter loses lock due to excessive phase shift on the $6F$ coil. The most effective combination of coil Q-factors (which determine the phase shift) has not been established to date. Nevertheless, certain facts are worth noting.

(i) The capture range of the detector depends on the Q-factor of the first $2F$ coil and the most prominent sideband frequency. For speech this is rather indeterminate, but ranges of ± 2 kHz are typical.

(ii) The hold range is determined by the aggregate of coil 1 and coil 2, i.e. the lock signal becomes diminished.

(iii) The demodulator output is not constant over hold range due to phase shifts ϕ_1 and ϕ_2.

(iv) The minimum detectable signal (threshold of demodulator) depends on the Q-factor of the first $2F$ coil and the N-path filter.

(v) The lower the threshold the smaller the capture range. This is because some signal must pass through the N-path filter in order to pull it onto frequency—the more selective the filter the less likely an off-tune signal will pass through the filter.

The time to capture the signal—and to lose lock—are also functions of the previous parameters.

Figure 8 gives some experimental information on this aspect, which is also discussed further below.

3.5 *A.G.C. Circuits*

For reasonably constant level transmitted signals such as d.s.b.-s.c. data signals, a.m., etc., envelope-derived a.g.c. is the obvious choice, and works well. For d.s.b.-s.c. speech, the incoming signal naturally comes in in bursts (the words) and the receiver cannot readily distinguish which is a change in signal level or a change in the transmission. At first sophisticated audio derived a.g.c. was attempted, akin to s.s.b. techniques. A new problem here is that when the signal first enters the receiver there may be no audio output because the demodulator needs a finite time to lock-in (Fig. 8). In this millisecond or so the receiver can quite easily limit which removes any amplitude fluctuation and hence the source of a.g.c. Various novel ways of overcoming this were tried, but simple envelope-derived a.g.c. appears most suitable. This was arranged to have a variable characteristic—selectable by the operator.

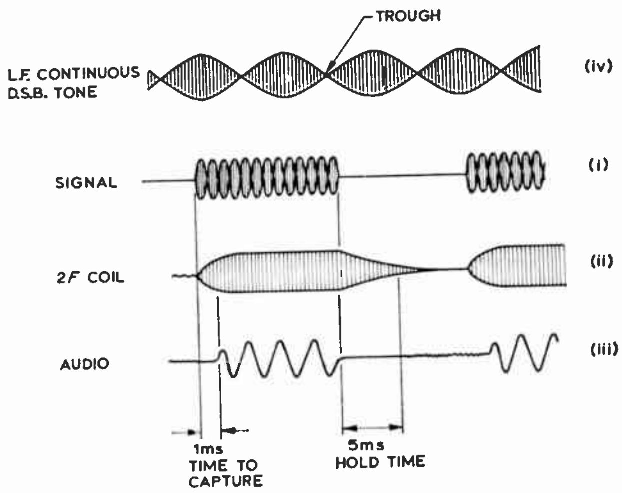

Fig. 8. Observed capture and hold time for demodulator.

The following characteristics were measured:

For 20 dB change in signal level, 10 → 100 μV

Fastest attack time 5 ms, decay time 11 ms

Slowest attack time 100 ms, decay time ~½ s

The fast position appears to be quite acceptable for mobile derived signals, whilst for a stationary (link) path, the slow mode is suitable.

4 Receiver Performance

A tuning head covering the range 55–260 MHz (ACL type SH–271P–1) was employed for most of the trials.

4.1 *Signal/Noise Ratios*

The most interesting performance data for the receiver are the audio (output) signal level versus input (r.f.) signal level, and the signal/noise ratio for the various modulation modes. These data are shown in Figs. 9–11 respectively. In Figs. 9 and 10 the audio output (1 kHz tone), normalized to 0 dBm receiver output, are plotted against input signal together with the noise plus total harmonic distortion. The ratio of these two data gives the SINAD ratios. For the f.m. and a.m. cases the receiver noise with signal but no modulation is also plotted. This gives information about signal/noise. For d.s.b.-s.c. the signal/noise ratio is the ratio of the signal to the noise with no signal (since no carrier) which is also plotted. At low signal levels the SINAD ratio and the signal/noise ratio are more or less equal, at high levels distortion effects show themselves. Figure 11 plots the SINAD ratios for the three modulations.

At low signal levels the decibel advantage of one modulation over another agrees with the theoretical differences—see Appendix 2. However, the experimental measurements of Figs. 9 and 10 indicate other interesting features. For example in a.m. and f.m. the carrier suppresses the receiver noise and one then hears the signal; in the case of d.s.b. however the signal appears to 'climb out' of the noise, once the input signal level is above demodulation threshold.

Depending on the a.g.c. action of the receiver a d.s.b.-s.c. signal may have various limiting signal/noise ratios. Thus in a signal-on/signal-off case, it is about 22 dB if the a.g.c. is very fast acting, but if slow acting it becomes about 36 dB. The addition of a partially suppressed carrier will improve the first figure. On the other hand the coherent (linear) d.s.b. demodulator provides a much improved signal distortion factor. Figure 12 illustrates this feature more clearly, by plotting the total harmonic distortion (%) of a tone modulated a.m. signal, as recovered by (i) a conventional envelope detector, and (ii) the coherent demodulator. It is usual to specify a.m.

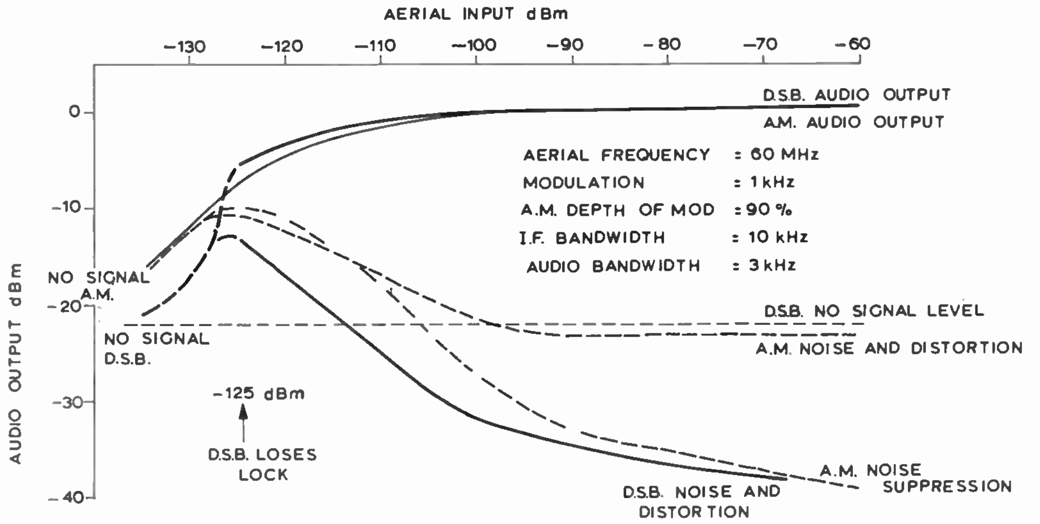

Fig. 9.
Comparison of d.s.b. and a.m. performance.

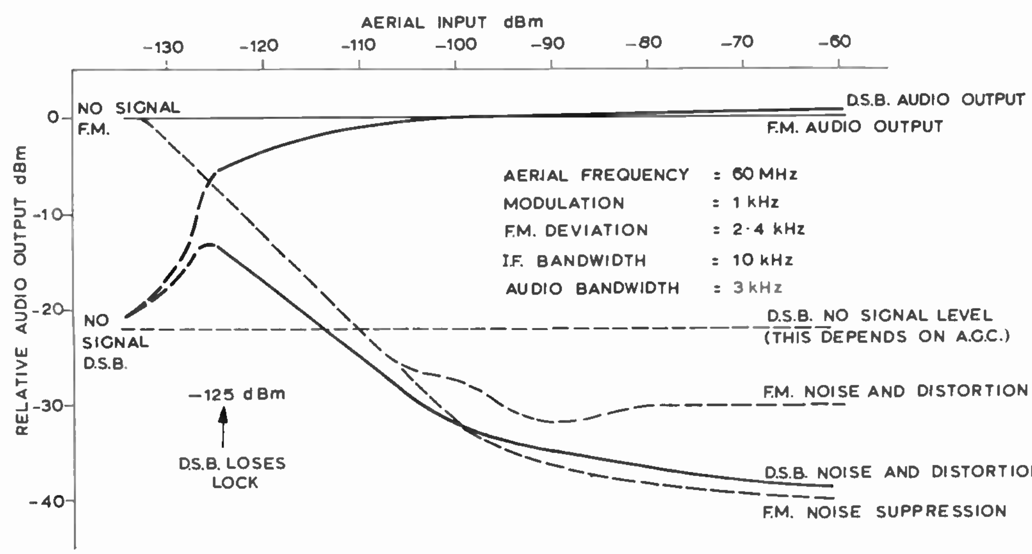

Fig. 10.
Comparison of d.s.b. and f.m. performance.

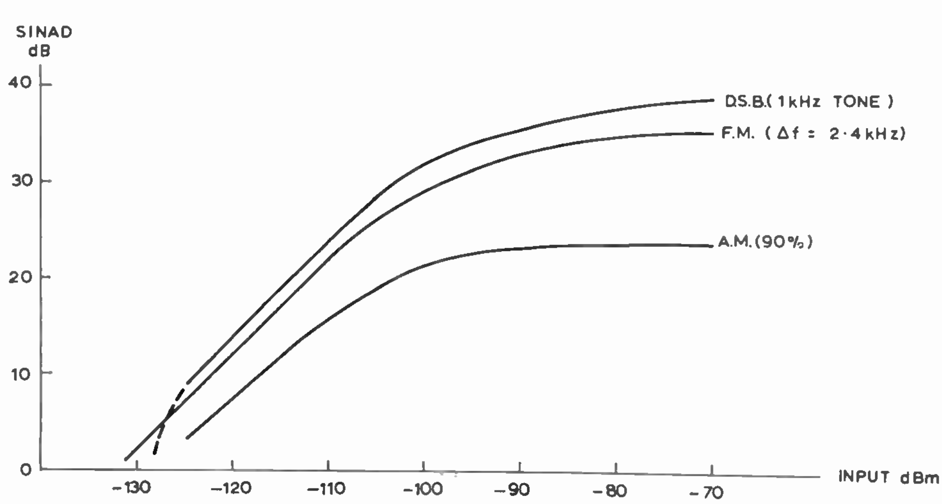

Fig. 11.
SINAD performance of ACL receiver for d.s.b., f.m. and a.m.

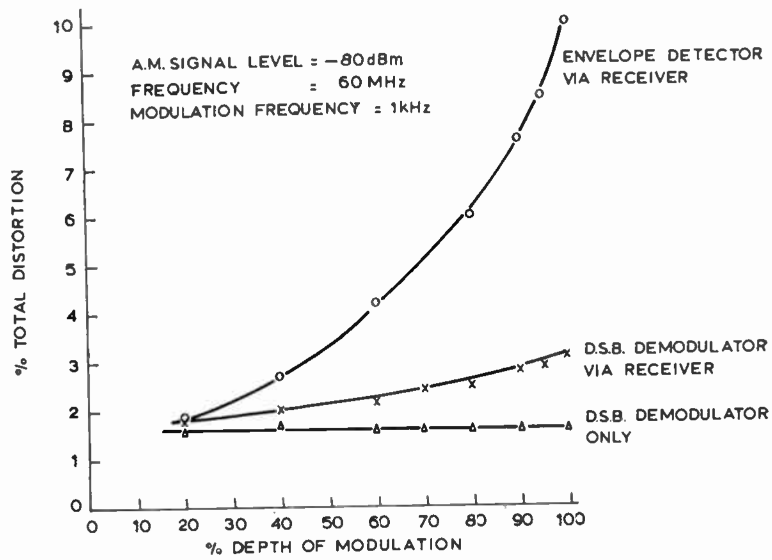

Fig. 12. Signal distortion improvement because of coherent demodulation.

communication systems overall distortion as being less than 10%. With a synchronous demodulator a worthwhile improvement is immediately available.

At very low signal levels (fractions of a microvolt in this case) demodulation threshold is critical. In the case of narrow band f.m. and a.m. there appears to be very little threshold deterioration, the signals gradually fade into the noise. In the case of d.s.b.-s.c. there is a more marked threshold—a function of demodulator employed here. This threshold is somewhat complicated and is best discussed with the aid of Fig. 13. This shows the threshold and signal/noise ratio for three d.s.b.-s.c. tones; 300 Hz, 1000 Hz and 3000 Hz and 0 Hz (carrier).

The reason the signal/noise is best for 1 kHz is because of the audio passband shaping, but a noticeable feature is the premature loss of lock for low frequency tones, i.e. those below 1000 Hz. This occurs because of the nature of a d.s.b.-s.c. waveform. The reader is referred back to Fig. 8. The first three waveforms illustrate the action of an on/off d.s.b.-s.c. signal on the receiver $2F$ signal (ii)

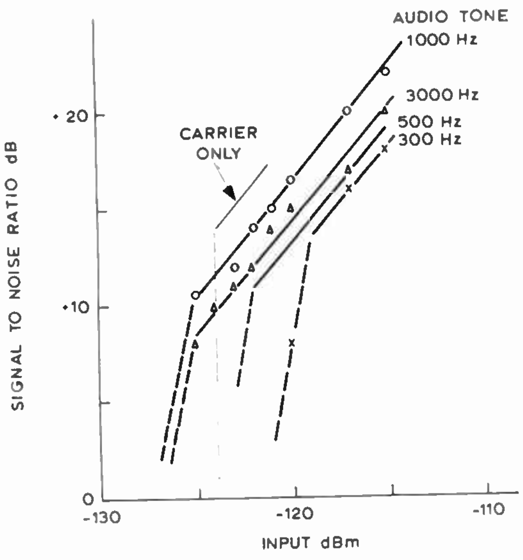

Fig. 13. Threshold performance.

and the audio output (iii). Because of the finite Q-factor of the $2F$ coil and circuit there is an acquisition, or lock-time, of approximately 1 ms (on-tune), whilst the system 'remembers' the signal frequency for approximately 5 ms. Referring now to waveform (iv) of Fig. 8, which is say a 300 Hz d.s.b.-s.c. tone, the long duration of the signal 'trough' is such that under low signal conditions the demodulator 'forgets' where the signal is, and loses lock prematurely.

A related feature is that whereas when tuning say a.m. across the i.f. passband, the audio output remains constant. For this particular synchronous demodulator, a coherent carrier is recovered as long as the sidebands are unaffected by the i.f. passband, but the audio response level falls off either side if the receiver is tuned off the nominal centre frequency. This is because of the phase shift in the $2F$ coils which causes the derived demodulating carrier to move out of phase with respect to the incoming signal (see Fig. 7).

4.2 *Other Features*

Continuing diminished carrier d.s.b. studies[10] in a mobile radio environment naturally led to the development of more specific receivers for the d.s.b.-d.c. mode, some of which have already been referred to.[6] In the main the subjective performances are similar except in so far that the receivers designed specifically to operate off a diminished carrier demonstrate a far less objectionable threshold than say does the present receiver operating with suppressed carrier speech.

Another important area of performance is response to vehicle ignition noise. Laboratory simulated tests here indicated a 3 dB noise advantage was realized with the receiver in the d.s.b. mode compared to the a.m. mode (envelope detector).† This fact also appears to be borne out when listening to low-level signals, though it must be recognized that the variation of d.s.b. and a.m. no-signal noise level discussed above can confuse the issue.

† Watkinson, S. (Swansea University College), Private communication.

The problem of ignition noise and ignition noise reduction in receivers has been recently discussed by Gosling[11] and since the receiver described here was in no way adjusted to minimize ignition noise no further measurements in this area were made.

5 Discussion and Conclusions

Although in the d.s.b. demodulation position the receiver responds perfectly well to a.m. transmissions, indeed with improved quality, full appreciation of d.s.b. operation cannot be gained without the proper transmission of d.s.b.-s.c. or d.s.b.-d.c. In this paper we do not consider the transmission aspect except in so much as to say that the receiver has been successfully employed in two areas of demonstration:

(i) Mobile-to-base d.s.b.-d.c. return path link.[10]

(ii) Ship-to-shore d.s.b.-s.c. data and telephony v.h.f. satellite link.[12]

It is worth noting the performance during these demonstrations confirmed the validity of the performance curves of Figs. 9–13.

For example, the conclusion of improved signal/noise ratio is very apparent in any received signal, but equally obvious is the loss of lock on d.s.b.-s.c. speech signals, because of the intermittent transmitted waveform. Yet on the other hand with, say, f.s.k. data signal, i.e. facsimile transmission, or an above-band audio selective calling tone (i.e. 3·5 kHz) with a speech transmissions, continuous unbroken output will occur once the signal exceeds the −125 dBm input level. It is perhaps these latter features which have perhaps made the possession of an experimental diminished carrier receiver so valuable since so little practical evidence of d.s.b.-s.c. or d.s.b.-d.c. systems appears to be available[13] prior to the Home Office Directorate programme of study.[1, 10]

Finally, the end of performance operation encourages new approaches to the design of diminished carrier systems. Clearly with the adopted receiver structure described here it is practical to introduce variations and immediately compare their improvement or likewise. For example, the demodulator described is of the 'homodyne' family[14] in that it derives the energy of the demodulating carrier directly from the incoming signal even though these may be only sidebands. Phase lock loop and other synchronous demodulators use the incoming signal to synchronize an internal reference carrier to the same phase and frequency of the received signal carrier.

A practical difference which results is that when the former demodulator is not in synchronism (or demodulating) it becomes noisy, whereas the synchrodyne type of detector gives a noticeable audible whistle when out of synchronism.

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7 Appendix 1: Analysis of the 2*F* Coil System

Figure 3 showed the components of the 2*F* coil demodulation circuit.

Let

$s(t)$ = input d.s.b.-s.c. signal $= A\cos\overline{(\omega_c \pm \omega_m)t}$

$p(t)$ = multiplier output $= [s(t)]^2$

$q(t)$ = $p(t)$ after LCR filtering

$\bar{q}(t)$ = $q(t)$ after limiting

$r(t)$ = $\bar{q}(t)$ after frequency division by 2

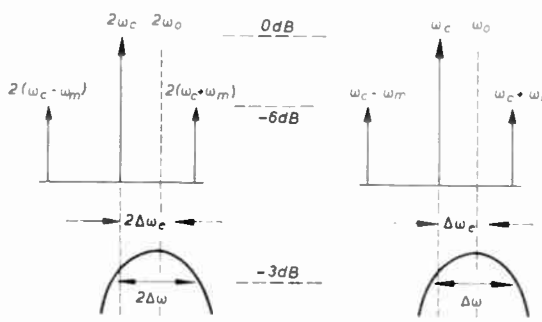
$m(t)$ = wanted output signal $= s(t).r(t)$.

From equation (2), for frequencies near the 2*F* frequency ($2\omega_c$),

$$s(t) = A^2 \cos 2\omega_c t + \tfrac{1}{2}A^2 \cos 2\overline{(\omega_c \pm \omega_m)}t. \qquad (7)$$

Figure 14 depicts the action of the 2*F* coil on what is effectively a 100% modulated a.m. signal. The coil is assumed to have 3 dB bandwidth of $2\Delta\omega$, i.e. Q-factor $= 2\omega_0/2\Delta\omega = \omega_0/\Delta\omega$. Further its resonant frequency $2\omega_0$ is assumed to be offset from the signal carrier (centred at $2\omega_c$) by a frequency error of $2\Delta\omega_e$. To avoid using the factor 2 in subsequent discussion, Fig. 14(b) shows the equivalent situation at a centre frequency of ω_c.

The effect of filtering is to produce the signal $q(t)$, which clearly according to Fig. 14 will have asymmetric sidebands. Amplitude modulated signals with asymmetric sidebands have been extensively discussed by Cherry,[15] and this work has been generally helpful in completing this Appendix.



(a) Mistuned circuit at $2\omega_0$. (b) Mistuned circuit at ω_0.

Fig. 14.

For example, the signal $q(t)$ consists of three terms

$$q(t) = A_L \cos[(\omega_c - \omega_m)t + \phi_L] + A_c \cos(\omega_c t + \phi_c) + A_U \cos[(\omega_c + \omega_m)t + \phi U] \qquad (8)$$

(L = lower sideband, c = carrier, U = upper sideband) where A_L, etc., and ϕ_L, etc., are the amplitude and phase terms of $p(t)$ as modified by the transfer function of the coil, namely

$$\text{Amplitude } A(\omega) = \frac{Q}{\left[1 + Q^2\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)^2\right]^{\frac{1}{2}}} \qquad (9)$$

$$\text{Phase } \phi(\omega) = \tan^{-1} Q\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)$$

where Q = coil Q-factor

Thus

$$\begin{aligned} A_L &= \frac{A^2 Q}{2}\left[1 + Q^2\left(\frac{\omega_c - \omega_m}{\omega_0} - \frac{\omega_0}{\omega_c - \omega_m}\right)^2\right]^{-\frac{1}{2}} \\ A_c &= \text{etc.} \\ \phi_L &= \tan^{-1} Q\left(\frac{\omega_c - \omega_m}{\omega_0} - \frac{\omega_0}{\omega_c - \omega_m}\right) \\ \phi_c &= \text{etc.} \end{aligned} \qquad (10)$$

According to Cherry the signal $q(t)$ may be expressed in the form

$$q(t) = [(X^2 + Y^2)]^{\frac{1}{2}} \cos(\omega_c t + \phi_c - \tan^{-1} Y/X) \quad (11)$$

where X and Y are the in-phase and quadrature components of the signal and are related to the terms of eqn. (10).

The action of the limiter in the demodulator is to act as a zero-crossing detector to produce a signal containing only the frequency and phase of $q(t)$, namely

$$\bar{q}(t) = \cos[\omega_c t + \phi_c - \psi(t)] \qquad (12)$$

where $\psi(t) = \tan^{-1} Y/X$.

The frequency of this signal (usually a square wave) is divided by two to produce

$$r(t) = \cos\left[\omega_c t + \frac{\phi_c - \psi(t)}{2}\right]. \qquad (13)$$

Note, here ω_c has been restored to read $2\omega_c$ before dividing by 2, so $r(t)$ is now the same frequency as $s(t)$.

The output of the demodulator is

$$m(t) = s(t)\, r(t)$$

which after low-pass filtering is

$$m(t) = A\cos\omega_m t \cos\frac{\phi_c - \psi(t)}{2}. \qquad (14)$$

If the 2*F* selection coil were very selective, as is the digital filter, the waveform $q(t)$ would contain negligible sidebands and the term $\psi(t)$ would be absent from eqn. (14). The effect of the phase shift ϕ_c is to cause the output amplitude to vary with tuning, as was discussed at the end of Section 4.

With a less selective coil, however, severe asymmetric sideband distortion, and even failure to demodulate, occurs due to the term $\psi(t)$. This onset of demodulation failure is now discussed in more detail.

For example the demodulator action depends critically on the limiter selecting a regular waveform $\bar{q}(t)$ from the zero-crossing of $q(t)$. Consideration of the three vectors

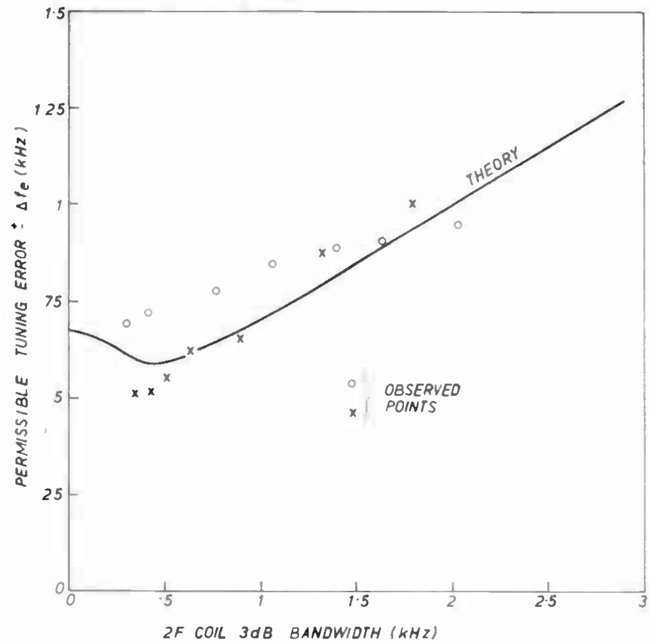


Fig. 15. Variation of tolerable carrier frequency error with $2F$ circuit Q-factor.

A_L, A_c and A_U indicates a zero of $q(t)$ can also occur if

$$A_c = A_L + A_U \tag{15}$$

which will, therefore, be the point of demodulation failure. From equation (10) the condition is

$$\left\{1+\frac{4\Delta\omega_c^2}{\Delta\omega^2}\right\}^{\frac{1}{2}} = \tfrac{1}{2}\left\{1+4\frac{(\omega_m+\Delta\omega_c)^2}{\Delta\omega^2}\right\}^{\frac{1}{2}} + \tfrac{1}{2}\left\{1+4\frac{(\omega_m-\Delta\omega_c)^2}{\Delta\omega^2}\right\}^{\frac{1}{2}} \tag{16}$$

where we have put in the expression for A_L, i.e. equation (10),

$$Q^2\left\{\frac{(\omega_c-\omega_m)^2-\omega_0^2}{\omega_0(\omega_c-\omega_m)}\right\}^2 = \frac{1}{\Delta\omega^2}\cdot\frac{(\omega_c-\omega_m-\omega_0)^2(\omega_c-\omega_m-\omega_0)^2}{(\omega_c-\omega_m)^2}$$

$$\simeq \frac{1}{\Delta\omega^2}\frac{(\omega_m-\Delta\omega_c)^2(2\omega_c-\omega_m)^2}{(\omega_c-\omega_m)^2}$$

$$\simeq \frac{4}{\Delta\omega^2}(\omega_m-\Delta\omega_c)^2, \text{ etc.} \tag{17}$$

Equation (16) may be further simplified by letting

$$B = \frac{2\Delta\omega_c}{\Delta\omega} \qquad C = \frac{2\omega_m}{\Delta\omega}$$

and noting

$$\cos\theta_B = \frac{1}{(1+B^2)^{\frac{1}{2}}}$$

$$\cos\theta_{C-B} = \frac{1}{[1+(C-B)^2]^{\frac{1}{2}}}$$

where θ_B, θ_{C-B} and θ_{C+B} are the angles subtended by B, $C-B$ and $C+B$. By this means condition equation (15) may be expressed as

$$2\cos B = \cos(C-B)+\cos(C+B). \tag{18}$$

This equation has been solved numerically and the solution is shown as Fig. 15, where the tuning error Δf_c is plotted versus the $2F$ coil 3 dB bandwidth Δf, both normalized to a modulation tone of 1 kHz. (Equation (18) gives a condition between $\Delta\omega_c/\Delta\omega$ and $\omega_m/\Delta\omega$ from which $\Delta f_c/f_m$ versus $\Delta f/f_m$ are obtained). The result is interesting in the sense that a very high Q-factor $2F$ coil is not necessarily the best arrangement from the viewpoint of receiver tuning tolerance, although a very selective coil has the best demodulation threshold characteristic. The result is supported by practical observations as indicated in Fig. 15.

There is also a further factor which suggests the $2F$ coil should not be too selective, which can occur due to finite limiter action in deriving $\bar{q}(t)$ from $q(t)$. In practice the limiter will fail at a certain minimum signal level, say, q_{min}, i.e. a condition is

$$|q(t)| > q_{min} \tag{19}$$

If we assume $q(t)$ mainly consists of the carrier term A_c, then

$$\frac{AQ}{\left[1+Q^2\left(\frac{\omega_c}{\omega_0}-\frac{\omega_0}{\omega_c}\right)^2\right]^{\frac{1}{2}}} \geqslant q_{min} \tag{20}$$

which can be written as

$$A > \frac{q_{min}}{Q}\left\{1+4\frac{\Delta\omega_c^2}{\Delta\omega}\right\}^{\frac{1}{2}}. \tag{21}$$

Equation (21) indicates that when the demodulator is centre-tuned one gains the usual Q magnification in the $2F$ coil, i.e.

$$A_{\text{on tune}} > \frac{q_{min}}{Q}. \tag{22}$$

Off-tune however the factor $\Delta f_c/\Delta f$ will increase this minimum value by an amount $[1+4(\Delta f_c/\Delta f)^2]^{\frac{1}{2}}$. From inspection of Fig. 15 this will however be small (about 2) unless a very selective circuit is used, in which case loss of demodulation, because of limiter failure, will occur prior to sideband interference failure.

8 Appendix 2: Comparative Signal/Noise Ratios

The method of measuring the signal/noise ratio for the various forms of modulation consisted of applying a signal of a known power level (dBm) to the receiver, and then deducing the signal/noise ratio of the audio output signal using a selective voltmeter or so-called distortion measuring meter. Since the same front end and i.f. passband width of the receiver were used in all cases it is assumed the noise power level remained the same in all instances.

Thus if the same input signal power level is used, any differences in the output signal/noise ratio—such as in Fig. 11—must be attributed to differences of demodulation improvement, etc. This Appendix considers the expected theoretical differences, a subject which is discussed in many texts, for example reference 16.

Table 1 assists in this matter. Here the three main modulation spectra are shown on, (i) an equal sideband power basis, and (ii) an equal carrier power basis, where applicable. As indicated in the row marked input power level, this gives 100% a.m. a 4·8 dB input signal power advantage, and f.m. a 3 dB advantage, respectively, over d.s.b.-s.c. Corresponding adjustment of the input signal levels should be made so that each form of modulation has the same signal input power.

In the case of d.s.b.-s.c. and a.m. the familiar 3 dB mark-up is obtained on demodulation. In the case of f.m., the improvement has been expressed as a ratio to a.m. output signal/noise ratio assuming equal carriers, according to the formula,[16]

$$\left(\frac{S}{N}\right)_{\text{f.m.}} = \frac{3\Delta f^2}{B^2}\left(\frac{S}{N}\right)_{\text{a.m.}} \tag{23}$$

where

Δf = f.m. deviation

B = receiver audio passband width.

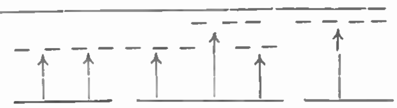
This correction is, therefore, added to the relative signal/noise ratio obtained for a.m.

Comparing d.s.b.-s.c. and a.m., the same demodulation improvement is obtained, but requires the a.m.. the input signal to be 4·8 dB above the d.s.b.-s.c. level, hence for equal input or received levels, the a.m. signal/noise output ratio lags 4·8 dB behind that of the d.s.b.-s.c. curve. In the practical measurement 90% a.m. is used (to avoid excessive distortion) and then a difference of 5·4 dB is expected, i.e. the ratio is $1+2/M^2$: 1, where M = modulation index. The figure of 5·4 dB is confirmed by Fig. 12, at low SINAD ratios.

For f.m., having set the input signal carrier equal to that of a.m., the f.m. advantage, equation (23), needs to be added to the a.m. result. Using Δf = 2·4 kHz (receiver bandwidth = 10 kHz), and a post-discrimination bandwidth B = 3·0 kHz, the advantage is + 2·8 dB which must now be added to the −4·8 dB, and so leads to a −2 dB difference when comparing d.s.b. and narrow band f.m., which is again confirmed by Fig. 12.

Table 1

Comparison of modulations

Modulation	d.s.b.-s.c.	100% a.m.	f.m. Δf = 2·4 kHz
Signal components			
Relative signal power	× 2·0	× 6·0	× 4·0
Input power relative to d.s.b.	0 dB	+ 4·8 dB	+ 3 dB
Signal power relative to a.m. carrier	− 3 dB	+ 1·8 dB	0 dB
Demodulation improvement	+ 3 dB	+ 3 dB	+ 2·8 dB†
Signal/noise ratio relative to d.s.b.-s.c.	0 dB	− 4·8 dB	− 2·0 dB

† Using equation (23).

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STANDARD FREQUENCY TRANSMISSIONS—May 1973

(Communication from the National Physical Laboratory)

May 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds NPL—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	−0·3	−0·1	0	721	629·3
2	+0·2	0	0	719	629·0
3	0	0	0	719	629·0
4	−0·1	0	0	720	628·6
5	+0·1	0	−0·1	719	628·4
6	+0·1	+0·1	0	718	627·8
7	+0·1	+0·1	0	717	627·0
8	+0·1	0	0	716	626·4
9	+0·1	+0·1	0	715	625·6
10	+0·1	+0·1	0	714	625·0
11	+0·1	+0·1	0	713	623·6
12	0	0	0	712	623·7
13	0	0	0	712	623·3
14	0	0	0	712	623·4
15	0	0	+0·1	712	623·7
16	0	0	0	712	624·4
17	−0·1	−0·1	0	713	625·1
18	−0·1	−0·1	0	714	625·8
19	−0·1	−0·1	−0·1	715	626·5
20	−0·1	−0·1	0	716	627·4
21	−0·1	−0.1	0	717	628·5
22	0	+0·1	0	717	627·7
23	+0·1	+0·1	0	716	626·9
24	0	+0·1	+0·1	716	626·2
25	+0·1	+0·1	+0·1	715	625·2
26	+0·1	0	+0·1	714	624·8
27	0	0	+0·1	714	624·4
28	0	0	0	714	624·4
29	0	0	0	714	625·4
30	0	0	0	714	624·4
31	0	0	0	714	624·1

All measurements in terms of H-P Caesium Standard No. 334, which agrees with the NPL Caesium Standard to 1 part in 10^{11}.

* Relative to UTC Scale; $(UTC_{NPL}$ − Station) = + 500 at 1500 UT 31st December 1968.

† Relative to AT Scale; $(AT_{NPL}$ − Station) = + 468·6 at 1500 UT 31st December 1968.

UDC 621.372.832.8 : 621.374.33

A technique for gating short microwave pulses

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SUMMARY

Short microwave pulses are generated by feeding c.w. microwave power into a closed, low loss resonant ring of waveguide so that the power circulates and builds up; this power may be released in the form of a short pulse of increased amplitude, by opening a gate controlled by p-i-n diode switch. An experimental system providing a 34 ns pulse with a 2·6 dB amplitude gain is described.

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1 Introduction

Waveguide resonant rings have been used for testing microwave components at high power levels without the need for high-power microwave sources. Such schemes involve an oscillator which couples its power into a closed waveguide ring by means of a directional coupler. If the path length around the closed ring is resonant, the input energy will continue to circulate and the power can build up to several times the input level. The circulating power level will be set by the resistive losses within the ring and may also be maximized by choosing the optimum coupling coefficient of the directional coupler.[1]

This paper describes a modified form of a waveguide resonant ring which includes a circulator, in which the spare port is terminated in a short circuit, produced by an appropriately biased diode (such as a p-i-n diode). If the ring is excited at resonance and the short circuit is removed by pulsing the diode, a short microwave pulse will emerge, having a magnitude equal to the circulating power within the ring and a time duration equal to the time delay around the ring. Such a pulse could be used as transmission waveform for a short-range, high-resolution radar system, thus providing the capability of short pulse gating for a microwave oscillator with a useful gain, while preserving phase coherence.

2 Experimental system

Figure 1 shows the schematic diagram of a X-band experimental waveguide resonant ring employing a circulator and a p-i-n diode gate. A conventional resonant ring is resonant for both directions of propagation around the ring and, although the directional coupler excites a travelling wave in one direction only, imperfections in coupling or small mismatches will cause resonant waves

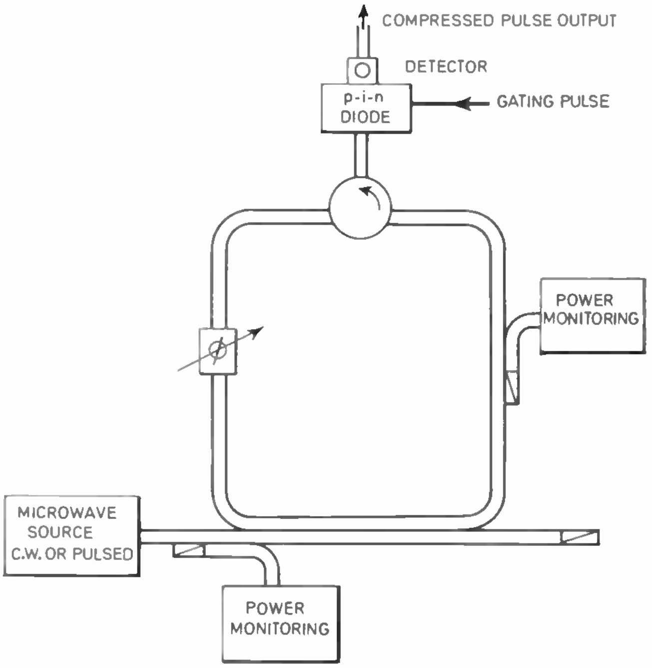

Fig. 1. Schematic diagram of experimental waveguide ring assembly with circulator and p-i-n diode gate.

to propagate in the reverse direction. However, the inclusion of the circulator and stub arm causes the electrical path length to be different for the two directions of propagation, since the stub section is not included in the path of a reverse wave. Consequently mismatches result in reverse waves which need not be resonant at the same frequencies.

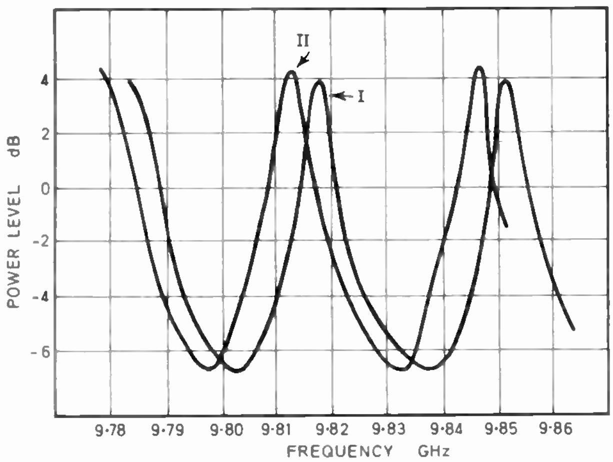

Fig. 2. Frequency response of resonant ring assembly. (0 dB represents input power level.)

Curve I with a waveguide short circuit terminating the stub arm.

Curve II when the resonant ring is continuous with the circulator and stub arm removed.

Although this technique is mainly applicable to generating very short pulses in the nanosecond region, the experimental demonstration utilized a rather long length of waveguide to study the system operation and display the waveforms more conveniently. The waveguide resonant ring was 6·7 metres long (corresponding to 161 λ at 9·784 GHz). Copper losses within the resonant ring were therefore far larger than for a short pulse duration design. Attenuation measurements within the ring gave a loss of 2·2 dB of which 0·35 dB was due to the circulator.

The frequency response of the ring was first measured with the stub arm terminated in a waveguide short circuit (Fig. 2, curve I). This response shows resonant maxima at 9·7845 GHz, 9·8115 GHz and 9·8515 GHz. Curve II is a corresponding response when the circulator and short circuit are removed to provide a continuous waveguide ring. The resonance maximum is increased by 0·48 dB owing to the reduced attenuation. The increase in power at resonance can be seen from Fig. 2 which is plotted relative to the input level taken as 0 dB. For the ring and stub arm (Curve I) this corresponds to a gain of 3·9 dB compared with a figure of 4·13 dB calculated from the measured ring attenuation. This is further modified by the imperfect diode short circuit for the subsequent pulse results.

Figure 3 (a) shows the typical output pulse resulting from gating the p-i-n diode. This was a type DH 582 diode providing a short-circuit reflexion coefficient of 0·99 and an open circuit transmission coefficient of 0·84. The rise of this pulse is 3 ns and the duration 34 ns. The effective amplification of the pulse is only 2·6 dB due to the loss in the waveguide and the effect of the diode. The good rectangular shape of the pulse is due to the fact that the duration of the diode gating signal has here been arranged to coincide with the duration of the output pulse (i.e. with the time delay around the ring).

Figure 3 shows the waveform when the p-i-n diode is left in the open circuit position for a period longer than the time duration of the pulse. Waveform I represents the pulse voltage applied to the p-i-n diode and waveform II the detected output of the compressed pulse. The lack of time coincidence of these two waveforms is principally due to different cable lengths to the display oscilloscope.

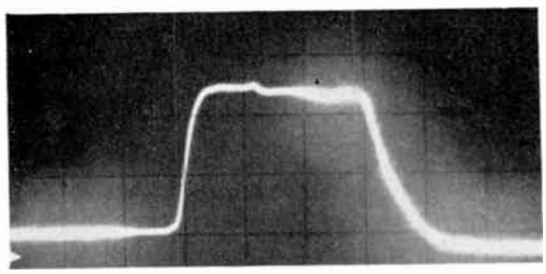

Fig. 3. (a) Waveform of detected output pulse from p-i-n diode gate, when the duration of the gating signal corresponds to the time delay around the waveguide ring. The time scale of the oscilloscope picture corresponds to 10 ns/cm.

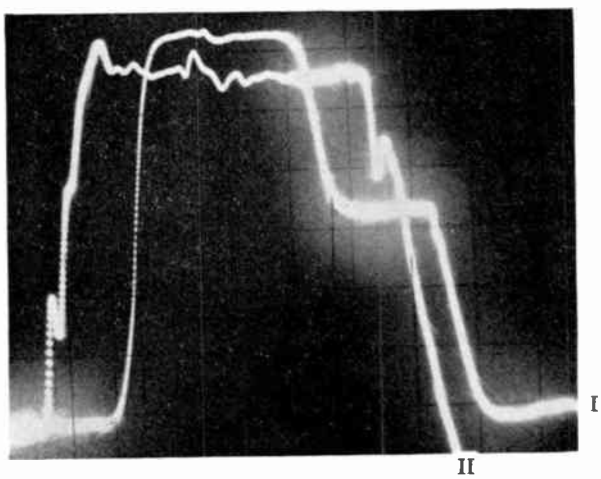

Fig. 3. (b) Waveform of the output detected pulse (waveform I) and of the gating pulse to the p-i-n diode (waveform II) when the diode gate is left open beyond the time delay around the ring. The time scale of the oscilloscope picture corresponds to 10 ns/cm.

The waveform of the compressed pulse shows an initial region with an amplitude gain of 2·6 dB and of duration given by the propagation time around the ring, the pulse then falls to about half its amplitude for a further 20 ns until the p-i-n diode switches the output off. This step on the trailing edge of the pulse is the combined effect of the power still being fed into the waveguide ring, and the power reflected from the imperfect open circuit of the p-i-n diode.

Hobson and Kocabiyikoglu[2] showed how it was possible to increase the peak output power of a Gunn effect diode by mounting it in a Q-switched waveguide cavity. The present proposal may also be regarded as a Q-switched resonant ring cavity. An advantage in the present arrangement arises from the inclusion of a circulator in the ring so that the forward and backward waves may be conveniently separated. In principle, it is also possible to cascade two separate resonant ring systems (employing different lengths of waveguide) to produce two stages of compression.

3 Conclusions

These microwave experiments demonstrate the feasibility of using a waveguide resonant ring with a gated short circuit to generate short microwave pulses. The gate needs to operate quickly in one direction only, from a short circuit to an open circuit condition, since the trailing edge of the pulse should represent an adequate turn off when the effective amplification in the ring is large. Amplitude gains larger than the present figure of 2·6 dB would be possible with waveguide rings of reduced time delay (and reduced attenuation).

4 References

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Electrolytic capacitors:

their fabrication and the interpretation of their operational behaviour

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SUMMARY

The various stages in the production of an aluminium electrolytic capacitor are outlined and these are then related to the equivalent circuit so as to establish an understanding of the preparation parameters that affect the performance. Particular reference is made to studies undertaken on the effects of etching prior to anodization, the anodization itself, the electrolyte used and the mechanical construction. In the case of sintered porous tantalum electrolytic capacitors, the parameters are not so well understood but comparisons will be drawn with aluminium capacitors wherever possible.

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1 Introduction

Electrolytic capacitors are basically made by the anodization of aluminium or tantalum metal, the metal either being in the form of thin foil or, in the case of tantalum, in the form of a sintered porous block. A recent review has been given on the preparation[1] and the properties of both aluminium and tantalum electrolytic capacitors and therefore it is unnecessary to repeat the majority of this material in the present paper. This paper, therefore, concentrates on the recent advances that have been made in the understanding of the behaviour of electrolytic capacitors, especially aluminium types, with particular reference to the equivalent circuit that can be drawn to represent the total behaviour of a capacitor, both with regard to impedance and frequency.

2 Preparation of Aluminium Electrolytic Capacitors

In order to analyse the equivalent circuit of an aluminium electrolytic capacitor it is proposed briefly to examine the preparation of these capacitors. This will identify the electrical parameters with the various stages of preparation and thus establish an understanding of the effects of the preparation parameters on performance.

The stages in the preparation of an aluminium electrolytic capacitor are shown diagrammatically in Fig. 1. The various stages can be identified as follows:

(1) *Pre-preparation*

Since it is important to have an impurity free surface, extremely pure aluminium foil (99·99% Al) is used as the starting material. Impurities that can be detrimental to the formation of a satisfactory oxide film are iron, copper, manganese, boron and silicon, and foil in general must be inspected for these materials prior to use, by chemical or spectrographic means, so as to establish that levels less than, or equal to, 10 parts per million are obtained. Other impurities that will affect the stability of the final capacitor such as sulphate or chloride ions have also to be assessed, although, in general, contamination due to these impurities arises from subsequent treatment rather than from the aluminium foil itself.

As part of the pre-preparation it is necessary to brush or chemically clean the surface. One of the effects of this is to give nucleation sites for the subsequent etching of the foil. However, it should be noted that brushing of foil with a metal wire brush (usually iron or steel) also has the effect of making the foil less brittle and therefore more easily windable. It has been found that such treatment is of use for making capacitors and chokes using aluminium foil even when the foil is not being subsequently etched or anodized, (i.e. the dielectric is a plastic sheet rather than an aluminium oxide film).

(2) *Etching*

In order to increase the surface area of the foil, the prepared aluminium is etched electrochemically. Etching results in a large number of narrow tunnels running into the foil[1] and these can increase the surface area by a factor of up to 50. It is found that the tunnels obtained are of varying widths and lengths. Thus, the surface area

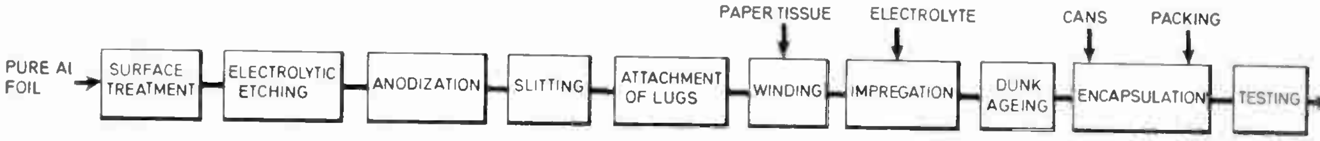

Fig. 1. Block diagram showing the preparation of an aluminium electrolytic capacitor.

is a function of the thickness of oxide required (i.e. the working voltage at which a foil is to operate), since as the thickness of oxide required is increased to allow for higher operating voltage more of the tunnels become filled and the surface area is therefore reduced. Figure 2 shows a typical relationship between surface gain and anodizing voltage obtained. As can be seen in Fig. 2, the effective surface area is very little more than that of a plain unetched foil if a sufficiently thick oxide is grown to block the pores.

Different etched structures with wider tunnels can be obtained for use with thick oxides (at high voltages).[1]

An electrolytic capacitor is generally constructed with an anode foil on which is formed the oxide applicable to the voltage required, and a cathode foil which in general is arranged to have a very large capacitance. This large cathode capacitance is effectively in series with the anode capacitance so that the larger the cathode capacitance the less effect it will have on the total value of capacitance obtained in the capacitor. To obtain as large a capacitance as possible the cathode foil needs to be etched to give a surface gain that is as large as possible and the oxide formed on the cathode needs to be very thin.

Great care must be taken to wash the foil subsequent to etching to remove all traces of contamination, particularly as sodium chloride is often used in the etching baths and chloride ions can cause subsequent corrosion of the aluminium. This can be done by the use of water jets but sometimes electrolytic techniques are employed.

The mechanisms of etching foils either for low or high voltage application are not at this stage fully understood. This is partly because of the commercial security surrounding the subject which prevents information being published.

(3) *Anodization*

Anodization is effected by applying a positive voltage to the foils immersed in a tank of an electrolyte. The second electrode in the tank, the cathode, is usually an aluminium or steel plate. A film will grow on the foil reaching a final thickness limited by the applied voltage (see Fig. 3). For a constant voltage applied, the limiting thickness obtained in the case of aluminium is 1·36 nm/V at an anodizing temperature of 20°C and 1·6–1·7 nm/V at 80°C. This latter temperature is the one at which anodization is normally performed.

Two limitations apply to the anodization of the foil. Firstly, in order to keep the anodization process efficient it is not possible to anodize it to the limiting thickness as the growth of the oxide will be relatively slow in the latter stages of formation. Therefore a limited time of immersion is allowed so that a film of oxide of smaller thickness will be formed. Thus it is necessary to define a 'working' voltage above which the anodized foil must not be used and this is not the same voltage as the 'forming' voltage which has been applied to the foil during the anodization. The 'working' voltage is usually around 70–80% of the anodization voltage. The second limitation is that it is not possible to anodize to voltages much above 650 V as thick oxide films will often break down electrically in subsequent use.

The oxide that is obtained by anodization is found to

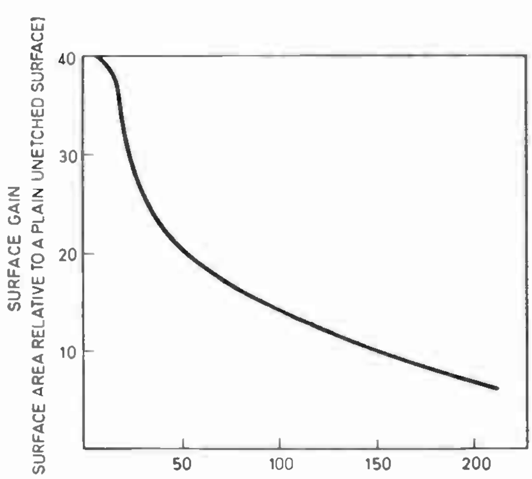

Fig. 2. Relationship between surface gain and anodizing voltage (for a low voltage type of foil etch).

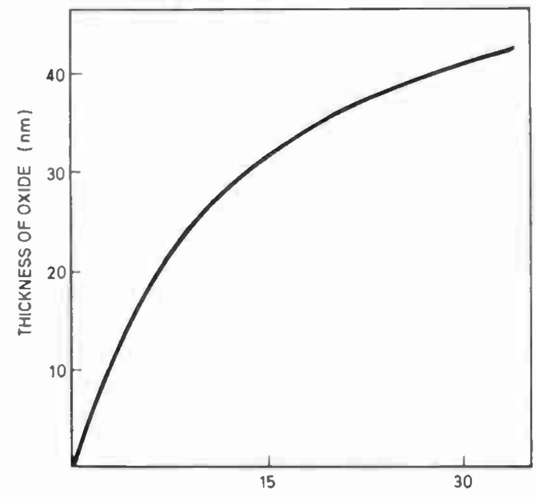

Fig. 3. Constant voltage growth of film of alumina on aluminium surface as a function of time (30 V applied).

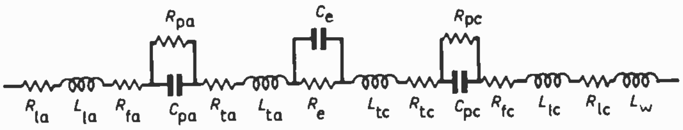

Fig. 4. Equivalent circuit of an electrolytic capacitor showing the effect of the various preparation stages.

be amorphous and has a dissipation factor of around 2%. This loss can be described physically as being due to electron hopping conduction between structural traps in the material[2] and recent work in this field has determined the relationship between the temperature coefficient of capacitance, permittivity and loss of such materials.[3]

It has been shown that the process of anodization is not as straightforward as might be expected. Boiling of the foil in water either before anodization or at some stage during the anodization process has been found to be important. It is known that this leads to the formation of hydrated alumina (Boehmite) which has the effect of giving a more stable film with a higher capacitance than would otherwise be obtained.[4]

(4) *Mechanical Construction*

Subsequent to anodization, the etched and anodized foils have to be mechanically treated. This involves slitting the foil to the correct width, attaching lugs and finally winding up the foils, both anode and cathode, together with the paper spacers between them which will be used as a wick to contain the working electrolyte. The slitting process will result in the exposure of an un-anodized edge to the foil and the lug attachment will damage the oxide which has been grown on the foil surface. The paper spacer itself must be free of sulphate and chloride impurities. Care must be taken not to damage the foils unnecessarily in winding. The winding will result in a cylindrical spiral construction for the completed unit.

(5) *Impregnation*

The wound units are impregnated with a working electrolyte usually using a vacuum technique, the actual electrolyte chosen being dependent on the 'working' voltage of the capacitor. The resistivity of the electrolyte will be between 2000 and 300 Ω cm.

It should be noted that the scintillation voltage in the electrolyte-oxide system is an important criteria. This is the voltage at which discharges start to occur at the oxide-electrolyte interface and is related to the composition of the electrolyte and its resistivity—a high scintillation voltage will be associated with a high resistivity.

(6) *Dunk Ageing*

In order to repair defects in the oxide produced during mechanical construction, i.e. the bare edges produced at slitting and faults produced by lug connexion and winding, the units are subjected to an ageing process in the working electrolyte. A voltage above the working voltage is applied so as to grow an oxide film on the defective regions.

(7) *Encapsulation*

The impregnated, aged units are finally placed in cans, packed to prevent movement of the unit, and sealed. As anodization of the foil will occur during the working life of the component, a large number of electrolytic capacitors, particularly those of large value, are supplied with a mechanical vent to allow the release of any hydrogen developed subsequently in the unit.

3 Lumped Component Equivalent Electrical Circuits

With the preparation stages in mind, it is possible to draw an equivalent circuit of an electrolytic capacitor and this is shown in Fig. 4. The following contributions to the equivalent circuit can be noted:

(1) *Connecting lugs to the anode foil*

The resistance and inductance due to the lugs connecting the terminals to the anode foil (R_{la}, L_{la}) are shown.

(2) *Anode foil*

Circuit elements associated with the anode foil are the resistance of the foil, R_{fa}, the capacitance of the oxide on the foil, C_{pa}, and the parallel resistance representing the loss of the oxide, R_{pa}. Two other elements have also been included with the foil. These are the resistance of the electrolyte in the etched tunnels in the foil R_{ta} and the inductance of the electrolyte in these tunnels L_{ta}. These last two elements are due both to the type of etching structure and the type of working electrolyte that is employed.

(3) *Elements due to the electrolyte*

The first two of these elements—the resistance and inductance of the electrolyte in the tunnels in the foil—have been referred to under item (2). It is also necessary however to include a further resistance which is due to the electrolyte in the paper wick, R_e and a parallel capacitance, C_e that represents the dispersion of the paper impregnated with electrolyte.[5,6]

(4) *Cathode foil*

The situation with regard to the circuit elements for the cathode foil is exactly similar to that discussed for the anode foil and components must be introduced to represent the resistance of the electrolyte in the tunnels in the cathode foil, R_{tc}, the inductance of the electrolyte in the tunnels in the cathode foil, L_{tc}, the resistance of the metal of the cathode foil, R_{fc}, the capacitance of the cathode, C_{pc}, and finally the parallel resistance, R_{pc}, representing the loss of the dielectric on the cathode. The cathode capacitance, C_{pc}, will in general be at least ten times the anode capacitance, C_{pa}.

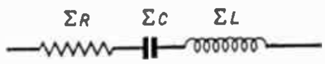

Fig. 5. Simplified equivalent circuit of electrolytic capacitor.

(5) *Connecting lugs to the cathode foil*

The connecting tabs to the cathode introduce a resistance, R_{lc}, and inductance, L_{lc}.

(6) *Winding inductance*

The final element that must be introduced is an inductance, L_w, representing the winding of the capacitor in a spiral formation.

Identification of these circuit elements with preparation parameters can now be undertaken. The elements due to anode and cathode lugs are a function of mechanical construction as are the resistances of the anode foil and the cathode foil, and the winding inductance. The anodization of the foil determines the anode and cathode capacitance and the anode and cathode parallel resistance. Etching of the foils governs the anode and cathode foil resistance and inductance. However, as has been noted above, the electrolyte (i.e. the impregnation), also influences the anode and cathode inductance and resistance. The electrolyte impregnation will affect the resistance due to the electrolyte of the paper wick and also the dispersion, i.e. the parallel capacitance, C_e.

Such a circuit can be simplified. It is possible to turn the parallel components representing the anode oxide, R_{pa} and C_{pa} into series resistance and capacitance values R_{sa} and C_{sa}. The series resistance will be given by

$$R_{sa} = R_{pa}[\tan^2\delta/(1+\tan^2\delta)] \qquad (1)$$

$\tan\delta$ is the loss of the oxide (i.e. $\tan\delta = 1/C_{pa}R_{pa}$).

The series capacitance C_{sa} is given by

$$C_{sa} = C_{pa}[1+\tan^2\delta] \qquad (2)$$

As it has been shown[1] that the loss of the dielectric is around 2%, the series capacitance, C_{sa}, is effectively equal to the parallel capacitance, C_{pa}.

A similar simplication is obtained with the cathode oxide.

The parallel components representing the dispersion of the electrolyte R_e and C_e, can be turned into series components. In this case the series capacitance obtained will be very large compared with that of the cathode and anode capacitance, C_{sc} and C_{sa} and can be ignored.

Therefore, all the components are now in series form and an equivalent circuit can be drawn as shown in Fig. 5. Such a circuit will have an impedance minimum as a function of frequency since the impedance Z is given by

$$Z = \sqrt{(\Sigma R)^2 + \left(\omega\Sigma L - \frac{1}{\omega\Sigma C}\right)^2} \qquad (3)$$

where

$$\Sigma R = R_{la}+R_{fa}+R_{sa}+R_{ta}+R_e+R_{tc}+R_{sc}+R_{fc}+R_{lc}$$

$$\Sigma C = C_{sa}.C_{sc}/(C_{sc}+C_{sa});$$

$$\Sigma L = L_{la}+L_{ta}+L_{tc}+L_{lc}+L_w.$$

Figure 6 shows a typical curve of impedance versus frequency for an electrolytic capacitor, (150 µF, 20 V at 85°C) showing the impedance minimum obtained.

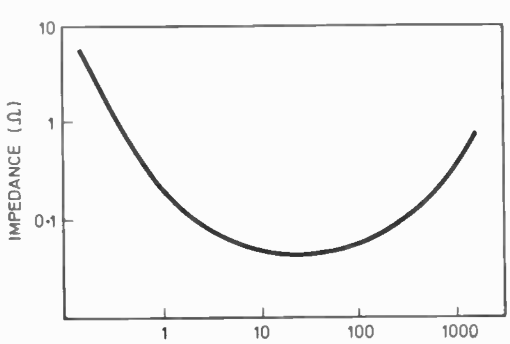

Fig. 6. Impedance versus frequency for 150 µF, 20 V capacitor (85°C).

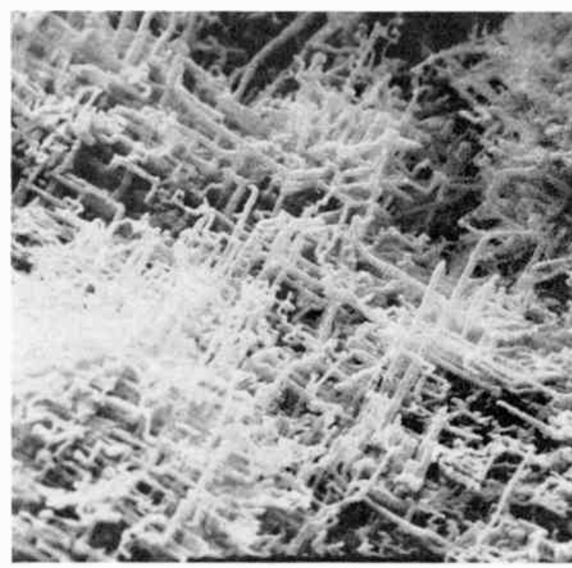

Fig. 7. Etched structure in aluminium foil. Anodized to 20 V (length of edge 40 µm).

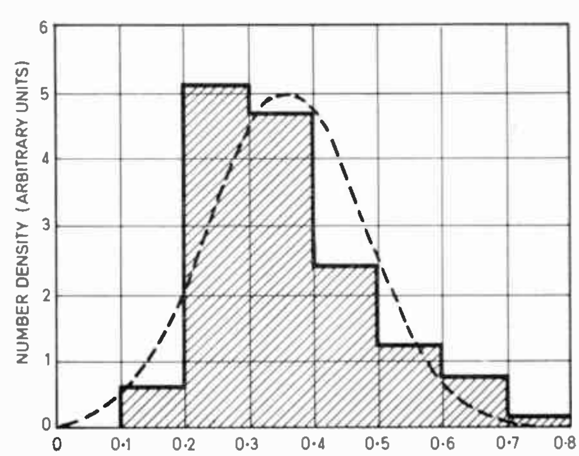

Fig. 8. Distribution of tunnel diameters for a practical low voltage foil (see Fig. 7).

4 Transmission Line Approach

The lumped equivalent components shown in Fig. 4 are however only an approximation to the practical situation that exists inside the electrolytic capacitor. Recent work has, therefore, been concerned with a transmission line approach to the problem to determine the effect of such an analysis on predicted performance, and furthermore to determine if the transmission line approach modifies the simple equivalent circuit.

There are two regions in which a transmission line analysis is possible. Firstly, it can be used to determine the effect of an etched tunnel structure filled with electrolyte, in both anode and cathode foils.[10] Secondly, it is possible to adapt a transmission line approach to the total wound component because of the distributed nature of the anode-electrolyte-cathode system.[12] Before dealing with this latter system, however, it is necessary to consider the transmission line equivalent of the anode and cathode tunnel structure.

4.1 *Transmission Line Approach to the Etched Anode and Cathode Foils*

The tunnels that are obtained by etching aluminium foil are in practice of variable length and variable width. However, it can be seen from the electron-micrograph of a typical etched structure, as used for an anode, shown in Fig. 7, that it is reasonable to assume that the tunnels are parallel sided. An estimation of the distribution of diameters can be obtained from such photographs. Figure 8 shows the distribution of tunnel diameters for such a practical foil used for low voltage applications. The length of the tunnels are difficult to measure from electron-micrographs but it has been found possible to assume a Gaussian distribution of lengths and a typical result is shown in Fig. 9.

To analyse the effect of this distribution on tunnel diameters and lengths, it is necessary in the first instance however to consider the behaviour of a single tunnel; Fig. 10 shows the parameters used in calculating the behaviour of the tunnel. The original radius of the tunnel is assumed to be a but on anodization part of the aluminium is converted to oxide increasing the final

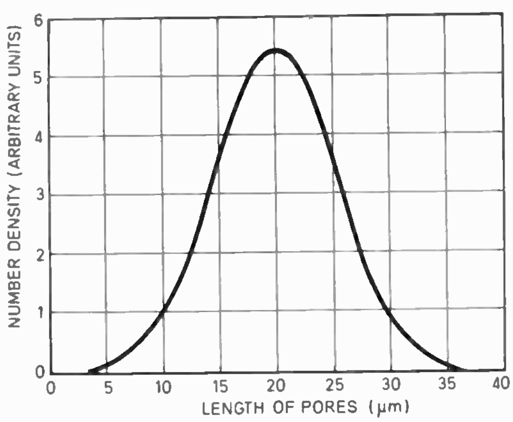

Fig. 9. Distribution of tunnel lengths for a practical low voltage foil (see Fig. 7).

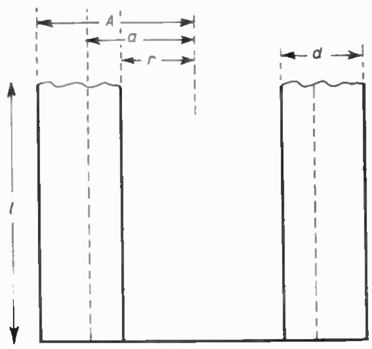

Fig. 10. Diagram of tunnel structure showing dimensions before and after anodization.

tunnel radius to A but reducing the remaining unoccupied radius to r, i.e. the thickness of oxide d is equal to $A-r$. It has been shown[7] that the volume of alumina produced to the volume of aluminium used is equal to 1·6 and therefore a new tunnel radius A and new inner radius can be calculated. These values are given by:

$$A = [(2{\cdot}56a^2 - 0{\cdot}6d^2)^{\frac{1}{2}} + d]/1{\cdot}6 \qquad (4)$$

$$r = [(2{\cdot}56a^2 - 0{\cdot}6d^2)^{\frac{1}{2}} - 0{\cdot}6d]/1{\cdot}6 \qquad (5)$$

The capacitance of a single tunnel, C_t, can then be calculated since

$$C_t = 2\pi\varepsilon_0 \varepsilon' l/\ln(A/r) \qquad (6)$$

where ε_0 is the permittivity of free space, ε' is the real part of the permittivity of the oxide and l is the tunnel length.

Similarly the resistance of the electrolyte, R_t in the tunnel of final radius r is equal to

$$R_t = \rho l/\pi r^2 \qquad (7)$$

where ρ is the resistivity of the electrolyte.

The inductance of a tunnel, L_t, for which $l \gg r$ is given by

$$L_t = \frac{2l\mu_0}{4\pi}\left[\ln\left(\frac{2l}{r}\right) - \frac{3}{4}\right] \qquad (8)$$

where μ_0 is the permeability of free space.

C_t, R_t and L_t are effectively distributed components.[8] It is however found in practice that the contributions of L_t to the transmission line total impedance is negligible and therefore the equivalent circuit for a single tunnel may be drawn as a resistive capacitive network only. This approach has been adopted previously by Broadbent.[9]

The equivalent transmission line of a single tunnel is shown in Fig. 11. The complex impedance Z^* of such a transmission line is given by

$$Z^* = \frac{R_t \coth \alpha l}{\alpha l} \qquad (9)$$

where

$$\alpha l = R_t C_t \omega(\mathrm{j} + \tan\delta').$$

$\tan\delta'$ is the real part of the loss of the oxide.

The transmission line may be simplified into a series capacitance and resistance but such terms will be

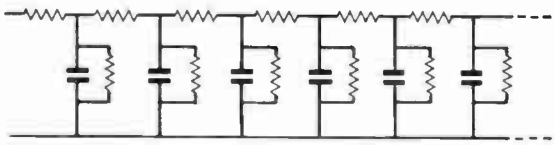

Fig. 11. Equivalent transmission line representation of a single tunnel.

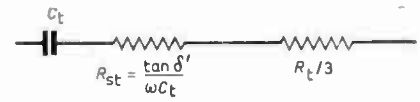

Fig. 12. Equivalent lumped circuit for representation of a single tunnel.

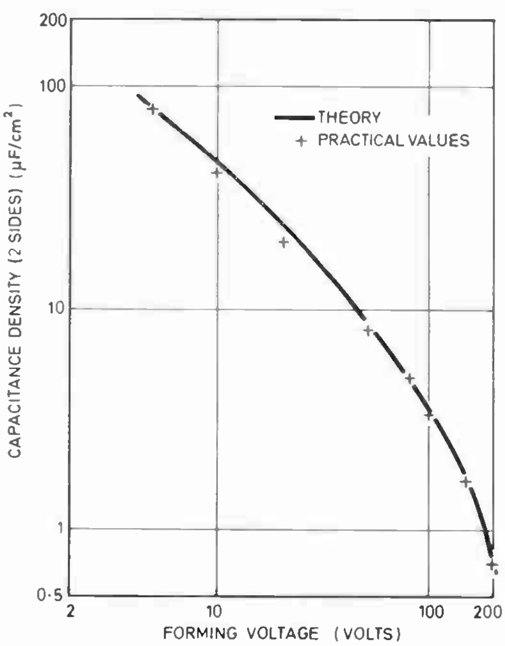

Fig. 13. Capacitance/unit area as a function of forming voltage (theory and practice).

frequency dependent. However, for low frequencies, less than the characteristic frequency f_c where f_c is given by[10]

$$f_c = 1/2\pi R_t C_t \qquad (10)$$

the capacitance is constant at a value of C_t. In practice, f_c is found to be greater than 1 kHz even for the thinnest oxide thickness (equivalent to 10 V forming voltage).

The loss of the transmission line, tan δ, is given by

$$\tan\delta = \tan\delta' + \omega C_t R_t/3 \qquad (11)$$

Equation (11) implies that the equivalent circuit for a simple tunnel is as shown in Fig. 12 where R_{st} is the series resistance of the tunnel oxide and tan δ' is equal to the loss tangent of the oxide.

It should be noted that such an approximation is not applicable above f_c in which case a full analysis must be undertaken. This had been done using a computer-aided analysis. Thus it is possible to determine the loss of a single tunnel at any frequency.

The analysis of a single tunnel may be extended to cover the distribution of tunnel diameters and tunnel lengths that are found in practice. The distribution of tunnel diameters has been shown in Fig. 8 and tunnel lengths in Fig. 9. The total number of tunnels, N_t at both sides of the surface must be calculated from the given capacitance per unit area at the lowest value of anodizing voltage. Figure 13 shows capacitance per unit area against forming voltage both theoretically and for a practical foil and it can be seen that there is good agreement between the theory using the distributions of Figs. 8 and 9 and practical values quoted by a manufacturer. It should be noted that the curve values fall below a straight line relationship between capacitance and voltage at the higher voltage levels due to the thinner pores filling in completely.

The equivalent circuit below a critical frequency, f'_c, for the summation of all the tunnels will be as shown in Fig. 14. The series resistance of the oxide in the tunnels is shown in Fig. 14 as being equivalent to the series resistance of the total anode oxide, as most of the oxide is in the tunnels. The critical frequency f'_c will be given by

$$f'_c = 1/2\pi \bar{R}_t \bar{C}_t$$

where $\bar{C}_t$ and $\bar{R}_t$ are the average tunnel capacitance and resistance.

The loss due to the summation of all the tunnels will be given by

$$\tan\delta = \tan\delta' + \omega \bar{C}_t \bar{R}_t/3 \qquad (12)$$

Figure 15 shows a plot of equation (12) for two frequencies as a function of anodizing voltage (i.e. of $\bar{C}_t$ and $\bar{R}_t$). The circuit of Fig. 14 and equation (12) is found to apply up to 180 V anodizing voltage (f'_c = 2·4 kHz). Above this voltage an exact solution must be calculated from full transmission line theory.

If the average value of length and tunnel diameter is taken it is found that calculated capacitance and loss are within 12% of full theory, below the critical frequency f'_c. f'_c is a function of capacitance and hence anodizing voltage, and Fig. 16 shows the curve of f'_c as a function of anodizing voltage.

The maximum in the critical frequency and the minimum in tan δ are both explained by the mathematical variation of the product $\bar{C}_t \bar{R}_t$ as a function of the forming voltage and hence oxide thickness. Initially, as the oxide is formed the capacitance contribution $\bar{C}_t$ decreases more rapidly than the resistance contribution $\bar{R}_t$ increases. However, at an oxide thickness $d = 0{\cdot}69a$ the roles reverse. This oxide thickness represents an optimum formation voltage, for not only does it represent a minimum tan δ and a maximum critical frequency, but also the point at which the gain starts to fall off rapidly with further increase in oxide thickness.

An exactly similar situation applies to the case of the cathode foil, with the simplification that in general only one thickness of oxide, i.e. forming voltage, is used.

4.2 *Transmission Line Approach to the Total Capacitor*

The total transmission line equivalent of the capacitor

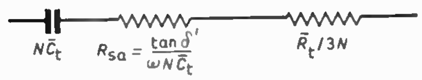

Fig. 14. Equivalent lumped circuit for representation of the summation of all the tunnels.

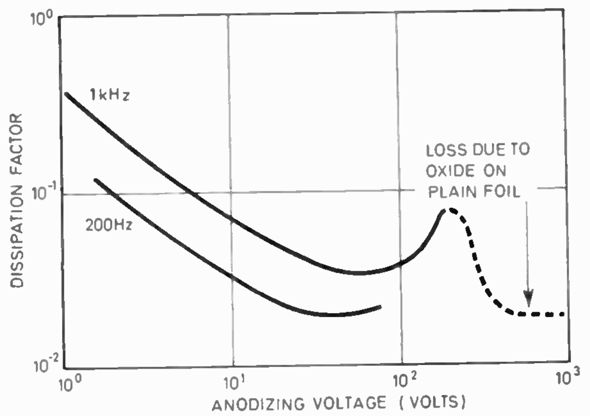

Fig. 15. Dissipation factor versus anodizing voltage for two frequencies.

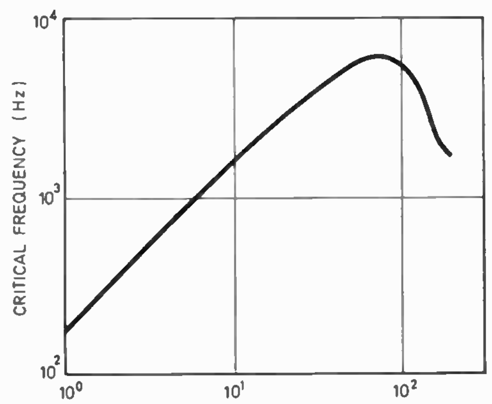

Fig. 16. f_c' versus volts.

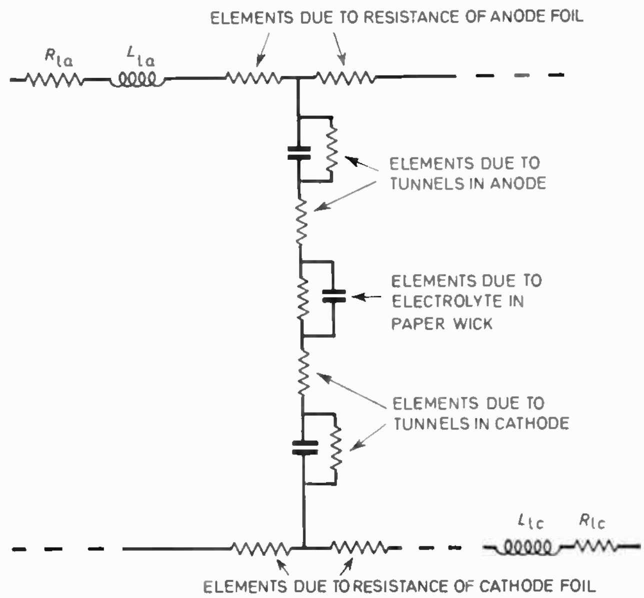

Fig. 17. Total transmission line equivalent of capacitor.

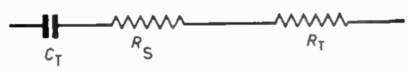

Fig. 18. Equivalent lumped circuit for representation of a complete capacitor.

can be drawn taking into account the distributed nature of the anode-cathode electrical parameters and the resistance of the anode and cathode foils. An element of this is shown in Fig. 17. Previous treatment has considered the transmission line effect of the pores[9] and other authors have considered the transmission line effect of the capacitor plate[11] but the effect of both analyses together has not been examined.

The characteristic frequency F_c of the total line will be given by

$$F_c = 1/2\pi(R_{fa}+R_{fc})C_T \tag{13}$$

where $R_{fa}+R_{fc}$ is the total series foil resistance and C_T is the total capacitance:

$$C_T = C_{sa}C_{sc}/(C_{sc}+C_{sa}) \tag{14}$$

Below F_c a simple equivalent circuit can be drawn and is shown in Fig. 18. R_s is the series resistance due to the dielectric

$$R_s = R_{sa}+R_{sc} \tag{15}$$

The remaining series resistance R_T will be given by

$$R_T = R_e + \left(\frac{\bar{R}_t}{3N}\right)_{\text{anode}} + \left(\frac{\bar{R}_t}{3N}\right)_{\text{cathode}} + R_{fa}/3 + R_{fc}/3 \tag{16}$$

Calculation of the terms in equation (16) is possible. R_e, the electrolyte resistance in the paper wick, is found by experiment to be approximately double that due to a similar volume of pure electrolyte, although it should be noted that higher figures have been obtained by other workers.[12] $(\bar{R}_t)_{\text{anode}}$ and $(\bar{R}_t)_{\text{cathode}}$ have been discussed previously. The resistance of the foils, R_{fa} and R_{fc} can be determined from the foil resistivity and dimensions. A factor k can be introduced as a multiplying factor to allow for the effect of the tunnel structures in the foil, but in practice it is found that k is so nearly equal to unity that it can be ignored.

4.3 *Dispersion in the Electrolyte*

It is found that the electrolyte used can show dispersion particularly if the resistivity is low. Figure 19 shows a practical curve obtained with a 300 Ω cm electrolyte in a kraft paper. However, the majority of electrolyte/paper combinations do not show as pronounced a change with frequency and therefore the need for introducing a parallel capacitance for this effect can in general be ignored.

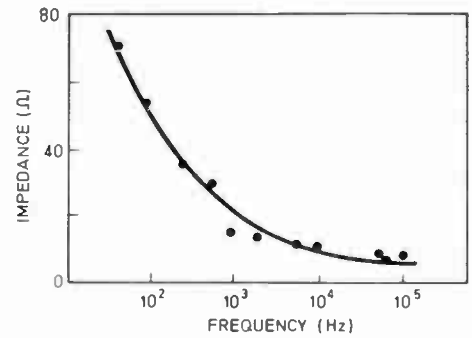

Fig. 19. Dispersion effects in electrolyte (300 Ω cm material).

4.4 *Winding*

As the final capacitor is wound up into a spiral it is necessary to examine the inductive effect of this construction. Previous workers have analysed a single foil construction in detail (i.e. an inductance with stray capacitance)[13] and a similar analysis can be undertaken on complete capacitors. However, the coupling between turns is extremely complex and the computer programmes involved are therefore lengthy. It would appear from initial results that the only inductive component that is of importance is that due to the lugs and it is this that gives the rise in impedance at high frequency (see Fig. 6).

4.5 *Total Circuit*

Taking all the effects into account an equivalent circuit can now be drawn provided the frequency is low enough for the lumped equivalent of the transmission lines used to analyse the effect of the tunnels and the effect of the distributed nature of capacitance and paper wick plus electrolyte resistance to hold. Under these circumstances a circuit such as shown in Fig. 20 is obtained. An important point to note is that the effect of the electrolyte in the tunnels and of the resistance of the foil are a third of that expected if lumped values had been considered.

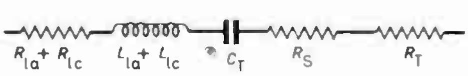

Fig. 20. Equivalent circuit of capacitor showing all the elements.

In Fig. 20 the components can be obtained as follows:

(i) R_{la} and R_{lc} from measurement.

(ii) L_{la} and L_{lc} from calculation.

(iii) C_T from equation (14) with C_{sa} and C_{sc} from knowledge of tunnel density, average tunnel length and diameter, and anodizing voltage.

(iv) R_s from equation (15) with R_{sa} and R_{sc} from the loss value of the oxide on the anode and cathode respectively and knowing C_{sa} and C_{sc}.

(v) R_T from equation (16).

The loss of the total capacitor, $\tan \delta_T$, will be given by

$$\tan \delta_T = \omega C_T(R_{sa}+R_{sc})+\omega C_T(R_T)+\omega C_T(R_{la}+R_{lc}) \quad (17)$$

5 Experimental Values

To examine the validity of the above analysis and in particular to determine the relative magnitude of the terms contributing to equations (14), (16) and (17) a particular production capacitor has been considered.

For a 300 μF, 7 V working capacitor which was constructed using foil with a tunnel structure as shown in Figs. 7, 8, and 9, the figures shown in Tables 1, 2, 3 and 4 were obtained.

The above figures show the extent of the agreement between theory and practice on capacitance and loss tangent measurements. The practical value of loss, shown in Table 4, is very subject to preparation conditions especially factors such as the attaching of the lugs to the foil, the resistivity of the electrolyte in the paper wick and the degree of impregnation. As an example the value of the resistivity of the electrolyte in the paper wick will affect the loss. A factor of 2 for the increase of resistivity of an electrolyte in paper has been taken, whereas if a value of 9 had been employed,[12] a final loss value of 6·6% would have been obtained. The calculations have assumed ideal preparation conditions. It should be noted that equations (16) and (17) do enable estimates to be made of the relative importance of changes from ideal preparation conditions.

6 Tantalum Electrolytic Capacitors

The preparation of tantalum capacitors has been discussed by various authors.[1] They can be prepared by methods exactly similar to those applicable to aluminium capacitors, or the metal body may be made by sintering to give a porous structure in which the anodized surfaces are contacted either by a normal wet electrolyte or by a manganese oxide layer to give a totally dry construction. The former of these structures can be analysed in a manner equivalent to that already presented for aluminium. However, no detailed analysis of foil has been undertaken and, in fact, the situation will not be so easy to analyse as the etched pores usually obtained in tantalum foil are not of a regular cylindrical nature.

An analysis of sintered porous structures has been made, where the effect of electrolyte in the pores has been analysed by transmission line techniques.[14] Such a treatment enabled the variation of capacitance and dissipation factor of cylindrical porous anode bodies to be calculated as a function of frequency. Figure 21 shows the type of result obtained.

In such a structure, the inductance effect due to a wound foil will be absent, but an impedance minimum will still be found due to the inductance of the contacts.

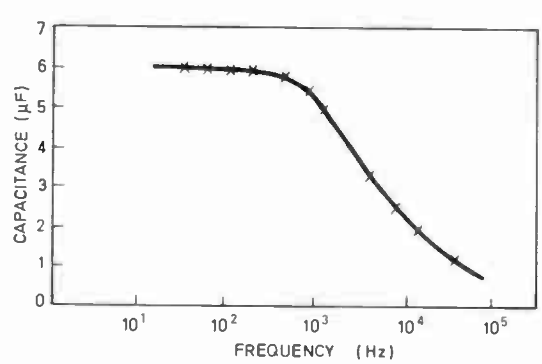

Fig. 21. Capacitance versus frequency for computed and experimental results on porous anode tantalum capacitor (after Morley[14]).

Table 1: Value obtained from structure measurements

	MEASURED	CALCULATED
Length of foil and paper wick	15 cm	—
Width of foil and paper wick	1·5 cm	—
Thickness of Al anode	50 μm	—
Thickness of Al cathode	25 μm	—
Thickness of paper wick	60 μm	—
Average length of anode tunnel	20 μm	—
Average length of cathode tunnel	—	10 μm
Radius of anode tunnel including oxide	0·175 μm	—
Radius of anode tunnel excluding oxide	—	0·150 μm
Radius of cathode tunnel including oxide	—	0·162 μm
Radius of cathode tunnel excluding oxide	—	0·158 μm
Inductance of capacitor	0·008 μH	—

Table 2: Values associated with capacitance measurements

	MEASURED	CALCULATED
Permittivity	8	—
Thickness of anode dielectric	25 nm	—
Thickness of cathode dielectric	4 nm	—
Capacitance of average anode tunnel	—	$5{\cdot}24 \times 10^{-8}$ μF
Gain of anode	30	—
Number of anode tunnels/cm²	—	$1{\cdot}6 \times 10^{8}$/cm²
Capacitance of anode tunnels/cm²	—	8·4 μF/cm²
Capacitance of unetched anode	—	0·28 μF/cm²
Total anode capacitance (C_{sa})	—	390 μF
Capacitance of average cathode tunnel	—	$17{\cdot}6 \times 10^{-8}$ μF
Gain of cathode	17	16
Capacitance of cathode tunnels/cm²	—	25·2 μF/cm²
Capacitance of unetched cathode	—	1·76 μF/cm²
Total cathode capacitance (C_{sc})	—	1210 μF
Value of C_T (eqn (14))	300 μF	294 μF

Table 3: Values associated with resistance measurements (equation (16))

	MEASURED	CALCULATED
Resistivity of Al	3×10^{-6} Ωcm	—
Resistivity of electrolyte	3×10^{2} Ωcm	—
Resistance of electrolyte		
in paper wick (R_e)	—	0·08 Ω
in anode tunnel ($\bar{R}_t$) anode	—	$0{\cdot}85 \times 10^{9}$ Ω
in cathode tunnel ($\bar{R}_t$) cathode	—	$0{\cdot}38 \times 10^{9}$ Ω
Resistance of anode foil (R_{fa}	6×10^{-3} Ω	6×10^{-3} Ω
Resistance of cathode foil (R_{fc})	13×10^{-3} Ω	12×10^{-3} Ω
Therefore remaining series resistance (R_T)	—	0·143 Ω

Table 4: Values associated with resistance measurements (equation (17))

	MEASURED	CALCULATED
Loss of anode dielectric (50 Hz)	2%	—
Series resistance of anode dielectric (R_{sa})	—	0·163 Ω
Loss of cathode dielectric (50 Hz)	5%	—
Series resistance of cathode dielectric (R_{sc})	—	0·132 Ω
Remaining series resistance (from Table 3)	—	0·143 Ω
Resistance of lugs (assumed) ($R_{la} + R_{lc}$)	0·003 Ω	—
Therefore tan δ_T (50 Hz)	7%	4%

7 Conclusions

It has been shown that it is possible to examine the construction of an aluminium electrolytic capacitor and determine the effect of construction on the various elements of a lumped equivalent circuit of a capacitor. The lumped values can be obtained from applying transmission line theory both to the behaviour of a single etched tunnel and to the total wound capacitor. Calculations have shown the relative importance of the lumped components in a typical aluminium electrolytic capacitor.

In the case of tantalum capacitors, similar analyses have not been made in detail, but reference is made to a transmission line approach that has been used to analyse the behaviour of sintered porous structures.

8 Acknowledgments

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Electron beam techniques for magnetic bubble device fabrication

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SUMMARY

By using the electron beam machine to expose an electron sensitive resist and then by use of the 'float-off' technique it is possible to produce the necessary permalloy overlay patterns for bubble devices with micron linewidths. Specification of the pattern is by computer program. At higher powers, the electron beam can also be used to damage the bubble material. This damage can be used to constrain the bubble motion.

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1 Introduction

A relative newcomer to the range of storage elements is the magnetic bubble device.[1,2] In such devices the presence or absence of a bubble—a cylindrical magnetic domain—at any memory location allows binary information to be stored. Bubbles can be moved between locations of the shift register by interaction with magnetic poles induced in a specially-shaped permalloy pattern by a rotating in-plane magnetic field. The permalloy is overlaid on the material supporting the bubble domains, which is conventionally an epitaxial film of a mixed rare earth iron garnet. The fabrication of the overlay pattern, with linewidths and spacings typically equal to the bubble radius and a thickness of a few hundred nanometres, is currently one of the limiting factors which determine the maximum storage packing density and to a certain extent the maximum speed of operation.

Some additional form of constraint upon the movement of the bubbles, as well as that provided by the permalloy overlay, would be desirable. This would be advantageous in the initial setting-up of a device, in performing logical functions, and also possibly in restricting the tendency of bubbles to be ejected from parts of the propagation circuit at high speed. The application of electron beam techniques provides a means for the fabrication of extremely small overlay structures and also a novel method of domain restraint.

2 Overlay Fabrication

Since typical linewidths and spacings of the permalloy propagation pattern are approximately half of the operational bubble diameter, then normal photolithographic techniques are barely adequate for forming high definition structures for use with bubbles of diameter much smaller than 5 μm.

To circumvent the problems of limiting optical resolution and undercutting during a chemical etching process, the technique of exposing a suitable resist using a beam of electrons and then forming the overlay pattern using the so-called 'float-off' or 'lift-off' method has recently been explored.[3,4] Electron beam exposure provides two advantages. Firstly, the effective wavelength of the electrons is substantially less than that of light and therefore the wavelength no longer sets the limit to the resolution. Second, a beam of electrons may be deflected with high precision, and this allows complex patterns to be exposed with greater accuracy. Furthermore, computerized control of the beam deflexion and exposure sequence, as described in this paper, facilitates easy and rapid change or modification of pattern which is lacking in photolithographic methods using an enlarged master drawing and repeated reduction stages.

2.1 *The Electron Beam Machine (e.b.m.)*

The e.b.m. is a purpose-built three-lens machine using an AEI type EM 6 gun operating typically at 10–20 kV during resist exposure. The specimen to be exposed is carried on a precision X-Y motion table assembly, movement of which is controlled by a pair of stepping motors. When focused the electron density over the beam cross-section has a Gaussian distribution. Defining the spot size by the half-height of the distribution curve, then spot diameters in the range 0·1 μm to 10 μm may be provided. The beam is deflected magnetically in X and Y by pairs of coils and can be blanked in less than 50 ns. The column of the machine is surrounded by magnetic shielding, and any remaining pickup (detected by operating the machine in the scanning electron microscope mode) finely trimmed out by injecting the appropriate cancelling signals into the octupole astigmator.

The basic working area of the machine is a square of side 1 mm. The table may be moved in 3 μm increments to extend the working area over a square of side 25 mm. Overall control of the system and the exposure sequence is provided by a computer with a direct link to the e.b.m. interface.

2.2 *Pattern Specification*

A pattern for exposure is specified by using a programming language and compiler developed for this purpose.[5]

Consider, for example, the simple Y-bar shift register shown in Fig. 1. When this structure of permalloy is subjected to a rotating field then in effect a moving magnetic pole pattern is set up which provides for bubble motion around the register. The square structure at bottom left is designed to inject new bubbles (i.e. new information) into the register by splitting an existing bubble into two, and is known as a bubble generator. A pattern element is specified by the coordinates of any single point on its outer edge and the appropriate horizontal, vertical and sloping lines necessary to develop its outline. If each element in the overall pattern was a different shape then each would need to be specified individually. But if, as is invariably the case in bubble devices, a few basic elements are repeated many times then a set of standard instructions can be used further to simplify the programming. These instructions allow the position of any element or any group of elements to be altered. They are, for instance; repeat n times, translate, rotate, mirror, delete, etc.

The program compiles the data, checking for syntax errors, and breaks the pattern down into basic shapes which the e.b.m. interface can accept. The breakdown of this pattern is shown in Fig. 2 the basic shapes being squares, rectangles, right-angled triangles and parallelograms. If the pattern is larger than 1 mm square the stepping motor instructions are produced for table motion. The interface then generates the appropriate scan raster to expose each basic shape with the correct size and positioning. The actual exposure is performed in a spot-wise fashion with the beam blanked between deflexion increments. The minimum standard deflexion increment is 0·25 μm on the present system (1 mm scan, 12-bit d-to-a converter).

If it is desired to use the electron-sensitive resist to mask against ordinary chemical etches it is necessary to expose the negative of the desired pattern. In this case the instruction NEGATIVE is inserted at the end of the data list, the resulting pattern breakdown being shown in Fig. 3.

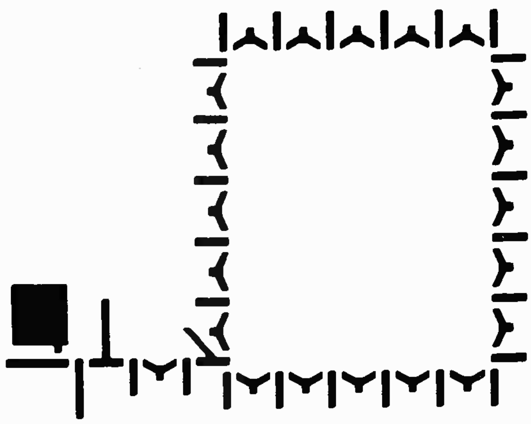

Fig. 1. Simple Y-bar shift register.

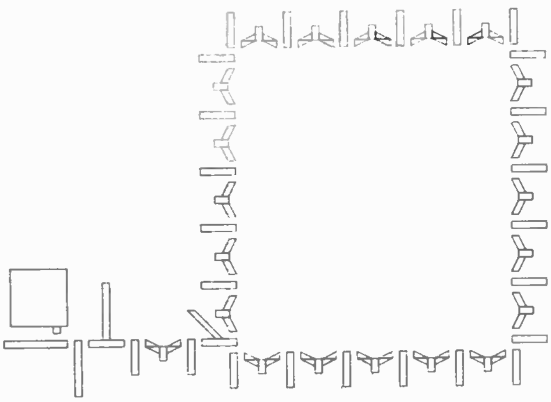

Fig. 2. Breakdown of pattern of Fig. 1 into basic shapes.

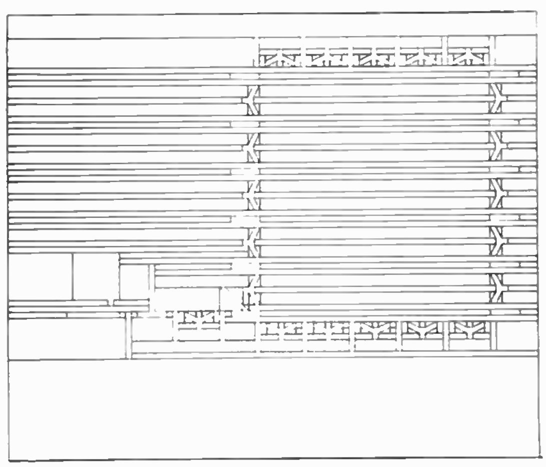

Fig. 3. Pattern breakdown to produce the negative of Fig. 1.

The ease of pattern programming for bubble devices is enhanced by the fact that practical memories usually consist of a few elementary shapes arranged in a high density array. Furthermore such standard shapes as T's, Y's, I's, chevrons, bars, bubble generators and annihilators and logic gates etc., can be held in a data library so that they do not need to be completely specified but can

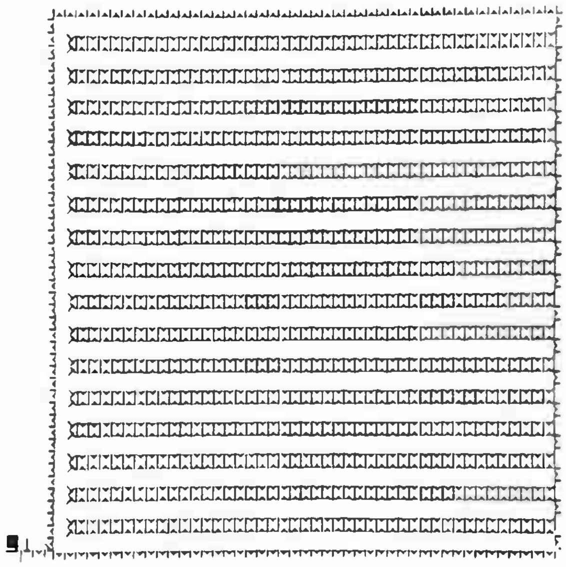

Fig. 4. Y-bar array.

be called up by name to the array as desired. Figure 4 shows such an example of a Y-bar array. This is a photograph taken from the display monitor with linewidths corresponding to approximately 2·5 μm in the finished circuit. The compiling of this register, which contains approximately 1500 data bits, took one minute of computing time and it would be exposed on line in approximately thirty seconds.

2.3 *The 'Lift-off' Process*

The electron-sensitive resist used in this investigation is poly (methyl methacrylate).[6] This is a positive-working degrading type of resist, insensitive to light. The resist is prepared by catalytic polymerization of the monomer. The polymer is then fractionated to produce a resist containing a narrow distribution of molecular weights for high resolution. The optimum molecular weight range lies between 50 000 and 500 000. Weights below this are removed to avoid pinhole formation in unexposed areas on development. The higher weights are removed to reduce the intrinsic viscosity in solution and hence improve the characteristics for spin coating.

The basic steps in the 'lift-off' process of overlay fabrication are shown in Fig. 5. A layer of resist is spun on to a substrate and air baked to improve adhesion. The required pattern is then exposed producing scissions in the polymer and greatly reducing its molecular weight. The exposed polymer is removed by development in a mixture of 4-methylpentan-2-one and propan-2-ol. Due to substrate backscatter and forward scattering in the polymer film the resist exhibits an undercut profile. If now a layer of permalloy is deposited overall, then the undercutting produces a physical separation between the permalloy adhering firmly to the substrate and that lying on top of the unexposed resist. The structure fabrication is completed when the unwanted permalloy is floated off by dissolving away the remaining resist in chloroform. It has been found worthwhile to use ultrasonic agitation during this last part of the process to ensure complete removal of the excess permalloy.

Because the polymer is an insulator, problems can arise due to surface charges produced during exposure which can deflect the beam and produce pattern distortion. This charge may be removed by evaporating a layer of aluminium approximately 30 nm thick over the resist before exposure. A layer of this thickness produces negligible defocusing of the beam and is removed before the resist is developed.

Rigorous cleaning of the specimen is necessary after resist development to ensure good adhesion of the thick permalloy film to the substrate. This is particularly the case since the substrate temperature during deposition must not exceed 100°C because of flowing of the resist above this temperature with consequent loss of the undercut profile.

Present effort is concentrated on overlays of 300 nm thick permalloy with 2·5 μm linewidths, since these dimensions seem optimum for devices using bubbles of diameter 5 μm to 6 μm. Edge and corner resolution is sufficiently good that the extension of this technique to sub-micron linewidths may be envisaged, enabling memories with greatly increased packing density to be produced. The resist is equally suitable for use as a mask to define permalloy shapes by electroplating. This latter method is now being explored as an even more convenient way of producing the overlays since, because of the low substrate temperature during vacuum deposition, the permalloy produced during the lift-off process sometimes needs to be annealed to provide acceptably low coercivity.

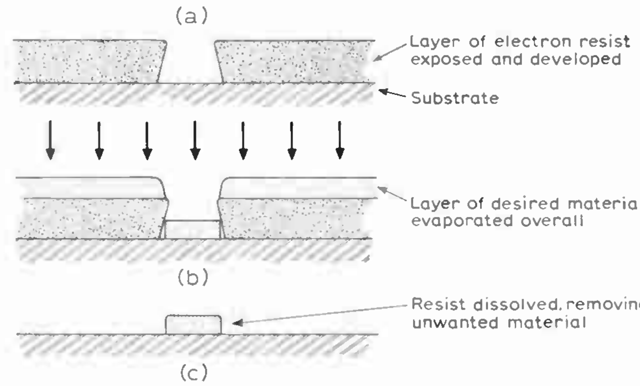

Fig. 5. Stages in the 'lift-off' process.

3 Electron Beam Damage

A further application of the e.b.m. to the production of bubble devices is the ability to constrain the motion of domains by introducing damage sites into the bubble-supporting material. It has been recognized for some time that naturally-occurring material defects impede domain motion,[7,8] and the aim in preparing and processing the bubble material is invariably to eliminate such defects. The operation of the e.b.m. at very high power densities enables damage of controlled extent and precise positioning to be introduced and provides a novel means of domain constraint.

Experiments have been conducted with flux-grown samarium terbium orthoferrite processed to a thickness of 50 μm and europium erbium gallium iron garnet

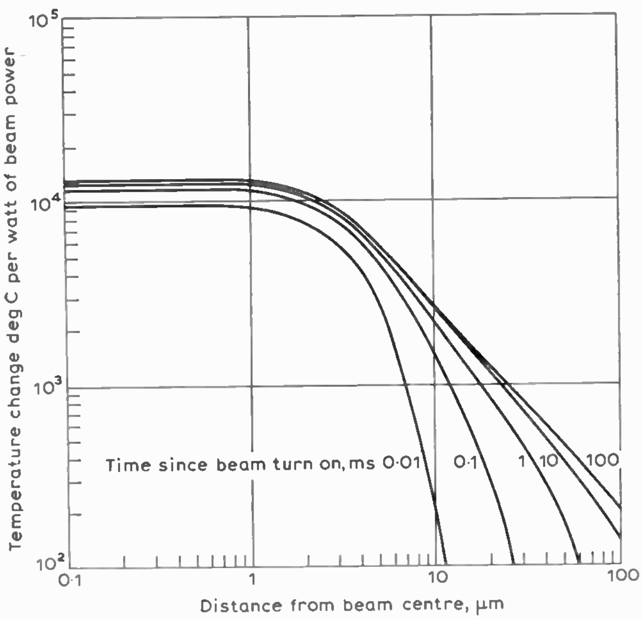

Fig. 6. Surface temperature rise for various exposure times of incident electron beam power.

grown by liquid phase epitaxy on a gadolinium gallium garnet substrate. In all cases no interaction was observed between the domains and the regions of the specimens exposed to the electron beam unless there was visible damage to the specimen. The onset and extent of thermal damage (i.e. melting of the slice surface) was found to be a complex function of beam diameter, beam current, accelerating voltage and exposure time.

Heating caused by the exposure of adjacent regions is another factor which determines the degree of damage. The use of a pulsed beam aids in localizing the damaged area.[9]

Calculations of the surface temperature rise under various beam conditions are published more fully elsewhere.[10] Figure 6 shows the temperature rise along the surface, per watt of incident beam power, for various exposure times, assuming an 8 μm diameter spot incident upon $Eu_2Er_1Ga_{0\cdot7}Fe_{4\cdot3}O_{12}$ material. To allow for electron penetration the appropriate correction factor must be applied, which for 20 kV accelerating potential is 0·7.

Thus the theory predicts that for an 8 μm spot, application of a 20 kV, 10 μA beam for a time in the range of 10 ms to 100 ms will produce a temperature rise of approximately 1800 degC at the centre of the beam. This temperature is in excess of the melting point of the garnet and leads to significant evaporation of material.

Figure 7 shows the interaction between domains and the damaged region when four such spots were exposed in $Sm_{0\cdot55}Tb_{0\cdot45}FeO_3$ with a spacing of approximately 50 μm. The intention was to capture a bubble between the spots as the bias field was raised through the strip-to-bubble transition, and thus to test whether this was a possible method of ensuring that a bubble was always present under a domain generator structure. However, only one spot was really necessary as can be seen from the sequence of photographs at increasing values of the bias field.

A point to note is that the wall of the domain is not attracted by the spot. This is in contrast to the results of other workers[7] who report strong attraction of the wall to defects such as twin boundaries, grain boundaries, dislocations etc. When a straight wall is swept through the region of a damaged spot, the wall is found to be repelled until it is forced into the actual damaged area. Only at this stage does the interaction become attractive with a coercivity between 160 Am^{-1} and 1·6 kAm^{-1} depending on the degree of damage. The mechanism of localized bubble formation on a damaged spot is evidently due to the reluctance of the domain walls to approach and pass through the spot, rather than for domain walls to be attracted to and held by the spot.

With a beam current of 7 μA, a spot size of 6 μm and an accelerating voltage of 25 kV a pattern in the form of the outline of a rectangle was exposed in epitaxial garnet. One of the corners of the rectangle is seen in Fig. 8. The

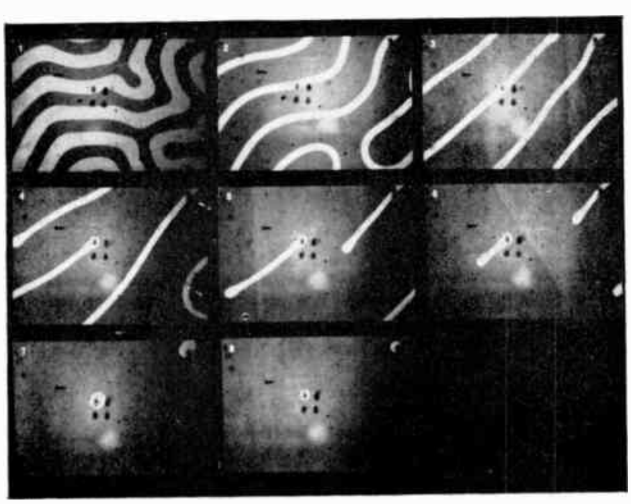

Fig. 7. Interaction between domains and damaged region in samarium terbium orthoferrite. Bias field increases in sequence 1–8.

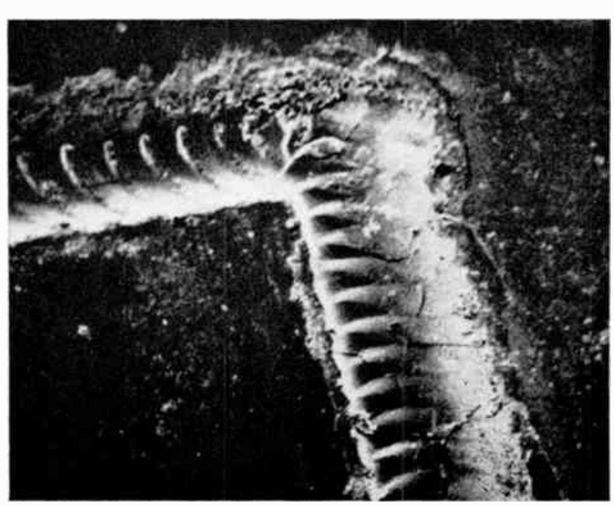

Fig. 8. Corner of rectangular pattern in epitaxial garnet.

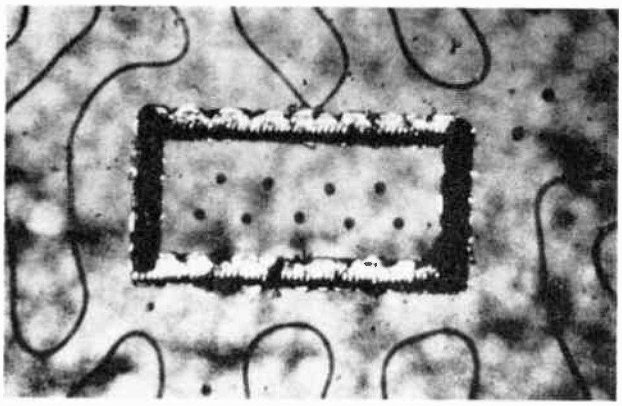

Fig. 9. Damage-domain interaction.

width across the machined track is approximately 20 μm. Fine cracking is apparent, though this is confined to the immediate region of exposure.

The repulsive nature of the damage-domain interaction is again demonstrated in Fig. 9, where the domain structure within the rectangle was split into bubbles using an oscillating bias field. The spacing of the bubbles is quite regular as the bubble's mutual repulsion is balanced by that of the damaged outline.

Such a repulsive interaction could be used in devices to perform logic functions or to channel the motion of domains along specified tracks, and the reliable capture of bubbles in specific positions as demonstrated in Fig. 7 could further contribute to logical operations.

4 Conclusion

The electron beam machine provides a powerful tool for the fabrication of bubble domain devices. It provides a means of overlay formation with linewidths in the micron region. Further, the specification of a desired pattern is readily and rapidly changed by computer program instead of by production of new artwork.

The use of the electron beam to damage the bubble supporting material provides a possible constraint on bubbles to ensure that generators are always loaded and to help steer bubbles in logic gates or at the corners of propagation paths.

5 Acknowledgment

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Circuit technology in a large computer system

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SUMMARY

In the design of a large high-speed computer, the size of the system leads to long cable delays for data transmitted between different parts of the machine. This problem and the interconnexion of a high-speed e.c.l. circuit family are discussed, and comment is made on future lines of development for technology in high-speed computers.

Priority circuits in a large asynchronous system also present difficulties not usually encountered in smaller machines and a discussion of how these difficulties arise is presented.

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1 Introduction

In the time between the design of the *Atlas* computer and the start of the *MU5* project at Manchester University,[1] considerable advances in circuit and memory technology have been made. The basic *Atlas* gate was a discrete component circuit with a propagation delay of approximately 16 ns and a power dissipation of 250 mW. These circuits were assembled on plug-in printed circuit cards so that the overall packing density was approximately 300 gates per ft^3 (10 600/m^3). The main store used a number of stacks of 2 μs ferrite core storage. The corresponding features of the *MU5* system are e.c.l. circuits containing 2–3 gates per chip with a propagation delay of approximately 2 ns per gate, and four stacks of plated wire store with a cycle time of 260 ns. All interconnexions in *MU5* are driven as matched transmission lines, which require relatively high currents in the gate outputs and the average power dissipation is 200 mW per gate.

In the period of 10 years between the *Atlas* and the *MU5* systems, the speed of the circuits has therefore increased by a factor of about 8:1, but the power dissipation per gate has not been significantly reduced. At the same time reductions on physical size due to the introduction of integrated circuit techniques have increased the packing density to over 2000 gates per ft^3 (70 000/m^3) in certain areas of *MU5*. This improvement in volume where the logic circuits are concentrated is not maintained in the overall system when the requirements of cooling and power supply distribution are taken into account, and this effect is more significant in *MU5* because of the high power dissipation per unit volume.

An increase in speed of 8:1 in both the logic circuits and the main store implies a corresponding improvement in the system speed, but a comparison of the access times to the store for the two systems gives 1·8 μs in *Atlas* and 585 ns in *MU5*, or a factor of just over 3:1.

These times include the delay time through the current page registers required for address translation in a paged system[2] and for the organization of four store stacks whose cycle times are overlapped to give a high rate of instructions and operands to the processor.

The relatively long access time in *MU5* is mainly due to the time taken for signals to travel between widely separated parts of the system, and also to a lesser extent the priority decision circuits required in an asynchronous system. While the volume occupied by the circuits has been considerably reduced, the average interconnexion length is related to the linear dimensions rather than the volume and an improvement of only about 2:1 in cable delay time has been achieved in practice.

1.1 *The MU5 System*

A block diagram of the *MU5* central processor is shown in Fig. 1. This is made up of a number of units each concerned with a part of the activity involved in executing an order. Instructions are obtained from the local store via the store access control (SAC) which contains the current page registers (CPRS) required to translate the virtual addresses used by the processor into

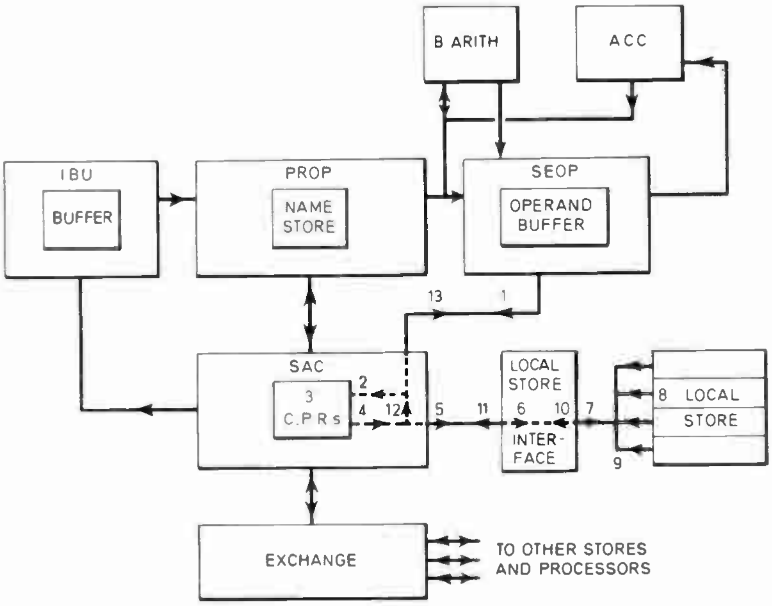

Fig. 1. The *MU5* central processor

the real addresses in the store. Instruction accesses are made in 128-bit groups, stored in an instruction buffer and passed in 16-bit groups to the primary operand unit (PROP).[3] This unit interprets the instruction, and accesses the operand specified directly. This operand may be used by the primary operand unit itself or by the index arithmetic unit (B-ARITH), or indirectly by the secondary operand unit (SEOP) as a data descriptor to access elements in data structures such as arrays which are processed in the accumulator (ACC).

The paths taken by the address and data in a typical store access for an element in a data structure are identified in the diagram and are as follows:

1. Cable delay from operand buffer to SAC	35 ns
2. Priority logic at input of SAC	65 ns
3. Address translation in CPRS	120 ns
4. Store overlapping priority logic	65 ns
5. Cable delay to store interface logic	25 ns
6. Address buffering in store interface	15 ns
7. Cable delay to store stack	25 ns
8. Store access	125 ns
9. Cable delay to interface logic	25 ns
10. Data buffering store interface	10 ns
11. Cable to SAC	25 ns
12. SAC	15 ns
13. Cable to operand buffer	35 ns
	585 ns

Cable delays and priority circuits thus represent over half the total access time.

In order to overcome this problem, and also to help achieve the design aim of a performance improvement at least a factor of 20 over *Atlas*, three separate buffer stores have been provided in the system to reduce the access time for the different types of data used. One buffer is provided for instruction, one for PROP named variables[4] and one for array elements, named variables and literals used in the accumulator. Studies of the patterns of accesses made by selected programs show that the majority of these accesses are for named variables within a routine, and that only a small number of these variables are in use at a given time.[5] The buffers can therefore be relatively small and it is estimated that a name store with a capacity of only 32 variables will be sufficient to hold the values required by over 99% of named primary operand accesses.[6] The rate of execution of instructions is also improved by means of a pipeline type of construction[3] in which several partially completed orders are in progress concurrently.

The use of pipeline techniques allows an average rate of one instruction executed every 120 ns and a peak rate of one every 40 ns. Most operands and instructions can be obtained from the small buffer stores situated close to the point of use so that the access time to each type of data is relatively short and the number of requests to the local store is reduced to an average of one 64-bit access every 210 ns for instructions and one 64-bit access every 800 ns for operands.

2 Circuit Family and Interconnexions

The range of integrated circuit devices used in *MU5* is shown in Fig. 2 which also indicates the number of each type used. The pipeline type of construction is reflected here by the relatively large number of dual flip-flop devices, mainly used as storage registers in the various stages of the pipeline. Between pipeline stages gating of data and decoding, etc., is mainly performed by multi-input and/or gates of the type shown in column 5. The most complex devices used are the 16-bit random access memory and 8-bit associative memory chips used in the associatively addressed buffer stores,[4] and translation circuits from e.c.l. levels to levels compatible with the associative devices account for most of the flat packs, discrete transistors and diodes.

The majority of the integrated circuits in the system are

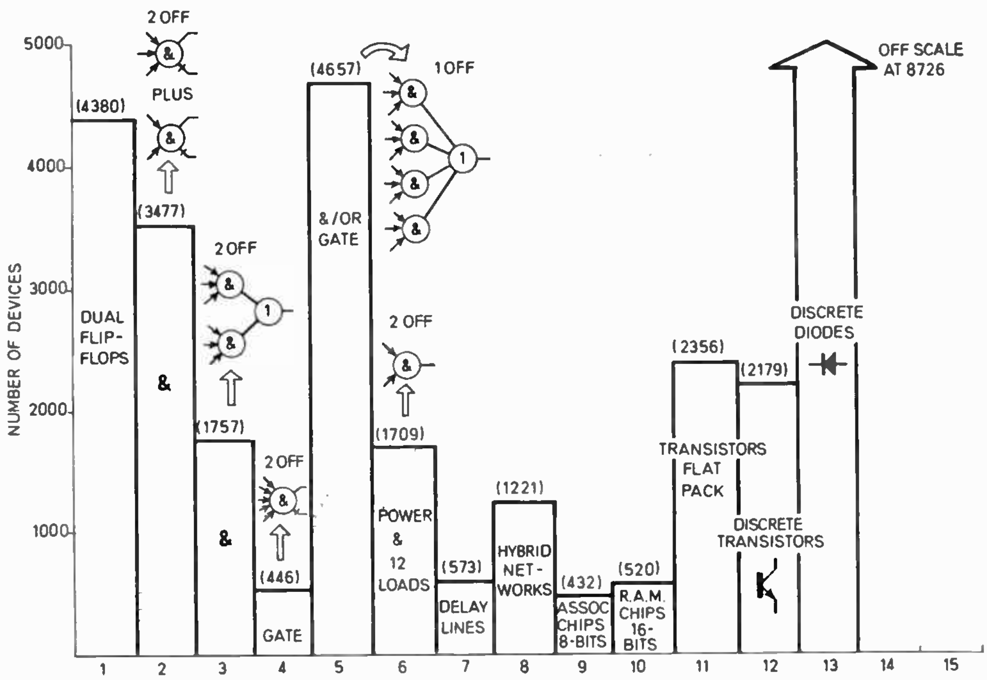

Fig. 2. *MU5* device types (excluding memory systems over 1k words (32 bits) capacity).

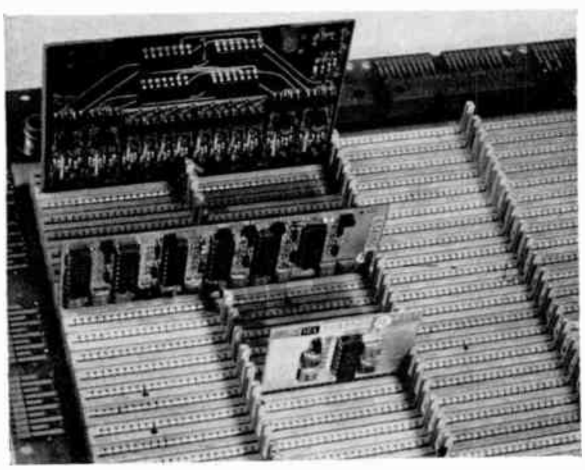

Fig. 3. *MU5* modules.

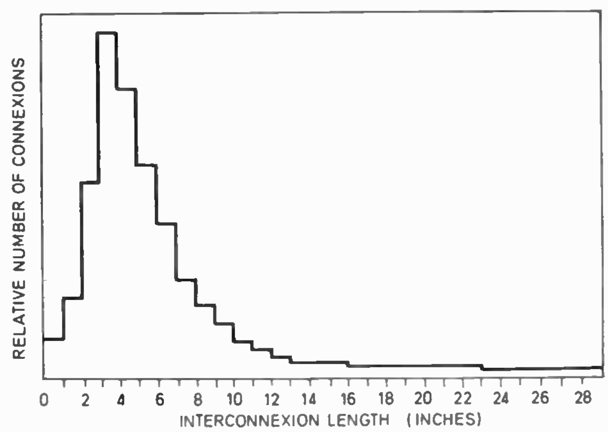

Fig. 4. Length of interconnexions in *MU5*

mounted on plug-in printed circuit modules 41 mm × 53 mm (1·6 in × 2·1 in) with 20 pins.[7] Up to 200 of these modules can be interconnected by means of a single 12-layer printed circuit platter as shown in Fig. 3. The packing density of circuits on these modules however is relatively poor and commonly used complex function macros such as adders[8] and a 32-bit associative store have been designed on 41 mm × 112 mm (1·6 in × 4·4 in) and 76 mm × 112 mm (3·0 in × 4·4 in) boards. These are also shown in Fig. 3. The platters in the *MU5* system measure approximately 33 cm × 41 cm and up to 33 platters can be contained within a logic bay. Within a bay 24 platters are mounted on movable doors to allow easy access for maintenance and a further nine are mounted on a fixed central plane. This arrangement gives relatively short runs for interconnexions whose source and destination are situated on the same platter but interconnexions, crossing platter boundaries must travel an average of 30 cm (12 in), and those passing from a fixed plane to a door, or from one side of a door to the other travel an average of 2·4 m (8 ft) along coaxial cables. A histogram of the distribution of interconnexion lengths in the *MU5* system, measured from the output pin of the source circuit to the input pin of the destination circuit is shown in Fig. 4.

Since a maximum of 3 integrated circuits can be mounted on a 20-pin module or 7 circuits on a 40-pin module relatively few interconnexions are between circuits on the same module, and a typical connexion travels a distance of about 25 mm (1 in) to reach the platter, 50 mm (2 in) on the platter and a further 25 mm from the platter to its destination on a second module.

Within the *MU5* system approximately 80% of all interconnexions are between integrated circuits on the same platter, 13% travel between adjacent platters, 5%

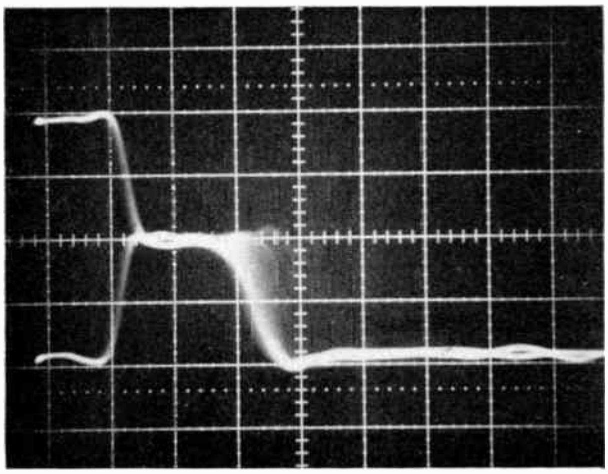

Fig. 5. E.c.l. flip-flop settling time.

go through cables to other platters in the same bay and 2% travel from bay to bay, distances of up to 15·2 m (50 ft).

E.c.l. circuits with typical edge speeds of 2 ns are used, driving interconnexions which are typically 50 mm in length but vary up to 15 m and average 15 cm between gates on the same platter. The relatively long delay time of these interconnexions dictates the use of a matched transmission line approach to minimize the possibility of reflexions. A series matching technique has been used in *MU5* in which the e.c.l. gates can drive two matched 75Ω lines from each output, and each line is capable of driving up to two inputs at the receiving end. Associated with every output is a group of three resistors, the output load resistor and two series matching resistors. These are fabricated on a ceramic chip in thick film technology and over 45,000 groups are required in the system. Because of the capacitance represented by each input and the series resistance of the line matching resistor the rise time at the input gate is degraded and an effective extra delay of 0·5 ns per load introduced. An average gate thus introduces a delay of 2 ns due to propagation through the e.c.l. circuit itself, 1 ns transmission time along the 15 cm interconnexion path and a further 1 ns delay from the input loading. Typical delay per gate in a system is therefore approximately 4 ns, and coaxial cables up to 15 m long are driven without difficulty by the circuits.

3 Asynchronous Timing

Each of the units shown in Fig. 1 has its own internal timing, functions and data being passed from one unit to another when the sending unit has the data available and when the receiving unit is not busy. This type of operation in which data transfers take place asynchronously generally allows the system to operate at a greater speed than a completely synchronous system where transfers can only take place at fixed times, but can lead to difficulties in the circuits controlling the transfers.

This problem occurs in a number of places in the system but can be illustrated by the paths from the three buffer stores through the store access control to the local store. Here three different units may request a store cycle at any time, and a decision must be taken by the store access control as to when the request can be accepted, so that the unit issuing the request can free its output address and data buffers for other activities.

In this particular example the requests are first staticized in flip-flops. On the basis of what type of requests are outstanding, and the state of the pipeline, a combinational circuit indicates which of these requests need to be serviced. This process takes a time of 15 ns from the initial receipt of a request, and 5 ns later the strobe for three decision flip-flops is set. When the unit next becomes free the strobe is removed and the state of the decision flip-flops indicates which requests require action at that time.

Because a second request may occur a short time after the first request it is possible for the inputs to the decision flip-flops to change state just before the strobe is removed, leaving the outputs midway between a '0' and a '1' level. Under these conditions the time taken for the flip-flop output to reach a constant level may be long compared with its normal propagation delay as shown in Fig. 5 and it is possible for subsequent circuits designed to select the highest priority request for servicing, to give an output inconsistent with either '0' or '1' inputs during this period.

Clearly sufficient time must be allowed here for the decision flip-flops to settle to a constant level or failure of the control circuits may occur. The settling time of the circuit used here is a function of its gain-bandwidth product, and the displacement of the output from the mid-level at the time the strobe is removed. In the case of an output starting from the exact mid-level, it is possible for an infinitely long settling time to be required, but the probability of such an occurrence is extremely low.

A graph of the mean time between failures of a typical control system against time allowed for settling is shown in Fig. 6. This data was obtained experimentally and indicates that an e.c.l. flip-flop with a propagation delay of 2·2 ns requires over 30 ns settling time for the failure rate to be reduced to an acceptable level. These results correlate well with the Appendix which shows that the mean time between failures is given by

$$\text{m.t.b.f.} \simeq \frac{T_1 T_2}{\tau'} \exp\left(\frac{t}{\tau'}\right)$$

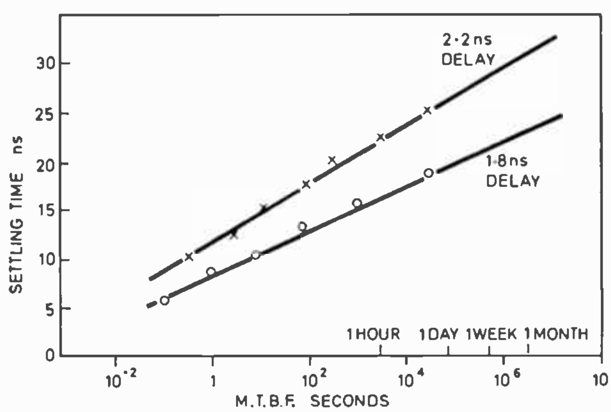

Fig. 6. Performance of e.c.l. flip-flops.

where T_1 = mean time between strobes
T_2 = mean time between requests
$1/2\pi\tau'$ = gain-band width product of flip-flop
t = settling time allowed

Since the number of decision flip-flops required here is relatively small considerable advantage can be gained in the system by using a special circuit with a higher gain-band width product than the standard device and a flip-flop with 1·8 ns propagation delay requires only 20 ns settling time here.

The total priority circuit time from the receipt of a request to its acceptance into the input buffer of the s.a.c. using the faster flip-flop is 65 ns and together with the cable delays still adds significantly to the communication time between the processor and the store.

4 Future Developments

Families of logic circuits are now available with propagation delays as low as 1 ns/gate, but a corresponding improvement in system performance will not necessarily be obtained by the use of these circuits alone. Other techniques must also be introduced to improve the packing density of circuits and reduce the interconnexion delay between gates. In addition to the interconnexion delay problem, system speed is often critically dependent on a relatively few areas such as fast priority circuits and accurate clock pulse drivers. In order to obtain the maximum performance in the system these control circuits should be faster than the standard circuits used in data paths.

The packing density of gates in a system can be improved by increasing the number of gates per integrated circuit package. Whilst the number of gates per package can be increased substantially by the use of medium and large-scale integration techniques, the number of connexions required to other packages is also a function of the number of gates in the package and has been quoted[9] as

$$p = 3{\cdot}5\, n^{0{\cdot}75}$$

for a high performance system, where p is the number of pins and n the number of gates. This expression correlates well with experinece in the *MU5* system, and Table 1 shows the relationship between the number of pins available for interconnexion and the number of gates actually accommodated on modules and platters.

Table 1

	Signal pins (p)	Gates (n)	Actual gates accommodated
Small module	18	9	7·5
Large module	36	25	20
Platter	360	480	500

$$p = 3{\cdot}5\, n^{0{\cdot}75}$$

Sixteen-pin dual-in-line packages are used in *MU5* to mount the integrated circuits and larger packages could be used to accommodate more complex functions, but the packing density is now limited by the number of interconnexion pins available on standard packages.

The number of signal pins available on typical dual-in-line packages is given in Table 2, together with the number of gates which can be supported by each package. This Table clearly shows that the improved density of gates per package given by large-scale integration is counterbalanced by an increase in package size so that the overall printed circuit board packing density has not been significantly improved, and interconnexion delays will still represent a large part of the total system delays.

In order to reduce the interconnexion delays to a level compatible with 1 ns gates, the packing density must be further increased, but the use of dual-in-line packages to accommodate the semiconductor chips presents a severe limitation on the number of interconnexion pins available. Packages containing 300 gates would typically require 250 pins, and the area occupied would not be significantly less than 10 individually packaged 30-gate arrays. Improvements in the packaging leading to a greater number of interconnexion pins per unit are required before full use of the potential of large scale integration can be made.

Table 2

Package	Signal pins (p)	Gates (n)	Package size	Board packing density (gates/in^2)
16-pin	14	6	0·3 in × 0·9 in	10·9
24-pin	22	10	0·6 in × 1·3 in	8·3
36-pin	34	21	0·6 in × 1·9 in	12·5

The use of more complex semiconductor devices in a high performance system requires a relatively large number of different circuit types, but the small number of each type used and the long production cycle involved for minor design changes make this approach economically unattractive.

Thick film hybrid circuit modules, such as that shown in Fig. 7, interconnecting a standard range of unpackaged

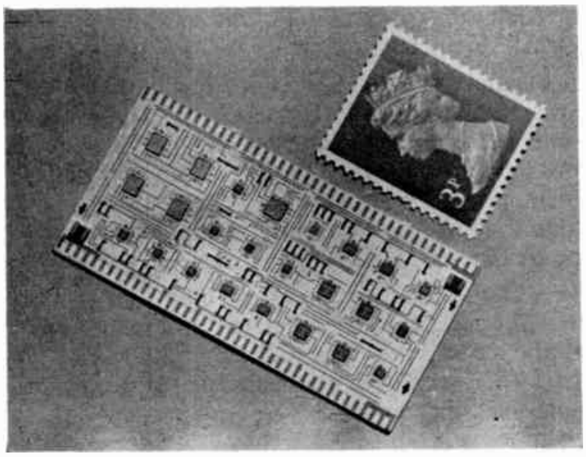

Fig. 7. Hybrid circuit module containing 23 integrated circuits.

semiconductor chips are another approach which allows high packing density and a relatively short production cycle for modification. This technique also allows the use of semiconductor devices in volume production, and the possibility of tailoring each circuit block to the system requirements with acceptable producton costs. Developments in this area may provide the best technology for future large fast computer systems.

5 Acknowledgments

The *MU5* project is supported by the Science Research Council and by ICL. The authors would like to thank all members of the University and ICL involved in the project for many helpful discussions and particularly Professors T. Kilburn and F. H. Sumner.

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7 Appendix: Flip-flop Settling Time

One of the problems in the design of priority circuits in asynchronous machines is the increase in settling time which can occur in flip-flop circuits operating under abnormal drive conditions.

In a simple latch-type flip-flop, the output normally follows the input while the clock or strobe signal is present. However, if the input changes a short time before the strobe is removed, the output may change to the new state or stay in the old state, depending upon the overlap time of the strobe and the input signal, and the response time of the flip-flop.

At a certain critical overlap time only just sufficient energy is received by the flip-flop to bias it to the switching level and subsequently the switching time will be dominated by its small signal response in the linear region around the switching level.

Most flip-flop circuits are designed as cross-coupled gates and, when the gates are biased to their switching level, can be regarded as linear amplifiers with 100% positive feedback.

In order to calculate the response time of this system, the transfer function of the amplifier must be known, and here a relatively simple function is assumed for the amplifier gain:

$$A = \frac{A_0}{1+S\tau}$$

where τ is the response time-constant and A_0 the low-frequency gain under small-signal conditions.

The overall gain with feedback is

$$G = \frac{A_0}{1-A_0+S\tau}$$

$$= \frac{A_0}{1-A_0}\left[\frac{1}{1+S\dfrac{\tau}{1-A_0}}\right]$$

Putting $\tau/(1-A_0) = -\tau'$:

$$\frac{\tau}{1-A_0} = -\tau': \quad G = \frac{A_0}{1-A_0}\left[\frac{1}{1-S\tau'}\right]$$

The response to a unit impulse is

$$\frac{A_0}{A_0-1} \cdot \frac{1}{\tau'}\exp\left(\frac{t}{\tau'}\right)$$

and response to an impulse of unit voltage and x seconds long; provided x is very small compared with τ' is

$$\frac{A_0}{A_0-1} \cdot \frac{x}{\tau'}\exp\left(\frac{t}{\tau'}\right)$$

Rearranging this equation, the time t taken to respond to an output of unit voltage after an impulse of x seconds is

$$t = \tau' \log_e \frac{A_0-1}{A_0} \cdot \frac{\tau'}{x}$$

Since the shape of the response curve is a positive exponential, the greater part of the response time is spent with a small deviation from the zero, or centre level, and the linear model should provide a reasonably accurate measure of the total time for the flip-flop to respond to a logic level when disturbed from the centre.

In order to calculate the time which must be allowed for the flip-flop to settle down in a given situation, the strobe frequency and the time between successive events which may change the flip-flop state must be known.

Assume that the time between successive strobes is T_2 and the time between events is T_1.

During the time T_1, T_1/T_2 strobes will occur, and the number of strobes occurring y seconds or less after a given event during T_1 is y/T_2.

The number of times this will happen in a long period N seconds is given by

$$\frac{N}{T_1} \cdot \frac{y}{T_2}$$

If it takes an impulse z seconds wide to bring the flip-flop to the linear region, the resulting disturb impulse difference x is given by $x = y-z$ and the number of times a strobe occurs within $\pm x$ of an event is given by

$$\frac{N}{T_1} \cdot \frac{2x}{T_2}$$

so that, on average 1 event will occur within $\pm x$ seconds of the time a flip-flop is biased to the centre when

$$x = \frac{T_1 . T_2}{2N}$$

Under these conditions the settling time of the flip-flop is given by

$$t = \tau' \log_e \frac{2\tau' . N}{T_1 T_2} \cdot \frac{A_0 - 1}{A_0}$$

In a practical situation only approximately half of the potential failures will cause false decisions to be taken since following gates will have a switching level either consistently above or below that of the decision flip-flop and in half of the cases the final output will be on the same side as the initial output after t seconds.

The mean time between failures will therefore be given by

$$2N = \frac{A_0}{A_0 - 1} \cdot \frac{T_1 T_2}{\tau'} \exp\left(\frac{t}{\tau'}\right)$$

and if A_0 is large compared with 1:

$$\text{m.t.b.f.} \simeq \frac{T_1 T_2}{\tau'} \exp\left(\frac{t}{\tau'}\right)$$

where

$$\tau' \simeq \frac{\tau}{A_0}$$

and is a measure of the gain-bandwidth product of the flip-flop.

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A direct conversion v.h.f. receiver

SALEH R. AL-ARAJI, Ph.D.*

and

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Based on a paper presented at the IERE Conference on Radio Receivers and Associated Systems held in Swansea from 4th to 6th July 1972.

SUMMARY

A direct conversion radio receiver is described which is characterized by many fewer spurious responses than the conventional superheterodyne, and lends itself to integration in monolithic form.

The major design problem derives from the need for an accurate two-phase tunable local oscillator. A circuit is described which uses a servo-controlled phase shifting network, and maintains the two outputs within $\pm 2°$ of quadrature from 66 to 95 MHz.

Complete a.m. and s.s.b. receivers are described, and problems of a.g.c. discussed. The expected relative freedom from spurious responses was confirmed experimentally.

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1 Principles of Direct Conversion Receivers

A direct conversion receiver is one in which the incoming r.f. signal is heterodyned with a local oscillator at or near the nominal carrier frequency of the signal. The output from such a conversion process is subjected to low-pass filtering to extract components in the a.f. spectrum. These a.f. waveforms usually correspond to the original modulating signal (possibly distorted in various ways) but sometimes further processing is required before the modulation is reconstituted.

From another point of view, a direct conversion receiver is often regarded as equivalent to a superheterodyne in which the i.f. has been reduced to, or near, zero. Interest in receivers of this type, also described as homodynes[1] or, in slightly modified form as synchrodynes,[2] is of very long standing. This is probably because they apparently offer the possibility of relatively simple realization, without the need of costly i.f. filters, and certainly are free from many of the spurious responses of the superheterodyne.

To take the latter point first, a superhet. receiver has two principal responses separated in frequency by twice the i.f.

There are in addition a number of spurious responses, associated with interactions between the harmonics both of the local oscillator and received signals. In each case these responses are 'twinned', corresponding to the two senses in which the interacting signals may differ in frequency by an amount equal to the i.f. If the i.f. is reduced to zero the twin responses coalesce, thus reducing by half the number of discrete frequencies at which a spurious response can occur. In addition, i.f. breakthrough responses are eliminated.

So far as simplicity of realization is concerned, it is obviously easier to obtain a large stable gain at a.f. than at r.f. or the conventional intermediate frequencies. There may not be much to choose between the cost of low-pass and band-pass filters of comparable performance, although the advantage, however small, is likely to be to the former. If it is desired to build the receiver in the form of a monolithic integrated circuit, however, clearly a design which uses a.f. amplification and active low-pass filters is more feasible than the conventional superhet., at least in terms of present day i.c. technology, and will also result in an i.c. with fewer external connexions.

The advantages in terms of front-end design are also significant. Because there are fewer spurious responses, and in particular no image response, the specification of any pre-mixer filter is much eased, and is only required to restrict the r.f. bandwidth sufficiently to make inter- and cross-modulation problems in the r.f. stages tractable.

The use of a direct conversion configuration thus holds out the prospect of a radio receiver completely integrated onto a single monolithic chip, with the exception of a relatively simple r.f. filter and some precision resistors and capacitors associated with active low-pass filters.

The advantages of direct conversion are obvious, but a difficulty has held back the application of the principle. In the case of a.m. signals, or double sideband signals with carrier suppressed, the received signal may be represented as a voltage $e(t)$, where

$$e(t) = y(t) \cos(\omega t + \phi). \quad (1)$$

Here $y(t)$ is the modulating function, ω is the carrier frequency and ϕ an arbitrary phase angle. The local oscillator generates a voltage $e'(t)$, where

$$e'(t) = a \cos \omega t \text{ and } a \text{ is a constant.}$$

After mixing (assumed equivalent to multiplication) and low-pass filtering a signal is obtained given by $e''(t)$, where

$$e''(t) = a' y(t) \cos \phi. \quad (2)$$

The third term on the right-hand side is a considerable embarassment: it can take any value between zero and unity since ϕ depends both on the distance between transmitter and receiver and on the highly variable characteristics of the propagation medium.

The conventional solution to this problem, adopted in the receiver to be described in this paper, is to build two receiving channels with local oscillators in phase quadrature. The outputs from these two channels (assumed of identical gain) will then be $e_1''(t)$ and $e_2''(t)$, where

$$\left.\begin{aligned} e_1''(t) &= a' y(t) \cos \phi \\ \text{and} \quad e_2''(t) &= a' y(t) \sin \phi \end{aligned}\right\}. \quad (3)$$

Utilizing these two outputs it is possible to obtain $y(t)$ independent of ϕ either:

(a) by combining e_1'' and e_2'' in a suitable network to eliminate dependence on ϕ, or

(b) by utilizing the relative magnitudes of e_1'' and e_2'' to drive ϕ to a standard value, say zero, in which event e_1'' is the required output.

Receivers which operate on the principle indicated in (a) above include a class described by Barber,[3] which is essentially equivalent to a band-pass filter and will be referred to as Barber receivers in what follows. The alternative approach described in (b) above has been exploited in a class of receivers described by, among others, Costas.[4]

Both types of receiver require a two-phase local oscillator with the phases in quadrature. In addition, the Barber circuit requires converters and subsequent amplifiers having identical gain as between the two channels of the receiver. For a Costas receiver it is convenient if the gain of the two channels is similar, but they need not be identical. In discrete component technology, exact equality of gain in two converter/amplifier chains would be difficult to maintain, but in monolithic technology this problem is much less severe.

Common to both types of receiver, however, is the problem of constructing a satisfactory two-phase oscillator, giving two outputs of equal amplitude and in phase quadrature yet capable of being tuned over the desired frequency range. These problems become progressively more acute at higher frequencies, yet it is precisely in the v.h.f. and u.h.f. ranges that the advantages of second-channel response suppression become most marked.

In what follows the design of a servo-controlled v.h.f. phase shifter will be described which made possible the realization of both Barber and Costas receivers having an approximately 30% tuning range at about 80 MHz. Extension to higher frequencies should present no insuperable difficulties. Before discussing the design in detail, however, a review of the properties of the Barber circuit will be given, leading to the derivation of a suitable specification for the phase-shifter.

2 Properties of the Barber Circuit

A block diagram of a Barber receiver is shown in Fig. 1. The incoming signal $V(t)$ passes to two modulators driven by phase quadrature local oscillators at a frequency ω_0. After low-pass filtering, which removes the sum term and any local oscillator component from the mixer output, in each channel the signal passes to a second mixer, driven at a frequency ω_1. In the original filter, described by Barber, $\omega_1 = \omega_0$ but for use as a radio receiver it is more convenient to make ω_1 relatively

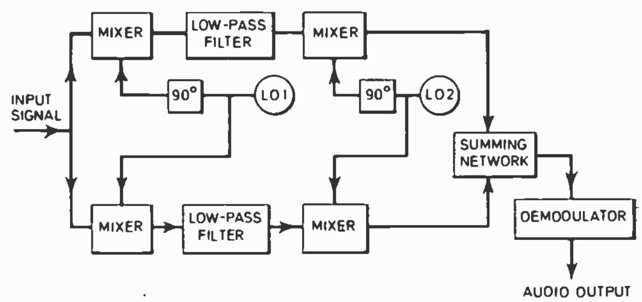

Fig. 1. The basic Barber receiver.

small. The complete mathematical analysis of the Barber receiver with $\omega_1 \neq \omega_0$ has been given elsewhere,[5] but for present purposes it is sufficient to note that if the input signal is a very general r.f. spectrum of the form

$$V(t) = \sum_{\text{all } r} a_r \cos(\omega_r t + \phi_r) \quad (4)$$

then the voltage $V'(t)$ appearing at the output terminal will be of the form

$$V'(t) = (A_1 + A_2)V_1(t) + (A_1 - A_2)V_2(t) \quad (5)$$

where A_1 and A_2 are the total gain in the first and second of the two channels and $V_1(t)$, $V_2(t)$ are given by

$$V_1(t) = \sum_{\text{all } r} a_r \cos[(\omega_r - \omega_0 + \omega_1)t + \phi_r] \quad (6)$$

$$V_2(t) = \sum_{\text{all } r} a_r \cos[(\omega_0 - \omega_r + \omega_1)t + \phi_r]. \quad (7)$$

The term in $V_1(t)$ in equation (5) represents the wanted receiver response; it has a spectrum exactly analogous to that of $V(t)$ but translated downward on the frequency axis by an amount $(\omega_0 - \omega_1)$ and subjected to band-pass filtering, the characteristic of which may be obtained from that of the low-pass filters used. If these have a characteristic voltage transfer ratio $F(\omega)$, then the band-pass filtering obtained can be shown to be equivalent to that given by $F(|\omega - \omega_0|)$, that is to say similar to the low-pass characteristic but with frequency zero translated to ω_0, and with a high-pass characteristic for frequencies less than ω_0 of identical shape to that of the low-pass characteristic for frequencies higher than ω_0. These results are represented in Fig. 2.

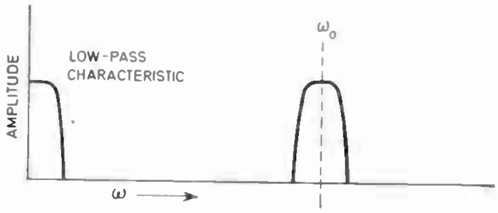

Fig. 2. Low-pass filter characteristic and the corresponding band-pass filter characteristic of the Barber receiver.

The term in equation (5) which corresponds to a spurious response is that in $V_2(t)$. This represents a signal similar to $V_1(t)$, but with its spectrum reversed, so that the highest frequency component becomes the lowest, and vice versa. The spurious response of this type of receiver is thus a reversed-spectrum version of the wanted signal, and may thus be regarded as a form of distortion. It does not result in interference from unwanted transmissions, as does the image response of a superhet. If A_1 is made identical to A_2 the spurious response is zero; practically the two amplitude coefficients may be adjusted to, say, 2% difference, giving a spurious response 40 dB down.

The above analysis assumes that the two phases of the local oscillators for both front and back end pairs of mixers are exactly in quadrature. Although this can be very nearly achieved in the latter case, the two-phase oscillator at v.h.f. may suffer from some phase error. Figure 3 shows the relationship[5] between such phase error and the degree of suppression of the spurious response. In practice an error of one or two degrees is permissible.

It will be seen that the Barber circuit acts as a band-pass filter and frequency convertor. The output $V'(t)$ must then be demodulated, and this can be accomplished in a conventional way. For example, in the case of a.m. transmissions an envelope or coherent detector would be employed, whilst for f.m. the second oscillator frequency ω_1 can be chosen to be low enough for a pulse-counting type of discriminator to be highly effective. In the case of s.s.b. transmissions, all that is needed is for the centre of the incoming spectrum to be translated to the middle of the a.f. band: demodulation in the conventional sense is not required. Used in this way, the Barber circuit is identical with the s.s.b. demodulator first proposed by Weaver.[6]

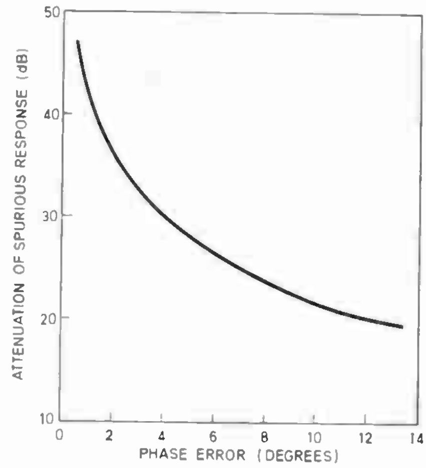

Fig. 3. Relationship between spurious response and total phase error of the quadrature oscillators.

3 The Phase Quadrature Oscillators

The two bi-phase oscillators are the elements of a Barber receiver which cause greatest difficulty in realization, and particularly so where operation in the v.h.f. region is required. There are a number of ways in which a 90° phase shift, constant over an extended range of frequencies, can be obtained. For example, passive phase shifting networks are widely used in s.s.b. practice. The design of such networks, giving a reasonably constant 90° shift over as much as a 10 : 1 frequency range, is well understood.[7] Unfortunately, at usable impedance levels the values of capacitor required to realize such networks at frequencies greater than about 30 MHz, are intractably small.

An alternative approach uses digital counters or shift registers to yield two square waveforms, one with a quarter period delay.[8] In this case, however, the digital circuits must operate from a clock frequency at least twice that of the bi-phase waveforms. For signal frequencies of, say, 100 MHz this digital approach is consequently unattractive.

The techniques finally adopted involves the use of a two section CR phase lag network, in which the capacitors are varactors, allowing the phase shift of the CR network to be electrically adjusted to be exactly 90° greater than that of a 'dummy' network consisting only of resistors.

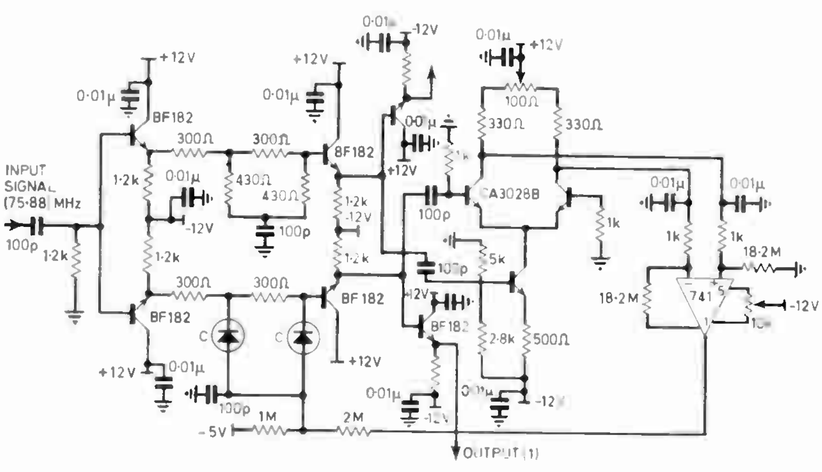

Fig. 4. A servo-controlled phase delay network producing equal amplitude quadrature outputs in the frequency range 66–100 MHz.

The dummy network values are then chosen to give exactly the same voltage transfer ratio as that of the CR network when the latter has the required phase shift. The circuit is shown in Fig. 4. The dummy network is at the top, preceeded and followed by common emitter buffer amplifiers, whilst a corresponding arrangement embodying the varactor CR networks appears below.

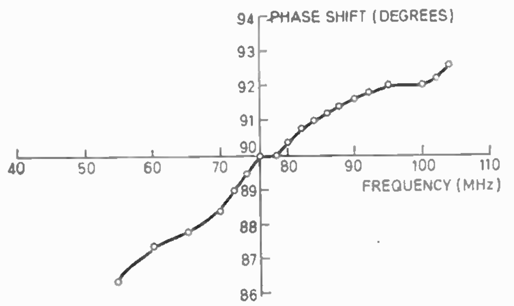

Fig. 5. Phase shift as a function of frequency for the SCPDN of Fig. 4.

The output of two channels is compared by means of a phase-sensitive rectifier of the long-tail pair type.[9] This produces a d.c. output proportional (for small angles) to the deviation of the phase relationship from true quadrature. The output from the p.s.r. is amplified by an operational amplifier and fed back to the varactor network to achieve the desired control of phase angle. Figure 5 shows the variation of phase angle measured with frequency varying between 70 and 100 MHz, from which it will be seen that acceptable angular errors were obtained over the operating range of 66–95 MHz.

The second local oscillator is operated at a frequency of only 30 kHz, hence in this case a wide variety of methods were available for deriving the required two-phase supply. That adopted was the use of a conventional two-phase sinusoidal CR oscillator at the required frequency.

4 The Complete Receiver

The complete receiver circuit, as used for a.m. reception, is shown in Fig. 6. The low-pass filters used were fourth-order Sallen and Key[10] types, having a characteristic as shown in Fig. 7. The selectivity curve of the receiver is shown in Fig. 8. Slight d.c. drift in both receiver channels and inadequate carrier suppression in the back-end mixers results in a standing carrier output in the absence of signal. The level of this output is indicated on the diagram. It is, of course, unmodulated but would interfere with the correct operation of a.g.c. circuits if the signal level were to fall sufficiently, and thus sets a limit to the dynamic range of the receiver. This is one of the principal areas in which further improvement of present levels of performance is to be sought.

The problem of a.g.c. in a receiver of this type merits further consideration. Application of a.g.c. to the amplifiers or mixers within the two channels of the Barber circuit itself is unpromising, since it is difficult to get exactly matched gain variation. Thus a.g.c. would normally be applied to the r.f. amplifier preceding the first mixers, or alternatively to an i.f. amplifier following the second mixers, or both. Arrangements for a.g.c. are not shown in Fig. 6.

Since the Barber receiver uses two local oscillators, the question of possible interaction (as commonly encountered in double superheterodyne configurations) arises. In the design here reported, the first oscillator was at v.h.f. and the second at 30 kHz, and no interaction effects of any kind were observed, provided that modest screening of the v.h.f. front end was employed. This is a distinct advantage of the Barber receiver over the double superheterodyne. The receiver as constructed had no observable spurious responses other than those at harmonics of the signal frequency. These are relatively easily suppressed by the r.f. filter.

For s.s.b. reception the cut-off frequency of the low-pass filters was reduced to 1·5 kHz and the second

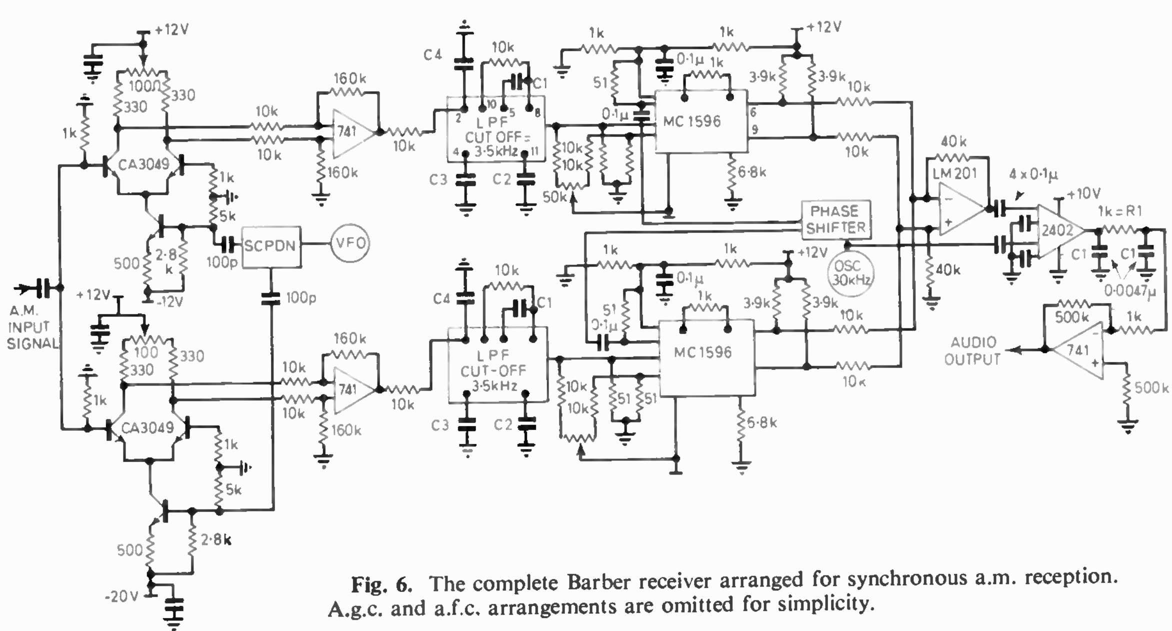

Fig. 6. The complete Barber receiver arranged for synchronous a.m. reception. A.g.c. and a.f.c. arrangements are omitted for simplicity.

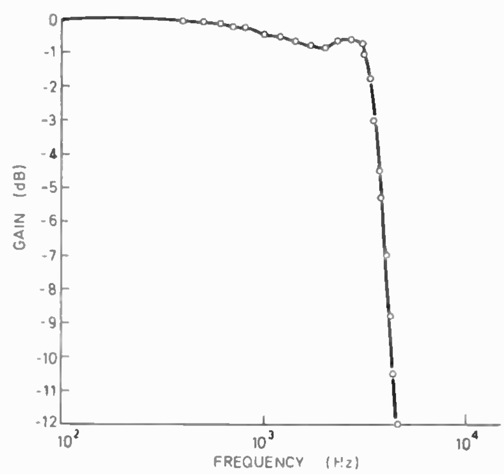

Fig. 7. Characteristic of 4th order low-pass filter used in Fig. 6.

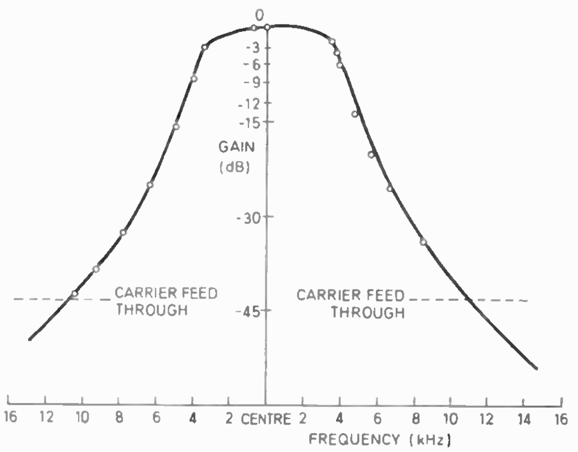

Fig. 8. Receiver selectivity curve. Centre frequency in the range 66–95 MHz.

mixer frequency to 1·7 kHz.[11] Thus by tuning the first mixer to a point at the centre of the received s.s.b. spectrum, a frequency translation to the range 200 Hz–3·2 kHz, is achieved. In this case it was found possible to a.c.-couple both channels of the receiver, consequently eliminating problems of spurious output from the back-end mixers due to d.c. drift in the preceeding channels. The a.c.-coupling obviously resulted in a band-stop characteristic centred at the middle of the audio spectrum. In one series of experiments the high-pass break frequency for both channels was arranged to be 250 Hz, giving an audio stop band centred on 1·7 kHz and 500 Hz wide between 3 dB points. Perhaps surprisingly, the effect of this stop band on the intelligibility of received speech transmissions was slight.

5 Discussion and Conclusions

The receiver described, although not yet developed to the limit of its potential, demonstrates clearly the use of the Barber circuit as a selective frequency converter. Image responses are avoided and adjacent channel selectivity is provided by the use of audio-frequency low-pass filters only. If radio receivers capable of complete integration on a single monolithic circuit are to be produced, the Barber configuration will be one approach which merits consideration.

One disadvantage of the present circuit is that the received carrier appears as a d.c. component after the first conversion process. For this reason the a.m. receiver has been d.c.-coupled to retain the carrier component after the second mixing process. The d.c. drift in the first mixers and subsequent amplifiers will, however, alter the magnitude and phase of the reconstituted carrier, so that signals below about 1 mV at the input terminal are not correctly received. A solution, to which reference has already been made, is the use of a.c.-coupling in both channels of the receiver. An s.s.b. receiver adopting this approach[11] successfully overcomes the d.c. drift problem, but at the cost of a 'null' in the middle of the pass-band, which is perceived as an audio stop band in the receiver output. In the case of an a.m. receiver, synchronously tuned, the effect of the stop-band would be to suppress the carrier, which is undesirable. A remedy is to incorporate an offset into the receiver a.f.c. system, so that the response 'null' falls between the carrier frequency and the lowest modulation sideband. In the case of speech transmissions the lowest sidebands of interest are 300 Hz from the carrier frequency, and thus a stop-band less than 300 Hz wide can be incorporated without loss of sidebands or carrier provided that the a.f.c. is offset by 150 Hz. A further description of this technique will be published shortly.

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Manuscript first received by the Institution on 29th March 1972 and in final form on 16th May 1973. (Paper No. 1534/Com 72.)

IERE News and Commentary

Institution Premiums and Awards

The Council of the Institution announces that authors of the following papers are to receive Premiums and Awards for papers published in *The Radio and Electronic Engineer* during 1972.

Clerk Maxwell Premium (*Value £30*)
(The most outstanding paper of the year)

'The Principles of Pulse Signal Recovery from Gravitational Antennas'
by Dr. M. J. Buckingham and Prof. E. A. Faulkner (University of Reading).
(Published in April).

J. Langham Thompson Premium (*Value £50*)
(Control Engineering)

'Predicting Servomechanism Dynamic Errors from Frequency Response Measurements'
by Mrs. M. J. Brown, Prof. D. R. Towill and Dr. P. A. Payne (UWIST Cardiff).
(January).

The Dr. Vladimir K. Zworykin Premium (*Value £50*)
(Medical and Biological Electronics)

'An Automatic Biochemical Analyser'
by R. Wyld (formerly with Vickers Medical Engineering).
(September).

The Heinrich Hertz Premium (*Value £20*)
(Mathematical or physical aspects of radio)

'Point-matched Solutions for Propagating Modes on Arbitrarily-shaped Dielectric Rods'
by Dr. J. R. James and I. N. L. Gallett (Royal Military College of Science)
and
'Engineering Approach to the Design of Tapered Dielectric-rod and Horn Antennas'
by Dr. J. R. James.
(June).

The A. F. Bulgin Premium (*Value £15*)
(Measurements)

'Absolute Measurement of Submillimetre and Far Infra-red Laser Frequencies'
by Dr. C. C. Bradley, Dr. G. Edwards and Dr. D. J. Knight (National Physical Laboratory).
(July).

The Dr. Norman Partridge Memorial Premium (*Value £10*)
(Audio frequency engineering)

'Statistical Stability in Spectrum Analysis'
by R. E. Bogner (Imperial College, London).
(September).

The Leslie McMichael Award (*Value £10*)
(Radio communication)

'An Experimental Adaptively Equalized Modem for Data Transmission over the Switched Telephone Network'
by R. J. Westcott (Post Office Research Department).
(November).

The Lord Brabazon Award (*Value £15*)
(Radar and navigational aids)

'A 16-Channel Digital Acoustic Telemetry System'
by D. Cattanach (Marine Laboratory, D.A.F.S., Aberdeen).
(March).

The Lord Rutherford Award (*Value £15*)
(Electronics associated with atomic physics)

'Ion Implantation in Semiconductor Device Technology'
by Dr. J. Stephen (A.E.R.E., Harwell).
(June).

The Marconi Award (*Value £10*)
(Engineering)

'Annular Resonant Structures and their uses as Microwave Filters'
by R. T. Irish (Royal Military College of Science).
(February).

The Charles Babbage Award (*Value £15*)
(Electronic Computers)

'Magnetic Bubbles and their Applications'
by R. D. Lock and Dr. J. M. Lucas (Bell Canada Northern Electric Research, Ottawa).
(October).

The Hugh Brennan Premium (*Value £15*)
(First read before a Local Section)

'A Special-purpose Computer for the Direct Digital Control of Processes'
by A. J. Allen and P. Atkinson (University of Reading).
(February).

The award of the following Premiums has been withheld as papers of suitable standard were not published during the year: Rediffusion Television Premium; P. Perring Thoms Premium; Arthur Gay Premium; Sir J. C. Bose Premium.

The authors will receive their Premiums and Awards from the President of the Institution at the Annual General Meeting to be held in London on Thursday, 25th October 1973.

Chief Officer of Technician Education Council

The Technician Education Council, which was established by the Secretary of State for Education and Science in March 1973 to develop a national pattern of schemes of technical education and examinations for persons at all levels of technician occupations in industry and elsewhere, has appointed Mr. Francis G. Hanrott as its Chief Officer. Mr. Hanrott will take up his appointment on 1st September 1973.

Mr. Hanrott has been the Registrar and Secretary of the Council for National Academic Awards since 1966; he was some years ago Staff Manager at the GEC Applied Electronics Laboratories.

ICE Luncheon Club

Membership of the Luncheon Club at the Institution of Civil Engineers, Great George Street, Westminster, S.W.1, has for some time been open to members of other CEI Institutions on personal application and subject to an annual subscription. The Council of the ICE has now approved the temporary suspension of the annual subscription and an amendment to the club rules so as not to require the issue of membership cards to members of sister Institutions.

Members of the IERE are now welcome to make full use of the facilities of the club and the only formality is that they may very occasionally be asked, as may Civil Engineers, to establish their identity and link with the IERE.

The Restaurant, Cafeteria, ICE Box Bar and Buttery and the Cocktail Bar at Great George Street are open daily from 12–2.30 (except at holiday periods). Members may bring guests (including ladies and children). It is advisable to book tables in the Restaurant—telephone 930 8373.

Engineering in North America

Proposals for Pensions for US Engineers

In 1972 for the first time, the US Congress paid serious attention to the pension problems of engineers. No legislation could be signed into law, but it is believed that the next Congress in 1973–74 may produce real results.

The incredibly high rate at which employees in general, and engineers in particular, move from employer to employer, and from pension plan to pension plan, is a worldwide problem. They are always being 'covered' or 'participating' but rarely vesting, and instead forfeit their pension rights time after time as they change jobs.

A bill guaranteeing against forfeitures after an employee worked eight years for the same employer was introduced in the Senate, but a favourable report by the Senate Labour Committee came too late for a full vote by the Senate.

But even eight-year vesting, although a tremendous step forward, is probably inadequate for most engineers, according to Mr. Frank Cummings, special Washington Counsel for IEEE, because many engineers and scientists regularly change jobs in less than that period of time.

Another bill dealing with problems of technological reconversion became the vehicle for a more specific attempt to deal with the pension problems of engineers and scientists. It included an amendment designed to protect the pension benefits of engineers working on Federal contracts. This bill passed the Senate but was too late for serious consideration in the House.

Civil Anti-trust Suit against US Engineers

The National Society of Professional Engineers began the New Year with a civil anti-trust suit filed against it by the US Justice Department. It charged the NSPE with eliminating price competition among members in the sale of engineering services.

NSPE is a non-profit membership corporation made up of approximately 67 000 professional engineers. The Anti-trust Division of the Justice Department has charged that the NSPE and members have for many years combined to violate Section I of the Sherman Act by agreeing that NSPE adopt, publish and distribute a Code of Ethics containing a provision prohibiting its members from submitting competitive bids for engineering services and that NSPE and its members abide by such a provision and police adherence to its terms.

The Justice Department claims these practices have eliminated price competition in the sale of engineering services and deprived customers of the benefits of free and open competition.

The complaint asked the court to cancel those provisions of the NSPE Code of Ethics that suppress competition among members. It also requested that NSPE be permanently enjoined from adopting in the future any rules or resolutions that suppress or eliminate such price competition.

New Canadian Immigration Regulations

The Federal Department of Manpower and Immigration in Ottawa has recently adopted new regulations involving a number of major changes to previous policy. Many of these rules will affect members of the engineering profession, and information interpreting the regulations has been received from the Canadian Council of Professional Engineers.

Engineers entering Canada as visitors will no longer be permitted to apply for or to gain landed immigrant status while in Canada, and they will not be permitted to work without a 'work visa'. Work visas are issued to engineers only when the Local Manpower Centre is satisfied that Canadian engineers are not available or willing to take up the employment. In the case of professional workers, the local Manpower Centre will examine the local demand records in its area and in surrounding areas, will discuss the case with the regional office to determine whether or not this demand should be checked on a regional and/or national basis, and will usually discuss the case with the appropriate professional association.

Work visas are normally requested by a prospective employer. If the Department is satisfied that the issuing of a work visa for a prospective employee is justified, it will be issued to the employer who, in turn, passes it to the prospective employee. Work visas are issued for specific periods of time which vary according to the type of work and other factors but may be renewed under the same conditions as originally granted. These rules also apply to staff members of foreign consulting firms and to employees of parent companies having subsidiaries in Canada.

Similarly, students entering Canada to study engineering will not be permitted to apply for or to obtain landed immigrant status during their stay in Canada, and will not be permitted to work in Canada without a 'work visa' during their stay as students. Special considerations will however be given to applications for work visas from students who are required to perform work as part of their educational process.

London Electronic Component Show 1973

Although this year's Electronic Component Show, held at Olympia from 22nd to 25th May, may not have been the largest in the series which started 40 years ago, its technical significance lies in indicating which of the developments of the research laboratories are finding commercial applications. The various processes for making integrated circuits, and new techniques in constructional methods were two areas of especial prominence and importance. These and a number of other points of interest are briefly described under the 'New Products' heading on pp (ix)–(x) of this issue.

Comments on the importance of the Exhibition and on the way in which the industry is going were made at the opening luncheon by H.R.H. the Duke of Kent who is a member of the National Electronics Council. He said:

'This Exhibition, whose character is truly international with one-fifth of the exhibitors being from overseas, is a particularly important one for the British electronic components industry —for two reasons. It is the first such show to be held since Britain became a member of the Common Market, and it falls at a time when the Electronics Industry as a whole is climbing out of a recession which was virtually world-wide. In the space of only a few months construction appears to have given way to expansion and in the present atmosphere of high demand and re-investment, the electronics industry, with its tentacles reaching out into so many fields—consumer products, avionics, communications, military equipment, industrial automation and many others—should stand to gain immense advantages.

'All this equipment, however—and its range and diversity are increasing almost daily—depends upon the quality of its component parts and in the end can only be as good as the hundreds and thousands of bits which go into it. These in themselves are becoming more complex, and hand in hand with the drive towards technical perfection and the trend to greater use of integrated circuits goes the essential increase in reliability. The degree of reliability needs to be staggeringly high—perhaps as great as one chance in a million of failure (of an individual item) if modern complex and vital pieces of equipment are to be acceptable.

'In this immensely important field of quality assurance and reliability the British electronics industry has set, and continues to set an enviable example and we are, I believe, fortunate in this country in having effective agencies such as the British Standards Institution and the Ministry of Defence's Electrical Quality Assurance Directorate concerned in the maintenance of these standards and also in promoting, as a longer-term objective, the adoption of a European standards system.

'I spoke just now of the drive towards technological improvement and in the electronics field this is both relentless and in some cases has already produced fairly spectacular results. To take an example, quite a small computer processor unit that only ten years ago weighed 60 pounds and cost about £30 000 can now be produced on a single chip costing £500. Such micro-computers could have a startling effect on many mundane and quite routine aspects of our lives. Not only could they as a matter of course control automatic transmissions, fuel injection and possibly navigation systems on road vehicles, but I believe it is not fanciful to suggest that one of these astonishing devices could manage all the systems of one's house—the heating, cooking, lighting, waste disposal and so on.

'The importance of this particular area of electronic development—integrated circuits—need hardly be emphasized in this company, but it lies, I believe, chiefly in the very widespread repercussions that such devices could have upon other industries and in other fields. Clear proof that this is now recognized has been given by the Department of Trade and Industry's welcome decision to undertake a programme of support for the micro-circuit industry over the next few years, support which can be seen as an investment in a vital area of potential growth. One interesting thought that does emerge from this is that the component designer of yesterday is rapidly evolving into today's system designer—and equally the rate of development is such that obsolescence can overtake the most advanced equipment alarmingly quickly. But I doubt if the industry need worry about that unduly just yet as I am certain it would be wrong to underestimate the continuing importance of the more conventional discrete components—rather less glamorous though these may be. Their variety and proliferation is enormous—there are after all more than 400 manufacturers exhibiting at this Show—and there seems to be every indication that demand will increase and no doubt specifications become even more rigorous.

'So the electronics industry, its engineers and scientists, continues to push at the boundaries of advanced technology and in spite of our limited money and resources it is encouraging—although perhaps not altogether surprising for a country that did after all give the world radar and television—that British companies continue to occupy a place in the forefront of this fast-moving development. They have not achieved this position by accident nor should anyone imagine it can be maintained except by a great deal of hard work and by the exercise of the utmost brainpower and skill in these areas where we have proved that we can excel.

'Forty years ago the British components industry comprised only some forty firms employing 5000 people and with a turnover of about £6M. Now in only four decades there are at least 200 firms in the components sector. Between them they employ 140 000 and have a turnover in excess of £400M. So this huge and vital industry has grown to its present size in a very short space of time and we salute the Radio and Electronic Components Manufacturers' Federation, on their fortieth anniversary.

'In 1933 it would have taken an extraordinarily good crystal-gazer to foresee the kind of miracles that we take for granted today. As to what could come about by the year 2013 it is probably only safe to predict that electronics will certainly have come to play a much more dominant role in our lives than it does now.

'You in the industry have a great deal to be proud of in your past achievements and I feel sure you must be contemplating the almost limitless opportunities which the future holds with keen anticipation. I wish all exhibitors and their customers every success at this London Electronic Component Show.'

Members' Appointments

CORPORATE MEMBERS

Mr. E. R. Friedlaender (Fellow 1944), who has been working as a free-lance technical journalist for the past 10 years, has retired from full-time work on reaching the age of 65. Mr. Friedlaender qualified as an Engineer in Berlin in 1928 and came to England in 1936. He was for many years concerned with magnetic dust core manufacture with Neosid Ltd. of Welwyn Garden City and with companies in the Manchester area. During this period, he was a member of the North Western Section Committee. In 1955 he moved to London as an Industrial Consultant, specializing on exploitation of patents. Mr. Friedlaender has on numerous occasions assisted Institution Committees and the Secretariat on technical translation problems.

Mr. Ronald Blakeley (Member 1972, Graduate 1962) is now a Lecturer in Electrical Engineering at Wilmorton College of Further Education, Derby. He returned to this country at the end of last year after working for Williamson Diamonds Ltd. in Tanzania as an Electronic Engineer since 1968.

Mr. A. G. Brown (Member 1967) was appointed Vice-Principal of Riversdale College of Technology, Liverpool, from 1st January 1973. He was previously Head of the Department of Electronics and Radio Engineering at the College.

Lt. Col. P. R. Gray, B.Sc., RA (Member 1967) has been appointed GSO 1(W) at the Royal Radar Establishment, Malvern. He previously held a similar Staff appointment at the Ministry of Defence.

Sqdn. Ldr. K. P. Hardman, B.Sc., (Member 1971) will take up the appointment of Head of the Department of General Studies at Riversdale College of Technology, Liverpool, on 1st September 1973. He is currently a Senior Education Officer at RAF Cosford.

Mr. A. C. Hayward, B.Eng. (Member 1971), who is currently Senior Lecturer in Electrical and Electronic Engineering at the Charles Trevelyan College, Newcastle-on-Tyne, moves to Riversdale College of Technology, Liverpool, as Head of the Department of Electronics and Radio Engineering on 1st September next.

Mr. K. C. Hills (Member 1973, Graduate 1970), who was formerly a Circuit Design Engineer with Kelvin Hughes, has recently been appointed a Lecturer in the Electrical Engineering Department of West Ham College of Technology.

Mr. C. F. H. Teed (Member 1971), formerly Manager of the Broadcasting Division of Marconi Communications Systems Ltd., has recently been appointed Marketing Director of the Company. Mr. Teed who joined the Marconi Company in 1953, returned to it in 1972 following four years as Chief Engineer of Independent Television News Ltd.

Mr. John D. Tucker (Member 1958) has resigned his position as Director of Operations, EMI Television Division, EMI Sound & Vision Equipment Ltd., to join Gresham Management & Investment Company Limited as Technical and Marketing Director. After service with the BBC, Mullard Equipment Ltd., and Associated Rediffusion Ltd., Mr. Tucker joined EMI in 1958 as Senior Sales Engineer of the Broadcast Equipment Division of EMI Electronics Ltd. He subsequently became Sales Manager, General Sales Manager and then Director of Operations.

Mr. Tucker is chairman of the Management Committee for the IBC and is one of EEA's representatives.

NON-CORPORATE MEMBERS

Mr. R. L. DeVille (Associate 1972), who was from 1964 to 1972 with Molins Ltd., latterly as an Electronic Designer, has recently joined Rank Xerox Ltd. as a Product Engineer.

Mr. H. Farthing (Graduate 1962), who was formerly with Standard Telephones & Cables Ltd., Newport, Monmouthshire, is now resident in Canada and employed by the Shawinigan Engineering Co. Ltd., a Montreal firm of Consulting Engineers.

Mr. A. C. F. Leadbitter (Associate 1962), is now General Manager of the Television Equipment Division of Belling & Lee Ltd. Two years ago he was appointed Divisional Manager of the Company's Interconnection Systems Division.

Mr. H. S. Manku (Graduate 1969, Student 1965) has joined Business Computers Ltd. of Brighton as a Pre-Production Engineer; he was formerly with Redifon Ltd., Crawley.

Mr. M. D. McPherson (Graduate 1970), who was for a number of years with British European Airways as a Communications Equipment Engineer concerned with software developments, is now working in Denmark for SPL, a software bureau associated with the PNR project of Scandinavian Airlines System.

Capt. K. G. Melton, REME (Graduate 1970) has been posted to the REME Wing of the Royal Artillery Range in the Hebrides as Officer in Charge of the Instrumentation Platoon. He was previously Training Officer at 8 Field Workshop (Airportable), REME, Colchester.

Mr. B. K. Price (Graduate 1970) has been promoted to Lecturer II at Riversdale College of Technology which he joined in 1971 following an appointment as a Technical Officer at Daresbury Nuclear Physics Laboratory.

Obituary

The Institution has learned with regret of the deaths of the following members.

Wing Commander George Ellams, O.B.E., A.F.C., RAF (Retd.) (Member 1958, Graduate 1952) died on 12th April 1973 at the age of 51 years. He leaves a widow.

Born in Liverpool and educated in Wallasey, Wing Cdr. Ellams joined the RAF as a Boy Entrant in 1938 and was commissioned in 1942. He was employed in flying and instructional duties as a Signals Leader from then until 1946. After a year as Signals Examiner for Aircrew he was transferred to the Technical Branch in 1947 and appointed Flying Wing Signals Officer, Singapore, and was personal radio operator to the C-in-C, FEAF. His next tour was as Signals and Radar Officer at Manston and subsequent appointments were as Senior Signals Officer at Shawbury, Aden, Topcliffe and Horsham St. Faith. In 1962 he returned to Singapore in the Headquarters Unit and in 1965 was appointed to RAF Cosford as Senior Training Officer. He was appointed an O.B.E. in 1966. Wing Cdr. Ellams retired from the RAF in 1968 but was subsequently re-appointed to Headquarters, Training Command, RAF Brampton, on ground training, where he was serving at the time of his sudden death.

William Francis Evans (Member 1954) died on 25th March 1973, aged 57 years. He leaves a widow and son.

Mr. Evans was born in Tavistock where he attended the Grammar School. After receiving training with Baird Television at Crystal Palace, he joined the Post Office Radio Branch, Dollis Hill, in 1937 as a Skilled Workman, subsequently progressing to Executive Engineer, the appointment he held at the time of his death. In the 1950s he worked on the development and commissioning of the first coaxial cable television link between London and Birmingham and on a number of other development projects on h.f. measuring sets and s.h.f. equipment. During the latter part of his career he was engaged on radio and line network planning. A man of diverse interests, Mr. Evans was particularly knowledgeable in the art of dowsing.

Bryan Frank Tindall, B.Sc. (Associate 1957), who died on 15th April 1973, was 49 years of age. He leaves a widow and two daughters.

Mr. Tindall was born at Ventnor and educated at Truro School, Cornwall. He served in the RAF from 1942–46 and was commissioned in the Signals Branch. In 1946 he joined the Air Ministry as a Scientific Assistant at the RAF Institute of Aviation Medicine at Farnborough. In 1954 Mr. Tindall obtained the B.Sc. (General) external degree of London University. He was promoted to Experimental Officer in charge of the Electronics Laboratory in 1954 and to Senior Experimental Officer in 1970. His more recent work was concerned with the instrumentation associated with investigations on vestibular physiology.

Colloquium on Recent Developments in Systems Performance Measurement

University College, London
10th April 1973

This one-day colloquium was organized by the Automation and Control Systems Group Committee of the IERE (Chairman: Professor D. R. Towill), in co-operation with the Ministry of Defence, Procurement Executive (Mr. A. R. Mellors).

The colloquium followed a highly successful meeting held at SRDE, Christchurch, in 1972 in which system designers were introduced to new techniques for system testing. At the London meeting, the emphasis was on the practical application of these techniques, and there was an opportunity for delegates to see an exhibition of equipment.

Nearly 200 engineers drawn from industry, universities and Government departments attended the meeting. The chairman for the morning session was Professor K. A. Hayes, O.B.E., formerly head of the Department of Electrical Science, and the afternoon session was chaired by Mr. I. J. Rhodes of the SRDE.

Session 1. Generalization of Correlation Techniques

Noise rejection properties of correlation measurements of dynamic response in control engineering

By Dr. J. D. Lamb and Dr. P. A. Payne (*UWIST, Cardiff*)

Methods of dynamic measurement in use at the current time are reviewed and the bases of their mechanization unified through an analysis of the cross-correlation-function. Experimental verification of the noise rejection properties of the cross-correlation technique in both the time and frequency domains is provided and a method of employing a single cross-correlator for measurements in both domains is examined.

Session 2. Time Domain Testing

The application of pseudo-random binary sequences to a gun-mounting system

By Cmdr. D. J. Kenner (*formerly Royal Navy; now Vice-Principal, Redbridge Technical College, Ilford, Essex*)

This paper describes some of the special problems encountered when using pseudo-random sequences to evaluate the dynamic behaviour of a gun-mounting servo-mechanism. Precautions taken to reduce the effects on non-linearities and the consequent effect on p.r.b.s. characteristics are indicated.

A special purpose digital computer for on-line determination of system impulse response

By H. Banasiewicz, Dr. S. E. Williamson and Professor W. F. Lovering (*University of Surrey, Guildford*)

The paper describes a low-priced computer for on-line computation of the impulse response of a system. A p.r.b.s. test sequence generated by the computer is used to excite the system under test and the response of the system to the sequence is logged in the computer store. The impulse response is computed by correlation techniques, stored as 63 digital numbers, and may be displayed repetitively on a c.r.t. display which is integral with the computer, or may be read out of store as a series of ordinates. Clock frequency is adjustable to suit a wide range of system time constants (from 0·1 seconds to 5 seconds). The paper also shows how modified p.r.b.s. sequences may be used to reduce the effects of backlash or stiction on the computed response.

Session 3. Fault Diagnosis

Fault diagnosis—a pragmatic approach

By R. D. Woodward and E. W. Carr (*Honeywell Limited, Hemel Hempstead, Herts.*)

It is suggested that an analysis of the symptoms which would be displayed by a control system as a result of the typical failure of each replaceable component can give an adequate probability of correct diagnosis, which can be updated by experience. Probable fault location, correlated with syndrome recognition may then become a main feature of the computer program for a diagnostic automatic test system. Such probability methods require an efficient data processing back-up, but this is needed equally for spares provisioning and should therefore be available.

Fault diagnosis using time domain measurements*

By Dr. H. Sriyananda (*University of Sri Lanka*), and Professor D. R. Towill (*UWIST, Cardiff*)

A voting technique is used to diagnose fault conditions down to component level in a feedback control system using only the input–output cross-correlation function measured at suitable time delays. These time delays are chosen using a new formula based on Bayes's theorem, which ranks the time delays in order of usefulness in fault diagnosis; it can be readily applied at the design stage, thus assisting the integration of system design and test functions. The fault conditions necessary to set up the scheme may be obtained by direct fault generation in an actual system, or by simulation of the system mathematical model. A learning approach to the design of a fault diagnosis scheme is described which makes full use of any available failure data, together with the ranking formula for time delay selection, in the creation of an optimum scheme.

Session 4. Frequency Domain Testing

Practical problems of digital and analogue measurement of frequency response

By P. D. Simmons, R. Kennedy and S. Urbanski (*SE Laboratories, Ltd., Feltham, Middlesex*)

The general requirements for both digital and analogue approaches to instrumentation for accurate measurement of frequency response characteristics are reviewed and a new basis for analogue transfer function analysers, the homodyne detector, is described. Results of using the new analogue instrument in the presence of noise and harmonic disturbances are reported and it is shown to offer good rejection properties despite the economic design employed. Some further results of employing a more sophisticated instrument for establishing harmonic content of signals from non-linear devices are discussed and the advantages of this form of measurement are set out.

* To be published in *The Radio and Electronic Engineer*.

The application of a commutated filter to the design of a frequency response analyser†

By C. J. PAULL (*University of Nottingham*) and W. A. EVANS (*University College of Swansea*)

A new structure is discussed for a frequency response analyser incorporating an *N*-path filter to execute the correlation process. Properties of the filter, not considered previously, are shown to justify further analysis. The instrument in question covers the frequency range 10 Hz to 100 kHz and incorporates other novel features including a new piecewise linear waveform synthesis technique. Advantages include higher frequency operation, increased dynamic range, and a voltage programmable sweep capability over a frequency range in excess of 1000 : 1.

Two-channel frequency response analysis by special frequency test methods

By Dr. A. M. FUCHS (*Bafco Inc., Pennsylvania, USA*)

In many instances the desired test result is a plot of amplitude ratio (dB) and phase shift vs. log frequency over a range of frequencies which define the control system. Methods of performing frequency response tests are reviewed for the possibility of direct plotting of results, while at the same time retain high noise and harmonic rejection. It is shown that Fourier Integral Analysis is particularly well suited to the goal. Further, two-channel mechanization of Fourier Integral Analysis is readily accomplished with the consequent plot of open loop data from closed loop tests. The effect of noise and harmonics on the plot is analysed and examples of simulated testing situations presented. A comparison is made with point-by-point frequency response measurement methods. Limitations of sweep speed on the accuracy of the results are discussed. Finally, this frequency response measurement procedure is equipped with the use of power and cross-power spectral analysis by F.F.T. methods.

Session 5. Test Automation

Servomechanism testing in air defence systems in the field

By R. PASSMORE (*Radar Branch, Technical Group, REME, RRE, Malvern, Worcs.*)

For over thirty years servomechanisms have been used in a wide range of Army equipments, from searchlights to guided weapons. The method of checking such equipments and consequently the test equipment employed are determined, in the main, by the following factors: (i) repair philosophy applied to the equipment, (ii) role of the equipment, (iii) technician skills available, (iv) repair philosophy applied to the test equipment. Tests and test equipments should be limited to those which will enable the technician to decide if the equipment is operating within its design parameters and if not diagnose the fault to the lowest item he is permitted to replace. An outline of the methods of testing a servo loop in the *Thunderbird II* Missile and the Optical Servo Group of *Rapier* System are given as examples.

† Published in *The Radio and Electronic Engineer* **43**, No. 6, pp. 369-78, June, 1973.

The desirable characteristics of a frequency response analyser for use in a practical measurement environment and in automatic test equipment

By A. J. MARTIN (*Solartron Electronic Group Ltd., Farnborough, Hants.*)

Conventional instrumentation such as voltmeters, counters and oscilloscopes are well understood and well documented in terms of how the specification and characteristics affect the accuracy and ease of measurement. The purpose of this paper is to attempt to lift frequency response analysers into this category. Special attention will be paid to the measurement environment and how the frequency response analyser: handles the 'real life' signals that occur in system test; is designed to provide minimum interference with the system under test; and provides access to the system under test through the system test points. In addition to a thorough review of desirable instrument features, the paper will show how changes in programming methods enables straightforward interfacing to digital computers to be performed, and software requirements simplified.

The software/hardware trade-off in automatic dynamic response testing

By J. B. IZATT and A. J. LEY (*Solartron Electronic Group Ltd., Farnborough, Hants.*)

Automatic dynamic response test systems usually consist of a general-purpose computer with varying quantities of special-purpose instrumentation. There is considerable choice in the degree in which functions may be carried out in software or hardware. Based on the particular characteristics of dynamic response testing an analysis is made of two systems, one where the real time calculations are made by a computer, the other by a computer-controlled dynamic response test instrument. System performance limitations caused by software problems are reviewed, and the factors affecting cost are discussed.

Conference on 'Hybrid Microelectronics' *(cont.)*

'Silver Transfer Through Thick Film Insulants'.
J. SAVAGE, *Atomic Weapons Research Establishment.*

HYBRID CIRCUIT APPLICATIONS

'Toward a High-Definition Visual Prosthesis'.
P. E. K. DONALDSON and E. SAYER, *M.R.C. Neurological Prostheses Unit.*

'Power Hybrid Thick Film Modules'.
W. G. ASHMAN, *Westinghouse Brake & Signal Co.*

'Hybrid Circuits in 120 Mbit/s Repeaters'.
D. S. LARNER, *Plessey Telecommunications Research.*

'A High Level Fast Limiting Switch as a Hybrid Module'.
C. H. JONES, *Signals Research and Development Establishment.*

'Large Thick-Film Multilayer Substrates for DIL-Packages'.
DR. H. DANIELSSON, *Saab-Scania AB, Sweden.*

'A Thick Film Circuit for a Guided Weapon Fin Control'.
A. BENNETT, *British Aircraft Corporation.*

'A Hybrid Module for the Computerized Monitoring of Electronic Equipment'.
DR. P. L. KIRBY, *Welwyn Electric.*

'The Thick Film Aspects of a High Thermal Dissipation Inter connection System for Use with Integrated Circuits'.
K. C. BINGHAM and Y. GURLER, *ICL.*

'Hybrid Forms of Construction for Computer Circuitry'.
K. C. BINGHAM and R. NAYLOR, *ICL.*

'Signal Interconnection Network for Integrated Circuit Chips'.
K. C. BINGHAM and A. G. A. GILLINGHAM, *ICL.*

Closing remarks by P. E. K. Donaldson, Chairman of the Conference Papers Committee.

Conference on

'Hybrid Microelectronics'

Organized by THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS with the association of The International Society for Hybrid Microelectronics (U.K.), The Institution of Electrical Engineers, The Institute of Physics and The Institute of Electrical and Electronics Engineers.

University of Kent at Canterbury, 25th to 27th September 1973

Hybrid techniques provide an invaluable complement to monolithic integrated circuits in the making of microelectronic equipment, since they are applicable in just those cases where monoliths are uneconomic or unusable. Microcircuits which do something other than the most standard functions, or are required to run from peculiar supply voltages or at microwave frequency, or are required quickly in moderate numbers only, are best realized by hybrid methods.

PROVISIONAL PROGRAMME

Tuesday, 25th September

Welcome by Professor J. C. Anderson, Chairman of the Conference Organizing Committee.

Opening Address by Dr. G. E. Weibel, Zenith Radio Corporation, U.S.A.

PRODUCTION AND ECONOMICS

'Technological and Economic Aspects in Producing Hybrid Circuits'.
Dr. G. Krüger, *Robert Bosch GmbH, Germany.*

'Hybrid Thick Film Circuits in the Telecommunication Industry'.
P. Moore, *Erie Electronics Ltd.*

DEPOSITION AND MATERIALS

'Improvements of Plasma Spraying Processes for Hybrid Microelectronics'.
M. Braguier, J. Bejat, R. Tueta, *C.N.E.T.* and M. Verna, G. Aubin, C. Naturel, *Desmarquest & C.E.C.*

'ZrB_2—A New Thin Film Resistor Material for Hybrid Microwave Technology'.
Dr. J. T. Calow, Dr. K. G. Knauff and G. Lütteke, *Philips, Aachen, Germany.*

'A New Low Cost Thick Film Resistor'.
A. S. Laurie, *Electrical Research Association.*

'Improved, Glass–Ceramic, Thick Film Capacitors'.
I. G. Bowkley, *Electrical Research Association.*

ADHESION AND BONDING

'Degradation of Thick Film Conductor Adhesion'.
A. C. Buckthorpe, *ITT Components Group.*

'Conducting Polymers for Die Bonding of Semiconductors to Thick Film Circuits'.
M. G. Harwood, J. Schofield and M. V. Burrell, *Mullard Mitcham.*

'The Feasibility of Ultrasonic Welding Flip Chips to Thick Film Conductors'.
K. I. Johnson, M. H. Scott, W. Battarbee and R. Griggs, *Welding Institute.*

'Low Temperature Methods of Attachment of Active Devices to Film Circuits'.
D. Hetherington and C. Marshall, *Newmarket Transistors.*

Wednesday, 26th September

THERMAL DESIGN

'Microcircuit Thermal Design Simplified by the Superposition Principle'.
D. J. Dean, *Atomic Weapons Research Establishment.*

'Thermal Considerations in Hybrid Constructions'.
J. T. Hughes and R. Naylor, *ICL.*

HIGH FREQUENCY INCLUDING MICROWAVE

'The Effect of Some Process and Material Variables on the Properties of Thick Film Microwave Transmission Lines'.
K. W. Woodcock and Dr. P. G. Barnwell, *Brighton Polytechnic.*

'The Radio Frequency Properties of Thick Film Components'.
Dr. P. G. Barnwell, *Brighton Polytechnic* and C. E. Tucker, *Brighton Polytechnic* and *Feedback Instruments.*

'Tunable Thick Film Circuits at Ultra High Frequencies'.
Dr. T. L. Harcombe, *Glamorgan Polytechnic.*

'Thick Film Oscillators in the 150–200 MHz Range with Thermostatic Control'.
D. P. Heywood, *Royal Aircraft Establishment.*

'Thick Film Technique for Hybrid Integrated Microwave Circuits'.
W. Funk and W. Schilz, *Philips, Hamburg.*

'Programmable Surface Acoustic Wave Devices Utilizing Hybrid Microelectronic Techniques'.
R. D. Lambert, P. M. Grant, D. P. Morgan and Professor J. H. Collins, *University of Edinburgh.*

'The Application of Lumped Element Techniques to High Frequency Hybrid Integrated Circuits'.
R. Chaddock, *Mullard Southampton.*

'Low Noise Hybrid Microwave Amplifier and Mixer Using Thick Film Techniques'.
R. J. Robertson and Dr. P. L. Bainbridge, *Dundee College of Technology.*

'Hybrid Thin-Film Technology for Microwave Integrated Circuits'.
M. Pilgrim and K. Winterbottom, *P.O. Telecommunications HQ.*

Thursday, 27th September

FAILURE AND RELIABILITY

'Evaluation Methods for the Examination of Thick Film Materials'.
M. V. Coleman, *Standard Telecommunications Laboratories.*

'The Production and Reliability Problems of Very Small Screen Printed Resistors'.
Miss J. B. McCloghrie and Dr. P. J. Holmes, *Royal Aircraft Establishment.*

'The Designers Need for Information on the Reliability of Hybrid Microcircuits'.
Miss P. S. Shove, *Admiralty Surface Weapons Establishment.*

'T.H.I. Measurements on Printed and Sprayed Thick Films'.
Professor J. C. Anderson, R. T. Smyth and E. J. Weidmann, *Imperial College of Science & Technology.*

(*cont. on opposite page*)

Industrial News

Another Golden Jubilee in the Electronics Industry

Hard on the heels of the 50th anniversary of Belling & Lee Limited, reported in the May 1973 Journal, another well known electronic component manufacturer has celebrated 50 years of existence. A. F. Bulgin & Co. Ltd. were founded as a private limited company in 1923 by Mr. A. F. Bulgin, O.B.E. (Fellow), in a small factory in Abbey Road, Barking. In 1939 the Company moved to a large new factory in Bye Pass Road, Barking, and during the war produced ten million components for all branches of the Armed Forces. With the return to peacetime activities, the Company's expansion continued and it went Public with a capital of £100 000 in 1948. Its capital today is £1 000 000 and it has subsequently extended its premises on the same site.

Bulgin's have specialized in such components as plugs, sockets, switches, valve holders, lamps, indicators and knobs of all types and their Research and Development Division includes a laboratory where products can be tested to meet all British and European standards.

The Company's founder, Mr. A. F. Bulgin, has continued as Chairman and Managing Director and he played a leading part in the formation in 1933 of the Radio and Electronic Components Manufacturers Federation. He was for many years the Chairman of the Federation's Exhibition Committee which organizes the Electronic Components Exhibition. Elected a Fellow of the Institution in 1943, he served for several years on the Membership Committee.

Fire Losses in British Engineering Industry

Over £9M was lost in fires in the British engineering and electrical industry last year, reports the Fire Protection Association. Provisional estimates reveal that there were 115 fires costing over £10 000 each during 1972—an average loss of £80 000 per fire. According to the FPA, the worst hit sectors of the industry were electrical engineering with a fire bill of over £2·6M and mechanical engineering firms with losses of £2·5M. These losses represent the direct cost of property damage and do not include the possibly more expensive losses due to business interruption.

Among the factors that contribute to these expensive losses is the use of hazardous plant in inadequately-protected areas, the large amounts of combustible packaging material used and high value of the property and equipment at risk. But the industry is at last taking some steps to improve its bad record. Fire safety is an integral part of the current safety campaign being conducted by the Engineering Employers' Federation, and a specific guide on fire control has been produced. Advice and information can also be obtained from the Fire Protection Association, Aldermary House, Queen Street, London EC4N 1TJ.

Queen's Awards to Industry for 1973

An Award for technological achievement has been made to the English Electric Valve Company in recognition of technological innovation in respect of ceramic hydrogen Thyratrons. In 1968 EEV first received The Queen's Award to Industry for technological achievement in respect of the image isocon television camera tube.

The Central Research Laboratories of EMI Limited have won an Award for the EMI-SCANNER—the company's computerized X-ray system for the diagnosis of brain disorders (described in the January-February 1973 Journal).

Among the Awards for outstanding export achievement are EASAMS Limited, an independent systems engineering and consultancy organization which is a member of GEC-Marconi Electronics Limited. This award is the first ever to be awarded to a pure consultancy organization in its particular field.

The first specialist data communications company to win a Queen's Award is Racal-Milgo. The company is active in the field of high-speed data transmission and has been instrumental in the achievement of the most complex international data communications networks in service today. Currently some 77% of its business is derived from overseas.

Sonicaid Limited of Bognor Regis, Sussex, which specializes in the design and manufacture of ultrasonic medical and industrial equipment, has also won an Award for outstanding export achievement.

THE ERIC MEGAW MEMORIAL AWARD

This year's competition sponsored by the Northern Ireland Section of the Institution for the Eric Megaw Memorial Award attracted an encouragingly high attendance at the Ashby Institute of the Queen's University of Belfast on 13th March last. The Department of Electrical Engineering of the University put forward six papers for presentation by their authors and adjudication was carried out by four judges. They were: Mr. D. Bromley, F.I.E.E. (Director of Studies, School of Electrical and Electronic Engineering, Ulster College); Mr. B. Slamin, M.I.E.R.E. (Head of Programme Services and Engineering, BBC, Northern Ireland); Mr. J. McC. Foye, M.I.E.E., M.I.M.C. (Head of Systems Department, Missile Systems Division, Short Brothers & Harland Limited); and Captain A. W. Allen, R.N., M.I.E.R.E. (Chairman, Northern Ireland Section of the Institution).

The prize winning paper was unanimously chosen by the adjudicators to be 'The Design, Construction and Application to HV Technology of Vari-focal Concave Mirror Cameras' by Mr. L. Bryans. In their report, the judges said that apart from excellent content of his paper, Mr. Bryans made the most of the opportunity to present his work in an articulate and at times humorous manner, normally found in more mature speakers.

The other five papers were:

'Automatic registering of a telephone caller's number' by H. H. Topping.

'Induction motor speed control by stator current modulation' by A. O'Donnell.

'Arithmetic processing of data originating in unit distance code format' by J. N. Magee.

'Quadraphonics' by J. Lavery.

'Stepping motor control for use in the fabrication of integrated circuit masks' by B. T. D. Smith.

The Award was established three years ago by the Institution's Northern Ireland Section to commemorate the memory of an eminent Ulster Radio Engineer, E. C. S. Megaw, M.B.E., D.Sc., who was a graduate of the Queen's University of Belfast. Dr. Megaw was on the Scientific Staff of the Research Laboratories of the General Electric Company from 1931 to 1946; for ten years prior to his death in 1956 he was with the Royal Naval Scientific Service, latterly as Director of Physical Research.

Forthcoming Conferences

The Electronics Industry and its Interface with Higher Education

The IERE is arranging a weekend conference on the above theme which will be held at the Royal Holloway College (University of London), Egham, Surrey, from Friday, 29th March to Sunday, 31st March, 1974. The Institution of Electrical Engineers is associated as a co-sponsor. The programme is now nearly completed and further details, including registration arrangements, will be published shortly.

International Conference on Radar

An international conference entitled 'Radar–present and future' is to be held in London from 23rd to 25th October 1973. Its aim is to provide a forum for the interchange of information on developments and probable trends in components, subsystems and complete radar systems design and performance.

The conference is being organized by the Electronics Division of the Institution of Electrical Engineers in association with the IERE and the Associazione Elettrotecnica ed Elettronica Italiana, the Institute of Electrical and Electronics Engineers (United Kingdom and Republic of Ireland Section), the Institute of Mathematics and its Applications, and the Société des Electriciens, des Electroniciens et des Radioélectriciens.

Many of the papers will relate to radar aspects associated with signal processing where the trend is to take advantage of modern digital techniques to reduce the impact of clutter. Papers will cover clutter and target characteristics, m. t. i. systems and plot extraction and tracking. Computers and computer technology play an important part especially in the field of extraction and tracking.

Other papers will be on steerable arrays and antennas and a session on low-angle tracking will be of particular interest to radar engineers involved in tracking systems, as papers will cover the mulitpath effects in primary and secondary radar systems.

Provisional programmes, registration forms, and further details are available from the Manager, Conference Department, IEE, Savoy Place, London WC2R 0BL.

Digital Instrumentation

A Conference on 'Digital Instrumentation' is to be held in London from 12th to 14th November 1973. Organized by the Electronics Division of the Institution of Electrical Engineers in association with the IERE and the Institute of Measurement and Control, the Conference aims to show the most significant techniques now available for digital instrumentation, and to illustrate their practical and possible uses.

Topics to be covered include:

Digital Application

Applications in which digital techniques are particularly relevant.

Frequency, Generation and Analysis: waveform synthesis, noise generators, spectrum analysers and correlators, counters, signal and pattern generators, phase and group delay.

Signal Magnitude Measurement and Generation, Multiplexing and sampling: convertors–analogue to digital and digital to analogue, pulse height analysers, integral measurements, chromatography.

Signal Processing: data acquisition, filters, laplace, Walsh and Fourier transform machines, algorithms for data conversion.

Transducers: digital encoders, distance gauges, sonic gauges.

Further information from the IEE, Conference Department, Savoy Place, London WC2R 0BL.

Computer Aided Design

A call for papers for the 1974 International Conference on Computer Aided Design has been issued. Organized by the Control and Automation, and Electronics Divisions of the Institution of Electrical Engineers with the association of the IERE and other professional bodies, the conference will be held from 8th–11th April 1974 at the University of Southampton. It is intended to discuss experience in applying computer aided design techniques to civil, mechanical and electrical engineering and to introduce new techniques.

Those wishing to offer contributions should send a synopsis of not more than 300 words to the IEE Conference Department Savoy Place, London WC2R 0BL by 3rd September 1973.

An associated exhibition, to be organized by the Electronic Engineering Association will be held concurrently with the conference, and details on participation in the Exhibition may be obtained from Mr F. A. Sherry, EEA, Leicester House, 8 Leicester Street, London WC2H 7BN

Symposium on Electromagnetic Wave Theory

A Symposium on Electromagnetic Wave Theory, sponsored by the Union Radio-Scientifique Internationale (URSI), is being held in London at the invitation of the Royal Society and the Institution of Electrical Engineers, from 9th to 12th July 1974 at the Imperial College of Science and Technology. The theme will be the effective use of computer methods in electromagnetic wave theory; at least one full session will be devoted exclusively to this topic. A number of speakers will be invited to present reviews in this area.

The programme will cover, as in previous electromagnetic wave theory symposia, the areas of electromagnetic wave propagation and antennas. Special attention will be given to: propagation under the earth's surface; open resonators; propagation in multi-conductor systems; optical waveguides; and phased arrays.

While the programme is devoted to electromagnetic waves, it is also considered appropriate to include acoustic surface wave propagation because of its similarity to electromagnetic wave propagation. The main emphasis of the symposium, as in previous years, will be on theoretical considerations, but papers with experimental results will be welcomed, particularly if they contribute to a better understanding of the theory.

Those wishing to offer contributions should submit abstracts of not more than three A4 pages (including illustrations) to the Conference Department, Institution of Electrical Engineers, Savoy Place, London, WC2R 0BL by December 1973.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 24th May and 12th June 1973 recommended to the Council the election and transfer of 87 candidates to Corporate Membership of the Institution and the election and transfer of 11 candidates to Graduateship and Associateship. In accordance with Bye-law 21, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details.

Meeting: 24th May 1973 (Membership Approval List No. 159)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

TIMMS, Robert Warren *Kirkcaldy, Fife.*

Transfer from Graduate to Member

DAWSON, John Graham *Claygate, Surrey.*
ECCLESTON, Stanley *Waterlooville, Hampshire.*
EDIRIWEERA, Dudley Pemasiri *North Wembley, Middlesex.*
EDWARDS, Arthur Raymond *Hinckley, Leicester*
GRIMSDALE, David Christopher Ian, Lieutenant RN *St. Germans, Saltash, Cornwall.*
HINCH, Terence Malcolm Harry *Lutterworth, Warwickshire.*
LANGLOIS, David *London, S.E.10.*
LEWIS, Michael Penrose *Watford, Hertfordshire.*
McCLUSKEY, Thomas Edward *Cosham, Hampshire.*
MACRAE, John Alick, Lieutenant RN *Woolston, Hampshire.*
MILLWARD, Alan George *Coventry, Warwickshire.*
NEWLAND, Anthony Thomas John *Braintree, Essex.*
PATTERSON, Girvan Leigh *Coventry, Warwickshire.*
PORTER, Frederick William *Moelfre, Anglesey.*
PRITCHARD, Robert, Flight Lieutenant RAF *Lyneham, Wiltshire.*
ROBINSON, William *Scunthorpe, Lincolnshire.*
WRIGHT, Raymond John, Lieutenant RN *Gosport, Hampshire.*

Transfer from Student to Member

BORSTNIK, Bozo *Scunthorpe, Lincolnshire.*

Direct Election to Member

EMBREY, Derek Morris *Cardiff, Glamorganshire.*
GEORGE, Peter Alan *Harpole, Northampton.*
GRIFFITHS, Kenneth Stuart, B.Sc.(Eng.), Major, REME *Arborfield, Berkshire.*
SANDMAN, Aubrey Max *Harlow, Essex.*
TAYLOR, Roy *Meopham, Kent.*
THOMPSON, Denis Harry, B.Sc. *Ramsey, Huntingdon.*
WARD, Jack Peter William, Lieutenant Commander, RN *Croydon, Surrey.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

CHIMONAS, Andreas Stylianou *London, N.4.*
KEMPTON, Michael Willis, B.Sc.(Eng.) *Ilford, Essex.*

STUDENTS REGISTERED

CAMBELL-CARR, Paul Martin St. John *Skerries, County Dublin.*
FITCH, John Lewis *Barnet, Hertfordshire.*
GASSON, Geoffrey Michael Hayward *Chelmsford, Essex.*
GOHIL, Hemat Lal *Cardiff.*
GRIMSHAW, Alan Paul *Ickenham, Uxbridge, Middlesex.*
NEWBY, Paul Leslie *Ardleigh, Colchester, Essex.*
PYE, Kevin John *Highfields, Stafford.*
SALLAM, Gars *Watford, Hertfordshire.*
SHORROCK, David John *Newcastle upon Tyne.*
THORNLEY, Nigel *London, W.C.1.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Member to Fellow

WHEATLEY, Norman *Kingston, Jamaica.*

Transfer from Graduate to Member

KONG, Shi Wei *North Point, Hong Kong.*
LEES, Frank Philip *Salisbury, Rhodesia.*

Direct Election to Member

ADEBANJO, Adebola Adebayo *Odunlami, Lagos, Nigeria.*
METCALFE, Christopher Shaun *Manly, N.S.W., Australia.*
WRIGHT, Anthony Frederick *Ottawa 5, Canada.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

ACHUTHAN, C. M. *Gopeng, Perak, Malaysia.*
LEE, Joo Tim *Singapore 20.*

STUDENTS REGISTERED

FAKHOURI, Salib Yassa *Khartoum, Sudan.*
HUI, Chuen-Shun *Kowloon, Hong Kong.*
NG, Man Fai *Kowloon, Hong Kong.*
TAN, Peng Hieng *Singapore 12.*

Meeting: 12th June 1973 (Membership Approval List No. 160)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Direct Election to Fellow

den BRINKER, Carl Siegmund *Oakley, Bedford.*
SATCHELL, Edward William John *Bath, Somerset.*

Transfer from Gradute to Member

ALEXANDER, Roger James *Newport, Essex.*
BELENKIN, Andrew Michael *Colchester, Essex.*
BRAND, Norman James *Dunmow, Essex.*
BRIGGS, David Edward *Wickford, Essex.*
CHANDLER, Michael John Hayden *London, N.W.3.*
COLLINS, Peter Cyril *Newmarket, Suffolk.*
COLLIS, Ronald *South Harrow, Middlesex.*
CRANFIELD, Ronald Frederick *Portishead, Bristol.*
CROTCH, John Frederick *Kilmarnock, Ayrshire.*
DAVIES, Allan Geoffrey *Fareham, Hampshire.*
EDWARDS, Robert Arthur *Liverpool, 14.*
EVES, Peter Edward *London, N.21.*
FENSOME, David Arthur, B.Sc. *Harpenden, Hertfordshire.*
FULLER, Peter Ernest Noel *Reading, Berkshire.*
GALLOP, David Leslie *Reading, Berkshire.*
GARNER, David Edward *Sutton, Surrey.*
GILLARD, David Edward *Farnborough, Hampshire.*
GRIFFITHS, Kenneth Edward *Liverpool 11.*
GRIPTON, Edward William *London, N.9.*
HARMEN, Joseph William, M.Sc. *Frimley, Surrey.*
McDOWELL, Edward Hunter *Larne, County Antrim.*
MARLOW-MANN, Philip David George *Billericay, Essex.*
MASON, Graham Ansley *Wimborne, Dorset.*
PAPADOPULO, Herbert James *Wimborne, Dorset.*
PARTRIDGE, Peter James *Benfleet, Essex.*
PATIENCE, Brian Michael, M.Phil. *Eastbourne, Sussex.*
PEARSON, Neil Wilfred *Southall, Middlesex.*
SKERRY, Christopher Scott *Havant, Hampshire.*
SKEWS, Robert William *Rainham, Essex.*
SQUIRES, John Wallace *Bordon, Hampshire.*
STEVENS, Brian Leslie *Cowfold, Sussex.*
STEVENSON, James Watt *Cambusbarron, Stirling.*
TARN, Colin *Stevenage, Hertfordshire.*
THROWER, Keith Leonard, Captain, R.E.M.E. *Horndean, Hampshire.*
WARREN, John Howard, B.Sc.(Eng.) *Maidenhead, Berkshire.*
WAY, Trevor Arthur *Cowes, Isle of Wight.*
WRIGHT, David Oxford *Eastleigh, Hampshire.*

Transfer from Student to Member

MICHEL, Gregory Charles *Camberley, Surrey.*

Direct Election to Member

CALLAWAY, Trevor John, B.Tech. *Chandler's Ford, Hampshire.*
PAINTER, Robin Peter Douglas Francome *Broadstone, Dorset.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

OLUSANYA, Olajide Owodunni, B.Sc. *Bolton, Lancashire.*

Direct Election to Graduate

DELNEVO, John Frederick Victor, B.Sc. *Witney, Oxfordshire.*
HOLDEN, David Alan, B.Sc. *Sevenoaks, Kent.*
McMANUS, John *Hillington, Glasgow.*

STUDENTS REGISTERED

ASPROMOUGOS, Demetrios *London, W. 14.*
BARROWS, Lynn Eric *West Ewell, Surrey.*

OVERSEAS

Transfer from Graduate to Member

AHMED VAHIDY, Nayyar *Jeddah, Saudi Arabia.*
AJANI, Samuel Olukunle *Lagos, Nigeria.*
BEG, Nazim *Karachi 8, Pakistan.*
ENE, Amaku Efiong *Nassau, Bahamas.*
HUNTER, George *Kabwe, Zambia.*
JEFFREY, William *Hainhausen, West Germany.*
KWONG, Wun-Cho *Montreal, Canada.*
MACDONALD, Ewen Alistair *Bundoora, Victoria, Australia.*
PINCHEN, Anthony Patrick *Lasalle, P.Q., Canada.*
SMITH, Bruce Douglas *Ann Arbor, Michigan, U.S.A.*
SWEETMAN, James Michael, B.Sc., Captain, R. Signals *B.F.P.O. 37.*
WIGWE, Shyngle Iwogbo Joseph *Port Harcourt, Nigeria.*

Direct Election to Member

BASTIKAR, Arvind R. *Kanata, Ontario, Canada.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

KUMAR, Anil Pratap, B.Sc., *Boston, Massachusetts, U.S.A.*

Direct Election to Graduate

BAROLE, William, B.Sc. *Tokyo, Japan.*

Transfer from Student to Associate

LEUNG YINKO, Jean Leung Sec Yon *Mauritius.*

STUDENTS REGISTERED

CHEW, Boon Kheng *Bukit Mertajam, Malaysia.*
TEOH, Teng Huan *Ipoh, Perak, Malaysia.*

Notice is hereby given that the elections and transfers shown on Lists 155, 156 and 157 have now been confirmed by the Council.